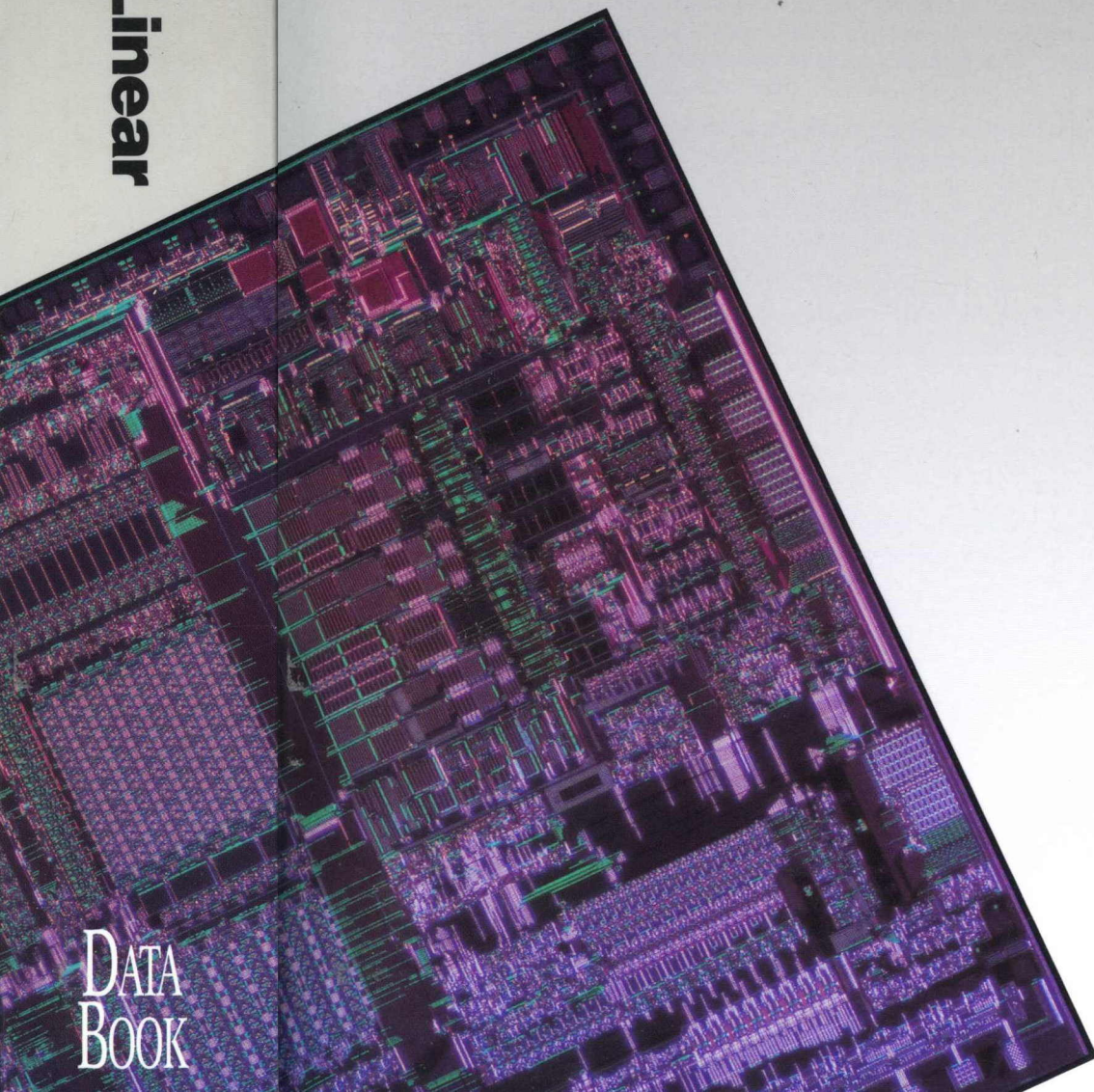


1993


Micro Linear

 **Micro Linear**
SEMI-STANDARD ANALOG



DATA
BOOK

1993
DATA BOOK

INTRODUCTION

Micro Linear Corporation, headquartered in San Jose, California, was founded in 1983 as an Analog USIC (user specified) integrated circuit manufacturer. The Corporation occupies 100,000 square feet of office and manufacturing space on six acres of land. The greater part of this space is used for state of the art engineering, test and wafer metallization facilities.

Micro Linear, a pioneer in analog semi-custom integrated circuits, continues to expand its catalog with a unique tile array design methodology. This advanced proprietary tile array approach allows custom designs, and standard products that can be quickly modified with minimum expense and risk to the end customer. The analog design skills and tile array methodology are strengths that separate Micro Linear from its competitors.

This standard product catalog has grown through the years to include more than 120 products and 25 arrays serving the following markets

- Data Communications
- Telecommunications
- Hard Disk Drive
- Motor Control
- Switch Mode Power Supply
- Data Acquisition
- Bus Products
- Mixed Signal USIC

Micro Linear is committed to supplying complex mixed signal integrated circuit solutions with the highest quality and best service possible.



SEMI-STANDARD INTEGRATED CIRCUITS

The ability to modify standard integrated circuit products offers a new dimension and power to users of integrated circuits. By modifying a standard product the customer gains a level of product optimization that otherwise would not exist. At the foundation of Micro Linear's business is this ability, along with the eagerness, to provide optimized integrated circuits to its customers. This is "Semi-Standard Integrated Circuits."

Semi-Standard can mean a change in the functionality of a device, an adjustment to a performance parameter, or even something as simple as a variation in the physical marking on the IC. The customer can define a change to one of Micro Linear's standard catalog products to gain a performance edge, power savings, or lower system cost. Micro Linear's goal is to provide the maximum amount of flexibility such that the customer can reap these benefits and be able to offer an improved end product for today's competitive environment.

Semi-Standard ICs offer a superior alternative over custom, and even semi-custom solutions. The level of technical risk is significantly reduced because only an incremental change is made to an already proven product. And faster time-to-market is enjoyed due to a combination of a shorter development time and the added benefit of being able to debug many of the system related issues by using the standard product.

Micro Linear offers this unique capability to enact functional changes to its standard products because of its Tile Array methodology. Tile Arrays are collections of uncommitted active and passive components arranged in patterns on an integrated circuit wafer. Circuits are implemented by designing the metal interconnect layers which are used in the final step of the wafer manufacturing process. The effectiveness of the Semi-Standard methodology is demonstrated in the fact that Micro Linear has been supplying very high volume standard products in fast moving dynamic markets since its inception.

Semi-Standard is a product. But it is also a way of doing business. The wherewithal to perform the customer changes is not enough. A successful supplier of Semi-Standard integrated circuits must also have the design methodology, mind set, and culture of Semi-Standard. Since Micro Linear was founded and structured as a supplier of semi-custom analog and analog/digital integrated solutions these traits are firmly rooted in the company.



Micro Linear Corporation Data Book 1993

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 **Micro Linear**

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1993 Product Data Book

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General Information

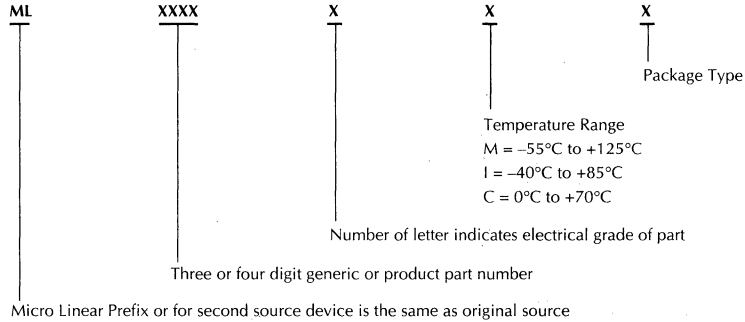
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Part Number and Package Type Explanation

PART NUMBER



PACKAGE TYPE

Letter Suffix	Description
D	Side Brazed Hermetic DIP
F	Flat Pack
J	Ceramic Hermetic DIP (CERDIP)
L	Ceramic Leadless Chip Carrier (LCC)
P	Plastic DIP
Q	Plastic Chip Carrier (PCC)
S	Small Outline (SOIC)
G	(PQFP) Plastic Quad Flat Pack
R	Shrink Small Outline Package (SSOP)
H	Thin Quad Flat Pack (TQFP)

Alternate Source Part Number

Analog Devices

Analog Devices Part Number	Micro Linear Direct Replacement ¹
AD7820BQ	ML2261CIJ
AD7820CQ	ML2261BIJ
AD7820KN	ML2261CCP
AD7820KP	ML2261CCQ
AD7820LN	ML2261BCP
AD7820LP	ML2261BCQ
AD7820TQ	ML2261CMJ
AD7820UQ	ML2261BMJ
AD7824BQ	ML2264CIJ
AD7824CQ	ML2264BIJ
AD7824KN	ML2264CCP
AD7824LN	ML2264BCP
AD7824TQ	ML2264CMJ
AD7824UQ	ML2264BMJ

Exar

Exar Part Number	Micro Linear Direct Replacement ¹
XR117R-2CP	ML117R-2CP
XR117R-4CP	ML117R-4CP
XR117R-4MD	ML117R-4CS
XR117R-6CJ	ML117R-6CQ
XR117R-6CP	ML117R-6CP
XR117R-6MD	ML117R-6CS
XR117-2CN	ML117-2CJ
XR117-2CP	ML117-2CP
XR117-2MD	ML117-2CS
XR117-4CN	ML117-4CJ
XR117-4CP	ML117-4CP
XR117-4MD	ML117-4CS
XR117-6CJ	ML117-6CQ
XR117-6CN	ML117-6CJ
XR117-6CP	ML117-6CP
XR117-6MD	ML117-6CS

Linear Technology

LTC Part Number	Micro Linear Direct Replacement ¹
LTC1060ACJ	ML2110BIJ ²
LTC1060ACN	ML2110BCP ²
LTC1060AMJ	ML2110BMJ ²
LTC1060CJ	ML2110CIJ ²
LTC1060CN	ML2110CCP ²
LTC1060MJ	ML2110CMJ ²

National Semiconductor

NSC Part Number	Micro Linear Direct Replacement ¹
ADC0808CJ	ADC0808CJ
	ML2258BMJ
ADC0808CCJ	ADC0808CCJ
	ML2258BIJ
ADC0808CCV	ADC0808CCV
	ML2258BIQ
ADC0809CCN	ADC0808CCN
	ML2258CIP
ADC0809CCV	ADC0809CCV
	ML2258CIQ
ADC0820BCJ	ML2261BIJ
ADC0820BCN	ML2261BCP
ADC0820BCV	ML2261BCQ
ADC0820CCJ	ML2261CIJ
ADC0820CCN	ML2261CCP
ADC0820CCV	ML2261CCQ
ADC0820CJ	ML2261CMJ
ADC0831BCJ	ADC0831BCJ
	ML2281BIJ
ADC0831BCN	ADC0831BCN
	ML2281BCP
ADC0831CCJ	ADC0831CCJ
	ML2281CIJ
ADC0831CCN	ADC0831CCN
	ML2281CCP
ADC0832BCJ	ADC0832BCJ
	ML2282BIJ
ADC0832BCN	ADC0832BCN
	ML2282BCP
ADC0832CCJ	ADC0832CCJ
	ML2282CIJ
ADC0832CCN	ADC0832CCN
	ML2282CCP
ADC0833BCJ	ADC0833BCJ
	ML2283BIJ
ADC0833BCN	ADC0833BCN
	ML2283BCP
ADC0833CCJ	ADC0833CCJ
	ML2283CIJ
ADC0833CCN	ADC0833CCN
	ML2283CCP
ADC0834BCJ	ADC0834BCJ
	ML2284BIJ
ADC0834BCN	ADC0834BCN
	ML2284BCP

Note 1. 100% pin-for-pin compatible with improved electrical specifications.

Note 2. Consult data sheet for electrical specifications that may vary from limit or conditions of alternate source.

Note 3. Alternate source ships -40°C to +85°C product as molded; Micro Linear does this on a customer need basis.

National Semiconductor (Continued)

NSC Part Number	Micro Linear Direct Replacement ¹
ADC0834CCJ	ADC0834CCJ ML2284CIJ
ADC0834CCN	ADC0834CCN ML2284CCP
ADC0838BCJ	ADC0838BCJ ML2288BIJ
ADC0838BCN	ADC0838BCN ML2288BCP
ADC0838BCV	ADC0838BCV ML2288BCQ
ADC0838CCJ	ADC0838CCJ ML2288CIJ
ADC0838CCN	ADC0838CCN ML2288CCP
ADC0838CCV	ADC0838CCV ML2288CCQ
ADC1061CIJ	ML2271CIJ
ADC1061CIN	ML2271CCP ³
ADC1061CIWM	ML2271CCS ³
ADC1061CMJ	ML2271CMJ
DP5016QC	ML501-6CQ
DP5016RQC	ML501R-6CQ
DP5018QC	ML501-8CQ
DP5018RQC	ML501R-8CQ
μA5016QC	ML501-6CQ
μA5016RQC	ML501R-6CQ
μA5018QC	ML501-8CQ
μA5018RQC	ML501R-8CQ
DP8464BN-3	ML8464B-3CP ²
DP8464BV-3	ML8464B-3CQ ²
DP8464BN-2	ML8464B-2CP ²
DP8464BV-2	ML8464B-2CQ ²
DP8464BN-2	ML8464B-2CP ²
DP8464BN-3	ML8464B-3CP ²
DP8464BV-2	ML8464B-2CQ ²
DP8464BV-3	ML8464B-3CQ ²
DP8468BTP-3	ML4568-3CQ ²
DP8468BTP-2	ML4568-2CQ ²
LMF100CCN	ML2111CCP
LMF100CCWM	ML2111CCS
MF10AJ	ML2110CMJ ²
MF10ACN	ML2110BCP ²
MF10CCJ	ML2110CIJ ²
MF10CCWM	ML2110CCS ²
MF10CCN	ML2110CCP ²

Silicon Systems, Inc.

SSI Part Number	Micro Linear Direct Replacement ¹
SSI 32P541-CH	ML541CQ
SSI 32P541-P	ML541CP
SSI 32P541A-CH	ML4042CQ
SSI 32P541A-P	ML4042CP
SSI 32P541B-CH	ML4042CQ
SSI 32P541B-P	ML4042CP
SSI 32R117R-2P	ML117R-2CP
SSI 32R117R-4F	ML117R-4CF
SSI 32R117R-4P	ML117R-4CP
SSI 32R117R-6F	ML117R-6CF
SSI 32R117R-6H	ML117R-6CQ
SSI 32R117R-6P	ML117R-6CP
SSI 32R117-2P	ML117-2CP
SSI 32R117-4F	ML117-4CF
SSI 32R117-4P	ML117-4CP
SSI 32R117-6F	ML117-6CF
SSI 32R117-6H	ML117-6CQ
SSI 32R117-6P	ML117-6CP
SSI 32R501R-6H	ML501R-6CQ
SSI 32R501R-8F	ML501R-8CF
SSI 32R501R-8H	ML501R-8CQ
SSI 32R501R-8P	ML501R-8CP
SSI 32R501-6H	ML501-6CQ
SSI 32R501-8F	ML501-8CF
SSI 32R501-8H	ML501-8CQ
SSI 32R501-8P	ML501-8CP
SSI 32R511R-4S	ML511R-4CS
SSI 32R511R-6H	ML511R-6CQ
SSI 32R511R-6P	ML511R-6CP
SSI 32R511R-6S	ML511R-6CS
SSI 32R511R-8H	ML511R-8CQ
SSI 32R511R-8P	ML511R-8CP
SSI 32R511R-8S	ML511R-8CS
SSI 32R511-4S	ML511-4CS
SSI 32R511-6H	ML511-6CQ
SSI 32R511-6P	ML511-6CP
SSI 32R511-6S	ML511-6CS
SSI 32R511-8H	ML511-8CQ
SSI 32R511-8P	ML511-8CP
SSI 32R511-8S	ML511-8CS

1

Note 1. 100% pin-for-pin compatible with improved electrical specifications.

Note 2. Consult data sheet for electrical specifications that may vary from limit or conditions of alternate source.

Note 3. Alternate source ships -40°C to +85°C product as molded; Micro Linear does this on a customer need basis.

Cross Reference Guide

Texas Instruments

NSC Part Number	Micro Linear Direct Replacement ¹
ADC0808MJ	ML2258BMJ
ADC0808FN	ML2258BCQ
ADC0808N	ML2258BIP
ADC0809FN	ML2258CCQ
ADC0809N	ML2258CIP
TLC0820ACN	ML2261CCP
TLC0820ACFN	ML2261CCQ
TLC0820BCN	ML2261BCP
TLC0820BCFN	ML2261BCQ
ADC0831ACP	ML2281CCP
ADC0831AIP	ML2281CIJ ³
ADC0831BCP	ML2281BCP
ADC0831BIP	ML2281BIJ ³
ADC0832ACP	ML2282CCP
ADC0832AIP	ML2282CIJ ³
ADC0832BCP	ML2282BCP
ADC0832BIP	ML2282BIJ ³
ADC0834ACN	ML2284CCP
ADC0834AIN	ML2284CIJ ³
ADC0834BCN	ML2284BCP
ADC0834BIN	ML2284BIJ ³
ADC0838ACN	ML2288CCP
ADC0838AIN	ML2288CIJ ³
ADC0838CCN	ML2288BCP
ADC0838BIN	ML2288BIJ ³

VTC

VTC Part Number	Micro Linear Direct Replacement ¹
VM117-2DK	ML117-2CJ
VM117-2PK	ML117-2CP
VM117-4FK	ML117-4CF
VM117-4PK	ML117-4CP
VM117-4DK	ML117-4CJ
VM117-6PK	ML117-6CP
VM117-6DK	ML117-6CJ
VM117-6PK	ML117-6CP
VM117-6PLK	ML117-6CQ
VM117R-2DK	ML117R-2CJ
VM117R-2PK	ML117R-2CP
VM117R-4FK	ML117R-4CF
VM117R-4PK	ML117R-4CP
VM117R-4DK	ML117R-4CJ
VM117R-6DK	ML117R-6CJ
VM117R-6PK	ML117R-6CP
VM117-6PLK	ML117R-6CQ
VM217-6PK	ML501-6CP
VM217-6PLK	ML501-6CQ
VM217-8PK	ML501-8CP
VM217-8PLK	ML501-8CQ

Unitrode

Unitrode Part Number	Micro Linear Direct Replacement ¹
UC1823J	ML4823MJ
UC1825J	ML4825MJ
UC2823N	ML4823IP
UC2823Q	ML4823IQ
UC2825N	ML4825IP
UC2825Q	ML4825IQ
UC3823N	ML4823IP
UC3823Q	ML4823CQ
UC3825N	ML4825IP
UC3825Q	ML4825CQ

Note 1. 100% pin-for-pin compatible with improved electrical specifications.

Note 2. Consult data sheet for electrical specifications that may vary from limit or conditions of alternate source.

Note 3. Alternate source ships -40°C to +85°C product as molded; Micro Linear does this on a customer need basis.

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A/D Converters, D/A Converters

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Part Number	Resolution (Bits)	Non Linearity (Max LSB)	Dynamic Performance Signal to Noise Ratio	Conversion Time (μ s)	Power Supplies (V)	Temperature Range ¹			Package	Comments
						C	I	M		
ADC0808	8	$\pm 1/2$		6.6	5 ($\pm 10\%$)	X	X	X	28-Pin DIP, 28-Pin PCC	μ P Comp, 8-CH
ADC0809	8	± 1		6.6	5 ($\pm 10\%$)	X	X	X	28-Pin DIP, 28-Pin PCC	μ P Comp, 8-CH
ADC0831B	8	$\pm 1/2$		6.0	5 ($\pm 10\%$)	X	X	X	8-Pin DIP	Serial I/O, Single CH
ADC0831C	8	± 1		6.0	5 ($\pm 10\%$)	X	X	X	8-Pin DIP	Serial I/O, Single CH
ADC0832B	8	$\pm 1/2$		6.0	5 ($\pm 10\%$)	X	X	X	8-Pin DIP	Serial I/O, 2-CH
ADC0832C	8	± 1		6.0	5 ($\pm 10\%$)	X	X	X	8-Pin DIP	Serial I/O, 2-CH
ADC0833B	8	$\pm 1/2$		6.0	5 ($\pm 10\%$)	X	X	X	14-Pin DIP	Serial I/O, 4-CH
ADC0833C	8	± 1		6.0	5 ($\pm 10\%$)	X	X	X	14-Pin DIP	Serial I/O, 4-CH
ADC0834B	8	$\pm 1/2$		6.0	5 ($\pm 10\%$)	X	X	X	14-Pin DIP	Serial I/O, 4-CH
ADC0834C	8	± 1		6.0	5 ($\pm 10\%$)	X	X	X	14-Pin DIP	Serial I/O, 4-CH
ADC0838B	8	$\pm 1/2$		6.0	5 ($\pm 10\%$)	X	X	X	20-Pin DIP, 20-Pin PCC	Serial I/O, 8-CH
ADC0838C	8	± 1		6.0	5 ($\pm 10\%$)	X	X	X	20-Pin DIP, 20-Pin PCC	Serial I/O, 8-CH
ML2200B	12 + Sign	$\pm 3/4$	12kHz, ± 2.5 V SINE, S/N 72dB	31.5	± 5 ($\pm 5\%$)	X			40-Pin DIP	4-CH Data Acq Peripheral
ML2200C	12 + Sign	± 1	12kHz, ± 2.5 V SINE, S/N 72dB	31.5	± 5 ($\pm 5\%$)	X			40-Pin DIP	4-CH Data Acq Peripheral
ML2200D	12 + Sign	± 1	8.5kHz, ± 2.5 V SINE, S/N 72dB	44	± 5 ($\pm 5\%$)	X			40-Pin DIP	4-CH Data Acq Peripheral
ML2208B	12 + Sign	$\pm 3/4$	12kHz, ± 2.5 V SINE, S/N 72dB	31.5	± 5 ($\pm 5\%$)	X			40-Pin DIP	8-CH Data Acq Peripheral
ML2208C	12 + Sign	± 1	12kHz, ± 2.5 V SINE, S/N 72dB	31.5	± 5 ($\pm 5\%$)	X			40-Pin DIP	8-CH Data Acq Peripheral
ML2208D	12 + Sign	± 1	8.5kHz, ± 2.5 V SINE, S/N 72dB	44	± 5 ($\pm 5\%$)	X			40-Pin DIP	8-CH Data Acq Peripheral
ML2221B*	12 + Sign	$\pm 3/4$	8.5kHz, ± 5.0 V SINE, S/N 72dB	44	± 5 ($\pm 5\%$)	X	X		16-Pin DIP, 20-Pin PCC	Serial, S.P.I.
ML2221C*	12 + Sign	± 1	8.5kHz, ± 5.0 V SINE, S/N 72dB	44	± 5 ($\pm 5\%$)	X	X		16-Pin DIP, 20-Pin PCC	Serial, S.P.I.
ML2222B*	12 + Sign	$\pm 3/4$	12kHz, ± 2.5 V SINE, S/N 72dB	35	± 5 ($\pm 5\%$)	X	X		16-Pin DIP, 20-Pin PCC	Serial, CODEC
ML2222C*	12 + Sign	± 1	12kHz, ± 2.5 V SINE, S/N 72dB	35	± 5 ($\pm 5\%$)	X	X		16-Pin DIP, 20-Pin PCC	Serial, CODEC

* Future Products

Note 1. Temperature Range:

C = 0°C to +70°C, I = -40°C to +85°C, M = -55°C to +125°C

A/D Converters

Part Number	Resolution (Bits)	Non Linearity (Max LSB)	Dynamic Performance Signal to Noise Ratio	Conversion Time (μ s)	Power Supplies (V)	Temperature Range ¹			Package	Comments
						C	I	M		
ML2223B*	12 + Sign	$\pm 3/4$	8.5kHz, ± 5.0 V SINE, S/N 72dB	44	± 5 ($\pm 5\%$)	X	X		16-Pin DIP, 20-Pin PCC	Serial, ASYNC
ML2223C*	12 + Sign	± 1	8.5kHz, ± 5.0 V SINE, S/N 72dB	44	± 5 ($\pm 5\%$)	X	X		16-Pin DIP, 20-Pin PCC	Serial, ASYNC
ML2230B	12 + Sign	$\pm 3/4$	12kHz, ± 2.5 V SINE, S/N 72dB	31.5	± 5 ($\pm 5\%$)	X			24-Pin DIP,	μ P Comp, 8-Bit Bus
ML2230C	12 + Sign	± 1	12kHz, ± 2.5 V SINE, S/N 72dB	31.5	± 5 ($\pm 5\%$)	X			24-Pin DIP,	μ P Comp, 8-Bit Bus
ML2230D	12 + Sign	± 1	8.5kHz, ± 2.5 V SINE, S/N 72dB	44	± 5 ($\pm 5\%$)	X			24-Pin DIP,	μ P Comp, 8-Bit Bus
ML2233B	12 + Sign	$\pm 3/4$	12kHz, ± 2.5 V SINE, S/N 72dB	31.5	± 5 ($\pm 5\%$)	X			28-Pin DIP,	μ P Comp, 16-Bit Bus
ML2233C	12 + Sign	± 1	12kHz, ± 2.5 V SINE, S/N 72dB	31.5	± 5 ($\pm 5\%$)	X			28-Pin DIP,	μ P Comp, 16-Bit Bus
ML2233D	12 + Sign	± 1	8.5kHz, ± 2.5 V SINE, S/N 72dB	44	± 5 ($\pm 5\%$)	X			28-Pin DIP,	μ P Comp, 16-Bit Bus
ML2252B	8	$\pm 1/2$	51kHz, 5V SINE, S/N 47dB	6.6	5 ($\pm 10\%$)	X	X	X	20-Pin DIP, 20-Pin PCC	μ P Comp, 2-CH
ML2252C	8	± 1	51kHz, 5V SINE, S/N 47dB	6.6	5 ($\pm 10\%$)	X	X	X	20-Pin DIP, 20-Pin PCC	μ P Comp, 2-CH
ML2258B	8	$\pm 1/2$	51kHz, 5V SINE, S/N 47dB	6.6	5 ($\pm 10\%$)	X	X	X	20-Pin DIP, 20-Pin PCC	μ P Comp, 8-CH
ML2258C	8	± 1	51kHz, 5V SINE, S/N 47dB	6.6	5 ($\pm 10\%$)	X	X	X	20-Pin DIP, 20-Pin PCC	μ P Comp, 8-CH
ML2259B	8	$\pm 1/2$	51kHz, 5V SINE, S/N 47dB	6.6	5 ($\pm 10\%$)	X	X	X	28-Pin DIP, 28-Pin PCC	μ P Comp, 8-CH
ML2259C	8	± 1	51kHz, 5V SINE, S/N 47dB	6.6	5 ($\pm 10\%$)	X	X	X	28-Pin DIP, 28-Pin PCC	μ P Comp, 8-CH
ML2261B	8	$\pm 1/2$	250kHz, 5V SINE, S/N 48dB	.67	5 ($\pm 5\%$)	X	X		20-Pin DIP, 20-Pin PCC	μ P Comp, RD/WR
ML2261C	8	± 1	250kHz, 5V SINE, S/N 48dB	.67	5 ($\pm 5\%$)	X	X		20-Pin DIP, 20-Pin PCC	μ P Comp, RD/WR
ML2264B	8	$\pm 1/2$	250kHz, 5V SINE, S/N 48dB	.68	5 ($\pm 5\%$)	X	X	X	24-Pin DIP, 24-Pin SOIC	μ P Comp, RD/WR, 4-CH Mux
ML2264C	8	± 1	250kHz, 5V SINE, S/N 48dB	.68	5 ($\pm 5\%$)	X	X	X	24-Pin DIP, 24-Pin SOIC	μ P Comp, RD/WR, 4-CH Mux
ML2271B*	10	$\pm 1/2$	150kHz, 5V SINE, S/N 60dB	1.5	5 ($\pm 5\%$)	X	X	X	20-Pin DIP, 20-Pin SOIC	μ P Comp, RD/WR
ML2271C*	10	± 1	150kHz, 5V SINE, S/N 60dB	1.5	5 ($\pm 5\%$)	X	X	X	20-Pin DIP, 20-Pin SOIC	μ P Comp, RD/WR
ML2280B	8	$\pm 1/2$	51kHz, 5V SINE, S/N 47dB	6.6	5 ($\pm 10\%$)	X	X	X	8-Pin DIP	Serial I/O, Single CH
ML2280C	8	± 1	51kHz, 5V SINE, S/N 47dB	6.6	5 ($\pm 10\%$)	X	X	X	8-Pin DIP	Serial I/O, Single CH
ML2281B	8	$\pm 1/2$	51kHz, 5V SINE, S/N 47dB	6.6	5 ($\pm 10\%$)	X	X	X	8-Pin DIP	Serial I/O, Single CH
ML2281C	8	± 1	51kHz, 5V SINE, S/N 47dB	6.6	5 ($\pm 10\%$)	X	X	X	8-Pin DIP	Serial I/O, Single CH
ML2282B	8	$\pm 1/2$	47.5kHz, 5V SINE, S/N 47dB	6.6	5 ($\pm 10\%$)	X	X	X	8-Pin DIP	Serial I/O, 2-CH
ML2282C	8	± 1	47.5kHz, 5V SINE, S/N 47dB	6.6	5 ($\pm 10\%$)	X	X	X	8-Pin DIP	Serial I/O, 2-CH

* Future Products

Note 1. Temperature Range:

C = 0°C to +70°C, I = -40°C to +85°C, M = -55°C to +125°C

Part Number	Resolution (Bits)	Non Linearity (Max LSB)	Dynamic Performance Signal to Noise Ratio	Conversion Time (μ s)	Power Supplies (V)	Temperature Range ¹			Package	Comments
						C	I	M		
ML2283B	8	$\pm 1/2$	36.5kHz, 5V SINE, S/N 47dB	6.6	5 ($\pm 10\%$)	X	X	X	14-Pin DIP	Serial I/O, 4-CH
ML2283C	8	± 1	36.5kHz, 5V SINE, S/N 47dB	6.6	5 ($\pm 10\%$)	X	X	X	14-Pin DIP	Serial I/O, 4-CH
ML2284B	8	$\pm 1/2$	39kHz, 5V SINE, S/N 47dB	6.6	5 ($\pm 10\%$)	X	X	X	14-Pin DIP	Serial I/O, 4-CH
ML2284C	8	± 1	39kHz, 5V SINE, S/N 47dB	6.6	5 ($\pm 10\%$)	X	X	X	14-Pin DIP	Serial I/O, 4-CH
ML2288B	8	$\pm 1/2$	36.5kHz, 5V SINE, S/N 47dB	6.6	5 ($\pm 10\%$)	X	X	X	20-Pin DIP, 20-Pin PCC	Serial I/O, 8-CH
ML2288C	8	± 1	36.5kHz, 5V SINE, S/N 47dB	6.6	5 ($\pm 10\%$)	X	X	X	20-Pin DIP, 20-Pin PCC	Serial I/O, 8-CH

* Future Products

Note 1. Temperature Range:

C = 0°C to +70°C, I = -40°C to +85°C, M = -55°C to +125°C

2

Selection Guide

Part Number	Resolution (Bits)	Non Linearity (Max LSB)	Settling Time (μ s Max)	Power Supplies (V)	Reference (V)	Output Voltage (V)	Temperature Range			Package	Comments
							C	I	M		
ML2340B	8 (11-bits with gain ranging)	$\pm 1/4$	5	Single 5 or 12, dual ± 5	2.25 or 4.50	Rail-to-rail, bipolar, unipolar	X	X	X	18-pin DIP 18-pin SOIC	Flow through, or single buffered data
ML2340C	8 (11-bits with gain ranging)	$\pm 1/2$	5	Single 5 or 12, dual ± 5	2.25 or 4.50	Rail-to-rail, bipolar, unipolar	X	X	X	18-pin DIP 18-pin SOIC	Flow through, or single buffered data
ML2341B	8 (11-bits with gain ranging)	$\pm 1/4$	5	Single 5 or 12, dual ± 5	2.25 or 4.50	Rail-to-rail, bipolar, unipolar	X	X	X	20-pin DIP 20-pin PCC	Double or single buffered data
ML2341C	8 (11-bits with gain ranging)	$\pm 1/2$	5	Single 5 or 12, dual ± 5	2.25 or 4.50	Rail-to-rail, bipolar, unipolar	X	X	X	20-pin DIP 20-pin PCC	Double or single buffered data
ML2350B	8 (11-bits with gain ranging)	$\pm 1/4$	5	Single 5 or 12, dual ± 5	2.50 or 5.00	Rail-to-rail, bipolar, unipolar	X	X	X	18-pin DIP 18-pin SOIC	Flow through, or single buffered data
ML2350C	8 (11-bits with gain ranging)	$\pm 1/2$	5	Single 5 or 12, dual ± 5	2.50 or 5.00	Rail-to-rail, bipolar, unipolar	X	X	X	18-pin DIP 18-pin SOIC	Flow through, or single buffered data
ML2351B	8 (11-bits with gain ranging)	$\pm 1/4$	5	Single 5 or 12, dual ± 5	2.50 or 5.00	Rail-to-rail, bipolar, unipolar	X	X	X	20-pin DIP 20-pin PCC	Double or single buffered data
ML2351C	8 (11-bits with gain ranging)	$\pm 1/2$	5	Single 5 or 12, dual ± 5	2.50 or 5.00	Rail-to-rail, bipolar, unipolar	X	X	X	20-pin DIP 20-pin PCC	Double or single buffered data

* Future Products

Note 1. Temperature Range:

C = 0°C to +70°C, I = -40°C to +85°C, M = -55°C to +125°C

12-Bit Plus Sign Data Acquisition Peripheral

GENERAL DESCRIPTION

The ML2200 and ML2208 Data Acquisition Peripherals (DAP) are monolithic CMOS data acquisition subsystems. These data acquisition peripherals feature an input multiplexer, a programmable gain instrumentation amplifier, a 2.5V bandgap reference, and a 12-bit plus sign A/D converter with built-in sample-and-hold. In addition to a general purpose 8-bit microprocessor interface, the ML2200 and ML2208 include a programmable processor, data buffering, a 16-bit timer, and limit alarms.

The ML2200B and ML2208B self-calibrating algorithmic A/D converters have a maximum non-linearity error over temperature of 0.018% of full-scale, while the ML2200C, ML2200D, ML2208C, and ML2208D have a maximum non-linearity error over temperature of 0.024%.

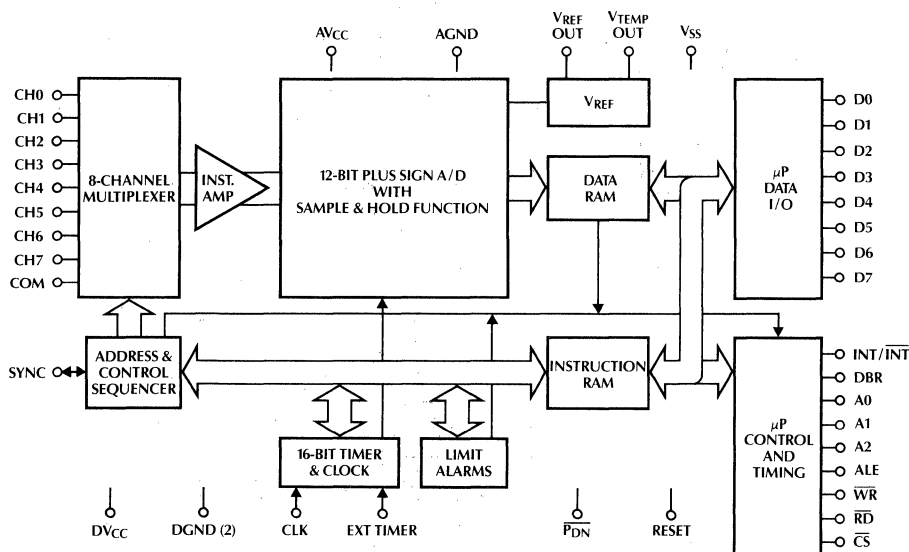
The ML2200 has a four channel differential input multiplexer and the ML2208 has an eight channel single ended input multiplexer.

The digital interface, with software-alterable configurations, is designed to off-load the microprocessor. Control of the DAP is autonomously handled through the control sequencer which receives its instructions from the instruction RAM.

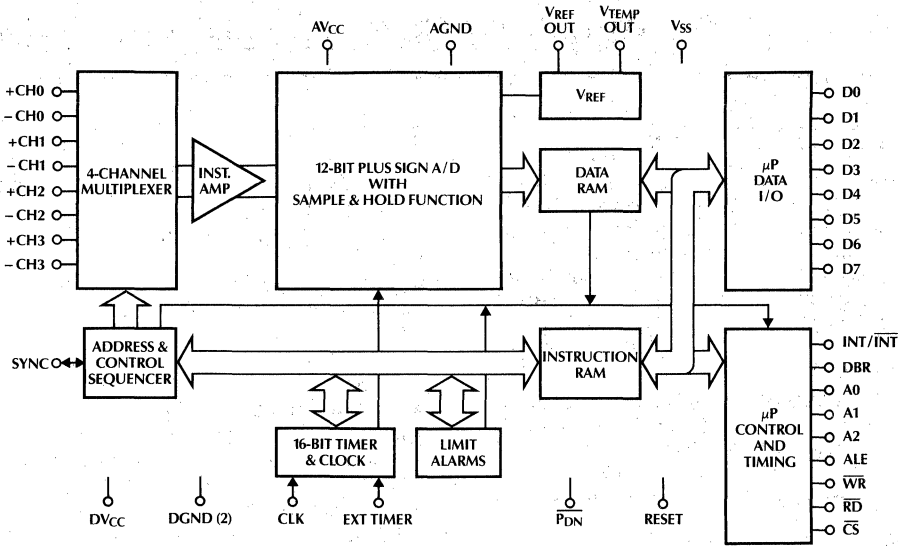
FEATURES

- Resolution 12 bits + sign
- Conversion time (including S/H acquisition) 31.5 μ s max
- Sample-and-hold acquisition 2.3 μ s max
- Non-linearity error $\pm 3/4$ LSB and ± 1 LSB max
- Low harmonic distortion 0.01%
- No missing codes
- Self-calibrating — maintains accuracy over time and temperature
- Inputs withstand $|7V|$ beyond supplies
- Internal voltage reference 2.5V \pm 2%
- Four differential or eight single-ended input channels
- Data buffering (8 word data RAM)
- Programmable limit alarm
- 8-Bit microprocessor interface — interrupt, DMA, or polling
- 16-Bit timer for programmable conversion rates
- Standard hermetic 40-pin DIP

ML2208 BLOCK DIAGRAM



ML2200 BLOCK DIAGRAM



BLOCK SCHEMATIC DIAGRAM

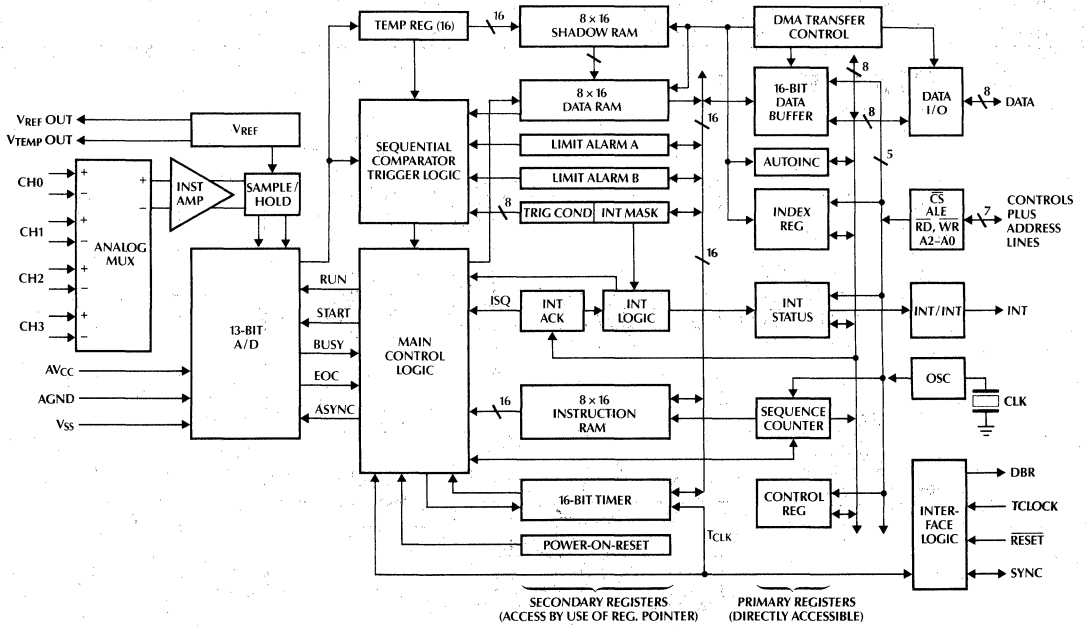


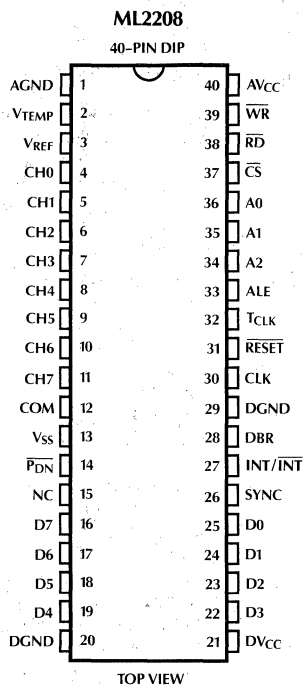
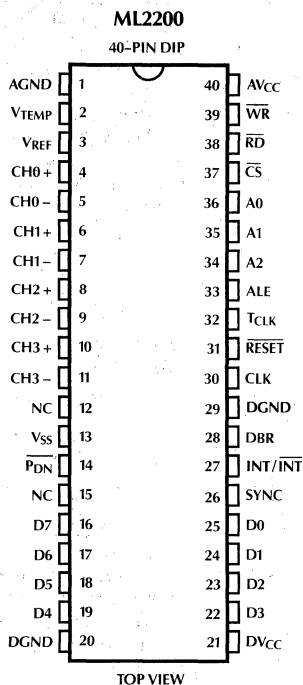
Figure 1. Block Schematic Diagram

PIN DESCRIPTIONS

PIN NO.	NAME	FUNCTION	PIN NO.	NAME	FUNCTION
1	AGND	Analog Ground.	28	DBR	Data Buffer Ready output active high indicates that a sequence of operations has completed and data is ready to transfer. DBR is not maskable. It can be used to generate an interrupt in addition to the INT pin when the DBR _{IE} bit in the interrupt mask register has not been enabled. DBR is the DMA request pin when DMA mode is enabled. DBR is not active unless in run mode and at least one sequence of operations has been completed. DBR remains active in the halt mode if not acknowledged; low during reset time and power-down.
2	V _{TEMP}	Voltage output proportional to the die temperature.	29	DGND	Digital Ground.
3	V _{REF}	Internal voltage reference output	30	CLK	Clock input. Drive with an external clock or crystal reference to DGND. The crystal must be parallel resonant with minimum capacitive loading (i.e., No bypass caps should be used and leads should be kept short).
4-11	CH	Analog Inputs. ML2200— Positive or negative input of four differential inputs ML2208— Eight single ended inputs referenced to common pin. Digitally selected by control sequencer.	31	RESET	Active low hardware reset with internal pull up resistor of 200K. Tie to system reset line or to grounded capacitor. The capacitor size (usually >6μF) is based on the time the power supplies stabilize, to the time reset voltage reaches 1.4V (>400ms).
12	NC COM	ML2200— No connection. ML2208— Negative common input for the eight input channels. Tie to analog ground or (V _{SS} +2.5) to (AV _{CC} -2.5V)	32	T _{CLK}	External timer, T _{CLK} is used as external clock input for the 16-bit timer when the T _{CLK} bit in the control register is set to one.
13	V _{SS}	Negative power supply; decouple to AGND.	33	ALE	Address latch enable, active low latches information on A0, A1, A2 and CS. Tie to AV _{CC} to disable use when separate address and data bus are used.
14	P _{DN}	Power-Down Input. When P _{DN} =0, device in power-down mode with register contents retained if AV _{CC} >2.0V.	34	A2	Address 2
15	NC	No Connection.	35	A1	Address 1
16-19	D7, D6, D5, D4	Bidirectional data bits.	36	A0	Address 0
20	DGND	Digital Ground.	37	CS	Chip select, active low
21	DV _{CC}	Digital power supply. Tie to AV _{CC} from same power supply.	38	RD	Read, active low enables ML2200 or ML2208 to drive data bus.
22-25	D3, D2, D1, D0	Bidirectional data bits.	39	WR	Write, active low allows writing into the registers.
26	SYNC	In the slave mode, SYNC is a positive edge triggered input used to start a conversion. In master mode, SYNC is an output and indicates a conversion has occurred.	40	A V _{CC}	Positive analog Power supply. Decouple to AGND. Tie to DV _{CC} from same power supply
27	INT	Interrupt output. A maskable interrupt programmable to be active high or low or will default to active high. INT will not clear until acknowledged in halt mode; not affected by the run or halt state. INT=0 during reset and inactive during P _{DN} .			

ML2200, ML2208

PIN CONNECTIONS



ABSOLUTE MAXIMUM RATINGS

(Note 1)

Supply Voltages (AV_{CC} and DV_{CC})	6.0V
Negative Supply Voltage (V_{SS})	-6.0V
Voltage at Analog Inputs	$V_{SS} - 7V$ to $AV_{CC} + 7V$
Voltage at V_{REF}	$V_{SS} - 7V$ to $AV_{CC} + 7V$
Input Current per Digital Pin	$\pm 10mA$
Input Current at Analog Inputs	$\pm 20mA$
Storage Temperature Range	-65°C to +150°C
Package Dissipation @ 25°C	1W
Lead Temperature (Soldering 10 sec.)	
Dual-In-Line Package (Ceramic)	300°C

OPERATING CONDITIONS

(Note 2)

Temperature Range	$T_{MIN} \leq T_A \leq T_{MAX}$
ML2200BCJ, ML2200CCJ, ML2200DCJ	0°C to 70°C
ML2208BCJ, ML2208CCJ, ML2208DCJ	0°C to 70°C
Supply Voltage (AV_{CC} and DV_{CC})	4.5V _{DC} to 6.0V _{DC}
Negative Supply Voltage (V_{SS})	-4.5V _{DC} to -6.0V _{DC}

ELECTRICAL CHARACTERISTICS

The following specifications apply for $AV_{CC} = DV_{CC} = +5V \pm 5\%$, $V_{SS} = -5V \pm 5\%$, $AGND = DGND = COM = CHX- = 0V$, $T_A = T_{MIN}$ to T_{MAX} unless otherwise specified. $C_L = 100pF$ for D0-D7, $C_L = 50pF$ for INT, DBR, and SYNC.

SYMBOL	PARAMETER	NOTES	CONDITIONS	MIN	TYP NOTE 3	MAX	UNITS
Converter Characteristics							
	Linearity Error ML2200BCJ, ML2208BCJ ML2200CCJ, ML2208CCJ ML2200DCJ, ML2208DCJ	4	$f_{CCLK} = 0.1 \leq 7MHz$ $f_{CCLK} = 0.1 \leq 7MHz$ $f_{CCLK} = 0.1 \leq 5MHz$			$\pm 3/4$ ± 1 ± 1	LSB LSB LSB
	Unadjusted Zero Error ML2200BCJ, ML2208BCJ ML2200CCJ, ML2208CCJ ML2200DCJ, ML2208DCJ	4				$\pm 3/4$ ± 2 ± 2	LSB LSB LSB
	Unadjusted Positive and Negative Full Scale Error	5				± 4	LSB
	Zero Error Temperature Coefficient				0.5		ppm/°C
	Gain Temperature Coefficient		External Reference		3		ppm/°C
	Common-Mode Rejection	13			80		dB
	Analog Input Range	5	All Analog Inputs	$V_{SS}-0.05$		$AV_{CC}+0.05$	V
	External Source Resistance for Analog Inputs	5 5	Channel = Analog Input Channel = Voltage Reference			2 0.5	kΩ kΩ
	Differential Analog Input Range	4	CHX referred to COM for ML2208 CHX+ referred to CHX- for ML2200	$-V_{REF}$		$+V_{REF}$	V
	Off Channel Leakage Current	5, 6	On Chan = 2.5V, Off Chan = -2.5V On Chan = -2.5V Off Chan = 2.5V	-100		+100	nA
	On Channel Leakage Current	5, 6	On Chan = -2.5V, Off Chan = 2.5V On Chan = 2.5V Off Chan = -2.5V	-100		+100	nA
	Gain Error		Gain = 2, 4, or 8		0.03		%
Voltage Reference and V_{TEMP} Characteristics							
	V_{REF} Absolute Value	4	Referred to AGND	2.45		2.55	V
	V_{REF} Output Pin Output Resistance	5		1		300	mΩ
	Minimum Load Resistance	5					kΩ
	Maximum Load Resistance	5				50	pF
	Temperature Coefficient				50		ppm/°C
	Line Regulation		$4.75 \leq AV_{CC} \leq 5.25$ $-4.75 \geq V_{SS} \geq -5.25$		1		mV
	Load Regulation		$1\mu A - 2.5mA$		1		mV
	Output Noise				100		μV_{RMS}
	V_{TEMP} Output Pin Absolute Value @ 25°C				$AV_{CC}-1.5$		V
	Volts per °C				5		mV/°C

ML2200, ML2208

ELECTRICAL CHARACTERISTICS (Continued)

The following specifications apply for $AV_{CC} = DV_{CC} = +5V \pm 5\%$, $V_{SS} = -5V \pm 5\%$, $AGND = DGND = COM = CHX = 0V$, $T_A = T_{MIN}$ to T_{MAX} unless otherwise specified. $C_L = 100pF$ for D0–D7, $C_L = 50pF$ for INT, DBR, and SYNC.

SYMBOL	PARAMETER	NOTES	CONDITIONS	MIN	TYP NOTE 3	MAX	UNITS
DC Characteristics							
	Power Supply Current AI_{CC} , Analog AV_{CC} DI_{CC} , Digital DV_{CC} I_{SS} , V_{SS}	4	$\overline{RD} = \overline{CS} = V_{IH}$		30	50	mA
		12			10		μA
I_{CC} I_{SS}	Standby Current $AV_{CC} + DV_{CC}$ Standby Current V_{CCPD} Minimum AV_{CC} and DV_{CC} for power-down data retention	4, 9	P_{DN} pin = GND		10	1000	μA
			P_{DN} pin = GND $V_{SS} = -5.25$ to GND	2	10	1000	μA V
	Power Supply Rejection AV_{CC}/DV_{CC} V_{SS}	7	DC DC to 25kHz, 200mV _{p-p} DC DC to 25kHz, 200mV _{p-p}		80		dB
					50		dB
					80		dB
					50		dB
V_{IL}	Input Low Voltage (except CLK, t_{CLK})	4				0.8	V
V_{IL1}	Input Low Voltage (CLK, t_{CLK})	4				0.8	V
V_{IH}	Input High Voltage (except CLK, t_{CLK})	4		2.0			V
V_{IH1}	Input High Voltage (CLK, t_{CLK})	4		3.5			V
V_{OL}	Output Low Voltage	4	$I_{OL} = 2.0mA$			0.45	V
V_{OH}	Output High Voltage	5	$I_{OH} = -1mA$	4.0			V
I_L	Input Leakage Current (except CLK and RESET)	4	$GND < V_{IN} < V_{CC}$			± 10	μA
I_{L1}	Input Leakage Current (CLK)	4	$GND < V_{IN} < V_{CC}$			± 200	μA
I_{L0}	Output Leakage Current (D0–D7)	4	$\overline{RD} = \overline{CS} = V_{IH}$			± 10	μA
I_{RST}	RESET Pin Source Current	4	RESET = 0V	15	50	100	μA
C_I	Input Capacitance (All Digital Inputs)				10		pF
C_O	Output Capacitance (All Outputs and D0–D7)				20		pF

AC Electrical Characteristics (Note 8)

t_C	Conversion Time	4, 9	CLK Mode = 0	$f_{CLK} = 7.0MHz$	31.5		μs
				$f_{CLK} = 5.0MHz$	44.0		μs
	Sample and Hold Acquisition	4, 9	CLK Mode = 0	$f_{CLK} = 7.0MHz$		2.3	μs
				$f_{CLK} = 5.0MHz$		3.2	μs
SNR	Signal-to-Noise Ratio		$V = 10kHz, 2.5V$ Sine. $f_{CLK} = 7MHz$ ($f_{SAMPLING} = 31.8kHz$). Noise is sum of all nonfundamental components up to $\frac{1}{2}$ of $f_{SAMPLING}$.		73		dB
THD	Total Harmonic Distortion		$V = 10kHz, 2.5V$ Sine. $f_{CLK} = 7MHz$ ($f_{SAMPLING} = 31.8kHz$). THD is sum of 2, 3, 4, 5 harmonics relative to fundamental.		-75		dB

ELECTRICAL CHARACTERISTICS (Continued)

The following specifications apply for $V_{CC} = DV_{CC} = +5V \pm 5\%$, $V_{SS} = -5V \pm 5\%$, $AGND = DGND = COM = CHX- = 0V$, $T_A = T_{MIN}$ to T_{MAX} unless otherwise specified. $C_L = 100pF$ for D0-D7, $C_L = 50pF$ for INT, DBR, and SYNC.

SYMBOL	PARAMETER	NOTES	CONDITIONS	MIN	TYP NOTE 3	MAX	UNITS
AC Electrical Characteristics (Note 8) (Continued)							
IMD	Intermodulation Distortion		$V_{IN} = f_A + f_B$, $f_A = 9kHz$, 1.25V sine. $f_B = 10kHz$, 1.25V sine. $f_{CLK} = 7MHz$ ($f_{SAMPLING} = 31.8kHz$). IMD is $(f_A + f_B)$, $(f_A - f_B)$, $(2f_A - f_B)$, $(2f_A - f_B)$, $(f_A + 2f_B)$, $(f_A - 2f_B)$ relative to fundamental.		-75		dB
FR	Frequency Response		$V_{IN} = 0$ to 10kHz, 2.5V sine relative to 1kHz		0.01		dB
f_{CLK}	CLK Frequency	4	(no crystal)	0.1		7	MHz
f_{CLKX}	CLK Frequency	4	(crystal)	3		7	MHz
f_{CLKI}	Internal CLK Frequency				1/2		f_{CLK} or f_{CLKX}
f_{CLKT}	CLK Frequency (t_{CLK} only)	4				f_{CLKI}	MHz
t_{CLKW}	Minimum Clock High/Low Width (CLK)	5		50			ns
t_{CLKWT}	Minimum Clock High/Low Width (t_{CLK})	5		75			ns
t_{RF}	Maximum Rise/Fall Times, All Inputs	5				25	ns
t_{RESET}	Minimum Reset Active Time	4, 10		10			f_{CLKI} Periods
t_{PDN}	Power-Up Time		Time After $P_{DN} = V_{IH}$		1		ms
Multiplexed Data Bus Timing							
t_{AL}	Address to ALE Setup Time	4		20			ns
t_{LA}	Address Hold Time After ALE	4		20			ns
t_{LC}	Latch to RD or WR Control	4		20			ns
t_{RD}	Valid Data Delay from Read	4				150	ns
t_{AD}	Address Stable to Valid Data	5		150			ns
t_{LL}	ALE Width	4		80			ns
t_{DF}	Data Bus Float After Read	4		10		50	ns
t_{CL}	Read or Write Control to ALE	4		20			ns
t_{CC}	Read or Write Control Width	4		150			ns
t_{DW}	Data Setup Time for Write	4		100			ns
t_{WD}	Data Hold Time for Write	4		0			ns
t_{RV}	Recovery Time Between Two Reads or Writes	4		250			ns
Non-Multiplexed Data Bus Timing							
t_{AD}	Address Stable to Valid Data	5		150			ns
t_{AR}	Address Stable Before Read	4		0			ns
t_{RA}	Address Hold Time for Read	4		0			ns
t_{RR}	Read Pulse Width	4		150			ns
t_{RD}	Data Delay from Read	4				150	ns
t_{DF}	Read to Data Float	4		10		50	ns

ML2200, ML2208

ELECTRICAL CHARACTERISTICS (Continued)

The following specifications apply for $V_{CC} = DV_{CC} = +5V \pm 5\%$, $V_{SS} = -5V \pm 5\%$, $AGND = DGND = COM = CHX- = 0V$, $T_A = T_{MIN}$ to T_{MAX} unless otherwise specified. $C_L = 100pF$ for D0-D7, $C_L = 50pF$ for INT, DBR, and SYNC.

SYMBOL	PARAMETER	NOTES	CONDITIONS	MIN	TYP NOTE 3	MAX	UNITS
Non-Multiplexed Data Bus Timing							
t_{RV}	Recovery Time Between Two Reads or Writes	4		250			ns
t_{AW}	Address Stable Before Write	4		0			ns
t_{WA}	Address Hold Time for Write	4		0			ns
t_{WW}	Write Pulse Width	4		150			ns
t_{DW}	Data Setup Time for Write	4		100			ns
t_{WD}	Data Hold Time for Write	4		0			ns

DMA Interrupt and SYNC Timings

t_{CKDBR}	Clock to DBR Assert	11, 4	DMA		120	180	ns
t_{RDD}	Read to DBR Negation on Last Byte	4			110	160	ns
t_{CKDBR}	Clock to DBR or t_{CKINT} , INT Assert	11, 4	Non-DMA		100	180	ns
t_{WRDBR}	Write to DBR or t_{WRINT} , INT Negation	11, 4			70	120	ns
t_{CKSYNC}	Clock to SYNC Delay	11, 4	Master Mode		150	200	ns
t_{SYNIN}	SYNC Input Width	5		3			f_{CLKI}
t_{SYNCK}	SYNC to Clock Setup	4	Slave: Mode 4 Only	50			ns
t_{SYNCO}	Minimum SYNC Output Width	4		4		4	f_{CLKI}

Note 1: Absolute maximum ratings are limits beyond which the life of the integrated circuit may be impaired. All voltages unless otherwise specified are measured with respect to ground.

Note 2: $0^{\circ}C$ to $+70^{\circ}C$ operating temperature range devices are 100% tested with temperature limits guaranteed by 100% testing, sampling, or by correlation with worst-case test conditions.

Note 3: Typicals are parametric norm at $25^{\circ}C$.

Note 4: Parameter guaranteed and 100% production tested.

Note 5: Parameter guaranteed. Parameters not 100% tested are not in outgoing quality level calculation.

Note 6: Leakage current is measured with the clock not switching.

Note 7: Power supply rejection is the ratio of the change in zero error to the change in power supply voltage.

Note 8: All parameters measured from 0.8V to 2.0V.

Note 9: Power-down current is with power-down pin at GND potential only. Any other level will dissipate more power. Other digital input pins may float but cannot be above V_{DD} or below GND.

Note 10: RESET should be held active for at least 10 internal clocks after power supplies have stabilized to within 5% of 5V.

Note 11: Since the internal master clock is the input clock divided by 2, this number can be either the maximum listed or the maximum listed plus $\frac{1}{2}$ the input clock period.

Note 12: When $RD = CS = V_{IL}$, the current into the DV_{CC} pin depends on the load on the data bus pins D0-D7.

Note 13: Common-Mode rejection is the ratio of the change in zero error to the change in common-mode input voltage.

TIMING DIAGRAMS

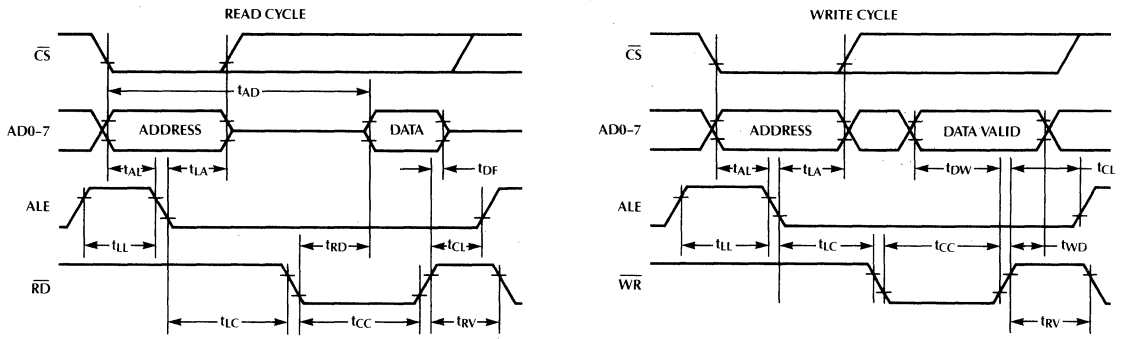


Figure 2. Multiplexed Bus

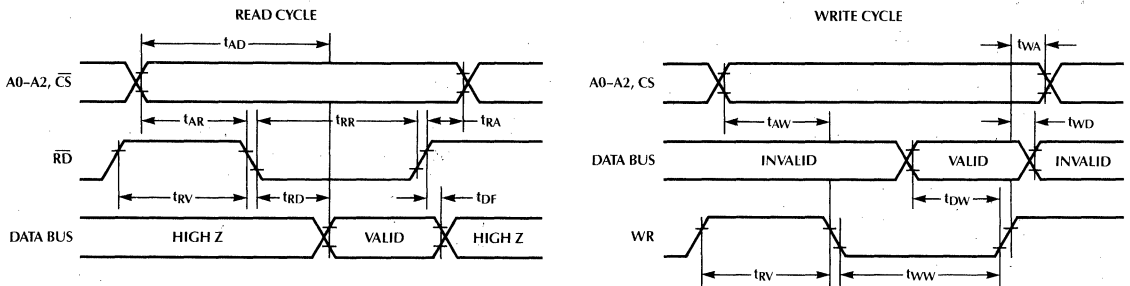


Figure 3. Non-Multiplexed Bus

TIMING DIAGRAMS (Continued)

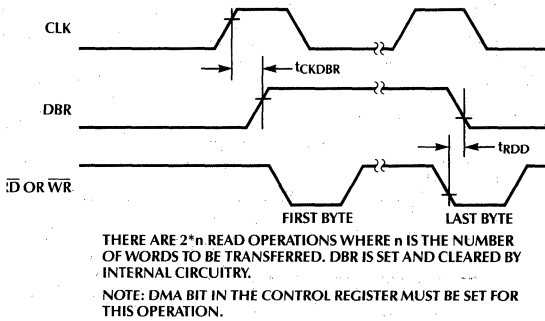


Figure 4. DMA Mode

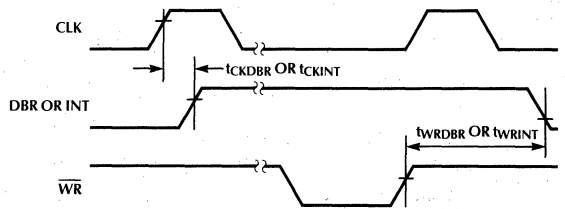


Figure 5. DBR and INT (Non-DMA Mode)

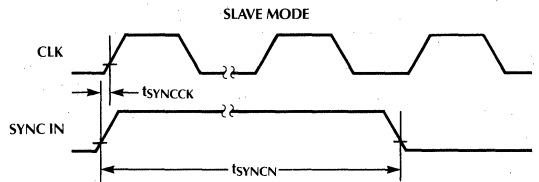
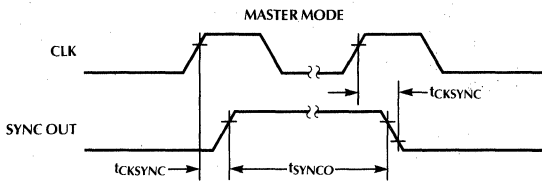


Figure 6. SYNC

1.0 FUNCTIONAL DESCRIPTION

1.1 Algorithmic A/D Converter

Micro Linear's algorithmic converter uses a successive approximation technique. Most of today's successive approximation converters use a DAC to feed back the approximated signal, however this technique requires more circuitry than algorithmic converters. In addition the values of all of the resistors or capacitors in the DAC must be matched to within the accuracy of the converter. This is difficult to do in silicon beyond 10 bits unless trimming is used. An algorithmic converter uses less circuitry and is more easily trimmed. Micro Linear's algorithmic converter is implemented using a $2\times$ amplifier, a sample/hold amp, and a comparator as shown in Figure 7.

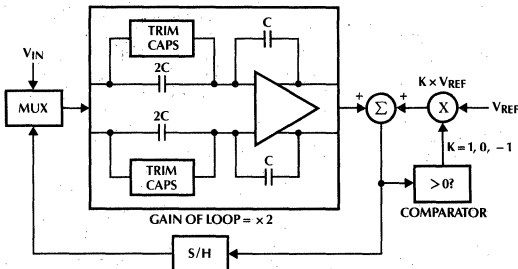


Figure 7. Self-Calibrating A/D Converter

The input sample is first multiplied by two then compared to the reference voltage. If the $2\times$ input voltage is greater than the reference, the MSB is a 1 and the reference voltage is subtracted from the $2\times$ input voltage. The remainder is stored in the sample-and-hold. If the $2\times$ input voltage is less than the reference, the MSB is a 0 and the $2\times$ input voltage is stored in the sample-and-hold. This process repeats again, however now the sample-and-hold voltage is multiplied by 2.

Self-Calibration

In order to maintain integral and differential linearity to the $1/2$ LSB level in an algorithmic converter, two critical parameters need to be controlled, loop offsets and the gain of the loop. Loop offsets are automatically nulled before each conversion using auto-zeroing circuitry on both the sampling amp and the $2\times$ amp. The gain of the loop is adjusted using self-calibration.

Self-calibrating the algorithmic converter, once the offsets have been nulled, is performed by measuring the $2\times$ gain of the loop and adjusting it. The gain can be measured by converting the reference voltage as the input as well as the reference (V_{REF}/V_{REF}), and examining the output code.

Converting V_{REF} should yield plus full-scale, since V_{REF}/V_{REF} should equal 1. If the gain of the loop is slightly less than 2, the resulting LSB of the conversion will be "0". If the magnitude bits of the resulting conversion are all "1s", the gain may

be too great, therefore the gain is reduced to the point where the threshold of the LSB is reached.

Adjustment of the $2\times$ gain is done with the binary weighted trim capacitor arrays connected to each of the 2C input capacitors. A small value of capacitance is either added to or subtracted from the 2C input caps until the gain of the loop is within 13-bit accuracy of 2.

1.2 Multiplexer Input

The input voltage is $\pm 2.5V$ relative to COM of the ML2208 or a CH- of the ML2200. The input voltages under normal operation must not exceed supply voltages by 0.05V. Each channel is selected by the programmable sequencer.

1.3 Internal Voltage Reference and V_{TEMP}

The internal bandgap voltage reference with a temperature coefficient of 50ppm/ $^{\circ}C$ has an external use current of 2.5mA.

The voltage reference V_{TEMP} output is directly proportional to the chip temperature.

1.4 Conversion Times

The following table lists the conversion times which include the sample-and-hold acquisition time. For a CALRD and CALWR no A/D conversion actually takes place.

Operation	Number of Internal Clocks*
8-bit A/D	80
13-bit A/D	110
CALWR	52
CALRD	80

*Internal clock is the external clock divided by two.

1.5 Sample-and-Hold Timing

Figure 8 shows the internal timing for the sample-and-hold circuitry. The relationship between the "Start of Conversion" and the input channel going into sample mode is fixed at 6

internal clocks, regardless of the Start Mode. Six internal clocks after the Start of Conversion, the Sample-and-Hold is switched into the sample mode, placing two 9pF capacitors in parallel with the input pins; one on CH+ and one on CH- for the ML2200, and CH and COM for the ML2208. The sample switch is kept in the sample mode for 8 internal clocks ($2.3\mu s$ at a 7MHz external clock), then placed in the hold mode. During the next 2 internal clocks the charge on the sample-and-hold is transferred into the A/D, after which the V_{REF} pin is sampled for 8 internal clocks.

Figure 8 also illustrates the timing of the SYNC pin in Master Mode during a conversion. SYNC is activated one internal clock cycle before the Start of Conversion and lasts for four internal clocks.

1.6 Analog Inputs

Differential Inputs and Common-Mode Rejection

The differential inputs of the ML2200 eliminate the effects of common-mode input noise (60Hz, for example), as CH+ and CH- are sampled at the same time.

Noise

The leads to the analog inputs should be kept as short as possible to minimize output noise. Noise as well as digital clocks can couple into the inputs and cause errors. Input filters can be used to reduce the effects of these sources.

Power Supply Decoupling

Low inductance tantalum capacitors of $1\mu F$ or greater and $0.01\mu F$ disc ceramic capacitors are recommended for bypassing AV_{CC} as well as V_{SS} to AGND. These capacitors should be placed close to the AV_{CC} and V_{SS} pins.

2.0 μP HARDWARE INTERFACE

The microprocessor interface is a byte-oriented structure which occupies eight memory or I/O locations in the microprocessor's address space. Each register is readable and writable via the chip select, read and write pins, three address lines, and 8-bit data bus.

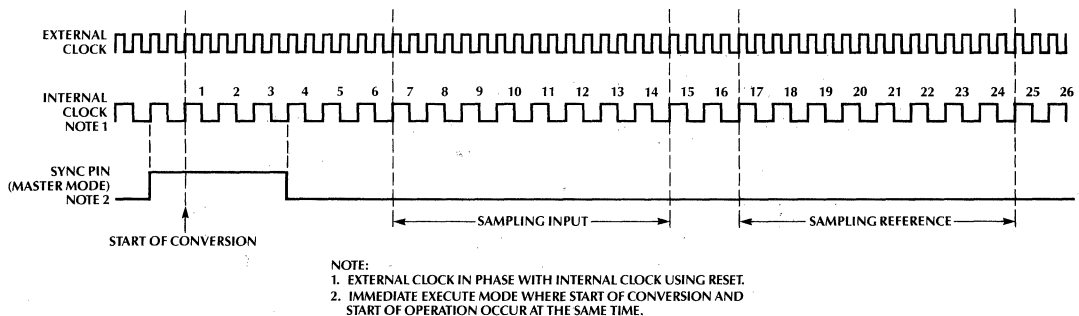


Figure 8. Sample-and-Hold Timing

ML2200, ML2208

Interfaces are shown for multiplexed address data bus in Figure 9 and Figure 10. When non-multiplexed interfaces are

used, ALE can be tied high. All internal address and chip select latches are transparent.

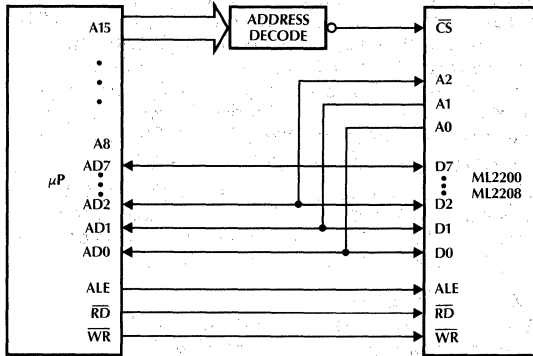


Figure 9. 8-Bit Multiplexed Bus Interface

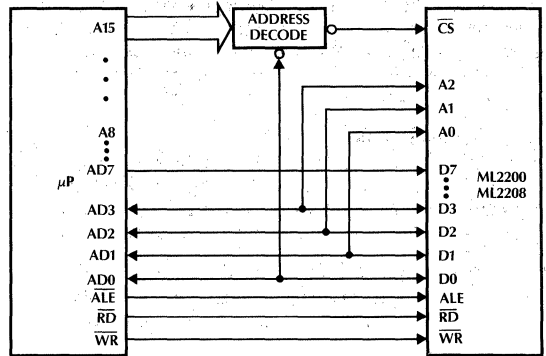


Figure 10. 16-Bit Multiplexed Bus Interface

2.1 Interrupts

The ML2200 and ML2208 provide two interrupt pins, one for control/status interrupts (INT), and one for data interrupts (DBR). The standard INT pin is maskable via an interrupt mask register while the DBR pin is always enabled to signify that data is available. DBR can be mapped into the INT pin if only one interrupt pin is desired.

The interrupt pin (INT) can be programmed, via the Interrupt Bit Mask register, to be active high, or active low. When programmed for active high, it is driven in both directions. When INT is programmed for active low, it is an open drain output, therefore an external pull-up resistor of 2.5kΩ or more should be used. The DAP's Status register can be read to determine whether its interrupt is active or not.

2.2 DMA

The separate DBR pin can also serve as a DMA request signal when DMA operation is enabled in the Control register. DBR goes active high when the data buffer is full and ready to be read. DBR remains high until the last byte in the data buffer has been read. This allows back-to-back DMA cycles or single cycle transfers depending on how the DMA controller is programmed. The data for the DMA cycle is transferred over the 8-bit data bus at address 0 (A0-A2=0). The ML2200 or ML2208 automatically places both high and low bytes of the 16-bit wide data buffer at address 0 or 1 for the DMA controller to read. The LOBYT bit in the Control register specifies whether the high or low byte is placed on the bus first. Figure 11 shows a block diagram interfacing to the 8237 DMA controller.

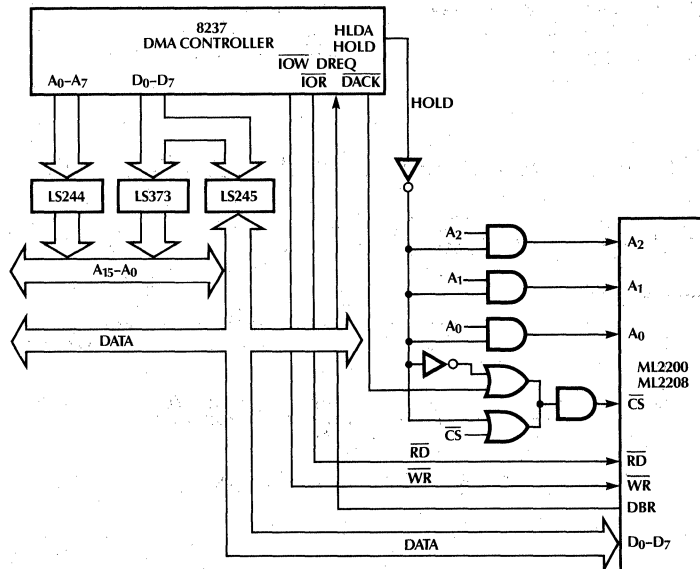


Figure 11. DMA Interface

3.0 REGISTER DEFINITIONS

These data acquisition peripherals contain six directly addressable 8-bit registers, and twenty indirectly addressable 16-bit registers. Figure 12 illustrates the register architecture while Figures 13, 14 & 15 illustrate the bit maps and addresses. The

first three primary registers (Window Low, Window High, and Index) are used to access the 20 secondary registers. Window Low and Window High provide read/write access to the low and high bytes of the secondary register pointed to by Index.

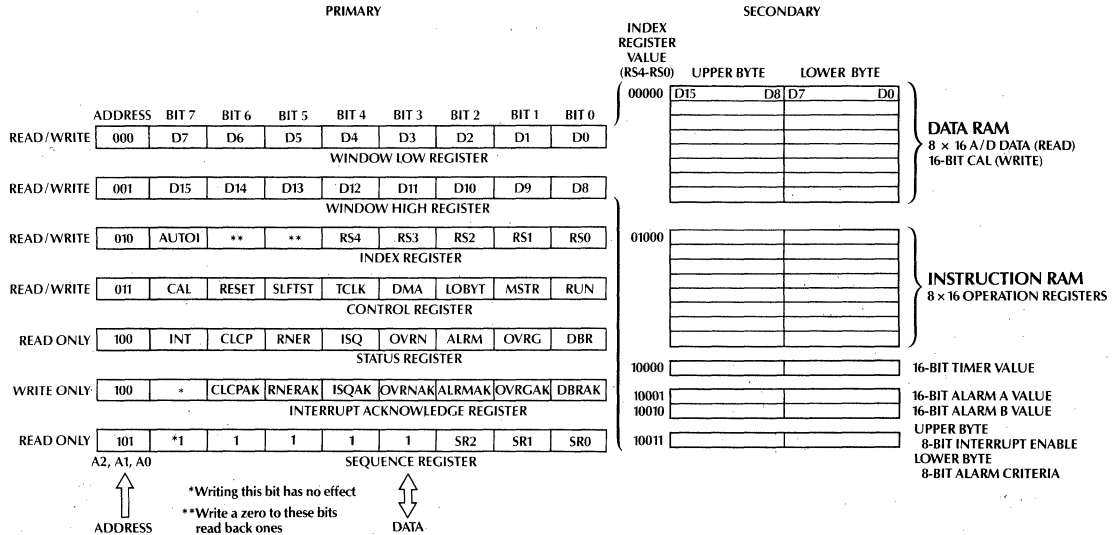


Figure 12. Register Architecture

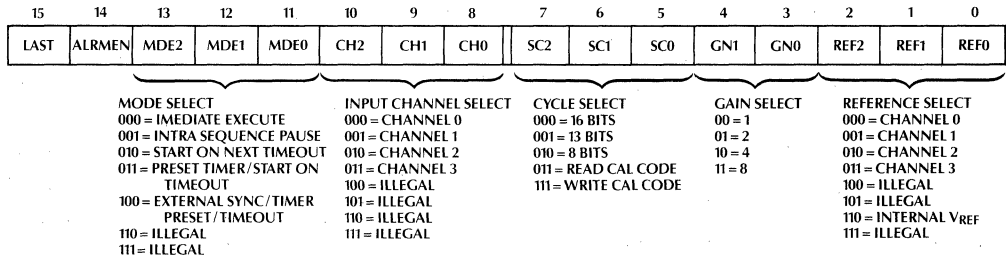


Figure 13. ML2200 Bit Map of Instruction RAM

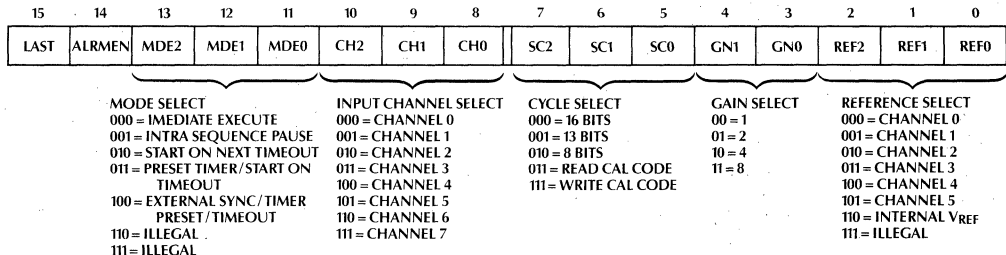


Figure 14. ML2208 Bit Map of Instruction RAM

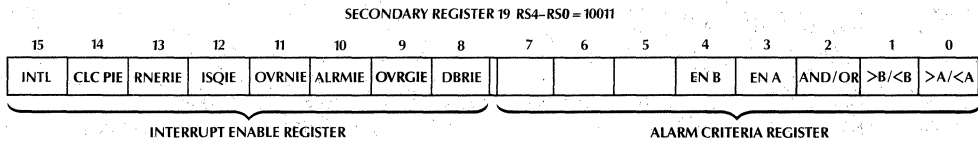


Figure 15. Interrupt Enable and Alarm Criteria Registers

3.1 Primary Registers

								READ/ WRITE
D7	D6	D5	D4	D3	D2	D1	D0	

Window Low Register

								READ/ WRITE
D15	D14	D13	D12	D11	D10	D9	D8	

Window High Register

Window Registers — Registers 0 and 1

These registers form a two-byte window into the secondary registers. Window Low is the low byte of the secondary 16-bit word, and Window High is the high byte. Any one of the 20 words in the secondary register set can be accessed by first setting a 5-bit address in the Index register, then reading from or writing to the Window registers. Sequential access of the secondary registers is also available without writing to the Index register via the AUTOI bit in the Index register.

Index Register — Register 2

							READ/ WRITE
AUTOI			RS4	RS3	RS2	RS1	RS0

Index Register

RSX = Secondary Register Address (Bits 0 to 4): The lower five bits of this register (RS0–RS4) define the location within the secondary register set that the window registers are positioned at.

Bits 5 and 6: Undefined. Writing to these bits have no effect, however a zero should be written; always read as ones.

AUTOI = autoincrement (Bit 7): Setting AUTOI signifies that the lower five addressing bits in the Index register are automatically incremented after either the Window Low or Window High register is accessed. Whether the auto-increment occurs when accessing Window Low or Window High register, is based on the LOBYT bit in the Control register.

Interrupt Operation Caution!!! - Using the auto-increment feature with interrupt driven software deserves special attention. A problem can arise when an interrupt service routine accessing the secondary registers, interrupts another routine accessing secondary registers. This problem can be avoided one of two ways: disable the interrupt in the main routine while accessing secondary registers, or reload the index register to its entry value when exiting the interrupt routine.

Note: The Index register is automatically cleared only under two conditions, one is a RESET, the other is when DMA mode is used. This register is reset to 0 in DMA mode just prior to the DMA request (DBR going active). DMA mode uses the index register for operation, so the index register should never be written to when RUN and DMA are set.

Control Register — Register 3

							READ/ WRITE
CAL	RESET	SLFST	t _{CLK}	DMA	LOBYT	MSTR	RUN

Control Register

RUN (Bit 0): Setting this bit to a one will cause the chip to start executing the operations defined in the Instruction RAM, beginning with location 0. This is referred to as the Run mode. Clearing this bit will place the ML2200 in the Halt mode. The run bit is initially cleared on power up or after a hardware or software reset. In order to properly start the chip operation, the RUN bit should be set after setting all other applicable bits in the control register. The act of halting the chip will always reset the sequence pointer to operation 0. Thus, the next time RUN is asserted, the chip starts from operation 0 again. Placing the chip in the Run or Halt mode has no effect on the Interrupt pins (INT and DBR), nor the status bits in the status register. It is recommended that secondary registers only be written to in the Halt mode. Writing to secondary registers in the Run mode will cause the RNER status bit to be set, indicating a run error. All of the status bits in the Status register should be acknowledged (cleared) before entering the Run mode.

MSTR = master (Bit 1): Indicates whether the SYNC pin will be an input or an output. If set the chip will enter the master mode of operation and the SYNC pin will become an output pin which puts out a sync pulse at the beginning of each operation. This serves as a signal for other slave chips that are used in a synchronous operating method. While in master mode, any operation requiring a sync input will not proceed, and the chip will “hang”, waiting for a sync that will never come. The chip default is slave mode with the SYNC pin as an input.

LOBYT = low byte first (Bit 2): This bit is used to indicate which byte is accessed first in AUTOI or DMA operation. When this bit is set, the index register is incremented on the read or write of the Window High register. When this bit is clear, the index register is incremented on the read or write of the Window Low register. If DMA operation is specified, then setting this bit will make the low byte be output first, then the high byte, after which the index register is incremented. Conversely, clearing this bit will output the high byte first, then the low byte, then increment the index register. The default is low.

DMA = DMA Mode (Bit 3): When set enables DMA operation. DMA operation proceeds as follows:

- 1) The DMA bit must be set after defining all other registers (Instruction RAM, Alarm etc.) but prior to setting the RUN bit. The RUN bit is then set.

- 2) The sequence of operations in the Instruction RAM is executed.
- 3) At the end of the sequence, the DBR pin goes true, requesting DMA service, and the Index register is automatically cleared, pointing to the first location of the data buffer.
- 4) Each read of either Window Low or Window High register outputs a byte from the data registers. The DMA controller can read Window Low register, or Window High register, or alternate between Window Low and Window High. The same data is placed in both Window Low and Window High registers, and updated in both of them when either one is read. The data is placed in the Window registers beginning with data word 0 and incrementing on up. The placement of the low byte/high byte order is based on the LOBYT bit in the Control register. The number of bytes transmitted equals twice the number of operations defined, since the words are 16 bits going over an 8-bit bus. DBR remains asserted until all of the bytes have been transmitted. It is negated on the leading edge of the last byte read pulse. DBR acknowledge (setting the DBRAK bit in the Status register) is not required when transferring bytes via DMA.

The AUTOI bit does not have to be set when in the DMA mode. Setting the DMA bit forces the Index register to be auto-incremented in the Run mode. However if AUTOI is not set, then when in Halt mode auto-increment will not be enabled. If the AUTOI bit and DMA bit are both set, the auto-increment will occur in both the Run mode and the Halt mode.

t_{CLK} = enable external timer clock (Bit 4): When set, will divert the clock input for the internal sixteen bit timer to the t_{CLK} pin. When reset to 0, the timer runs at the internal chip clock frequency, which is 1/2 of that generated at the CLK pin.

SLFTST = self test (Bit 5): When set, the function of the input multiplexer is modified to enable self test operations. This bit

also redefines the Instruction Word, specifically the CHAN field of the instruction word (See Figure 16 for the redefinition of the Instruction Word when SLFTST = 1). With SLFTST set the CHAN bits now specify which of four self tests is to be performed as shown below.

Instruction Word CHAN Field	Function	Description
000	System Offset	Inputs shorted together and shorted to ground
001	Internal Reference	Convert internal V _{REF} plus side tied to V _{REF} minus side tied to AGND
010	Minus Internal Reference	Convert internal V _{REF} minus side tied to V _{REF} plus side tied to AGND
011	Common Mode	Both inputs of the converter are tied to V _{REF}
100-111	Illegal	

2

These self-test results are useful for user confidence at power-on. The default on reset is 0, normal mode of operation.

RESET = soft reset (Bit 6): Is a software reset of the chip. This bit does not have to be cleared once set. The microprocessor should read this bit back to determine if the reset operation has completed, especially if a slow clock rate is being used. It takes at least 4 internal clocks for the reset bit to clear. Microprocessor communication with the chip should be held off until this bit is read back as cleared. When issuing a hardware reset, communication with the chip should be held off until the RESET pin goes inactive. The chip will be in the Halt mode (RUN bit cleared) after a reset. See RESET/Power-On Conditions (Section 4.2) for chip register conditions after a reset.

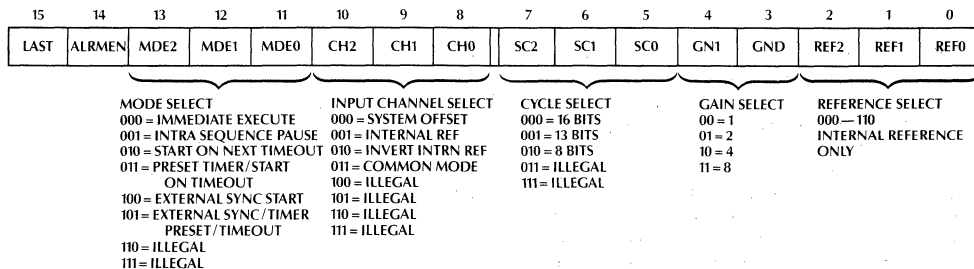


Figure 16. Bit Map of Instruction Word When SLFTST = 1

CAL = calibration start (Bit 7): When set, a self-calibration of the A/D converter will begin. Reading the CAL bit indicates whether the chip has been calibrated since the last reset or power-on condition. If CAL is a 1, then a calibration of the A/D converter has been performed since the last reset or power-up. When setting CAL, the user should not write a 0 back to clear it. **Writing a 0 to the CAL bit has no effect;** this will not clear it if it was previously set. Attempting to set the RUN bit without this bit being set will result in a run error condition, in which the RNER status bit will be set, and an interrupt being generated if it was enabled in the mask register. The amount of time required for calibration is 8,260 internal clocks, or 16,520 external clocks. To determine when a calibration is complete, the microprocessor should enable the calibration complete interrupt (CLCPIE) in the interrupt mask register, and wait for the interrupt to occur. Interrupt servicing of the calibration complete interrupt is done in a normal manner, in which the interrupt is acknowledged by setting the CLCPAK bit in the interrupt acknowledge register. All I/O to the ML2200 should be avoided during calibration (i.e., 16,520 external clocks after the CAL bit is set), because accessing the chip during calibration could adversely affect the calibration. If an interrupt is not desired, the microprocessor can read the Status register to verify completion 16,520 external clocks after the CAL bit is set. When the CAL bit is set, all other bits in the Control register should be cleared. **DO NOT** set the CAL and RUN bits simultaneously.

Status Register — Register 4

								READ ONLY
INT	CLCP	RNER	ISQ	OVRN	ALRM	OVRG	DBR	

Status Register

Register 4 serves as the status register of the various conditions that can occur. The bits in the Status register will be updated regardless of the Mask register. The status bits are updated any time within or at the end of a sequence of operations. The bits in the Status register are cleared by setting the corresponding bits in the Interrupt Acknowledge register. The status register can be polled at any time without fear of clearing the status bits. This register is not cleared at HALT time. When entering the Run mode, all of the old status bits should be cleared (acknowledged).

DBR = Data Buffer Ready (Bit 0): Is set when the chip has gone through one complete sequence of operations and has filled the data registers with the converted results. This bit signifies that the microprocessor should read all locations in the data registers that have relevant data. Reading all loaded data locations will clear DBR. If all loaded data locations are not read, DBRAK in the Interrupt Acknowledge register should be set to clear DBR, else OVRN will be set. The DBR pin is logically the same as the DBR status bit. The DBR pin is ALWAYS enabled and cannot be masked out. The DBR status bit is the only condition that can cause the DBR pin to be asserted. The DBR status bit can be enabled to assert the INT pin through the Interrupt Mask register.

OVRG = overrange interrupt (Bit1): Is set at the end of an operation when an underflow or overflow of the A/D converter has occurred (underflow and overflow are the most negative and most positive number, respectively, that is representable in the chip according to the specified cycle length). The overflow and underflow conditions apply to ALL incoming A/D converted data.

ALRM = limit alarm (Bit 2): Is the limit alarm status bit. It is set whenever the alarm criteria specified in the alarm registers is satisfied by a conversion from an operation where the ALRMEN bit is enabled. The limit alarm test only applies to an operation in which the ALRMEN bit is set.

Note that OVRG and ALRM can be enabled without enabling the DBR interrupt so that the microprocessor can be left alone until an overflow/underflow or limit alarm occurs. This is done to search for a limit condition first without taking any data into the microprocessor. Doing this, however, will set the OVRN (overrun error) bit in the status register, indicating that the microprocessor has not read any data from previous sequences.

OVRN = overrun error (Bit 3): The OVRN bit indicates that the microprocessor has missed from one byte to several blocks of data. Even if an overrun error occurs, the ML2200 or ML2208 continues converting the inputs and updating the data registers with the new conversions.

This bit may intentionally become set as a result of searching for the overflow/underflow or limit alarm criteria without reading the data.

The setting of the OVRN bit also occurs in DMA mode if all data has not been read by the completion of the next sequence. (Note: DBRAK should not be set in DMA mode, since DBR is automatically cleared by the chip.) If OVRN occurs in DMA mode, DBR will not be reactivated once all of the data from the sequence which was overrun is read; OVRN automatically disables DBR re-activation. Acknowledging OVRN (setting OVRNAK in the Interrupt Acknowledge register) will re-enable the DBR pin, however the OVRN bit may immediately be set again before the DMA controller can read the entire buffer. Therefore, it is recommended that in DMA mode if OVRN gets set, put the ML2200 or ML2208 in the Halt state, acknowledge the overrun and the DBR, then place the chip back in the Run mode.

ISQ = intra-sequence pause (Bit 4): Indicates that the chip has halted operation within a sequence as a result of choosing the ISQ op code in the mode field of the Instruction word. Setting the ISQAK bit in the interrupt acknowledge register will then re-start the operation within the sequence. This lets the microprocessor achieve timing control of individual operations within a sequence.

RNER = run error (Bit 5): Indicates that an error occurred either entering or operating in the Run state. The following operational errors cause the RNER bit to get set

1. Entering the Run state without having performed a self-calibration after the most recent Reset or power-up. The status of whether a calibration was executed or not is indicated by the CAL bit in the control register. If the CAL bit in the Control register is a one, the chip has already been calibrated.
2. Writing to any secondary registers other than the data registers during Run mode. All secondary register locations are readable during Run time.

CLCP = calibration complete (Bit 6): Is set at the end of a calibration sequence. The purpose of this bit is to notify the microprocessor that a self-calibration has completed.

INT = interrupt (Bit 7): Is identical to the state of the INT pin. The INT status bit and pin is an OR of the status bits enabled in the Interrupt Mask register. While the polarity of the INT pin can be defined in the interrupt mask register, this bit is positive true only.

Interrupt Acknowledge Register — Register 4

								WRITE ONLY
	CLCPAK	RNERAK	ISQAK	OVRNAK	ALRWAK	OVRGAK	DBRAK	

Interrupt Acknowledge Register

The status bits in the status register can only be cleared by setting the appropriate bit in this register; writing a zero has no effect. The relative bit positions in the Interrupt Acknowledge register are identical to the Status register except for bit 7, which is valid for reads (see explanation in Status Register) and undefined for writes (user must write a zero to this bit to be software-compatible for possible future redefinitions).

Sequence Register — Register 5

								READ ONLY
1	1	1	1	1	SR2	SR1	SR0	

Sequence Register

During the RUN mode, this register can be read back to indicate the current operation in progress. This is especially useful for examining interrupts when multiple intra-sequence pauses are specified. Bits 3-7 always reads 1s.

Registers 6 and 7—these registers are reserved for future use.

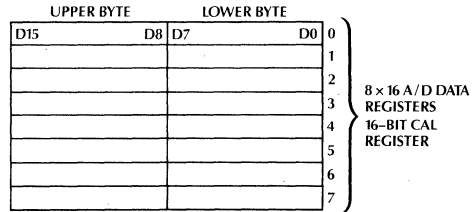
3.2 Secondary Registers

There are twenty 16-bit wide secondary registers containing the Data RAM, Instruction RAM, Timer, Alarms, Alarm Criteria Register, and Interrupt Mask. Except for the Data RAM, the secondary registers should only be accessed on initialization, or when the chip is placed in the Halt mode.

Secondary Registers 0 to 7

Data RAM (read only)

Calibration Holding Register (write only)



The Data RAM consists of eight 16-bit wide registers that hold the output results from the latest conversion sequence. Each word in the Data RAM has a one-for-one correspondence with a word in the Instruction RAM. The Data RAM is also referred to as the data output registers.

The data output registers are double buffered and readable by the microprocessor at any time. The A/D converter fills a "shadow" bank of registers during conversions, while the microprocessor is free to read the output registers. When the sequence is done, the "shadow" bank information is transferred to the output registers for the microprocessor to read, after which time DBR is asserted. Therefore, the microprocessor has essentially one sequence time to drain the data buffer. This time varies according to the number of operations defined, the system clock frequency, the mode field for each operation, and the cycle length (number of bits to be converted). Refer to Conversion Times for more information.

Data Format

All data is returned from the converter in 16-bit two's complement format, right hand justified, with the sign bit extended across the most significant unused bits.

Cycle	+Max	-Max	Mid-Range
16	7FFF	8000	0000
13	0FFF	F000	0000
8	007F	FF80	0000

Calibration Holding Register —

This register is for diagnostic purposes only. It is one 16-bit wide register mapped into the write only secondary address space 0 to 7 (i.e., a write to any of the secondary addresses 0-7 will load the Calibration Holding register). This register is write only and cannot be read back directly. It is used when the mode field in the Instruction Word is set to CAL Write, and the Instruction is executed. This command takes the contents of the Calibration Holding register and loads it into the Calibration register of the A/D converter. Note that this will change the calibration of the A/D converter, and a calibration of the A/D converter should be done after a CAL Write command is issued.



Instruction RAM—Secondary Registers 8 to 15 (Read/Write)

OP	0	8
OP	1	9
OP	2	10
OP	3	11
OP	4	12
OP	5	13
OP	6	14
OP	7	15

} 8 x 16 OPERATION REGISTERS

The Instruction RAM, sometimes referred to as the Operation registers, consists of eight 16-bit wide registers broken up into seven different fields (see Figures 10 and 10A). Each Instruction or Operation defines a single conversion, where the converted data result is stored in the corresponding data output register. Note that when the SLFTST bit in the Control register is set, the Instruction Word is redefined for diagnostic mode. Figure 12 illustrates the redefinition when SLFTST is set. The following section defines the seven different fields making up the Instruction word when SLFTST = 0.

D15	D14	D13-11	D10-8	D7-5	D4,3	D2-0
LAST	ALRMEN	MODE	CHAN	CYCLE	GAIN	REF

REF (Bits 2, 1, 0—Voltage Reference Selection) REF specifies the source of the voltage reference used for the A/D conversion.

GAIN (Bits 4 and 3—Gain Settings) GAIN defines the gain of the precision instrumentation amplifier. The gain can be either 1, 2, 4, or 8. Each gain factor of 2 adds an additional 4 internal clock cycles ($1/f_{CLKI}$) to the conversion time. Therefore a gain of 8 adds an additional 12 internal clock cycles to the conversion time.

CYCLE (Bits 7, 6, 5—Cycle Select) CYCLE defines one of five different cycles: 8-, 13-, or 16-bit conversions, and READ or WRITE CAL CODE. Choosing 8-, 13- or 16-bit cycles determines how many bits the A/D converter will convert. However, even though the converter has a 16-bit cycle, it may not have 16 bits of useful resolution. The useful resolution of the converter can be determined from the linearity specs.

Since the algorithmic converter is a successive approximation type of converter, an 8-bit cycle requires the least amount of time to convert, and the 16-bit cycle requires the most. Refer

to Sampling Rates and Conversion Times for the exact number of clocks each cycle takes.

READ CAL CODE and WRITE CAL CODE cycles are for diagnostic purposes only. READ CAL CODE reads the Calibration register in the A/D converter and loads it into the corresponding data output register. WRITE CAL CODE transfers the contents of the Calibration Holding register into the A/D converter's Calibration register. The transfer is complete after the operation is executed. Refer to Diagnostics for more information on READ and WRITE CAL CODE.

CHAN (Bits 10, 9, 8—Input Channel Number) defines the input channel to be converted.

ALRMEN (Bit 14—Alarm Enable) When this bit is set the alarm criteria for the operation is enabled, otherwise the alarm is disabled for this operation. If ALRMEN is set and the alarm condition is met, the ALRM bit in the Status register will be set at the end of the operation.

LAST (Bit 15—Last Operation) signifies that this operation is the last operation of the sequence. The chip will return to and begin the first operation of the sequence after execution of the current operation. If all eight operations are specified, the last one MUST have this bit set.

MODE (Bits 13, 12, 11—Mode Selection) defines the condition that must be met for the operation to proceed. The mode field also has an effect on the Operation Execution Time.

000	Immediate Execute
001	Intra-Sequence Pause
010	Start on Next Time out
011	Preset Timer/Start on Time out
100	External Sync Start
101	External Sync/Timer Preset/Time out
110	ILLEGAL
111	ILLEGAL

Events That Occur Within an Operation

To better understand six modes of the ML2200 or ML2208 one must first understand the events that occur during an operation. This can be aided by referring to Figure 17.

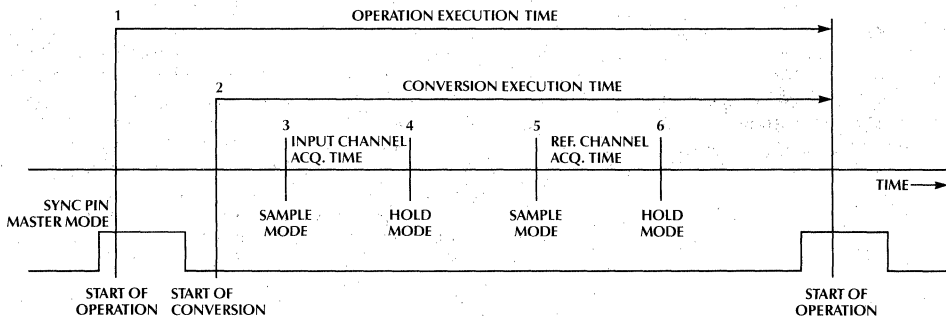


Figure 17. Events Within an Operation

The first event that occurs in the Operation is the Start of Operation. This may or may not be the beginning of the conversion, depending on the mode selected. The time between the Start of Operation and Start of Conversion is variable. When the conditions of the mode have been met, the Start of Conversion occurs.

The Conversion Execution time includes the input and reference acquisition times, the gain time, and the successive approximation conversion time. Shortly after the Start of Conversion the S/H goes into sample mode acquiring the input channel for eight internal clocks. After the input has been acquired the S/H goes into hold mode, disconnecting the S/H from the input channel, and transferring the charge into the A/D converter. A couple of clocks later the same S/H goes into sample mode on the reference voltage, either the internal V_{REF} or one of the input channels. The reference acquisition time for all six modes is the same; eight internal clocks. After the S/H goes into hold mode the successive approximation A/D conversion begins. When the conversion is complete the next operation begins.

Immediate Execute (000) — The Start of Conversion begins at the Start of Operation. In other words, the conversion begins the instant the operation begins. There is no gating item delaying the conversion. This mode allows the chip to convert at its maximum rate with no unnecessary delays. As an example of calculating the sequence time, if all eight operations used Immediate Execute with a gain of 1 and a 13-bit conversion, the time to execute one sequence (all eight operations) would be $8 * 110 = 880$ internal clocks.

Intra-Sequence Pause (001) — This mode provides a way for the microprocessor to initiate conversions, rather than the other modes which either initiate conversions from internal timings or an external pulse. At the Start of Operation the ISQ status bit is set. The microprocessor will recognize the setting of the ISQ status bit either by polling the Status register, or having enabled the ISQ interrupt. The Start of Conversion is delayed until the ISQAK bit in the Interrupt Acknowledge register is set.

Start on Next Time out (010) — After the Start of Operation occurs the Start of Conversion is delayed until the internal timer decrements from 1 to 0. When using this mode the timer will be free-running. This means that the timer is initialized in the Halt mode and left alone to decrement and reload automatically when in the Run mode. This mode can be used to establish a specific sampling rate. Note that the timer value must be greater than the conversion time, therefore this mode can only slow the sampling rate down from its maximum rate. In the case where several operations are used, and only one of them uses this mode, the timer value must be greater than all the other Operation Execution times plus the current operation conversion time.

Preset Timer/Start on Time out (011) — At the Start of Operation the timer is loaded with its pre-programmed count. The delay between the Start of Operation and the Start of Conversion is the pre-programmed count. Execution time of the operation is the pre-programmed timer count plus the conversion time. As opposed to mode 2, the timer can be any value between 2 and 2^{16} ; i.e., there is no restriction on the timer value being greater than the conversion time. One

application of this mode would be when an external analog event is triggered by the SYNC pulse, and the conversion needs to be delayed by a programmable amount of time.

Using the External SYNC Input — The following description applies to modes 4 and 5, since these two modes use the external SYNC input. These modes should only be used when the SYNC pin is programmed as an input (MSTR bit in Control register is 0). If the external SYNC signal arrives before the Start of Operation, it may be latched depending upon the arrival time. If it arrives 22 clocks after the previous operation's Start of Conversion, external SYNC will be latched; any time before will miss the pulse. Therefore the external SYNC pulse rate should not be any faster than the frequency of the operations which use this mode, otherwise there will be more external SYNC pulses than conversions.

External SYNC Start (100) — After the Start of Operation, the Start of Conversion is delayed until the rising edge of the SYNC pulse and the next rising edge of the internal clock. Unless the rising edge of the external SYNC is synchronized with the internal clock (See t_{SYNCCK} Spec), the aperture uncertainty is one internal clock.

External SYNC/Timer Preset/Time out (101) — For this mode, the external SYNC pulse presets the timer, and when the timer times out the Start of Conversion begins. The timer is preset after the rising edge of the external SYNC and the next rising edge of the internal clock. When the timer transitions from 1 to 0, the Start of Conversion begins. As in the previous mode, unless the rising edge of the external SYNC is synchronized with the internal clock, the aperture uncertainty is one internal clock.

Timer Functions of the Different Modes — The on-chip timer is started when RUN is asserted. It then free-runs, pre-loads and restarts itself at the pre-programmed count unless one of the modes in an operation word specifies a timer preset. If

“Start on Next Timeout” mode is selected for all operations, the timer free-runs and subsequently starts conversions on regular intervals, without the inclusion of any variable overhead timing requirements of any specific operation. The “preset timer” function that can be specified in any operation, functions as a “one-shot” time out feature; however it can upset the regularity of conversions. The use of the external SYNC start allows flexibility with asynchronous conditions outside the chip. In addition, the use of time out with external SYNC allows synchronous operation of multiple Micro Linear chips with interleaved operation. If a different rate is desired other than increments of one master clock cycle ($1/2$ the CLK pin frequency) or if external events need to be counted before starting an operation, then setting the t_{CLK} bit in the control register will divert the timer to the t_{CLK} pin for all operations.

Timer Holding Register — Secondary Register 16 — This register holds the pre-programmed value of the timer. The value is in 1 internal clock increments, or the period of t_{CLK} if this input is used. The timer is a countdown timer, therefore the realized delay will be the number loaded into the Timer Holding register multiplied by the clock period. The value is written as a 16-bit binary word, and either high or low bytes can be written first. These registers are both writable and

2

readable, with register writes executed only when the chip is in the Halt mode (RUN bit cleared in the control register). Reading the Timer Holding register will return the pre-programmed value for the timer, it will not provide the actual timer value. Timer Holding register values of 1 or 0 are illegal and will "hang up" the timer when placed in Run mode. Therefore the minimum value that can be loaded into the Timer holding register is 2. The timer is decrementing when in Run mode and idle when in Halt mode. When the chip is placed in Run mode, the timer is automatically loaded with the value in the Timer Holding Register, and begins to count down.

Alarm Registers—Secondary Registers 17, 18 (Read/Write)—These registers specify the alarm criteria against which the converted data of a current operation is compared. The comparison occurs only when the ALRMEN bit is set within the operation. Secondary register 17 is Alarm A and secondary register 18 is Alarm B. These values are written in two's complement format, right justified and sign extended (refer to Data Format for more information).

Alarm Criteria Register—Secondary Register 19 lower byte (Read/Write)—Specifies the compare criteria to be used with alarm registers A and B. Bit 0 specifies whether the comparison of alarm word A is to be greater than (setting the bit) or less than equal to (clearing the bit). Similarly, bit 1 specifies the same criteria for alarm word B. The criteria of the two groups can be "ANDed" or "ORed" together by clearing (OR) or setting (AND) bit 2. Bits 3 and 4 enable the alarm comparison for words A and B, respectively. Bits 5, 6, and 7 are unused and can be any value when written, always read as ones. The following table illustrates all of the possible combinations, X signifies don't care.

Bit Number					Test Done:
4	3	2	1	0	
ENB	ENA	AND	GB	GA	
0	0	X	X	X	No Test
0	1	X	X	0	≤A
0	1	X	X	1	>A
1	0	X	0	X	≤B
1	0	X	1	X	>B
1	1	0	0	0	≤B or ≤A
1	1	0	0	1	≤B or >A
1	1	0	1	0	>B or ≤A
1	1	0	1	1	>B or >A
1	1	1	0	0	≤B and ≤A
1	1	1	0	1	≤B and >A
1	1	1	1	0	>B and ≤A
1	1	1	1	1	>B and >A

Using the various criteria, the chip can discern whether a certain channel is inside or outside a band, or greater than or less than a value. Notifying the microprocessor can be done through an interrupt or by polling the status register.

Interrupt Mask—Secondary Register 19 Upper Byte (Read/Write)

D15	D14	D13	D12	D11	D10	D9	D8
INTL	CLCPIE	RNERIE	ISQIE	OVRNIE	ALRMIE	OVRGIE	DBRIE

This register is used to define which interrupt conditions are capable of setting the hardware interrupt pin and the INT bit of the Status register. The bits in the Interrupt Mask register are interrupt enable bits, meaning when the bits are set they enable the corresponding status bit to activate the hardware interrupt pin as well as the INT bit in the Status register. The INTL bit determines the polarity of the INT pin. If set, the INT pin becomes active low, with an open drain output. If clear, the INT pin becomes active high, with driving capability in both directions.

Secondary Registers 20 to 31—Undefined

These registers are undefined and will cause unpredictable results if read or written to.

4.0 SAMPLING RATES AND CONVERSION TIMES

To determine the sampling rate, one must first determine the sequence execution time. A sequence is defined as the number of operations or instructions used. Therefore the sequence execution time equals the sum of the individual operation execution times. The simplest case for determining the sampling rate is when only one operation is used in the sequence. Then the sampling period is the operation execution time. If all eight instructions are used in the sequence, the sampling rate would be the sequence rate multiplied by the number of times the channel was sampled in the sequence.

It is possible to sample one channel more frequently than another. For example, if every other operation sampled channel 0, while the remaining operations sampled channels 1, 2, and 3, the sampling rate for channel 0 would be four times the sampling rate of the other channels. If periodic sampling is important, one must be careful when sampling a channel multiple times in a sequence since different operations can have different execution times.

Example: Sampling Four Channels in a Burst Every 10ms

Using Mode 2 "Start on Next Time out" for Instruction 0 will establish the 10ms sampling rate, once the clock is initialized properly. Instructions 1, 2, and 3 can each use Mode 0 "Immediate Execution". For the ML2200, each instruction can sample a different channel, thus covering all four channels in a burst. For the ML2208, the same holds true except all eight channels can be sampled in a burst, periodically.

Assuming the external clock is 7MHz and each conversion is 13 bits with a gain of 1, the conversion time for each operation will be $110 \times 286 \text{ ns} = 31.4 \mu\text{s}$. Therefore four instructions will require $4 \times 31.4 \mu\text{s} = 125.7 \mu\text{s}$. The execution time is much less than the sampling rate, thus the timer can be used to set the sampling rate. The timer value for a 10ms sample rate is: $10 \text{ ms} / 286 \text{ ns} = 35,000$ decimal or 88B8H.

Operation or Instruction Execution Time

Figure 17 illustrates the Operation Execution Time. The time between the Start of Operation and Start of Conversion is variable and depends on the Mode chosen. For more information on how to determine the time between Start of Operation and Start of Conversion refer to the Secondary registers Mode field description in the Instruction RAM.

The Conversion Execution time depends on the Cycle, the Gain, and the Mode chosen in the instruction word. Modes 0–5 all behave the same way when it comes to Conversion Execution Time. To help determine the Conversion Execution Time the following table gives the number of internal clocks used for Modes 0–5 based on the Cycle chosen.

Cycle	Number of Internal System Clocks Needed ($1/f_{CLKI}$)
16-Bit	128
13-Bit	110
8-Bit	80
Read CAL	80
Write CAL	52

Add 4 extra clocks to the Cycle time for each gain of 2 (including Read CAL and Write CAL). For a gain of 2 add 4 extra clocks, for gain of 4 add 8 extra clocks, for gain of 8 add 12 extra clocks. Example: Modes 0–5, Cycle = 13-bit conversion with a gain of 8. Conversion Execution time is 122 internal clocks.

5.0 MICROPROCESSOR INITIALIZATION PROCEDURE

The following sequence of steps is recommended when initializing the ML2200 from the microprocessor:

- 1) Keep reset active for at least 10 internal clock cycles after power supplies have stabilized. If a software reset is issued, hold off microprocessor communications with the chip until the RESET bit in the control register is read back as cleared, which takes 4 internal clock cycles.
- 2) If desired, check the data register path by performing a write and read of the calibration register for all 8 operations. (This step is optional, but does provide user assurance of the integrity of the on-chip data paths.) The calibration register is a full 16-bit data path.
- 3) Perform a calibration by first enabling the CLCP interrupt in the Interrupt Mask register, then start the calibration by asserting the CAL bit in the Control register. Alternately, if an interrupt driven system is not desired, the interrupt status register can be polled 8260 internal clocks after the CAL bit has been set. The chip should not be polled during calibration.
- 4) Upon receiving the CLCP interrupt, acknowledge it. If desired, read back the calibration code to verify a successful calibration. Other diagnostics may be run at this time, however diagnostics are optional and not required.

- 5) Load the Instruction RAM, alarm criteria, interrupt conditions, and timer. Set the proper data transfer mode up (DMA, interrupt driven or polled mode.) Clear all status bits before setting the RUN bit.
- 6) Start the chip running by setting the RUN bit in the Control register. This may be done by ORing the RUN bit with the other bits already configured in the Control register; *however* do not set the CAL bit again or another calibration will take place. Writing a 0 to the CAL bit has no effect; it will still read 1.

5.1 Reset/Power-On Conditions

When applying power to the ML2200, DV_{CC} and AV_{CC} should never be powered-on at different times.

It is OK to assert both \overline{RD} and \overline{WR} during RESET time, but not legal to do so otherwise; this may damage the chip internally.

The following list specifies the affected registers on the chip after a reset is performed. Note that both hardware and software reset of the chip have identical effects.

All registers shown below are cleared (all bits 0):

Primary Registers:

- Index register (register 2)
- Control register (register 3)
- Status register (register 4)
- Sequence status (register 5)

Secondary Registers:

- Interrupt bit mask (upper half, register 19)
- Alarm criteria register (lower half, register 19)

All other registers will have random data in them after power-on. If a hardware or software reset is performed later, registers which are not listed above will be unchanged.

Re-calibration after a hardware or software reset is not necessary, since the calibration register remains the same after a reset. Only after a power-up is a calibration necessary. However the CAL bit in the Control Register will be cleared after a reset. Setting the RUN bit while the CAL bit is clear will cause the RNER bit to be set. But, if a calibration had been done before the reset, the RNER may be ignored.

5.2 Timer

If any of the operations require a timer function, (either a one-shot or regular conversion interval) then the timer value must be written. This is done by writing the index register value to 10 hexadecimal and writing the proper 16-bit time value to the window registers. The timer value must be greater than 1. If using mode 2 "Start on Next Time out" the timer value must be greater than the conversion time.

5.3 Limit Alarm Operation

The chip may be set up to watch for certain data conditions by enabling the proper interrupt bits in the Interrupt Mask register. These conditions include A/D overrange/underrange and user-defined alarm criteria. In order to use the alarms, the A and B alarm values must be defined. Note that since alarm registers A and B are 16 bits wide, 13-bit two's complement sign extended values must be loaded. (Refer to Data Format for more information). In order to further qualify alarm registers A and B, the Alarm Criteria register must be initialized.

5.4 Defining Interrupt Conditions

If the chip is used in polled situations, the interrupt mask bits need not be set unless the "OR" of the interrupt conditions, bit 7 in the Status register is used.

If the chip is used in interrupt mode rather than polled mode, the desired interrupt conditions should be considered. In addition to the interrupts specified for data comparison operations, several other interrupts can be defined in the Interrupt Mask register. The DBR bit can be set if the DBR pin is not used. This enables interrupts at the end of sequences for data transfer via the INT pin. The intra-sequence interrupt bit should be enabled if intra-sequence pauses are desired in any of the operations. Overrun error and run-time error bits should be enabled if trapping of these errors is desired.

Note that alarm A and B and overrange interrupts occur at any time within the sequence of operations. Due to the interrupt latency time of the microprocessor, multiple interrupts of this type within a sequence may be indistinguishable from each other. The A and B alarms should generally be used on only one operation so that its source can be determined with no ambiguity. Overrange interrupts can be handled by examining the data in the chip at the end of the sequence.

The INT pin polarity can be defined to be active high (bit 15 cleared in the Interrupt Mask register) or active low (bit 15 set). When active low is chosen, the INT pin is open drain without a pull-up. When active high, the INT pin is driven actively in both directions. The default condition is active high, and the INT pin is actively driven low during reset time.

6.0 METHODS OF DATA TRANSFER TO THE MICROPROCESSOR

There are several ways to handle the data output; polling, interrupt, or DMA. If interrupts are the method chosen, method 5) may be preferable. Method 5) DMA/Interrupt mode, does not require a DMA controller. It simply uses the DMA mode of the ML2200 or ML2208 which can be interfaced to an interrupt controller.

- 1) Intra-Sequence pause instruction is used when the microprocessor is not going to periodically/continuously read the data, but it will read the data at arbitrary times. The Table 1 below shows the op codes to sample all eight channels.

Table 1. Channels in an ML2208 at Arbitrary Times

	Last	ALRMEN	Mode	CHAN	Cycle	Gain	REF
SEQ0	0	0	Intra Sequence Pause	CH0	13	1	Internal
SEQ1	0	0	Immed Execute	CH1	13	1	Internal
SEQ2	0	0	Immed Execute	CH2	13	1	Internal
SEQ3	0	0	Immed Execute	CH3	13	1	Internal
SEQ4	0	0	Immed Execute	CH4	13	1	Internal
SEQ5	0	0	Immed Execute	CH5	13	1	Internal
SEQ6	0	0	Immed Execute	CH6	13	1	Internal
SEQ7	1	0	Immed Execute	CH7	13	1	Internal

Using these instructions the program begins when the RUN bit is set in the control register. Immediately after RUN is set, before the first conversion takes place, the ISQ bit in the status register is set. This indicates that the sequencer has paused. When the microprocessor wants to read a value on one or more of the channels it sets the ISQAK bit in the Interrupt Acknowledge register. The ML2208 then performs eight conversions back-to-back, jumps back to sequence 0, and sets the ISQ and DBR bits in the status register. The data from all eight channels is now available in the Data RAM. The next time a conversion is desired, once again the microprocessor sets ISQAK in the interrupt acknowledge register.

- 2) **Polled mode transfer** is done simply by polling the status register and examining the DBR bit to see if a sequence has been completed. The DBR1E interrupt mask bit need not be set, but an acknowledge should be done by setting DBRAK in the Interrupt Acknowledge register, otherwise an overrun error will occur. The CPU can just poll the INT bit in the Status register. Only the bits which are enabled in the Interrupt Mask register will set the INT status bit. When the INT bit is set, the CPU can examine the other status bits to determine which requests are active.
- 3) **Interrupt mode** can be implemented using the INT pin and enabling the desired interrupt conditions in the Interrupt Mask register. The polarity of the INT pin can be selected at the same time. If desired, DBR can be used as a second interrupt pin to signify the transfer of data only. This may be useful in systems with multiple and prioritized interrupt structures. If DBR is used, the DBR mask bit in the interrupt mask register should be disabled or cleared.
- 4) **DMA mode** can be implemented by setting the DMA enable bit in the control register and selecting high byte or low byte first by setting or clearing the LOBYT bit. The DBR pin is utilized as the DMA request, and will remain asserted until all data from the sequence is read.
- 5) **DMA/Interrupt mode.** DMA mode can also be used in non-DMA applications. Although this appears to be unconventional, it may actually be preferred over the interrupt mode because of its convenience and speed. One way to do this would be to use the DBR pin as an interrupt request but enable DMA mode in the DAP. When data is ready DBR interrupts the microprocessor. The

microprocessor then reads either window register the required number of times to drain the Data RAM. Using the DMA mode interrupt method over non-DMA mode interrupt method saves a lot of overhead. For example in non-DMA mode interrupt method (assuming AUTOI is set), the index register would have to be set on entry, and the DBRAK bit would have to be set each service routine. In DMA interrupt mode, neither the Index register nor the DBRAK bit would have to be set. These are handled automatically in DMA mode.

7.0 POWER-DOWN MODE

The chip can be powered-down by asserting the P_{DN} pin. It is advisable to place the chip in HALT mode first by clearing the RUN bit in the control register, however the chip will automatically go into Halt mode when powered-down. All analog circuits are powered-off; digital circuits are left in an idle state. All registers within the chip will retain their values down to a level of 2V between V_{CC} and GND.

Powering-up the chip is done by bringing P_{DN} high. The chip will be in Halt mode upon power-up. Note, however, that the

first 10ms of chip operation after a power-up will not be valid due to the settling of quiescent bias conditions within the on-chip's analog circuits. Any data that is returned for this period after power-up should be considered invalid. The user has the choice of either throwing away the first 10ms of data or waiting for 10ms and then setting the chip in RUN mode. The on-chip timer can be used for this purpose, if desired, by defining a sequence of dummy operations that last for the required delay, then re-writing the required operations for normal use.

Acknowledge register. D_{BR}AK should also be set sometime before the next sequence to prevent the OVRN bit from being set, however this is not necessary.

Note that the microprocessor cannot let the ML2200 sequencer run continuously, i.e., SEQ 0 would be changed to Immediate Execute and asynchronously read the Data RAM. The problem in this case would be that the microprocessor may read the data at the same time that the chip is updating it. That is why either polling, interrupt, or DMA transfer is required in a continuous run mode of operation.

APPENDIX A

Diagnostics

The ML2200 and ML2208 may be run through a diagnostic routine after power-up. The DAP provides software programmable diagnostics so that no external hardware is necessary. Diagnostics are not necessary. They are provided as an option to the user.

Self-Test Mode

Setting the SLFTST bit in the Control register redefines the CHAN field in the Instruction Word. This in effect changes the input to the Sample-and-Hold from the multiplexer input channels to internal points within the chip; such as V_{REF} and AGND. Conversions in the Self-Test Mode allow the user to determine how the Sample-and-Hold and A/D converter behave with known input signals. This can be useful as a diagnostic routine for a product in the field, or as a debugging feature during product development. Figure 16 illustrates the re-definition of the instruction word when SLFTST = 1.

1. **System Offset** - The positive and negative inputs to the Sample-and-Hold are tied to analog ground. With this setting, converted data will give the offset of the A/D converter and Sample-and-Hold combination.

2. **Internal Reference** - Connects the positive input of the Sample-and-Hold to V_{REF} and the negative input of the Sample-and-Hold to analog ground. The result of converting in this test mode is a value near positive full scale.
3. **Invert Internal Reference** - Connects the negative input of the Sample-and-Hold to V_{REF} and the positive input of the Sample-and-Hold to analog ground. The result of converting in this test mode is a value near negative full scale.
4. **Common Mode** - Both the positive and negative inputs of the Sample-and-Hold are tied to the internal V_{REF}. The result of a conversion in this test mode indicates how well the converter is rejecting a common mode signal.

Since setting the SLFTST bit merely changes the input to the Sample-and-Hold, conversions must be executed in order to read the results. This means placing the chip in the RUN mode and reading the results from the Data RAM. It is possible to run one sequence then halt the sequencer and read the results. The sequencer can be put in a "pause" via the Intra Sequence Pause Mode instruction. The following instructions accomplish this:

	LAST	ALRMEN	MODE	CHAN	CYCLE	GAIN	REF
SEQ 0	0	0	INTRA SEQ PAUSE	SYSTEM OFFSET	13	1	0
SEQ 1	0	0	IMMED EXECUTE	INT REF	13	1	0
SEQ 2	0	0	IMMED EXECUTE	MINUS INT REF	13	1	0
SEQ 3	1	0	IMMED EXECUTE	COMMON MODE	13	1	0

After the RUN bit is set, the ISQ bit in the status register is immediately set. Setting the ISQAK bit in the Interrupt Acknowledge register will allow the sequencer to continue. The next time the ISQ bit is set, the results may be read from the Data RAM.

Reading and Writing to the Calibration Register

The ML2200 and ML2208 architecture provides a way for the microprocessor to indirectly read and write to the A/D converter; specifically the Calibration register and the A/D's Data

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register. Figure 18 illustrates this architecture.

The instructions that cause these transfers are READ CAL CODE and WRITE CAL CODE; selected in the Cycle field of the instruction word when SLFTST = 0. WRITE CAL CODE transfers the contents of the Calibration Holding register into the A/D converter's Calibration register. READ CAL CODE transfers the contents of the Calibration Holding register through the A/D's Data register, into the Data Output register with the same location as the operation.

As a result of providing READ and WRITE CAL, it is possible to execute digital loopbacks through the Calibration register, A/D registers, and all 8 Data Output registers. These loopbacks provides user assurance that all of the paths are clear and there are no stuck bits.

Writing to the Calibration register changes the calibration of the A/D converter. Therefore a self calibration should be performed after executing a WRITE CAL CODE to ensure the A/D is properly calibrated.

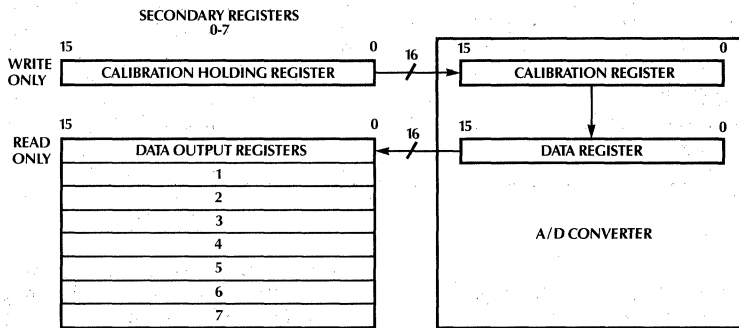


Figure 18. Digital Loopback Architecture

DIGITAL LOOPBACK ARCHITECTURE

Reading the calibration register provides a way for the microprocessor to determine that the self calibration was successful. The microprocessor configures the DAP to execute a READ CAL CODE after a self calibration has been performed. If the lower byte of data from the READ CAL CODE is anything other than all 1s, then the calibration was successful.

Even though the calibration register itself is a 16-bit register, and is capable of holding a 16-bit result, only the lower 9 bits are significant in determining the calibration code. These 9 bits have a sign magnitude format; in other words the 9th bit (MSB of the 9-bit word) is the sign bit, and the other eight bits are magnitude bits. An easy way to determine whether the calibration has passed or failed is to read the lower data byte after a READ CAL is executed. If it's not all 1s then the calibration was a success.

APPLICATIONS

Utilizing instruction RAM bits 0, 1, and 2, any of the differential input channels of the ML2200 can be programmed to sense the external reference (See Figure 13.) Only single ended channels 0 thru 5 can be used on the ML2208 (See Figure 14.)



Figure 19. Using a 2.5V External Reference

The system gain errors can be nulled by applying 2.4991V (the full-scale voltage minus 1.5LSB) to one of the input channels and adjusting R1 until the digital output toggles between 0 1111 1111 1110 and 0 1111 1111 1111. If offset is not adjusted the full-scale voltage will be shifted by the amount of this unadjusted offset voltage.

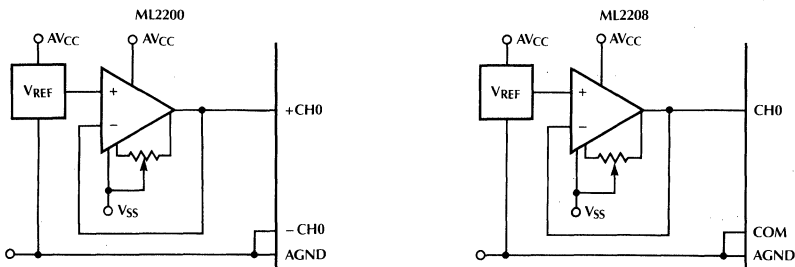


Figure 20. Adjusting Full-Scale Error

APPLICATIONS (Continued)

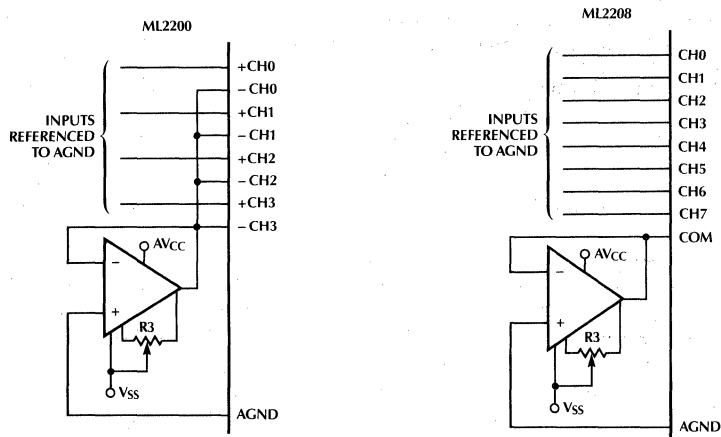


Figure 21. Adjusting Zero Error

An op amp with an offset adjustment with a range of at least ± 1.3 mV is required, like an OP-27. The Zero Error can be nulled by first applying $305 \mu\text{V}$ to one of the input channels (referenced to AGND.) $305 \mu\text{V}$ is equivalent to $1/2$ LSB which is the ideal input voltage which should cause the digital output to toggle from 0 0000 0000 0000 to 0 0000 0000 0001. Adjust R3 until this occurs.

If an external reference is also being used, it should be referenced to AGND, while the COM or negative inputs are tied to the offset op amp as shown above. In this configuration, the offset adjustment will effect the gain setting and so should be set first.

The Channel to Channel Zero Error and Full-Scale Error are very low and do not need to be adjusted separately. If, however, the input signal sources have their own different offsets, a separate op amp, with an offset adjustment, can be placed at each channel input.

APPLICATIONS (Continued)

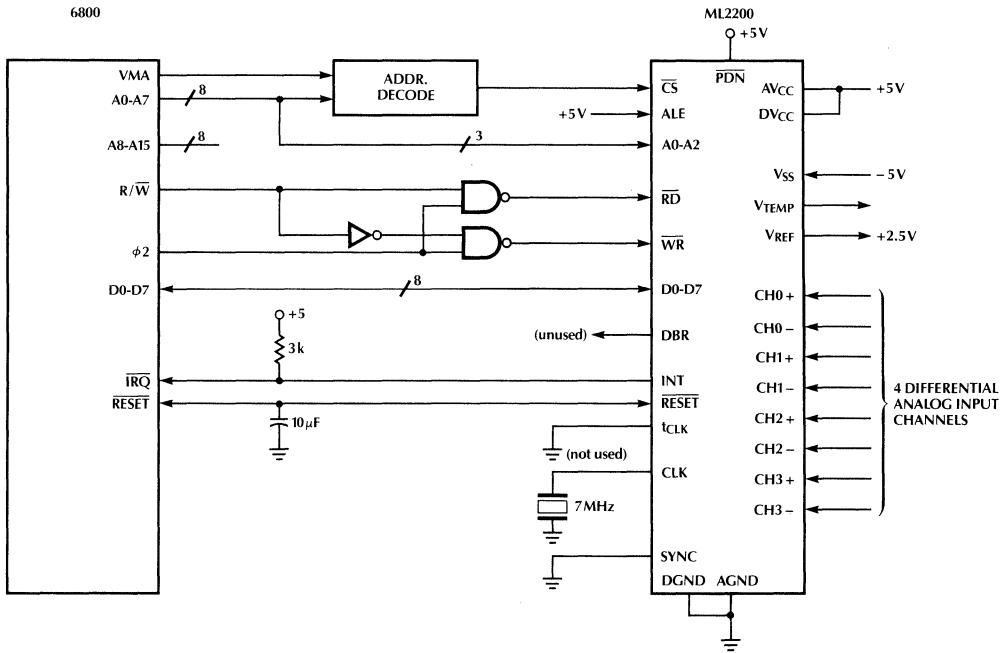


Figure 22. Interfacing ML2200 to 6800 Type Microprocessors

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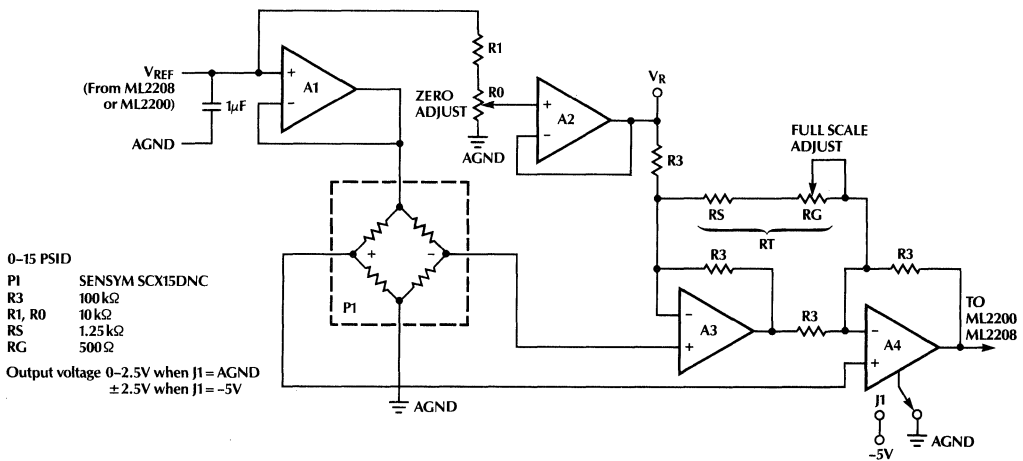


Figure 23. Pressure Sensor Application

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ORDERING INFORMATION

PART NUMBER	LINEARITY ERROR	MINIMUM CONVERSION	PACKAGE	TEMPERATURE RANGE
Four Differential Analog Inputs				
ML2200BCJ ML2200CCJ ML2200DCJ	$\pm 3/4$ LSB ± 1 LSB ± 1 LSB	31.5 μ s 31.5 μ s 44.0 μ s	Hermetic DIP (J40)	0°C to +70°C
Eight Single Ended Analog Inputs				
ML2208BCJ ML2208CCJ ML2208DCJ	$\pm 3/4$ LSB ± 1 LSB ± 1 LSB	31.5 μ s 31.5 μ s 44.0 μ s	Hermetic DIP (J40)	0°C to +70°C

Serial Peripheral Interface (SPI) 12-Bit Plus Sign A/D Converter with S/H

GENERAL DESCRIPTION

The ML2221 is a member of Micro Linear's 12-bit plus sign CMOS A/D converter family utilizing a self-calibrating algorithmic SAR technique. All errors of the sample-and-hold are accounted for in the analog-to-digital converter's accuracy specification.

These A/D converters have a maximum nonlinearity error over temperature of $\pm 0.009\%$ or $\pm 0.012\%$ of minus full scale to plus full scale.

The serial interface is compatible with industry standard serial interfaces. The ML2221 has 4 modes of operation: gated serial data clock, gated chip select, chip select to initiate conversion with serial out data controlled by ML2221, and free run mode.

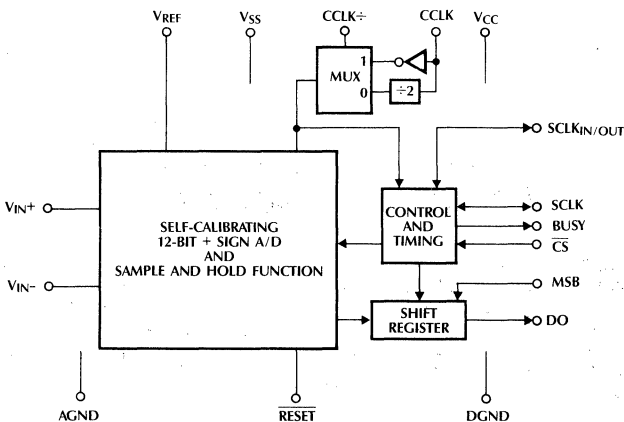
The serial interface allows either MSB or LSB first data with 2's complement output coding. For easy interface to microprocessors and shift registers the output data word is 16 bits.

FEATURES

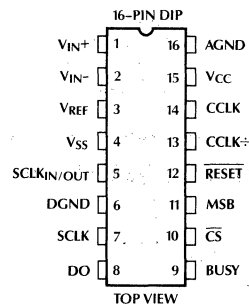
- Nonlinearity error $\pm 3/4$ LSB and ± 1 LSB max
- Conversion time (including S/H acquisition) $44\mu\text{s}$ max
- Harmonic distortion 0.01%
- No missing codes
- Inputs withstand $|7V|$ beyond supplies
- Bipolar $-5V$ to $+5V$ analog input range
- Controlled or free run operation
- Direct 4-wire interface to μP (MPU) with synchronous serial formats
- 0°C to $+70^\circ\text{C}$, -40°C to $+85^\circ\text{C}$ temperature range
- 16-pin DIP

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BLOCK DIAGRAM



PIN CONNECTION



ML2221

PIN DESCRIPTION

PIN NO.	NAME	FUNCTION	PIN NO.	NAME	FUNCTION
1	V _{IN+}	Positive Differential Analog Input; range = V _{SS} ≤ V _{IN+} ≤ V _{CC} ; (V _{IN+}) - (V _{IN-}) ≤ V _{REF} .	11	MSB	Most Significant Bit is transmitted first if MSB is tied to V _{CC} ; Least Significant Bit transmitted first if MSB is tied to DGND.
2	V _{IN-}	Negative Differential Analog Input; range = V _{SS} ≤ V _{IN-} ≤ V _{CC} ; (V _{IN+}) - (V _{IN-}) ≤ V _{REF} .	12	RESET	Active Low Reset. The RESET period is set by the time constant of the internal 50K pull up resistor and an external capacitor. After the RESET period the converter will be ready for accepting requests or will automatically start conversions/transmissions based upon the mode.
3	V _{REF}	Voltage Reference Input; referenced to analog ground.	13	CCLK÷	Sets CCLK equal to internal clock if tied to 5V. If tied to 0V the internal clock equals CCLK/2.
4	V _{SS}	Negative Supply -5V ± 5%; decouple to AGND.	14	CCLK	With CCLK equal to the internal CLK the user can synchronize to all internal timing events, although CCLK duty cycle must be controlled to meet the minimum clock high and low times specified.
5	SCLK _{IN/OUT}	SCLK mode select SCLK _{IN/OUT} = 5V; SCLK is an input serial CLK. SCLK _{IN/OUT} = 0V; SCLK is an output serial CLK.	15	V _{CC}	Clock Input. Internal clock can be generated by tying a crystal from this pin to DGND or applying a clock directly to the pin.
6	DGND	Digital Ground.	16	AGND	Positive Supply. +5V ± 5% decouple to AGND.
7	SCLK	Bi-Directional Serial Data Clock. Serial data is transmitted by the clock present at SCLK.			Analog Ground 0 Volts. Common mode reference point of the internal differential circuitry.
8	DO	Data Out. Digital output which contains result of A/D conversion. The serial data is clocked out on falling edges of SCLK.			
9	BUSY	Three-state active high BUSY status output. Normally low. Goes high to indicate that a conversion is in progress; de-asserted when conversion is complete and data is available from the conversion just completed. A pull-down resistor is recommended on this pin.			
10	CS	Active Low Chip Select, starts a conversion and brings the BUSY and DO _{out} of the three-state mode. CS is used in modes where conversion or transmission timing is controlled; held low in gated SCLK and FREERUN modes.			

ABSOLUTE MAXIMUM RATINGS

(Note 1)

Supply Voltage (V _{CC})	6.0V
Negative Supply Voltage (V _{SS})	-6.0V
Voltage at Analog Inputs	V _{SS} - 7V to V _{CC} + 7V
Voltage at V _{REF}	V _{SS} - 7V to V _{CC} + 7V
Input Current per Digital Pin	±10mA
Input Current at Analog Inputs	±20mA
Storage Temperature Range	-65°C to +150°C
Package Dissipation at 25°C (Board Mount)	875mW
Lead Temperature (soldering 10 seconds)	
Dual-In-Line Package (Molded)	260°C
Dual-In-Line Package (Ceramic)	300°C

OPERATING CONDITIONS

(Note 2)

Temperature Range	T _{MIN} ≤ T _A ≤ T _{MAX}
ML2221BIJ, ML2221CIJ	-40°C to +85°C
ML2221BCP, ML2221CCP	0°C to +70°C
Supply Voltage (V _{CC})	4.5V _{DC} to 6.0V _{DC}
Negative Supply Voltage (V _{SS})	-4.5V _{DC} to -6.0V _{DC}
Reference Voltage (V _{REF})	V _{CC}

ELECTRICAL CHARACTERISTICS

The following specifications apply for $V_{CC} = +5V \pm 5\%$, $V_{SS} = -5V \pm 5\%$, $V_{REF} = +4.75V$, $V_{IN-} = AGND$, $V_{IN+} = -4.75V$ to $+4.75V$, $T_A = T_{MIN}$ to T_{MAX} unless otherwise specified.

PARAMETER	NOTES	CONDITIONS	ML2221BIJ, ML2221CIJ			ML2221BCP, ML2221CCP			UNITS
			MIN	TYP (Note 3)	MAX	MIN	TYP (Note 3)	MAX	
Converter Characteristics									
Linearity Error ML2221BXX ML2221CXX	4	$f_{CCLK} = 0.1$ to $5MHz$			$\pm\frac{3}{4}$ ± 1			$\pm\frac{3}{4}$ ± 1	LSB LSB
Unadjusted Zero Error ML2221BXX ML2221CXX	4				$\pm\frac{3}{4}$ ± 2			$\pm\frac{3}{4}$ ± 2	LSB LSB
Unadjusted Positive and Negative Full-Scale Error	4				± 5			± 4	LSB
Zero Error Temperature Coefficient				0.5			0.5		ppmFS/ °C
Gain Temperature Coefficient				10			10		ppmFS/ °C
Common Mode Rejection	5, 6		80			80			dB
Analog Input Source Resistance	4				2			2	k Ω
Analog Input Range	4	V_{IN+} Referred to V_{IN-}	$-V_{REF}$		$+V_{REF}$	$-V_{REF}$		$+V_{REF}$	V
Analog Input Leakage Current	4				100			100	nA
Voltage Reference Input Source Impedance	4				0.5			0.5	k Ω
Reference Input Leakage Current	4				100			100	nA
Digital and DC Characteristics									
Power Supply Current I_{CC}, V_{CC} I_{SS}, V_{SS}	4			30 18	50 30		30 18	50 30	mA mA
Power Supply Rejection V_{CC} V_{SS}	7	DC DC to 25kHz DC DC to 25kHz		80 50 80 50			80 50 80 50		dB dB dB dB
V_{ILCLK} , Clock Input Low Voltage	4				0.8			0.8	V
V_{IHCLK} , Clock Input High Voltage	4		3.5		V_{CC}	3.5		V_{CC}	V
I_{L1} , Input Leakage Current (CCLK)	4	$AGND \leq V_{IN} \leq V_{CC}$			± 200			± 200	μA
V_{IL} , Input Low Voltage	4				0.8			0.8	V
V_{IH} , Input High Voltage	4		2.0		V_{CC}	2.0		V_{CC}	V
V_{OL} , Output Low Voltage	4	$I_{OL} = 2.0mA$			0.45			0.45	V
V_{OH} , Output High Voltage	4	$I_{OH} = -400\mu A$	2.4			2.4			V
I_L , Input Leakage Current (except CCLK)	4	$DGND \leq V_{IN} \leq V_{CC}$			± 10			± 10	μA
I_{HFZ} , Output Leakage Current	4	$\overline{CS} \geq V_{IH}$			± 10			± 10	μA
C_I , Input Capacitance (all digital inputs)	5				10			10	pF
C_O , Output Capacitance (all digital outputs)	5				10			10	pF

ELECTRICAL CHARACTERISTICS (Continued)

SYMBOL	PARAMETER	NOTES	CONDITIONS	MIN	TYP (Note 3)	MAX	UNITS
AC Electrical Characteristics (Note 8)							
t_C	Conversion Time	4, 9	$f_{CCLK} = 5\text{MHz}$ (CCLK $\div = "0"$)	44			μs
	Sample and Hold Acquisition	4, 9	$f_{CCLK} = 5\text{MHz}$ (CCLK $\div = "0"$)	3.2			μs
f_{CCLK0}	Clock Frequency	5, 9	Crystal (CCLK $\div = "0"$)	3		5	MHz
			Driven (CCLK $\div = "0"$)	0.1		5	MHz
f_{CCLK}	Clock Duty Cycle	5, 9	Driven	40		60	%
f_{CCLK0}	Clock Width	5, 9	Driven (CCLK $\div = "0"$)	High	50		ns
				Low	50		ns
f_{CCLK1}	Clock Frequency	5, 9	Driven (CCLK $\div = "1"$)	0.05		2.5	MHz
f_{CCLK1}	Clock Width	5	Driven (CCLK $\div = "1"$)	High	150		ns
				Low	150		ns
t_{CSB}	\overline{CS} Low to BUSY Driven	4				85	ns
t_{CSBHZ}	\overline{CS} High to BUSY, Hi-Z	4				85	ns
$t_{SCLKB\bar{A}}$	SCLK High to BUSY	5	Gated SCLK			270	ns
$t_{CCLKB\bar{D}}$	CCLK Low to BUSY, Deassert	5				160	ns
$t_{SCLK, DO}$	Serial Clock Low to DO Valid/Hold	4				190	ns
$t_{CS, DO}$	\overline{CS} Low to DO Driven	4				85	ns
$t_{CS, DOHZ}$	\overline{CS} High to DO Hi-Z	4				85	ns
$t_{CS, CCLK}$	\overline{CS} Low Setup Time to CCLK	4	Immediate Conversion Start	0			ns
$t_{CS, SCLK}$	\overline{CS} Low Setup to SCLK Low for No-Delay Data Transmit	5				75	ns
$t_{CCLK, SCLK}$	CCLK to SCLK Output Delay	5	SCLK $_{IN/OUT} = "0"$			225	ns

Note 1: Absolute maximum ratings are limits beyond which the life of the integrated circuit may be impaired. All voltages unless otherwise specified are measured with respect to ground.

Note 2: 0°C to +70°C and -40°C to +85°C operating temperature range devices are 100% tested with temperature limits guaranteed by 100% testing, sampling, or by correlation with worst-case test conditions.

Note 3: Typicals are parametric norm at 25°C.

Note 4: Parameter guaranteed and 100% production tested.

Note 5: Parameter guaranteed. Parameters not 100% tested are not in outgoing quality level calculation.

Note 6: Common mode rejection is the ratio of the change in zero error to the change in common mode input voltage.

Note 7: Power supply rejection is the ratio of the change in zero error to the change in power supply voltage.

Note 8: All parameters measured from 0.8V to 2.0V, $C_L = 50\text{pF}$.

Note 9: Maximum frequency is $1/t_{CLK1}(\text{high}) + t_{CLK1}(\text{low}) + \text{rise} + \text{fall times}$, which must be $\leq 2.5\text{ MHz}$.

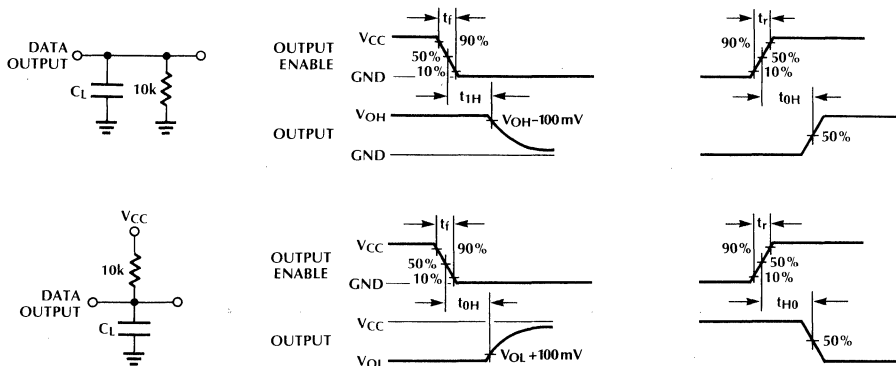


Figure 1. High Impedance Test Circuits and Waveforms

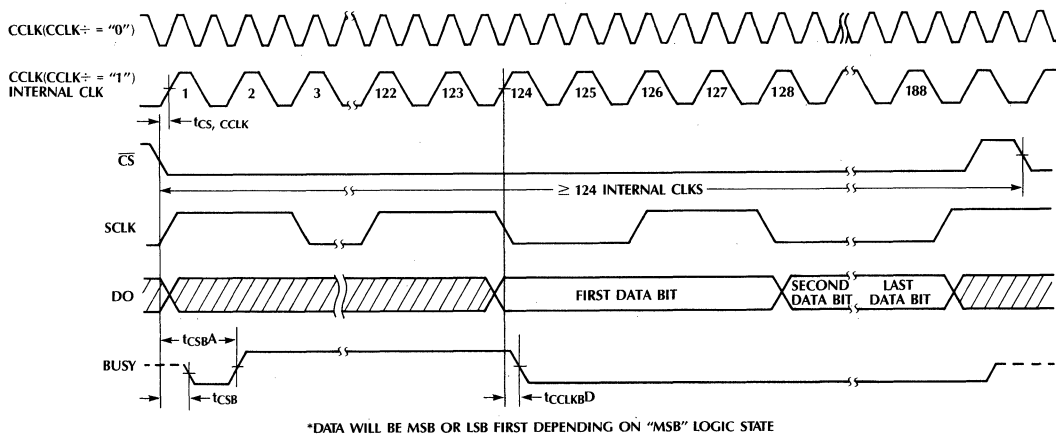


Figure 2. \overline{CS} , SCLK Sourced Mode

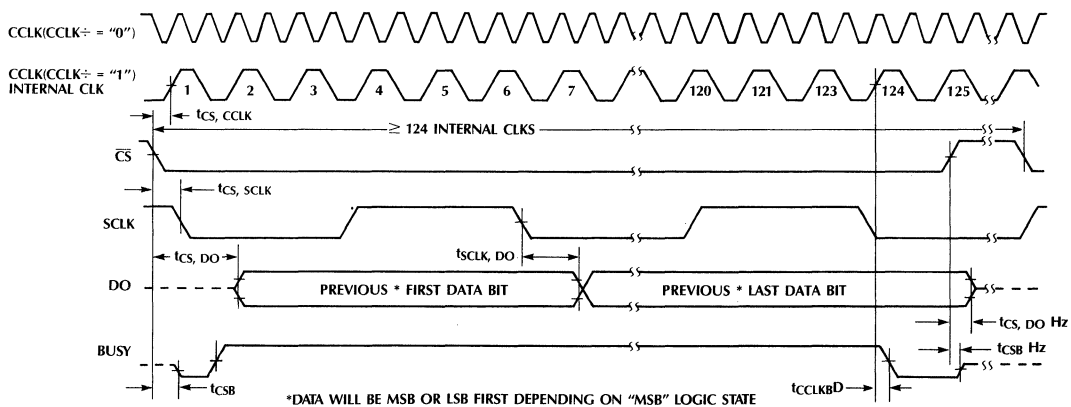


Figure 3. \overline{CS} , SCLK External Mode

2

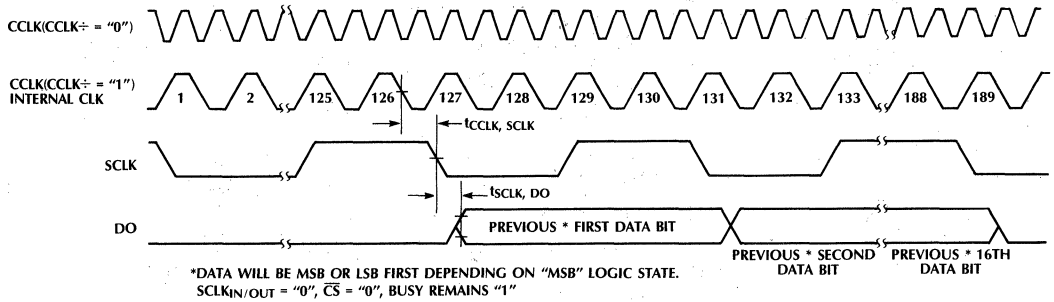


Figure 4. FREERUN Mode

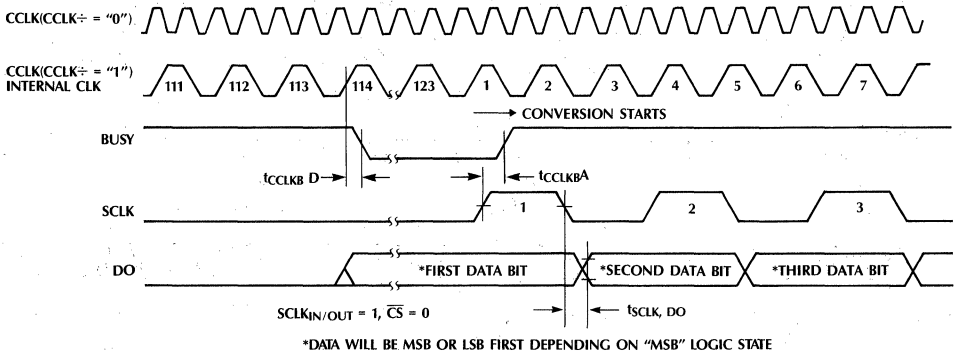


Figure 5. Gated SCLK Mode

1.0 FUNCTIONAL DESCRIPTION

1.1 ALGORITHMIC A/D CONVERTER

Micro Linear's algorithmic converter uses a successive approximation technique. Most of today's successive approximation converters use a DAC to feedback the approximated signal, however this technique requires more circuitry than algorithmic converters. In addition the values of all of the resistors or capacitors in the DAC must be matched to within the accuracy of the converter. This is difficult beyond 10 bits unless trimming is used. An algorithmic converter uses less circuitry and is more easily trimmed. Micro Linear's algorithmic converter is implemented using a 2x amplifier, a sample/hold amplifier and a comparator as shown in Figure 6.

The input sample is first multiplied by two then compared to the reference voltage. If the 2x input voltage is greater than the reference, the MSB is a 1 and the reference voltage is subtracted from the 2x input voltage. The remainder is stored in the sample and hold. If the 2x input voltage is less than the reference, the MSB is a 0 and the 2x input voltage is stored in the sample and hold. This process repeats again, however now the sample and hold voltage is multiplied by 2.

The algorithm involves multiplication by 2, comparison, and possibly subtraction. Referring to Figure 6, the algorithm for the circuit can be described as follows:

- Step 1 If $(2 \times V_{IN}) - V_{REF} \geq 0$
 then MSB = 1
 $(2 \times V_{IN}) - V_{REF} \rightarrow S/H$
 else MSB = 0
 $(2 \times V_{IN}) \rightarrow S/H$
- Step 2 If $(2 \times S/H) - V_{REF} \geq 0$
 then next bit = 1
 $(2 \times S/H) - V_{REF} \rightarrow S/H$
 else next bit = 0
 $(2 \times S/H) \rightarrow S/H$
- Step 3 Repeat Step 2 until conversion complete.

Since the A/D converter handles bipolar inputs, negative inputs are handled slightly differently using the same principle.

1.1.1 Self Calibration

In order to maintain integral and differential linearity in an algorithmic converter, two critical parameters need to be controlled, loop offsets and the gain of the loop. Loop offsets are automatically nulled before each conversion using auto-zeroing circuitry on both the sampling amplifier and the 2x amplifier. The gain of the loop is adjusted using self calibration.

Self calibrating the algorithmic converter, once the offsets have been nulled, is performed by measuring the 2x gain of the loop and adjusting it. The gain can be measured by converting the reference voltage at the input as well as the reference (V_{REF}/V_{REF}), and examining the output code. Converting V_{REF} should yield plus full scale, since V_{REF}/V_{REF} should equal 1. If the gain of the loop is slightly less than 2, the resulting LSB of the conversion will be "0". If the magnitude bits of the resulting conversion are all "1s", the gain may be too great, therefore the gain is reduced to the point where the threshold of the LSB is reached.

Adjustment of the 2x gain is done with the binary weighted trim capacitor arrays connected to each of the 2C input capacitors. A small value of capacitance is either added to or subtracted from the 2C input caps until the gain of the loop is within 13-bit accuracy.

Self calibration is done at the factory. The calibration process is not available at the finished product level.

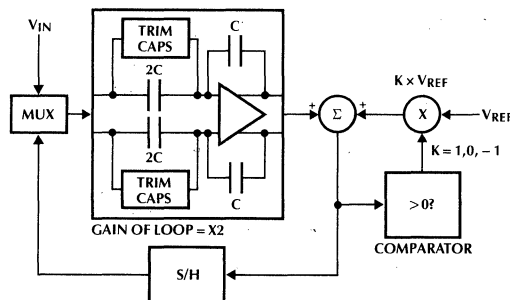


Figure 6. Self Calibrating A/D Converter

1.1.2 Conversion Times

The following table lists the conversion times which include the sample and hold acquisition time.

OPERATION MODE	INTERNAL CLOCKS*
\overline{CS} , SCLK External	124
\overline{CS} , SCLK Sourced	124
FREERUN	110
Gated SCLK	124

1.1.3 Sample and Hold Timing

Figure 7 shows the internal timing for the sample and hold circuitry. The relationship between the start of conversion and the input channel going into sample mode is fixed at 6 internal clocks*, regardless of the start mode. Six internal clocks after the start of conversion the sample and hold is switched into the sample mode, placing two 9pF capacitors in parallel with the input pins; one on V_{IN+} and one on V_{IN-} . The sample switch is kept in the sample mode for 8 internal clocks (3.2 μ s at a 5MHz external clock, if $CLK\div = 0$), then placed in the hold mode. During the next 2 internal clocks the charge on the sample and hold is transferred into the A/D, after which the V_{REF} pin is sampled for 8 internal clocks.

1.2 ANALOG INPUTS

1.2.1 Differential Inputs and Common Mode Rejection

The differential inputs of the ML2221 eliminate the effects of common mode input noise (60Hz for example), as V_{IN+} and V_{IN-} are sampled at the same time.

* For a description of internal clocks see Clock section.

1.2.2 Noise

The leads to the analog inputs should be kept as short as possible to minimize input noise. Noise as well as digital clocks can couple into the inputs and cause errors. Input filters can be used to reduce the effects of these sources.

1.2.3 Power Supply Decoupling

Low inductance tantalum capacitors of 1 μ F or greater and 0.01 μ F disc ceramic capacitors are recommended for bypassing V_{CC} as well as V_{SS} to AGND. These capacitors should be placed close to the V_{CC} and V_{SS} pins.

1.3 CONVERTER CLOCK

The CCLK input can be driven with an external clock or a crystal referenced to DGND. The crystal must be parallel resonant with minimum capacitive loading. (i.e., no bypass caps should be used and leads should be kept short)

If driven with external clock and if the CCLK pin is tied to V_{CC} the frequency must be between 50KHz to 2.5MHz with the requirement that clock LOW (t_{CCLKL}) and clock HIGH (t_{CCLKH}) durations must be more than 150ns. If the CCLK pin is tied to ground then the frequency can be from 100KHz to 5.0MHz.

For crystal operation with the divide by two flip flop bypassed, and there is a 30 to 70% variation in duty cycle of the oscillator, the maximum crystal frequency is 2.0MHz to insure that the minimum clock high and low times are greater than 150 nsec.

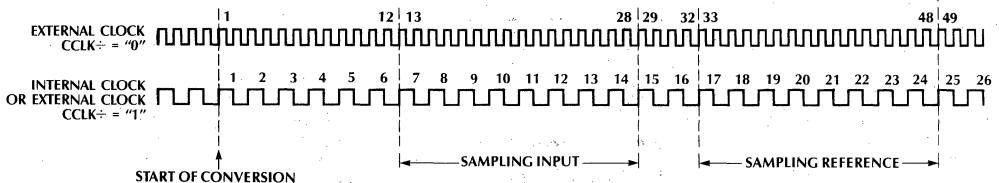


Figure 7. Sample and Hold Timing

1.4 RESET

The RESET pin has an internal 100K pullup resistor. Power supplies must be stable to within a ±5% tolerance before the reset condition is removed.

The active low hardware reset can be performed by a capacitor value (usually >6µF) tied to the RESET pin or by driving it with the system reset signal.

1.5 DIGITAL INTERFACE

All four synchronous interface modes of operation are determined by CS during reset period as follows:

Logic Level of CS During Reset	SCLK Mode Select (SCLK _{IN/OUT})	Serial Interface Mode
0	0	FREERUN
0	1	Gated SCLK
1	0	CS, SCLK Sourced
1	1	CS, SCLK External

After the reset time, the SCLK_{IN/OUT} pin can be changed to switch between either (FREERUN and Gated SCLK) or (CS, SCLK Sourced and CS, SCLK External).

The logic level of CS will not change the mode of operation of the ML2221 once the mode of operation is programmed during the RESET period.

1.5.1 CS, SCLK External Mode

CS starts a conversion. The SCLK is continuously driven into the ML2221 and data from the previous conversion is shifted out at the SCLK rate starting at the first SCLK falling edge from the CS assertion. CS is normally kept low for all 16 bits of data, but can be brought back high after the desired number of bits have been shifted out. CS should be held low for a minimum of 124 internal clocks (see Figure 8) for the conversion to complete.

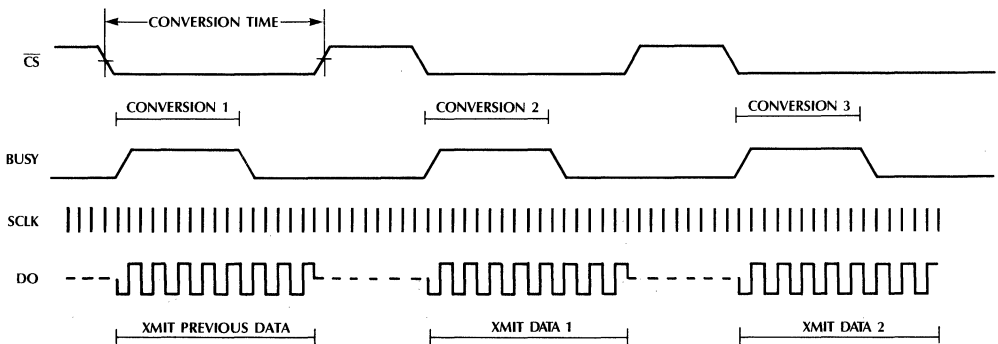
It takes 110 internal clocks to convert an analog signal into 13 bits of data plus 13 more clock periods to make data available. At a 5.0MHz clock and CCLK÷ = 0V, the maximum conversion rate is 49.6 microseconds or 124 internal converter clocks.

When CS is asserted (LOW) a conversion begins and the DO output becomes active. The ML2221 is ready to shift out the data serially.

The BUSY output is in the high impedance state when the ML2221 is not selected. When CS input goes low, the BUSY output is driven high or low depending on if a conversion is in progress. Once a conversion begins, BUSY is held active for 123 internal converter clocks.

The DO output is high impedance when the ML2221 is not selected. When CS input goes low, it is driven with the first bit of data initially, and then begins to put out all subsequent data bits on each FALLING edges of the serial clock (SCLK). Data is always output in 16 bit format: if the LSB is output first, the data is sign extended after 13 bits; if the MSB is output first, the data is zero-filled after 13 bits. DO remains driven as long as CS remains low.

a. Serial Transmission < Conversion Time



NOTE: CONVERSION TIME EQUALS 124 INTERNAL CLOCK OR CCLK'S IF CCLK÷ = "1"

Notes:

1. Use 10k pulldown resistor on BUSY pin to get "true" convert busy.
2. If CS is brought high in the middle of a serial data transmission, the data transmission is aborted and the data is reloaded into the output shifter.

Figure 8. CS, SCLK External Mode

1.5.2 \overline{CS} , SCLK Sourced Mode

Conversion is initiated by \overline{CS} . In this mode, SCLK is sourced by the ML2221. At the end of the conversion, the device will provide a packet of 16 SCLKs to transmit the 16 bits data stream (see Figure 9). The data rate at which the data is being transmitted is (internal clock)/4. For example, when $CCLK \div = 1$, $CCLK = 256kHz$, the data rate is 64Kbps.

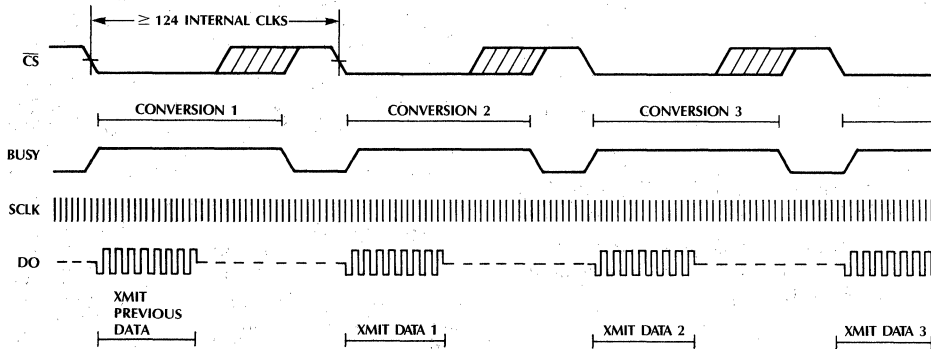
\overline{CS} should be held low during the entire conversion and the transmission sequence. The time required to convert an analog signal is 110 clocks, with additional

13 clocks to shift out the data. The total conversion time is therefore 123 clocks plus one bit delay. Data transmission will need 64 (16-bit x 4) internal clocks. Hence, the total clock cycles to complete one operation in this mode is 188 clocks. \overline{CS} therefore should be held low for a minimum of 188 clocks.

Example: If $CCLK = 5\text{ Mhz}$, $CCLK \div = 0$, the maximum conversion time will be $75.2\mu s$. Therefore, the maximum frequency for \overline{CS} is 13.3kHz.

In this mode, the data transmitted is always the current data.

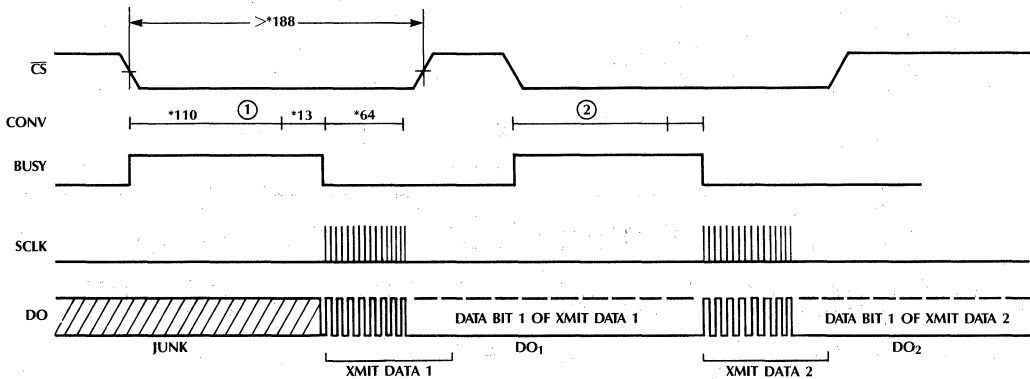
b. Serial Transmission > Conversion Time



Notes:

1. Use 10K pull-down resistor on BUSY pin to get "TRUE" convert BUSY status.
2. If \overline{CS} is brought HIGH in the middle of a serial data transmission, the data transmission is aborted and the data is reloaded into the output shifter.

Figure 8. \overline{CS} , SCLK External Mode



*NUMBER OF INTERNAL CONVERTER CLOCKS OR C CLK_S WITH C CLK_÷ = "1"

Notes:

1. Use 10K pull-down resistor on BUSY pin to get "TRUE" convert BUSY status.
2. If \overline{CS} is brought HIGH in the middle of a serial data transmission, the data transmission is aborted and the data is reloaded into the output shifter.

Figure 9. \overline{CS} , SCLK Sourced Mode

1.5.3 FREERUN Mode

The FREERUN mode executes continuous back-to-back conversions at the rate of 110 internal converter clocks per conversion, and outputs 16 bits of data and 16 corresponding SCLKs at the rate of 4 internal converter clocks per bit (see Figure 10). The ML2221 immediately begins converting after reset and starts outputting data after the first conversion. A conversion rate of 44 microseconds can be achieved by using the maximum CCLK frequency.

In the FREERUN mode, SCLK can not be sourced externally. The SCLK provided internally by the device is equal to (internal clock)/4. Since the converter is performing continuous conversion, BUSY is therefore always asserted.

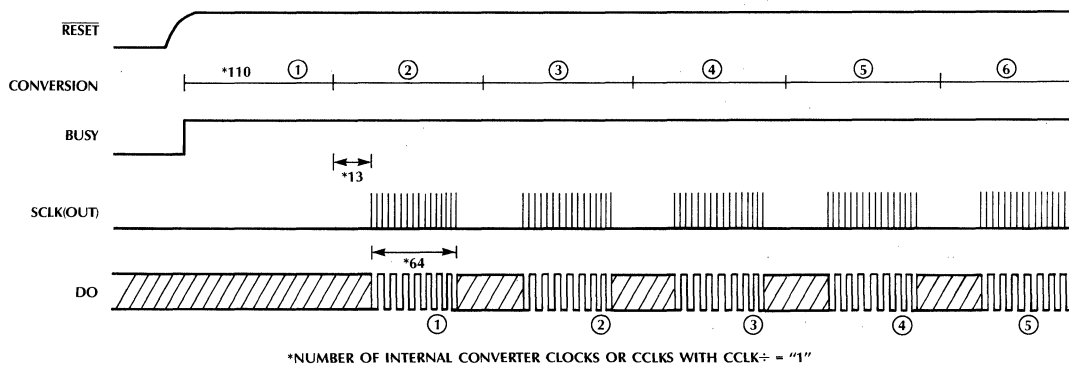
1.5.4 Gated SCLK Mode

In this mode of operation, an external SCLK source must be used. This external SCLK should be a 16 clock

packet which will be used by the converter to send out the data and initiate the conversion simultaneously (see Figure 11).

The data transmitted by ML2221 is the data from the previous conversion (see Figure 11). Therefore, in order to ensure integrity of the first data byte, the first SCLK signal should be initiated after a minimum of 124 internal clocks after reset. After the reception of the first SCLK signal, the converter will start the conversion process which is 124 clock as mentioned. Therefore the minimum time required between initiation of conversion by the SCLK should be no less than 124 clocks. In the case of maximum CCLK at 5MHz, the minimum time interval between two packets of SCLK should be 49.6 μ s.

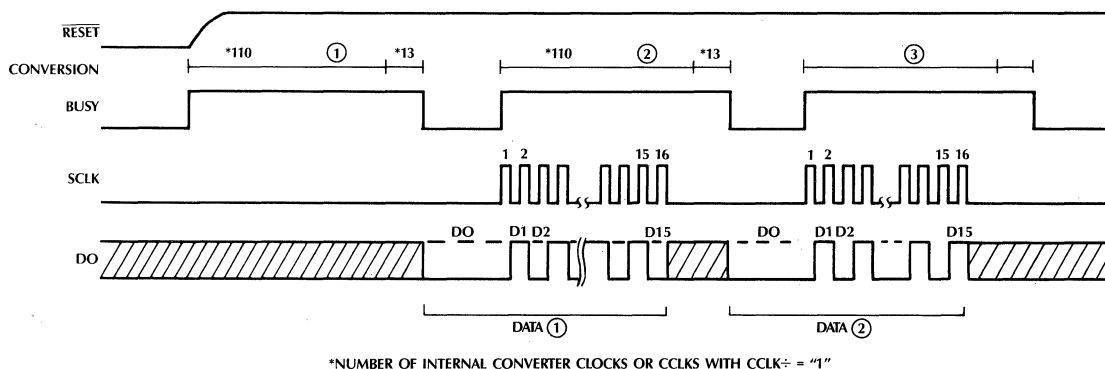
The BUSY output never floats and is asserted at the first SCLK and deasserted after 123 internal converter clocks. DO is always driven.



Note: DO is always driven.

Figure 10. FREERUN Mode

a. Serial Data Transmission < Conversion Time

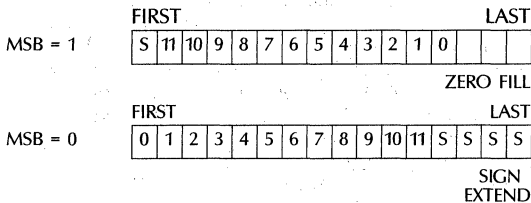


Note: Time from first SCLK₁ to seventeenth SCLK₁ must be greater than 124 internal converter clocks.

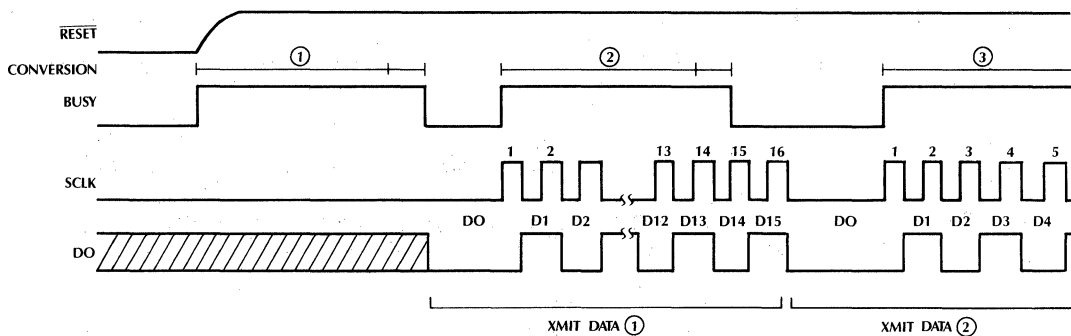
Figure 11. Gated SCLK Mode

1.6 DATA FORMAT

The MSB pin determines if the MSB or LSB data is transmitted first and in the following format. If more than 13 SCLK's occur.



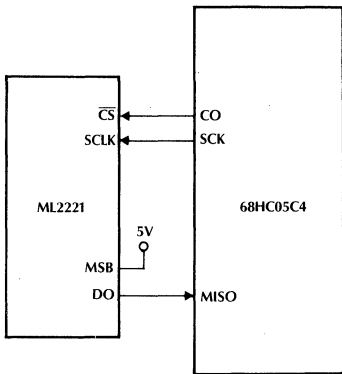
b. Serial Data Transmission > Conversion Time



Note: Time from first SCLK1 to seventeenth SCLK1 must be greater than 124 internal converter clocks.

Figure 11. Gated SCLK Mode

APPLICATIONS



START	MNEMONIC	INSTRUCTION
START	BCLRn	Bit 0 Port C goes low (\overline{CS} goes low)
	LDA	Load contents of SPI data register into Acc. (D_{OUT} MSBs)
	STA	Start next SPI cycle
	AND	Clear 3 MSBs of first D_{OUT} word
	STA	Store in memory location A (MSBs)
	TST	Test status of SPIF
	BPL	Loop to previous instruction if not done with transfer
	BSETn	Set B0 of Port C (\overline{CS} goes high)
	LDA	Load contents of SPI data register into Acc. (D_{OUT} LSBs)
	STA	Store in memory location A + 1 (LSBs)

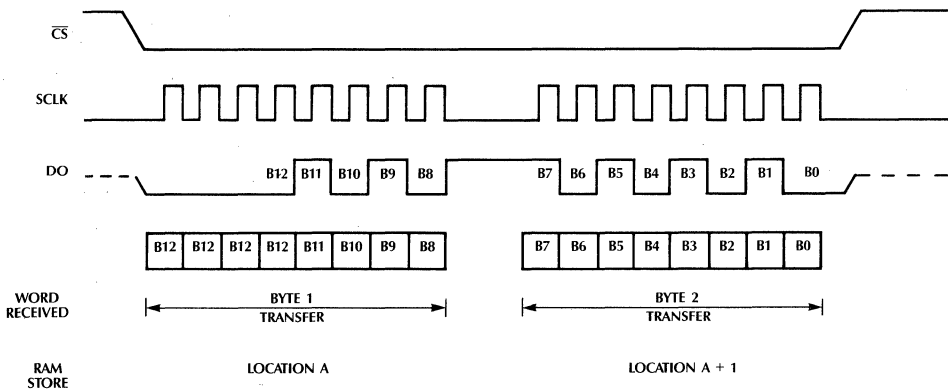


Figure 12. Interfacing to 68HC05C4 with a Dedicated Serial Port

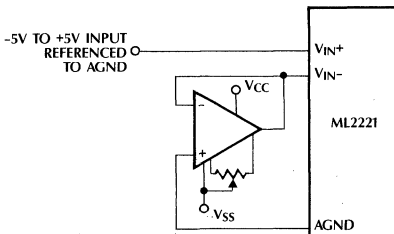


Figure 13. Adjusting Zero Error

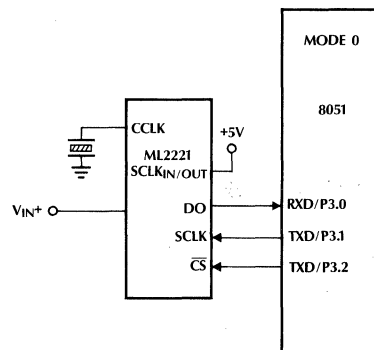


Figure 14. 1 Mbps 8051 Interface

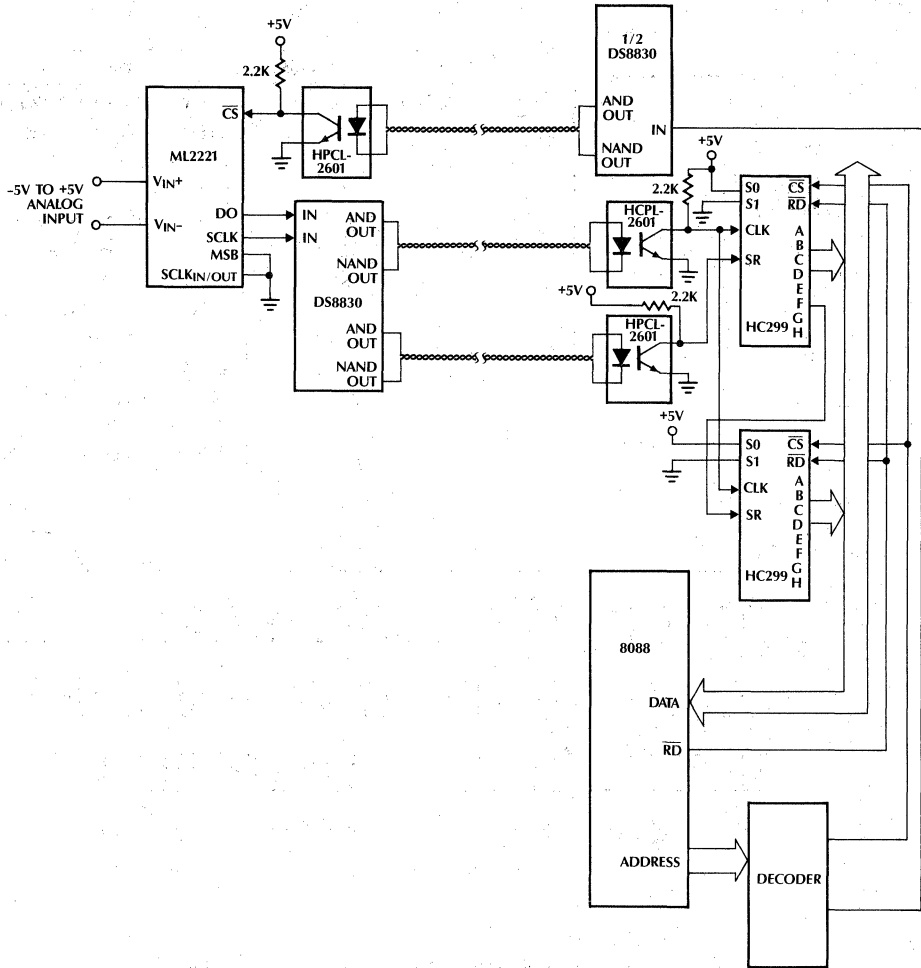


Figure 15. Optical Isolated 8088 Interface

ORDERING INFORMATION

PART NUMBER	LINEARITY ERROR	TOTAL UNADJUSTED ERROR	TEMPERATURE RANGE	PACKAGE
ML2221BCP	±1 LSB	±2½	0°C to +70°C	MOLDED DIP (P16)
ML2221BIJ			-40°C to +85°C	HERMETIC DIP (J16)
ML2221CCP	±¼ LSB	±1½	0°C to +70°C	MOLDED DIP (P16)
ML2221CIJ			-40°C to +85°C	HERMETIC DIP (J16)

Serial, CODEC/DSP Interface 12-Bit Plus Sign A/D Converter with S/H

GENERAL DESCRIPTION

The ML2222 is a member of Micro Linear's 12-bit plus sign CMOS A/D converter family utilizing a self-calibrating algorithmic SAR technique. All errors of the sample-and-hold are accounted for in the analog-to-digital converter's accuracy specification.

These A/D converters have a maximum nonlinearity error over temperature of $\pm 0.012\%$ or $\pm 0.024\%$ of full scale.

The CODEC serial interface is compatible with the communication industry standard protocol of PCM (Pulse Code Modulation). The ML2222 upon receiving the transmit frame synchronization pulse (FSX), shifts 16 bits of data.

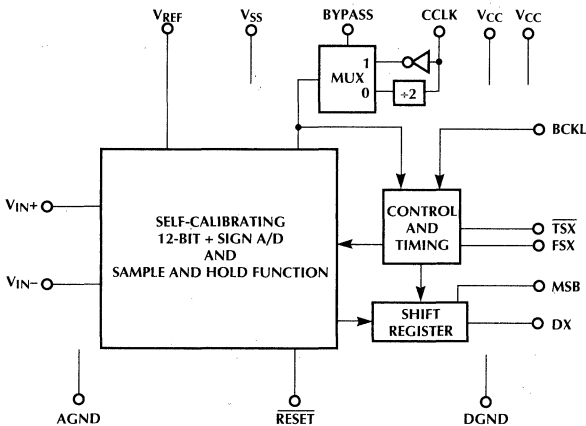
The transmit clock may vary from 64 KHz to 2.048 MHz.

The serial interface allows either MSB or LSB first data with 2's complement output coding. For easy interface to microprocessors and shift registers the output data word is 16 bits.

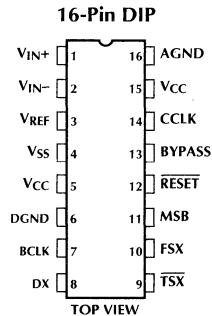
FEATURES

- Standard communication industry protocol for timing and frame sync
- Transmit clock from 64 KHz to 2.048 MHz
- Nonlinearity error $\pm 1/2$ LSB and ± 1 LSB max
- Conversion time (including S/H acquisition) 31.5 μ s max
- Harmonic distortion 0.01%
- No missing codes
- Self-calibrating — maintains accuracy over time and temperature
- Inputs withstand 17V! beyond supplies
- Bipolar $-2.5V$ to $+2.5V$ analog input range
- $0^{\circ}C$ to $+70^{\circ}C$, $-40^{\circ}C$ to $+85^{\circ}C$ temperature range
- 16-pin DIP

BLOCK DIAGRAM



PIN CONNECTION



Asynchronous Serial Interface 12-Bit Plus Sign A/D Converter with S/H

GENERAL DESCRIPTION

The ML2223 is a member of Micro Linear's 12-bit plus sign CMOS A/D converter family utilizing a self calibrating algorithmic SAR technique. All errors of the sample-and-hold are accounted for in the analog-to-digital converter's accuracy specification.

These A/D converters have a maximum nonlinearity error over temperature of $\pm 0.009\%$ or $\pm 0.012\%$ of full scale.

For easy interface to microprocessors, the ML2223 is designed to transmit data into RS-232 type ports.

The ML2223 operates in an asynchronous mode of operation. In this mode, the A/D continuously transmits 2 bytes in a 24-bit stream, inserting 8 idle bits between transmissions. When CURR input pin is tied high, transmission of the previous data begins immediately upon receiving a conversion start request. When CURR is low, transmission is started after a new conversion is completed.

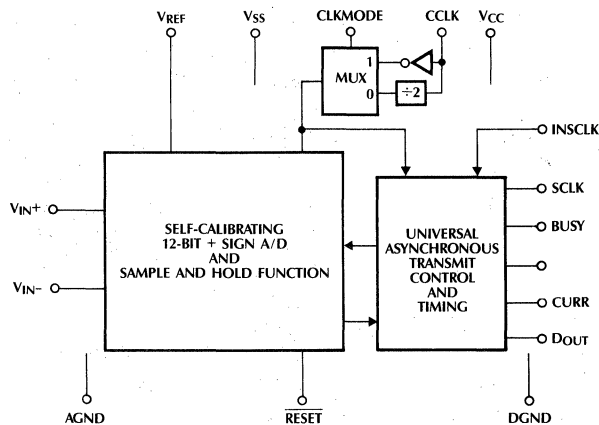
The serial data clock can be generated by the ML2223 or it can be provided by an external source.

The serial interface provides LSB first data with 2's complement output coding.

FEATURES

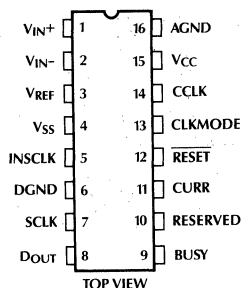
- RS-232 compatible asynchronous interface
- One-wire data transmission
- Continuous conversions
- Nonlinearity error $\pm 3/4$ LSB and ± 1 LSB max
- Conversion time (including S/H acquisition) $45.6\mu\text{s}$ max
- Bipolar -5V to $+5\text{V}$ analog input range with $\pm 5\text{V}$ power supplies
- Harmonic distortion 0.01%
- No missing codes
- Self calibrating — maintains accuracy over time and temperature
- Inputs withstand $|7\text{V}|$ beyond supplies
- 0°C to $+70^\circ\text{C}$, -40°C to $+85^\circ\text{C}$ temperature range
- Standard .3" 16-pin DIP

BLOCK DIAGRAM



PIN CONNECTION

16-Pin DIP



PIN DESCRIPTION

PIN NO.	NAME	FUNCTION	PIN NO.	NAME	FUNCTION
1	V_{IN+}	Positive Differential Analog Input; range = $V_{SS} \leq V_{IN+} \leq V_{CC}$; $ (V_{IN+}) - (V_{IN-}) \leq V_{REF}$.	9	BUSY	Three-state active high BUSY status output. Normally low. Goes high to indicate that a conversion is in progress; de-asserted when conversion is complete and data is available from the conversion just completed. A pull-down resistor is recommended on this pin.
2	V_{IN-}	Negative Differential Analog Input; range = $V_{SS} \leq V_{IN-} \leq V_{CC}$; $ (V_{IN+}) - (V_{IN-}) \leq V_{REF}$.	10	RESERVED	This pin should be tied to ground.
3	V_{REF}	Voltage Reference Input; referenced to analog ground.	11	CURR	Current or Previous Data Mode pin. When this pin is tied high, the data will be transmitted at the start of a conversion (previous data mode). When CURR is tied low, the data will then be transmitted at the completion of the conversion.
4	V_{SS}	Negative Supply $-5V \pm 5\%$; decouple to AGND.	12	$\overline{\text{RESET}}$	Active Low Reset. The RESET period is set by the time constant of the internal 100K pull up resistor and an external capacitor. After the RESET period the converter will be ready for accepting requests or will automatically start conversions/transmissions based upon the mode.
5	INSCLK	SCLK Mode Select This pin is used to select SCLK pin as an input or an output. When INSCLK is high, SCLK is an input. When INSCLK is low, SCLK becomes an output pin. SCLK will then provide a clock at 1/128 or 1/256 depending on how the CLKMODE pin is set-up.	13	CCLKMODE	Clock mode pin. When CLKMODE pin = 1, the internal converter clock = CCLK. When CLKMODE pin is tied low, the internal converter clock = CCLK/2.
6	DGND	Digital Ground.	14	CCLK	Clock Input. Internal clock can be generated by tying a crystal from this pin to DGND or applying a clock directly to the pin.
7	SCLK	Serial Data Transmit Clock. The serial data will always be transmitted at the frequency of the clock present at this pin. The SCLK pin can be programmed as an input or an output by using the INSCLK pin (pin 5). When the SCLK is used as an output pin. The data rate will be the internal converter clock divided by 128. When CLKMODE = 1, SCLK = CCLK/128. When CLKMODE = 0, SCLK = CCLK/256	15	V_{CC}	Positive Supply. $+5V \pm 5\%$ decouple to AGND.
8	DO	Data Out. Digital output which contains result of A/D conversion. The serial data is clocked out on falling edges of SCLK.	16	AGND	Analog Ground 0 Volts. Common mode reference point of the internal differential circuitry.

ABSOLUTE MAXIMUM RATINGS

(Note 1)

Supply Voltage (V_{CC})	6.0V
Negative Supply Voltage (V_{SS})	-6.0V
Voltage at Analog Inputs	$V_{SS} - 7V$ to $V_{CC} + 7V$
Voltage at V_{REF}	$V_{SS} - 7V$ to $V_{CC} + 7V$
Input Current per Digital Pin	$\pm 10mA$
Input Current at Analog Inputs	$\pm 20mA$
Storage Temperature Range	$-65^{\circ}C$ to $+150^{\circ}C$
Package Dissipation at $25^{\circ}C$ (Board Mount)	875mW
Lead Temperature (soldering 10 seconds)	
Dual-In-Line Package (Molded)	260 $^{\circ}C$
Dual-In-Line Package (Ceramic)	300 $^{\circ}C$

OPERATING CONDITIONS

(Note 2)

Temperature Range	$T_{MIN} \leq T_A \leq T_{MAX}$
ML2223BIJ, ML2223CIJ	$-40^{\circ}C$ to $+85^{\circ}C$
ML2223BCP, ML2223CCP	$0^{\circ}C$ to $+70^{\circ}C$
Supply Voltage (V_{CC})	$4.5V_{DC}$ to $6.0V_{DC}$
Negative Supply Voltage (V_{SS})	$-4.5V_{DC}$ to $-6.0V_{DC}$
Reference Voltage (V_{REF})	V_{CC}

ML2223

ELECTRICAL CHARACTERISTICS

The following specifications apply for $V_{CC} = +5V \pm 5\%$, $V_{SS} = -5V \pm 5\%$, $V_{REF} = +4.75V$, $V_{IN-} = AGND$, $V_{IN+} = -4.75V$ to $+4.75V$, $T_A = T_{MIN}$ to T_{MAX} unless otherwise specified.

PARAMETER	NOTES	CONDITIONS	ML2223BIJ, ML2223CIJ			ML2223BCP, ML2223CCP			UNITS
			MIN	TYP (Note 3)	MAX	MIN	TYP (Note 3)	MAX	
Converter Characteristics									
Linearity Error ML2223BXX ML2223CXX	4	$f_{CCLK} = 0.1 \leq 5\text{MHz}$			$\pm\frac{3}{4}$ ± 1			$\pm\frac{3}{4}$ ± 1	LSB LSB
Unadjusted Zero Error ML2223BXX ML2223CXX	4				$\pm\frac{3}{4}$ ± 2			$\pm\frac{3}{4}$ ± 2	LSB LSB
Unadjusted Positive and Negative Full-Scale Error	4				± 5			± 4	LSB
Zero Error Temperature Coefficient				0.5			0.5		ppmFS/ °C
Gain Temperature Coefficient				10			10		ppmFS/ °C
Common Mode Rejection	5, 6		80			80			dB
Analog Input Source Resistance	4				2			2	k Ω
Analog Input Range	4	V_{IN+} Referred to V_{IN-}	$-V_{REF}$		$+V_{REF}$	$-V_{REF}$		$+V_{REF}$	V
Analog Input Leakage Current	4				100			100	nA
Voltage Reference Input Source Impedance	4				0.5			0.5	k Ω
Reference Input Leakage Current	4				100			100	nA
Digital and DC Characteristics									
Power Supply Current I_{CC}, V_{CC} I_{SS}, V_{SS}	4				30 18	50 30		30 18 50 30	mA mA
Power Supply Rejection V_{CC} V_{SS}	7	DC DC to 25kHz DC DC to 25kHz			80 50 80 50			80 50 80 50	dB dB dB dB
V_{ILCLK} , Clock Input Low Voltage	4				0.8			0.8	V
V_{IHCLK} , Clock Input High Voltage	4		3.5		V_{CC}	3.5		V_{CC}	V
I_{L1} , Input Leakage Current (CCLK)	4	$DGND \leq V_{IN} \leq V_{CC}$			± 200			± 200	μA
V_{IL} , Input Low Voltage	4				0.8			0.8	V
V_{IH} , Input High Voltage	4		2.0		V_{CC}	2.0		V_{CC}	V
V_{OL} , Output Low Voltage	4	$I_{OL} = 2.0\text{mA}$			0.45			0.45	V
V_{OH} , Output High Voltage	4	$I_{OH} = -400\mu\text{A}$	2.4			2.4			V
I_L , Input Leakage Current (except CLK)	4	$DGND \leq V_{IN} \leq V_{CC}$			± 10			± 10	μA
I_{HI-Z} , Output Leakage Current	4	$\overline{CS} \geq V_{IH}$			± 10			± 10	μA
C_I , Input Capacitance (all digital inputs)	5				10			10	pF
C_O , Output Capacitance (all digital outputs)	5				10			10	pF

ELECTRICAL CHARACTERISTICS (Continued)

SYMBOL	PARAMETER	NOTES	CONDITIONS	MIN	TYP (Note 3)	MAX	UNITS
AC Electrical Characteristics (Note 8)							
t_C	Conversion Time	4, 9	$f_{CCLK} = 5\text{MHz}$ (CCLKMODE = "0")	45.6			μs
f_{CCLK0}	Clock Frequency	5, 9	Crystal (CCLKMODE = "0")	.3		5	MHz
			Driven (CCLKMODE = "0")	.1		5	MHz
f_{CCLK0}	Clock Width	5, 9	Driven (CCLKMODE = "0")	High	50		ns
				Low	50		ns
f_{CCLK1}	Clock Frequency	5	Driven (CCLKMODE = "1")	.05		2.5	MHz
f_{CCLK1}	Clock Width	5	Driven (CCLKMODE = "1")	High	150		ns
				Low	150		ns
$t_{SCLK, DO}$	Serial Clock Low to DO Valid/Hold	4				190	ns

Note 1: Absolute maximum ratings are limits beyond which the life of the integrated circuit may be impaired. All voltages unless otherwise specified are measured with respect to ground.

Note 2: 0°C to +70°C and -40°C to +85°C operating temperature range devices are 100% tested with temperature limits guaranteed by 100% testing, sampling, or by correlation with worst-case test conditions.

Note 3: Typicals are parametric norm at 25°C.

Note 4: Parameter guaranteed and 100% production tested.

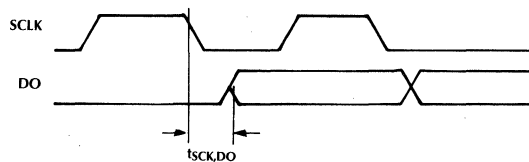
Note 5: Parameter guaranteed. Parameters not 100% tested are not in outgoing quality level calculation.

Note 6: Common mode rejection is the ratio of the change in zero error to the change in common mode input range.

Note 7: Power supply rejection is the ratio of the change in zero error to the change in power supply voltage.

Note 8: All parameters measured from 0.8V to 2.0V, $C_L = 50\text{pF}$.

Note 9: Maximum frequency is $1/t_{CLK1}(\text{high}) + t_{CLK1}(\text{low}) + \text{rise} + \text{fall times}$, which must be $\leq 2.5\text{ MHz}$.



Serial Clock to Data Out Delay

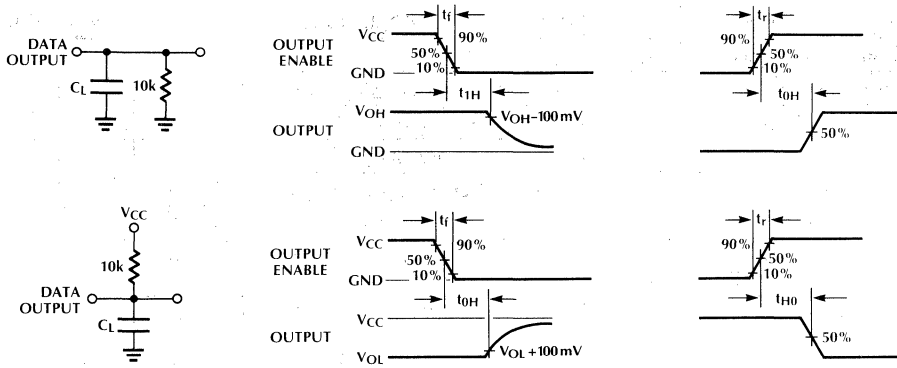


Figure 1. High Impedance Test Circuits and Waveforms

1.0 FUNCTIONAL DESCRIPTION

1.1 ALGORITHMIC A/D CONVERTER

Micro Linear's algorithmic converter uses a successive approximation technique. Most of today's successive approximation converters use a DAC to feedback the approximated signal, however this technique requires more circuitry than algorithmic converters. In addition the values of all of the resistors or capacitors in the DAC must be matched to within the accuracy of the converter. This is difficult beyond 10 bits unless trimming is used. An algorithmic converter uses less circuitry and is more easily trimmed. Micro Linear's algorithmic converter is implemented using a 2x amplifier, a sample/hold amplifier and a comparator as shown in Figure 2.

The input sample is first multiplied by two then compared to the reference voltage. If the 2x input voltage is greater than the reference, the MSB is a 1 and the reference voltage is subtracted from the 2x input voltage. The remainder is stored in the sample and hold. If the 2x input voltage is less than the

reference, the MSB is a 0 and the 2x input voltage is stored in the sample and hold. This process repeats again, however now the sample and hold voltage is multiplied by 2.

The algorithm involves multiplication by 2, comparison, and possibly subtraction. Referring to Figure 2, the algorithm for the circuit can be described as follows:

- Step 1 If $(2 \times V_{IN}) - V_{REF} \geq 0$
 then MSB = 1
 $(2 \times V_{IN}) - V_{REF} \rightarrow S/H$
 else MSB = 0
 $(2 \times V_{IN}) \rightarrow S/H$
- Step 2 If $(2 \times S/H) - V_{REF} \geq 0$
 then next bit = 1
 $(2 \times S/H) - V_{REF} \rightarrow S/H$
 else next bit = 0
 $(2 \times S/H) \rightarrow S/H$
- Step 3 Repeat Step 2 until conversion complete.

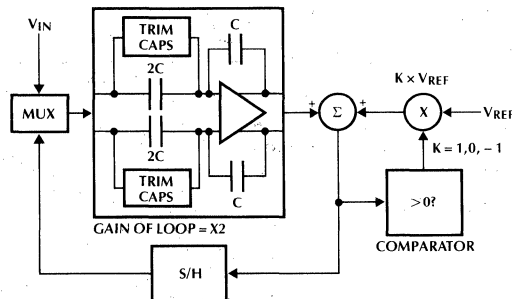


Figure 2. Self-Calibrating A/D Converter

Since the A/D converter handles bipolar inputs, negative inputs are handled slightly differently using the same principle.

1.1.1 Self Calibration

In order to maintain integral and differential linearity in an algorithmic converter, two critical parameters need to be controlled, loop offsets and the gain of the loop. Loop offsets are automatically nulled before each conversion using auto-zeroing circuitry on both the sampling amplifier and the 2x amplifier. The gain of the loop is adjusted using self calibration.

Self calibrating the algorithmic converter, once the offsets have been nulled, is performed by measuring the 2x gain of the loop and adjusting it. The gain can be measured by converting the reference voltage at the input as well as the reference (V_{REF}/V_{REF}), and examining the output code. Converting V_{REF} should yield plus full scale, since V_{REF}/V_{REF} should equal 1. If the gain of the loop is slightly less than 2, the resulting LSB of the conversion will be "0". If the magnitude bits of the resulting conversion are all "1s", the gain may be too great, therefore the gain is reduced to the point where the threshold of the LSB is reached.

Adjustment of the 2x gain is done with the binary weighted trim capacitor arrays connected to each of the 2C input capacitors. A small value of capacitance is either added to or subtracted from the 2C input caps until the gain of the loop is within 13-bit accuracy.

The self calibration and trimming is performed in the factory at wafer sort. This procedure is not available as a finished product.

1.1.2 Conversion Times

The following table lists the conversion times which include the sample and hold acquisition time.

OPERATION MODE	INTERNAL CLOCKS*
FREERUN	113

1.1.3 Sample and Hold Timing

Figure 3 shows the internal timing for the sample and hold circuitry. The relationship between the start of conversion and the input channel going into sample mode is fixed at 6 internal clocks.* Six internal clocks after the start of conversion the sample and hold is switched into the sample mode, placing two 9pF capacitors in parallel with the input pins; one on V_{IN+}

and one on V_{IN-} . The sample switch is kept in the sample mode for 8 internal clocks (3.2 μ s at a 5MHz external clock), then placed in the hold mode. During the next 2 internal clocks the charge on the sample and hold is transferred into the A/D, after which the V_{REF} pin is sampled for 8 internal clocks.

* For a description of internal clocks see Clock section.

1.2 ANALOG INPUTS

1.2.1 Differential Inputs and Common Mode Rejection

The differential inputs of the ML2223 eliminate the effects of common mode input noise (60Hz for example), as V_{IN+} and V_{IN-} are sampled at the same time.

1.2.2 Noise

The leads to the analog inputs should be kept as short as possible to minimize output noise. Noise as well as digital clocks can couple into the inputs and cause errors. Input filters can be used to reduce the effects of these sources.

1.2.3 Power Supply Decoupling

Low inductance tantalum capacitors of 1 μ F or greater and 0.01 μ F disc ceramic capacitors are recommended for bypassing V_{CC} as well as V_{SS} to AGND. These capacitors should be placed close to the V_{CC} and V_{SS} pins.

1.3 CONVERTER CLOCK

The CCLK input can be driven with an external clock or a crystal referenced to DGND. The crystal must be parallel resonant with minimum capacitive loading. (i.e., no bypass caps should be used and leads should be kept short)

If driven with external clock and if the CLKMODE pin is tied to V_{CC} , the frequency must be between 50KHz to 2.5MHz with the requirement that clock LOW (t_{CCLKL}) and clock HIGH (t_{CCLKH}) durations must be more than 150ns. If the CCLKMODE pin is tied to ground then the frequency can be from 100KHz to 5.0MHz.

For crystal operation with the divide by two flip flop bypassed, and there is a 40 to 60% variation in duty cycle of the oscillator, the maximum crystal frequency is 2.5MHz to insure that the minimum clock high and low times are greater than 150 nsec.

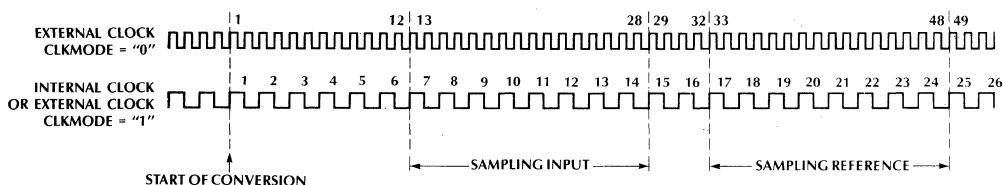


Figure 3. Sample and Hold Timing

1.4 RESET

The RESET pin has an internal 100K pullup resistor. Power supplies must be stable to within a $\pm 5\%$ tolerance before the reset condition is removed.

The active low hardware reset can be performed by a capacitor value (usually $>6\mu\text{F}$) tied to the RESET pin or by driving it with the system reset signal.

1.5 SCLK

The SCLK is used to clock out the data to the transmission line via the D_{OUT} pin. The SCLK can be supplied either externally or internally through the use of the INSCLK pin. Maximum SCLK frequency is 625kHz.

In applications where the internal SCLK is used, the SCLK is generated by dividing the internal clock by 128. For example, when 2.4576MHz crystal is used to generate CCLK, this will result in a 19.2Kbps data transmission rate (CLKMODE = 1). In the case where CLKMODE = 0, then the transmission rate will be 9.6Kbps.

1.6 CONVERTER OPERATION

There are two basic types of operation. By programming the CURR pin to "1," the device will be operating in the

"Previous Data Mode" and when CURR pin = "0," the device is in the "Current Data Mode."

The ML2223 performs continuous conversion. The conversion takes 110 clock periods and an additional 13 clock periods are required for the device to shift out the data. In the Current Data Mode, the data are transmitted after the conversion is completed. 24-bit time of the SCLK (transmit clock) is required to transmit the full data frame and an additional 8-bit time of idle is needed before the next conversion begins. A total of 32-bit time delay of the SCLK is therefore needed between conversions (see Figure 4).

In the Previous Data Mode, the data is transmitted at the start of the conversion (see Figure 5).

1.7 DATA FORMAT

The converter data is output in two data byte frame. Each frame has one start bit and two stop bits and each data byte consists of 8 bits of data and one parity bit. The data is transmitted LSB first (see Figure 6).

The first data byte transmitted is the least significant byte with even parity bit and the second byte is the most significant byte with odd parity. Thus, the UART can identify the lower byte or the upper byte by observing the parity error flag in accordance to the parity check it has set-up.

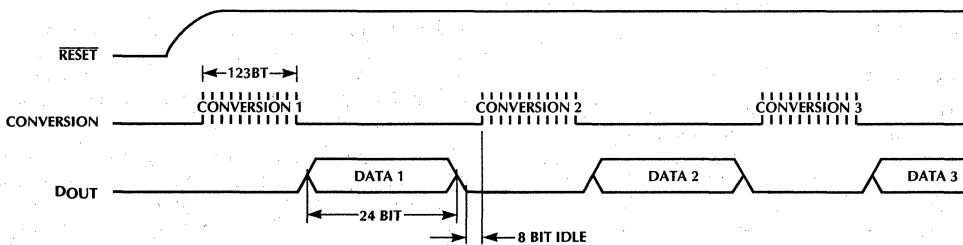


Figure 4. Current Data Mode.

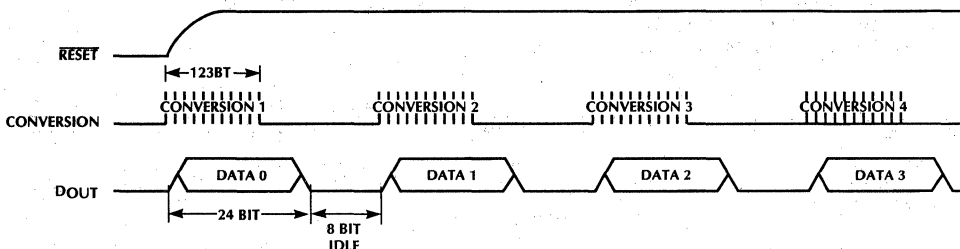


Figure 5. Previous Data Mode.

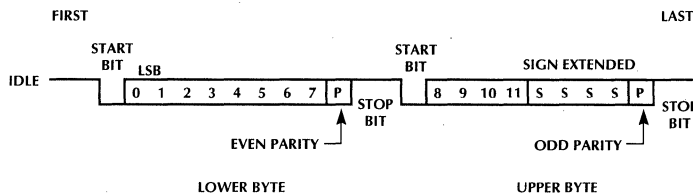


Figure 6. Data Format.

APPLICATIONS

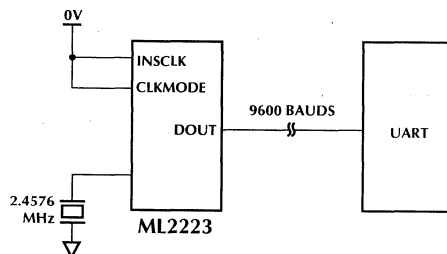


Figure 7. Remote Monitor System.

ORDERING INFORMATION

PART NUMBER	LINEARITY ERROR	TOTAL UNADJUSTED ERROR	TEMPERATURE RANGE	PACKAGE
ML2223BCP	±3/4 LSB	±1½	0°C to +70°C	MOLDED DIP (P16)
ML2223BIJ			-40°C to +85°C	HERMETIC DIP (J16)
ML2223CCP	±1 LSB	±2½	0°C to +70°C	MOLDED DIP (P16)
ML2223CIJ			-40°C to +85°C	HERMETIC DIP (J16)

ML2230

μ P Compatible 12-Bit Plus Sign A/D Converter with Sample and Hold

GENERAL DESCRIPTION

The ML2230 is a member of Micro Linear's 12-bit plus sign CMOS A/D converter family utilizing a self-calibrating algorithmic technique. The sample-and-hold, incorporated on the ML2230, has a differential input for noise immunity and power supply rejection. All errors of the sample-and-hold are accounted for in the analog-to-digital converter's accuracy specification.

The ML2230B has a maximum non-linearity error over temperature of 0.018% of full-scale, and the ML2230C and ML2230D have a maximum non-linearity error over temperature of 0.024% of full scale.

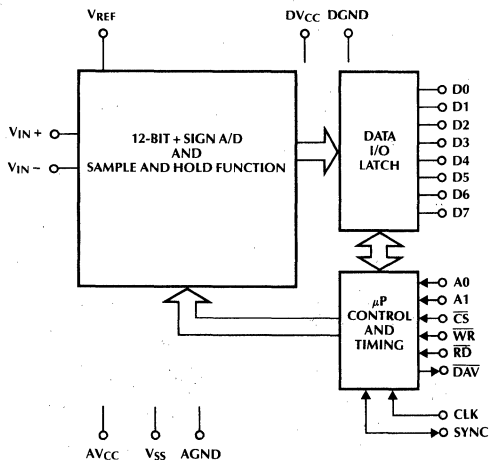
Designed to interface to an 8-bit microprocessor bus without additional components, the ML2230 outputs the 13-bit data result in two 8-bit bytes. Data format is 2's complement. All digital signals are fully TTL and CMOS compatible.

For interfacing to a 16-bit microprocessor bus the ML2233 provides a 13-bit data result.

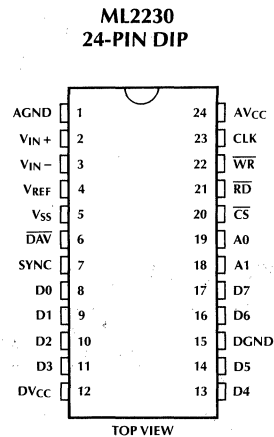
FEATURES

- Resolution 12-bits + sign
- Conversion time (including S/H acquisition) 31.5 μ s max
- Sample and hold acquisition 2.3 μ s max
- Non-linearity error $\pm 3/4$ LSB and ± 1 LSB max
- Low harmonic distortion 0.01%
- No missing codes
- Self-calibrating—maintains accuracy over time and temperature
- Inputs withstand |7V| beyond supplies
- Data transfer options—interrupt, DMA, or polling
- Outputs data in two 8-bit bytes
- Standard 24-pin DIP

BLOCK DIAGRAM



PIN CONNECTIONS



PIN DESCRIPTION

DIP	PCC	NAME	FUNCTION	DIP	PCC	NAME	FUNCTION
1	1	AGND	Analog ground.	13	16	D4	Bidirectional data bit.
2	2	V _{IN+}	Positive differential analog input; range = $V_{SS} \leq V_{IN+} \leq AV_{CC}$, $ (V_{IN+}) - (V_{IN-}) \leq V_{REF}$.	14	17	D5	Bidirectional data bit.
3	3	V _{IN-}	Negative differential analog input; range = $V_{SS} \leq V_{IN-} \leq AV_{CC}$, $ (V_{IN+}) - (V_{IN-}) \leq V_{REF}$.	15	18, 19	DGND	Digital ground.
4	4	V _{REF}	Voltage reference input; referenced to analog ground.	16	20	D6	Bidirectional data bit.
5	5	V _{SS}	Negative power supply; decouple to AGND.	17	21	D7	Bidirectional data bit.
6	8	\overline{DAV}	Data available; indicates a conversion has completed and data is available or calibration completed.	18	22	A1	Address for the microprocessor interface to access any one of the four registers.
7	9	SYNC	In the slave mode, SYNC is a positive edge triggered input used to start a conversion. In master mode, SYNC is an output and indicates a conversion has occurred.	19	23	A0	Address for the microprocessor interface to access any one of the four registers.
8	10	D0	Bidirectional data bit.	20	24	\overline{CS}	Chip select; enables writing to or reading from.
9	11	D1	Bidirectional data bit.	21	25	\overline{RD}	Read; enables ML2230 to drive data bus.
10	13	D2	Bidirectional data bit.	22	26	\overline{WR}	Write; allows writing into the registers.
11	14	D3	Bidirectional data bit.	23	27	CLK	Clock input. Drive with an external clock or crystal referenced to DGND. The crystal must be parallel resonant with minimum capacitive loading. (i.e., no bypass caps should be used and leads should be kept short.)
12	15	DV _{CC}	Digital power supply. Tie to AV _{CC} from same power supply.	24	28	AV _{CC}	Positive analog power supply. Decouple to AGND. Tie to DV _{CC} from same power supply.

ABSOLUTE MAXIMUM RATINGS
(Note 1)

Supply Voltages (AV _{CC} and DV _{CC})	6.0V
Negative Supply Voltage (V _{SS})	-6.0V
Voltage at Analog Inputs	V _{SS} - 7V to AV _{CC} + 7V
Voltage at V _{REF}	V _{SS} - 7V to AV _{CC} + 7V
Input Current per Digital Pin	±10mA
Input Current at Analog Inputs	±20mA
Storage Temperature Range	-65°C to +150°C
Package Dissipation @ 25°C	875mW
Lead Temperature soldering, Dual-In-Line Package (Ceramic)	300°C

OPERATING CONDITIONS (Note 2)

Temperature Range	0°C to 70°C
Supply Voltage (AV _{CC} and DV _{CC})	4.5V _{DC} to 6.0V _{DC}
Negative Supply Voltage (V _{SS})	-4.5V _{DC} to -6.0V _{DC}
Reference Voltage (V _{REF})	2.60V

ML2230

ELECTRICAL CHARACTERISTICS

The following specifications apply for $V_{CC} = +5V \pm 5\%$, $V_{SS} = -5V \pm 5\%$, $V_{REF} = +4.75V$, $V_{IN-} = AGND$, $V_{IN+} = -4.75V$ to $+4.75V$, $T_A = T_{MIN}$ to T_{MAX} unless otherwise specified.

PARAMETER	NOTES	CONDITIONS	MIN	TYP (Note 3)	MAX	UNITS
Converter Characteristics						
Linearity Error ML2230BXX ML2230CXX ML2230DXX	4	$f_{CCLK} = 0.1 \leq 7MHz$ $f_{CCLK} = 0.1 \leq 7MHz$ $f_{CCLK} = 0.1 \leq 5MHz$			$\pm 3/4$ ± 1 ± 1	LSB LSB LSB
Unadjusted Zero Error ML2230BXX ML2230CXX ML2230DXX	4				$\pm 3/4$ ± 2 ± 2	LSB LSB LSB
Unadjusted Positive and Negative Full Scale Error	5				± 4	LSB
Zero Error Temperature Coefficient				0.5		ppm/°C
Gain Temperature Coefficient				10		ppm/°C
Common-Mode Rejection	5, 6		80			dB
Analog Input Source Resistance	5				2	k Ω
Analog Input Range	4	V_{IN+} Refer to V_{IN-}	$-V_{REF}$		$+V_{REF}$	V
Analog Input Leakage Current	4				100	nA
Voltage Reference Input Source Impedance	5				0.5	k Ω
Reference Input Leakage Current	4				100	nA
Digital and DC Characteristics						
Power Supply Current I_{CC} , Analog V_{CC} I_{CC} , Digital V_{CC} I_{SS} , V_{SS}	4			30 10 18	50 30	mA μ A mA
Power Supply Rejection AV_{CC} V_{SS}	7	DC DC to 25kHz DC DC to 25kHz		80 50 80 50		dB dB dB dB
V_{ILCLK} , Clock Input Low Voltage	4				0.8	V
V_{IHCLK} , Clock Input High Voltage	4		3.5		AV_{CC}	V
I_{L1} , Input Leakage Current (CLK)	4	$AGND \leq V_{IN} \leq AV_{CC}$		± 200		μ A
V_{IL} , Input Low Voltage	4				0.8	V
V_{IH} , Input High Voltage	4		2.0		DV_{CC}	V
V_{OL} , Output Low Voltage	4	$I_{OL} = 2.0mA$			0.45	V
V_{OH} , Output High Voltage	4	$I_{OH} = -400\mu A$	2.4			V
I_L , Input Leakage Current (except CLK)	4	$AGND \leq V_{IN} \leq AV_{CC}$			± 10	μ A
I_{HI-Z} , Output Leakage Current (D0-D7)	4	$RD = CS = V_{IH}$			± 10	μ A
C_I , Input Capacitance (all digital inputs)				10		pF
C_O , Output Capacitance (outputs D0 to D7, and DAV)				10		pF

ELECTRICAL CHARACTERISTICS (Continued)

SYMBOL	PARAMETER	CONDITIONS		MIN	TYP (Note 3)	MAX	UNITS	
AC Electrical Characteristics (Note 8)								
t_C	Conversion Time	4, 9	CLK Mode = 0	$f_{CLK} = 70\text{MHz}$ $f_{CLK} = 5.0\text{MHz}$	31.5 44.0		μs μs	
	Sample and Hold Acquisition	4, 9	CLK Mode = 0	$f_{CLK} = 70\text{MHz}$ $f_{CLK} = 5.0\text{MHz}$		2.3 3.2	μs μs	
f_{CLK0}	Clock Frequency	5, 9	Crystal (CLK Mode = 0) Driven (CLK = Mode 0)		3 1	7 7	MHz MHz	
t_{CLK0}	Clock Width	5, 9	Driven (CLK Mode = 0)	High Low	50 50		ns ns	
f_{CLK1}	Clock Frequency	5, 10	Driven (CLK Mode = 1)		0.5	(Note 11)	MHz	
t_{CLK1}	Clock Width	5, 10	Driven (CLK Mode = 1)	High Low	125 125		ns ns	
t_{AD}	Address Stable to Valid Data	4			150		ns	
t_{AR}	Address Stable Before Read	4			0		ns	
t_{RA}	Address Hold After Read	4			0		ns	
t_{RR}	Read Pulse Width	5			150		ns	
t_{RD}	Read Access	4				150	ns	
t_{1Z}, t_{0Z}	Data Read to Hi-Z	4			0	50	ns	
t_{RV}	Recovery Between Two Reads or Writes	5			250		ns	
t_{RDCK}	Read to Clock Setup Time	5, 12			40		ns	
t_{AW}	Address Stable Before Write	4			0		ns	
t_{WA}	Address Hold After Write	4			0		ns	
t_{WW}	Write Pulse Width	4			150		ns	
t_{DW}	Data Setup Before Write Trailing Edge	4			100		ns	
t_{WD}	Data Hold After Write Trailing Edge	4			0		ns	
t_{WRCK}	Write to Clock Setup Time	5, 12			40		ns	
t_{CKDAV}	Clock to DAV Assert	4, 13	$C_L = 50\text{pF}$			120	220	ns
t_{SYNCK}	SYNC Input to Clock Setup	5, 12			40		ns	
t_{SYNCN}	SYNC Input Width	5	(CLK Mode = 0) (CLK Mode = 1)		6 3		$1/f_{CLK0}$ $1/f_{CLK1}$	
t_{CKSYNC}	External Clock to SYNC Output Delay	5, 13	$C_L = 50\text{pF}$			150	200	ns
t_{SYNCO}	SYNC Output Pulse Width	5, 13	(CLK Mode = 0) (CLK Mode = 1)			8 4	$1/f_{CLK0}$ $1/f_{CLK1}$	
t_{WRDAV}	Write Reg2 to DAV Rising Edge	4, 14	$C_L = 50\text{pF}$			170	ns	
t_{RDDAV}	Read Reg0 to DAV Rising Edge	4, 15	$C_L = 50\text{pF}$			170	ns	
t_r, t_f	Rise and Fall		All Inputs			25	ns	

Note 1: Absolute maximum ratings are limits beyond which the life of the integrated circuit may be impaired. All voltages unless otherwise specified are measured with respect to ground.

Note 2: Devices are 100% tested with temperature limits guaranteed by 100% testing, sampling or by correlation with worst-case test conditions.

Note 3: Typicals are parametric norm at 25°C.

Note 4: Parameter guaranteed and 100% production tested.

Note 5: Parameter guaranteed. Parameters not 100% tested are not in outgoing quality level calculation.

Note 6: Common mode rejection is the ratio of the change in zero error to the change in common mode input range.

Note 7: Power supply rejection is the ratio of the change in zero error to the change in power supply voltage.

Note 8: All parameters measured from 0.8V to 2.0V, $C_L = 100\text{pF}$.

Note 9: CK1X bit in control register = 0.

Note 10: CK1X bit in control register = 1.

Note 11: Maximum frequency is $1/t_{CLK1}$ (high) + t_{CLK1} (low) + rise + fall times and $\leq 3.5\text{MHz}$.

Note 12: Setup time required for synchronous start of conversion.

Note 13: In CLK mode = 0 (CK1X bit in control register = 0) start of conversion will occur at specified time; or time plus one f_{CLK0} period (see Figure 5).

Note 14: Writing a control register bit 0 with a one will acknowledge the DAV condition and de-assert DAV output.

Note 15: In start mode = 1, a read from location "0" will start the next conversion and de-assert the DAV output.

TIMING DIAGRAMS

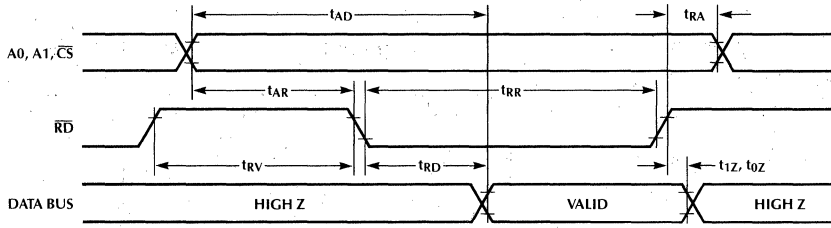


Figure 1. Read Cycle

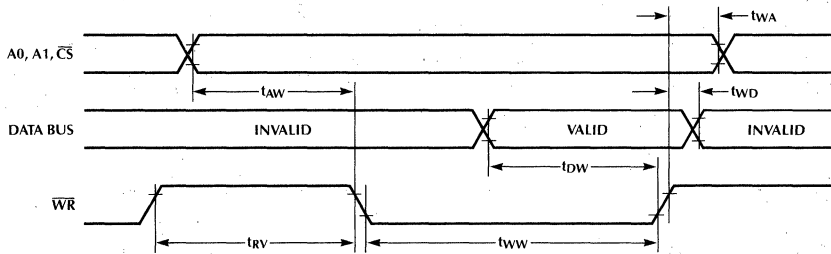
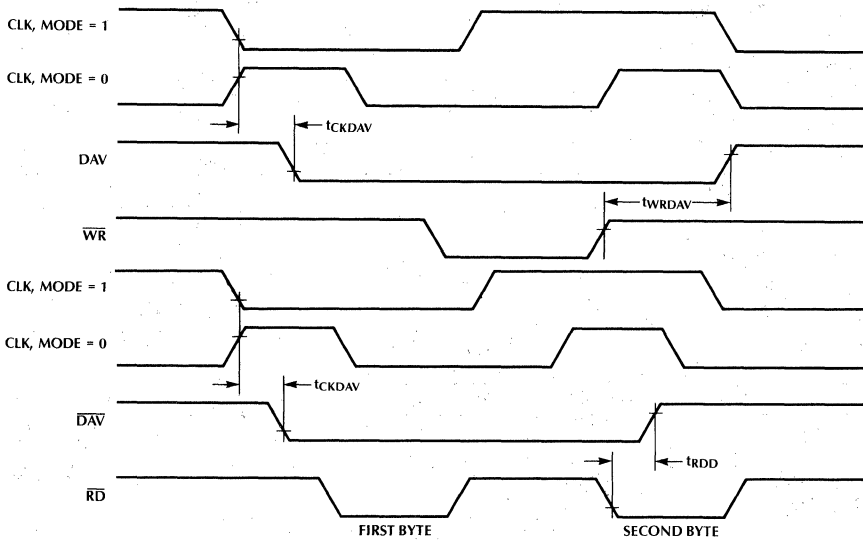


Figure 2. Write Cycle



DAV IS SET AND CLEARED BY INTERNAL CIRCUITRY.
NOTE: DMA BIT IN THE CONTROL REGISTER MUST BE SET FOR THIS OPERATION.

Figure 3. Data Available

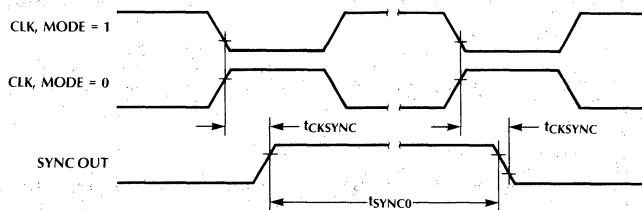
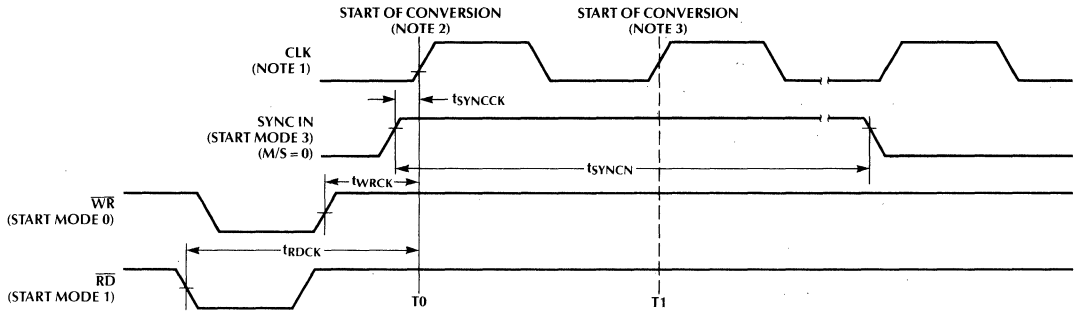


Figure 4. SYNC Output

TIMING DIAGRAMS (Continued)



- NOTES:
 1. CLK IS THE CLOCK DRIVEN AT THE CLOCK PIN.
 2. IN CLK MODE 1, WILL ALWAYS OCCUR AT T0 IF SETUP TIMES ARE MET.
 3. IN CLOCK MODE 0, WILL OCCUR EITHER AT T0 OR T1 IF SETUP TIMES ARE MET.

Figure 5. Start of Conversion (Start Mode 0,1,3)

BLOCK DIAGRAM

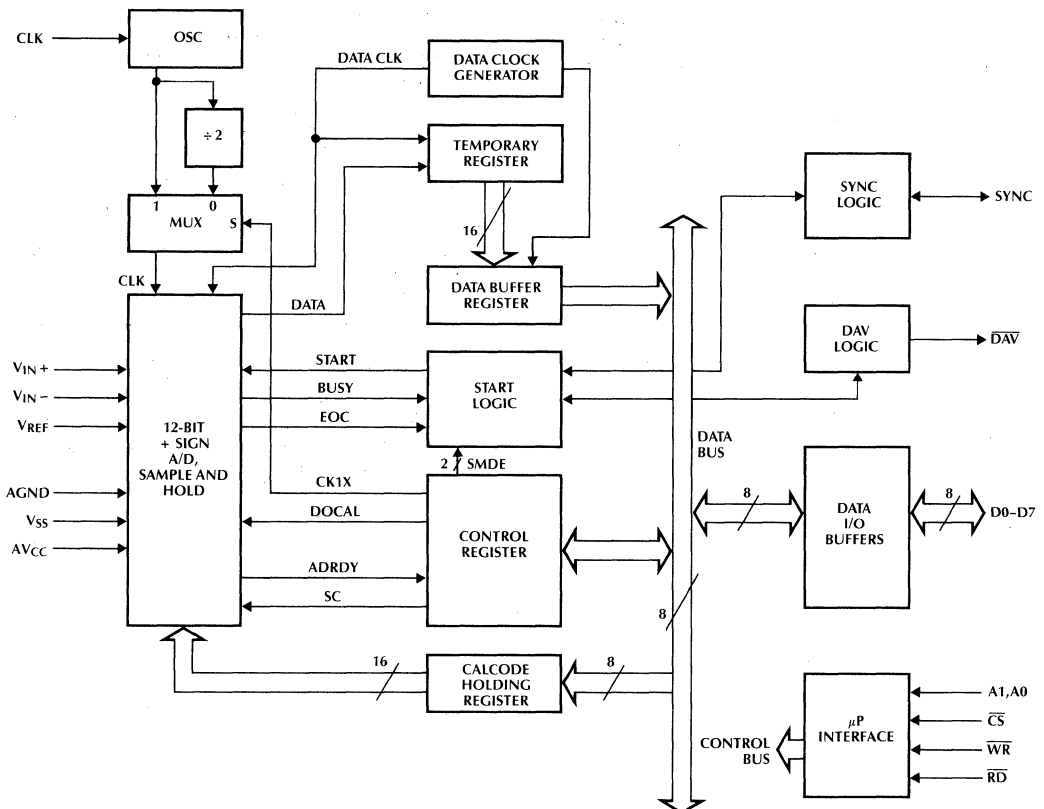


Figure 6. Block Schematic Diagram

FUNCTIONAL DESCRIPTION

ALGORITHMIC A/D CONVERTER

Micro Linear's algorithmic converter uses a successive approximation technique. Most of today's successive approximation converters use a DAC to feedback the approximated signal, however this technique requires more circuitry than algorithmic converters. In addition the values of all of the resistors or capacitors in the DAC must be matched to within the accuracy of the converter. This is difficult to do in silicon beyond 10 bits unless trimming is used. An algorithmic converter uses less circuitry and is more easily trimmed. Micro Linear's algorithmic converter is implemented using a 2x amplifier, a sample/hold amp, and a comparator as shown in Figure 7.

The input sample is first multiplied by two then compared to the reference voltage. If the 2x input voltage is greater than the reference, the MSB is a 1 and the reference voltage is subtracted from the 2x input voltage. The remainder is stored in the sample and hold. If the 2x input voltage is less than the reference, the MSB is a 0 and the 2x input voltage is stored in the sample and hold. This process repeats again, however now the sample and hold voltage is multiplied by 2.

The algorithm involves multiplication by 2, comparison, and possibly subtraction. Referring to Figure 6, the algorithm for the circuit can be described as follows:

Step 1 If $(2 \times V_{IN}) - V_{REF} \geq 0$
 then MSB = 1
 $(2 \times V_{IN}) - V_{REF} \rightarrow S/H$
 else MSB = 0
 $(2 \times V_{IN}) \rightarrow S/H$

Step 2 If $(2 \times S/H) - V_{REF} \geq 0$
 then next bit = 1
 $(2 \times S/H) - V_{REF} \rightarrow S/H$
 else next bit = 0
 $(2 \times S/H) \rightarrow S/H$

Step 3 Repeat Step 2 until conversion complete.

Since the A/D converter handles bipolar inputs, negative inputs are handled slightly differently using the same principle.

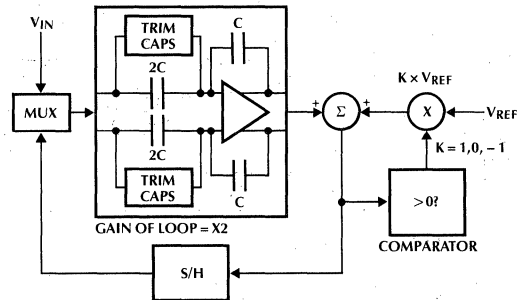


Figure 7. Self-Calibrating A/D Converter

SELF CALIBRATION

In order to maintain integral and differential linearity to the 1/2 LSB level in an algorithmic converter, two critical parameters need to be controlled, loop offsets and the gain of the loop. Loop offsets are automatically nulled before each conversion using auto-zeroing circuitry on both the sampling amp and the 2x amp. The gain of the loop is adjusted using self calibration.

Self-calibrating the algorithmic converter, once the offsets have been nulled, is performed by measuring the 2x gain of the loop and adjusting it. The gain can be measured by converting the reference voltage as the input as well as the reference (V_{REF}/V_{REF}), and examining the output code. Converting V_{REF} should yield plus full scale, since V_{REF}/V_{REF} should equal 1. If the gain of the loop is slightly less than 2, the resulting LSB of the conversion will be "0". If the magnitude bits of the resulting conversion are all "1s", the gain may be too great, therefore the gain is reduced to the point where the threshold of the LSB is reached.

Adjustment of the 2x gain is done with the binary weighted trim capacitor arrays connected to each of the 2C input capacitors. A small value of capacitance is either added to or subtracted from the 2C input caps until the gain of the loop is within 13 bit accuracy of 2.

FUNCTIONAL DESCRIPTION (Continued)

CONVERSION TIMES

The following table lists the conversion times which include the sample and hold acquisition time. For a CALRD and CALWR no A/D conversion actually takes place.

OPERATION	# OF INTERNAL CLOCKS*
8 bit A/D	80
13 bit A/D	110
CALWR	52
CALRD	80

SAMPLE AND HOLD TIMING

Figure 8 shows the internal timing for the sample and hold circuitry. The relationship between the "Start of Conversion" and the input channel going into sample mode is fixed at 6 internal clocks*, regardless of the Start Mode. Six internal clocks after the Start of Conversion the Sample and Hold is switched into the sample mode, placing two 9pF capacitors in parallel with the inputs pins; one on V_{IN+} and one on V_{IN-} . The sample switch is kept in the sample mode for 8 internal clocks (2.3 μ s at a 7MHz external clock), then placed in the hold mode. During the next 2 internal clocks the charge on the sample and hold is transferred into the A/D, after which the V_{REF} pin is sampled for 8 internal clocks.

Figure 8 also illustrates the timing of the SYNC pin in Master Mode during a conversion (M/S=1 Control High Byte register) and Start Mode 0, 1, or 2. SYNC is activated one internal clock cycle after the Start of Conversion and lasts for four internal clocks.

*For a description of internal clocks see Clock section.

ANALOG INPUTS

DIFFERENTIAL INPUTS AND COMMON MODE REJECTION

The differential inputs of the ML2230 eliminate the effects of common mode input noise (60Hz for example), as V_{IN+} and V_{IN-} are sampled at the same time.

NOISE

The leads to the analog inputs should be kept as short as possible to minimize output noise. Noise as well as digital clocks can couple into the inputs and cause errors. Input filters can be used to reduce the effects of these sources.

POWER SUPPLY DECOUPLING

Low inductance tantalum capacitors of 1 μ F or greater and 0.01 μ F disc ceramic capacitors are recommended for bypassing AV_{CC} as well as V_{SS} to AGND. These capacitors should be placed close to the AV_{CC} and V_{SS} pins.

MICROPROCESSOR INTERFACE

There are four 8 bit directly addressable registers; two Data Buffer registers, and two Control registers. The data buffer registers provide the conversion results. The data registers are double buffered, allowing one result to be read while the next sample is being converted. The data registers also allow access to the algorithmic converter's calibration code. Normally the ML2230 is operated without ever accessing these registers. (Refer to Diagnostics for more information). The two Control registers provide complete control and status information. The four registers are addressed by pins A0 and A1.

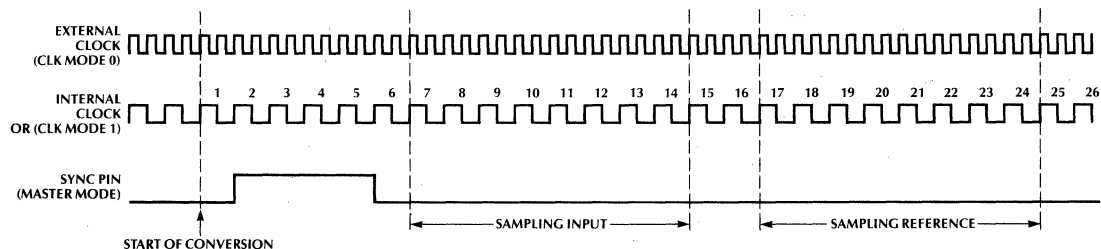


Figure 8. Sample and Hold Timing

FUNCTIONAL DESCRIPTION (Continued)

All data is returned from the converter in sixteen bit two's complement format, right hand justified, with the sign bit extended across the most significant bits.

Cycle	+ Max	- Min	Zero
13	0FFF	F000	0000
8	007F	FF80	0000

REGISTER DESCRIPTION

Register 0—Data Buffer Low Byte:

Register 0 contains the low byte result of the latest conversion when read. Depending on the Start Mode selected, reading or writing to this register may start the next conversion.

Register 1—Data Buffer High Byte:

Register 1 contains the high byte result of the latest conversion when read.

Register 2—CONTROL Register Low Byte:

Bit 0 (DAV status when READ/DAVACK acknowledge when a ONE is written):
Reading DAV = 1 indicates that new data is available or a calibration is complete. If both data bytes have been read, DAV will be cleared automatically. This bit can be explicitly acknowledged by writing a ONE to it; writing a zero has no effect. The DAV output pin always reflects the DAV status bit.

Bit 1 (BUSY status when READ/RESET when a ONE is written):
Reading BUSY = 1 indicates that a conversion or calibration is in progress. Writing a ONE will force a chip reset. Writing a zero has no effect.

RESET Default Conditions:

Both Control registers will automatically be cleared. Both Data Buffer registers will be unchanged. The Calibration register is not cleared after a reset, however the ADDRDY bit is cleared. Since the DAV status bit is cleared, the DAVB output is inactivated (high). The SYNC pin is forced to be an input as a result of clearing the M/S bit in the Control High Byte register.

Bit 2 (ADDRDY status when READ/DOCAL request when a ONE is written):

Reading ADDRDY = 0 indicates that the converter has not been calibrated since the last reset, and ADDRDY = 1 indicates that it has been calibrated since the last reset. Writing a ONE will force the converter to do a calibration; writing a zero has no effect.

Bit 3 (SC: Short cycle select):

Selects 8 or 13 bit conversions.

SC = 0: 13-bit conversion (default)
SC = 1: 8-bit conversion (short cycle)

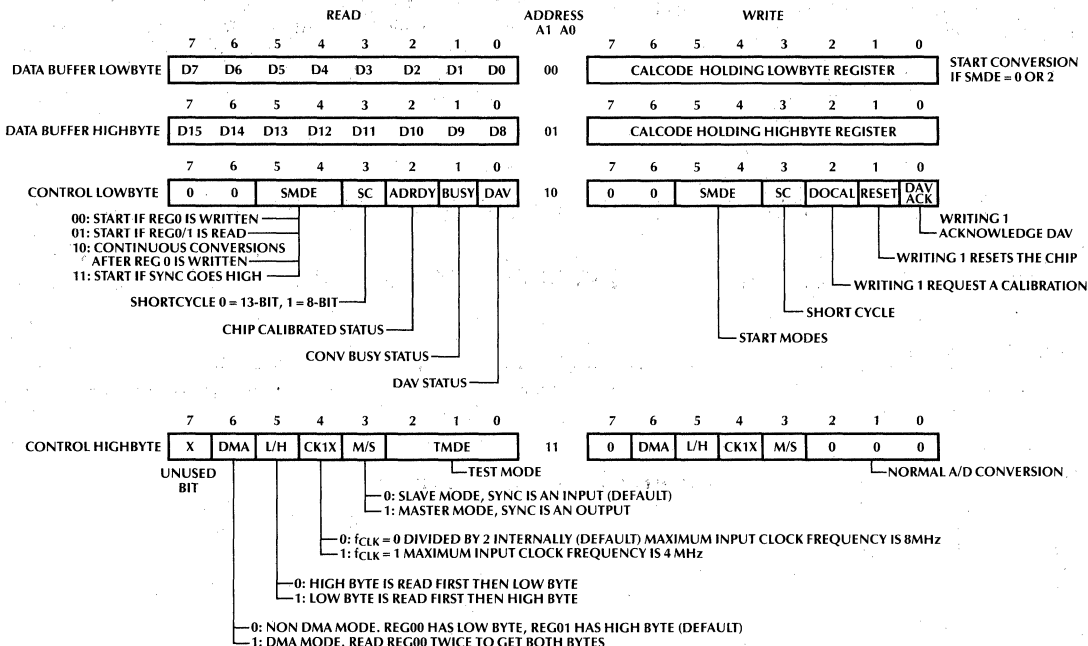


Figure 9. Register Description

FUNCTIONAL DESCRIPTION (Continued)

Note: For 8-bit conversions in non-DMA mode, only one byte needs to be read. This can be accomplished by setting L/H = 0, DMA = 0 and reading the Data Low Byte register. In DMA mode both bytes need to be read.

Bits 4,5 (SMDE: Start Mode):
Defines Start Conversion mode.

Bits 5,4

00	Start Conversion upon writing to register 0 (default)
01	Start Conversion upon reading register 0 if L/H = 0, or Start Conversion upon reading register 1 if L/H = 1. In DMA mode both bytes need to be read. The second byte read will Start Conversion.
10	Start Continuous Conversions upon writing to register 0.
11	Start on external SYNC input going high (Requires Slave mode: M/S = 0)

Bits 7,6 (reserved):

These bits are reserved by Micro Linear and must be written as zero.

Register 3 (Control Register High Byte):

Bits 2,1,0 (TMDE: test mode select bits)

These bits are used for diagnostic purposes only and normally not accessed during operation. The default value of TMDE is 000 which selects a normal A/D conversion. See Diagnostics for more information.

TMDE	Description
000	Normal A/D Conversion
001	Reserved by Micro Linear (Do Not Use)
010	CALWR Operation
011	CALRD Operation
100	System Offset
101	Common-mode
110	Plus Full Scale
111	Minus Full Scale

Bit 3 (M/S: Master/Slave bit):

Dictates whether the SYNC pin is an input or an output. Upon RESET, this bit is cleared.

M/S = 0: Slave Mode	SYNC is an input which is used to trigger a conversion if SMDE = 11.
M/S = 1: Master Mode	SYNC is an output. At the beginning of every conversion, SYNC is high for 4 internal clocks.

Bit 4 (CK1X: clock select bit):

Selects whether the external clock will be divided by two or used directly as the internal clock. See Clock section for a detailed explanation.

CK1X = 0:	the external clock is divided by two and used as the internal clock. This is referred to as CLK Mode = 0.
-----------	---

CK1X = 1:	the external clock input is used directly as the internal clock. This is referred to as CLK Mode = 1.
-----------	---

Bit 5 (L/H: Low Byte/High Byte):

In non-DMA mode the L/H bit defines whether DAVB is deactivated by reading the Data Low Byte or Data High Byte. In DMA mode, the L/H bit defines the order in which the Low/High Data Bytes are presented to the data bus. DMA mode automatically deactivates DAVB after both bytes are read.

*non-DMA mode: DMA = 0

L/H = 0:	reading register 0 (Low Byte) will de-assert DAVB
L/H = 1:	reading register 1 (High Byte) will de-assert DAVB

*DMA mode: DMA = 1

L/H = 0:	the first read is the Data High Byte, and the second read is the Data Low Byte, then DAVB output is de-asserted
----------	---

L/H = 1:	the first read is the Data Low Byte, and the second read is the Data High Byte, then DAVB output is de-asserted
----------	---

Bit 6 (DMA: DMA mode bit):

This bit allows both high and low bytes from the 13 bit conversion to be read from one address; either Data Buffer Low Byte or Data Buffer High Byte registers.

DMA = 0:	The high byte of the conversion will always be read from the Data Buffer High Byte register and the low byte of the conversion will always be read from the Data Buffer Low Byte Register.
----------	--

DMA = 1:	Both high and low bytes of the conversion can be read from either the Data Buffer High or Low Byte Registers. A DMA controller, microprocessor, or other I/O device can use a single I/O address to read both the low and high bytes of the conversion. The order in which the high and low data bytes are presented is defined by the L/H control bit.
----------	---

Note: This feature is not restricted to DMA controllers. It is an I/O option which may be used by a DMA controller, microprocessor, or any other type of I/O device.

Bit 7 (Reserved by Micro Linear)

This bit is not used. When written use zero.

FUNCTIONAL DESCRIPTION (Continued)

GENERAL OPERATING INFORMATION CONVERSION-START PROTOCOL

There are four different ways to start a conversion. They are defined by SMDE bits 4 and 5 in the Control Low Byte Register.

SMDE Bits 5,4

- 00: A write to register 0 will start a conversion. During a conversion, if another write is issued to register 0, the "Start Conversion" command will be latched and another conversion will immediately follow the current one. To insure that the second write will be latched, it must occur at least 3 internal clocks after the first write. Only one additional write will be latched; multiple writes within a conversion will only yield one more conversion.
- 01: Reading the data from the previous conversion starts the next conversion. Start Conversion upon reading register 0 if L/H = 0, or Start Conversion upon reading register 1 if L/H = 1. In DMA mode both bytes need to be read. The second byte read will start the Conversion.
- 10: This mode causes continuous conversions; the next conversion begins immediately after the previous conversion ends. Writing to register 0 will start the first conversion; thereafter the converter runs continuously. This mode yields the maximum conversion rate.
- 11: The Sync input triggers the start of a conversion. The M/S bit in the Control High Byte Register must be cleared, placing the chip in the slave mode.

Note: The external activation signals for Start Modes 0, 1, and 3 are synchronized internally to the system clock. If periodic sampling is required using these Start Modes, the SYNC, RD, or WR pulses must be synchronized to the system clock. Start Mode 2 guarantees periodic sampling.

DOUBLE-BUFFERED DATA REGISTER

The A/D conversion result is double-buffered using the Data Buffer registers and the A/D Data register. The actual End-Of-Conversion (EOC) does not correspond with the DAVB output going low. The DAVB output goes low 16 internal clocks after the EOC. From the time DAVB output goes LOW, the user has one full conversion time (80 or 110 internal clocks) minus 16 internal clocks to read two data bytes as shown in Figure 10.

SELF CALIBRATION

Setting the DOCAL bit issues a calibration request to the chip. When calibration is done, the DAV status bit is set and the DAVB output goes low.

A calibration requires 8,260 internal clocks. Using a 7MHz clock (CLK Mode = 0), this is approximately 2 ms. Power supplies and external voltage reference must be stable before issuing a request for calibration.

The ML2230 should be calibrated before any conversions are attempted. Calibrations must not be performed simultaneously with conversions. Before requesting a calibration, the user may want to read the Busy status bit to make sure that the converter is idle. Polling the chip while the calibration is in progress is not recommended.

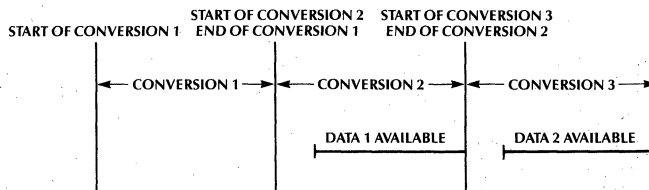


Figure 10

FUNCTIONAL DESCRIPTION (Continued)

CLOCK

The ML2230 has the option of dividing the clock at the CLK pin by 2, or using it directly to drive the internal logic. This option is selected through the CK1X bit in the Control register. When CK1X = 0 the clock is divided by 2. This is referred to as CLK Mode = 0. The clock at the CLK pin is referred to as the External clock, and the Internal Clock is the External clock divided by 2. When CK1X = 1, the clock at the CLK pin drives the internal logic directly, therefore this clock is referred to as the Internal clock. This is also known as CLK Mode = 1. All internally clocked logic is positive edge triggered.

CLK Mode = 0:

There are two advantages to CLK Mode 0. This is the only Mode that allows an external crystal to be used. CLK Mode 1 cannot operate with an external crystal, the CLK pin must be driven. The second advantage of CLK Mode 0 is that the duty cycle for a driven clock is less stringent than in CLK Mode 1. (Refer to t_{CLK0} and t_{CLK1} in AC Electrical Characteristics for CLK Mode 0 and 1 timing requirements, respectively.)

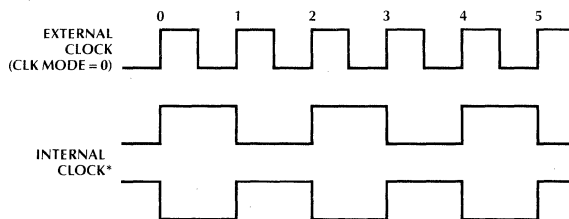
On power up the state of the divide by two flip-flop is indeterminate. Therefore the relationship between the internal clock and the external clock at the CLK pin can have one of two possibilities as shown in Figure 11. As a result the following should be considered.

t_{WRCK} , t_{RDCK} , and t_{SYNCK} specs, \overline{RD} , \overline{WR} , and SYNC setup times to Start of Conversion), will be as shown in the data sheet, or the data sheet specs plus one external clock period. Since these specifications are with respect to the rising edge of the external clock, it is not known whether this rising edge corresponds to the rising edge or falling edge of the internal clock. Therefore there is an uncertainty of one external clock period.

If periodic sampling is necessary and Start Mode 0, 1, or 3 is used, the external start pulse (either \overline{RD} , \overline{WR} , or SYNC) must be synchronous to the external clock, meet the setup time, and be an even number of external clock periods. If the start pulse were an odd number of external clock periods, half the pulses would correspond with the rising edge of the internal clock, and the other half would correspond with the falling edge of the internal clock. Therefore the sampling period would change by one external clock period every sample. Start Mode 2 guarantees periodic sampling regardless of the CLK mode.

CLK Mode = 1:

This mode eliminates the requirement that external start pulses must be an even number of external clock periods. However periodic sampling still requires that the start pulse be synchronous to the external clock, and the setup time must be met. CLK Mode 1 also eliminates the uncertainty of the t_{WRCK} , t_{RDCK} , and t_{SYNCK} requirements.



*INTERNAL CLOCK MAY BE ONE OF THE TWO ABOVE IN CLK MODE = 0

Figure 11

FUNCTIONAL DESCRIPTION (Continued)

DIAGNOSTICS

Diagnostic routines may be run after power up or any other time to ensure proper operation. The diagnostic features, which are software selectable, don't require external hardware. Both the analog and digital sections can be tested.

The ML2230 is placed in the diagnostic mode via the TMDE field in the Control High Byte Register. Once the ML2230 is placed in one of the diagnostic modes, a conversion must be executed before the results can be read. As with all conversions, DAVB will be activated upon completion.

ANALOG CONVERSION DIAGNOSTICS

TMDE = 000: Normal Operation

Selects normal A/D conversion. Default condition after a software reset.

TMDE = 001: Reserved by Micro Linear.

TMDE = 010: CALWR operation

The data in Write register 0 and 1 (CALCODE Holding Register), are transferred into the converter's Calibration register when a "Start Conversion" is issued. A dummy conversion occurs and the DAVB output goes LOW to indicate that the operation is complete.

TMDE = 011: CALRD operation

The contents of the Calibration register are transferred through the A/D Data register and loaded into the Data Buffer registers. A dummy 8-bit conversion occurs and DAVB output goes LOW to indicate that the CALRD operation is complete.

TMDE = 100: System Offset

The positive and negative inputs to the Sample and Hold are tied to analog ground. With this setting, converted data will give the offset of the A/D converter and Sample/Hold combination. The V_{IN+} and V_{IN-} pins will remain in a high impedance state while in this mode.

TMDE = 101: Common-mode

Both the positive and negative inputs of the Sample and Hold are tied to V_{REF} . The results of a conversion in this test mode indicates how well the converter is rejecting a common mode signal.

TMDE = 110: Positive Full Scale

This test mode connects the positive input of the Sample and Hold to V_{REF} and the negative input of the Sample and Hold to analog ground. The result of converting in this test mode is a value near positive full scale.

TMDE = 111: Negative Full Scale

This test mode connects the positive input of the Sample and Hold to analog ground and the negative input to V_{REF} . The result of converting in this test mode is a value near negative full scale.

DIGITAL LOOPBACK

The ML2230's architecture provides a way for the micro-processor to indirectly read and write to the A/D converter's calibration register and data register via a CALRD and CALWR. Figure 12 illustrates this architecture. This in effect allows a digital loopback.

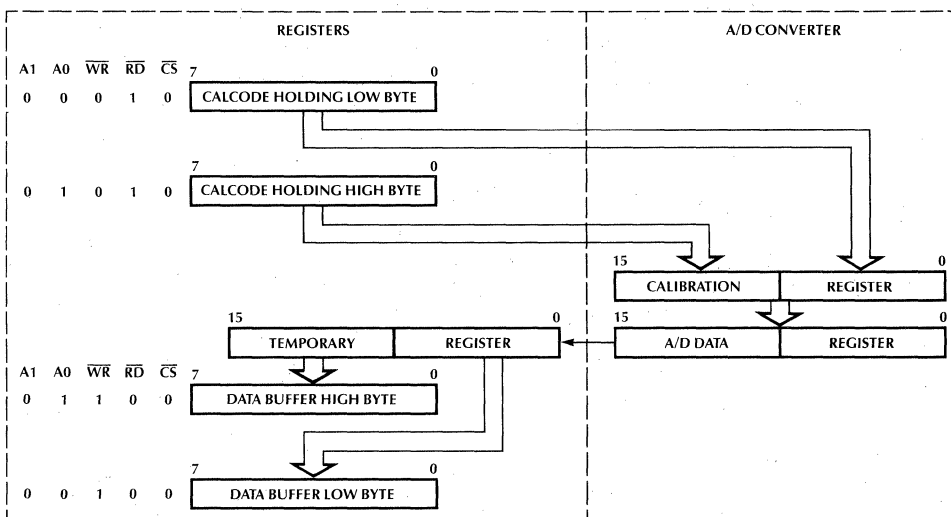


Figure 12. Digital Loopback

FUNCTIONAL DESCRIPTION (Continued)

When the TMDE bits are set to 010 CAL WRITE (CALWR), and a Start Conversion is issued in any one of the four modes, the contents of the CALCODE Holding Low Byte and High Byte registers are transferred into the A/D converter's Calibration register. When the TMDE bits are set to 011 CAL READ (CALRD), and a Start Conversion is issued, the contents of the Calibration register are transferred through the A/D's Data register into the Data Buffer Low Byte and Data Buffer High Byte registers. The result of these two operations is a complete loopback from the CALCODE Holding registers through the A/D converter and back into the Data buffer registers. This loopback provides user assurance that all the paths are clear and there are no stuck bits.

Note: When a CALWR is done, the previous calibration value is lost. The correct calibration value must be restored before the converter is used to convert data.

CALIBRATION PASS/FAIL TEST

The CALRD can be used as a way to verify a successful calibration. After a calibration is completed, the CALRD may be issued in order to read the contents of the Calibration register. If the Low Byte of the data buffer register is all ones after executing a CALRD, the calibration failed; otherwise the calibration is successful.

2

FUNCTIONAL DESCRIPTION (Continued)

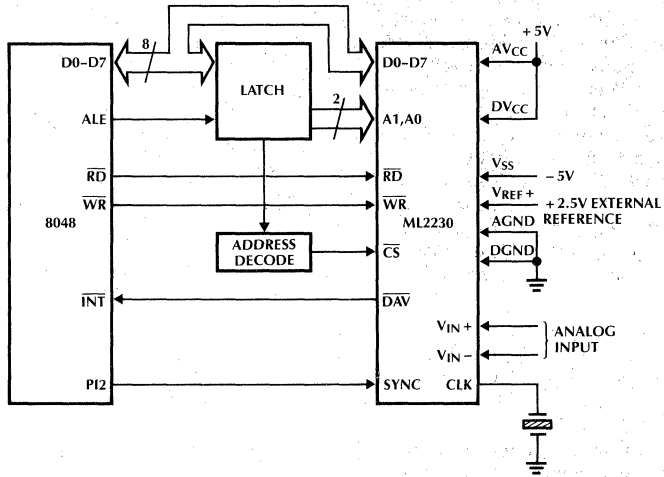


Figure 13. Interfacing to 8048 Microcontroller

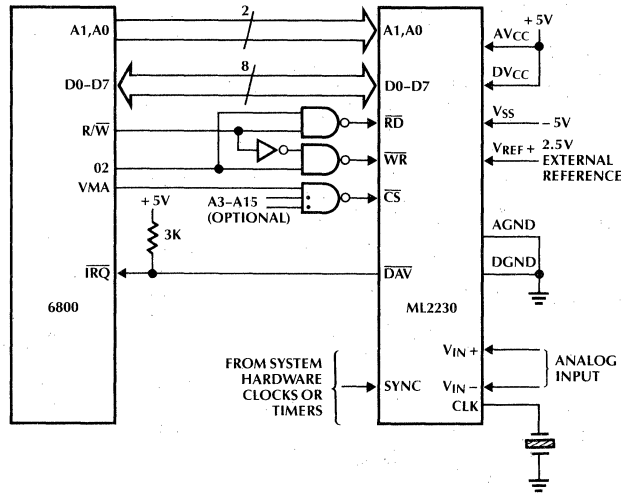


Figure 14. Interfacing to 6800 Microprocessor

ORDERING INFORMATION

PART NUMBER	LINEARITY ERROR	MINIMUM CONVERSION	TEMPERATURE RANGE	PACKAGE
ML2230BCJ	$\pm\frac{3}{4}$ LSB	31.5 μ s	0°C to +70°C	Hermetic DIP (J24)
ML2230CCJ	± 1 LSB	31.5 μ s	0°C to +70°C	Hermetic DIP (J24)
ML2230DCJ	± 1 LSB	44.0 μ s	0°C to +70°C	Hermetic DIP (J24)

μP Compatible 12-Bit Plus Sign A/D Converter with Sample and Hold

GENERAL DESCRIPTION

The ML2233 is a member of Micro Linear's 12-bit plus sign CMOS A/D converter family utilizing a self calibrating algorithmic technique. The sample-and-hold, incorporated on the ML2233, has a differential input for noise immunity and power supply rejection. All errors of the sample-and-hold are accounted for in the analog-to-digital converter's accuracy specification.

The ML2233B has a maximum non-linearity error over temperature of 0.018% of full-scale, and the ML2233C and ML2233D have a maximum non-linearity error over temperature of 0.024% of full scale.

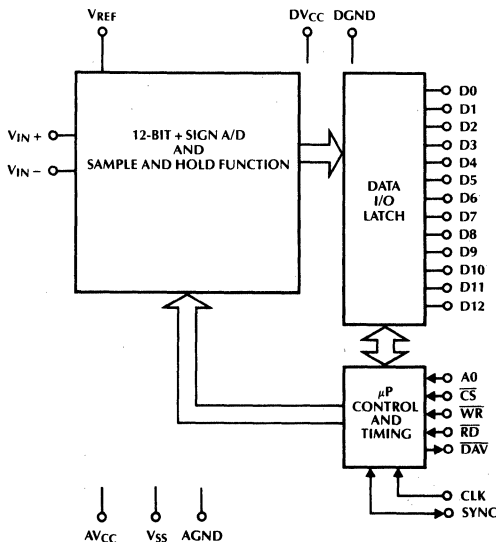
Designed to interface to a 16-bit microprocessor bus without additional components, the ML2233 outputs the 13-bit data result in one word. Data format is 2's complement. All digital signals are fully TTL and CMOS compatible.

For interfacing to an 8-bit microprocessor bus the ML2230 provides a 13-bit data result in two 8-bit bytes.

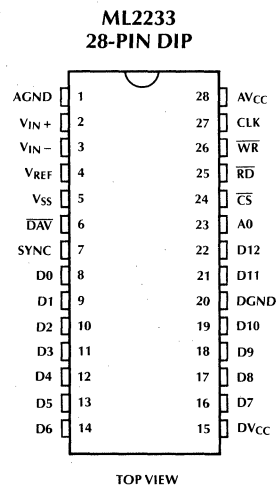
FEATURES

- Resolution 12-bits + sign
- Conversion time (including S/H acquisition) 31.5μs max
- Sample and hold acquisition 2.3μs max
- Non-linearity error ± $\frac{3}{4}$ LSB and ±1LSB max
- Low harmonic distortion 0.01%
- No missing codes
- Self calibrating—maintains accuracy over time and temperature
- Inputs withstand |7V| beyond supplies
- Data transfer options—interrupt, DMA, or polling
- 13-bit result for 16-bit bus interface
- Standard 28-pin DIP

BLOCK DIAGRAM



PIN CONNECTIONS



PIN DESCRIPTION

PIN NO.	NAME	FUNCTION	PIN NO.	NAME	FUNCTION
1	AGND	Analog ground.	16	D7	Bidirectional data bit.
2	V _{IN+}	Positive differential analog input; range = $V_{SS} \leq V_{IN+} \leq AV_{CC}$, $ (V_{IN+}) - (V_{IN-}) \leq V_{REF}$.	17	D8	Bidirectional data bit.
3	V _{IN-}	Negative differential analog input; range = $V_{SS} \leq V_{IN-} \leq AV_{CC}$, $ (V_{IN+}) - (V_{IN-}) \leq V_{REF}$.	18	D9	Bidirectional data bit.
4	V _{REF}	Voltage reference input; referenced to analog ground.	19	D10	Bidirectional data bit.
5	V _{SS}	Negative power supply; decouple to AGND.	20	DGND	Digital ground.
6	\overline{DAV}	Data available; indicates a conversion has completed and data is available or calibration completed.	21	D11	Bidirectional data bit.
7	SYNC	In the slave mode, SYNC is a positive edge triggered input used to start a conversion. In master mode, SYNC is an output and indicates conversion start.	22	D12	Bidirectional data bit.
8	D0	Bidirectional data bit.	23	A0	Address for the microprocessor interface to access registers.
9	D1	Bidirectional data bit.	24	\overline{CS}	Chip select; enables writing to or reading from.
10	D2	Bidirectional data bit.	25	\overline{RD}	Read; enables ML2233 to drive data bus.
11	D3	Bidirectional data bit.	26	\overline{WR}	Write; allows writing into the registers.
12	D4	Bidirectional data bit.	27	CLK	Clock input. Driven with an external clock or crystal referenced to DGND. The crystal must be parallel resonant with minimum capacitive loading. (i.e., no bypass caps should be used and leads should be kept short.)
13	D5	Bidirectional data bit.	28	AV _{CC}	Positive analog power supply. Decouple to AGND. Tie to DV _{CC} from same power supply.
14	D6	Bidirectional data bit.			
15	DV _{CC}	Digital power supply.			

ABSOLUTE MAXIMUM RATINGS
(Note 1)

Supply Voltages (AV _{CC} and DV _{CC})	6.0V
Negative Supply Voltage (V _{SS})	-6.0V
Voltage at Analog Inputs	V _{SS} -7V to AV _{CC} +7V
Voltage at V _{REF}	V _{SS} -7V to AV _{CC} +7V
Input Current per Digital Pin	±10mA
Input Current at Analog Inputs	±20mA
Storage Temperature Range	-65°C to +150°C
Package Dissipation @ 25°C	875mW
Lead Temperature (soldering, 10 seconds)	
Dual-In-Line Package (Ceramic)	300°C

OPERATING CONDITIONS (Note 2)

Temperature Range	0°C to 70°C
Supply Voltage (AV _{CC} and DV _{CC})	4.5V _{DC} to 6.0V _{DC}
Negative Supply Voltage (V _{SS})	-4.5V _{DC} to -6.0V _{DC}
Reference Voltage (V _{REF})	2.60V

ML2233

ELECTRICAL CHARACTERISTICS

The following specifications apply for $AV_{CC} = DV_{CC} = 5V \pm 5\%$, $V_{SS} = -5V \pm 5\%$, $V_{REF} = 2.500V$, $V_{IN-} = AGND$, $V_{IN+} = -2.5V$ to $+2.5V$, $T_A = T_{MIN}$ to T_{MAX} unless otherwise specified.

PARAMETER	NOTES	CONDITIONS	MIN	TYP (Note 3)	MAX	UNITS
Converter Characteristics						
Linearity Error ML2233BCJ ML2233CCJ ML2233DCJ	4	$f_{CCLK} = 0.1 \leq 7MHz$ $f_{CCLK} = 0.1 \leq 7MHz$ $f_{CCLK} = 0.1 \leq 5MHz$			$\pm\frac{3}{4}$ ± 1 ± 1	LSB LSB LSB
Unadjusted Zero Error ML2233BCJ ML2233CCJ ML2233DCJ	4				$\pm\frac{3}{4}$ ± 2 ± 2	LSB LSB LSB
Unadjusted Positive and Negative Full Scale Error	5				± 4	LSB
Zero Error Temperature Coefficient				0.5		ppm/°C
Gain Temperature Coefficient				10		ppm/°C
Common-Mode Rejection	5, 6		80			dB
Analog Input Source Resistance	5				2	k Ω
Analog Input Range	4	V_{IN+} Referred to V_{IN-}	$-V_{REF}$		$+V_{REF}$	V
Analog Input Leakage Current	4				100	nA
Voltage Reference Input Source Impedance	5				0.5	k Ω
Reference Input Leakage Current	4				100	nA
Digital and DC Characteristics						
Power Supply Current I_{CC} , Analog V_{CC} I_{CC} , Digital V_{CC} I_{SS} , V_{SS}	4			30 10 18	50 30	mA μ A mA
Power Supply Rejection AV_{CC} V_{SS}	7	DC DC to 25kHz DC DC to 25kHz		80 50 80 50		dB dB dB dB
V_{ILCLK} , Clock Input Low Voltage	4				0.8	V
V_{IHCLK} , Clock Input High Voltage	4		3.5		AV_{CC}	V
I_{LI} , Input Leakage Current (CLK)	4	$AGND \leq V_{IN} \leq AV_{CC}$			± 200	μ A
V_{IL} , Input Low Voltage	4				0.8	V
V_{IH} , Input High Voltage	4		2.0		DV_{CC}	V
V_{OL} , Output Low Voltage	4	$I_{OL} = 2.0mA$			0.45	V
V_{OH} , Output High Voltage	4	$I_{OH} = -400\mu A$	2.4			V
I_L , Input Leakage Current (except CLK)	4	$AGND \leq V_{IN} \leq AV_{CC}$			± 10	μ A
I_{HI-Z} , Output Leakage Current (D0–D12)	4	$\overline{RD} = \overline{CS} = V_{IH}$			± 10	μ A
C_I , Input Capacitance (all digital inputs)				10		pF
C_O , Output Capacitance (outputs D0 to D12, SYNC and \overline{DAV})				10		pF

ELECTRICAL CHARACTERISTICS (Continued)

SYMBOL	PARAMETER	CONDITIONS		MIN	TYP (Note 3)	MAX	UNITS
AC Electrical Characteristics (Note 8)							
t_C	Conversion Time	4, 9	CLK Mode = 0	$f_{CLK} = 70\text{MHz}$ $f_{CLK} = 5.0\text{MHz}$	31.5 44.0		μs μs
	Sample and Hold Acquisition	4, 9	CLK Mode = 0	$f_{CLK} = 70\text{MHz}$ $f_{CLK} = 5.0\text{MHz}$		2.3 3.2	μs μs
f_{CLK0}	Clock Frequency	5, 9	CLK Mode = 0	Crystal-Driven	3 1	7 7	MHz MHz
t_{CLK0}	Clock Width	5, 9	Driven (CLK Mode = 0)	High Low	50 50		ns ns
f_{CLK1}	Clock Frequency	5, 10	Driven (CLK Mode = 1)		0.5	(Note 11)	MHz
t_{CLK1}	Clock Width	5, 10	Driven (CLK Mode = 1)	High Low	125 125		ns ns
t_{AD}	Address Stable to Valid Data	4			150		ns
t_{AR}	Address Stable Before Read	4			0		ns
t_{RA}	Address Hold After Read	4			0		ns
t_{RR}	Read Pulse Width	4			150		ns
t_{RD}	Read Access	4				150	ns
t_{1Z}, t_{0Z}	Data Read to Hi-Z	4			0	50	ns
t_{RV}	Recovery Between Two Reads or Writes	5			250		ns
t_{RDCK}	Read to Clock Setup Time	5, 12			40		ns
t_{AW}	Address Stable Before Write	4			0		ns
t_{WA}	Address Hold After Write	4			0		ns
t_{WW}	Write Pulse Width	4			150		ns
t_{DW}	Data Setup Before Write Trailing Edge	4			100		ns
t_{WD}	Data Hold After Write Trailing Edge	4			0		ns
t_{WRCK}	Write to Clock Setup Time	5, 12			40		ns
t_{CKDAV}	Clock to DAV Assert	4, 13	$C_L = 50\text{pF}$			120	220 ns
t_{SYNCK}	SYNC Input to Clock Setup	5, 12			40		ns
t_{SYNCN}	SYNC Input Width	5	(CLK Mode = 0) (CLK Mode = 1)		6 3		$1/f_{CLK0}$ $1/f_{CLK1}$
t_{CKSYNC}	External Clock to SYNC Output Delay	5, 13	$C_L = 50\text{pF}$			150	200 ns
t_{SYNCO}	SYNC Output Pulse Width	5, 13	(CLK Mode = 0) (CLK Mode = 1)			8 4	$1/f_{CLK0}$ $1/f_{CLK1}$
t_{WRDAV}	Write Reg2 to $\overline{\text{DAV}}$ Rising Edge	4, 14	$C_L = 50\text{pF}$			170	ns
t_{RDDAV}	Read Reg0 to $\overline{\text{DAV}}$ Rising Edge	4, 15	$C_L = 50\text{pF}$			170	ns
t_r, t_f	Rise and Fall		All Inputs			25	ns

Note 1: Absolute maximum ratings are limits beyond which the life of the integrated circuit may be impaired. All voltages unless otherwise specified are measured with respect to ground.

Note 2: Devices are 100% tested with temperature limits guaranteed by 100% testing, sampling or by correlation with worst-case test conditions.

Note 3: Typicals are parametric norm at 25°C.

Note 4: Parameter guaranteed and 100% production tested.

Note 5: Parameter guaranteed. Parameters not 100% tested are not in outgoing quality level calculation.

Note 6: Common mode rejection is the ratio of the change in zero error to the change in common mode input range.

Note 7: Power supply rejection is the ratio of the change in zero error to the change in power supply voltage.

Note 8: All parameters measured from 0.8V to 2.0V, $C_L = 100\text{pF}$.

Note 9: CK1X bit in control register = 0.

Note 10: CK1X bit in control register = 1.

Note 11: Maximum frequency is $1/t_{CLK1}(\text{high}) + t_{CLK1}(\text{low}) + \text{rise} + \text{fall times}$ and $\leq 3.5\text{MHz}$.

Note 12: Setup time required for synchronous start of conversion.

Note 13: In CLK mode = 0 (CK1X bit in control register = 0) start of conversion will occur at specified time; or time plus one f_{CLK0} period (see Figure 5).

Note 14: Writing a control register bit 0 with a one will acknowledge the $\overline{\text{DAV}}$ condition and de-assert $\overline{\text{DAV}}$ output.

Note 15: In start mode = 1, a read from location "0" will start the next conversion and de-assert the $\overline{\text{DAV}}$ output.

TIMING DIAGRAMS

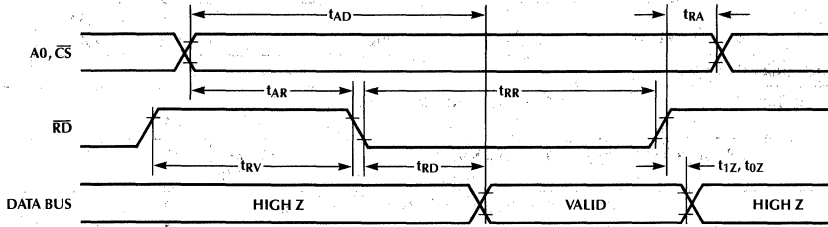


Figure 1. Read Cycle

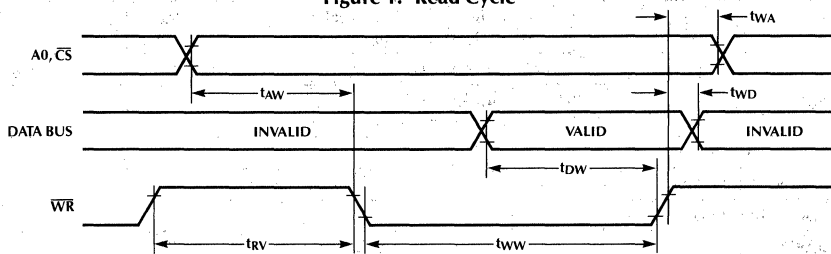
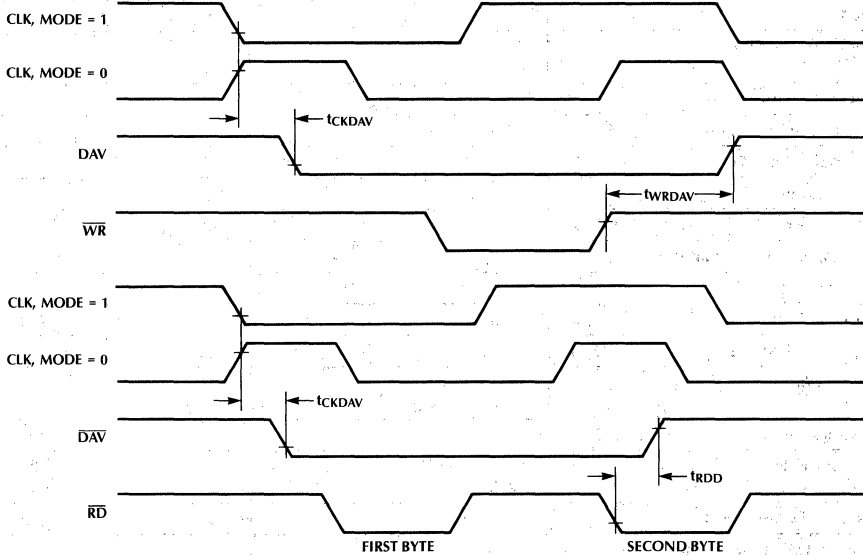


Figure 2. Write Cycle



DAV IS SET AND CLEARED BY INTERNAL CIRCUITRY.
NOTE: DMA BIT IN THE CONTROL REGISTER MUST BE SET FOR THIS OPERATION.

Figure 3. Data Available

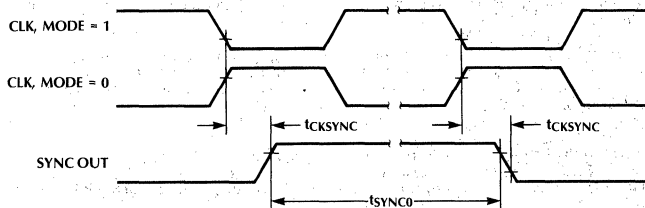
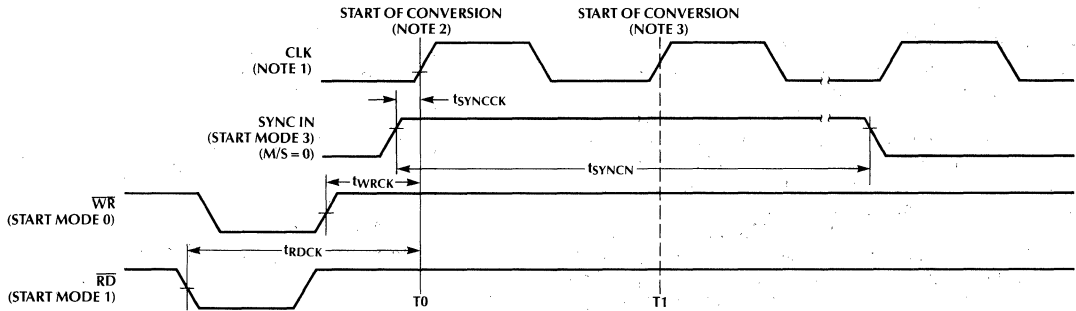


Figure 4. SYNC Output

TIMING DIAGRAMS (Continued)



- NOTES:
 1. CLK IS THE CLOCK DRIVEN AT THE CLOCK PIN.
 2. IN CLK MODE 1, WILL ALWAYS OCCUR AT T0 IF SETUP TIMES ARE MET.
 3. IN CLK MODE 0, WILL OCCUR EITHER AT T0 OR T1 IF SETUP TIMES ARE MET.

Figure 5. Synchronous Start of Conversion (Start Mode 0,1,3)

BLOCK DIAGRAM

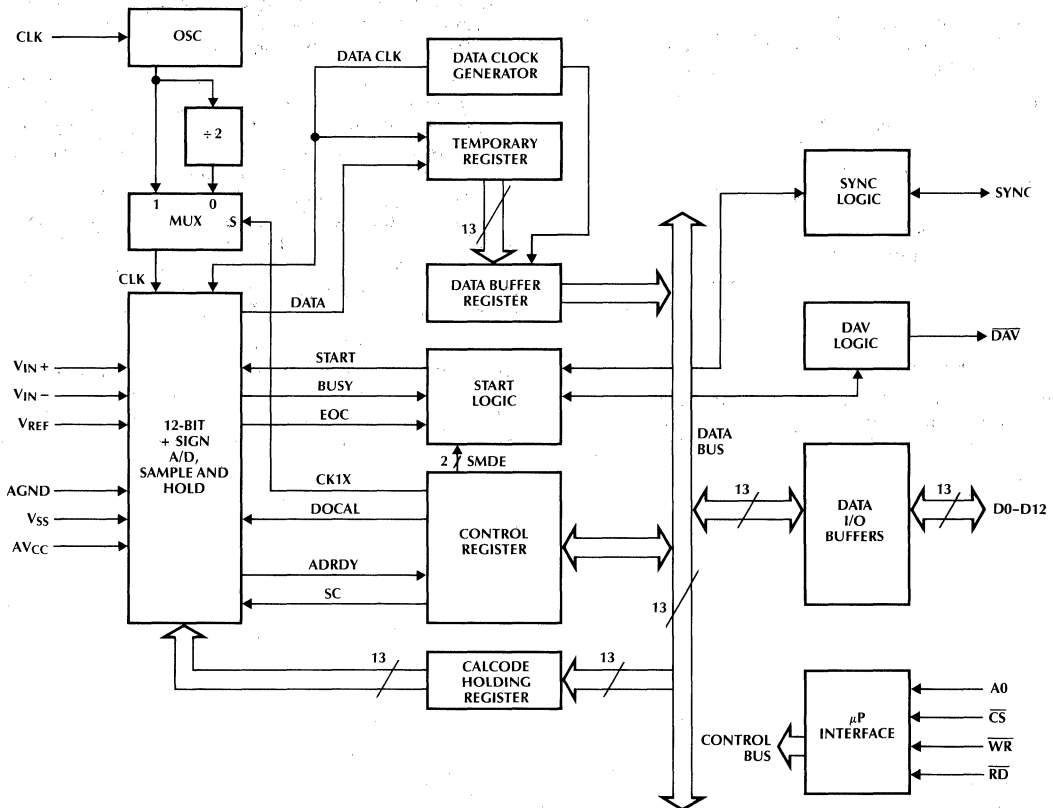


Figure 6. Block Schematic Diagram

FUNCTIONAL DESCRIPTION

ALGORITHMIC A/D CONVERTER

Micro Linear's algorithmic converter uses a successive approximation technique. Most of today's successive approximation converters use a DAC to feedback the approximated signal, however this technique requires more circuitry than algorithmic converters. In addition the values of all of the resistors or capacitors in the DAC must be matched to within the accuracy of the converter. This is difficult to do in silicon beyond 10 bits unless trimming is used. An algorithmic converter uses less circuitry and is more easily trimmed. Micro Linear's algorithmic converter is implemented using a 2x amplifier, a sample/hold amp, and a comparator as shown in Figure 7.

The input sample is first multiplied by two then compared to the reference voltage. If the 2x input voltage is greater than the reference, the MSB is a 1 and the reference voltage is subtracted from the 2x input voltage. The remainder is stored in the sample and hold. If the 2x input voltage is less than the reference, the MSB is a 0 and the 2x input voltage is stored in the sample and hold. This process repeats again, however now the sample and hold voltage is multiplied by 2.

The algorithm involves multiplication by 2, comparison, and possibly subtraction. Referring to Figure 6, the algorithm for the circuit can be described as follows:

Step 1 If $(2 \times V_{IN}) - V_{REF} \geq 0$
 then MSB = 1
 $(2 \times V_{IN}) - V_{REF} \rightarrow S/H$
 else MSB = 0
 $(2 \times V_{IN}) \rightarrow S/H$

Step 2 If $(2 \times S/H) - V_{REF} \geq 0$
 then next bit = 1
 $(2 \times S/H) - V_{REF} \rightarrow S/H$
 else next bit = 0
 $(2 \times S/H) \rightarrow S/H$

Step 3 Repeat Step 2 until conversion complete.

Since the A/D converter handles bipolar inputs, negative inputs are handled slightly differently using the same principle.

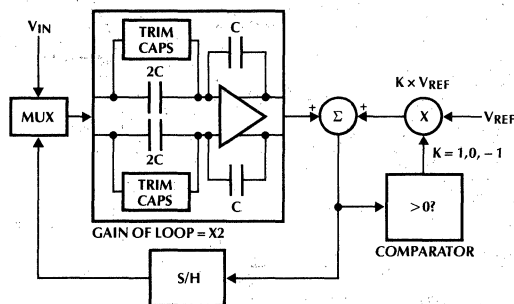


Figure 7. Self-Calibrating A/D Converter

SELF CALIBRATION

In order to maintain integral and differential linearity to the 1/2 LSB level in an algorithmic converter, two critical parameters need to be controlled, loop offsets and the gain of the loop. Loop offsets are automatically nulled before each conversion using auto-zeroing circuitry on both the sampling amp and the 2x amp. The gain of the loop is adjusted using self calibration.

Self-calibrating the algorithmic converter, once the offsets have been nulled, is performed by measuring the 2x gain of the loop and adjusting it. The gain can be measured by converting the reference voltage as the input as well as the reference (V_{REF}/V_{REF}), and examining the output code. Converting V_{REF} should yield plus full scale, since V_{REF}/V_{REF} should equal 1. If the gain of the loop is slightly less than 2, the resulting LSB of the conversion will be "0". If the magnitude bits of the resulting conversion are all "1s", the gain may be too great, therefore the gain is reduced to the point where the threshold of the LSB is reached.

Adjustment of the 2x gain is done with the binary weighted trim capacitor arrays connected to each of the 2C input capacitors. A small value of capacitance is either added to or subtracted from the 2C input caps until the gain of the loop is within 13 bit accuracy of 2.

FUNCTIONAL DESCRIPTION (Continued)

CONVERSION TIMES

The following table lists the conversion times which include the sample and hold acquisition time. For a CALRD and CALWR no A/D conversion actually takes place.

OPERATION	# OF INTERNAL CLOCKS*
8 bit A/D	80
13 bit A/D	110
CALWR	52
CALRD	80

SAMPLE AND HOLD TIMING

Figure 8 shows the internal timing for the sample and hold circuitry. The relationship between the "Start of Conversion" and the input channel going into sample mode is fixed at 6 internal clocks*, regardless of the Start Mode. Six internal clocks after the Start of Conversion the Sample and Hold is switched into the sample mode, placing two 9pF capacitors in parallel with the inputs pins; one on V_{IN+} and one on V_{IN-} . The sample switch is kept in the sample mode for 8 internal clocks (2.3µs at a 7MHz external clock), then placed in the hold mode. During the next 2 internal clocks the charge on the sample and hold is transferred into the A/D, after which the V_{REF} pin is sampled for 8 internal clocks.

Figure 8 also illustrates the timing of the SYNC pin in Master Mode during a conversion (M/S = 1 Control register) and Start Mode 0, 1, or 2. SYNC is activated one internal clock cycle after the Start of Conversion and lasts for four internal clocks.

*For a description of internal clocks see Clock section.

ANALOG INPUTS

DIFFERENTIAL INPUTS AND COMMON MODE REJECTION

The differential inputs of the ML2233 eliminate the effects of common mode input noise (60Hz for example), as V_{IN+} and V_{IN-} are sampled at the same time.

NOISE

The leads to the analog inputs should be kept as short as possible to minimize output noise. Noise as well as digital clocks can couple into the inputs and cause errors. Input filters can be used to reduce the effects of these sources.

POWER SUPPLY DECOUPLING

Low inductance tantalum capacitors of 1µF or greater and 0.01µF disc ceramic capacitors are recommended for bypassing AV_{CC} as well as V_{SS} to AGND. These capacitors should be placed close to the AV_{CC} and V_{SS} pins.

MICROPROCESSOR INTERFACE

There are two 13-bit directly addressable registers; a Data Buffer register and a Control register. The data buffer register provides the conversion results. The data register is double buffered, allowing one result to be read while the next sample is being converted. The data register also allows access to the algorithmic converter's calibration code. Normally the ML2233 is operated without ever accessing these registers. (Refer to Diagnostics for more information). The Control register provides complete control and status information. The two registers are addressed by pin A0.

2

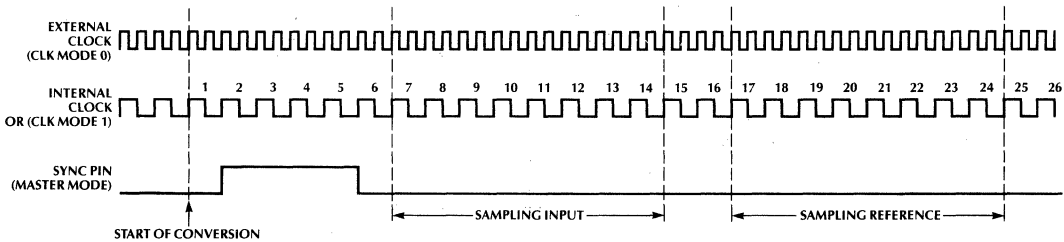


Figure 8. Sample and Hold Timing

FUNCTIONAL DESCRIPTION (Continued)

All data is returned from the converter in two's complement format.

Cycle	+Max	-Min	Zero
13	0FFF	1000	0000
8	007F	1F80	0000

REGISTER DESCRIPTION

Register 0—Data Buffer:

Register 0 contains the results of the latest conversion when read. Depending on the Start Mode selected, reading or writing to this register may start the next conversion.

Register 1—CONTROL Register:

Bit 0 (DAV status when READ/DAVACK acknowledged when a ONE is written):

Reading DAV = 1 indicates that new data is available or a calibration is complete. DAV will be cleared automatically when the data is read. This bit can be explicitly acknowledged by writing a ONE to it; writing a zero has no effect. The DAV output pin always reflects the DAV status bit.

Bit 1 (BUSY status when READ/RESET when a ONE is written):

Reading BUSY = 1 indicates that a conversion or calibration is in progress. Writing a ONE will force a chip reset. Writing a zero has no effect.

RESET Default Conditions:

The Control register will automatically be cleared. The Data Buffer register will be unchanged. The Calibration register is not cleared after a reset, however the ADRDY bit is cleared. Since the DAV status bit is cleared, the DAV output is inactivated (high). The SYNC pin is forced to be an input as a result of clearing the M/S bit in the Control register.

Bit 2 (ADRDY status when READ/DOCAL request when a ONE is written):

Reading ADRDY = 0 indicates that the converter has not been calibrated since the last reset, and ADRDY = 1 indicates that it has been calibrated since the last reset. Writing a ONE will force the converter to do a calibration; writing a zero has no effect.

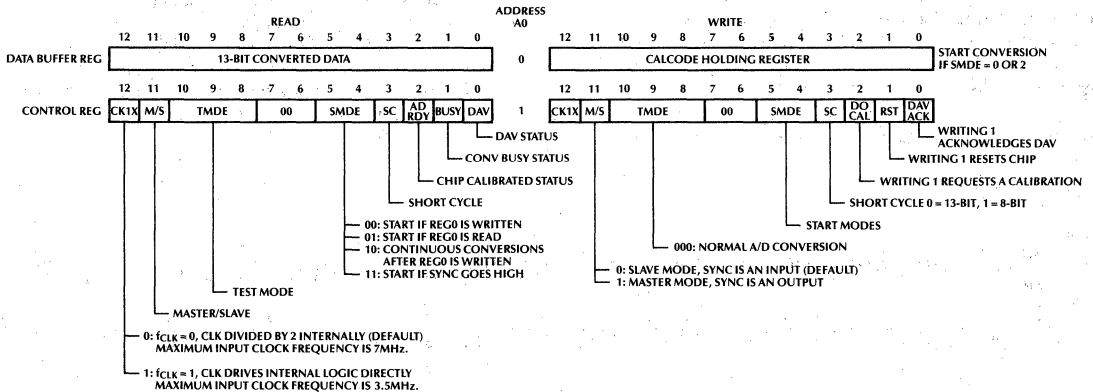


Figure 9. Register Description

FUNCTIONAL DESCRIPTION (Continued)

Bit 3 (SC: Short cycle select):

Selects 8 or 13 bit conversions.

SC = 0: 13-bit conversion (default)

SC = 1: 8-bit conversion (short cycle)

Bits 4,5 (SMDE: Start Mode):

Defines Start Conversion mode.

Bits 5,4

00	Start Conversion upon writing to register 0 (default)
01	Start Conversion upon reading register 0
10	Start Continuous Conversions upon writing to register 0.
11	Start on external SYNC input going high (Requires Slave mode: M/S = 0)

Bits 7,6 (reserved):

These bits are reserved by Micro Linear and must be written as zero.

Bits 10,9,8 (TMDE: test mode select bits)

These bits are used for diagnostic purposes only and normally not accessed during operation. The default value of TMDE is 000 which selects a normal A/D conversion. See Diagnostics for more information.

TMDE	Description
000	Normal A/D Conversion
001	Reserved by Micro Linear (Do Not Use)
010	CALWR Operation
011	CALRD Operation
100	System Offset
101	Common-mode
110	Plus Full Scale
111	Minus Full Scale

Bit 11 (M/S: Master/Slave bit):

Dictates whether the SYNC pin is an input or an output. Upon RESET, this bit is cleared.

M/S = 0: Slave Mode SYNC is an input which is used to trigger a conversion if SMDE = 11.

M/S = 1: Master Mode SYNC is an output. At the beginning of every conversion, SYNC is high for 4 internal clocks.

Bit 12 (CK1X: clock select bit):

Selects whether the external clock will be divided by two or used directly as the internal clock. See Clock section for a detailed explanation.

CK1X = 0: the external clock is divided by two and used as the internal clock. This is referred to as CLK Mode = 0.

CK1X = 1: the external clock input is used directly as the internal clock. This is referred to as CLK Mode = 1.

2

FUNCTIONAL DESCRIPTION (Continued)

GENERAL OPERATING INFORMATION CONVERSION-START PROTOCOL

There are four different ways to start a conversion. They are defined by SMDE bits 4 and 5 in the Control Register.

SMDE Bits 5,4

- 00: A write to register 0 will start a conversion. During a conversion, if another write is issued to register 0, the "Start Conversion" command will be latched and another conversion will immediately follow the current one. To insure that the second write will be latched, it must occur at least 3 internal clocks after the first write. Only one additional write will be latched; multiple writes within a conversion will only yield one more conversion.
- 01: Reading the data from the previous conversion starts the next conversion.
- 10: This mode causes continuous conversions; the next conversion begins immediately after the previous conversion ends. Writing to register 0 will start the first conversion; thereafter the converter runs continuously. This mode yields the maximum conversion rate.
- 11: The SYNC input triggers the start of a conversion. The M/S bit in the Control Register must be cleared, placing the chip in the slave mode.

Note: The external activation signals for Start Modes 0, 1, and 3 are synchronized internally to the system clock. If

periodic sampling is required using these Start Modes, the SYNC, \overline{RD} , or \overline{WR} pulses must be synchronized to the system clock. Start Mode 2 guarantees periodic sampling.

DOUBLE-BUFFERED DATA REGISTER

The A/D conversion result is double-buffered using the Data Buffer register and the A/D Data register. The actual End-Of-Conversion (EOC) does not correspond with the \overline{DAV} output going low. The \overline{DAV} output goes low 16 internal clocks after the EOC. From the time \overline{DAV} output goes LOW, the user has one full conversion time (80 or 110 internal clocks) minus 16 internal clocks to read the data as shown in Figure 10.

SELF CALIBRATION

Setting the DOCAL bit issues a calibration request to the chip. When calibration is done, the \overline{DAV} status bit is set and the \overline{DAV} output goes low.

A calibration requires 8,260 internal clocks. Using a 7MHz clock (CLK Mode=0), this is approximately 2 ms. Power supplies and external voltage reference must be stable before issuing a request for calibration.

The ML2233 should be calibrated before any conversions are attempted. Calibrations must not be performed simultaneously with conversions. Before requesting a calibration, the user may want to read the Busy status bit to make sure that the converter is idle. Polling the chip while the calibration is in progress is not recommended.

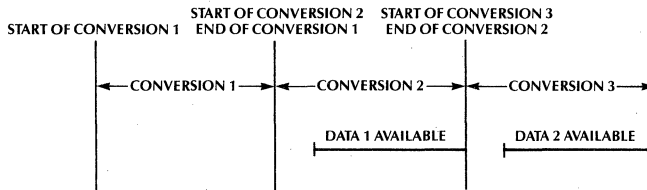


Figure 10

FUNCTIONAL DESCRIPTION (Continued)

CLOCK

The ML2233 has the option of dividing the clock at the CLK pin by 2, or using it directly to drive the internal logic. This option is selected through the CK1X bit in the Control register. When CK1X = 0 the clock is divided by 2. This is referred to as CLK Mode = 0. The clock at the CLK pin is referred to as the External clock, and the Internal Clock is the External clock divided by 2. When CK1X = 1, the clock at the CLK pin drives the internal logic directly, therefore this clock is referred to as the Internal clock. This is also known as CLK Mode = 1. All internally clocked logic is positive edge triggered.

CLK Mode = 0:

There are two advantages to CLK Mode 0. This is the only Mode that allows an external crystal to be used. CLK Mode 1 cannot operate with an external crystal, the CLK pin must be driven. The second advantage of CLK Mode 0 is that the duty cycle for a driven clock is less stringent than in CLK Mode 1. (Refer to t_{CLK0} and t_{CLK1} in AC Electrical Characteristics for CLK Mode 0 and 1 timing requirements, respectively.)

On power up the state of the divide by two flip-flop is indeterminate. Therefore the relationship between the internal clock and the external clock at the CLK pin can have one of two possibilities as shown in Figure 11. As a result the following should be considered.

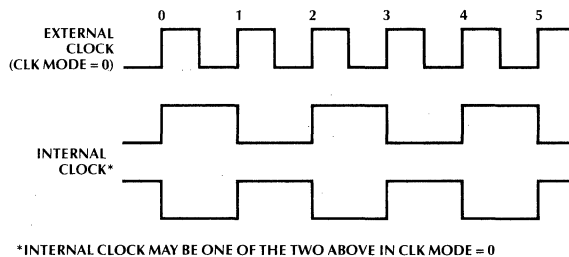


Figure 11

t_{WRCK} , t_{RDCK} , and t_{SYNCK} specs, (\overline{RD} , \overline{WR} , and SYNC setup times to Start of Conversion), will be as shown in the data sheet, or the data sheet specs plus one external clock period. Since these specifications are with respect to the rising edge of the external clock, it is not known whether this rising edge corresponds to the rising edge or falling edge of the internal clock. Therefore there is an uncertainty of one external clock period.

If periodic sampling is necessary and Start Mode 0, 1, or 3 is used, the external start pulse (either \overline{RD} , \overline{WR} , or SYNC) must be synchronous to the external clock, meet the setup time, and be an even number of external clock periods. If the start pulse were an odd number of external clock periods, half the pulses would correspond with the rising edge of the internal clock, and the other half would correspond with the falling edge of the internal clock. Therefore the sampling period would change by one external clock period every sample. Start Mode 2 guarantees periodic sampling regardless of the CLK mode.

CLK Mode = 1:

This mode eliminates the requirement that external start pulses must be an even number of external clock periods. However periodic sampling still requires that the start pulse be synchronous to the external clock, and the setup time must be met. CLK Mode 1 also eliminates the uncertainty of the t_{WRCK} , t_{RDCK} , and t_{SYNCK} requirements.

FUNCTIONAL DESCRIPTION (Continued)

DIAGNOSTICS

Diagnostic routines may be run after power up or any other time to ensure proper operation. The diagnostic features, which are software selectable, don't require external hardware. Both the analog and digital sections can be tested.

The ML2233 is placed in the diagnostic mode via the TMDE field in the Control Register. Once the ML2233 is placed in one of the diagnostic modes, a conversion must be executed before the results can be read. As with all conversions, $\overline{\text{DAV}}$ will be activated upon completion.

ANALOG CONVERSION DIAGNOSTICS

TMDE=000: Normal Operation

Selects normal A/D conversion. Default condition after a software reset.

TMDE=001: Reserved by Micro Linear.

TMDE=010: CALWR operation

The data in Write register 0 (CALCODE Holding Register), is transferred into the converter's Calibration register when a "Start Conversion" is issued. A dummy conversion occurs and the $\overline{\text{DAV}}$ output goes LOW to indicate that the operation is complete.

TMDE=011: CALRD operation

The contents of the Calibration register are transferred through the A/D Data register and loaded into the Data Buffer register. A dummy 8-bit conversion occurs and $\overline{\text{DAV}}$ output goes LOW to indicate that the CALRD operation is complete.

TMDE = 100: System Offset

The positive and negative inputs to the Sample and Hold are tied to analog ground. With this setting, converted data will give the offset of the A/D converter and Sample/ Hold combination. The $V_{\text{IN}+}$ and $V_{\text{IN}-}$ pins will remain in a high impedance state while in this mode.

TMDE = 101: Common-mode

Both the positive and negative inputs of the Sample and Hold are tied to V_{REF} . The results of a conversion in this test mode indicates how well the converter is rejecting a common mode signal.

TMDE = 110: Positive Full Scale

This test mode connects the positive input of the Sample and Hold to V_{REF} and the negative input of the Sample and Hold to analog ground. The result of converting in this test mode is a value near positive full scale.

TMDE = 111: Negative Full Scale

This test mode connects the positive input of the Sample and Hold to analog ground and the negative input to V_{REF} . The result of converting in this test mode is a value near negative full scale.

DIGITAL LOOPBACK

The ML2233's architecture provides a way for the microprocessor to indirectly read and write to the A/D converter's calibration register and data register via a CALRD and CALWR. Figure 12 illustrates this architecture. This in effect allows a digital loopback.

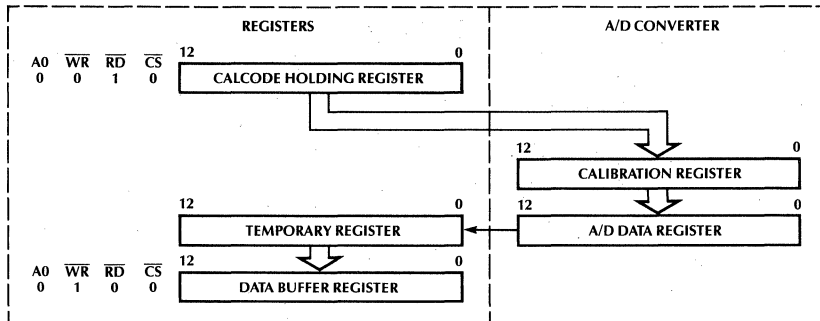


Figure 12. Digital Loopback

FUNCTIONAL DESCRIPTION (Continued)

When the TMDE bits are set to 010 CAL WRITE (CALWR), and a Start Conversion is issued in any one of the four modes, the contents of the CALCODE Holding register is transferred into the A/D converter's Calibration register. When the TMDE bits are set to 011 CAL READ (CALRD), and a Start Conversion is issued, the contents of the Calibration register are transferred through the A/D's Data register into the Data Buffer register. The result of these two operations is a complete loopback from the CALCODE Holding register through the A/D converter and back into the Data buffer register. This loopback provides user assurance that all the paths are clear and there are no stuck bits.

Note: When a CALWR is done, the previous calibration value is lost. The correct calibration value must be re-stored before the converter is used to convert data.

CALIBRATION PASS/FAIL TEST

The CALRD can be used as a way to verify a successful calibration. After a calibration is completed, the CALRD may be issued in order to read the contents of the Calibration register. If the Low Byte (lower 8 bits) of the data buffer register are ones after executing a CALRD, the calibration failed; otherwise the calibration is successful.

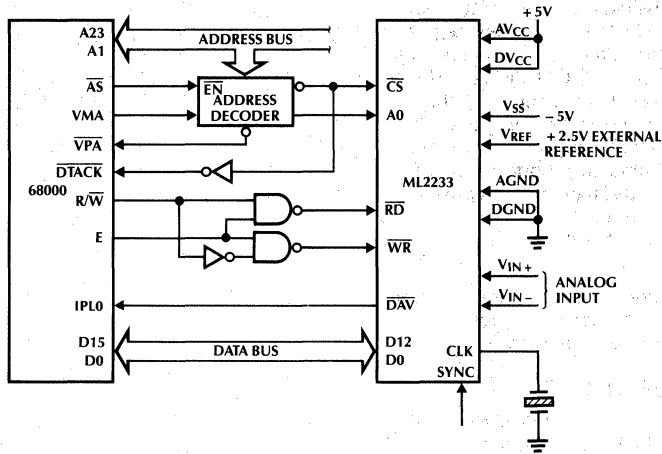


Figure 13. Interfacing to 68000 Microprocessor

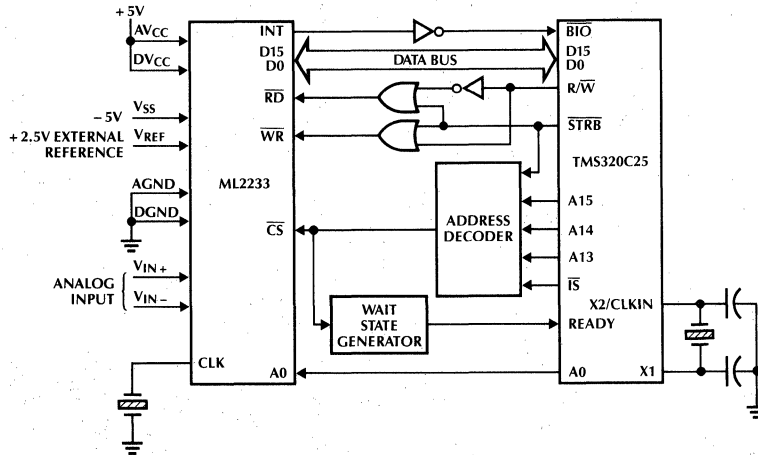


Figure 14. Interfacing to TMS320C25 Digital Signal Processor

ORDERING INFORMATION

PART NUMBER	MAXIMUM LINEARITY ERROR	MAXIMUM TOTAL UNADJUSTED ERROR	MINIMUM CONVERSION	PACKAGE
ML2233BCJ	$\pm 3/4$ LSB	$\pm 1 1/2$ LSB	31.5 μ s	Hermetic DIP (J28)
ML2233CCJ	± 1 LSB	$\pm 2 1/2$ LSB	31.5 μ s	Hermetic DIP (J28)
ML2233DCJ	± 1 LSB	$\pm 2 1/2$ LSB	44.0 μ s	Hermetic DIP (J28)

μ P Compatible 8-Bit A/D Converters with 2- or 8-Channel Multiplexer

GENERAL DESCRIPTION

The ML2252 and ML2259 combine an 8-bit A/D converter, 2- or 8-channel analog multiplexer, and a microprocessor compatible 8-bit parallel interface and control logic in a single monolithic CMOS device.

Easy interface to microprocessors is provided by the latched and decoded multiplexer address inputs and a double buffered three-state data bus. These analog-to-digital converters allow the microprocessor to operate completely asynchronous to the converter clock.

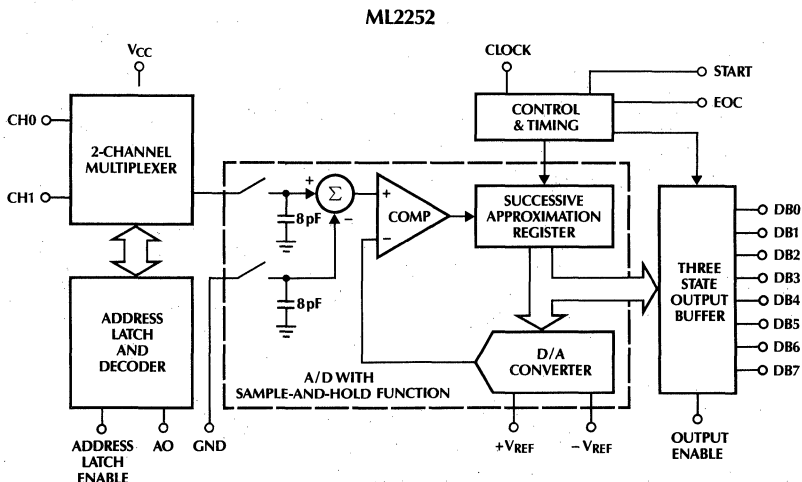
The built in sample and hold function provides the ability to digitize a 5V, 50kHz sine wave to 8-bit accuracy. The differential comparator design provides low power supply sensitivity to DC and AC variations. The voltage reference can be externally set to any value between ground and V_{CC} , thus allowing a full conversion over a relatively small span if desired. All parameters are guaranteed over temperature with a power supply voltage of $5V \pm 10\%$.

The device is suitable for a wide range of applications from process and machine control to consumer, automotive, and telecommunication applications.

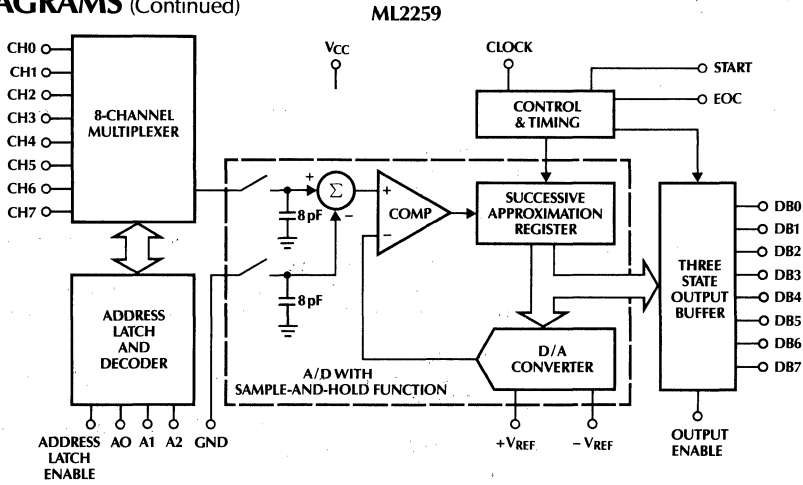
FEATURES

- Conversion time ($f_{CLK} = 1.46 \text{ MHz}$) 6.6 μ s
- Total unadjusted error $\pm 1/2 \text{ LSB}$ or $\pm 1 \text{ LSB}$
- No missing codes
- Sample and hold 390ns acquisition
- Capable of digitizing a 5V, 50kHz sine wave
- 2- or 8-channel input multiplexer
- 0V to 5V analog input range with single 5V power supply
- Operates ratiometrically or with up to 5V voltage reference
- No zero or full scale adjust required
- Analog input protection 25 mA (min) per input
- Continuous conversion mode
- Low power dissipation 15mW MAX
- TTL and CMOS compatible digital inputs and outputs
- Standard 20-pin or 28-pin DIP or PCC
- Temperature range 0°C to $+70^{\circ}\text{C}$,
or -40°C to $+85^{\circ}\text{C}$,
or -55°C to $+125^{\circ}\text{C}$

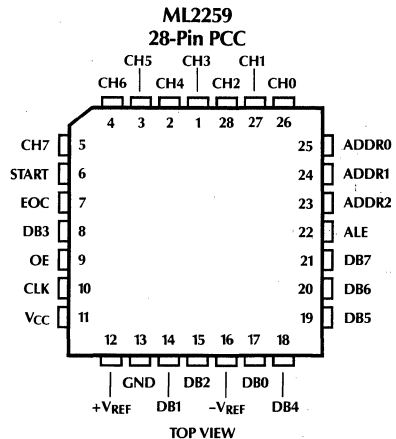
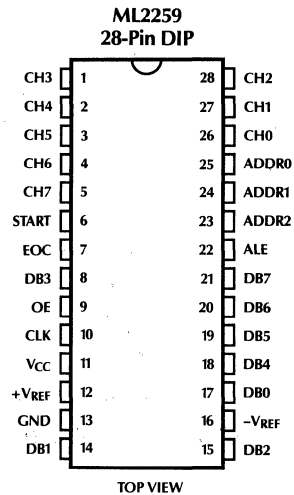
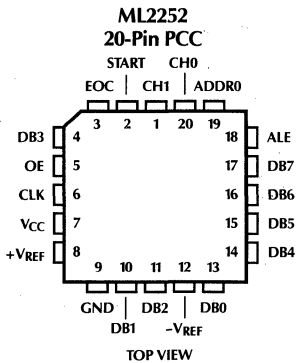
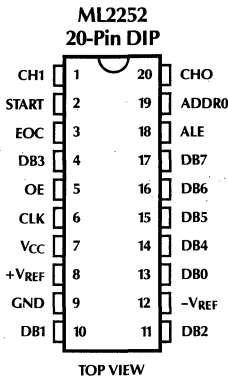
BLOCK DIAGRAMS



BLOCK DIAGRAMS (Continued)



PIN CONNECTIONS



ML2252, ML2259

PIN DESCRIPTION

PIN #		NAME	FUNCTION
ML2252	ML2259		
	1	CH3	Analog input 3.
	2	CH4	Analog input 4.
	3	CH5	Analog input 5.
	4	CH6	Analog input 6.
	5	CH7	Analog input 7.
2	6	START	Start of conversion. Active high digital input pulse initiates conversion.
3	7	EOC	End of conversion. This output goes low after a START pulse occurs, stays low for the entire A/D conversion, and goes high after conversion is completed. Data on DB0-DB7 is valid on rising edge of EOC and stays valid until next EOC rising edge.
4	8	DB3	Data output 3.
5	9	OE	Output enable input. When OE = 0, DB0-DB7 are in high impedance state; OE = 1, DB0-DB7 are active outputs.
6	10	CLK	Clock. Clock input provides timing for A/D converter, S/H, and digital interface.
7	11	V _{CC}	Positive supply. 5V ± 10%.
8	12	+V _{REF}	Positive reference voltage.
9	13	GND	Ground. 0V, all analog and digital inputs or outputs are referenced to this point.
10	14	DB1	Data output 1.
11	15	DB2	Data output 2.
12	16	-V _{REF}	Negative reference voltage.
13	17	DB0	Data output 0.
14	18	DB4	Data output 4.
15	19	DB5	Data output 5.
16	20	DB6	Data output 6.
17	21	DB7	Data output 7.
18	22	ALE	Address latch enable. Input to latch in the digital address (ADDR2-0) on the rising edge of the multiplexer.
	23	ADDR2	Address input 2 to multiplexer. Digital input for selecting analog input.
	24	ADDR1	Address input 1 to multiplexer. Digital input for selecting analog input.
19	25	ADDR0	Address input 0 to multiplexer. Digital input for selecting analog input.
20	26	CH0	Analog input 0.
1	27	CH1	Analog input 1.
	28	CH2	Analog input 2.

ABSOLUTE MAXIMUM RATINGS

(Note 1)

Supply Voltage, V _{CC}	6.5V
Voltage	
Logic Inputs	-0.3V to V _{CC} +0.3V
Analog Inputs	-0.3V to V _{CC} +0.3V
Input Current per Pin (Note 2)	±25mA
Storage Temperature	-65°C to +150°C
Package Dissipation	
at T _A = 25°C (Board Mount)	875mW
Lead Temperature (Soldering 10 sec.)	
Dual-In-Line Package (Plastic)	260°C
Dual-In-Line Package (Ceramic)	300°C
Molded Chip Carrier Package	
Vapor Phase (60 sec.)	215°C
Infrared (15 sec.)	220°C

OPERATING CONDITIONS

Supply Voltage, V _{CC}	4.5V _{DC} to 6.3V _{DC}
Temperature Range (Note 3)	T _{MIN} ≤ T _A ≤ T _{MAX}
ML2252BMJ, ML2252CMJ	-55°C to +125°C
ML2259BMJ, ML2259CMJ	
ML2252BIJ, ML2252CIJ	-40°C to +85°C
ML2259BIJ, ML2259CIJ	
ML2252BCP, ML2259BCP	0°C to +70°C
ML2252BCQ, ML2259BCQ	
ML2252CCP, ML2259CCP	
ML2252CCQ, ML2259CCQ	

ELECTRICAL CHARACTERISTICS

Unless otherwise specified $T_A = T_{MIN}$ to T_{MAX} , $V_{CC} = +V_{REF} = 5V \pm 10\%$, $-V_{REF} = GND$ and $f_{CLK} = 1.46\text{MHz}$.

PARAMETER	NOTES	CONDITIONS	ML2252B, ML2259B			ML2252C, ML2259C			UNITS
			MIN	TYP NOTE 4	MAX	MIN	TYP NOTE 4	MAX	
CONVERTER AND MULTIPLEXER CHARACTERISTICS									
Total Unadjusted Error	5, 7	$V_{REF} = V_{CC}$			$\pm 1/2$			± 1	LSB
+ V_{REF} Voltage Range	6		$-V_{REF}$		$V_{CC} + 0.1$	$-V_{REF}$		$V_{CC} + 0.1$	V
- V_{REF} Voltage Range	6		$GND - 0.1$		+ V_{REF}	$GND - 0.1$		+ V_{REF}	V
Reference Input Resistance	5		14	20	28	14	20	28	k Ω
Analog Input Range	5, 8		$GND - 0.1$		$V_{CC} + 0.1$	$GND - 0.1$		$V_{CC} + 0.1$	V
Power Supply Sensitivity	6	DC $V_{CC} = 5V \pm 10\%$		$\pm 1/32$	$\pm 1/4$		$\pm 1/32$	$\pm 1/4$	LSB
		100 mV _{p,p} 100 kHz Sine on V_{CC} , $V_{IN} = 0$		$\pm 1/16$			$\pm 1/16$		LSB
I_{Off} , Off Channel Leakage Current (Note 9)	5, 9	On Channel = V_{CC} Off Channel = 0V	-1			-1			μA
		On Channel = 0V Off Channel = V_{CC}			1		1		μA
I_{On} , On Channel Leakage Current (Note 9)	5, 9	On Channel = 0V Off Channel = V_{CC}	-1			-1			μA
		On Channel = V_{CC} Off Channel = 0V			1		1		μA

SYMBOL	PARAMETER	NOTES	CONDITIONS	MIN	TYP NOTE 4	MAX	UNITS
DIGITAL AND DC CHARACTERISTICS							
$V_{IN(1)}$	Logical "1" Input Voltage	5		2.0			V
$V_{IN(0)}$	Logical "0" Input Voltage	5				0.8	V
$I_{IN(1)}$	Logical "1" Input Current	5	$V_{IN} = V_{CC}$			1	μA
$I_{IN(0)}$	Logical "0" Input Current	5	$V_{IN} = 0V$	-1			μA
$V_{OUT(1)}$	Logical "1" Output Voltage	5	$I_{OUT} = -2\text{mA}$	4.0			V
$V_{OUT(0)}$	Logical "0" Output Voltage	5	$I_{OUT} = 2\text{mA}$			0.4	V
I_{OUT}	Three-State Output Current	5	$V_{OUT} = 0V$	-1			μA
			$V_{OUT} = V_{CC}$			1	μA
I_{CC}	Supply Current	5			1.5	3	mA

AC AND DYNAMIC PERFORMANCE CHARACTERISTICS (Note 10)

t_{ACQ}	Sample & Hold Acquisition				1/2		$1/f_{CLK}$
f_{CLK}	Clock Frequency	5		10		1460	kHz
t_C	Conversion Time				8.5	$8.5 + 250\text{ns}$	$1/f_{CLK}$
SNR	Signal to Noise Ratio		$V_{IN} = 51\text{kHz}, 5V$ Sine. $f_{CLK} = 1.46\text{MHz}$ ($f_{SAMPLING} \approx 150\text{kHz}$). Noise is Sum of All Nonfundamental Components up to $1/2$ of $f_{SAMPLING}$		47		dB
THD	Total Harmonic Distortion		$V_{IN} = 51\text{kHz}, 5V$ Sine. $f_{CLK} = 1.46\text{MHz}$ ($f_{SAMPLING} \approx 150\text{kHz}$). THD is Sum of 2, 3, 4, 5 Harmonics Relative to Fundamental		-60		dB

ELECTRICAL CHARACTERISTICS (Continued)

Unless otherwise specified $T_A = T_{MIN}$ to T_{MAX} , $V_{CC} = +V_{REF} = 5V \pm 10\%$, $-V_{REF} = GND$ and $f_{CLK} = 1.46\text{ MHz}$.

SYMBOL	PARAMETER	NOTES	CONDITIONS	MIN	TYP NOTE 4	MAX	UNITS
AC AND DYNAMIC PERFORMANCE CHARACTERISTICS (Note 10) (Continued)							
IMD	Intermodulation Distortion		$V_{IN} = f_A + f_B$, $f_A = 49\text{ kHz}$, 2.5 V Sine , $f_B = 47.8\text{ kHz}$, 2.5 V Sine , $f_{CLK} = 1.46\text{ MHz}$ ($f_{SAMPLING} \cong 150\text{ kHz}$). IMD is ($f_A + f_B$), ($f_A - f_B$), ($2f_A + f_B$), ($2f_A - f_B$), ($f_A + 2f_B$), ($f_A - 2f_B$) Relative to Fundamental		-60		dB
FR	Frequency Response		$V_{IN} = 0$ to 50 kHz . 5 V Sine Relative to 1 kHz		0.1		dB
t_{DC}	Clock Duty Cycle	6, 11		40		60	%
t_{EOC}	End of Conversion Delay	5			1/2	1/2 + 250 ns	1/ f_{CLK}
t_{WS}	Start Pulse Width	5		50			ns
t_{SS}	Start Pulse Setup Time	6, 12	Synchronous Only	40			ns
t_{WALE}	Address Latch Enable Pulse Width	5		50			ns
t_S	Address Setup	5		0			ns
t_H	Address Hold	5		50			ns
t_{HI} , t_{HO}	Output Enable for DB0-DB7	6	Figure 1, $C_L = 50\text{ pF}$			100	ns
		6	Figure 1, $C_L = 10\text{ pF}$			50	ns
t_{IH} , t_{OH}	Output Disable for DB0-DB7	6	Figure 1, $C_L = 50\text{ pF}$			100	ns
		6	Figure 1, $C_L = 10\text{ pF}$			50	ns
C_{IN}	Capacitance of Logic Input				5		pF
C_{OUT}	Capacitance of Logic Outputs				10		pF

Note 1: Absolute maximum ratings are limits beyond which the life of the integrated circuit may be impaired. All voltages unless otherwise specified are measured with respect to ground.

Note 2: When the input voltage (V_{IN}) at any pin exceeds the power supply rails ($V_{IN} < GND - 0.1\text{ V}$ or $V_{CC} + 0.1\text{ V}$) the absolute value of current at that pin should be limited to 25 mA or less.

Note 3: -55°C to $+125^\circ\text{C}$ operating temperature range devices are 100% tested at temperature extremes with worst-case test conditions. -40°C to $+85^\circ\text{C}$ operating temperature range devices are 100% tested with temperature limits guaranteed by 100% testing, sampling, or by correlation with worst-case test conditions.

Note 4: Typicals are parametric norm at 25°C .

Note 5: Parameter guaranteed and 100% production tested.

Note 6: Parameter guaranteed. Parameters not 100% tested are not in outgoing quality level calculation.

Note 7: Total unadjusted error includes offset, full-scale, linearity, multiplexer and sample and hold errors.

Note 8: For $-V_{REF} \geq V_{IN}(+)$ the digital output code will be 0000 0000. Two on-chip diodes are tied to each analog input which will forward conduct for analog input voltages one diode drop below ground or one diode drop greater than the V_{CC} supply. Be careful, during testing at low V_{CC} levels (4.5 V), as high level analog inputs (5 V) can cause this input diode to conduct—especially at elevated temperatures, and cause errors for analog inputs near full-scale. The spec allows 100 mV forward bias of either diode. This means that as long as the analog V_{IN} or V_{REF} does not exceed the supply voltage by more than 100 mV , the output code will be correct. To achieve an absolute $0V_{DC}$ to $5V_{DC}$ input voltage range will therefore require a minimum supply voltage of $4.900V_{DC}$ over temperature variations, initial tolerance and loading.

Note 9: Leakage current is measured with the clock not switching.

Note 10: $C_L = 50\text{ pF}$, timing measured at 50% point.

Note 11: A 40% to 60% clock duty cycle range insures proper operation at all clock frequencies. In the case that an available clock has a duty cycle outside of these limits, the minimum time the clock is high or the minimum time the clock is low must be at least 40 ns . The maximum time the clock can be high or low is $60\mu\text{s}$.

Note 12: The conversion start setup time requirement only needs to be satisfied if a conversion must be synchronized to a given clock rising edge. If the setup time is not met, start conversion will have an uncertainty of one clock pulse.

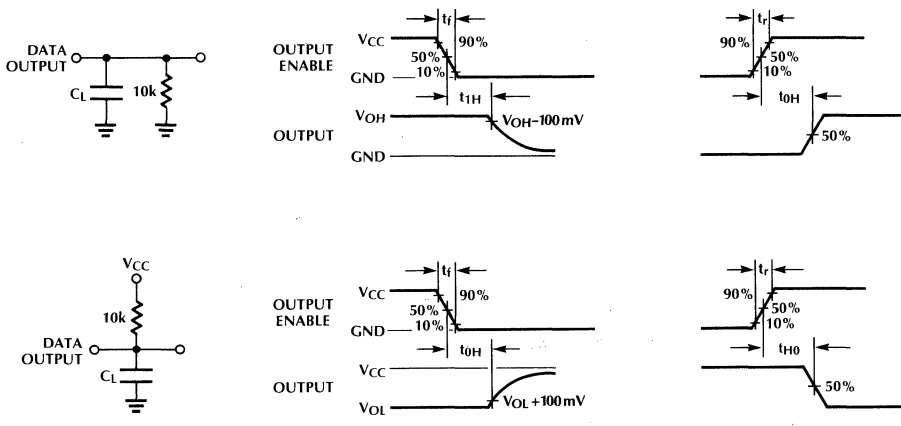


Figure 1. High Impedance Test Circuits and Waveforms

2

TYPICAL PERFORMANCE CURVES

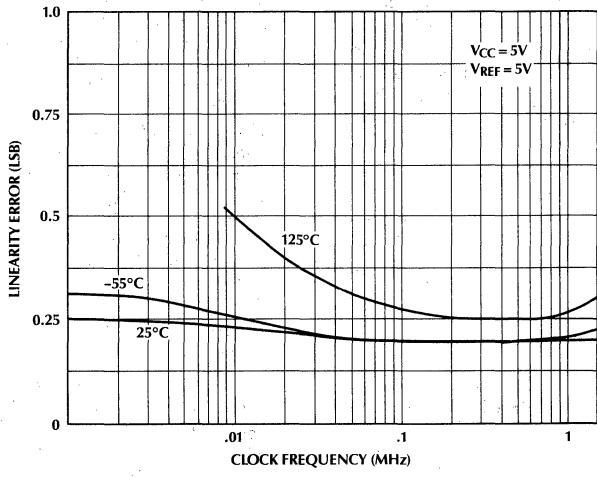


Figure 2. Linearity Error vs f_{CLK}

TYPICAL PERFORMANCE CURVES (Continued)

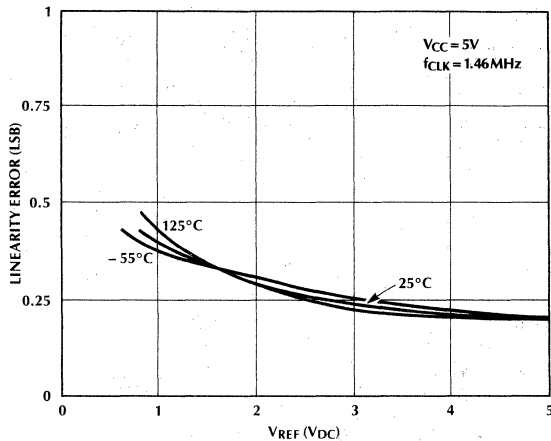


Figure 3. Linearity Error vs V_{REF} Voltage

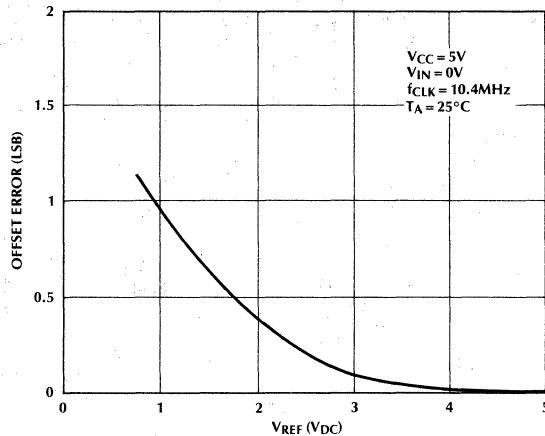


Figure 4. Unadjusted Offset Error vs V_{REF} Voltage

1.0 FUNCTIONAL DESCRIPTION

1.1 Multiplexer Addressing

The ML2252 and ML2259 contain a single ended analog multiplexer. A particular input channel is selected by using the address decoder. The relationship between the address inputs, ADDR0-ADDR2, and the analog input selected is shown in Table 1. The address inputs are latched into the decoder on the rising edge of the address latch signal ALE.

Table 1. Multiplexer Address Decoding
ML2252

Selected Analog Channel	Address Input
CH0	0
CH1	1

ML2259

Selected Analog Channel	Address Input		
	ADDR2	ADDR1	ADDR0
CH0	0	0	0
CH1	0	0	1
CH2	0	1	0
CH3	0	1	1
CH4	1	0	0
CH5	1	0	1
CH6	1	1	0
CH7	1	1	1

1.2 A/D Converter

The A/D converter uses successive approximation to perform the conversion. The converter is composed of the successive approximation register, the DAC and the comparator.

The DAC generates the precise levels that determine the linearity and accuracy of the conversion. The DAC is composed of a capacitor upper array and a resistor lower array. The capacitor upper array generates the 4 MSB decision levels while the series resistor lower array generates the 4 LSB decision levels. A switch decoder tree is used to decode the proper level from both arrays.

The capacitor/resistor array offers fast conversion, superior linearity and accuracy since matching is only required between $2^4 = 16$ elements (as opposed to $2^8 = 256$ elements in conventional designs). And since the levels are based on the ratio of capacitors to capacitors and resistors to resistors, the accuracy and long term stability of the converter is improved. This also guarantees monotonicity and no missing codes, as well as eliminating any linearity temperature or power supply dependence.

The successive approximation register is a digital block used to store the bit decisions from the conversion.

The comparator design is unique in that it is fully differential and auto zeroed. The fully differential architecture provides excellent noise immunity, excellent power supply rejection, and wide common mode range. The comparator is auto zeroed at the start of each conversion in order to remove any DC offset and full-scale gain error, thus improving accuracy and linearity.

Another advantage of the capacitor array approach used in the ML2252 and ML2259 is the inherent sample-and-hold function. This true S/H allows an accurate conversion to be done on the input even if the analog signal is not stable.

1.0 FUNCTIONAL DESCRIPTION (Continued)

Linearity and accuracy are maintained for analog signals up to 1/2 the sampling frequency. As a result, input signals up to 50 kHz can be converted without degradation in linearity or accuracy.

The sequence of events during a conversion is shown in Figure 5. The rising edge of a START pulse resets the internal registers and initiates a conversion on the next rising edge of CLK providing that (t_{ss}) start pulse setup time is satisfied. If this setup time is not met, start conversion will have an uncertainty of one clock pulse. The input is then sampled for the next half CLK period until EOC goes low. EOC goes low on the falling edge of the next CLK pulse indicating that the conversion is now beginning. The actual conversion now takes place for the next eight CLK pulses, one bit for each CLK pulse. After the conversion is done, the data is updated on DB0-DB7 and EOC goes high on the rising edge of the 9th CLK pulse, indicating that the conversion has been completed and data is valid on DB0-DB7. The data will stay valid on DB0-DB7 until the next conversion updates the data word on the next rising edge of EOC.

A conversion can be interrupted and restarted at any time by a new START pulse.

1.3 Analog Inputs and Sample/ Hold

The ML2252 and ML2259 have a true sample-and-hold circuit which samples both the selected input and ground simultaneously. These analog to digital converters can reject AC common mode signals from DC-50 kHz as well as maintain linearity for signals from DC-50 kHz.

The plot below (Figure 6) shows a 2048 point FFT of the ML2259 converting a 50 kHz, 0 to 5 V, low distortion sine wave input. The ML2252 and ML2259 sample and digitize at their specified accuracy, dynamic input signals with frequency components up to the Nyquist frequency (one-half the sampling rate). The output spectra yields precise measurements of input signal level, harmonic components, and signal to noise ratio up to the 8-bit level. The near ideal signal to noise ratio is maintained independent of increasing analog input frequencies to 50 kHz.

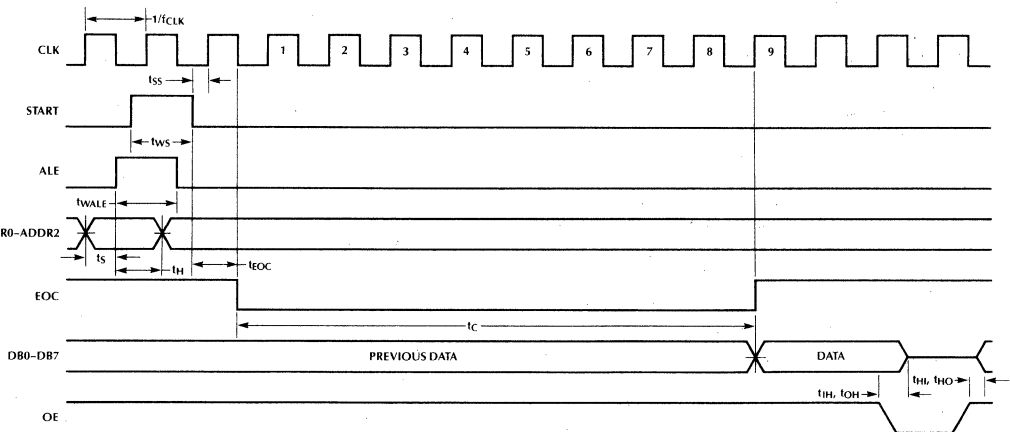


Figure 5. Timing Diagram

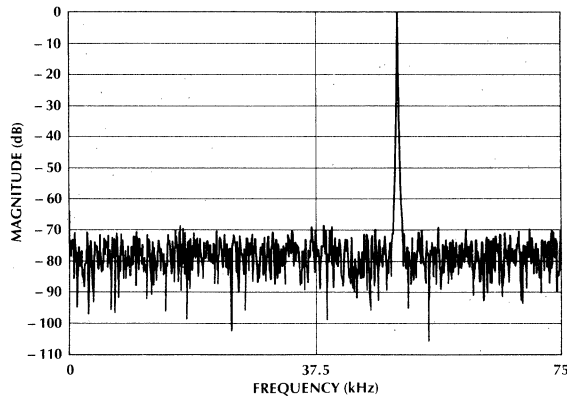


Figure 6. Output Spectrum

1.0 FUNCTIONAL DESCRIPTION (Continued)

The signal at the analog input is sampled during the interval when the sampling switch is open prior to conversion start. The sampling window (S/H acquisition time) is one half CLK period long and occurs one half CLK period after START goes low. When the sampling switch closes at the start of the S/H acquisition time, 8 pF of capacitance is thrown onto the analog input. One half CLK period later, the sampling switch opens, the signal present at analog input is stored and conversion starts. Since any error on the analog input at the end of the S/H acquisition time will cause additional conversion error, care should be taken to insure adequate settling and charging time from the source. If more charging or settling time is needed to reduce these analog input errors, a longer CLK period can be used.

Each analog input has dual diodes to the supply rails, and a minimum of ± 25 mA (± 100 mA typically) can be injected into each analog input without causing latchup.

1.4 Reference

The voltage applied to the $+V_{REF}$ and $-V_{REF}$ inputs defines the voltage span of the analog input (the difference between V_{INMAX} and V_{INMIN}) over which the 256 possible output codes apply. The devices can be used in either ratiometric applications or in systems requiring absolute accuracy. The reference pins must be connected to a voltage source capable of driving the reference input resistance, typically 20k.

In a ratiometric system, the analog input voltage is proportional to the voltage used for the A/D reference. This voltage is typically the system power supply, so the $+V_{REF}$ pin can be tied to V_{CC} and $-V_{REF}$ tied to GND. This technique relaxes the stability requirements of the system reference as the analog input and A/D reference move together maintaining the same output code for a given input condition.

For absolute accuracy, where the analog input varies between specific voltage limits, the reference pins can be biased with a time and temperature stable voltage source.

$+V_{REF}$ and $-V_{REF}$ can be at any voltage between V_{CC} and GND. In addition, the difference between $+V_{REF}$ and $-V_{REF}$ can be set to small values for conversions over smaller voltage ranges. Particular care must be taken with regard to noise pickup, circuit layout and system error voltage sources when operating with a reduced span due to the increased sensitivity of the converter.

1.5 Power Supply and Reference Decoupling

A 10 μ F electrolytic capacitor is recommended to bypass V_{CC} to GND, using as short a lead length as possible. In addition, with clock frequencies above 1MHz, a 0.1 μ F ceramic disc capacitor should be used to bypass V_{CC} to GND.

If REF+ and REF- inputs are driven by long lines, they should be bypassed by 0.1 μ F ceramic disc capacitors at the reference input pins (pins 12, 16).

1.6 Dynamic Performance

Signal-to-Noise Ratio

Signal-to-noise ratio (SNR) is the measured signal to noise at the output of the converter. The signal is the rms magnitude of the fundamental. Noise is the rms sum of all the nonfunda-

mental signals up to half the sampling frequency. SNR is dependent on the number of quantization levels used in the digitization process; the more the levels, the smaller the quantization noise. The theoretical SNR for a sine wave is given by

$$SNR = (6.02N + 1.76)dB$$

where N is the number of bits. Thus for ideal 8-bit converter, SNR = 49.92 dB.

Harmonic Distortion

Harmonic distortion is the ratio of the rms sum of harmonics to the fundamental. Total harmonic distortion (THD) of the ML2252 or ML2259 are defined as

$$20 \log = \frac{(V_2^2 + V_3^2 + V_4^2 + V_5^2)^{1/2}}{V_1}$$

where V_1 is the rms amplitude of the fundamental and V_2, V_3, V_4, V_5 are the rms amplitudes of the individual harmonics.

Intermodulation Distortion

With inputs consisting of sine waves at two frequencies, f_A and f_B , any active device with nonlinearities will create distortion products, of order $(m + n)$, at sum and difference frequencies of $mf_A + nf_B$, where $m, n = 0, 1, 2, 3 \dots$ Intermodulation terms are those for which m or n is not equal to zero. The (IMD) intermodulation distortion specification includes the second order terms $(f_A + f_B)$ and $(f_A - f_B)$ and the third order terms $(2f_A + f_B)$, $(2f_A - f_B)$, $(f_A + 2f_B)$, and $(f_A - 2f_B)$ only.

1.7 Digital Interface

The analog inputs are selected by the digital addresses, ADDR0-ADDR2, and latched on the rising edge of ALE. This is described in the Multiplexer Addressing section.

A conversion is initiated by the rising edge of a START pulse. As long as this pulse is high, the internal logic is reset.

The sampling interval starts with the following CLK rising edge after a START falling edge and ends on the falling edge of CLK. The conversion starts and EOC goes low. The sampling clock is at least one half CLK period wide. Each bit conversion in the successive approximation process takes 1 CLK period. On the rising edge of the ninth CLK pulse, the digital output of the conversion is updated on the outputs DB0-DB7 and EOC goes high indicating the conversion is done and data on DB0-DB7 is valid.

One feature of the ML2252 and ML2259 is that the data is double buffered. This means that the outputs DB0-DB7 will stay valid until updated at the end of the next conversion and will not become invalid when the next conversion starts. This facilitates interfacing with external logic of μ P.

The signal OE drives the data bus, DB0-DB7, into the high impedance state when held low. This allows the ML2252 and ML2259 to be tied directly to a μ P system bus without any latches or buffers.

1.0 FUNCTIONAL DESCRIPTION (Continued)

1.7.1 Restart During Conversion

If the A/D is restarted (start goes low and returns high) during a conversion, the converter is reset and a new conversion is started. The output data latch is not updated if the conversion in process is not allowed to be completed. EOC will remain low and the output data latch is not updated.

1.7.2 Continuous Conversions

In the free-running, continuous conversion mode, the start input is tied to the (Figure 7) EOC output. An initialization pulse, following power-up, of momentarily forcing a logic high level is required to guarantee operation.

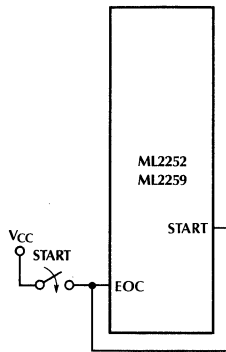


Figure 7. Continuous Conversion Mode

2

2.0 TYPICAL APPLICATIONS

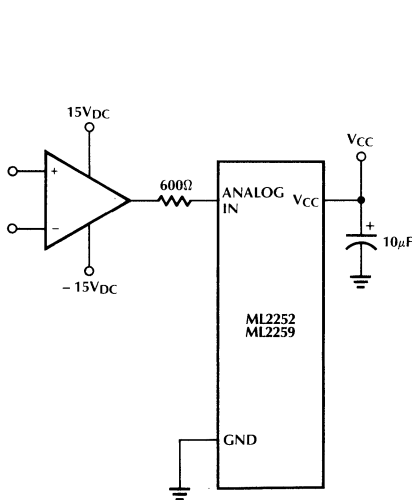


Figure 8. Protecting the Input

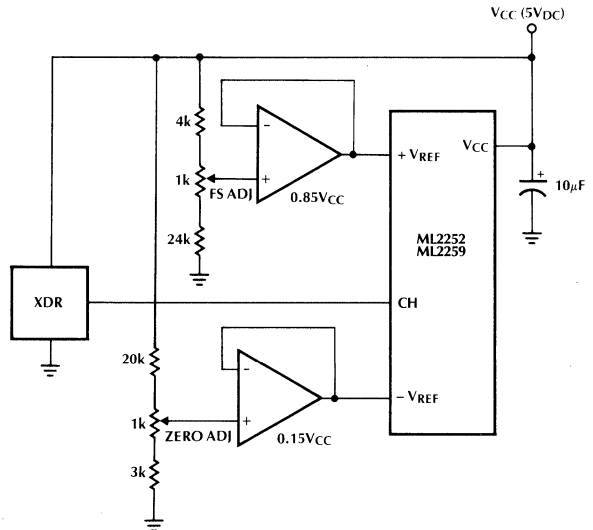


Figure 9. Operating with Ratiometric Transducers 15% of $V_{CC} \le V_{XDR} \le 85\%$ of V_{CC}

ML2252, ML2259

ORDERING INFORMATION

PART NUMBER	TOTAL UNADJUSTED ERROR	TEMPERATURE RANGE	PACKAGE
TWO ANALOG INPUTS, 20-PIN PACKAGE			
ML2252BMJ ML2252BIJ ML2252BCP ML2252BCQ ML2252CIJ ML2252CCP ML2252CCQ	$\pm 1/2$ LSB ± 1 LSB	-55°C to $+125^{\circ}\text{C}$ -40°C to $+85^{\circ}\text{C}$ 0°C to $+70^{\circ}\text{C}$ 0°C to $+70^{\circ}\text{C}$ -40°C to $+85^{\circ}\text{C}$ 0°C to $+70^{\circ}\text{C}$ 0°C to $+70^{\circ}\text{C}$	HERMETIC DIP (J20) HERMETIC DIP (J20) MOLDED DIP (P20) MOLDED PCC (Q20) HERMETIC DIP (J20) MOLDED DIP (P20) MOLDED PCC (Q20)
EIGHT ANALOG INPUTS, 28-PIN PACKAGE			
ML2259BMJ ML2259BIJ ML2259BCP ML2259BCQ ML2259CIJ ML2259CCP ML2259CCQ	$\pm 1/2$ LSB ± 1 LSB	-55°C to $+125^{\circ}\text{C}$ -40°C to $+85^{\circ}\text{C}$ 0°C to $+70^{\circ}\text{C}$ 0°C to $+70^{\circ}\text{C}$ -40°C to $+85^{\circ}\text{C}$ 0°C to $+70^{\circ}\text{C}$ 0°C to $+70^{\circ}\text{C}$	HERMETIC DIP (J28) HERMETIC DIP (J28) MOLDED DIP (P28) MOLDED PCC (Q28) HERMETIC DIP (J28) MOLDED DIP (P28) MOLDED PCC (Q28)

μP Compatible 8-Bit A/D Converter with 8-Channel Multiplexer

GENERAL DESCRIPTION

The ML2258 combines an 8-bit A/D converter, 8-channel analog multiplexer, and a microprocessor compatible 8-bit parallel interface and control logic in a single monolithic CMOS device.

Easy interface to microprocessors is provided by the latched and decoded multiplexer address inputs and latched three-state outputs.

The device is suitable for a wide range of applications from process and machine control to consumer, automotive, and telecommunication applications.

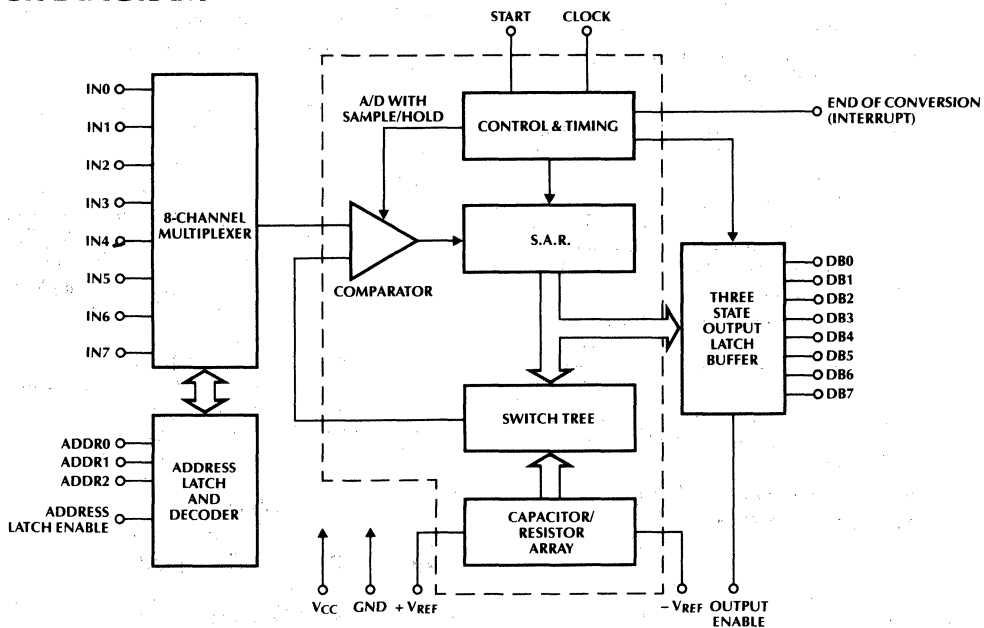
The ML2258 is an enhanced pin compatible second source for the industry standard ADC0808/ADC0809. The ML2258 enhancements are faster conversion time, true sample and hold function, superior power supply rejection, wider reference range, and a double buffered data bus as well as faster digital timing. All parameters are guaranteed over temperature with a power supply voltage of $5V \pm 10\%$.

FEATURES

- Conversion time 6.6μs
- Total unadjusted error $\pm 1/2\text{LSB}$ or $\pm 1\text{LSB}$
- No missing codes
- Sample and hold 390ns acquisition
- Capable of digitizing a 5V, 50kHz sine wave
- 8-input multiplexer
- 0V to 5V analog input range with single 5V power supply
- Operates ratiometrically or with up to 5V voltage reference
- No zero or full scale adjust required
- Analog input protection 25mA per input min
- Low power dissipation 3mA max
- TTL and CMOS compatible digital inputs and outputs
- Standard 28-pin DIP or surface mount PCC
- Superior pin compatible replacement for ADC0808 and ADC0809

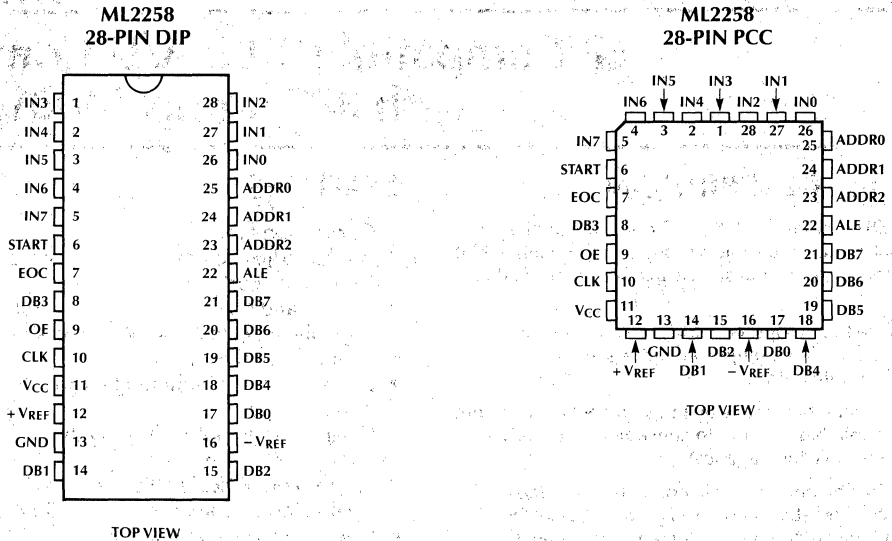
2

BLOCK DIAGRAM



ML2258

PIN CONNECTIONS



PIN DESCRIPTION

PIN NO.	NAME	FUNCTION	PIN NO.	NAME	FUNCTION
1	IN3	Analog input 3.	13	GND	Ground. 0V, all analog and digital inputs or outputs are reference to this point.
2	IN4	Analog input 4.	14	DB1	Data output 1.
3	IN5	Analog input 5.	15	DB2	Data output 2.
4	IN6	Analog input 6.	16	-VREF	Negative reference voltage.
5	IN7	Analog input 7.	17	DB0	Data output 0.
6	START	Start of conversion. Active high digital input pulse initiates conversion.	18	DB4	Data output 4.
7	EOC	End of conversion. This output goes low after a START pulse occurs, stays low for the entire A/D conversion, and goes high after conversion is completed. Data on DB0-DB7 is valid on rising edge of EOC and stays valid until next EOC rising edge.	19	DB5	Data output 5.
8	DB3	Data output 3.	20	DB6	Data output 6.
9	OE	Output enable input. When OE = 0, DB0-DB7 are in high impedance state; OE = 1, DB0-DB7 are active outputs.	21	DB7	Data output 7.
10	CLK	Clock. Clock input provides timing for A/D converter, S/H, and digital interface.	22	ALE	Address latch enable. Input to latch in the digital address (ADDR2-0) on the rising edge of the multiplexer.
11	VCC	Positive supply. 5V ± 10%.	23	ADDR0	Address input 0 to multiplexer. Digital input for selecting analog input.
12	+VREF	Positive reference voltage.	24	ADDR1	Address input 1 to multiplexer. Digital input for selecting analog input.
			25	ADDR2	Address input 2 to multiplexer. Digital input for selecting analog input.
			26	IN0	Analog input 0.
			27	IN1	Analog input 1.
			28	IN2	Analog input 2.

ABSOLUTE MAXIMUM RATINGS

(Note 1)

Supply Voltage, V_{CC}	6.5V
Voltage	
Logic Inputs	-0.3V to $V_{CC} + 0.3V$
Analog Inputs	-0.3V to $V_{CC} + 0.3V$
Input Current per Pin (Note 2)	$\pm 25mA$
Storage Temperature	-65°C to +150°C
Package Dissipation	
at $T_A = 25^\circ C$ (Board Mount)	875mW
Lead Temperature (Soldering 10 sec.)	
Dual-In-Line Package (Plastic)	260°C
Dual-In-Line Package (Ceramic)	300°C
Molded Chip Carrier Package	
Vapor Phase (60 sec.)	215°C
Infrared (15 sec.)	220°C

OPERATING CONDITIONS

Supply Voltage, V_{CC}	4.5V _{DC} to 6.3V _{DC}
Temperature Range (Note 3)	$T_{MIN} \leq T_A \leq T_{MAX}$
ML2258BMJ, ML2258CMJ	-55°C to +125°C
ML2258BIJ, ML2258BIP	
ML2258BIQ, ML2258CIJ	
ML2258CIP, ML2258CIQ	-40°C to +85°C

ELECTRICAL CHARACTERISTICS

Unless otherwise specified, $T_A = T_{MIN}$ to T_{MAX} , $V_{CC} = +V_{REF} = 5V \pm 10\%$, $-V_{REF} = GND$ and $f_{CLK} = 10.24MHz$

PARAMETER	NOTES	CONDITIONS	ML2258B			ML2258C			UNITS
			MIN	TYP (NOTE 4)	MAX	MIN	TYP (NOTE 4)	MAX	
Converter and Multiplexer									
Total Unadjusted Error	5,7	$V_{REF} = V_{CC}$			$\pm 1/2$			± 1	LSB
+ V_{REF} Voltage Range	6		$-V_{REF}$		$V_{CC} + 0.1$	$-V_{REF}$		$V_{CC} + 0.1$	V
- V_{REF} Voltage Range	6		GND-0.1		+ V_{REF}	GND-0.1		+ V_{REF}	V
Reference Input Resistance	5		14	20	35	14	20	28	k Ω
Analog Input Range	5,8		GND-0.1		$V_{CC} + 0.1$	GND-0.1		$V_{CC} + 0.1$	V
Power Supply Sensitivity	6	DC $V_{CC} = 5V \pm 10\%$		$\pm 1/32$	$\pm 1/4$		$\pm 1/32$	$\pm 1/4$	LSB
		100mVp-p 100kHz Sine on V_{CC} , $V_{IN} = 0$		$\pm 1/16$			$\pm 1/16$		LSB
I_{OFF} , Off Channel Leakage Current (Note 9)	5,9	On Channel = V_{CC} Off Channel = 0V	-1			-1			μA
		On Channel = 0V Off Channel = V_{CC}			1			1	μA
I_{ON} , On Channel Leakage Current (Note 9)	5,9	On Channel = 0V Off Channel = V_{CC}	-1			-1			μA
		On Channel = V_{CC} Off Channel = 0V			1			1	μA
Digital and DC									
$V_{IN(1)}$, Logical "1" Input Voltage	5		2.0			2.0			V
$V_{IN(0)}$, Logical "0" Input Voltage	5				0.8			0.8	V
$I_{IN(1)}$, Logical "1" Input Current	5	$V_{IN} = V_{CC}$			1			1	μA
$I_{IN(0)}$, Logical "0" Input Current	5	$V_{IN} = 0V$	-1			-1			μA
$V_{OUT(1)}$, Logical "1" Output Voltage	5	$I_{OUT} = -2mA$	4.0			4.0			V
$V_{OUT(0)}$, Logical "0" Output Voltage	5	$I_{OUT} = 2mA$			0.4			0.4	V
I_{OUT} , Three-State Output Current	5	$V_{OUT} = 0V$	-1			-1			μA
		$V_{OUT} = V_{CC}$			1			1	μA
I_{CC} , Supply Current	5			1.5	3		1.5	3	mA

ELECTRICAL CHARACTERISTICS (Continued)

SYMBOL	PARAMETER	NOTES	CONDITIONS	MIN	TYP (NOTE 4)	MAX	UNITS
AC and Dynamic Performance Characteristics (Note 10)							
t_{ACQ}	Sample and Hold Acquisition				4		$1/f_{CLK}$
f_{CLK}	Clock Frequency	5		100		10240	kHz
t_C	Conversion Time	5			67	67 + 250ns	$1/f_{CLK}$
SNR	Signal to Noise Ratio		$V_{IN} = 51\text{kHz}$, 5V Sine. $f_{CLK} = 10.24\text{MHz}$ ($f_{SAMPLING} \cong 150\text{kHz}$). Noise is Sum of All Nonfundamental Components up to $1/2$ of $f_{SAMPLING}$		47		dB
THD	Total Harmonic Distortion		$V_{IN} = 51\text{kHz}$, 5V Sine. $f_{CLK} = 10.24\text{MHz}$ ($f_{SAMPLING} \cong 150\text{kHz}$). THD is Sum of 2, 3, 4, 5 Harmonics Relative to Fundamental		-60		dB
IMD	Intermodulation Distortion		$V_{IN} = f_A + f_B$, $f_A = 49\text{kHz}$, 2.5V Sine. $f_B = 47.8\text{kHz}$, 2.5V Sine, $f_{CLK} = 10.24\text{MHz}$ ($f_{SAMPLING} \cong 150\text{kHz}$). IMD is ($f_A + f_B$), ($f_A - f_B$), ($2f_A + f_B$), ($2f_A - f_B$), ($f_A + 2f_B$), ($f_A - 2f_B$) Relative to Fundamental		-60		dB
FR	Frequency Response		$V_{IN} = 0$ to 50kHz. 5V Sine Relative to 1kHz		0.1		dB
t_{DC}	Clock Duty Cycle	6, 11		40		60	%
t_{EOC}	End of Conversion Delay	5			8	8 + 250ns	$1/f_{CLK}$
t_{WS}	Start Pulse Width	5		50			ns
t_{SS}	Start Pulse Setup Time	6, 12	Synchronous Only	40			ns
t_{WALE}	Address Latch Enable Pulse Width	5		50			ns
t_S	Address Setup	5		0			ns
t_H	Address Hold	5		50			ns
$t_{HI, HO}$	Output Enable for DB0-DB7	6	Figure 1, $C_L = 50\text{pF}$			100	ns
		6	Figure 1, $C_L = 10\text{pF}$			50	ns
$t_{IH, OH}$	Output Disable for DB0-DB7	6	Figure 1, $C_L = 50\text{pF}$			200	ns
		6	Figure 1, $C_L = 10\text{pF}$			100	ns
C_{IN}	Capacitance of Logic Input				5		pF
C_{OUT}	Capacitance of Logic Outputs				10		pF

Note 1: Absolute maximum ratings are limits beyond which the life of the integrated circuit may be impaired. All voltages unless otherwise specified are measured with respect to ground.

Note 2: When the input voltage (V_{IN}) at any pin exceeds the power supply rails ($V_{IN} < V^-$ or $V_{IN} > V^+$) the absolute value of current at that pin should be limited to 25mA or less.

Note 3: -55°C to +125°C operating temperature range devices are 100% tested at temperature extremes with worst-case test conditions. -40°C to +85°C operating temperature range devices are 100% tested with temperature limits guaranteed by 100% testing, sampling, or by correlation with worst-case test conditions.

Note 4: Typicals are parametric norm at 25°C.

Note 5: Parameter guaranteed and 100% production tested.

Note 6: Parameter guaranteed. Parameters not 100% tested are not in outgoing quality level calculation.

Note 7: Total unadjusted error includes offset, full scale, linearity, multiplexer and sample and hold errors.

Note 8: For $-V_{REF} \geq V_{IN}(+)$ the digital output code will be 0000 0000. Two on-chip diodes are tied to each analog input which will forward conduct for analog input voltages one diode drop below ground or one diode drop greater than the V_{CC} supply. Be careful, during testing at low V_{CC} levels (4.5V), as high level analog inputs (5V) can cause this input diode to conduct—especially at elevated temperatures, and cause errors for analog inputs near full scale. The spec allows 100mV forward bias of either diode. This means that as long as the analog V_{IN} or V_{REF} does not exceed the supply voltage by more than 100mV, the output code will be correct. To achieve an absolute $0V_{DC}$ to $5V_{DC}$ input voltage range will therefore require a minimum supply voltage of $4.900V_{DC}$ over temperature variations, initial tolerance and loading.

Note 9: Leakage current is measured with the clock not switching.

Note 10: $C_L = 50\text{pF}$, timing measured at 50% point.

Note 11: A 40% to 60% clock duty cycle range insures proper operation at all clock frequencies. In the case that an available clock has a duty cycle outside of these limits, the minimum time the clock is high or the minimum time the clock is low must be at least 40ns. The maximum time the clock can be high or low is 60 μ s.

Note 12: The conversion start setup time requirement only needs to be satisfied if a conversion must be synchronized to a given clock rising edge. If the setup time is not met, start conversion will have an uncertainty of one clock pulse.

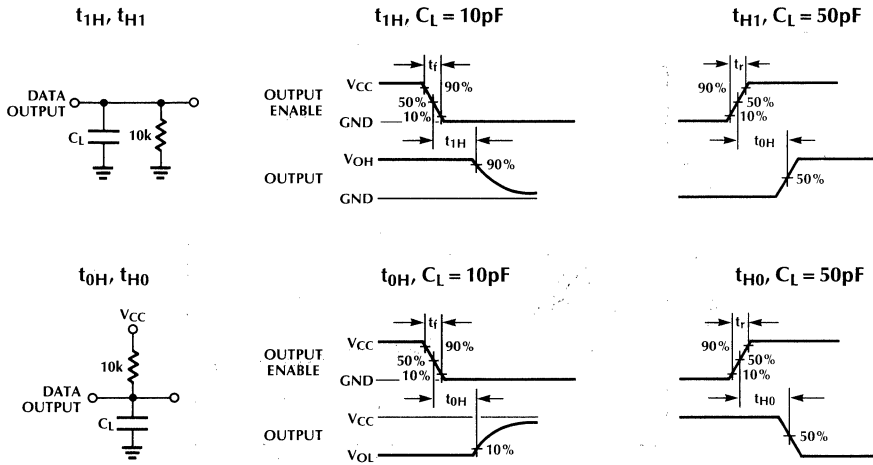


Figure 1. High Impedance Test Circuits and Waveforms

TYPICAL PERFORMANCE CURVES

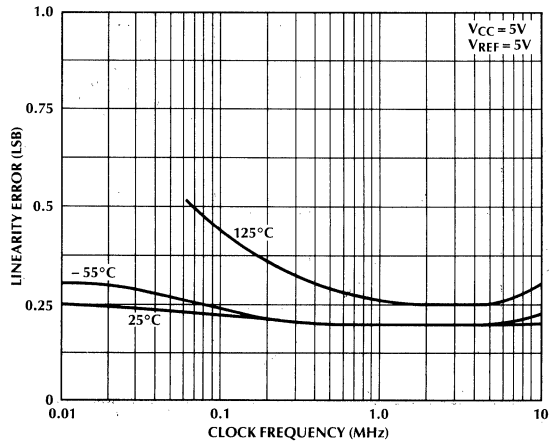


Figure 2. Linearity Error vs f_{CLK}

TYPICAL PERFORMANCE CURVES (Continued)

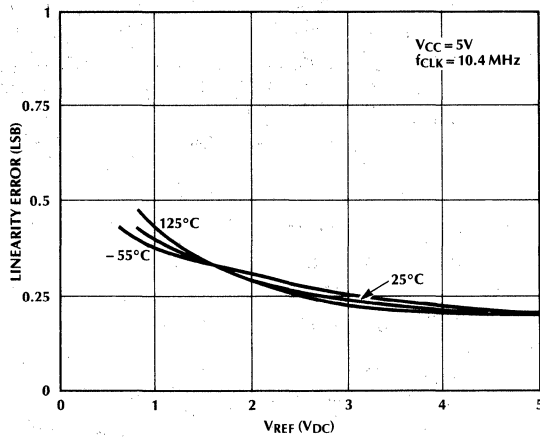


Figure 3. Linearity Error vs VREF Voltage

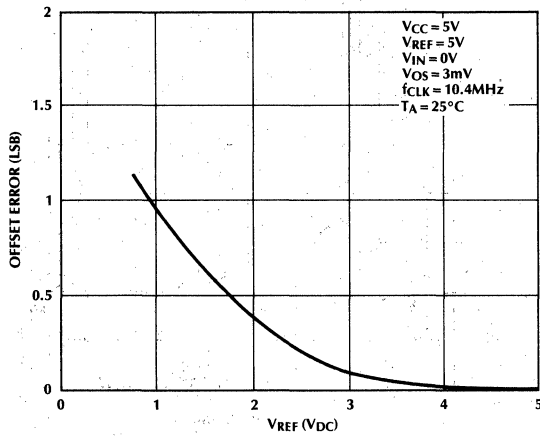


Figure 4. Unadjusted Offset Error vs VREF Voltage

1.0 FUNCTIONAL DESCRIPTION

1.1 MULTIPLEXER ADDRESSING

The ML2258 contains an 8-channel single ended analog multiplexer. A particular input channel is selected by using the address decoder. The relationship between the address inputs, ADDR0–ADDR2, and the analog input selected is shown in Table 1. The address inputs are latched into the decoder on the rising edge of the address latch signal ALE.

Table 1. Multiplexer Address Decoding

SELECTED ANALOG CHANNEL	ADDRESS INPUT		
	ADDR2	ADDR1	ADDR0
IN0	0	0	0
IN1	0	0	1
IN2	0	1	0
IN3	0	1	1
IN4	1	0	0
IN5	1	0	1
IN6	1	1	0
IN7	1	1	1

1.2 A/D CONVERTER

The A/D converter uses successive approximation to perform the conversion. The converter is composed of the successive approximation register, the DAC and the comparator.

The DAC generates the precise levels that determine the linearity and accuracy of the conversion. The DAC is composed of a capacitor upper array and a resistor lower array. The capacitor upper array generates the 4 MSB decision levels while the series resistor lower array generates the 4 LSB decision levels. A switch decoder tree is used to decode the proper level from both arrays.

The capacitor/resistor array offers fast conversion, superior linearity and accuracy since matching is only required between $2^4 = 16$ elements (as opposed to $2^8 = 256$ elements in conventional designs). And since the levels are based on the ratio of capacitors to capacitors and resistors to resistors, the accuracy and long term stability of the converter

is improved. This also guarantees monotonicity and no missing codes, as well as eliminating any linearity temperature or power supply dependence.

The successive approximation register is a digital block used to store the bit decisions from the conversion.

The comparator design is unique in that it is fully differential and auto zeroed. The fully differential architecture provides excellent noise immunity, excellent power supply rejection, and wide common mode range. The comparator is auto zeroed at the start of each conversion in order to remove any DC offset and full scale gain error, thus improving accuracy and linearity.

Another advantage of the capacitor array approach used in the ML2258 over conventional designs is the inherent sample and hold function. This true S/H allows an accurate conversion to be done on the input even if the analog signal is not stable. Linearity and accuracy are maintained for analog signals up to 1/2 the sampling frequency. As a result, input signals up to 75kHz can be converted without degradation in linearity or accuracy.

The sequence of events during a conversion is shown in Figure 5. The rising edge of a START pulse resets the internal registers and the falling edge initiates a conversion on the next rising edge of CLK. Four CLK pulses later, sampling of the analog input begins. The input is then sampled for the next four CLK periods until EOC goes low. EOC goes low on the rising edge of the 8th CLK pulse indicating that the conversion is now beginning. The actual conversion now takes place for the next 56 CLK pulses, one bit for each 7 CLK pulses. After the conversion is done, the data is updated on DB0–DB7 and EOC goes high on the rising edge of the 67th CLK pulse, indicating that the conversion has been completed and data is valid on DB0–DB7. The data will stay valid on DB0–DB7 until the next conversion updates the data word on the next rising edge of EOC.

A conversion can be interrupted and restarted at any time by a new START pulse.

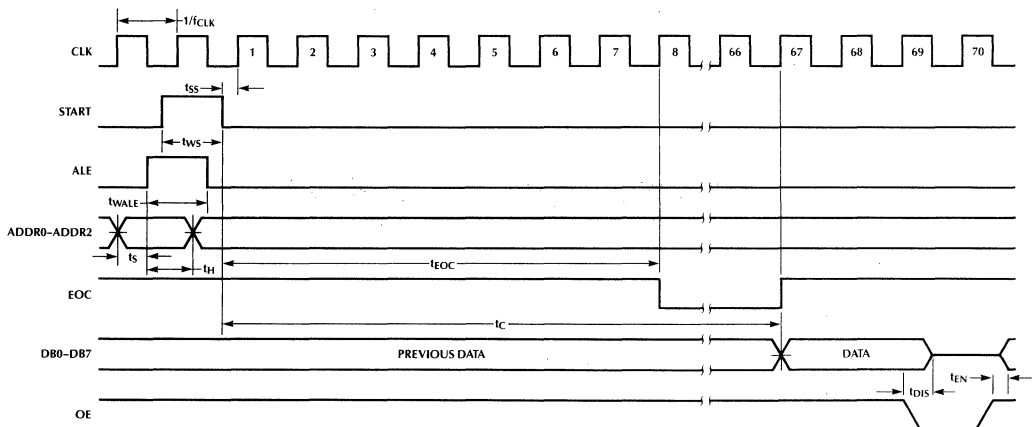


Figure 5. Timing Diagram

1.0 FUNCTIONAL DESCRIPTION (Continued)

1.3 ANALOG INPUTS AND SAMPLE/HOLD

The ML2258 has a true sample and hold circuit which samples both the selected input and ground simultaneously. This simultaneous sampling with a true S/H will give common mode rejection and AC linearity performance that is superior to devices where the two input terminals are not sampled at the same instant and where true sample and hold capability does not exist. Thus, the ML2258 can reject AC common mode signals from DC–50kHz as well as maintain linearity for signals from DC–50kHz.

The plot below (Figure 6) shows a 2048 point FFT of the ML2258 converting a 50kHz, 0 to 5V, low distortion sine wave input. The ML2258 samples and digitizes, at its specified accuracy, dynamic input signals with frequency components up to the Nyquist frequency (one-half the sampling rate). The output spectra yields precise measurements of input signal level, harmonic components, and signal to noise ratio up to the 8-bit level. The near ideal signal to noise ratio is maintained independent of increasing analog input frequencies to 50kHz.

The signal at the analog input is sampled during the interval when the sampling switch is open prior to conversion start. The sampling window (S/H acquisition time) is 4 CLK periods long and occurs 4 CLK periods after START goes low. When the sampling switch closes at the start of the S/H acquisition time, 8pF of capacitance is thrown onto the analog input. 4 CLK periods later, the sampling switch opens, the signal present at analog input is stored and conversion starts. Since any error on the analog input at the end of the S/H acquisition time will cause additional conversion error, care should be taken to insure adequate settling and charging time from the source. If more charging or settling time is needed to reduce these analog input errors, a longer CLK period can be used.

The ML2258 has improved latchup immunity. Each analog input has dual diodes to the supply rails, and a minimum of $\pm 25\text{mA}$ ($\pm 100\text{mA}$ typically) can be injected into each analog input without causing latchup.

1.4 REFERENCE

The voltage applied to the $+V_{\text{REF}}$ and $-V_{\text{REF}}$ inputs defines the voltage span of the analog input (the difference between V_{INMAX} and V_{INMIN}) over which the 256 possible output codes apply. The devices can be used in either ratiometric applications or in systems requiring absolute accuracy. The reference pins must be connected to a voltage source capable of driving the reference input resistance, typically 20k.

In a ratiometric system, the analog input voltage is proportional to the voltage used for the A/D reference. This voltage is typically the system power supply, so the $+V_{\text{REF}}$ pin can be tied to V_{CC} and $-V_{\text{REF}}$ tied to GND. This technique relaxes the stability requirements of the system reference as the analog input and A/D reference move together maintaining the same output code for a given input condition.

For absolute accuracy, where the analog input varies between specific voltage limits, the reference pins can be biased with a time and temperature stable voltage source.

In contrast to the ADC0808 and ADC0809, the ML2258 $-V_{\text{REF}}$ and $+V_{\text{REF}}$ reference values do not have to be symmetric around one half of the supply. $+V_{\text{REF}}$ and $-V_{\text{REF}}$ can be at any voltage between V_{CC} and GND. In addition, the difference between $+V_{\text{REF}}$ and $-V_{\text{REF}}$ can be set to small values for conversions over smaller voltage ranges. Particular care must be taken with regard to noise pickup, circuit layout and system error voltage sources when operating with a reduced span due to the increased sensitivity of the converter.

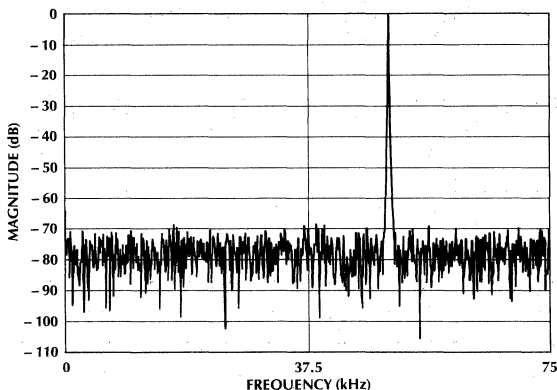


Figure 6. Output Spectrum

1.0 FUNCTIONAL DESCRIPTION (Continued)

1.5 POWER SUPPLY AND REFERENCE DECOUPLING

A 10 μ F electrolytic capacitor is recommended to bypass V_{CC} to GND, using as short a lead length as possible. In addition, with clock frequencies above 1MHz, a 0.1 μ F ceramic disc capacitor should be used to bypass V_{CC} to GND.

If REF+ and REF- inputs are driven by long lines, they should be bypassed by 0.1 μ F ceramic disc capacitors at the reference input pins (pins 12, 16).

1.6 DYNAMIC PERFORMANCE

Signal-to-Noise Ratio

Signal-to-noise ratio (SNR) is the measured signal to noise at the output of the converter. The signal is the rms magnitude of the fundamental. Noise is the rms sum of all the nonfundamental signals up to half the sampling frequency. SNR is dependent on the number of quantization levels used in the digitization process; the more levels, the smaller the quantization noise. The theoretical SNR for a sine wave is given by

$$\text{SNR} = (6.02N + 1.76)\text{dB}$$

where N is the number of bits. Thus for ideal 8-bit converter, SNR = 49.92dB.

Harmonic Distortion

Harmonic distortion is the ratio of the rms sum of harmonics to the fundamental. Total harmonic distortion (THD) of the ML2258 is defined as

$$20 \log \frac{(V_2^2 + V_3^2 + V_4^2 + V_5^2)^{1/2}}{V_1}$$

where V_1 is the rms amplitude of the fundamental and V_2 , V_3 , V_4 , V_5 are the rms amplitudes of the individual harmonics.

Intermodulation Distortion

With inputs consisting of sine waves at two frequencies, f_A and f_B , any active device with nonlinearities will create distortion products, of order $(m+n)$, at sum and difference frequencies of $mf_A + nf_B$, where $m, n = 0, 1, 2, 3, \dots$. Intermodulation terms are those for which m or n is not equal to zero. The ML2258 (IMD) intermodulation distortion specification includes the second order terms $(f_A + f_B)$ and $(f_A - f_B)$ and the third order terms $(2f_A + f_B)$, $(2f_A - f_B)$, $(f_A + 2f_B)$ and $(f_A - 2f_B)$ only.

1.7 DIGITAL INTERFACE

The analog inputs are selected by the digital addresses, ADDR0-ADDR2, and latched on the rising edge of ALE. This is described in the Multiplexer Addressing section.

A conversion is initiated by the rising edge of a START pulse. As long as this pulse is high, the internal logic is reset.

The sampling interval starts with the 4th CLK rising edge after a START falling edge and ends on the 8th rising edge of CLK, 4 CLK periods later. On the rising edge of the 8th CLK pulse, the conversion starts and EOC goes low.

Each bit conversion in the successive approximation process takes 7 CLK periods. On the rising edge of the 64 CLK pulse, the digital output of the conversion is updated on the outputs DB0-DB7. On the rising edge of the 65th CLK pulse, EOC goes high indicating the conversion is done and data on DB0-DB7 is valid.

One feature of the ML2258 over conventional devices is that the data is double buffered. This means that the outputs DB0-DB7 will stay valid until updated at the end of the next conversion and will not become invalid when the next conversion starts. This facilitates interfacing with external logic of μ P.

The signal OE drives the data bus, DB0-DB7, into the high impedance state when held low. This allows the ML2258 to be tied directly to a μ P system bus without any latches or buffers.

2

ML2258

2.0 TYPICAL APPLICATIONS

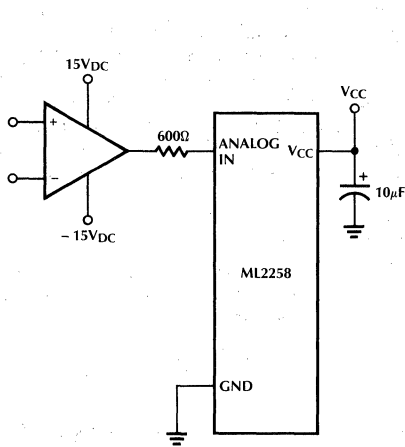


Figure 7. Protecting the Input

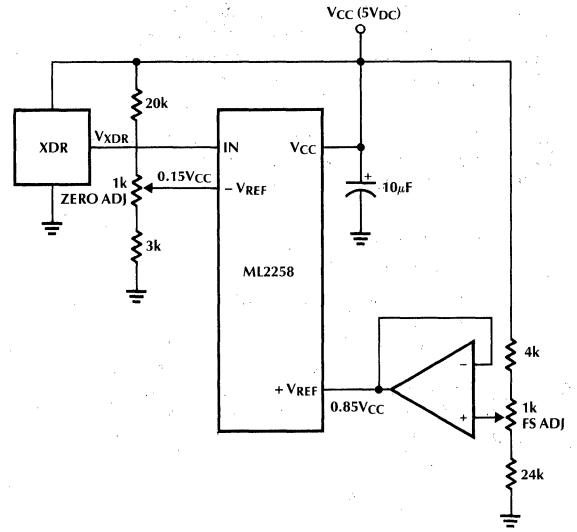


Figure 8. Operating with Ratiometric Transducers 15% of $V_{CC} \leq V_{XDR} \leq 85\%$ of V_{CC}

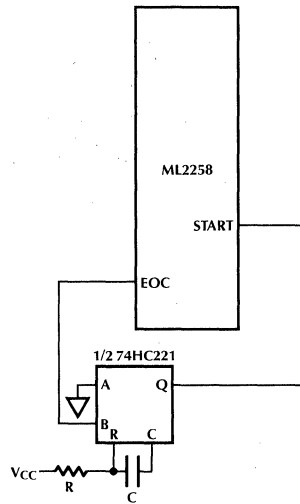


Figure 9. Continuous Conversion Mode

ORDERING INFORMATION

PART NUMBER	ALTERNATE PART NUMBER	TOTAL UNADJUSTED ERROR	TEMPERATURE RANGE	PACKAGE
ML2258BMJ ML2258BIJ ML2258BIP ML2258BIQ	ADC0808CJ ADC0808CCJ ADC0808CCN ADC0808CCV	$\pm 1/2\text{LSB}$	-55°C to +125°C -40°C to +85°C -40°C to +85°C -40°C to +85°C	Hermetic DIP (J20) Hermetic DIP (J20) Molded DIP (P20) Molded PCC (Q20)
ML2258CIJ ML2258CIP ML2258CIQ	ADC0809CCN ADC0809CCV	$\pm 1\text{LSB}$	-40°C to +85°C -40°C to +85°C -40°C to +85°C	Hermetic DIP (J20) Molded DIP (P20) Molded PCC (Q20)

μ P Compatible High-Speed 8-Bit A/D Converter with T/H (S/H)

GENERAL DESCRIPTION

The ML2261 is a high-speed, μ P compatible 8-bit A/D converter with a conversion time of 670ns over the operating temperature range and supply voltage tolerance. The ML2261 operates from a single 5V supply and has an analog input range from GND to V_{CC} .

The ML2261 has two different pin selectable modes. The T/H mode has an internal track and hold. The S/H mode has a true internal sample and hold and can digitize 0 to 5V sinusoidal signals as high as 500kHz. Timing is compatible with the AD7821.

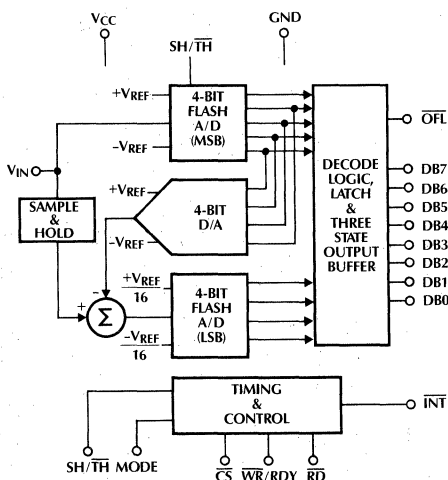
The ML2261 digital interface has been designed so that the device appears as a memory location or I/O port to a μ P.

The ML2261 is an enhanced, pin compatible second source for the industry standard ADC0820 and AD7820. The ML2261 enhancements are faster conversion time, parameters guaranteed over the supply tolerance and temperature range, improved digital interface timing, superior power supply rejection, and better latchup immunity on analog inputs.

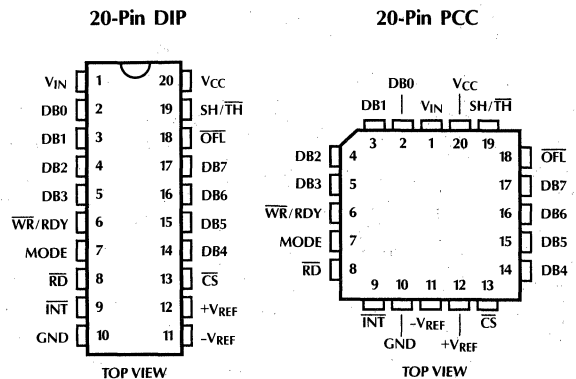
FEATURES

- Conversion time, WR-RD mode over temperature and supply voltage tolerance
 - Track & Hold Mode 850ns max
 - Sample & Hold Mode 700ns max
- Total unadjusted error $\pm 1/2$ LSB or ± 1 LSB
- Digitizes a 5V, 250kHz sine wave to 8-bit accuracy
- No missing codes
- 0V to 5V analog input range with single 5V power supply
- No zero or full scale adjust required
- Analog input protection 25mA min
- Operates ratiometrically or with up to 5V voltage reference
- No external clock required
- Easy interface to μ P, or operates stand alone
- Power-on reset circuitry
- Low power 75mW
- Standard 20-pin DIP or surface mount PCC
- Superior pin compatible replacement for ADC0820 and AD7820

BLOCK DIAGRAM



PIN CONNECTIONS



PIN DESCRIPTION

PIN NO.	NAME	FUNCTION	PIN NO.	NAME	FUNCTION
1	V _{IN}	Analog input.	10	GND	Ground.
2	DB0	Data output — bit 0 (LSB).	11	-V _{REF}	Negative reference voltage for A/D converter.
3	DB1	Data output — bit 1.	12	+V _{REF}	Positive reference voltage for A/D converter.
4	DB2	Data output — bit 2.	13	$\overline{\text{CS}}$	Chip select input. This pin must be held low for the device to perform a conversion.
5	DB3	Data output — bit 3.	14	DB4	Data output — bit 4.
6	$\overline{\text{WR/RDY}}$	Write input or ready output. In WR-RD mode, this pin is WR input. In RD mode, this pin is RDY open drain output. See Digital Interface section.	15	DB5	Data output — bit 5.
7	MODE	Mode select input. MODE = GND: RD mode MODE = V _{CC} : WR-RD mode Pin has internal current source pulldown to GND.	16	DB6	Data output — bit 6.
8	$\overline{\text{RD}}$	Read input. In RD mode, this pin initiates a conversion. In WR-RD mode, this pin latches data into output latches. See Digital Interface section.	17	DB7	Data output — bit 7 (MSB).
9	$\overline{\text{INT}}$	Interrupt output. This output signals the end of a conversion and indicates that data is valid on the data outputs. See Digital Interface section.	18	$\overline{\text{OFL}}$	Overflow output. This output goes low at end of conversion if V _{IN} is greater than +V _{REF} - 1/2LSB.
			19	SH/ $\overline{\text{TH}}$	S/H, T/H mode select. When SH/ $\overline{\text{TH}}$ = V _{CC} the device is in sample and hold mode. When SH/ $\overline{\text{TH}}$ = GND, the device is in track and hold mode. Pin has internal pulldown current source to GND.
			20	V _{CC}	Positive supply. +5 volts ± 5%.

ABSOLUTE MAXIMUM RATINGS

(Note 1)

Supply Voltage, V _{CC}	6.5V
Voltage	
Logic Inputs	-0.3V to V _{CC} + 0.3V
Analog Inputs	-0.3V to V _{CC} + 0.3V
Input Current per Pin (Note 2)	±25mA
Storage Temperature	-65°C to +150°C
Package Dissipation	
at T _A = 25°C (Board Mount)	875mW
Lead Temperature (Soldering 10 sec.)	
Dual-In-Line Package (Plastic)	260°C
Dual-In-Line Package (Ceramic)	300°C
Molded Chip Carrier Package	
Vapor Phase (60 sec.)	215°C
Infrared (15 sec.)	220°C

OPERATING CONDITIONS

Supply Voltage, V _{CC}	4.5V _{DC} to 6.0V _{DC}
Temperature Range (Note 3)	T _{MIN} ≤ T _A ≤ T _{MAX}
ML2261BIJ, ML2261CIJ	-40°C to +85°C
ML2261BCQ, ML2261CCQ	
ML2261BCP, ML2261CCP	0°C to +70°C

ML2261

ELECTRICAL CHARACTERISTICS

Unless otherwise specified, $T_A = T_{MIN}$ to T_{MAX} , $V_{CC} = +V_{REF} = 5V \pm 5\%$, and $-V_{REF} = GND$

PARAMETER	NOTES	CONDITIONS	ML2261XCX			ML2261XIX			UNITS
			MIN	TYP (Note 4)	MAX	MIN	TYP (Note 4)	MAX	
Converter									
Total Unadjusted Error ML2261BXX ML2261CXX	5, 7	$V_{REF} = V_{CC}$			$\pm 1/2$ ± 1			$\pm 1/2$ ± 1	LSB LSB
+ V_{REF} Voltage Range	6		$-V_{REF}$		$V_{CC}+0.1$	$-V_{REF}$		$V_{CC}+0.1$	V
- V_{REF} Voltage Range	6		$GND-0.1$		$+V_{REF}$	$GND-0.1$		$+V_{REF}$	V
Reference Input Resistance	5		1	2	3	1	2	3	k Ω
Analog Input Range	5, 8		$GND-0.1$		$V_{CC}+0.1$	$GND-0.1$		$V_{CC}+0.1$	V
Power Supply Sensitivity	5	DC $V_{CC} = 5V \pm 5\%$, $V_{REF} = 4.75V$		$\pm 1/32$	$\pm 1/4$		$\pm 1/32$	$\pm 1/4$	LSB
		100mVp-p 100kHz sine on V_{CC} . $V_{IN} = 0$		$\pm 1/16$			$\pm 1/16$		LSB
Analog Input Leakage Current	5, 9	Converter Idle	-1		+1	-1		+1	μA
Analog Input Capacitance		During Acquisition Period		45			45		pF
Digital and DC									
$V_{IN(1)}$, Logical "1" Input Voltage	5	\overline{WR} , \overline{RD} , \overline{CS}	2.0			2.0			V
		MODE, SH/ \overline{TH}	$V_{CC}-0.5$			$V_{CC}-0.5$			V
$V_{IN(0)}$, Logical "0" Input Voltage	5	\overline{WR} , \overline{RD} , \overline{CS}			0.8			0.8	V
		MODE, SH/ \overline{TH}			0.5			0.5	V
$I_{IN(1)}$, Logical "1" Input Current	5	$V_{IH} = V_{CC}$	\overline{WR} , \overline{RD} , \overline{CS}			1		1	μA
			MODE, SH/ \overline{TH}	15	50	150	15	50	150
$I_{IN(0)}$, Logical "0" Input Current	5	$V_{IL} = GND$	\overline{WR} , \overline{RD} , \overline{CS}	-1			-1		μA
			MODE, SH/ \overline{TH}	-20			-20		μA
$V_{OUT(1)}$, Logical "1" Output Voltage	5	$I_{OUT} = -2mA$	4.0			4.0			V
$V_{OUT(0)}$, Logical "0" Output Voltage	5	$I_{OUT} = 2mA$			0.4			0.4	V
I_{OUT} , Three-State Output Current	5	$V_{OUT} = 0V$	-1			-1			μA
		$V_{OUT} = V_{CC}$			1			1	μA
C_{OUT} , Logic Output Capacitance				5			5		pF
C_{IN} , Logic Input Capacitance				5			5		pF
I_{CC} Supply Current	5	$\overline{CS} = \overline{WR} = \overline{RD} = "1"$ No Output Load		8	14		8	15.5	mA

ELECTRICAL CHARACTERISTICS (Continued)

Unless otherwise specified, $T_A = T_{MIN}$ to T_{MAX} , $V_{CC} = +V_{REF} = 5V \pm 5\%$, $-V_{REF} = GND$, and timing measured at 1.4V, $C_L = 100pF$.

PARAMETER	NOTES	CONDITIONS	ML2261XCX			ML2261XIX			UNITS
			MIN	TYP (Note 4)	MAX	MIN	TYP (Note 4)	MAX	
AC and Dynamic Performance (Note 9)									
t_{CRD} , Conversion Time, Read Mode	5	RD to INT, MODE = 0V			1060			1100	ns
t_{CWR-RD} , Conversion Time, Write-Read Mode	5, 9	WR Falling Edge to INT, $t_{RD} < t_{INT}$, MODE = V_{CC}	SH/TH = V_{CC}	650	700		690	740	ns
			SH/TH = GND			850		920	ns
SNR, Signal to Noise Ratio		$V_{IN} = 5V$, 250kHz Noise is sum of all nonfundamental components from 0-500kHz. SH/TH = V_{CC} , MODE = V_{CC} $f_{SAMPLING} = 1$ MHz		48			48		dB
HD, Harmonic Distortion		$V_{IN} = 5V$, 250kHz THD is sum of 2-5th harmonics relative to fundamental. SH/TH = V_{CC} , MODE = V_{CC} $f_{SAMPLING} = 1$ MHz		-63			-63		dB
IMD, Intermodulation Distortion		$f_a = 2.5V$, 250kHz $f_b = 2.5V$, 248kHz IMB is $(f_a + f_b)$, $(f_a - f_b)$, $(2f_a + f_b)$, $(2f_a - f_b)$, $(f_a + 2f_b)$, or $(f_a - 2f_b)$ relative to fundamental. SH/TH = V_{CC} , MODE = V_{CC} $f_{SAMPLING} = 1$ MHz		-60			-60		dB
FR, Frequency Response		$V_{IN} = 5V$, 0-250kHz Relative to 1kHz SH/TH = V_{CC} , MODE = V_{CC} $f_{SAMPLING} = 1$ MHz		± 0.1			± 0.1		dB
SR, Slew Rate Tracking	6	SH/TH = V_{CC}			4.0		4.0		V/ μ s
		SH/TH = GND			.25		.25		V/ μ s
AC Performance Read Mode (Pin 7 = 0V), Figure 2									
t_{RDY} , CS to RDY Delay	5		0		65	0		70	ns
t_{RDD} , RD Low to RDY Delay	5, 10	Figure 1			1060			1100	ns
t_{CSS} , CS to RD, WR Setup Time	5		0			0			ns
t_{CSH} , CS to RD, WR Hold Time	5		0			0			ns
t_{CRD} , Conversion Time — RD Low to INT Low	5, 10				1060			1100	ns

ML2261

ELECTRICAL CHARACTERISTICS (Continued)

Unless otherwise specified, $T_A = T_{MIN}$ to T_{MAX} , $V_{CC} = +V_{REF} = 5V \pm 5\%$, $-V_{REF} = GND$, and timing measured at 1.4V, $C_L = 100pF$

PARAMETER	NOTES	CONDITIONS	ML2261XCX			ML2261XIX			UNITS
			MIN	TYP (Note 4)	MAX	MIN	TYP (Note 4)	MAX	
AC Performance Read Mode (Pin 7 = 0V), Figure 2 (Continued)									
t_{ACC0} , Data Access Time RD to Data Valid	5		t_{CRD}		$t_{CRD}+30$	t_{CRD}		$t_{CRD}+30$	ns
t_{RDPW} , RD Pulse Width	5		$t_{CRD}+30$			$t_{CRD}+30$			ns
t_{INTH} , RD to INT Delay	5, 10		0		65	0		70	ns
t_{DH} , Data Hold Time — RD Rising Edge to Data High Impedance State	6, 10	Figure 1	0		50	0		60	ns
t_p , Delay Time Between Conversions — INT Low to RD Low	5, 10	Sample & Hold Mode, SH/TH = V_{CC}	300			325			ns
		Track & Hold Mode, SH/TH = GND	240			260			ns
AC Performance Write-Read Mode (Pin 7 = 5V), Figures 3 and 4									
t_{CSS} , CS to RD, WR Setup Time	5		0			0			ns
t_{CSH} , CS to RD, WR Hold Time	5		0			0			ns
t_{WR} , WR Pulse Width	5	SH/TH = V_{CC}	170		50K	180		50K	ns
	6	SH/TH = GND	320		50K	360		50K	ns
t_{RD} , Read Time — WR High to RD Low Delay	5	$t_{RD} < t_{INTL}$	275			290			ns
t_{RI} , RD to INT Delay	5, 10	$t_{RD} < t_{INTL}$	0		255	0		270	ns
t_{ACL} , Data Access Time — RD Low to Data Valid	5	$t_{RD} < t_{INTL}$	0		260	0		280	ns
$t_{CVR,RD}$, Conversion Time — WR Falling Edge to INT Low	5,9,10	$t_{RD} < t_{INTL}$, SH/TH = V_{CC}		650	700		690	740	ns
	6,9,10	$t_{RD} < t_{INTL}$, SH/TH = GND			850			920	ns
t_{INTL} , Internal Comparison Time — WR Rising Edge to INT Low	5, 10	$t_{RD} > t_{INTL}$			650			670	ns
t_{ACC2} , Data Access Time — RD to Data Valid	5	$t_{RD} > t_{INTL}$	0		50	0		60	ns
t_{DH} , Data Hold Time — RD Rising Edge to Data High Impedance State	6, 10	Figure 1	0		50	0		60	ns
t_{INTH} , RD1 to INT1 Delay	5, 10		0		65	0		70	ns
t_p , Delay Time Between Conversions — INT Low to WR Low	5, 10	Sample & Hold Mode, SH/TH = V_{CC}	300			325			ns
		Track & Hold Mode, SH/TH = GND	240			260			ns
t_{IHDR} , WR1 to INT1 Delay	5, 10	Standalone Mode	0		100	0		110	ns
t_{ID} , INT1 to Data Valid Delay	5, 10	Standalone Mode	0		20	0		30	ns

- Note 1:** Absolute maximum ratings are limits beyond which the life of the integrated circuit may be impaired. All voltages unless otherwise specified are measured with respect to ground.
- Note 2:** When the voltage at any pin exceeds the power supply rails ($V_{IN} < GND$ or $V_{IN} > V_{CC}$) the absolute value of current at that pin should be limited to 25mA or less.
- Note 3:** 0°C to +70°C and -40°C to +85°C operating temperature range devices are 100% tested with temperature limits guaranteed by 100% testing, sampling, or by correlation with worst-case test conditions.
- Note 4:** Typicals are parametric norm at 25°C.
- Note 5:** Parameter guaranteed and 100% production tested.
- Note 6:** Parameter guaranteed. Parameters not 100% tested are not in outgoing quality level calculation.
- Note 7:** Total unadjusted error includes offset, full scale, linearity, and sample and hold errors. Total unadjusted error is tested at the minimum specified times for WR, RD, t_{RI} , and t_p . For example, for the ML2261XCX in the sample and hold mode, WR/RD mode: $t_{WR} = 170ns$, $t_{RD} = 275ns$ with a frequency of 1.00MHz (cycle time of 1000ns).
- Note 8:** For $-V_{REF} \geq V_{IN}$ the digital output code will be 0000 0000. Two on-chip diodes are tied to the analog input which will forward conduct for analog input voltages one diode drop below ground or one diode drop greater than the V_{CC} supply. Be careful, during testing at low V_{CC} levels (4.5V), as high level analog inputs (5V) can cause this input diode to conduct — especially at elevated temperatures, and cause errors for analog inputs near full scale. The spec allows 100mV forward bias of either diode. This means that as long as the analog V_{IN} or V_{REF} does not exceed the supply voltage by more than 100mV, the output code will be correct. To achieve an absolute 0V_{DC} to 5V_{DC} input voltage range will therefore require a minimum supply voltage of 4.900V_{DC} over temperature variations, initial tolerance and loading.
- Note 9:** Conversion time, write-read mode = $t_{WR} + t_{RD} + t_{RI}$.
- Note 10:** Defined from the time an output crosses 0.8V or 2.4V.

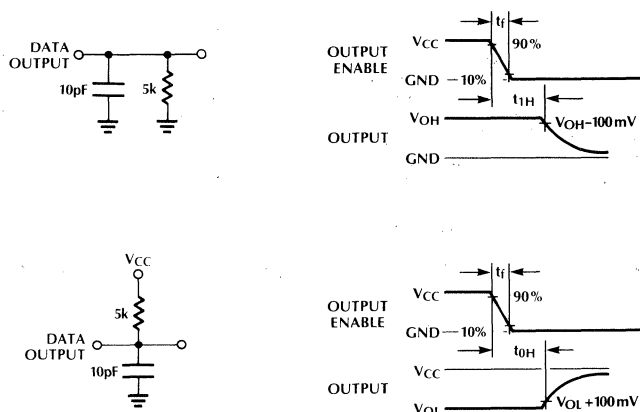


Figure 1. High Impedance Test Circuits and Waveforms

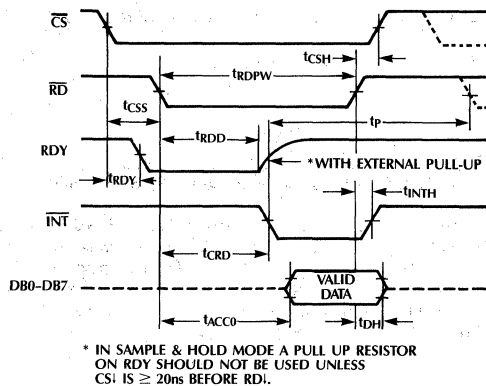


Figure 2. RD Mode Timing

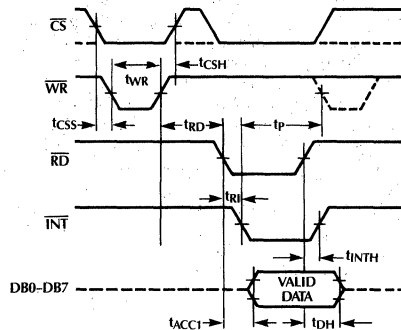


Figure 4. WR-RD Mode Timing ($t_{RD} < t_{INTL}$)

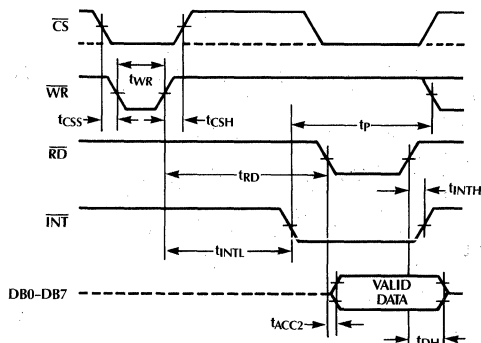


Figure 3. WR-RD Mode Timing ($t_{RD} > t_{INTL}$)

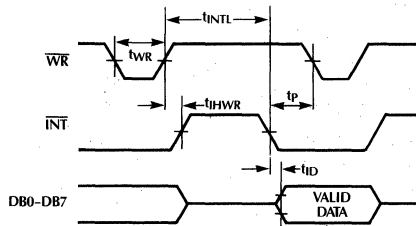


Figure 5. WR-RD Mode Stand-Alone Timing $\overline{CS} = \overline{RD} = 0$

1.0 FUNCTIONAL DESCRIPTION

The ML2261 uses a two stage flash technique for A/D conversion. This technique first performs a 4 bit flash conversion on V_{IN} to determine the 4 MSB's. These 4 MSB's are then cycled through an internal DAC to recreate the analog input. This reconstructed analog input signal from the DAC is then subtracted from the input, and the difference voltage is converted by a second 4 bit flash conversion, providing the 4 LSB's of the output data word. An additional overrange function detects if V_{IN} is greater than $+V_{REF} - \frac{1}{2}LSB$.

1.1 ANALOG INPUT

The analog input on the ML2261 behaves differently than inputs on conventional converters. The analog input current requirements change while the conversion is in progress, and the amount of input current depends on what cycle the converter is in.

The equivalent input circuit for the converter is shown in Figure 6. When the conversion starts in the T/H mode (WR1 in the WR-RD mode or RD1 in the RD mode) S1, S4 and S6 close and S3 opens. This period is known as the acquisition period where the MSB flash converter tracks the input signal and the LSB flash converter samples it. During this period, V_{IN} is connected to the 16 MSB and 15 LSB comparators. In addition, there is a stray capacitance of approximately 11 pF that needs to be charged through the external source resistance R_S . This period ends in the WR-RD mode when WR1 or by an internal timer in the RD mode. At this point S1 and S4 open and the analog input at V_{IN} is no longer being sampled; thus during this time the analog voltage on V_{IN} does not affect converter performance.

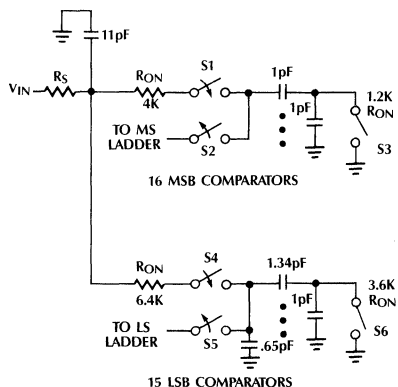


Figure 6. Converter Equivalent Input Circuit

As shown above, the critical period for charging up the analog input occurs when the MSB and LSB comparators are sampling the input, known as the acquisition period. The source of the external signal on V_{IN} must adequately charge up the analog voltage during the acquisition period. To do this, the input must settle within the required analog accuracy tolerance at least 50ns before the end of the acquisition period so that the MSB comparators have adequate time to make the correct decision. If more time is needed due to finite charging or settling time of the external source, the WR low period can be extended in WR-RD mode. In RD mode, since the acquisition time is fixed by internal delays, the burden is on the external source to charge up and settle the input adequately.

When the ML2261 operates in the S/H mode (pin 19 = V_{CC}) both the MSB and the LSB flash converter perform a true sample and hold operation during the acquisition or sampling period. This period starts after the falling edge of INT and ends with the falling edge of WR in the WR-RD mode or the falling edge of RD in the RD mode. The duration of this period is user controlled and must satisfy a minimum of t_p .

During this period S1, S3, S4 and S6 close, therefore 46 pF of input capacitance must be charged up in addition to the 11 pF of stray capacitance.

1.2 TRACK AND HOLD vs. SAMPLE AND HOLD

The MSB Flash Converter of the ML2261 in T/H mode has a track and hold mechanism for sampling the input. The input is attached to the MSB comparators directly in the MSB compare cycle, or acquisition period. When the MSB compare cycle ends, the state of the MSB comparators is latched. The LSB Flash Converter always performs a S/H operation. Thus, the analog input signal can be changing during the MSB compare cycle, or acquisition period, and the MSB comparators will be tracking it as long as the slew rate of the analog input is slow enough so that the MSB comparators can respond. The ML2261 can track and hold signals with slew rates as high as $.25V/\mu s$ (16kHz @ 5 volts) without sacrificing conversion accuracy.

The ML2261 in S/H mode does not have the slew rate limitation of the T/H mode since an internal sample and hold acquires the analog signal, holds it internally, and then performs a conversion. Since this is a true sample and hold function, the S/H mode can theoretically digitize signals of frequencies much higher than the T/H mode. The ML2261 in S/H mode can digitize signals of frequencies as high as 250kHz @ 5V (slew rates as high as $4V/\mu s$) without sacrificing conversion accuracy. In most applications, the S/H mode is more desirable than T/H mode because of the better dynamic performance.

1.2.1 CONVERTER — T/H MODE

The operating sequence for the WR-RD mode is illustrated in Figure 7a. Initially, the internal comparators are auto-zeroed while WR is high. A conversion is initiated by the falling edge of WR. While WR is low, the MSB comparators are tracking the analog input and comparing this voltage against voltages from the internal resistor ladder. At the same time, the input is being acquired or sampled by LSB comparators. On the rising edge of WR, the MSB comparator results are latched, and the LSB acquisition time is ended by closing the sampling switch to the LSB comparators. While WR is high, the LSB comparators then compare the residual input voltage against internal voltages from the resistor ladder to determine the 4 LSB's. When the LSB comparison or conversion is complete, INT goes low and latches the conversion result into the output latches. Then, the comparators are auto-zeroed while WR is high before another conversion can start.

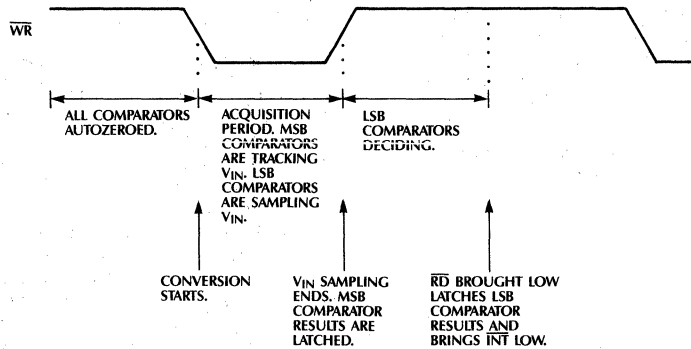
The operating sequence for RD mode, is similar to that described above for the WR-RD mode, except the conversion is initiated by the falling edge of RD, and the MSB and LSB conversions are generated by internal clock edges that are generated while RD is low.

1.2.2 CONVERTER — S/H MODE

The operating sequence for S/H mode is illustrated in Figure 7b. Notice that it is similar to T/H mode described above except this mode has a true sample and hold function. The falling edge of INT closes the sampling switch and starts the acquisition period where the analog input is sampled at the same time all comparators are auto-zeroed. The falling edge of WR opens the internal sampling switch, ends the acquisition period, and starts the conversion on the internally sample and held signal. The MSB comparators make their decisions while WR is low. On the rising edge of WR, the MSB comparator results are latched. The LSB comparators make their decision when WR is high. When the LSB comparison or conversion is complete, INT goes low and latches the conversion result into the output buffers. Then, the acquisition period begins again and the converter is ready for the next conversion.

The operating sequence for the RD mode is the same as the WR-RD mode, except the conversion is initiated by the falling edge of RD, and the MSB and LSB conversions are generated by internal clock edges that are generated while RD is low.

a. T/H Mode



b. S/H Mode

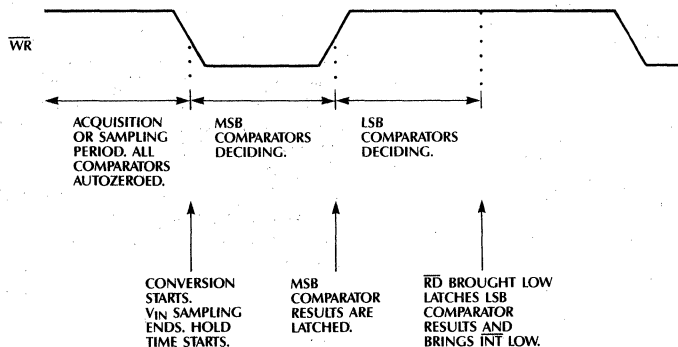


Figure 7. Operating Sequence (WR-RD Mode)

1.3 REFERENCE

The $+V_{REF}$ and $-V_{REF}$ inputs are the reference voltages that determine the full scale and zero input voltages, respectively, for the A/D converter. Thus, $+V_{REF}$ defines the analog input which produces a full scale output and $-V_{REF}$ defines the analog input which produces an output code of all zeroes. The transfer function for the A/D converter is shown in Figure 8.

$+V_{REF}$ and $-V_{REF}$ can be set to any voltage between GND and V_{CC} . This means that the reference voltages can be offset from GND and the difference between $+V_{REF}$ and $-V_{REF}$ can be made small to increase the resolution of the conversion. Note that the total unadjusted error increases when $[+V_{REF} - (-V_{REF})]$ decreases.

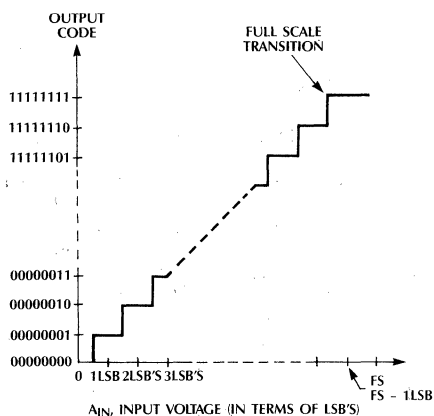


Figure 8. A/D Transfer Characteristic

1.4 POWER SUPPLY AND REFERENCE DECOUPLING

A $0.1\mu\text{F}$ ceramic disc capacitor is recommended to bypass V_{CC} to GND, using as short a lead length as possible.

If REF+ and REF- inputs are driven by long lines, they should be bypassed by $0.1\mu\text{F}$ ceramic disc capacitors at the reference input pins.

1.5 DYNAMIC PERFORMANCE

1.5.1 SINUSOIDAL INPUTS

Since the ML2261 has an internal sample and hold, the device can digitize high frequency sinusoids with little or no signal degradations. Using the Nyquist criteria, the highest frequency input to the converter could theoretically be $1/2$ the sampling rate (f_s). Any frequency components above $f_s/2$ will be aliased below $f_s/2$. In most applications, these aliased components cause unacceptable distortion and must be filtered out of the input. If the input frequency is too close to $f_s/2$, then the requirements on the anti-alias filter become difficult

to impossible to realize with standard component and tolerances. In most practical applications, the highest input frequency has to be limited to $1/3$ to $1/4$ of f_{max} in order to relax the filtering requirements enough to make a realizable anti-alias filter.

The maximum sampling rate (f_{max}) for the ML2261 in the WR-RD mode, ($t_{RD} < t_{INTL}$) can be calculated as follows:

$$f_{max} = \frac{1}{t_{WR} + t_{RD} + t_{RI} + t_p}$$

$$f_{max} = \frac{1}{170\text{ns} + 275\text{ns} + 255\text{ns} + 300\text{ns}}$$

$$f_{max} = 1.00 \text{ MHz}$$

t_{WR} = Write Pulse Width

t_{RD} = Delay Time between \overline{WR} and \overline{RD} Pulses

t_{RI} = \overline{RD} to \overline{INT} Delay

t_p = Delay Time between Conversions

This permits a maximum sampling rate of 1MHz for the ML2261. The dynamic performance specifications (SNR, HD, IMD, and FR) for the ML2261 are all specified at 250kHz, which is approximately $1/4$ of the sampling rate, f_s .

In applications where aliased frequency components are acceptable and filtering of the input signal is not needed or where a filter with a steep amplitude response is available, the user can apply an input sinusoid higher than 250kHz to the device. Note, however, that as the input frequency increases above 500kHz, dynamic performance degradation will occur due to the finite bandwidth of the internal sample and hold.

The Figure 9 plots are 4096 point FFT's of the ML2261 converting a 257kHz and a 491kHz, 0 to 4.5V, low distortion sine wave input. The ML2261 samples and digitizes at its specified accuracy, dynamic input signals with frequency components up to the Nyquist frequency (one-half the sampling rate). The output spectra yields precise measurements of the input signal level, harmonic components, and signal to noise ratio up to the 8-bit level. The near ideal signal to noise ratio is maintained independent of increasing analog input frequencies to 500kHz.

1.5.2 SIGNAL-TO-NOISE RATIO

Signal-to-noise ratio (SNR) is the measured signal to noise at the output of the converter. The signal is the rms magnitude of the fundamental. Noise is the rms sum of all the nonfundamental signals up to half the sampling frequency. SNR is dependent on the number of quantization levels used in the digitization process; the more the levels, the smaller the quantization noise. The theoretical SNR for a sine wave is given by

$$\text{SNR} = (6.02N + 1.76) \text{ dB}$$

where N is the number of bits. Thus for ideal 8-bit converter, SNR = 49.92 dB.

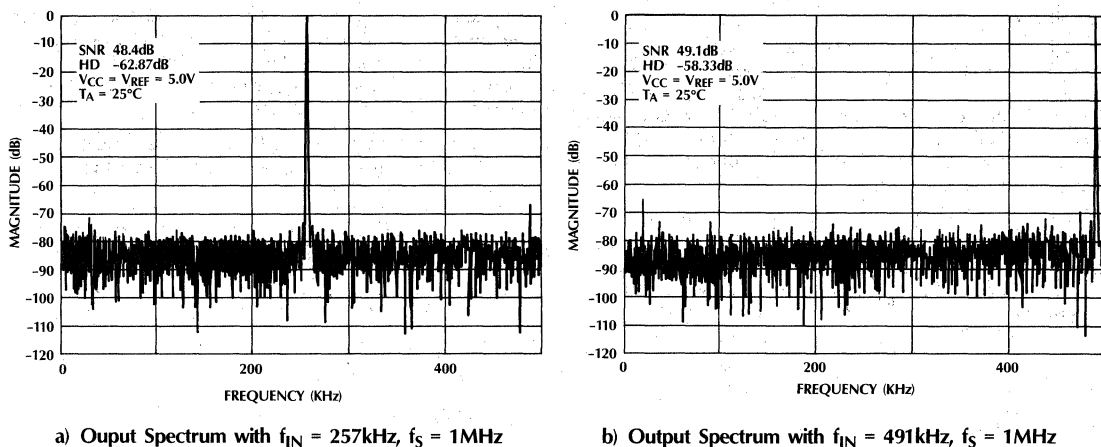


Figure 9. Dynamic Performance, Sample and Hold Mode

1.5.3 HARMONIC DISTORTION

Harmonic distortion is the ratio of the rms sum of harmonics to the fundamental. Total harmonic distortion (THD) of the ML2261 is defined as

$$20 \log = \frac{(V_2^2 + V_3^2 + V_4^2 + V_5^2)^{1/2}}{V_1}$$

where V_1 is the rms amplitude of the fundamental and V_2, V_3, V_4, V_5 are the rms amplitudes of the individual harmonics.

1.5.4 INTERMODULATION DISTORTION

With inputs consisting of sine waves at two frequencies, f_A and f_B , any active device with nonlinearities will create distortion products, of order $(m + n)$, at sum and difference frequencies of $mf_A + nf_B$, where $m, n = 0, 1, 2, 3 \dots$. Intermodulation terms are those for which m or n is not equal to zero. The (IMD) intermodulation distortion specification includes the second order terms $(f_A + f_B)$ and $(f_A - f_B)$ and the third order terms $(2f_A + f_B)$, $(2f_A - f_B)$, $(f_A + 2f_B)$, and $(f_A - 2f_B)$ only.

1.6 DIGITAL INTERFACE

The ML2261 has two basic interface modes, RD and WR-RD, which are selected by the MODE input pin.

1.6.1 RD MODE

In the RD mode, the $\overline{\text{WR}}/\text{RDY}$ pin is configured as the RDY output. The read mode performs a conversion with a single RD pulse. This allows the μP to start a conversion, wait, and then read data with a single read instruction.

The timing for the RD mode is shown in Figure 2. To do a conversion, CS must be low to select the device. After CS goes low, the RDY output goes low indicating that the device is ready to do a conversion. The conversion starts on the falling edge of RD. While RD is low, the MSB and LSB decisions are made with internally generated clock edges. When the conversion is complete, RDY goes high and $\overline{\text{INT}}$ goes low signaling the end of the conversion. After $\overline{\text{INT}}$ goes low, the data outputs go from high impedance to active state with valid output data. Data stays valid until either RD or CS goes high. When either signal goes high, the output data lines return to the high impedance state and $\overline{\text{INT}}$ returns high. A pull up resistor on RDY in the sample and hold mode will cause clock injection, degrading the total unadjusted error, unless CS \dagger is $\geq 20\text{ns}$ before RD \dagger .

1.6.2 WR-RD MODE

In the WR-RD mode, the $\overline{\text{WR}}/\text{RDY}$ pin is configured as the WR input. In this mode, WR initiates the conversion and RD controls reading the output data. This can be done in several ways, described below.

1.6.3 WR-RD MODE — USING INTERNAL DELAY ($t_{RD} > t_{INTL}$)

The timing is shown in Figure 3. To do a conversion, CS must be low to select the device. Then, WR falling edge triggers the conversion. While WR is low, the MSB comparison is made. When WR returns high the LSB decision is made. After some internal delay, $\overline{\text{INT}}$ goes low indicating end of conversion. Valid data will appear on DB0-7 when RD is pulled low. $\overline{\text{INT}}$ is then reset by the rising edge of either CS or RD.

1.6.4 WR-RD MODE — READING BEFORE DELAY ($t_{RD} < t_{INTL}$)

The internally generated delay for the LSB decision when $t_{RD} > t_{INTL}$ is longer than necessary due to circuit design tolerances of t_{INTL} delay. If desired, a faster conversion will result without loss of accuracy by bringing RD low within the minimum time specified for t_{RD} . The timing diagram for this mode is shown in Figure 4. WR is the same as when $t_{RD} > t_{INTL}$. But in this case, RD is brought low t_{RD} ns after WR rising edge and before INT. INT goes low indicating an end of conversion after the falling edge of RD and is reset on the rising edge of RD or CS. When RD is brought low before INT goes low the data bus always remains in the high-impedance state until INT.

1.6.5 WR-RD MODE — STAND ALONE OPERATION

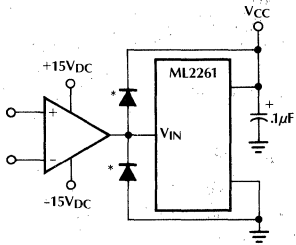
Stand alone operation can be implemented by tying CS and RD low as shown in Figure 5. WR initiates a conversion as before. When WR is low, the MSB comparison is made. When WR goes high, the LSB comparison is made. Since RD is already low, the output data will appear automatically at end of conversion. Since RD is always low, INT is reset on rising edge of WR and goes low at end of conversion.

1.6.6 POWER-ON RESET

When power is first applied, an internal power-on reset and timer circuit inhibits the CS input and resets the internal circuitry to prevent the ML2261 from starting in an unknown state. During this period of approximately $3\mu s$, INT remains high and the data bus is in the high-impedance state.

2

2.0 TYPICAL APPLICATIONS



* NO PROTECTION IS REQUIRED IF INPUT CURRENT < 25ma

Figure 10. Protecting the Input

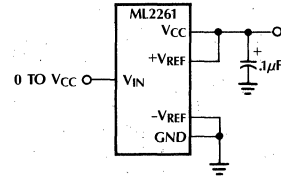


Figure 11. Using VCC as Reference for Ratiometric Operation

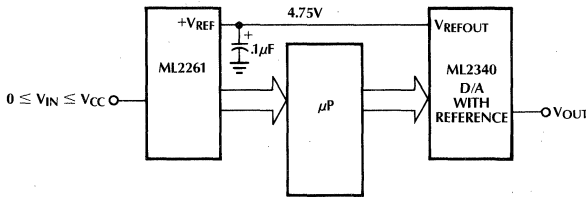


Figure 12. Using External Reference of D/A

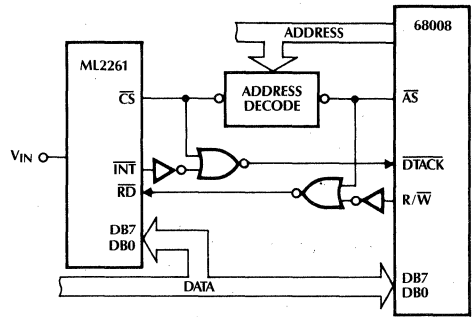


Figure 13. 68000 Type Interface to ML2261

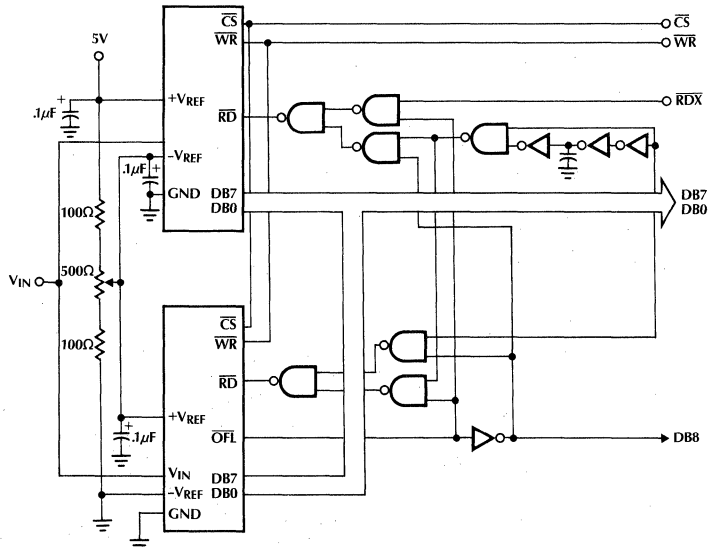


Figure 14. 9-Bit Resolution

2.0 TYPICAL APPLICATIONS (Continued)

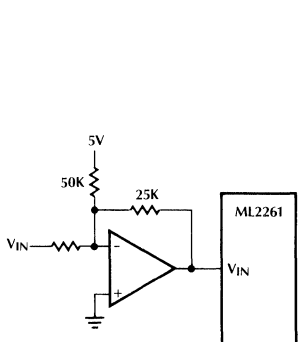


Figure 15. ±2.5V Analog Input Range

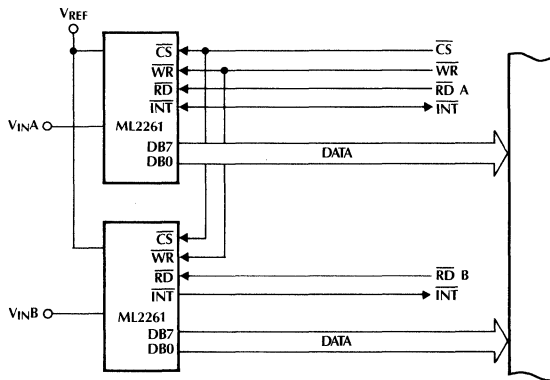


Figure 16. Simultaneous Sampling of Two Variables

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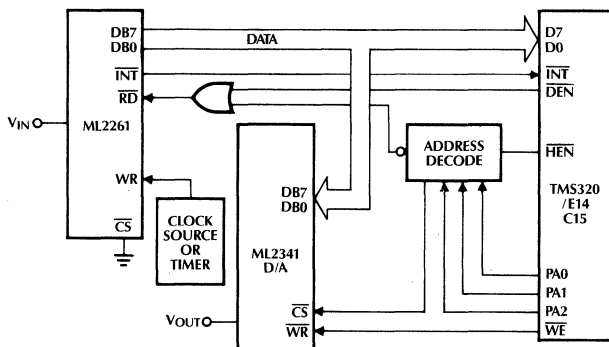


Figure 17. TMS320 Interface with D/A Output

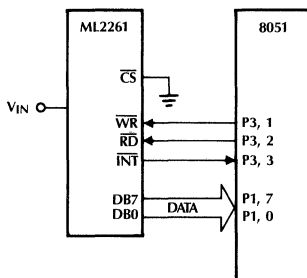


Figure 18. 8051 Interface to ML2261

ML2261

TYPICAL APPLICATIONS (Continued)

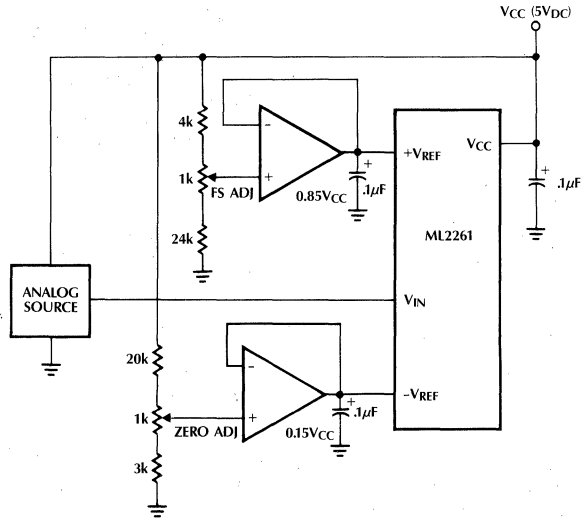


Figure 19. Operating with a Ratiometric Analog Signal of 15% of V_{CC} to 85% of V_{CC}

ORDERING INFORMATION

PART NUMBER	TOTAL UNADJUSTED ERROR	TEMPERATURE RANGE	PACKAGE
ML2261BIJ	±½ LSB	-40°C to +85°C	HERMETIC DIP (J20)
ML2261BCP		0°C to +70°C	MOLDED DIP (P20)
ML2261BCQ		0°C to +70°C	MOLDED PCC (Q20)
ML2261CIJ	±1 LSB	-40°C to +85°C	HERMETIC DIP (J20)
ML2261CCP		0°C to +70°C	MOLDED DIP (P20)
ML2261CCQ		0°C to +70°C	MOLDED PCC (Q20)

4-Channel High-Speed 8-Bit A/D Converter with T/H (S/H)

GENERAL DESCRIPTION

The ML2264 is a high-speed, μ P compatible, 4-channel 8-bit A/D converter with a conversion time of 680ns over the operating temperature range and supply voltage tolerance. The ML2264 operates from a single 5V supply and has an analog input range from GND to V_{CC} .

The ML2264 has two different pin selectable modes. The T/H mode has an internal track and hold. The S/H mode has a true internal sample and hold and can digitize 0 to 5V sinusoidal signals as high as 500kHz.

The ML2264 digital interface has been designed so that the device appears as a memory location or I/O port to a μ P. Analog input channels are selected by the latched and decoded multiplexer address inputs.

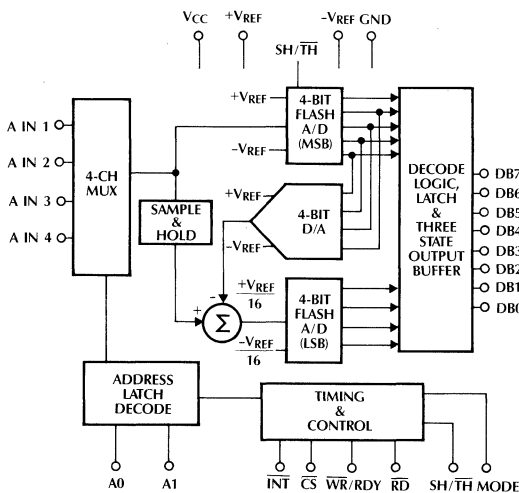
The ML2264 is an enhanced, pin compatible second source for the industry standard AD7824. The ML2264 enhancements are faster conversion time, parameters guaranteed over the supply tolerance and temperature range, improved digital interface timing, superior power supply rejection, and better latching immunity on analog inputs.

FEATURES

- Conversion time, WR-RD mode over temperature and supply voltage tolerance
 - Track & Hold Mode 830ns max
 - Sample & Hold Mode 700ns max
- Total unadjusted error $\pm 1/2$ LSB or ± 1 LSB
- Capable of digitizing a 5V, 250kHz sine wave
- 4-analog input channels
- No missing codes
- 0V to 5V analog input range with single 5V power supply
- No zero or full scale adjust required
- Analog input protection 25mA min
- Operates ratiometrically or with up to 5V voltage reference
- No external clock required
- Power-on reset circuitry
- Low power 100mW
- Narrow 24-pin DIP or surface mount SOIC
- Superior pin compatible replacement for AD7824

2

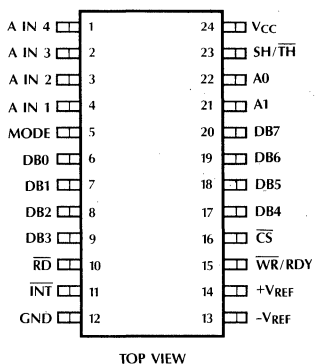
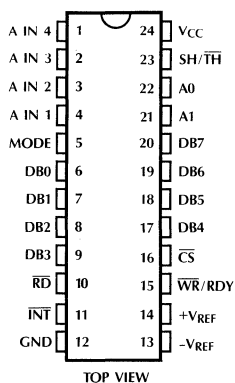
BLOCK DIAGRAM



PIN CONNECTIONS

24-Pin DIP

24-Pin SOIC



ML2264

PIN DESCRIPTION

PIN NO.	NAME	FUNCTION	PIN NO.	NAME	FUNCTION
1	A IN 4	Analog input 4.	15	WR/RDY	Write input or ready output. In WR-RD mode, this pin is WR input. In RD mode, this pin is RDY open drain output. See Digital Interface section.
2	A IN 3	Analog input 3.			
3	A IN 2	Analog input 2.			
4	A IN 1	Analog input 1.			
5	MODE	Mode select input. MODE = GND: RD mode MODE = V _{CC} : WR-RD mode Pin has internal current source pulldown to GND.	16	$\overline{\text{CS}}$	Chip select input. This pin must be held low for the device to perform a conversion.
6	DB0	Data output — bit 0 (LSB).	17	DB4	Data output — bit 4.
7	DB1	Data output — bit 1.	18	DB5	Data output — bit 5.
8	DB2	Data output — bit 2.	19	DB6	Data output — bit 6.
9	DB3	Data output — bit 3.	20	DB7	Data output — bit 7 (MSB).
10	$\overline{\text{RD}}$	Read input. In RD mode, this pin initiates a conversion. In WR-RD mode, this pin latches data into output latches. See Digital Interface section.	21	A1	Digital address input 1 that selects analog input channel. See multiplexer addressing section.
11	$\overline{\text{INT}}$	Interrupt output. This output signals the end of a conversion and indicates that data is valid on the data outputs. See Digital Interface section.	22	A0	Digital address input 0 that selects analog input channel. See multiplexer addressing section.
12	GND	Ground.	23	SH/ $\overline{\text{TH}}$	S/H, $\overline{\text{T/H}}$ mode select. When SH/ $\overline{\text{TH}}$ = V _{CC} , the device is in sample and hold mode. When SH/ $\overline{\text{TH}}$ = GND, the device is in track and hold mode. Pin has internal pulldown current source to GND.
13	-V _{REF}	Negative reference voltage for A/D converter.	24	V _{CC}	Positive supply. +5 volts ± 5%.
14	+V _{REF}	Positive reference voltage for A/D converter.			

ABSOLUTE MAXIMUM RATINGS

(Note 1)

Supply Voltage, V _{CC}	6.5V
Voltage	
Logic Inputs	-0.3V to V _{CC} + 0.3V
Analog Inputs	-0.3V to V _{CC} + 0.3V
Input Current per Pin (Note 2)	±25mA
Storage Temperature	-65°C to +150°C
Package Dissipation	
at T _A = 25°C (Board Mount)	875mW
Lead Temperature (Soldering 10 sec.)	
Dual-In-Line Package (Plastic)	260°C
Dual-In-Line Package (Ceramic)	300°C
SOIC	
Vapor Phase (60 sec.)	215°C
Infrared (15 sec.)	220°C

OPERATING CONDITIONS

Supply Voltage, V _{CC}	4.5V _{DC} to 6.0V _{DC}
Temperature Range (Note 3)	T _{MIN} ≤ T _A ≤ T _{MAX}
ML2264BMJ, ML2264CMJ	-55°C to +125°C
ML2264BIJ, ML2264CIJ	-40°C to +85°C
ML2264BCS, ML2264CCS	
ML2264BCP, ML2264CCP	0°C to +70°C

ELECTRICAL CHARACTERISTICS

Unless otherwise specified, $T_A = T_{MIN}$ to T_{MAX} , $V_{CC} = +V_{REF} = 5V \pm 5\%$, and $-V_{REF} = GND$

PARAMETER	NOTES	CONDITIONS	ML2264XCX			ML2264XIX, ML2264XMX			UNITS
			MIN	TYP (Note 4)	MAX	MIN	TYP (Note 4)	MAX	
Converter									
Total Unadjusted Error ML2264BXX ML2264CXX	5, 7	$V_{REF} = V_{CC}$			$\pm 1/2$ ± 1			$\pm 1/2$ ± 1	LSB LSB
Integral Linearity Error ML2264BXX ML2264CXX	5, 7	$V_{REF} = V_{CC}$			$\pm 1/2$ ± 1			$\pm 1/2$ ± 1	LSB LSB
Differential Linearity Error ML2264BXX ML2264CXX	5	$V_{REF} = V_{CC}$			$\pm 1/2$ ± 1			$\pm 1/2$ ± 1	LSB LSB
Full Scale Error ML2264BXX ML2264CXX	5				$\pm 1/2$ ± 1			$\pm 1/2$ ± 1	LSB LSB
Zero Scale Error ML2264BXX ML2264CXX	5				$\pm 1/2$ ± 1			$\pm 1/2$ ± 1	LSB LSB
Channel to Channel Mismatch	5				$\pm 1/4$			$\pm 1/4$	LSB
+ V_{REF} Voltage Range	6		$-V_{REF}$		$V_{CC}+0.1$	$-V_{REF}$		$V_{CC}+0.1$	V
- V_{REF} Voltage Range	6		$GND-0.1$		$+V_{REF}$	$GND-0.1$		$+V_{REF}$	V
Reference Input Resistance	5		1	2.5	4	1	2.5	4	k Ω
Analog Input Range	5, 8		$GND-0.1$		$V_{CC}+0.1$	$GND-0.1$		$V_{CC}+0.1$	V
Power Supply Sensitivity	5	DC $V_{CC}=5V \pm 5\%$, $V_{REF}=4.50V$		$\pm 1/32$	$\pm 1/4$		$\pm 1/32$	$\pm 1/4$	LSB
		100mVp-p 100kHz sine on V_{CC} $V_{IN} = 0$		$\pm 1/16$			$\pm 1/16$		LSB
Analog Input Leakage Current, OFF Channel	5	ON Channel = V_{CC} OFF Channel = 0V	-1			-1			μA
		ON Channel = 0V OFF Channel = V_{CC}			1			1	μA
Analog Input Leakage Current, ON Channel	5	ON Channel = 0V OFF Channel = V_{CC}	-1			-1			μA
		ON Channel = V_{CC} OFF Channel = 0V			1			1	μA
Analog Input Capacitance		During Acquisition Period		45			45		pF
Digital and DC									
$V_{IN(1)}$ Logical "1" Input Voltage	5	\overline{WR} , \overline{RD} , \overline{CS} , A0, A1	2.0			2.0			V
		MODE, SH/TH	$V_{CC}-0.5$			$V_{CC}-0.5$			V
$V_{IN(0)}$ Logical "0" Input Voltage	5	\overline{WR} , \overline{RD} , \overline{CS} , A0, A1			0.8			0.8	V
		MODE, SH/TH			0.5			0.5	V

ML2264

ELECTRICAL CHARACTERISTICS (Continued)

Unless otherwise specified, $T_A = T_{MIN}$ to T_{MAX} , $V_{CC} = +V_{REF} = 5V \pm 5\%$, $-V_{REF} = GND$, and timing measured at 1.4V, $C_L = 100pF$.

PARAMETER	NOTES	CONDITIONS	ML2264XCX			ML2264XIX, ML2264XMX			UNITS
			MIN	TYP (Note 4)	MAX	MIN	TYP (Note 4)	MAX	
Digital and DC (Continued)									
$I_{IN(1)}$, Logical "1" Input Current	5	$V_{IH} = V_{CC}$	$\overline{WR}, \overline{RD}, \overline{CS}, A0, A1$			1		1	μA
			$\overline{MODE}, \overline{SH}/\overline{TH}$	15	50	150	15	50	150
$I_{IN(0)}$, Logical "0" Input Current	5	$V_{IL} = GND$	$\overline{WR}, \overline{RD}, \overline{CS}$	-1			-1		μA
			$\overline{MODE}, \overline{SH}/\overline{TH}$	-20			-20		μA
$V_{OUT(1)}$, Logical "1" Output Voltage	5	$I_{OUT} = -2mA$		4.0			4.0		V
$V_{OUT(0)}$, Logical "0" Output Voltage	5	$I_{OUT} = 2mA$			0.4			0.4	V
I_{OUT} , Three-State Output Current	5	$V_{OUT} = 0V$		-1			-1		μA
		$V_{OUT} = V_{CC}$			1			1	μA
C_{OUT} , Logic Output Capacitance				5			5		pF
C_{IN} , Logic Input Capacitance				5			5		pF
I_{CC} , Supply Current	5	$\overline{CS} = \overline{WR} = \overline{RD} = "1"$ No Output Load				18		20	mA
AC and Dynamic Performance (Note 9)									
t_{CRD} , Conversion Time, Read Mode	5	\overline{RD} to INT, $MODE = 0V$				1020		1100	ns
t_{CWR-RD} , Conversion Time, Write-Read Mode	5, 9	\overline{WR} Falling Edge to INT, $t_{RD} < t_{INT}$, $MODE = V_{CC}$	$\overline{SH}/\overline{TH} = V_{CC}$			700		775	ns
			$\overline{SH}/\overline{TH} = GND$			830		930	ns
SNR, Signal to Noise Ratio		$V_{IN} = 5V, 250kHz$ Noise is sum of all nonfundamental components from 0-500kHz. $\overline{SH}/\overline{TH} = V_{CC}, MODE = V_{CC}$ $f_{SAMPLING} = 1.0 MHz$		48			48		dB
HD, Harmonic Distortion		$V_{IN} = 5V, 250kHz$ THD is sum of 2-5th harmonics relative to fundamental. $\overline{SH}/\overline{TH} = V_{CC}, MODE = V_{CC}$ $f_{SAMPLING} = 1.0 MHz$		-63			-63		dB
IMD, Intermodulation Distortion		$f_a = 2.5V, 250kHz$ $f_b = 2.5V, 248kHz$ IMB is $(f_a + f_b), (f_a - f_b), (2f_a + f_b), (2f_a - f_b), (f_a + 2f_b),$ or $(f_a - 2f_b)$ relative to fundamental. $\overline{SH}/\overline{TH} = V_{CC}, MODE = V_{CC}$ $f_{SAMPLING} = 1.0 MHz$		-60			-60		dB

ELECTRICAL CHARACTERISTICS (Continued)

Unless otherwise specified, $T_A = T_{MIN}$ to T_{MAX} , $V_{CC} = +V_{REF} = 5V \pm 5\%$, $-V_{REF} = GND$, and timing measured at 1.4V, $C_L = 100pF$

PARAMETER	NOTES	CONDITIONS	ML2264XCX			ML2264XIX, ML2264XMX			UNITS
			MIN	TYP (Note 4)	MAX	MIN	TYP (Note 4)	MAX	
AC and Dynamic Performance (Note 9) (Continued)									
FR, Frequency Response		$V_{IN} = 5V$, 0–250kHz Relative to 1kHz $SH/\overline{TH} = V_{CC}$, $MODE = V_{CC}$ $f_{SAMPLING} = 1.0$ MHz		± 0.1			± 0.1		dB
SR, Slew Rate Tracking	6	$SH/\overline{TH} = V_{CC}$			4.0			4.0	V/ μ s
		$SH/\overline{TH} = GND$.25			.25	V/ μ s
t_{AS} , Multiplexer Address Setup Time	5	$SH/\overline{TH} = GND$, Figure 1 (Track & Hold Operation)	0			0			ns
t_{AH} , Multiplexer Address Hold Time	5	$SH/\overline{TH} = GND$, Figure 1 (Track & Hold Operation)	60			70			ns
t_{AS} , Multiplexer Address Setup Time	5	$SH/\overline{TH} = V_{CC}$, Figure 2 (Sample & Hold Operation)	225			245			ns
t_{AH} , Multiplexer Address Hold Time	5	$SH/\overline{TH} = V_{CC}$, Figure 2 (Sample & Hold Operation)	60			70			ns
AC Performance Read Mode (Pin 5 = 0V), Figure 4									
t_{RDY} , \overline{CS} to RDY Delay	5		0		60	0		65	ns
t_{RDD} , \overline{RD} Low to RDY Delay	5, 10	Figure 3			1020			1100	ns
t_{CSS} , \overline{CS} to \overline{RD} , \overline{WR} Setup Time	5		0			0			ns
t_{CSH} , \overline{CS} to \overline{RD} , \overline{WR} Hold Time	5		0			0			ns
t_{CRD} , Conversion Time — \overline{RD} Low to INT Low	5, 10				1020			1100	ns
t_{ACCD} , Data Access Time RD to Data Valid	5		t_{CRD}		$t_{CRD}+30$	t_{CRD}		$t_{CRD}+30$	ns
t_{RDPW} , \overline{RD} Pulse Width	5		$t_{CRD}+30$			$t_{CRD}+30$			ns
t_{INTH} , \overline{RD} to INT Delay	5, 10		0		65	0		75	ns
t_{DH} , Data Hold Time — \overline{RD} Rising Edge to Data High Impedance State	6, 10	Figure 3	0		50	0		60	ns
t_p , Delay Time Between Conversions — INT Low to RD Low	5, 10	Sample & Hold Mode, $SH/\overline{TH} = V_{CC}$	300			325			ns
		Track & Hold Mode, $SH/\overline{TH} = GND$	240			260			ns
AC Performance Write-Read Mode (Pin 5 = 5V), Figures 5 and 6									
t_{CSS} , \overline{CS} to \overline{RD} , \overline{WR} Setup Time	5		0			0			ns
t_{CSH} , \overline{CS} to \overline{RD} , \overline{WR} Hold Time	5		0			0			ns
t_{WR} , \overline{WR} Pulse Width	5	$SH/\overline{TH} = V_{CC}$	190		50K	205		50K	ns
	6	$SH/\overline{TH} = GND$	320		50K	360		50K	ns

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ELECTRICAL CHARACTERISTICS (Continued)

Unless otherwise specified, $T_A = T_{MIN}$ to T_{MAX} , $V_{CC} = +V_{REF} = 5V \pm 5\%$, $-V_{REF} = GND$, and timing measured at 1.4V, $C_L = 100pF$

PARAMETER	NOTES	CONDITIONS	ML2264XCX			ML2264XIX, ML2264XMX			UNITS
			MIN	TYP (Note 4)	MAX	MIN	TYP (Note 4)	MAX	
AC Performance Write-Read Mode (Pin 5 = 5V), Figures 5 and 6 (Continued)									
t_{RD} , Read Time — WR High to RD Low Delay	5	$t_{RD} < t_{INTL}$	275			300			ns
t_{RI} , RD to INT Delay	5, 10	$t_{RD} < t_{INTL}$	0		235	0		270	ns
t_{ACC1} , Data Access Time — RD Low to Data Valid	5	$t_{RD} < t_{INTL}$	0		240	0		300	ns
t_{CWR-RD} , Conversion Time — WR Falling Edge to INT Low	5, 9, 10	$t_{RD} < t_{INTL}$, SH/TH = V_{CC}			700			775	ns
	6, 9, 10	$t_{RD} < t_{INTL}$, SH/TH = GND			830			930	ns
t_{INTL} , Internal Comparison Time — WR Rising Edge to INT Low	5, 10	$t_{RD} > t_{INTL}$			620			670	ns
t_{ACC2} , Data Access Time — RD to Data Valid	5	$t_{RD} > t_{INTL}$	0		50	0		60	ns
t_{DHI} , Data Hold Time — RD Rising Edge to Data High Impedance State	6, 10	Figure 3	0		50	0		60	ns
t_{INTH} , RD1 to INT1 Delay	5, 10		0		65	0		75	ns
t_p , Delay Time Between Conversions — INT Low to WR Low	5, 10	Sample & Hold Mode, SH/TH = V_{CC}	300			325			ns
		Track & Hold Mode, SH/TH = GND	240			260			ns
t_{IHWR} , WR1 to INT1 Delay	5, 10	Standalone Mode	0		90	0		100	ns
t_{ID} , INT1 to Data Valid Delay	5, 10	Standalone Mode	0		20	0		30	ns

Note 1: Absolute maximum ratings are limits beyond which the life of the integrated circuit may be impaired. All voltages unless otherwise specified are measured with respect to ground.

Note 2: When the voltage at any pin exceeds the power supply rails ($V_{IN} < GND$ or $V_{IN} > V_{CC}$) the absolute value of current at that pin should be limited to 25mA or less.

Note 3: $-55^{\circ}C$ to $+125^{\circ}C$ operating temperature range devices are 100% tested at temperature extremes with worst-case test conditions. $0^{\circ}C$ to $+70^{\circ}C$ and $-40^{\circ}C$ to $+85^{\circ}C$ operating temperature range devices are 100% tested with temperature limits guaranteed by 100% testing, sampling, or by correlation with worst-case test conditions.

Note 4: Typicals are parametric norm at $25^{\circ}C$.

Note 5: Parameter guaranteed and 100% production tested.

Note 6: Parameter guaranteed. Parameters not 100% tested are not in outgoing quality level calculation.

Note 7: Total unadjusted error includes offset, full scale, linearity, sample and hold, and multiplexer errors. Total unadjusted error is tested at the minimum specified times for WR, RD, t_{RI} , and t_p . For example, for the ML2264XCX in the sample and hold mode, WR/RD mode: $t_{WR} = 190ns$, $t_{RD} = 275ns$ with a frequency of 1.000MHz (cycle time of 1000ns).

Note 8: For $-V_{REF} \geq V_{IN}$ the digital output code will be 0000 0000. Two on-chip diodes are tied to the analog input which will forward conduct for analog input voltages one diode drop below ground or one diode drop greater than the V_{CC} supply. Be careful, during testing at low V_{CC} levels (4.5V), as high level analog inputs (5V) can cause this input diode to conduct — especially at elevated temperatures, and cause errors for analog inputs near full scale. The spec allows 100mV forward bias of either diode. This means that as long as the analog V_{IN} or V_{REF} does not exceed the supply voltage by more than 100mV, the output code will be correct. To achieve an absolute $0V_{DC}$ to $5V_{DC}$ input voltage range will therefore require a minimum supply voltage of $4.900V_{DC}$ over temperature variations, initial tolerance and loading.

Note 9: Conversion time, write-read mode = $t_{WR} + t_{RD} + t_{RI}$.

Note 10: Defined from the time an output crosses 0.8V or 2.4V.

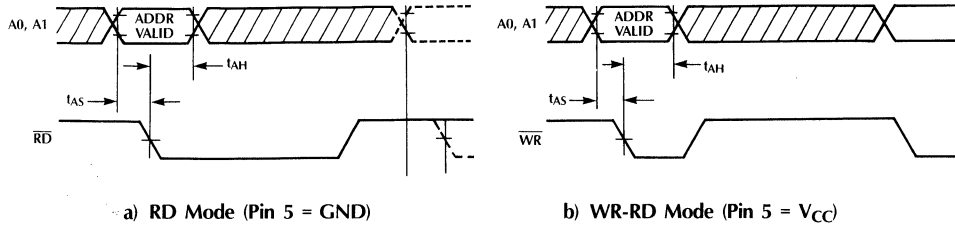


Figure 1. Analog Multiplexer Address Timing for Track & Hold Mode (Pin 23 = GND)

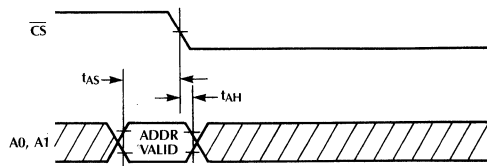


Figure 2. Analog Multiplexer Address Timing for Sample & Hold Mode (Pin 23 = VCC)

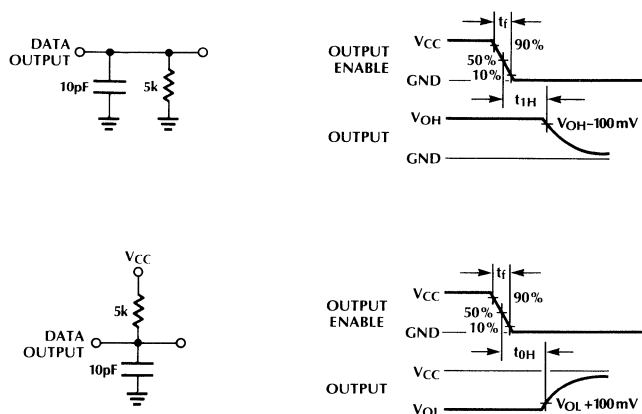
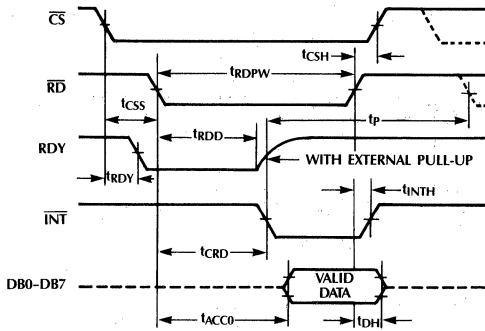


Figure 3. High Impedance Test Circuits and Waveforms



* In SAMPLE & HOLD mode a pull up resistor on RDY should not be used unless CSI is ≥ 20 ns before RD1.

Figure 4. RD Mode Timing

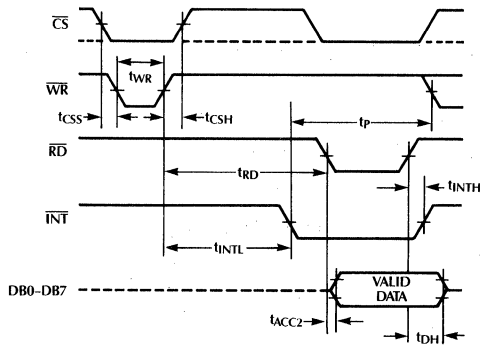


Figure 5. WR-RD Mode Timing ($t_{RD} > t_{INTL}$)

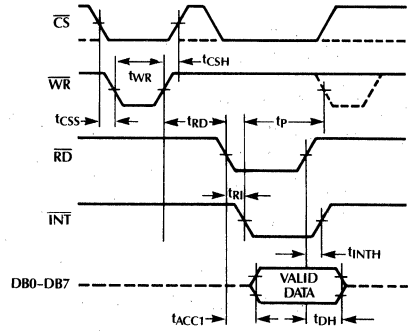


Figure 6. WR-RD Mode Timing ($t_{RD} < t_{INTL}$)

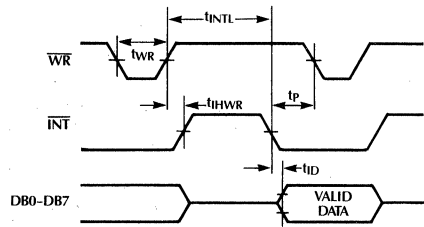


Figure 7. WR-RD Mode Stand-Alone Timing $\overline{CS} = \overline{RD} = 0$

1.0 FUNCTIONAL DESCRIPTION

The ML2264 uses a two stage flash technique for A/D conversion. This technique first performs a 4 bit flash conversion on V_{IN} to determine the 4 MSB's. These 4 MSB's are then cycled through an internal DAC to recreate the analog input. This reconstructed analog input signal from the DAC is then subtracted from the input, and the difference voltage is converted by a second 4 bit flash conversion, providing the 4 LSB's of the output data word.

1.1 MULTIPLEXER ADDRESSING

The ML2264 contains a 4-channel single ended analog multiplexer. A particular input channel is selected by using the address inputs A0 and A1. The relationship between the address inputs, A0 and A1, and the analog input selected is shown in Table 1.

Selected Analog Channel	Address Input	
	A0	A1
A IN 1	0	0
A IN 2	1	0
A IN 3	0	1
A IN 4	1	1

Table 1. Multiplexer Address Decoding

The address inputs are latched into the ML2264 on the falling edge of the RD, WR, or CS depending on the state of pins SH/TH and mode as shown in Table 2.

Address Latching Signal	Mode	Operation Mode
\overline{RD}	GND	GND
\overline{WR}	V_{CC}	GND
\overline{CS}	GND	V_{CC}
\overline{CS}	V_{CC}	V_{CC}

Table 2.

In the Sample & Hold mode of operation \overline{CS} is used as the address latch enable, allowing for continuous conversions without addressing a given analog input for each conversion.

The Track & Hold mode of operation requires an analog input to be addressed and latched for each conversion that the ML2264 performs.

1.2 ANALOG INPUTS

The analog input on the ML2264 behaves differently than inputs on conventional converters. The analog input current requirements change while the conversion is in progress, and the amount of input current depends on what cycle the converter is in.

The equivalent input circuit for the converter is shown in Figure 8. When the conversion starts in the T/H mode (WR1 in the WR-RD mode or RD1 in the RD mode) S1, S4 and S6 close and S3 opens. This period is known as the acquisition period where the MSB flash converter tracks the input signal and the LSB flash converter samples it. During this period, V_{IN} is connected to the 16 MSB and 15 LSB comparators. Thus 38 pF of input capacitance must be charged up through the combined R_{ON} resistance of the internal analog switches plus any external source resistance, R_S . In addition, there is a stray capacitance of approximately 11 pF that needs to be charged through the external source resistance R_S . This period ends in the WR-RD mode when WR1 or by an internal timer in the RD mode. At this point S1 and S4 open and the analog input at V_{IN} is no longer being sampled; thus during this time the analog voltage on V_{IN} does not affect converter performance.

As shown above, the critical period for charging up the analog input occurs when the MSB and LSB comparators are sampling the input, known as the acquisition period. The source of the external signal on V_{IN} must adequately charge up the analog voltage during the acquisition period. To do this, the input must settle within the required analog accuracy tolerance at least 50ns before the end of the acquisition period so that the MSB comparators have adequate time to make the correct decision. If more time is needed due to finite charging or settling time of the external source, the WR low period can be extended in WR-RD mode. In RD mode, since the acquisition time is fixed by internal delays, the burden is on the external source to charge up and settle the input adequately.

When the ML2264 operates in the S/H mode (pin 23 = V_{CC}) both the MSB and the LSB flash converter perform a true sample and hold operation during the acquisition or sampling period. This period starts after the falling edge of INT and ends with the falling edge of WR in the WR-RD mode or the falling edge of RD in the RD mode. The duration of this period is user controlled and must satisfy a minimum of t_p .

During this period S1, S3, S4 and S6 close, therefore 46 pF of input capacitance must be charged up in addition to the 11 pF of stray capacitance.

1.3 TRACK AND HOLD vs. SAMPLE AND HOLD

The MSB Flash Converter of the ML2264 in T/H mode has a track and hold mechanism for sampling the input. The input is attached to the MSB comparators directly in the MSB compare cycle, or acquisition period. When the MSB compare cycle ends, the state of the MSB comparators is latched. The LSB Flash Converter always performs a S/H operation. Thus, the analog input signal can be changing during the MSB compare cycle, or acquisition period, and the MSB comparators will be

2

tracking it as long as the slew rate of the analog input is slow enough so that the MSB comparators can respond. The ML2264 can track and hold signals with slew rates as high as $.25V/\mu s$ (16kHz @ 5 volts) without sacrificing conversion accuracy.

The ML2264 in S/H mode does not have the slew rate limitation of the T/H mode since an internal sample and hold acquires the analog signal, holds it internally, and then performs a conversion. Since this is a true sample and hold function, the S/H mode can theoretically digitize signals of frequencies much higher than the T/H mode. The ML2264 in S/H mode can digitize signals of frequencies as high as 250kHz @ 5V (slew rates as high as $4V/\mu s$) without sacrificing conversion accuracy. In most applications, the S/H mode is more desirable than T/H mode because of the better dynamic performance.

1.3.1 CONVERTER — T/H MODE

The operating sequence for the WR-RD mode is illustrated in Figure 9a. Initially, the internal comparators are auto-zeroed while WR is high. A conversion is initiated by the falling edge of WR. While WR is low, the MSB comparators are tracking the analog input and comparing this voltage against voltages from the internal resistor ladder. At the same time, the input is being acquired or sampled by LSB comparators. On the rising edge of WR, the MSB comparator results are latched, and the LSB acquisition time is ended by closing the sampling switch to the LSB comparators. While WR is high, the LSB comparators then compare the residual input voltage against internal voltages from the resistor ladder to determine the 4 LSB's. When the LSB comparison or conversion is complete, INT goes low and latches the conversion result into the output latches. Then, the comparators are auto-zeroed while WR is high before another conversion can start.

The operating sequence for RD mode, is similar to that described above for the WR-RD mode, except the conversion is initiated by the falling edge of RD, and the MSB and LSB conversions are generated by internal clock edges that are generated while RD is low.

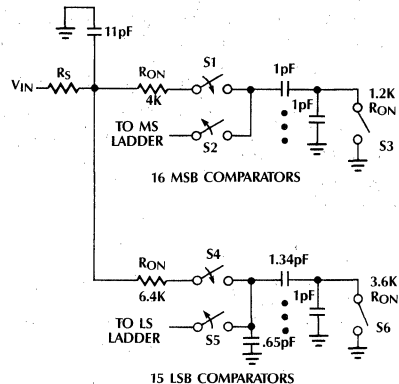


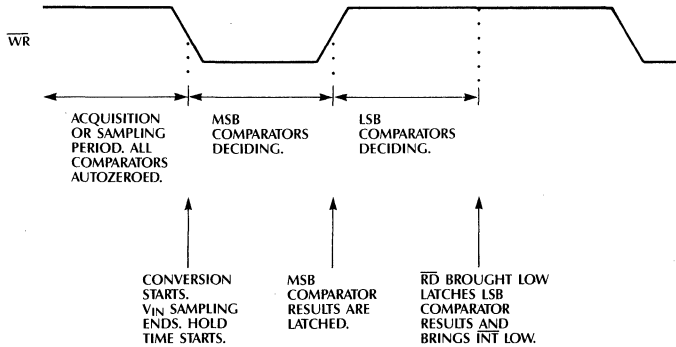
Figure 8. Converter Equivalent Input Circuit

1.3.2 CONVERTER — S/H MODE

The operating sequence for S/H mode is illustrated in Figure 9b. Notice that it is similar to T/H mode described above except this mode has a true sample and hold function. The falling edge of INT closes the sampling switch and starts the acquisition period where the analog input is sampled at the same time all comparators are auto-zeroed. The falling edge of WR opens the internal sampling switch, ends the acquisition period, and starts the conversion on the internally sample and held signal. The MSB comparators make their decisions while WR is low. On the rising edge of WR, the MSB comparator results are latched. The LSB comparators make their decision when WR is high. When the LSB comparison or conversion is complete, INT goes low and latches the conversion result into the output buffers. Then, the acquisition period begins again and the converter is ready for the next conversion.

The operating sequence for the RD mode is the same as the WR-RD mode, except the conversion is initiated by the falling edge of RD, and the MSB and LSB conversions are generated by internal clock edges that are generated while RD is low.

a). T/H Mode



b). S/H Mode

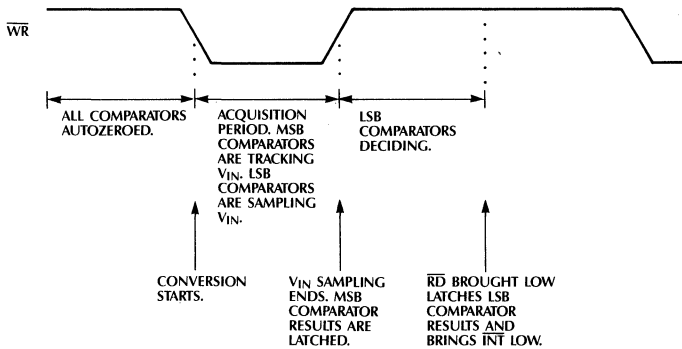


Figure 9. Operating Sequence (WR-RD Mode)

1.4 REFERENCE

The $+V_{REF}$ and $-V_{REF}$ inputs are the reference voltages that determine the full scale and zero input voltages, respectively, for the A/D converter. Thus, $+V_{REF}$ defines the analog input which produces a full scale output and $-V_{REF}$ defines the analog input which produces an output code of all zeroes. The transfer function for the A/D converter is shown in Figure 10.

$+V_{REF}$ and $-V_{REF}$ can be set to any voltage between GND and V_{CC} . This means that the reference voltages can be offset from GND and the difference between $+V_{REF+}$ and $-V_{REF-}$ can be made small to increase the resolution of the conversion. Note that the total unadjusted error increases when $[+V_{REF} - (-V_{REF})]$ decreases.

1.5 POWER SUPPLY AND REFERENCE DECOUPLING

A $0.1\mu F$ ceramic disc capacitor is recommended to bypass V_{CC} to GND, using as short a lead length as possible.

If REF+ and REF- inputs are driven by long lines, they should be bypassed by $0.1\mu F$ ceramic disc capacitors at the reference input pins.

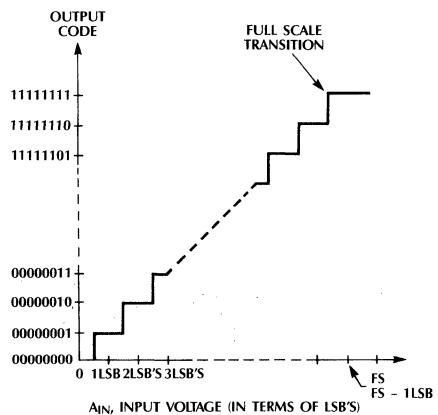


Figure 10. A/D Transfer Characteristic

1.6 DYNAMIC PERFORMANCE

1.6.1 SINUSOIDAL INPUTS

Since the ML2264 has an internal sample and hold, the device can digitize high frequency sinusoids with little or no signal degradations. Using the Nyquist criteria, the highest frequency input to the converter could theoretically be 1/2 the sampling rate (f_s). Any frequency components above $f_s/2$ will be aliased below $f_s/2$. In most applications, these aliased components cause unacceptable distortion and must be filtered out of the input. If the input frequency is too close to $f_s/2$, then the requirements on the anti-alias filter become difficult to impossible to realize with standard component and tolerances. In most practical applications, the highest input frequency has to be limited to 1/3 to 1/4 of f_{MAX} in order to relax the filtering requirements enough to make a realizable anti-alias filter.

The maximum sampling rate (f_{max}) for the ML2264 in the WR-RD mode, ($t_{RD} < t_{INTL}$) can be calculated as follows:

$$f_{max} = \frac{1}{t_{WR} + t_{RD} + t_{RI} + t_p}$$

$$f_{max} = \frac{1}{190ns + 275ns + 235ns + 300ns}$$

$$f_{max} = 1.000 \text{ MHz}$$

t_{WR} = Write Pulse Width

t_{RD} = Delay Time between WR and RD Pulses

t_{RI} = RD to INT Delay

t_p = Delay Time between Conversions

This permits a maximum sampling rate of 1MHz for the ML2264. The dynamic performance specifications (SNR, HD, IMD, and FR) for the ML2264 are all specified at 250kHz, which is approximately 1/4 of the sampling rate, f_s .

In applications where aliased frequency components are acceptable and filtering of the input signal is not needed, or where a filter with a steep amplitude response is available, the user can apply an input sinusoid higher than 250kHz to the device. Note, however, that as the input frequency increases above 500kHz, dynamic performance degradation will occur due to the finite bandwidth of the internal sample and hold.

The Figure 11 plots are 4096 point FFT's of the ML2264 converting a 257kHz and a 491kHz, 0 to 4.5V, low distortion sine wave input. The ML2264 samples and digitizes at its specified accuracy, dynamic input signals with frequency components up to the Nyquist frequency (one-half the sampling rate). The output spectra yields precise measurements of the input signal level, harmonic components, and signal to noise ratio up to the 8-bit level. The near ideal signal to noise ratio is maintained independent of increasing analog input frequencies to 500kHz.

1.6.2 SIGNAL-TO-NOISE RATIO

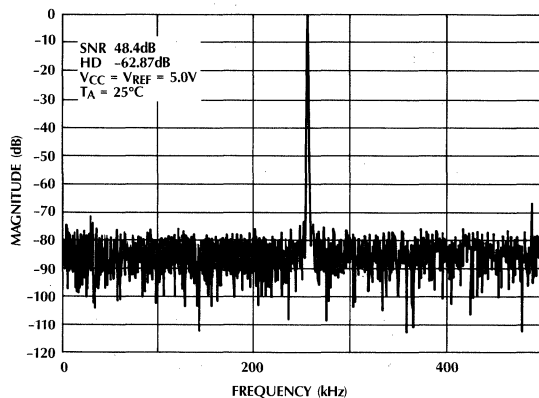
Signal-to-noise ratio (SNR) is the measured signal to noise at the output of the converter. The signal is the rms magnitude of the fundamental. Noise is the rms sum of all the nonfundamental signals up to half the sampling frequency. SNR is dependent on the number of quantization levels used in the digitization process; the more the levels, the smaller the quantization noise. The theoretical SNR for a sine wave is given by

$$SNR = (6.02N + 1.76) \text{ dB}$$

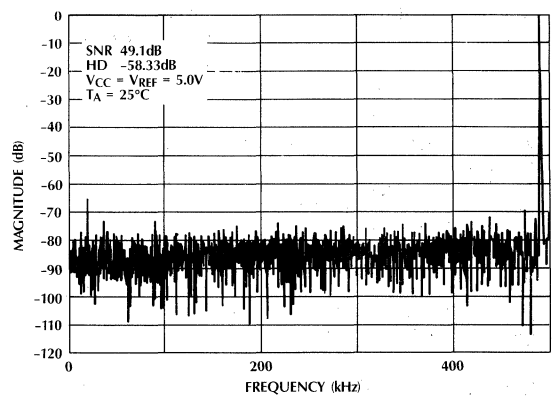
where N is the number of bits. Thus for ideal 8-bit converter, SNR = 49.92 dB.

1.6.3 HARMONIC DISTORTION

Harmonic distortion is the ratio of the rms sum of harmonics to the fundamental. Total harmonic distortion (THD) of the ML2264 is defined as



a) Output Spectrum with $f_{IN} = 257\text{kHz}$, $f_S = 1\text{MHz}$



b) Output Spectrum with $f_{IN} = 491\text{kHz}$, $f_S = 1\text{MHz}$

Figure 11. Dynamic Performance, Sample and Hold Mode

$$20 \log = \frac{(V_2^2 + V_3^2 + V_4^2 + V_5^2)^{1/2}}{V_1}$$

where V_1 is the rms amplitude of the fundamental and V_2, V_3, V_4, V_5 are the rms amplitudes of the individual harmonics.

1.6.2 SIGNAL-TO-NOISE RATIO

Signal-to-noise ratio (SNR) is the measured signal to noise at the output of the converter. The signal is the rms magnitude of the fundamental. Noise is the rms sum of all the nonfundamental signals up to half the sampling frequency. SNR is dependent on the number of quantization levels used in the digitization process; the more the levels, the smaller the quantization noise. The theoretical SNR for a sine wave is given by

$$\text{SNR} = (6.02N + 1.76) \text{ dB}$$

where N is the number of bits. Thus for ideal 8-bit converter, $\text{SNR} = 49.92 \text{ dB}$.

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Harmonic distortion is the ratio of the rms sum of harmonics to the fundamental. Total harmonic distortion (THD) of the ML2264 is defined as

$$20 \log = \frac{(V_2^2 + V_3^2 + V_4^2 + V_5^2)^{1/2}}{V_1}$$

where V_1 is the rms amplitude of the fundamental and V_2, V_3, V_4, V_5 are the rms amplitudes of the individual harmonics.

1.6.4 INTERMODULATION DISTORTION

With inputs consisting of sine waves at two frequencies, f_A and f_B , any active device with nonlinearities will create distortion products, of order $(m + n)$, at sum and difference frequencies of $mf_A + nf_B$, where $m, n = 0, 1, 2, 3 \dots$. Intermodulation terms are those for which m or n is not equal to zero. The (IMD) intermodulation distortion specification includes the second order terms $(f_A + f_B)$ and $(f_A - f_B)$ and the third order terms $(2f_A + f_B)$, $(2f_A - f_B)$, $(f_A + 2f_B)$, and $(f_A - 2f_B)$ only.

1.7 DIGITAL INTERFACE

The ML2264 has two basic interface modes, RD and WR-RD, which are selected by the MODE input pin.

1.7.1 RD MODE

In the RD mode, the $\overline{\text{WR/RDY}}$ pin is configured as the RDY output. The read mode performs a conversion with a single RD pulse. This allows the μP to start a conversion, wait, and then read data with a single read instruction.

The timing for the RD mode is shown in Figure 4. To do a conversion, $\overline{\text{CS}}$ must be low to select the device. After $\overline{\text{CS}}$ goes low, the RDY output goes low indicating that the device is ready to do a conversion. The conversion starts on the falling edge of RD. While RD is low, the MSB and LSB decisions are made with

internally generated clock edges. When the conversion is complete, RDY goes high and $\overline{\text{INT}}$ goes low signaling the end of the conversion. After $\overline{\text{INT}}$ goes low, the data outputs go from high impedance to active state with valid output data. Data stays valid until either RD or CS goes high. When either signal goes high, the output data lines return to the high impedance state and $\overline{\text{INT}}$ returns high.

1.7.2 WR-RD MODE

In the WR-RD mode, the $\overline{\text{WR/RDY}}$ pin is configured as the WR input. In this mode, WR initiates the conversion and RD controls reading the output data. This can be done in several ways, described below.

1.7.3 WR-RD MODE — USING INTERNAL DELAY ($t_{\text{RD}} > t_{\text{INTL}}$)

The timing is shown in Figure 5. To do a conversion, CS must be low to select the device. Then, WR falling edge triggers the conversion. While WR is low, the MSB comparison is made. When WR returns high the LSB decision is made. After some internal delay, INT goes low indicating end of conversion. Valid data will appear on DB0-7 when RD is pulled low. INT is then reset by the rising edge of either CS or RD.

1.7.4 WR-RD MODE — READING BEFORE DELAY ($t_{\text{RD}} < t_{\text{INTL}}$)

The internally generated delay for the LSB decision when $t_{\text{RD}} > t_{\text{INTL}}$ is longer than necessary due to circuit design tolerances of t_{INTL} delay. If desired, a faster conversion will result without loss of accuracy by bringing RD low within the minimum time specified for t_{RD} . The timing diagram for this mode is shown in Figure 6. WR is the same as when $t_{\text{RD}} > t_{\text{INTL}}$. But in this case, RD is brought low t_{RD} ns after WR rising edge and before INT. INT goes low indicating an end of conversion after the falling edge of RD and is reset on the rising edge of RD or CS. When RD is brought low before INT goes low the data bus always remains in the high-impedance state until INT \uparrow .

1.7.5 WR-RD MODE — STAND ALONE OPERATION

Stand alone operation can be implemented by tying $\overline{\text{CS}}$ and RD low as shown in Figure 7. WR initiates a conversion as before. When WR is low, the MSB comparison is made. When WR goes high, the LSB comparison is made. Since RD is already low, the output data will appear automatically at end of conversion. Since RD is always low, INT is reset on rising edge of WR and goes low at end of conversion.

1.7.6 POWER-ON RESET

When power is first applied, an internal power-on reset and timer circuit inhibits the $\overline{\text{CS}}$ input and resets the internal circuitry to prevent the ML2264 from starting in an unknown state. During this period of approximately $3\mu\text{s}$, INT remains high and the data bus is in the high-impedance state.

2.0 TYPICAL APPLICATIONS

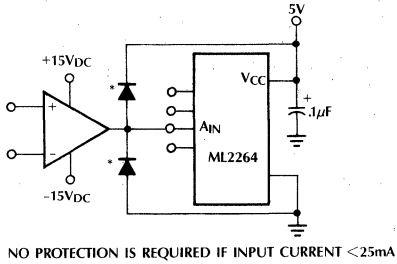


Figure 12. Protecting the Input

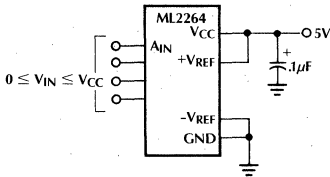


Figure 13. Using V_{CC} as Reference for Ratiometric Operation

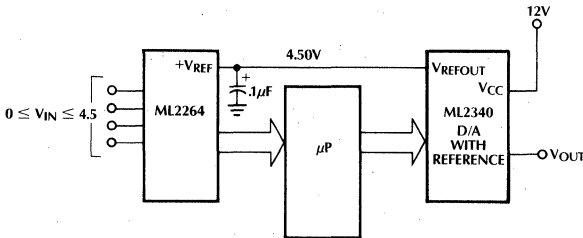


Figure 14. Using External Reference of D/A

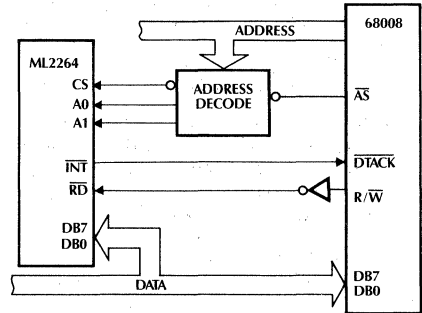


Figure 15. 68000 Type Interface to ML2264

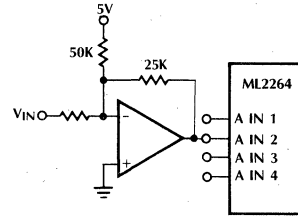


Figure 16. ±2.5V Analog Input Range

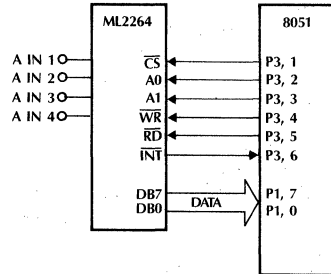


Figure 17. 8051 Interface to ML2264

μ P Compatible High-Speed 10-Bit A/D Converter with S/H

GENERAL DESCRIPTION

The ML2271 is a high speed, μ P compatible 10-bit A/D converter. A three step flash technique is used to achieve a conversion time of 1.45μ s. The ML2271 operates from a single 5V supply and has an analog input range from GND to V_{CC} .

The ML2271 has a true internal sample and hold and can digitize sinusoid signals as high as 150kHz without conversion errors.

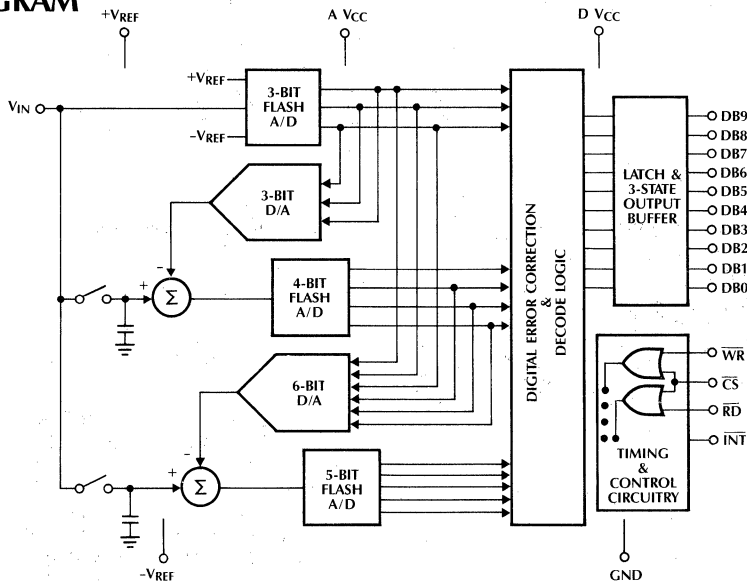
The ML2271 digital interface has been designed so that the device appears as a memory location or I/O port to a μ P, eliminating the need for external interfacing logic. The data outputs are latched and have three state control, allowing direct connection to a μ P bus or I/O port. The addition of an internal timing generator also allows the device to easily operate in stand alone applications.

The ML2271 is pin and function compatible with the ADC1061.

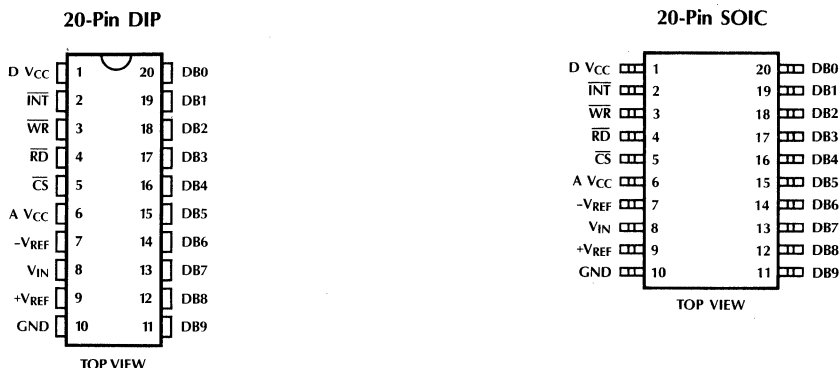
FEATURES

- Conversion time over temperature and supply voltage tolerance 1.5 μ s
- Linearity error $\pm 1/2$ LSB or ± 1 LSB
- Full scale error $\pm 1/2$ LSB or ± 1 LSB
- Zero error $\pm 1/2$ LSB or ± 1 LSB
- Capable of digitizing a 5V, 150kHz sine wave
- No missing codes
- 0V to 5V analog input range with single 5V power supply
- Analog input protection 25mA min
- Operates ratiometrically or with up to 5V voltage reference
- No external clock required
- Easy interface to μ P, or operates standalone
- Latched, 3-state data outputs
- Power-on reset circuitry
- Low power 180mW max
- Standard 20-pin DIP or surface mount SOIC
- 0°C to 70°C, -40°C to +85°C, -55°C to +125°C operating temperature range

BLOCK DIAGRAM



PIN CONNECTIONS



PIN DESCRIPTION

PIN NO.	NAME	FUNCTION	PIN NO.	NAME	FUNCTION
1	D V _{CC}	Digital supply. +5V ± 5%. Connect to A V _{CC} .	7	-V _{REF}	Negative reference input voltage for A/D converter.
2	$\overline{\text{INT}}$	Interrupt output. This output signals the end of a conversion and indicates that data is valid on the data outputs. See Digital Interface section.	8	V _{IN}	Analog input.
3	$\overline{\text{WR}}$	Write input. Input which initiates a conversion. See Digital Interface section.	9	+V _{REF}	Positive reference input voltage for A/D converter.
4	$\overline{\text{RD}}$	Read input. This input latches data into the output latches. See Digital Interface section.	10	GND	Ground.
5	$\overline{\text{CS}}$	Chip select input. This input must be held low during WR and RD for the device to perform a conversion.	11	DB9	Data output — bit 9 (MSB).
6	A V _{CC}	Analog supply. +5V ± 5%. Connect to D V _{CC} .	12	DB8	Data output — bit 8.
			13	DB7	Data output — bit 7.
			14	DB6	Data output — bit 6.
			15	DB5	Data output — bit 5.
			16	DB4	Data output — bit 4.
			17	DB3	Data output — bit 3.
			18	DB2	Data output — bit 2.
			19	DB1	Data output — bit 1.
			20	DB0	Data output — bit 0 (LSB).

ABSOLUTE MAXIMUM RATINGS

(Note 1)

Supply Voltage, A V _{CC} , D V _{CC}	6.5V
Voltage	
Logic Inputs	-0.3V to V _{CC} + 0.3V
Analog Inputs	-0.3V to V _{CC} + 0.3V
Input Current per Pin (Note 2)	±25mA
Storage Temperature	-65°C to +150°C
Package Dissipation	
at T _A = 25°C (Board Mount)	875mW
Lead Temperature (Soldering 10 sec.)	
Dual-In-Line Package (Molded)	260°C
Dual-In-Line Package (Ceramic)	300°C
Molded Small Outline IC Package	
Vapor Phase (60 sec.)	215°C
Infrared (15 sec.)	220°C

OPERATING CONDITIONS

Temperature Range (Note 3)	T _{MIN} ≤ T _A ≤ T _{MAX}
ML2271BMJ, ML2271CMJ	-55°C to +125°C
ML2271BIJ, ML2271CIJ	-40°C to +85°C
ML2271BCS, ML2271CCS	
ML2271BCP, ML2271CCP	0°C to +70°C

ML2271

ELECTRICAL CHARACTERISTICS

Unless otherwise specified, $T_A = T_{MIN}$ to T_{MAX} , $D V_{CC} = A V_{CC} = +V_{REF} = 5V \pm 5\%$, and $-V_{REF} = GND$

PARAMETER	NOTES	CONDITIONS	ML2271XCX			ML2271XIX, ML2271XMX			UNITS
			MIN	TYP (Note 4)	MAX	MIN	TYP (Note 4)	MAX	
Converter									
Integral Linearity Error ML2271BXX ML2271CXX	5, 7	$V_{REF} = V_{CC}$			$\pm 1/2$ ± 1			$\pm 1/2$ ± 1	LSB LSB
Differential Linearity Error ML2271BXX ML2271CXX	5	$V_{REF} = V_{CC}$			$\pm 1/2$ ± 1			$\pm 1/2$ ± 1	LSB LSB
Full Scale Error ML2271BXX ML2271CXX	5				$\pm 1/2$ ± 1			$\pm 1/2$ ± 1	LSB LSB
Zero Scale Error ML2271BXX ML2271CXX	5				$\pm 1/2$ ± 1			$\pm 1/2$ ± 1	LSB LSB
Total Unadjusted Error ML2271BXX ML2271CXX	5				$\pm 3/4$ $\pm 1\frac{1}{2}$			$\pm 3/4$ $\pm 1\frac{1}{2}$	LSB LSB
+ V_{REF} Voltage Range	6		$-V_{REF}$		$V_{CC}+0.1$	$-V_{REF}$		$V_{CC}+0.1$	V
- V_{REF} Voltage Range	6		$GND-0.1$		$+V_{REF}$	$GND-0.1$		$+V_{REF}$	V
Reference Input Resistance	5		.9	1.3	1.7	.9	1.3	1.7	k Ω
Analog Input Range	5, 8		$-V_{REF}$		$+V_{REF}$	$-V_{REF}$		$+V_{REF}$	V
Power Supply Sensitivity	5	DC $V_{CC} = 5V \pm 5\%$, $V_{REF} = 4.75V$		$\pm 1/32$	$\pm 1/4$		$\pm 1/32$	$\pm 1/4$	LSB
		100mVp-p 100kHz sine on V_{CC} $V_{IN} = 0$		$\pm 1/16$			$\pm 1/16$		LSB
Analog Input Leakage Current	5, 9	Converter Idle	-2		+2	-2		+2	μA
Analog Input Capacitance		During Acquisition Period		25			25		pF
Digital and DC									
$V_{IN(1)}$, Logical "1" Input Voltage	5		2.0			2.0			V
$V_{IN(0)}$, Logical "0" Input Voltage	5				0.8			0.8	V
$I_{IN(1)}$, Logical "1" Input Current	5	$V_{IN} = V_{CC}$			1			1	μA
$I_{IN(0)}$, Logical "0" Input Current	5	$V_{IN} = 0V$	-1			-1			μA
$V_{OUT(1)}$, Logical "1" Output Voltage	5	$I_{OUT} = -2mA$	4.0			4.0			V
$V_{OUT(0)}$, Logical "0" Output Voltage	5	$I_{OUT} = 2mA$			0.4			0.4	V
I_{OUT} , Three-State Output Current	5	$V_{OUT} = 0V$	-1			-1			μA
		$V_{OUT} = V_{CC}$			1			1	μA

ELECTRICAL CHARACTERISTICS (Continued)

Unless otherwise specified, $T_A = T_{MIN}$ to T_{MAX} , $D V_{CC} = A V_{CC} = +V_{REF} = 5V \pm 5\%$, $-V_{REF} = GND$, and timing measured at 1.4V, $C_L = 100pF$.

PARAMETER	NOTES	CONDITIONS	ML2271XCX			ML2271XIX, ML2271XMX			UNITS
			MIN	TYP (Note 4)	MAX	MIN	TYP (Note 4)	MAX	
Digital and DC (Continued)									
C_{OUT} , Logic Output Capacitance				5			5		pF
C_{IN} , Logic Input Capacitance				5			5		pF
I_{CC} , Supply Current, Analog Plus Digital	5	$\overline{CS} = \overline{WR} = \overline{RD} = 0$ No Output Load			32			35	mA
AC and Dynamic Performance (Note 9)									
t_{CONV} , Conversion Time, Interrupt Mode	6	Figure 2			1450			1600	ns
t_{CONV} , Conversion Time, Write-Read Mode	5	Figure 3			1450			1600	ns
t_{CONV} , Conversion Time, Read Mode	6	Figure 4			1450			1600	ns
SNR, Signal to Noise Ratio		$V_{IN} = 5V, 150kHz$ Noise is sum of all nonfundamental components from 0–300kHz. $f_{SAMPLING} = 600kHz$		60			60		dB
HD, Harmonic Distortion		$V_{IN} = 5V, 150kHz$ THD is sum of 2–5th harmonics or aliases relative to fundamental. $f_{SAMPLING} = 600kHz$		-60			-60		dB
IMD, Intermodulation Distortion		$f_a = 2.5V, 150kHz$ $f_b = 2.5V, 148kHz$ IMB is $(f_a + f_b)$, $(f_a - f_b)$, $(2f_a + f_b)$, $(2f_a - f_b)$, $(f_a + 2f_b)$, or $(f_a - 2f_b)$ relative to fundamental. $f_{SAMPLING} = 600kHz$		-60			-60		dB
FR, Frequency Response		$V_{IN} = 5V, 0-150kHz$ Relative to 1kHz $f_{SAMPLING} = 600kHz$		± 0.1			± 0.1		dB
SR, Slew Rate Tracking				2.36			2.36		V/ μs
AC Performance, Figures 2, 3, 4, and 5									
t_{CSS} , \overline{CS} to \overline{RD} , \overline{WR} Setup Time	5		0			0			ns
t_{CSH} , \overline{CS} to \overline{RD} , \overline{WR} Hold Time	5		0			0			ns
t_{WR} , \overline{WR} Pulse Width	5		250		50K	300		50K	ns

ELECTRICAL CHARACTERISTICS (Continued)

Unless otherwise specified, $T_A = T_{MIN}$ to T_{MAX} , $D V_{CC} = A V_{CC} = +V_{REF} = 5V \pm 5\%$, and $-V_{REF} = GND$, and timing measured at 1.4V, $C_L = 100$ pF.

PARAMETER	NOTES	CONDITIONS	ML2271XCX			ML2271XIX, ML2271XMX			UNITS
			MIN	TYP (Note 4)	MAX	MIN	TYP (Note 4)	MAX	
AC Performance, Figures 2, 3, 4, and 5 (Continued)									
t_{ACC2} , WR to Data Valid	5				1500			1600	ns
t_{RD} , Read Pulse Width	5		100			120			ns
t_{WRL} , $\overline{WR1}$ to $\overline{RD1}$	6		0			0			ns
t_{INTH} , $\overline{RD1}$ to $\overline{INT1}$	5		10		50	10		50	ns
t_{RIL} , \overline{RD} to $\overline{INT1}$	5		0		50	0		50	ns
t_{ACCI} , Data Access Time, $\overline{RD1}$ to Data Valid	5		0		55	0		60	ns
t_{ID} , Data Access Time, $\overline{INT1}$ to Data Valid	5		0		55	0		60	ns
t_{IH} , t_{OH} , $\overline{RD1}$ to Data High Impedance State	5	Figure 1	10		50	10		60	ns
t_p , Delay Time Between Conversions	5		300			300			ns

Note 1: Absolute maximum ratings are limits beyond which the life of the integrated circuit may be impaired. All voltages unless otherwise specified are measured with respect to ground.

Note 2: When the voltage at any pin exceeds the power supply rails ($V_{IN} < GND$ or $V_{IN} > V_{CC}$) the absolute value of current at that pin should be limited to 25mA or less.

Note 3: -55°C to $+125^{\circ}\text{C}$ operating temperature range devices are 100% tested at temperature extremes with worst-case test conditions. 0°C to $+70^{\circ}\text{C}$ and -40°C to $+85^{\circ}\text{C}$ operating temperature range devices are 100% tested with temperature limits guaranteed by 100% testing, sampling, or by correlation with worst-case test conditions.

Note 4: Typicals are parametric norm at 25°C .

Note 5: Parameter guaranteed and 100% production tested.

Note 6: Parameter guaranteed. Parameters not 100% tested are not in outgoing quality level calculation.

Note 7: Total unadjusted error includes offset, full scale, linearity, and sample and hold errors.

Note 8: For $-V_{REF} \geq V_{IN}$ the digital output code will be 0000 0000. Two on-chip diodes are tied to the analog input which will forward conduct for analog input voltages one diode drop below ground or one diode drop greater than the V_{CC} supply. Be careful, during testing at low V_{CC} levels (4.5V), as high level analog inputs (5V) can cause this input diode to conduct — especially at elevated temperatures, and cause errors for analog inputs near full scale. The spec allows 100mV forward bias of either diode. This means that as long as the analog V_{IN} or V_{REF} does not exceed the supply voltage by more than 100mV, the output code will be correct. To achieve an absolute $0V_{DC}$ to $5V_{DC}$ input voltage range will therefore require a minimum supply voltage of $4.900V_{DC}$ over temperature variations, initial tolerance and loading.

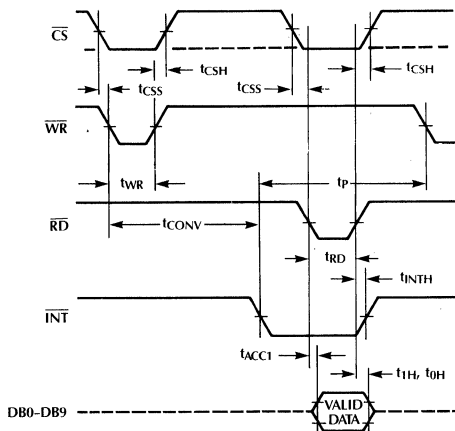
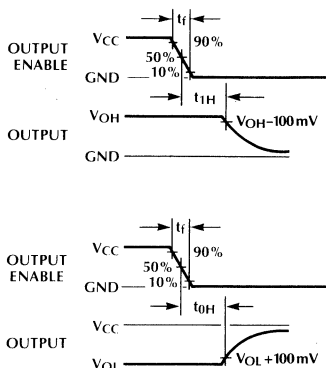
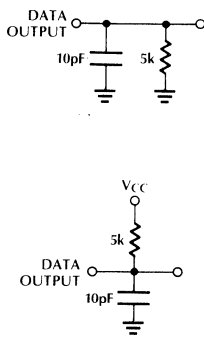


Figure 1. High Impedance Test Circuits and Waveforms

Figure 2. Interrupt Mode Timing

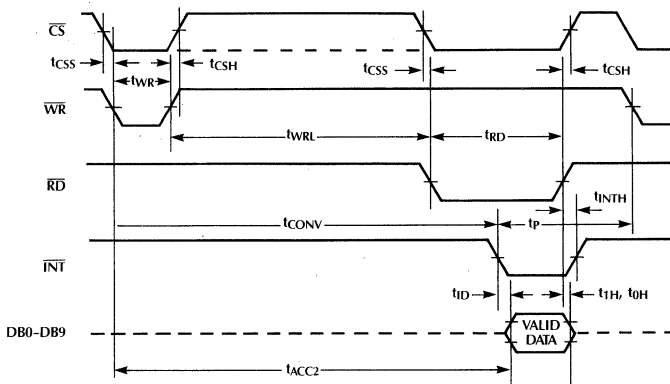
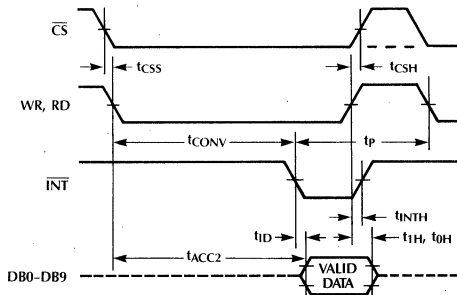


Figure 3. WR-RD Mode Timing



READ IS TIED TO WRITE

Figure 4. RD Mode Timing

1.0 FUNCTIONAL DESCRIPTION

The ML2271 uses a three step flash technique for A/D conversion. This technique first performs a 3 bit flash conversion on V_{IN} to determine the 3 most significant bits (MSB decision). These 3 MSB's are then cycled through an internal DAC to recreate the analog input. This reconstructed analog input signal from the DAC is then subtracted from the input, and the difference voltage is converted by a second 3 bit flash conversion providing the next 3 significant bits, called intermediate significant bits (ISB decision). This procedure is then performed again to provide the final 4 least significant bits (LSB decision).

The ML2271 has a true internal sample and hold. The internal operating sequence is shown in Figure 5. The falling edge of \overline{WR} opens the S/H sampling switch, ends the acquisition time for the analog input, and starts the conversion on the internally sample and held signal. Then the MSB, ISB, and LSB decisions are made. \overline{INT} goes low at end of conversion and \overline{RD} controls the data outputs. This falling edge of \overline{INT} also closes the sampling switch and starts the acquisition period for the next conversion.

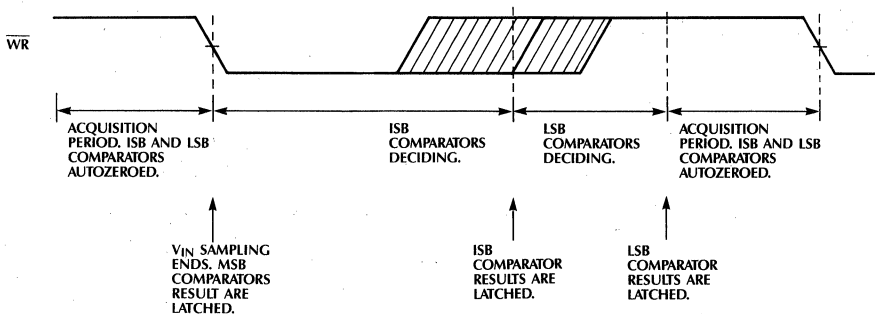


Figure 5. Operating Sequence

1.1 ANALOG INPUT

The analog input on the ML2271 behaves differently than inputs on conventional converters. The analog input current requirements change while the conversion is in progress, and the amount of input current depends on what cycle the converter is in.

The input circuit for the converter is shown in Figure 6A with the equivalent input circuit shown in Figure 6B. The acquisition period for the S/H starts on \overline{INT} falling edge and ends on \overline{WR} falling edge.

The critical period for charging up the analog input occurs during the acquisition period and the source of the external signal on V_{IN} must adequately charge up the analog voltage during this time. To do this, the input must settle within the required analog accuracy tolerance 100ns before the end of the acquisition period so that the sampling capacitors have adequate time to store the input signal. If more time is needed due to finite charging or settling time of the external source, the \overline{WR} high period can be extended as long as is required.

1.2 SAMPLE AND HOLD

The ML2271 does not have the limitation of an equivalent circuit implemented with a track/hold. An internal sample and hold acquires the analog signal, holds it internally, and then a conversion is performed on the sample and held signal. Since this is a true sample and hold function, the ML2271 can sample and hold signals with frequencies as high as 150kHz @ 5V (slew rates as high as 2.36V/ μ s) without sacrificing conversion accuracy.

1.3 REFERENCE

The $+V_{REF}$ and $-V_{REF}$ inputs are the reference voltages that determine the full scale and zero input voltages, respectively, for the A/D converter. Thus, $+V_{REF}$ defines the analog input which produces a full scale output and $-V_{REF}$ defines the analog input which produces an output code of all zeroes. The transfer function for the A/D converter is shown in Figure 7.

+V_{REF} and -V_{REF} can be set to any voltage between GND and V_{CC}. This means that the reference voltages can be offset from GND and the difference between +V_{REF} and -V_{REF} can be made small to increase the resolution of the conversion. Note that the linearity error increases when [+V_{REF} - (-V_{REF})] decreases.

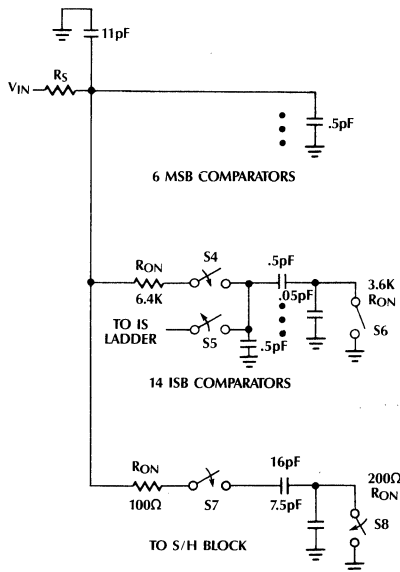


Figure 6A. Converter Input Circuit

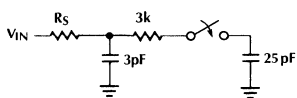


Figure 6B. Converter Equivalent Input Circuit

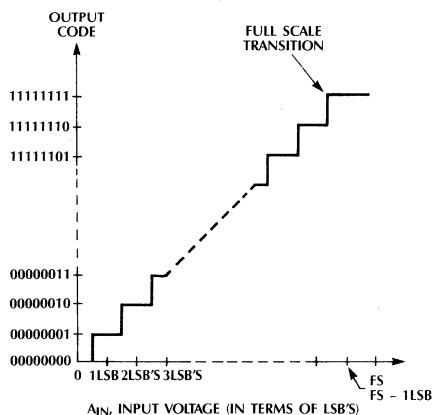


Figure 7. A/D Transfer Characteristic

1.4 POWER SUPPLY AND REFERENCE DECOUPLING

0.1μF in parallel with 0.01μF ceramic disc capacitors are recommended to bypass A V_{CC} to GND, as well as D V_{CC} to GND, using the shortest lead lengths possible.

If +V_{REF} and -V_{REF} inputs are driven by long lines, they should be bypassed by 0.1μF in parallel with 0.01μF ceramic disc capacitors at the reference input pins.

1.5 DYNAMIC PERFORMANCE

1.5.1 SINUSOIDAL INPUTS

Since the ML2271 has an internal sample and hold, the device can digitize high frequency sinusoids with little or no signal degradations. Using the Nyquist criteria, the highest frequency input to the converter could theoretically be 1/2 the sampling rate (f_s). Any frequency components above f_s/2 will be aliased below f_s/2. In most applications, these aliased components cause unacceptable distortion and must be filtered out of the input. If the input frequency is too close to f_s/2, then the requirements on the antialias filter become difficult or impossible to realize with standard component and tolerances. In most practical applications, the highest input frequency has to be limited to 1/3 to 1/4 of f_s in order to relax the filtering requirements enough to make a realizable antialias filter.

The maximum sampling rate (f_{MAX}) for the ML2271 can be calculated as follows:

$$f_{max} = \frac{1}{t_{CONV} + t_p}$$

$$f_{max} = \frac{1}{1.45\mu s + 0.300\mu s}$$

$$f_{max} = 570kHz$$

t_{WR} = Write Pulse Width

t_{WRD} = Write to Data Delay

t_p = Delay Time between Conversions

Note that the dynamic performance specifications (SNR, HD, IMD, and FR) for the ML2271 are all specified at 150kHz, which is less than 1/3 of the sampling rate, f_s. This allows adequate margin between the input frequency and the aliased components to allow antialias filtering if needed.

In applications where aliased frequency components are acceptable and filtering of the input signal is not needed, the user can apply an input sinusoid higher than 150kHz to the device. Note, however, that as the input frequency increases above 150kHz, dynamic performance degradation will occur due to the finite bandwidth of the internal sample and hold.

1.5.2 SIGNAL-TO-NOISE RATIO

Signal-to-noise ratio (SNR) is the measured signal to noise at the output of the converter. The signal is the rms magnitude of the fundamental. Noise is the rms sum of all the nonfundamental signals up to half the sampling frequency. SNR is dependent on the number of quantization levels used in the digitization process; the more the levels, the smaller the quantization noise. The theoretical SNR for a sine wave is given by

$$\text{SNR} = (6.02N + 1.76) \text{ dB}$$

where N is the number of bits. Thus for ideal 10-bit converter, SNR = 61.96 dB.

1.5.3 HARMONIC DISTORTION

Harmonic distortion is the ratio of the rms sum of harmonics to the fundamental. Total harmonic distortion (THD) of the ML2271 is defined as

$$20 \log = \frac{(V_2^2 + V_3^2 + V_4^2 + V_5^2)^{1/2}}{V_1}$$

where V_1 is the rms amplitude of the fundamental and V_2, V_3, V_4, V_5 are the rms amplitudes of the individual harmonics.

1.5.4 INTERMODULATION DISTORTION

With inputs consisting of sine waves at two frequencies, f_A and f_B , any active device with nonlinearities will create distortion products, of order $(m + n)$, at sum and difference frequencies of $mf_A + nf_B$, where $m, n = 0, 1, 2, 3 \dots$. Intermodulation terms are those for which m or n is not equal to zero. The (IMD) intermodulation distortion specification includes the second order terms $(f_A + f_B)$ and $(f_A - f_B)$ and the third order terms $(2f_A + f_B)$, $(2f_A - f_B)$, $(f_A + 2f_B)$, and $(f_A - 2f_B)$ only.

1.6 DIGITAL INTERFACE

Depending on the way the external signals are applied to the ML2271, the timing of the conversion and resultant digital interface can be configured in three different modes.

While the operation for each mode is described below, there are some general rules that dictate the general relationships between $\overline{\text{CS}}$, $\overline{\text{WR}}$, $\overline{\text{RD}}$, $\overline{\text{INT}}$, and DB0-DB9 . The falling edge of $\overline{\text{WR}}$ terminates the acquisition period and initiates a conversion. $\overline{\text{INT}}$ is forced low when a conversion is internally completed. $\overline{\text{INT}}$ is reset high by the $\overline{\text{RD}}$ rising edge. DB0-DB9 is in the high impedance state except when both $\overline{\text{RD}}$ and $\overline{\text{INT}}$ are low. $\overline{\text{RD}}$ low period does not affect the internal conversion but only determines when the digital signals DB0-DB9 are active; thus, $\overline{\text{RD}}$ can occur anytime. $\overline{\text{CS}}$ is used to select the device and needs to be low only while $\overline{\text{WR}}$ is low or when $\overline{\text{RD}}$ is low.

1.6.1 INTERRUPT MODE

Timing for the Interrupt Mode is shown in Figure 2. To do a conversion, $\overline{\text{CS}}$ must be low to select the device. $\overline{\text{INT}}$ falling edge starts the acquisition period. The falling edge of $\overline{\text{WR}}$ ends the acquisition period and the MSB comparison is made. Then, the (Intermediate Significant Bits) ISB and LSB decisions are made with internal timing signals. After the conversion is complete, $\overline{\text{INT}}$ goes low indicating end of conversion. When $\overline{\text{RD}}$ goes low, DB0-DB9 goes from high impedance to the active state with the digital result of the conversion. $\overline{\text{INT}}$ is reset high and DB0-DB9 is reset to high impedance on the rising edge of $\overline{\text{RD}}$.

Interrupt Operation is intended to be used in interrupt driven systems or applications where $\overline{\text{INT}}$ signals the transfer of data.

1.6.2 WRITE-READ MODE

Write-Read Operation is the same as Interrupt Operation except that $\overline{\text{RD}}$ is brought low before the internal conversion is completed (before $\overline{\text{INT}}$ goes low).

Timing for Write-Read Operation is shown in Figure 3. To perform a conversion, $\overline{\text{CS}}$ must be low to select the device. $\overline{\text{INT}}$ falling edge starts the acquisition period. The falling edge of $\overline{\text{WR}}$ ends the acquisition period and the MSB decision is made. Then, the ISB and LSB decisions are made by internal timing signals. In this mode, $\overline{\text{RD}}$ is brought low before the internal conversion is completed. When the internal conversion is completed, $\overline{\text{INT}}$ will be forced low and data will appear on DB0-DB9 as long as $\overline{\text{RD}}$ is still low. $\overline{\text{INT}}$ is reset high and DB0-DB9 is reset to high impedance on the rising edge of $\overline{\text{RD}}$.

Write-Read Operation is intended for applications where $\overline{\text{RD}}$ controls the transfer of data to a microprocessor.

1.6.3 READ MODE

Read Mode Operation is implemented by tying $\overline{\text{RD}}$ to $\overline{\text{WR}}$ and keeping $\overline{\text{RD}}$ and $\overline{\text{WR}}$ low long enough so that the conversion time is totally determined by the internal timing signals.

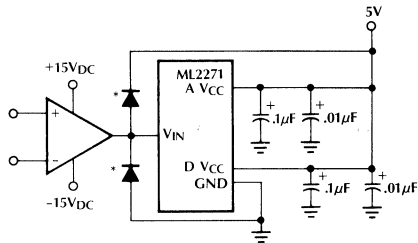
Timing for the Read Mode is shown in Figure 4. To do a conversion, $\overline{\text{CS}}$ must be low to select the device. The $\overline{\text{RD}}$ and $\overline{\text{WR}}$ falling edge starts the conversion. $\overline{\text{RD}}$ and $\overline{\text{WR}}$ is held low for the entire internal conversion. Thus, the MSB, ISB, and LSB comparisons along with the end of the acquisition period are made by internally generated timing signals. After the conversion is complete, $\overline{\text{INT}}$ goes low. Since $\overline{\text{RD}}$ is fixed low, DB0-DB9 will go from high impedance to active state as soon as $\overline{\text{INT}}$ goes low. $\overline{\text{INT}}$ is reset high and DB0-DB9 is reset to high impedance on rising edge of $\overline{\text{WR}}$ and $\overline{\text{RD}}$.

Read Mode Operation allows a conversion to be done with the device's own internal timing and thus, no external timing is needed.

1.6.4 POWER-ON RESET

When power is first applied, an internal power-on reset and timer circuit inhibits the CS input and resets the internal circuitry to prevent the ML2271 from starting in an unknown state. During this period of approximately 50μs, INT remains high and the data bus is in the high-impedance state.

2.0 TYPICAL APPLICATIONS



PROTECTION IS REQUIRED IF INPUT CURRENT > 25mA

Figure 8. Protecting the Input

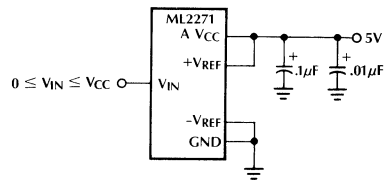


Figure 9. Using V_{CC} as Reference for Ratiometric Operation

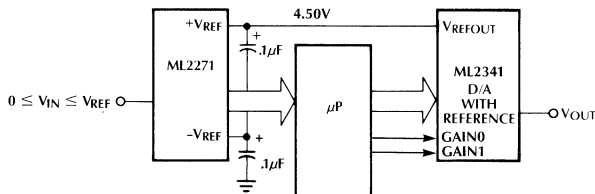


Figure 10. Using External Reference of D/A

2.0 TYPICAL APPLICATIONS (Continued)

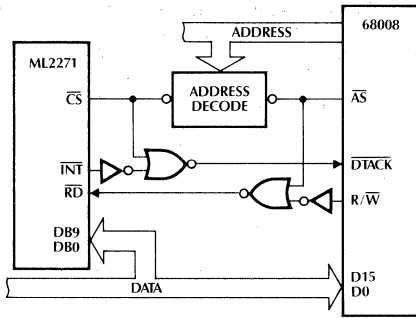


Figure 11. 68000 Type Interface to ML2271

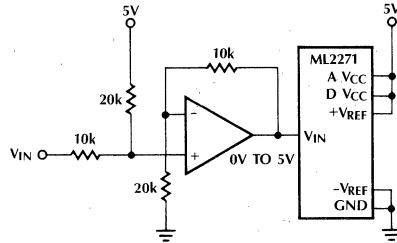


Figure 12. ±2.5V Analog Input Range

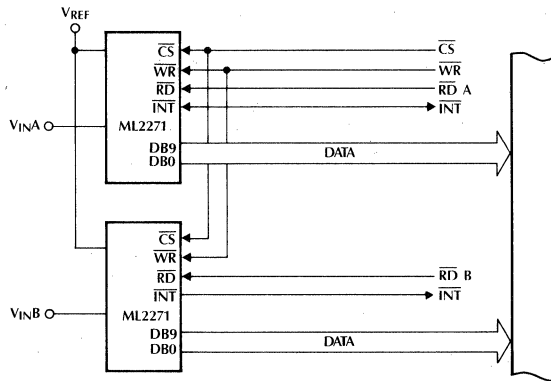


Figure 13. Simultaneous Sampling of Two Variables

2.0 TYPICAL APPLICATIONS (Continued)

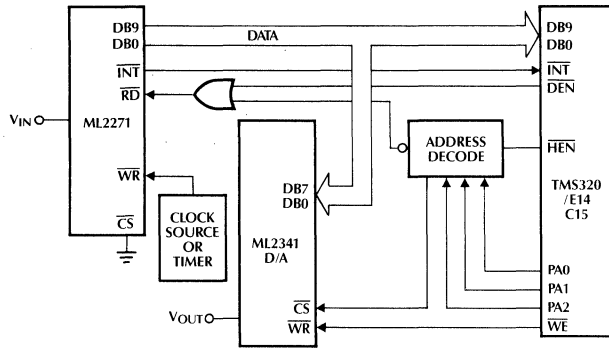


Figure 14. TMS320 Interface with D/A Output

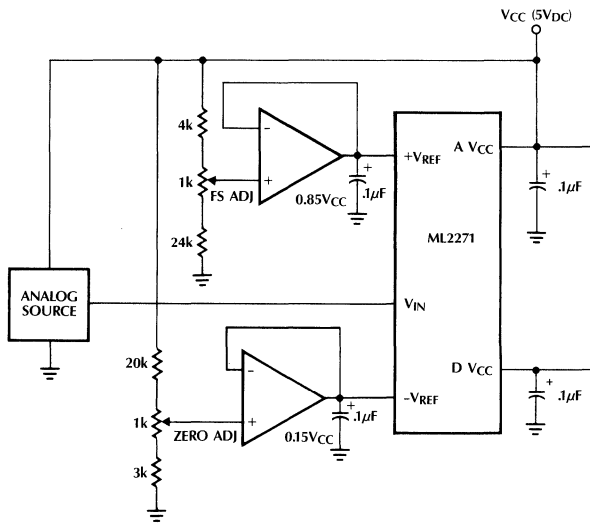


Figure 15. Operating with a Ratiometric Analog Signal of 15% of V_{CC} to 85% of V_{CC}

ML2271

ORDERING INFORMATION

PART NUMBER	LINEARITY ERROR	TEMPERATURE RANGE	PACKAGE
ML2271BMJ	$\pm\frac{1}{2}$ LSB	-55°C to +125°C	HERMETIC DIP (J20)
ML2271BIJ		-40°C to +85°C	HERMETIC DIP (J20)
ML2271BCP	± 1 LSB	0°C to +70°C	MOLDED DIP (P20)
ML2271BCS		0°C to +70°C	MOLDED SOIC (S20)
ML2271CMJ		-55°C to +125°C	HERMETIC DIP (J20)
ML2271CIJ		-40°C to +85°C	HERMETIC DIP (J20)
ML2271CCP		0°C to +70°C	MOLDED DIP (P20)
ML2271CCS		0°C to +70°C	MOLDED SOIC (S20)

Serial I/O 8-Bit A/D Converters

GENERAL DESCRIPTION

The ML2280 and ML2283 are 8-bit successive approximation A/D converters with serial I/O and configurable input multiplexers with up to 4 input channels.

All errors of the sample-and-hold incorporated on the ML2280 and ML2283, are accounted for in the analog-to-digital converters accuracy specification.

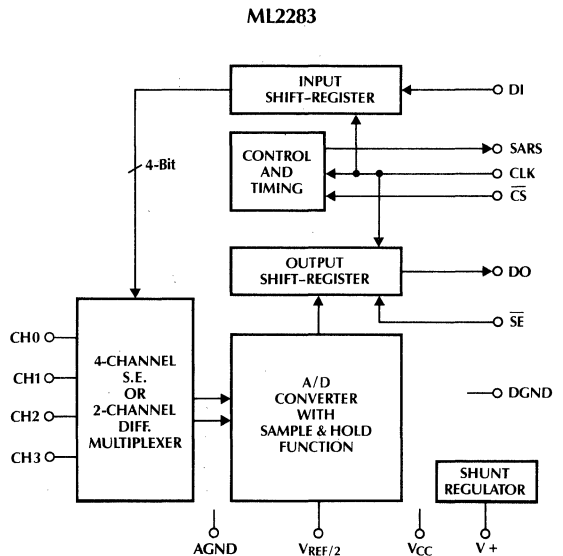
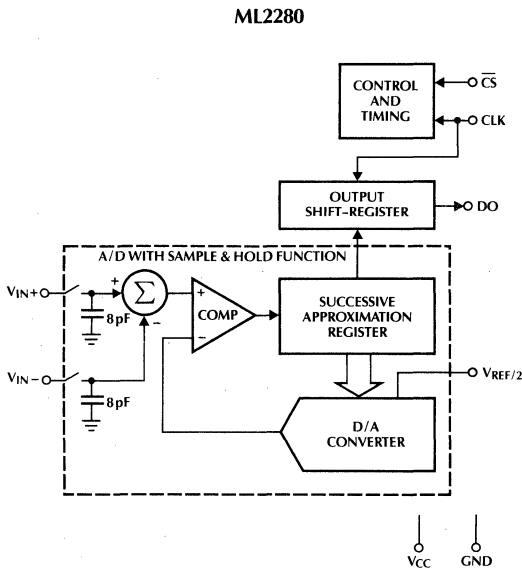
The voltage reference can be externally set to any value between GND and V_{CC} , thus allowing a full conversion over a relatively small voltage span if desired.

The ML2283 is an enhanced double polysilicon CMOS pin compatible second source for the ADC0833 A/D converter. All parameters are guaranteed over temperature with a power supply voltage of $5V \pm 10\%$.

FEATURES

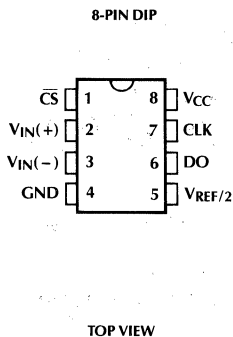
- Conversion time $6\mu s$
- ML2280 capable of digitizing a 5V, 40kHz sine wave
- Total unadjusted error with external reference $\pm 1/2$ LSB or ± 1 LSB
- Sample-and-hold 375 ns acquisition
- 0 to 5V analog input range with single 5V power supply
- 2.5V reference provides 0 to 5V analog input range
- No zero or full-scale adjust required
- Low power 12.5mW MAX
- Analog input protection 25 mA (min) per input
- Differential analog voltage inputs
- 0.3" width 8- or 14-pin DIP
- 4-channel input MUX option
- Superior pin compatible replacement for ADC0833

BLOCK DIAGRAMS

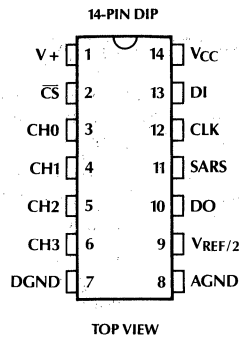


PIN CONNECTIONS

ML2280 Single Differential Input



ML2283 4-Channel MUX



PIN DESCRIPTION

NAME	FUNCTION
V _{CC}	Positive supply. 5volts \pm 10%
DGND	Digital ground. 0volts. All digital inputs and outputs are referenced to this point.
AGND	Analog ground. The negative reference voltage for A/D converter.
GND	Combined analog and digital ground.
CH0, V _{IN+} , V _{IN-}	Analog inputs. Digitally selected to be single ended (V _{IN}) or; V _{IN+} or V _{IN-} of a differential input. Analog range = GND \leq V _{IN} \leq V _{CC}
V _{REF/2}	Reference. The analog input range is twice the positive reference voltage value applied to this pin.
V+	Input to the Shunt Regulator.
DO	Data out. Digital output which contains result of A/D conversion. The serial data is clocked out on falling edges of CLK.

NAME	FUNCTION
SARS	Successive approximation register status. Digital output which indicates that a conversion is in progress. When SARS goes to 1, the sampling window is closed and conversion begins. When SARS goes to 0, conversion is completed. When CS = 1, SARS is in high impedance state.
CLK	Clock. Digital input which clocks data in on DI on rising edges and out on DO on falling edges. Also used to generate clocks for A/D conversion.
DI	Data input. Digital input which contains serial data to program the MUX and channel assignments.
CS	Chip select. Selects the chip for multiplexer and channel assignment and A/D conversion. When CS = 1, all digital outputs are in high impedance state. When CS = 0, normal A/D conversion takes place.

ABSOLUTE MAXIMUM RATINGS

(Note 1)

Current into V+	15 mA
Supply Voltage, V _{CC}	6.5 V
Voltage	
Logic Inputs	-7V to V _{CC} +7V
Analog Inputs	-0.3V to V _{CC} +0.3V
Input Current per Pin (Note 2)	±25 mA
Storage Temperature	-65°C to +150°C
Package Dissipation	
at T _A = 25°C (Board Mount)	800 mW
Lead Temperature (Soldering 10 sec.)	
Dual-In-Line Package (Molded)	260°C
Dual-In-Line Package (Ceramic)	300°C

OPERATING CONDITIONS

Supply Voltage, V _{CC}	4.5V _{DC} to 6.3V _{DC}
Temperature Range (Note 3)	T _{MIN} ≤ T _A ≤ T _{MAX}
ML2280BMJ, ML2283BMJ	-55°C to +125°C
ML2280BIJ, ML2283BIJ	-40°C to +85°C
ML2280CIJ, ML2283CIJ	
ML2280BCP, ML2283BCP	0°C to +70°C
ML2280CCP, ML2283CCP	

ELECTRICAL CHARACTERISTICS

Unless otherwise specified T_A = T_{MIN} to T_{MAX}, V_{CC} = 5V ± 10%, f_{CLK} = 1.333 MHz, and V_{REF/2} = 2.5V.

PARAMETER	NOTES	CONDITIONS	ML2280B, ML2283B			ML2280C, ML2283C			UNITS
			MIN	TYP NOTE 4	MAX	MIN	TYP NOTE 4	MAX	
CONVERTER AND MULTIPLEXER CHARACTERISTICS									
Total Unadjusted Error	5, 7	V _{REF/2} = 2.5V V _{REF/2} not Connected			± 1/2 ± 2			± 1 ± 2	LSB LSB
Reference Input Resistance, V _{REF/2}	5		3	5	7.5	3	5	7.5	kΩ
Common-Mode Input Range	5, 8		GND -0.05		V _{CC} +0.05	GND -0.05		V _{CC} +0.05	V
DC Common-Mode Error	6	Common Mode Voltage GND to V _{CC/2}		± 1/16	± 1/4		± 1/16	± 1/4	LSB
AC Common-Mode Error	6	Common Mode Voltage GND to V _{CC} , 0 to 50 kHz			± 1/4			± 1/4	LSB
DC Power Supply Sensitivity	6	V _{CC} = 5V ± 10% V _{REF} ≤ V _{CC} + 0.1V		± 1/32	± 1/4		± 1/32	± 1/4	LSB
AC Power Supply Sensitivity	6	100 mV _{p-p} , 25 kHz Sine on V _{CC}			± 1/4			± 1/4	LSB
Change in Zero Error from V _{CC} = 5V to Internal Zener Operation	6	15 mA into V+ V _{CC} = N.C. V _{REF/2} = 2.5V		± 1/2			± 1/2		LSB
V _Z , Internal Diode Regulated Breakdown (at V+)		15 mA into V+		6.9			6.9		V
V+ Input Resistance	5		20	35		20	35		kΩ
I _{Off} , Off Channel Leakage Current	5, 9	On Channel = V _{CC} Off Channel = 0V	-1			-1			μA
		On Channel = 0V Off Channel = V _{CC}			+1			+1	μA
I _{On} , On Channel Leakage Current	5, 9	On Channel = 0V Off Channel = V _{CC}	-1			-1			μA
		On Channel = V _{CC} Off Channel = 0V			+1			+1	μA

2

ML2280, ML2283

ELECTRICAL CHARACTERISTICS (Continued)

Unless otherwise specified $T_A = T_{MIN}$ to T_{MAX} , $V_{CC} = 5V \pm 10\%$, $f_{CLK} = 1.333\text{MHz}$, and $V_{REF/2} = 2.5V$.

PARAMETER	NOTES	CONDITIONS	MIN	TYP NOTE 4	MAX	LIMIT UNITS
DIGITAL AND DC CHARACTERISTICS						
$V_{IN(1)}$, Logical "1" Input Voltage	5		2.0			V
$V_{IN(0)}$, Logical "0" Input Voltage	5				0.8	V
$I_{IN(1)}$, Logical "1" Input Current	5	$V_{IN} = V_{CC}$			1	μA
$I_{IN(0)}$, Logical "0" Input Current	5	$V_{IN} = 0V$	-1			μA
$V_{OUT(1)}$, Logical "1" Output Voltage	5	$I_{OUT} = -2\text{mA}$	4.0			V
$V_{OUT(0)}$, Logical "0" Output Voltage	5	$I_{OUT} = 2\text{mA}$			0.4	V
I_{OUT} , Hi-Z Output Current	5	$V_{OUT} = 0V$ $V_{OUT} = V_{CC}$	-1		1	μA μA
I_{SOURCE} , Output Source Current	5	$V_{OUT} = 0V$	-6.5			mA
I_{SINK} , Output Sink Current	5	$V_{OUT} = V_{CC}$			8.0	mA
I_{CC} , Supply Current	5			1.3	2.5	mA
AC ELECTRICAL CHARACTERISTICS						
f_{CLK} , Clock Frequency	5		10		1333	kHz
t_{ACQ} , Sample-and-Hold Acquisition				1/2		$1/f_{CLK}$
t_C , Conversion Time		Not including MUX Addressing Time		8		$1/f_{CLK}$
SNR, Signal to Noise Ratio ML2280	12	$V_{IN} = 40\text{kHz}$, 5V Sine. $f_{CLK} = 1.333\text{MHz}$ ($f_{SAMPLING} \cong 120\text{kHz}$). Noise is Sum of All Nonfundamental Components up to $1/2$ of $f_{SAMPLING}$		47		dB
THD, Total Harmonic Distortion ML2280	12	$V_{IN} = 40\text{kHz}$, 5V Sine. $f_{CLK} = 1.333\text{MHz}$ ($f_{SAMPLING} \cong 120\text{kHz}$). THD is Sum of 2, 3, 4, 5 Harmonics Relative to Fundamental		-60		dB
IMD, Intermodulation Distortion ML2280	12	$V_{IN} = f_A + f_B$, $f_A = 40\text{kHz}$, 2.5V Sine. $f_B = 39.8\text{kHz}$, 2.5V Sine, $f_{CLK} = 1.333\text{MHz}$ ($f_{SAMPLING} \cong 120\text{kHz}$). IMD is ($f_A + f_B$), ($f_A - f_B$), ($2f_A + f_B$), ($2f_A - f_B$), ($f_A + 2f_B$), ($f_A - 2f_B$) Relative to Fundamental		-60		dB
Clock Duty Cycle	5, 10		40		60	%
t_{SETUP} , CS Falling Edge or Data Input Valid to CLK Rising Edge	5		130			ns
t_{HOLD} , Data Input Valid after CLK Rising Edge	5		80			ns

ELECTRICAL CHARACTERISTICS (Continued)Unless otherwise specified $T_A = T_{MIN}$ to T_{MAX} , $V_{CC} = 5V \pm 10\%$, $f_{CLK} = 1.333\text{MHz}$, and $V_{REF/2} = 2.5V$

PARAMETER	NOTES	CONDITIONS	MIN	TYP NOTE 4	MAX	LIMIT UNITS
AC ELECTRICAL CHARACTERISTICS						
t_{pd1} , t_{pd0} — CLK Falling Edge to Output Data Valid	5, 11	$C_L = 100\text{pF}$ Data MSB First		90	200	ns
		Data LSB First		50	110	ns
t_{IH} , t_{OH} — Rising Edge of CS to Data Output and SARS Hi-Z	6	$C_L = 10\text{pF}$, $R_L = 10\text{k}$ (see High Impedance Test Circuits)		40	90	ns
	6	$C_L = 100\text{pF}$, $R_L = 2\text{k}$		80	160	ns
C_{IN} , Capacitance of Logic Input				5		pF
C_{OUT} , Capacitance of Logic Outputs				5		pF

Note 1: Absolute maximum ratings are limits beyond which the life of the integrated circuit may be impaired. All voltages unless otherwise specified are measured with respect to ground.

Note 2: When the input voltage (V_{IN}) at any pin exceeds the power supply rails ($V_{IN} < \text{GND}$ or $V_{IN} > V_{CC}$) the absolute value of current at that pin should be limited to 25mA or less.

Note 3: -55°C to $+125^\circ\text{C}$ operating temperature range devices are 100% tested at temperature extremes with worst-case test conditions. 0°C to 70°C and -40°C to $+85^\circ\text{C}$ operating temperature range devices are 100% tested with temperature limits guaranteed by 100% testing, sampling, or by correlation with worst-case test conditions.

Note 4: Typicals are parametric norm at 25°C .

Note 5: Parameter guaranteed and 100% production tested.

Note 6: Parameter guaranteed. Parameters not 100% tested are not in outgoing quality level calculation.

Note 7: Total unadjusted error includes offset, full-scale, linearity, multiplexer and sample-and-hold errors.

Note 8: For $V_{IN}(-) \geq V_{IN}(+)$ the digital output code will be 0000 0000. Two on-chip diodes are tied to each analog input (See Block Diagram) which will forward conduct for analog input voltages one diode drop below ground or one diode drop greater than the V_{CC} supply. Be careful, during testing at low V_{CC} levels (4.5V), as high level analog inputs (5V) can cause this input diode to conduct-especially at elevated temperatures, and cause errors for analog inputs near full-scale. The spec allows 50mV forward bias of either diode. This means that as long as the analog V_{IN} or V_{REF} does not exceed the supply voltage by more than 50mV, the output code will be correct. To achieve an absolute 0V to 5V input voltage range will therefore require a minimum supply voltage of $4.950 V_{DC}$ over temperature variations, initial tolerance and loading.

Note 9: Leakage current is measured with the clock not switching.

Note 10: A 40% to 60% clock duty cycle range insures proper operation at all clock frequencies. In the case that an available clock has a duty cycle outside of these limits, the minimum, time the clock is high or the minimum time the clock is low must be at least 300ns. The maximum time the clock can be high or low is $60\mu\text{s}$.

Note 11: Since data, MSB first, is the output of the comparator used in the successive approximation loop, an additional delay is built in (See Block Diagram) to allow for comparator response time.

Note 12: Because of multiplexer addressing, test conditions for the ML2283 is $V_{IN} = 30\text{kHz}$, 5V sine ($f_{\text{SAMPLING}} \approx 89\text{kHz}$).

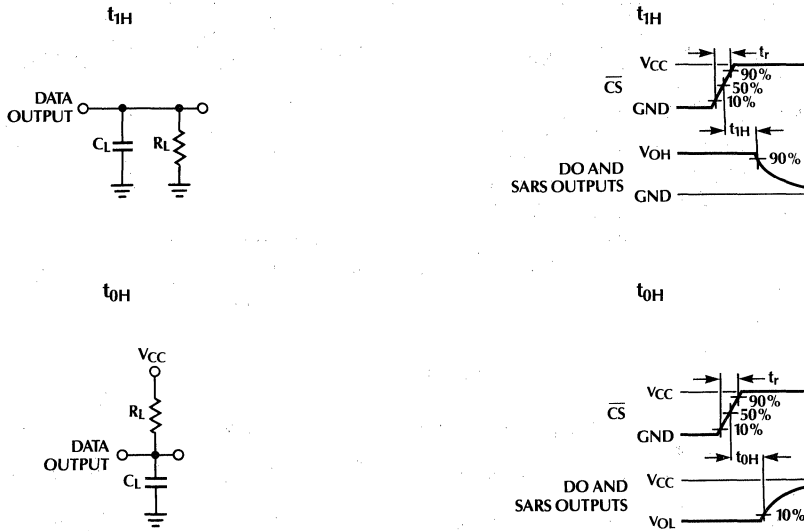


Figure 1. High Impedance Test Circuits and Waveforms

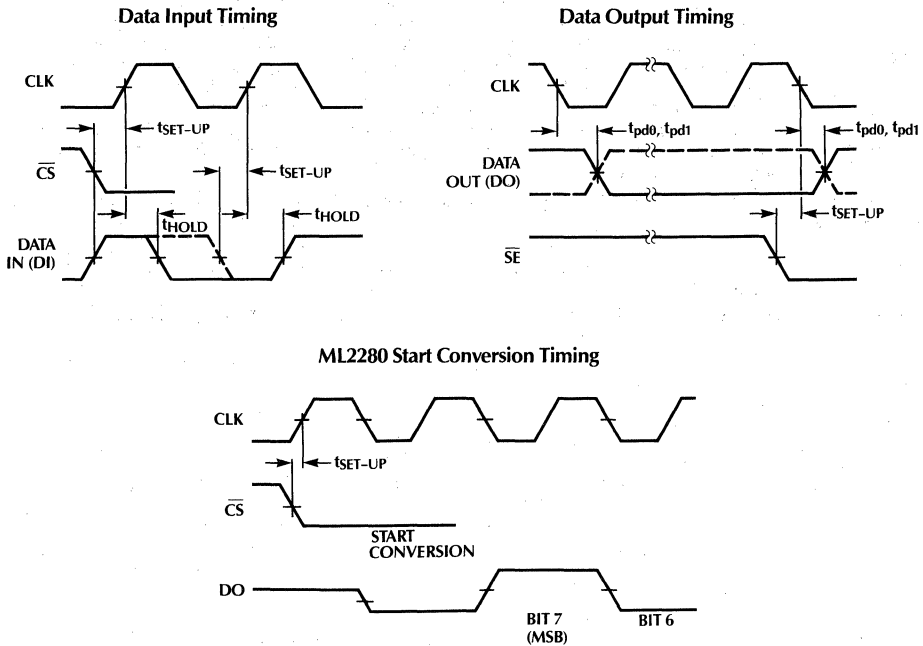
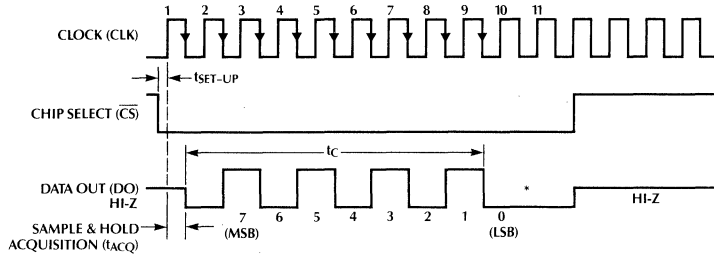


Figure 2. Timing Diagrams

ML2280 Timing



*LSB FIRST OUTPUT NOT AVAILABLE ON ML2280

2

ML2283 Timing

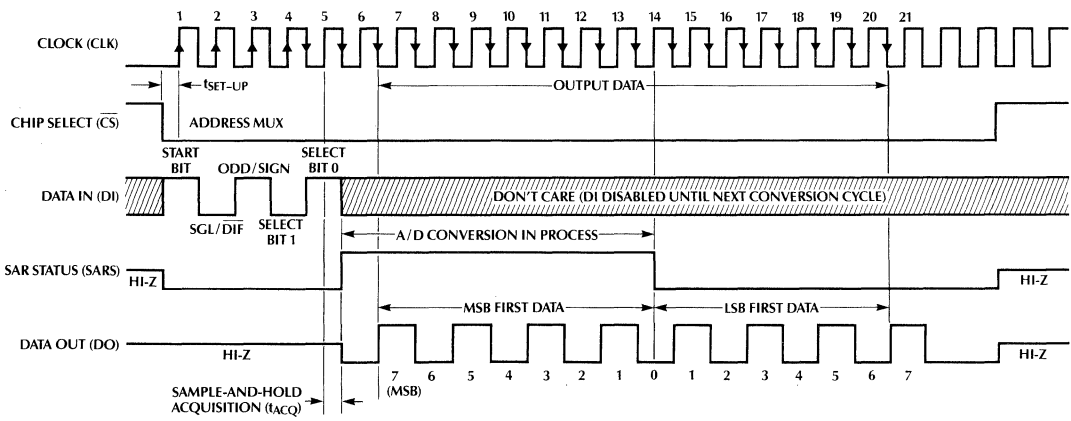


Figure 2. Timing Diagrams (Continued)

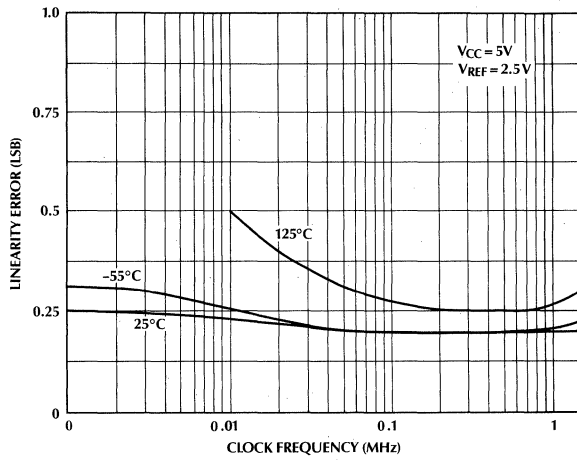


Figure 3. Linearity Error vs f_{CLK}

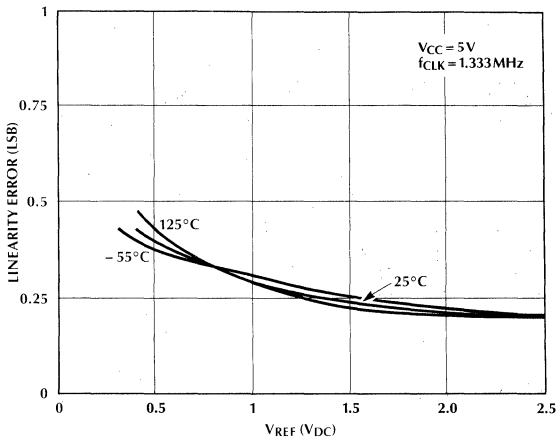


Figure 4. Linearity Error vs V_{REF} Voltage

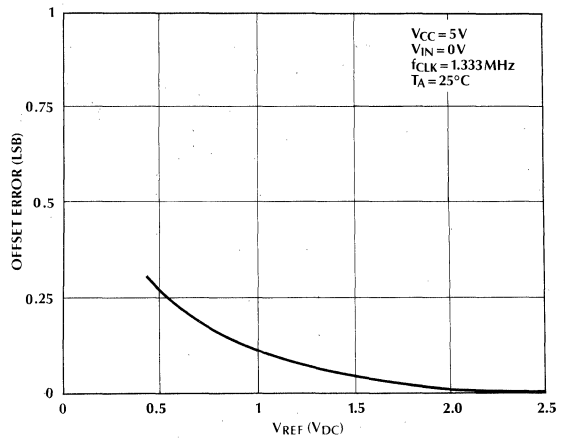
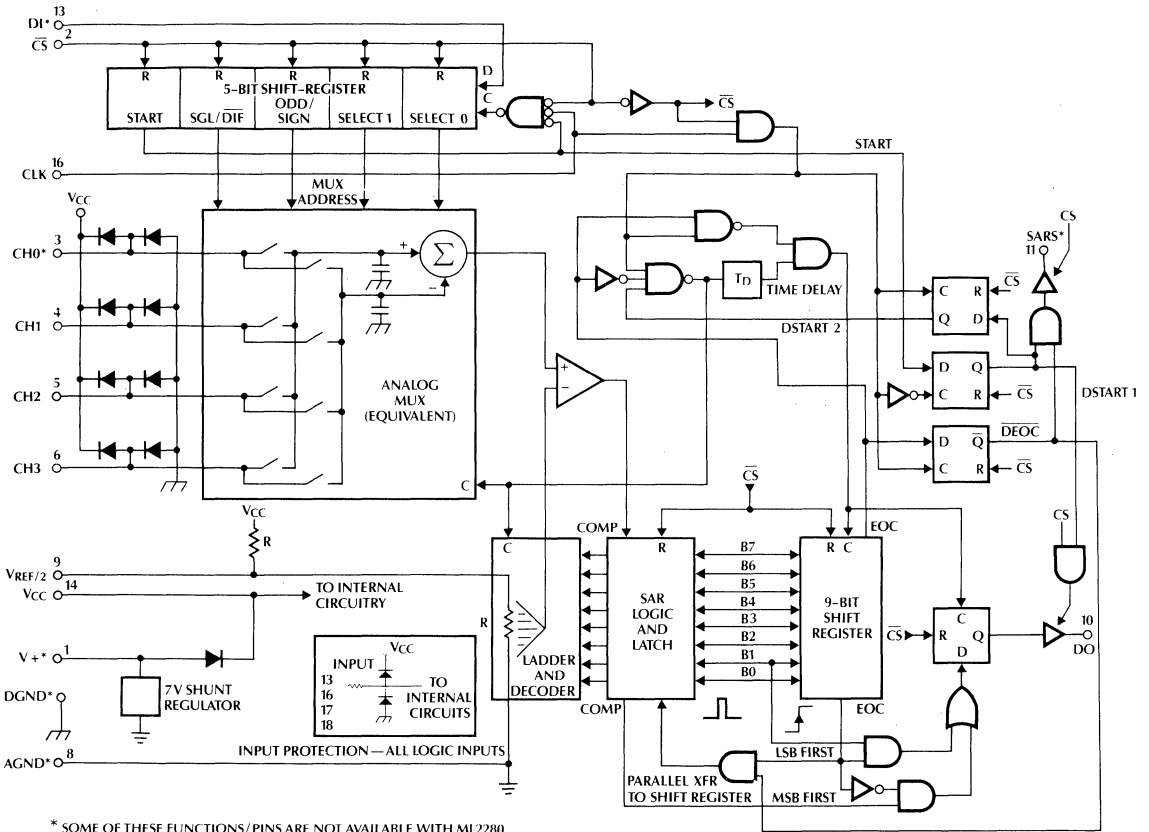


Figure 5. Unadjusted Offset Error vs V_{REF} Voltage



* SOME OF THESE FUNCTIONS/PINS ARE NOT AVAILABLE WITH ML2280.

Figure 6. ML2283 Functional Block Diagram

1.0 FUNCTIONAL DESCRIPTION

1.1 Multiplexer Addressing

The design of these converters utilizes a sample data comparator structure which provides for a differential analog input to be converted by a successive approximation routine.

The actual voltage converted is always the difference between an assigned "+" input terminal and a "-" input terminal. The polarity of each input terminal of the pair being converted indicates which line the converter expects to be the most positive. If the assigned "+" input is less than the "-" input, the converter responds with an all zeros output code.

A unique input multiplexing scheme has been utilized to provide multiple analog channels with software configurable single ended, or differential options.

A particular input configuration is assigned during the MUX addressing sequence, prior to the start of a conversion. The MUX address selects which of the analog inputs are to be enabled and whether this input is single ended or differential. In the differential case, it also assigns the polarity of the analog channels. Differential inputs are restricted to adjacent channel pairs. For example, channel 0 and channel 1 may be selected as a different pair but channel 0 or channel 1 cannot act differentially with any other channel. In addition to selecting the differential mode, the sign may also be selected. Channel 0 may be selected as the positive input and channel 1 as the negative input or vice versa. This programmability is illustrated by the MUX addressing codes shown in *Table 1*.

The MUX address is shifted into the converter via the DI input. Since the ML2280 contains only one differential input channel with a fixed polarity assignment, it does not require addressing.

Since the input configuration is under software control, it can be modified, as required, at each conversion. A channel can be treated as a single-ended, ground referenced input for one conversion; then it can be reconfigured as part of a differential channel for another conversion. *Figure 7* illustrates these different input modes.

Table 1. ML2283 MUX Addressing 4 Single-Ended or 2 Differential Channel

MUX Address				Channel #			
SGL/ DIF	ODD/ SIGN	SELECT		0	1	2	3
		1	0				
1	0	0	1	+			
1	0	1	1			+	
1	1	0	1		+		
1	1	1	1				+

MUX Address				Channel #			
SGL/ DIF	ODD/ SIGN	SELECT		0	1	2	3
		1	0				
0	0	0	1	+	-		
0	0	1	1			+	-
0	1	0	1	-	+		
0	1	1	1			-	+

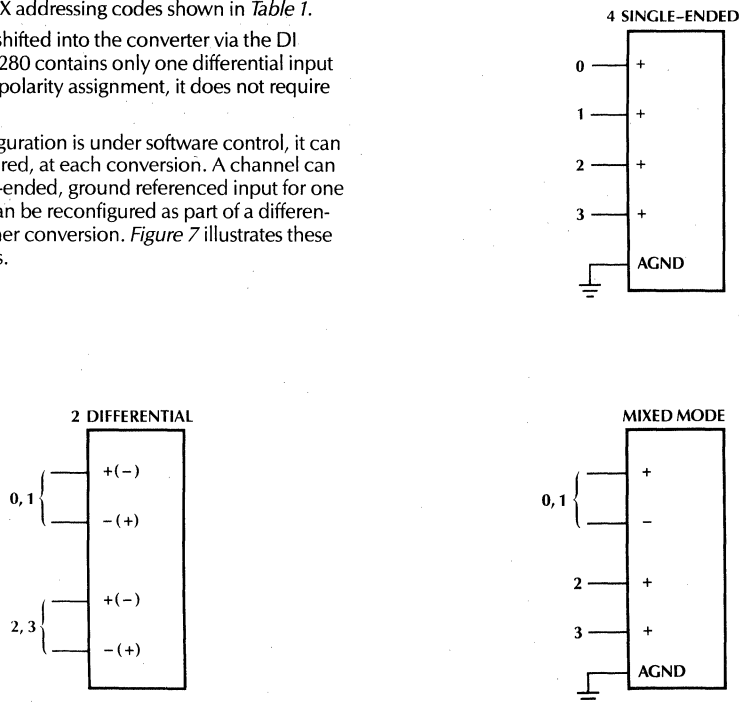


Figure 7. Analog Input Multiplexer Functional Options for ML2283

1.2 Digital Interface

The block diagram and timing diagrams in *Figures 2-5* illustrate how a conversion sequence is performed.

A conversion is initiated when \overline{CS} is pulsed low. This line must be held low for the entire conversion. The converter is now waiting for a start bit and its MUX assignment word.

A clock is applied to the CLK input. On each rising edge of the clock, the data on DI is clocked into the MUX address shift register. The start bit is the first logic "1" that appears on the DI input (all leading edge zeros are ignored). After the start bit, the device clocks in the next 2 to 4 bits for the MUX assignment word.

When the start bit has been shifted into the start location of the MUX register, the input channel has been assigned and a conversion is about to begin. An interval of 1/2 clock period is used for sample-and-hold settling through the selected MUX channels. The SAR status output goes high at this time to signal that a conversion is now in progress and the DI input is ignored.

The DO output comes out of high impedance and provides a leading zero for this one clock period.

When the conversion begins, the output of the comparator, which indicates whether the analog input is greater than or less than each successive voltage from the internal DAC, appears at the DO output on each falling edge of the clock. This data is the result of the conversion being shifted out (with MSB coming first) and can be read by external logic or μP immediately.

After 8 clock periods, the conversion is completed. The SAR status line returns low to indicate this 1/2 clock cycle later.

The serial data is always shifted out MSB first during the conversion. After the conversion has been completed, the data is shifted out a second time with LSB first. The ML2280 data is shifted out only once, MSB first.

All internal registers are cleared when the \overline{CS} input is high. If another conversion is desired, \overline{CS} must make a high to low transition followed by address information.

The DI input and DO output can be tied together and con-

trolled through a bidirectional μP I/O bit with one connection. This is possible because the DI input is only latched in during the MUX addressing interval while the DO output is still in the high impedance state.

1.3 Reference

The ML2280 and ML2283 are intended primarily for use in circuits requiring absolute accuracy. In this type of system, the analog inputs vary between very specific voltage limits and the reference voltage for the A/D converter must remain stable with time and temperature. For ratiometric applications, see the ML2281 and ML2284 which have a V_{REF} input that can be tied to V_{CC} .

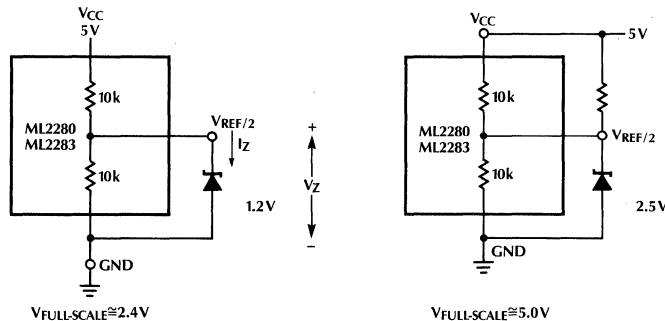
The voltage applied to the $V_{REF/2}$ pin defines the voltage span of the analog input (the difference between V_{IN+} and V_{IN-}) over which the 256 possible output codes apply. A full-scale conversion (an all 1s output code) will result when the voltage difference between a selected "+" input and "-" input is approximately twice the voltage at the $V_{REF/2}$ pin. This internal gain of 2 from the applied reference to the full-scale input voltage allows biasing a low voltage reference diode from the $5V_{DC}$ converter supply. To accommodate a 5V input span, only a 2.5V reference is required. The output code changes in accordance with the following equation:

$$\text{Output Code} = 256 \left(\frac{V_{IN(+)} - V_{IN(-)}}{2(V_{REF/2})} \right)$$

where the output code is the decimal equivalent of the 8-bit binary output (ranging from 0 to 255) and the term $V_{REF/2}$ is the voltage to ground.

The $V_{REF/2}$ pin is the center point of a two resistor divider (each resistor is 10k Ω) connected from V_{CC} to ground. Total ladder input resistance is the parallel combination of these two equal resistors. As shown in *Figure 8*, a reference diode with a voltage less than $V_{CC}/2$ can be connected without requiring an external biasing resistor if its current requirements meet the indicated level.

The minimum value of $V_{REF/2}$ can be quite small (See Typical Performance Curves) to allow direct conversions of transducer outputs providing less than a 5V output span. Particu-



NOTE: NO EXTERNAL BIASING RESISTOR NEEDED IF $V_Z < \frac{V_{CC}}{2}$ and $I_Z \text{ min.} < \frac{V_{CC}/2 - V_Z}{5k\Omega}$

Figure 8. Reference Biasing

lar care must be taken with regard to noise pickup, circuit layout and system error voltage sources when operating with a reduced span due to the increased sensitivity of the converter (1LSB equals $V_{REF}/256$).

1.4 Analog Inputs and Sample/Hold

An important feature of the ML2280 and ML2283 is that they can be located at the source of the analog signal and then communicate with a controlling μP with just a few wires. This avoids bussing the analog inputs long distances and thus reduces noise pickup on these analog lines. However, in some cases, the analog inputs have a large common mode voltage or even some noise present along with the valid analog signal.

The differential input of these converters reduces the effects of common mode input noise. Thus, if a common mode voltage is present on both “+” and “-” inputs, such as 60Hz, the converter will reject this common mode voltage since it only converts the difference between “+” and “-” inputs.

The ML2280 and ML2283 have a true sample-and-hold circuit which samples both “+” and “-” inputs simultaneously. This simultaneous sampling with a true S/H will give common mode rejection and AC linearity performance that is superior to devices where the two input terminals are not sampled at the same instant and where true sample-and-hold capability does not exist. Thus, these A/D converters can reject AC common mode signals from DC-50kHz as well as maintain linearity for signals from DC-50kHz.

The signal at the analog input is sampled during the interval when the sampling switch is closed prior to conversion start. The sampling window (S/H acquisition time) is $1/2$ CLK period wide and occurs $1/2$ CLK period before DO goes from high impedance to active low state. When the sampling switch closes at the start of the S/H acquisition time, 8 pF of capacitance is thrown onto the analog input. $1/2$ CLK period later, the sampling switch is opened and the signal present at the analog input is stored. Any error on the analog input at the end of the S/H acquisition time will cause additional conversion error. Care should be taken to allow adequate charging or settling time from the source. If more charging or settling time is needed to reduce these analog input errors, a longer CLK period can be used.

For latch-up immunity each analog input has dual diodes to the supply rails, and a minimum of ± 25 mA (± 100 mA typically) can be injected into each analog input without causing latch-up.

1.5 Zero Error Adjustment

The zero of the A/D does not require adjustment. If the minimum analog input voltage value, V_{INMIN} is not ground, a zero offset can be done. The converter can be made to out-

put 00000000 digital code for this minimum input voltage by biasing any V_{IN-} input at this V_{INMIN} value. This utilizes the differential mode operation of the A/D.

The zero error of the A/D converter relates to the location of the first riser of the transfer function and can be measured by grounding the V_{IN-} input and applying a small magnitude positive voltage to the V_{IN+} input. Zero error is the difference between the actual DC input voltage which is necessary to just cause an output digital code transition from 00000000 to 00000001 and the ideal $1/2$ LSB value ($1/2$ LSB = 9.8 mV for $V_{REF/2} = 2.500V_{DC}$).

1.6 Full-Scale Adjustment

The full-scale adjustment can be made by applying a differential input voltage which is $1/2$ LSB down from the desired analog full-scale voltage range and then adjusting the magnitude of the $V_{REF/2}$ input for a digital output code which is just changing from 11111110 to 11111111.

1.7 Adjustment for an Arbitrary Analog Input Voltage Range

If the analog zero voltage of the A/D is shifted away from ground (for example, to accommodate an analog input signal which does not go to ground), this new zero reference should be properly adjusted first. A V_{IN+} voltage which equals this desired zero reference plus $1/2$ LSB (where the LSB is calculated for the desired analog span, $1\text{LSB} = \text{analog span}/256$) is applied to selected “+” input and the zero reference voltage at the corresponding “-” input should then be adjusted to just obtain the 00000000 to 00000001 code transition.

The full-scale adjustment should be made by forcing a voltage to the V_{IN+} input which is given by:

$$V_{IN+} \text{ fs adjust} = V_{MAX} - 1.5 * [(V_{MAX} - V_{MIN}) / 256]$$

where V_{MAX} = high end of the analog input range
 V_{MIN} = low end (offset zero) of the analog range

The V_{REF} or V_{CC} voltage is then adjusted to provide a code change from 11111110 to 11111111.

1.8 Shunt Regulator

A unique feature of the ML2283 is the inclusion of a shunt regulator connected from $V+$ terminal to ground which also connects to the V_{CC} terminal (which is the actual converter supply) through a silicon diode as shown in Figure 9. When the regulator is turned on, the $V+$ voltage is clamped at $11 V_{BE}$ set by the internal resistor ratio. The typical I-V curve of the shunt regulator is shown in Figure 10. It should be noted that before $V+$ voltage is high enough to turn on the shunt regulator (which occurs at about 5.5V), 35 k Ω of resistance is observed between $V+$ and GND. When the shunt regulator is not used, $V+$ pin should be either left floating or tied to GND. The temperature coefficient of the regulator is -22 mV/ $^{\circ}$ C.

2.0 APPLICATIONS

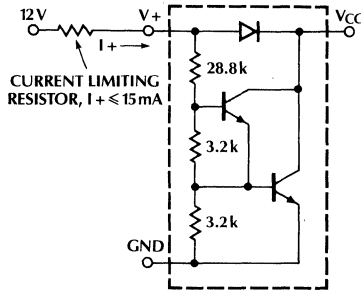


Figure 9. Shunt Regulator

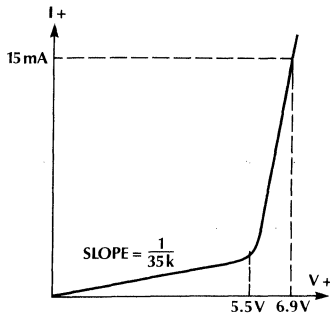
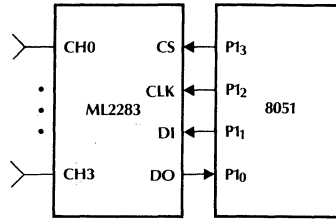


Figure 10. I-V Characteristic of the Shunt Regulator

8051 Interface and Controlling Software

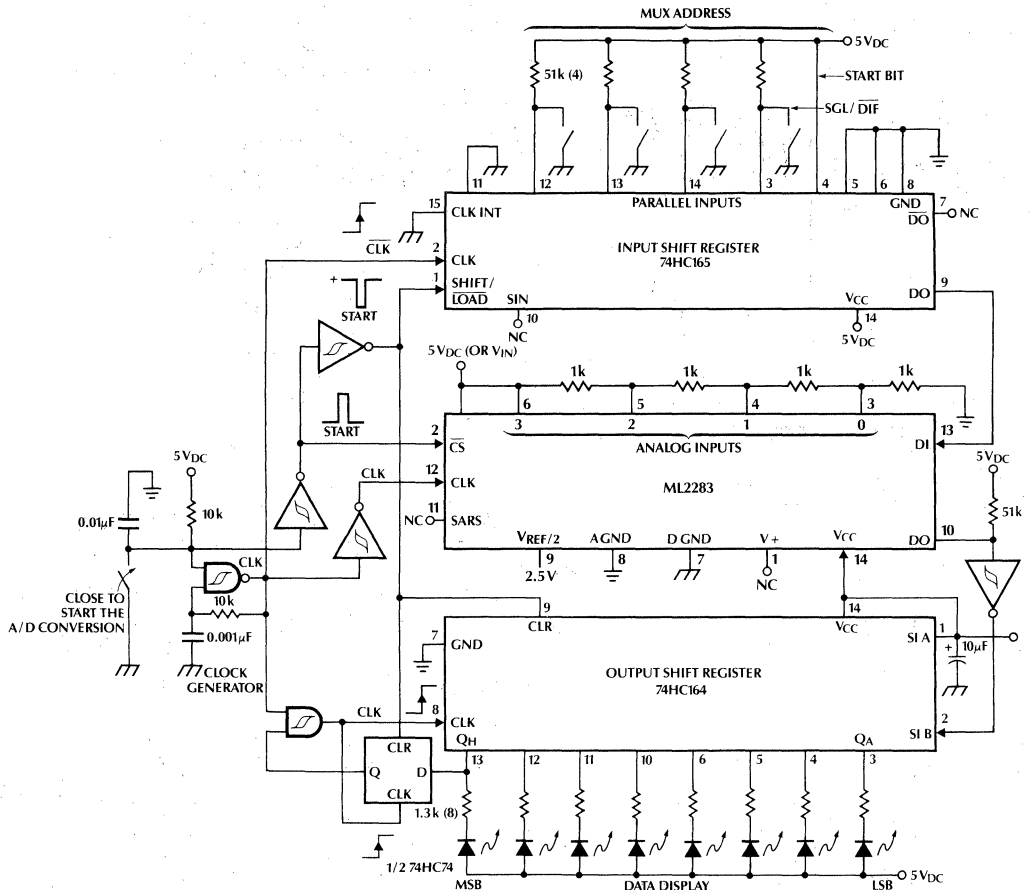


2

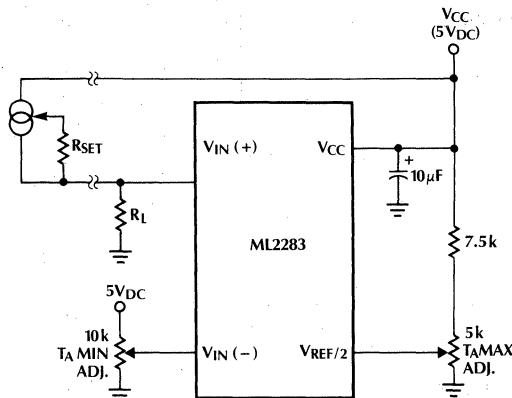
	Mnemonic		Instruction
START	ANL	P1, #0F7H	;SELECT A/D (CS = 0)
	MOV	B, #5	;BIT COUNTER ← 5
	MOV	A, #ADDR	;A ← MUX BIT
LOOP 1:	RRC	A	;CY ← ADDRESS BIT
	JC	ONE	;TEST BIT
			;BIT = 0
ZERO:	ANL	P1, #0FEH	;DI ← 0
	SJMP	CONT	;CONTINUE
			;BIT = 1
ONE:	ORL	P1, #1	;DI ← 1
CONT:	ACALL	PULSE	;PULSE SK 0 → 1 → 0
	DJNZ	B, LOOP 1	;CONTINUE UNTIL DONE
	ACALL	PULSE	;EXTRA CLOCK FOR SYNC
	MOV	B, #8	;BIT COUNTER ← 8
LOOP 2:	ACALL	PULSE	;PULSE SK 0 → 1 → 0
	MOV	A, P1	;CY ← DO
	RRC	A	
	RRC	A	
	MOV	A, C	;A ← RESULT
	RLC	A	;A(O) ← BIT AND SHIFT
	MOV	C, A	;C ← RESULT
	DJNZ	B, LOOP 2	;CONTINUE UNTIL DONE
RETI			;PULSE SUBROUTINE
PULSE:	ORL	P1, #04	;SK ← 1
	NOP		;DELAY
	ANL	P1, #0FBH	;SK ← 0
	RET		

APPLICATIONS (Continued)

ML2283 "Stand-Alone" or Evaluation Circuit

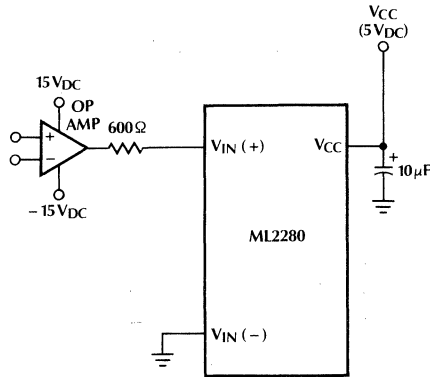


Low-Cost Remote Temperature Sensor



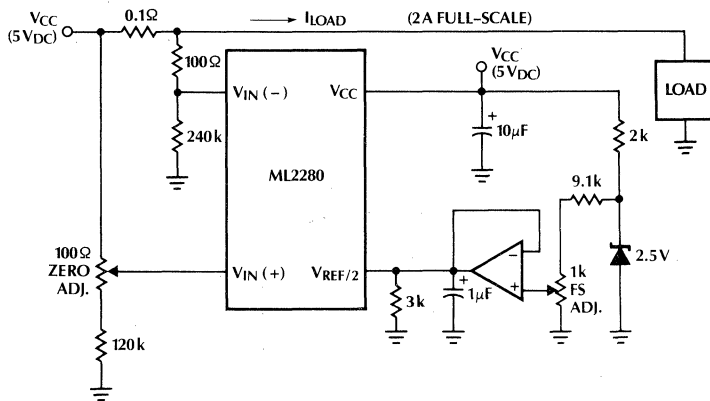
APPLICATIONS (Continued)

Protecting the Input



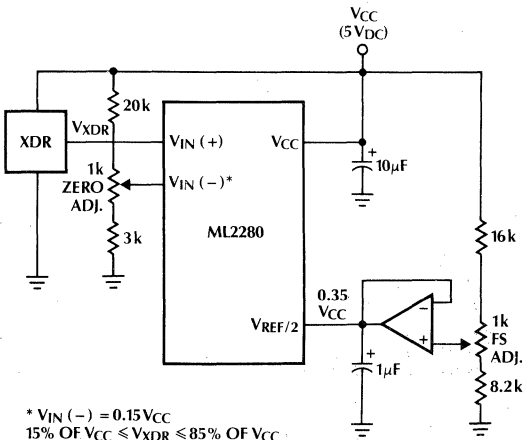
DIODE CLAMPING IS NOT NEEDED
IF CURRENT IS LIMITED TO 25mA

Digitizing a Current Flow

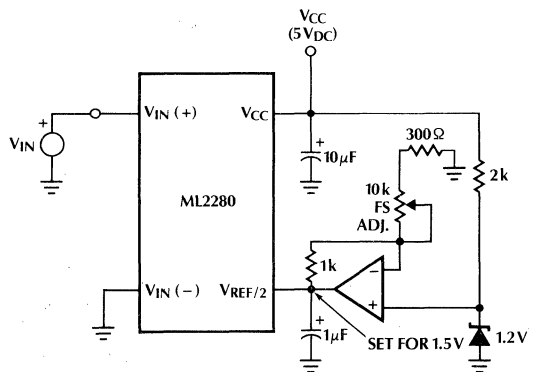


Operating with Ratiometric Transducers

Span Adjust: $0V \leq V_{IN} \leq 3V$



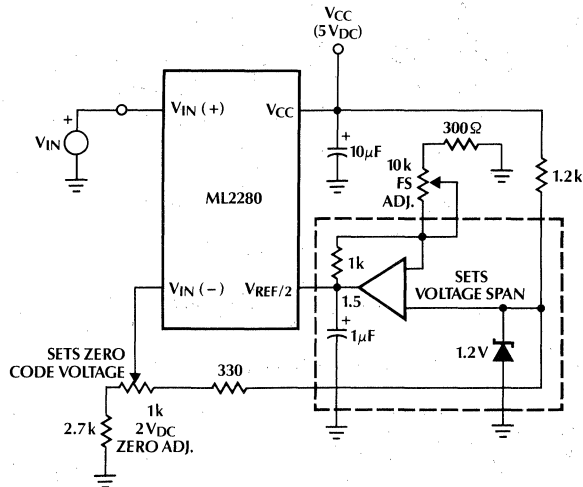
* $V_{IN} (-) = 0.15V_{CC}$
 $15\% \text{ OF } V_{CC} \leq V_{XDR} \leq 85\% \text{ OF } V_{CC}$



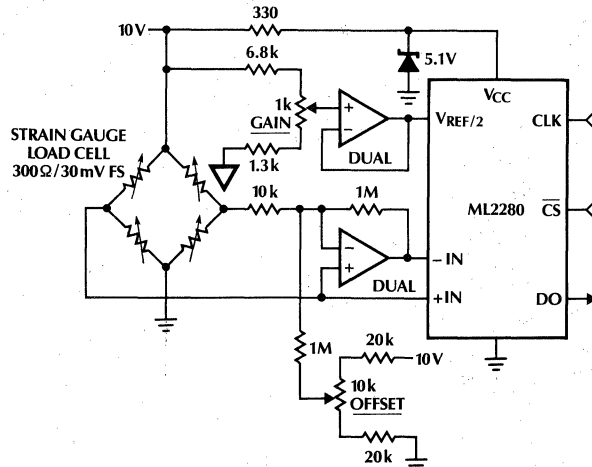
ML2280, ML2283

APPLICATIONS (Continued)

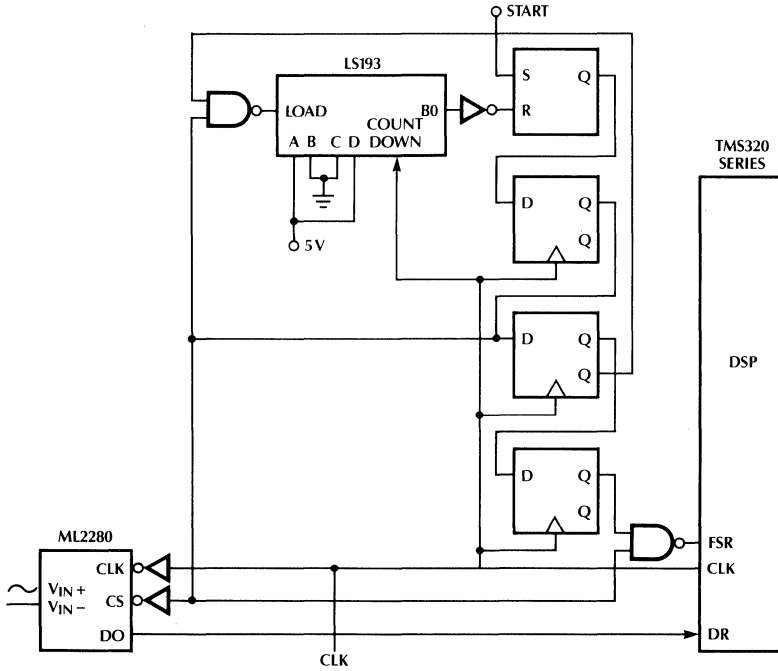
Zero-Shift and Span Adjust: $2V \leq V_{IN} \leq 5V$



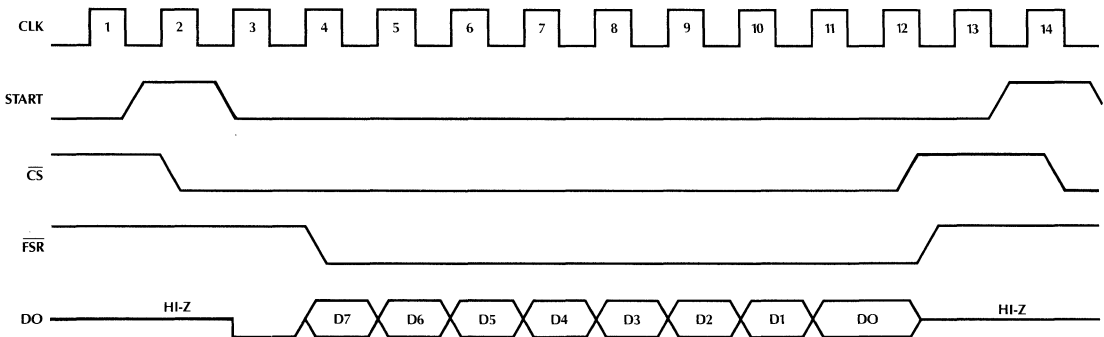
Digital Load Cell



- USES ONE MORE WIRE THAN LOAD CELL ITSELF
- TWO MINI-DIPs COULD BE MOUNTED INSIDE LOAD CELL FOR DIGITAL OUTPUT TRANSDUCER
- ELECTRONIC OFFSET AND GAIN TRIMS RELAX MECHANICAL SPECS FOR GAUGE FACTOR AND OFFSET
- LOW LEVEL CELL OUTPUT IS CONVERTED IMMEDIATELY FOR HIGH NOISE IMMUNITY



Sampling Rate 111KHz, Data Rate 1.33MHz



Interfacing ML2280 to TMS320 Series

ML2280, ML2283

ORDERING INFORMATION

PART NUMBER	ALTERNATE PART NUMBER	TOTAL UNADJUSTED ERROR	TEMPERATURE RANGE	PACKAGE
SINGLE ANALOG INPUT, 8-PIN PACKAGE				
ML2280BMJ ML2280BIJ ML2280BCP ML2280CIJ ML2280CCP		$\pm 1/2$ LSB	-55°C to +125°C -40°C to +85°C 0° to +70°C	HERMETIC DIP (J08) HERMETIC DIP (J08) MOLDED DIP (P08)
		± 1 LSB	-40°C to +85°C 0°C to +70°C	HERMETIC DIP (J08) MOLDED DIP (P08)
FOUR ANALOG INPUTS, 14-PIN PACKAGE				
ML2283BMJ ML2283BIJ ML2283BCP ML2283CIJ ML2283CCP	ADC0833BJ ADC0833BCJ ADC0833BCN ADC0833CCJ ADC0833CCN	$\pm 1/2$ LSB	-55°C to +125°C -40°C to +85°C 0° to +70°C	HERMETIC DIP (J14) HERMETIC DIP (J14) MOLDED DIP (P14)
		± 1 LSB	-40°C to +85°C 0°C to +70°C	HERMETIC DIP (J14) MOLDED DIP (P14)

Serial I/O 8-Bit A/D Converters with Multiplexer Options

GENERAL DESCRIPTION

The ML2281 family are 8-bit successive approximation A/D converters with serial I/O and configurable input multiplexers with up to 8 input channels.

All errors of the sample-and-hold, incorporated on the ML2281 family are accounted for in the analog-to-digital converters accuracy specification.

The voltage reference can be externally set to any value between GND and V_{CC} , thus allowing a full conversion over a relatively small voltage span if desired.

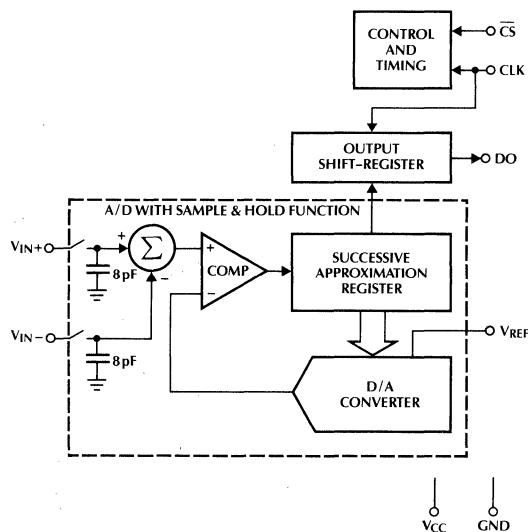
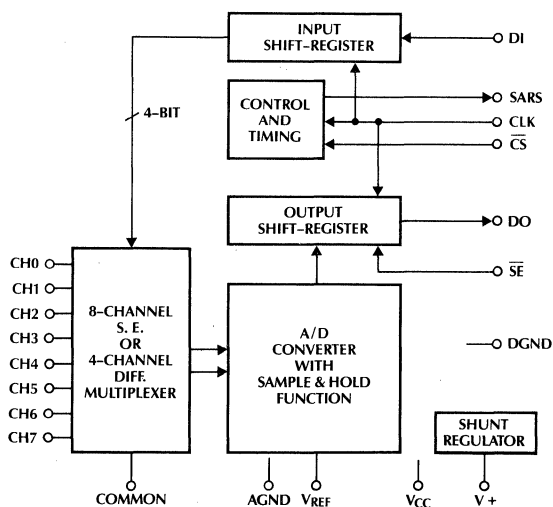
The ML2281 family is an enhanced double polysilicon CMOS pin compatible second source for the ADC0831, ADC0832, ADC0834, and ADC0838 A/D converters. The ML2281 series enhancements are faster conversion time, true sample-and-hold function, superior power supply rejection, improved AC common mode rejection, faster digital timing, and lower power dissipation. All parameters are guaranteed over temperature with a power supply voltage of $5V \pm 10\%$.

FEATURES

- Conversion time 6 μ s
- Total unadjusted error $\pm 1/2$ LSB or ± 1 LSB
- Sample-and-hold 375 ns acquisition
- 2, 4, or 8-input multiplexer options
- 0 to 5V analog input range with single 5V power supply
- Operates ratiometrically or with up to 5V voltage reference
- No zero or full-scale adjust required
- ML2281 capable of digitizing a 5V, 40kHz sine wave
- Low power 12.5mW MAX
- 0.3" width 8-, 14-, or 20-pin DIP
- 20-pin surface mount PCC ML2288
- Superior pin compatible replacement for ADC0831, ADC0832, ADC0834, and ADC0838
- Analog input protection 25mA (min) per input

2

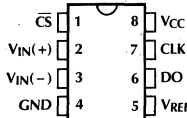
BLOCK DIAGRAMS

ML2281

ML2288


ML2281, ML2282, ML2284, ML2288

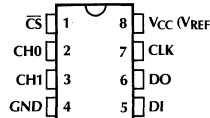
PIN CONNECTIONS

ML2281 Single Differential Input
8-PIN DIP



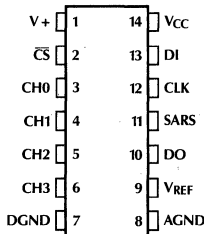
TOP VIEW

ML2282 2-Channel MUX
8-PIN DIP



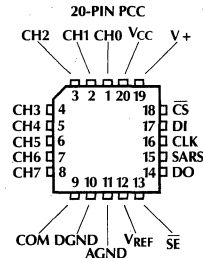
TOP VIEW

ML2284 4-Channel MUX
14-PIN DIP



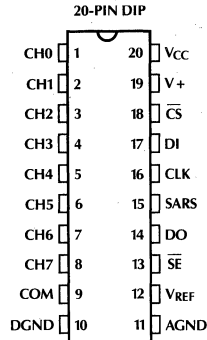
TOP VIEW

ML2288 8-Channel MUX



TOP VIEW

ML2288 8-Channel MUX



TOP VIEW

PIN DESCRIPTION

NAME	FUNCTION	NAME	FUNCTION
V_{CC}	Positive supply. 5 volts \pm 10%	DO	Data out. Digital output which contains result of A/D conversion. The serial data is clocked out on falling edges of CLK.
DGND	Digital ground. 0volts. All digital inputs and outputs are referenced to this point.	SARS	Successive approximation register status. Digital output which indicates that a conversion is in progress. When SARS goes to 1, the sampling window is closed and conversion begins. When SARS goes to 0, conversion is completed. When $\overline{CS}=1$, SARS is in high impedance state.
AGND	Analog ground. The negative reference voltage for A/D converter.	CLK	Clock. Digital input which clocks data in on DI on rising edges and out on DO on falling edges. Also used to generate clocks for A/D conversion.
CH0-7, V_{IN+} , V_{IN-}	Analog inputs. Digitally selected to be single ended (V_{IN-}) or; V_{IN+} or V_{IN-} of a differential input. Analog range = $GND \leq V_{IN} \leq V_{CC}$	DI	Data input. Digital input which contains serial data to program the MUX and channel assignments.
COM	Common reference point for analog inputs. A/D conversion is performed on voltage difference between analog input and this common reference point if single-end conversion is specified.	\overline{CS}	Chip select. Selects the chip for multiplexer and channel assignment and A/D conversion. When $\overline{CS}=1$, all digital outputs are in high impedance state. When $\overline{CS}=0$, normal A/D conversion takes place.
V_{REF}	Reference. The positive reference voltage for A/D converter.		
\overline{SE}	Shift enable. Input controls whether LSB first bit stream is shifted out on serial output DO. If $\overline{SE}=1$, MSB first is shifted out only. If $\overline{SE}=0$, an MSB first bit stream with LSB first is shifted out after end of conversion.		
$V+$	Input to the Shunt Regulator.		

ABSOLUTE MAXIMUM RATINGS

(Note 1)

Current into V+	15 mA
Supply Voltage, V _{CC}	6.5 V
Voltage	
Logic Inputs	-7 V to V _{CC} +7 V
Analog Inputs	-0.3 V to V _{CC} +0.3 V
Input Current per Pin (Note 2)	±25 mA
Storage Temperature	-65°C to +150°C
Package Dissipation	
at T _A = 25°C (Board Mount)	800 mW
Lead Temperature (Soldering 10 sec.)	
Dual-In-Line Package (Molded)	260°C
Dual-In-Line Package (Ceramic)	300°C
Molded Chip Carrier Package	
Vapor Phase (60 sec.)	215°C
Infrared (15 sec.)	220°C

OPERATING CONDITIONS

Supply Voltage, V _{CC}	4.5 V _{DC} to 6.3 V _{DC}
Temperature Range (Note 3)	T _{MIN} ≤ T _A ≤ T _{MAX}
ML2281/2/4/8 BMJ	-55°C to +125°C
ML2281/2/4/8 CMJ	
ML2281/2/4/8 BIJ	-40°C to +85°C
ML2281/2/4/8 CIJ	
ML2281/2/4/8 BCP	0°C to +70°C
ML2281/2/4/8 CCP	

ELECTRICAL CHARACTERISTICS

Unless otherwise specified T_A = T_{MIN} to T_{MAX}, V_{CC} = V_{REF} = 5 V ± 10%, and f_{CLK} = 1.333 MHz

PARAMETER	NOTES	CONDITIONS	ML2281B, ML2282B ML2284B, ML2288B			ML2281C, ML2282C ML2284C, ML2288C			UNITS
			MIN	TYP NOTE 4	MAX	MIN	TYP NOTE 4	MAX	
CONVERTER AND MULTIPLEXER CHARACTERISTICS									
Total Unadjusted Error	5, 7	V _{REF} = V _{CC}			± 1/2			± 1	LSB
Reference Input Resistance	5, 8		6	10	15	6	10	15	kΩ
Common-Mode Input Range	5, 9		GND -0.05		V _{CC} +0.05	GND -0.05		V _{CC} +0.05	V
DC Common-Mode Error	6	Common Mode Voltage GND to V _{CC/2}		± 1/16	± 1/4		± 1/16	± 1/4	LSB
AC Common-Mode Error	6	Comon Mode Voltage GND to V _{CC/2} , 0 to 50 kHz			± 1/4			± 1/4	LSB
DC Power Supply Sensitivity	6	V _{CC} = 5 V ± 10% - V _{REF} ≤ V _{CC} + 0.1 V		± 1/32	± 1/4		± 1/32	± 1/4	LSB
AC Power Supply Sensitivity	6	100 mV _{p-p} , 25 kHz sine on V _{CC}			± 1/4			± 1/4	LSB
Change in Zero Error from V _{CC} = 5 V to Internal Zener Operation	6	15 mA into V+ V _{CC} = N.C. V _{REF} = 5 V		± 1/2			± 1/2		LSB
V _Z , Internal Diode Regulated Breakdown (at V+)		15 mA into V+		6.9			6.9		V
V+ Input Resistance	5		20	35		20	35		kΩ
I _{Off} , Off Channel Leakage Current	5, 10	On Channel = V _{CC} Off Channel = 0 V	-1			-1			μA
		On Channel = 0 V Off Channel = V _{CC}			+1			+1	μA
I _{On} , On Channel Leakage Current	5, 10	On Channel = 0 V Off Channel = V _{CC}	-1			-1			μA
		On Channel = V _{CC} Off Channel = 0 V			+1			+1	μA

2

ML2281, ML2282, ML2284, ML2288

ELECTRICAL CHARACTERISTICS (Continued)

Unless otherwise specified $T_A = T_{MIN}$ to T_{MAX} , $V_{CC} = V_{REF} = 5V \pm 10\%$, and $f_{CLK} = 1.333\text{MHz}$

PARAMETER	NOTES	CONDITIONS	ML2281B, ML2282B ML2284B, ML2288B			ML2281C, ML2282C ML2284C, ML2288C			UNITS
			MIN	TYP NOTE 4	MAX	MIN	TYP NOTE 4	MAX	
DIGITAL AND DC CHARACTERISTICS									
$V_{IN(1)}$, Logical "1" Input Voltage	5		2.0			2.0			V
$V_{IN(0)}$, Logical "0" Input Voltage	5				0.8			0.8	V
$I_{IN(1)}$, Logical "1" Input Current	5	$V_{IN} = V_{CC}$			1			1	μA
$I_{IN(0)}$, Logical "0" Input Current	5	$V_{IN} = 0V$	-1			-1			μA
$V_{OUT(1)}$, Logical "1" Output Voltage	5	$I_{OUT} = -2\text{mA}$	4.0			4.0			V
$V_{OUT(0)}$, Logical "0" Output Voltage	5	$I_{OUT} = 2\text{mA}$			0.4			0.4	V
I_{OUT} , Hi-Z Output Current	5	$V_{OUT} = 0V$ $V_{OUT} = V_{CC}$	-1		1	-1		1	μA μA
I_{SOURCE} , Output Source Current	5	$V_{OUT} = 0V$	-6.5			-6.5			mA
I_{SINK} , Output Sink Current	5	$V_{OUT} = V_{CC}$			8.0			8.0	mA
I_{CC} , Supply Current ML2281, ML2284, ML2288	5			1.3	2.5		1.3	2.5	mA
ML2282	5	Includes Ladder Current		1.8	3.5		1.8	3.5	mA

PARAMETER	NOTES	CONDITIONS	MIN	TYP NOTE 4	MAX	LIMIT UNITS
AC ELECTRICAL CHARACTERISTICS						
f_{CLK} , Clock Frequency	5		10		1333	kHz
t_{ACQ} , Sample-and-Hold Acquisition				1/2		$1/f_{CLK}$
t_C , Conversion Time		Not including MUX Addressing Time		8		$1/f_{CLK}$
SNR, Signal to Noise Ratio ML2281	12	$V_{IN} = 40\text{kHz}$, 5V Sine. $f_{CLK} = 1.333\text{MHz}$ ($f_{SAMPLING} \cong 120\text{kHz}$). Noise is Sum of All Nonfundamental Components up to $1/2$ of $f_{SAMPLING}$		47		dB
THD, Total Harmonic Distortion ML2281	12	$V_{IN} = 40\text{kHz}$, 5V Sine. $f_{CLK} = 1.333\text{MHz}$ ($f_{SAMPLING} \cong 120\text{kHz}$). THD is Sum of 2, 3, 4, 5 Harmonics Relative to Fundamental		-60		dB

ELECTRICAL CHARACTERISTICS

Unless otherwise specified $T_A = T_{MIN}$ to T_{MAX} , $V_{CC} = V_{REF} = 5V \pm 10\%$, and $f_{CLK} = 1.333\text{ MHz}$

PARAMETER	NOTES	CONDITIONS	MIN	TYP NOTE 4	MAX	LIMIT UNITS
AC ELECTRICAL CHARACTERISTICS (Continued)						
IMD, Intermodulation Distortion ML2281	12	$V_{IN} = f_A + f_B$, $f_A = 40\text{ kHz}$, 2.5V Sine. $f_B = 39.8\text{ kHz}$, 2.5V Sine, $f_{CLK} = 1.333\text{ MHz}$ ($f_{SAMPLING} \approx 120\text{ kHz}$). IMD is $(f_A + f_B)$, $(f_A - f_B)$, $(2f_A + f_B)$, $(2f_A - f_B)$, $(f_A + 2f_B)$, $(f_A - 2f_B)$ Relative to Fundamental		-60		dB
Clock Duty Cycle	5, 11		40		60	%
t_{SETUP} , CS Falling Edge or Data Input Valid to CLK Rising Edge	5		130			ns
t_{HOLD} , Data Input Valid after CLK Rising Edge	5		80			ns
t_{pd1} , t_{pd0} — CLK Falling Edge to Output Data Valid	5, 13	$C_L = 100\text{ pF}$ Data MSB First Data LSB First		90 50	200 110	ns ns
t_{IH} , t_{OH} , — Rising Edge of CS to Data Output and SARS Hi-Z	6	$C_L = 10\text{ pF}$, $R_L = 10\text{ k}$ (see High Impedance Test Circuits)		40	90	ns
	5	$C_L = 100\text{ pF}$, $R_L = 2\text{ k}$		80	160	ns
C_{IN} , Capacitance of Logic Input				5		pF
C_{OUT} , Capacitance of Logic Outputs				5		pF

Note 1: Absolute maximum ratings are limits beyond which the life of the integrated circuit may be impaired. All voltages unless otherwise specified are measured with respect to ground.

Note 2: When the input voltage (V_{IN}) at any pin exceeds the power supply rails ($V_{IN} < GND$ or $V_{IN} > V_{CC}$) the absolute value of current at that pin should be limited to 25mA or less.

Note 3: -55°C to $+125^\circ\text{C}$ operating temperature range devices are 100% tested at temperature extremes with worst-case test conditions. 0°C to 70°C and -40°C to $+85^\circ\text{C}$ operating temperature range devices are 100% tested with temperature limits guaranteed by 100% testing, sampling, or by correlation with worst-case test conditions.

Note 4: Typicals are parametric norm at 25°C .

Note 5: Parameter guaranteed and 100% production tested.

Note 6: Parameter guaranteed. Parameters not 100% tested are not in outgoing quality level calculation.

Note 7: Total unadjusted error includes offset, full-scale, linearity, multiplexer and sample-and-hold errors.

Note 8: Cannot be tested for ML2282.

Note 9: For $V_{IN(-)} \geq V_{IN(+)}$ the digital output code will be 0000 0000. Two on-chip diodes are tied to each analog input (see Block Diagram) which will forward conduct for analog input voltages one diode drop below ground or one diode drop greater than the V_{CC} supply. Be careful, during testing at low V_{CC} levels (4.5V), as high level analog inputs (5V) can cause this input diode to conduct—especially at elevated temperatures, and cause errors for analog inputs near full-scale. The spec allows 50mV forward bias of either diode. This means that as long as the analog V_{IN} or V_{REF} does not exceed the supply voltage by more than 50mV, the output code will be correct. To achieve an absolute 0V to 5V input voltage range will therefore require a minimum supply voltage of $4.950 V_{DC}$ over temperature variations, initial tolerance and loading.

Note 10: Leakage current is measured with the clock not switching.

Note 11: A 40% to 60% clock duty cycle range insures proper operation at all clock frequencies. In the case that an available clock has a duty cycle outside of these limits, the minimum, time the clock is high or the minimum time the clock is low must be at least 300ns. The maximum time the clock can be high or low is $60\mu\text{s}$.

Note 12: Because of multiplexer addressing, test conditions for the ML2282 would be $V_{IN} = 34\text{ kHz}$, 5V sine ($f_{SAMPLING} \approx 102\text{ kHz}$); ML2284 $V_{IN} = 32\text{ kHz}$, 5V sine ($f_{SAMPLING} \approx 95\text{ kHz}$); ML2288 $V_{IN} = 30\text{ kHz}$, 5V sine ($f_{SAMPLING} \approx 89\text{ kHz}$).

Note 13: Since data, MSB first, is the output of the comparator used in the successive approximation loop, an additional delay is built in (see Block Diagram) to allow for comparator response time.

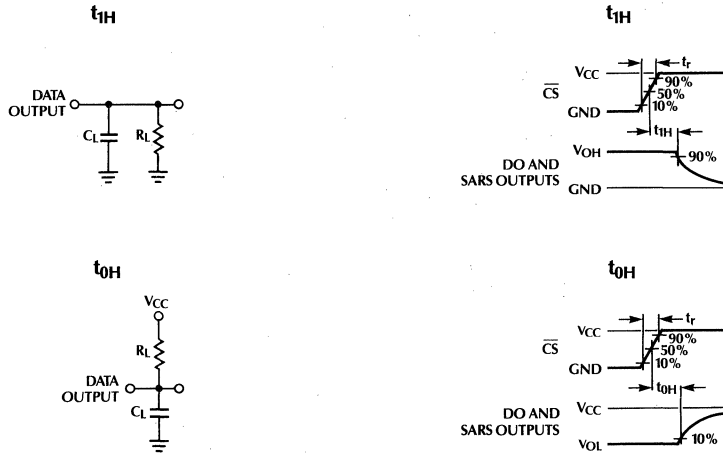


Figure 1. High Impedance Test Circuits and Waveforms

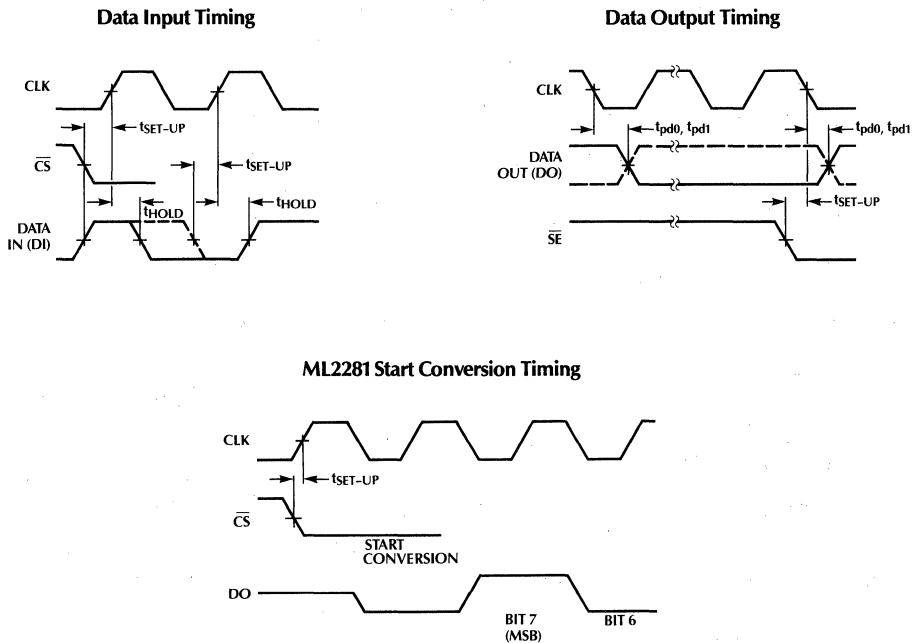
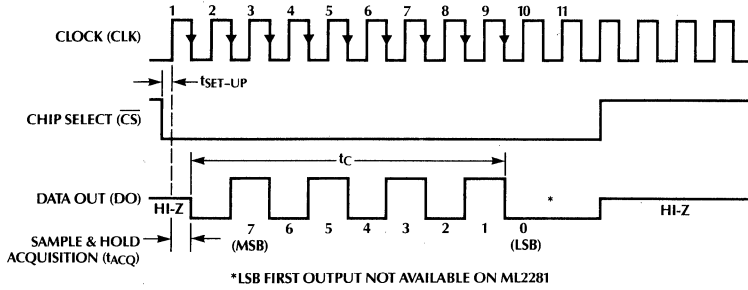
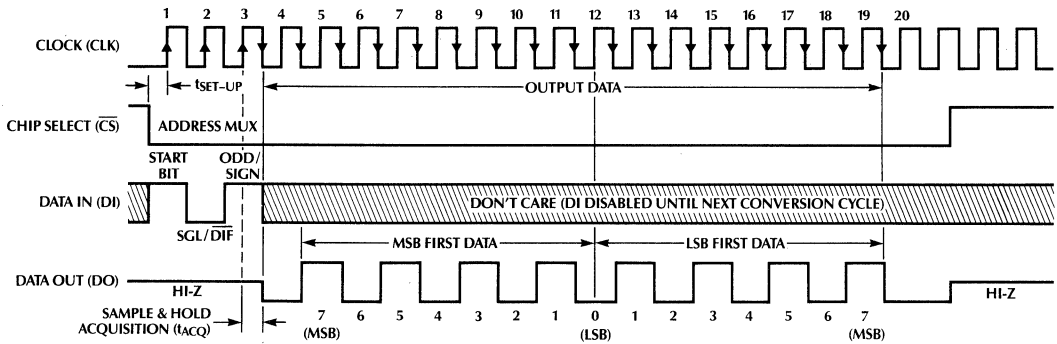


Figure 2. Timing Diagrams

ML2281 Timing



ML2282 Timing



ML2284 Timing

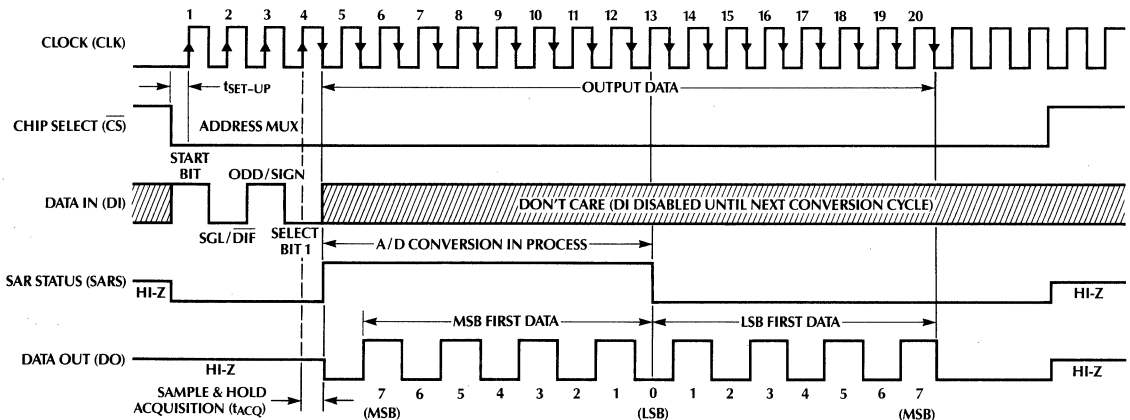


Figure 2. Timing Diagrams (Continued)

ML2288 Timing

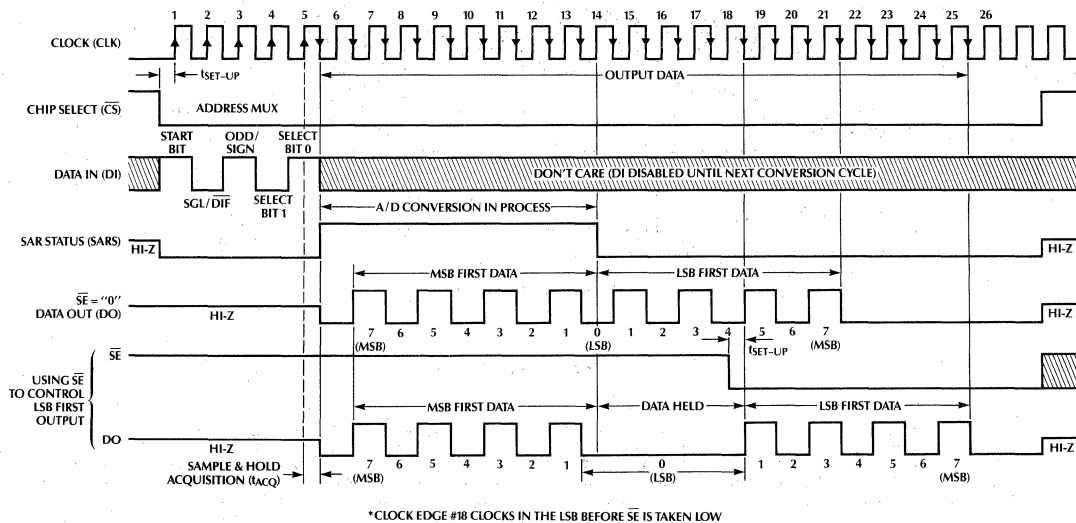


Figure 2. Timing Diagrams (Continued)

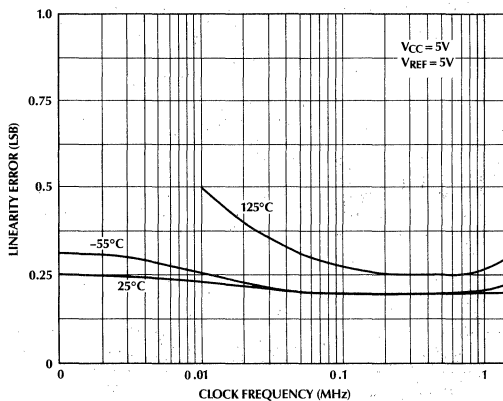


Figure 3. Linearity Error vs f_{CLK}

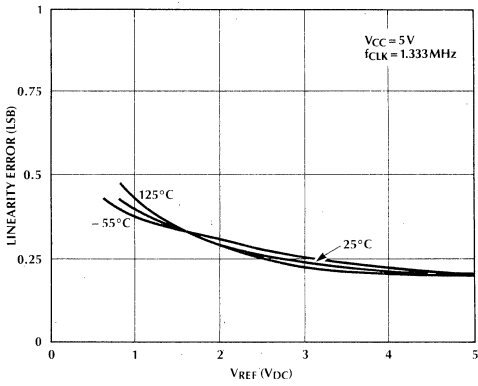


Figure 4. Linearity Error vs V_{REF} Voltage

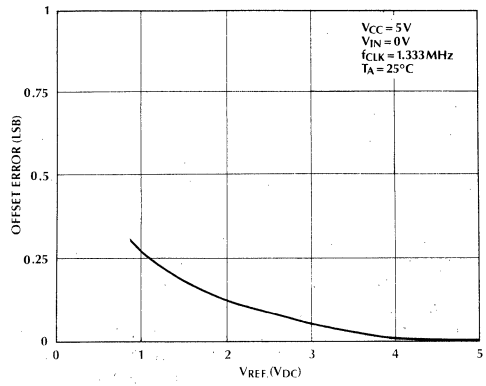
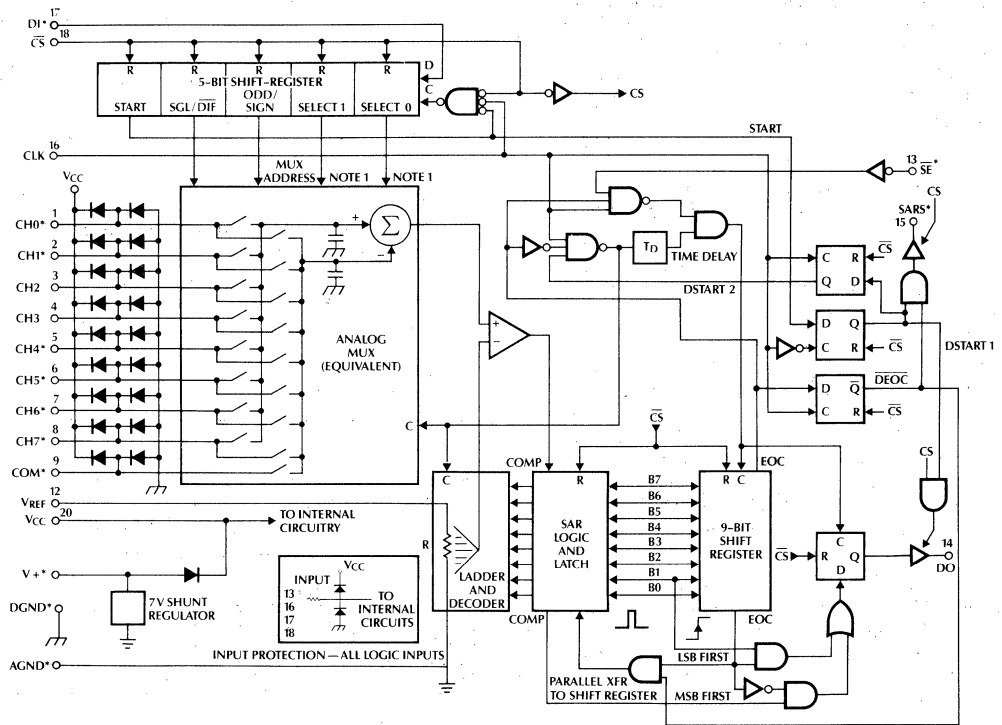


Figure 5. Unadjusted Offset Error vs V_{REF} Voltage

2



*SOME OF THESE FUNCTIONS/PINS ARE NOT AVAILABLE WITH OTHER OPTIONS.
 NOTE 1: FOR THE ML2284 DI IS INPUT DIRECTLY TO THE D INPUT OF SELECT 1. SELECT 0 IS FORCED TO A "1". FOR THE ML2282, DI IS INPUT DIRECTLY TO THE D INPUT OF ODD/SIGN. SELECT 0 IS FORCED TO A "1" AND SELECT 1 IS FORCED TO A "0".

Figure 6. ML2288 Functional Block Diagram

1.0 FUNCTIONAL DESCRIPTION

1.1 Multiplexer Addressing

The design of these converters utilizes a sample data comparator structure which provides for a differential analog input to be converted by a successive approximation routine.

The actual voltage converted is always the difference between an assigned “+” input terminal and a “-” input terminal. The polarity of each input terminal of the pair being converted indicates which line the converter expects to be the most positive. If the assigned “+” input is less than the “-” input, the converter responds with an all zeros output code.

A unique input multiplexing scheme has been utilized to provide multiple analog channels with software configurable single ended, differential, or pseudo differential options. The pseudo differential option will convert the difference between the voltage at any analog input and a common terminal. One converter package can now accommodate ground referenced inputs and true differential inputs as well as signals with some arbitrary reference voltage.

A particular input configuration is assigned during the MUX addressing sequence, prior to the start of a conversion. The MUX address selects which of the analog inputs are to be enabled and whether this input is single ended or differential. In the differential case, it also assigns the polarity of the analog channels. Differential inputs are restricted to adjacent channel pairs. For example, channel 0 and channel 1 may be selected as a different pair but channel 0 or channel 1 cannot act differentially with any other channel. In addition to selecting the differential mode, the sign may also be selected. Channel 0 may be selected as the positive input and channel 1 as the negative input or vice versa. This programmability is illustrated by the MUX addressing codes shown in *Tables 1, 2, and 3*.

The MUX address is shifted into the converter via the DI input. Since the ML2281 contains only one differential input channel with a fixed polarity assignment, it does not require addressing.

The common input line on the ML2288 can be used as a pseudo differential input. In this mode, the voltage on the COM pin is treated as the “-” input for any of the other input channels. This voltage does not have to be analog ground; it can be any reference potential which is common to all of the inputs. This feature is most useful in single supply applications where the analog circuitry may be biased at a potential other than ground and the output signals are all referred to this potential.

Since the input configuration is under software control, it can be modified, as required, at each conversion. A channel can be treated as a single-ended, ground referenced input for one conversion; then it can be reconfigured as part of a differential channel for another conversion. *Figure 7* illustrates these different input modes.

Table 1. ML2288 MUX Addressing 8 Single-Ended or 4 Differential Channels

MUX Address				Single-Ended MUX Mode									
SGL/DIF	ODD/SIGN	SELECT		Analog Single-Ended Channel #									
		1	0	0	1	2	3	4	5	6	7	COM	
1	0	0	0	+									-
1	0	0	1			+							-
1	0	1	0					+					-
1	0	1	1								+		-
1	1	0	0		+								-
1	1	0	1				+						-
1	1	1	0						+				-
1	1	1	1									+	-

MUX Address				Differential MUX Mode									
SGL/DIF	ODD/SIGN	SELECT		Analog Differential Channel-Pair #									
		1	0	0	1	2	3	4	5	6	7		
0	0	0	0	+	-								
0	0	0	1			+	-						
0	0	1	0					+	-				
0	0	1	1								+	-	
0	1	0	0	-	+								
0	1	0	1			-	+						
0	1	1	0					-	+				
0	1	1	1								-	+	

Table 2. ML2284 MUX Addressing 4 Single-Ended or 2 Differential Channel

MUX Address				Single-Ended MUX Mode			
SGL/DIF	ODD/SIGN	SELECT		Channel #			
		1	0	0	1	2	3
1	0	0	0	+			
1	0	1	0			+	
1	1	0	0		+		
1	1	1	0				+

COM is internally tied to A GND

MUX Address				Differential MUX Mode			
SGL/DIF	ODD/SIGN	SELECT		Channel #			
		1	0	0	1	2	3
0	0	0	0	+	-		
0	0	1	0			+	-
0	1	0	0	-	+		
0	1	1	0			-	+

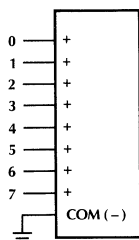
Table 3. ML2282 MUX Addressing 2 Single-Ended or 1 Differential Channel

MUX Address		Channel #	
SGL/DIF	ODD/SIGN	0	1
1	0	+	
1	1		+

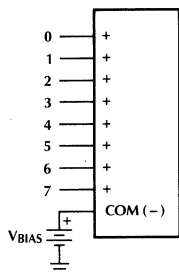
COM is internally tied to GND

MUX Address		Channel #	
SGL/DIF	ODD/SIGN	0	1
0	0	+	-
0	1	-	+

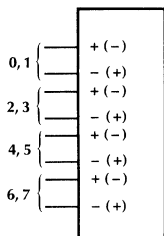
8 Single-Ended



8 Pseudo-Differential



4 Differential



Mixed Mode

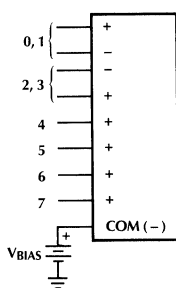


Figure 7. Analog Input Multiplexer Functional Options for ML2288

1.2 Digital Interface

The block diagram and timing diagrams in *Figures 2-5* illustrate how a conversion sequence is performed.

A conversion is initiated when \overline{CS} is pulsed low. This line must be held low for the entire conversion. The converter is now waiting for a start bit and its MUX assignment word.

A clock is applied to the CLK input. On each rising edge of the clock, the data on DI is clocked into the MUX address shift register. The start bit is the first logic "1" that appears on the DI input (all leading edge zeros are ignored). After the start bit, the device clocks in the next 2 to 4 bits for the MUX assignment word.

When the start bit has been shifted into the start location of the MUX register, the input channel has been assigned and a conversion is about to begin. An interval of $1/2$ clock period is used for sample & hold settling through the selected MUX channels. The SAR status output goes high at this time to signal that a conversion is now in progress and the DI input is ignored.

The DO output comes out of High impedance and provides a leading zero for this one clock period.

When the conversion begins, the output of the comparator, which indicates whether the analog input is greater than or less than each successive voltage from the internal DAC, appears at the DO output on each falling edge of the clock. This data is the result of the conversion being shifted out (with MSB coming first) and can be read by external logic or μP immediately.

After 8 clock periods, the conversion is completed. The SAR status line returns low to indicate this $1/2$ clock cycle later.

The serial data is always shifted out MSB first during the conversion. After the conversion has been completed, the data can be shifted out a second time with LSB first, depending on level of \overline{SE} input. For the case of ML2288, if $\overline{SE} = 1$, the data is shifted out MSB first during the conversion only. If \overline{SE} is brought low before the end of conversion (which is signalled by the high to low transition of \overline{SARS}), the data is shifted out again immediately after the end of conversion; this time LSB first. If \overline{SE} is brought low after end of conversion, the LSB first data is shifted out on falling edges of clock after \overline{SE} goes low. For ML2282 and 2284, \overline{SE} is internally tied low, so data is shifted out MSB first, then shifted out a second time LSB first at end of conversion. For ML2281, \overline{SE} is internally tied high, so data is shifted out only once MSB first.

All internal registers are cleared when the \overline{CS} input is high. If another conversion is desired, \overline{CS} must make a high to low transition followed by address information.

The DI input and DO output can be tied together and controlled through a bidirectional μP I/O bit with one connection. This is possible because the DI input is only latched in during the MUX addressing interval while the DO output is still in the high impedance state.

1.3 Reference

The voltage applied to the reference input to these converters defines the voltage span of the analog input (the difference between V_{INmax} and V_{INmin}) over which the 256 possible output codes apply. The devices can be used in either ratiometric applications or in systems requiring absolute accuracy. The reference pin must be connected to a voltage source capable of driving the reference input resistance, typically 10k. This pin is the top of a resistor divider string used for the successive approximation conversion.

In a ratiometric system, the analog input voltage is proportional to the voltage used for the A/D reference. This voltage is typically the system power supply, so the V_{REF} pin can be tied to V_{CC} . This technique relaxes the stability requirements of the system reference as the analog input and A/D reference move together maintaining the same output code for a given input condition.

For absolute accuracy, where the analog input varies between specific voltage limits, the reference pin can be biased with a time and temperature stable voltage source.

The maximum value of the reference is limited to the V_{CC} supply voltage. The minimum value, however, can be quite small to allow direct conversion of inputs with less than 5 volts of voltage span. Particular care must be taken with regard to noise pickup, circuit layout and system error voltage sources when operating with a reduced span due to the increased sensitivity of the converter.

1.4 Analog Inputs and Sample/Hold

An important feature of the ML2281 family of devices is that they can be located at the source of the analog signal and then communicate with a controlling μP with just a few wires. This avoids bussing the analog inputs long distances and thus reduces noise pickup on these analog lines. However, in some cases, the analog inputs have a large common mode voltage or even some noise present along with the valid analog signal.

The differential input of these converters reduces the effects of common mode input noise. Thus, if a common mode voltage is present on both "+" and "-" inputs, such as 60Hz, the converter will reject this common mode voltage since it only converts the difference between "+" and "-" inputs.

The ML2281 family have a true sample and hold circuit which samples both "+" and "-" inputs simultaneously. This simultaneous sampling with a true S/H will give common mode rejection and AC linearity performance that is superior to devices where the two input terminals are not sampled at the same instant and where true sample and hold capability does not exist. Thus, the ML2281 family of devices can reject AC common mode signals from DC-50kHz as well as maintain linearity for signals from DC-50kHz.

The signal at the analog input is sampled during the interval when the sampling switch is closed prior to conversion start. The sampling window (S/H acquisition time) is $1/2$ CLK period wide and occurs $1/2$ CLK period before DO goes from high impedance to active low state. When the sampling switch closes at the start of the S/H acquisition time, 8pF of

capacitance is thrown onto the analog input. $1/2$ CLK period later, the sampling switch is opened and the signal present at the analog input is stored. Any error on the analog input at the end of the S/H acquisition time will cause additional conversion error. Care should be taken to allow adequate charging or settling time from the source. If more charging or settling time is needed to reduce these analog input errors, a longer CLK period can be used.

The ML2281X family has improved latchup immunity. Each analog input has dual diodes to the supply rails, and a minimum of ± 25 mA (± 100 mA typically) can be injected into each analog input without causing latchup.

1.5 Dynamic Performance

Signal-to-Noise-Ratio

Signal-to-noise ratio (SNR) is the measured signal-to-noise at the output of the converter. The signal is the RMS magnitude of the fundamental. Noise is the RMS sum of all the nonfundamental signals up to half the sampling frequency. SNR is dependent on the number of quantization levels used in the digitization process; the more levels, the smaller the quantization noise. The theoretical SNR for a sine wave is given by

$$SNR = (6.02N + 1.76) \text{ dB}$$

where N is the number of bits. Thus for ideal 8-bit converter, $SNR = 49.92$ dB.

Harmonic Distortion

Harmonic distortion is the ratio of the RMS sum of harmonics to the fundamental. Total harmonic distortion (THD) of the ML2281 Series is defined as

$$20 \log \frac{(V_2^2 + V_3^2 + V_4^2 + V_5^2)^{1/2}}{V_1}$$

where V_1 is the RMS amplitude of the fundamental and V_2, V_3, V_4, V_5 are the RMS amplitudes of the individual harmonics.

Intermodulation Distortion

With inputs consisting of sine waves at two frequencies, f_A and f_B , any active device with nonlinearities will create distortion products, of order $(m+n)$, at sum and difference frequencies of $mf_A + nf_B$, where $m, n = 0, 1, 2, 3, \dots$. Intermodulation terms are those for which m or n is not equal to zero. The (IMD) intermodulation distortion specification includes the second order terms $(f_A + f_B)$ and $(f_A - f_B)$ and the third order terms $(2f_A + f_B)$, $(2f_A - f_B)$, $(f_A + 2f_B)$ and $(f_A - 2f_B)$ only.

1.6 Zero Error Adjustment

The zero of the A/D does not require adjustment. If the minimum analog input voltage value, V_{INmin} is not ground, a zero offset can be done. The converter can be made to output 00000000 digital code for this minimum input voltage by biasing any V_{IN-} input at this V_{INmin} value. This utilizes the differential mode operation of the A/D.

The zero error of the A/D converter relates to the location of the first riser of the transfer function and can be measured by grounding the V_{IN-} input and applying a small magnitude positive voltage to the V_{IN+} input. Zero error is the difference between the actual DC input voltage which is necessary to just cause an output digital code transition from 00000000 to 00000001 and the ideal $1/2$ LSB value ($1/2$ LSB = 9.8 mV for $V_{REF} = 5.000V_{DC}$).

1.7 Full-Scale Adjustment

The full-scale adjustment can be made by applying a differential input voltage which is $1/2$ LSB down from the desired analog full-scale voltage range and then adjusting the magnitude of the V_{REF} input or V_{CC} for a digital output code which is just changing from 11111110 to 11111111.

1.8 Adjustment for an Arbitrary Analog Input Voltage Range

If the analog zero voltage of the A/D is shifted away from ground (for example, to accommodate an analog input signal which does not go to ground), this new zero reference should be properly adjusted first. A V_{IN+} voltage which equals this desired zero reference plus $1/2$ LSB (where the LSB is calculated for the desired analog span, $1 \text{ LSB} = \text{analog span} / 256$) is applied to selected “+” input and the zero reference voltage at the corresponding “-” input should then be adjusted to just obtain the 00000000 to 00000001 code transition.

The full-scale adjustment should be made by forcing a voltage to the V_{IN+} input which is given by:

$$V_{IN+} \text{ fs adjust} = V_{\text{max}} - 1.5 * [(V_{\text{max}} - V_{\text{min}}) / 256]$$

where V_{max} = high end of the analog input range
 V_{min} = low end (offset zero) of the analog range

The V_{REF} or V_{CC} voltage is then adjusted to provide a code change from 11111110 to 11111111.

1.9 Shunt Regulator

A unique feature of ML2288 and ML2284 is the inclusion of a shunt regulator connected from $V+$ terminal to ground which also connects to the V_{CC} terminal (which is the actual converter supply) through a silicon diode as shown in Figure 8. When the regulator is turned on, the $V+$ voltage is clamped at $11 V_{BE}$ set by the internal resistor ratio. The typical I-V curve of the shunt regulator is shown in Figure 9. It should be noted that before $V+$ voltage is high enough to turn on the shunt regulator (which occurs at about 5.5V), 35 k Ω of resistance is observed between $V+$ and GND. When the shunt regulator is not used, $V+$ pin should be either left floating or tied to GND. The temperature coefficient of the regulator is $-22 \text{ mV}/^\circ\text{C}$.

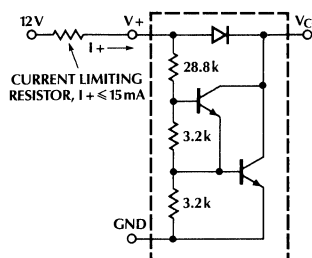


Figure 8. Shunt Regulator

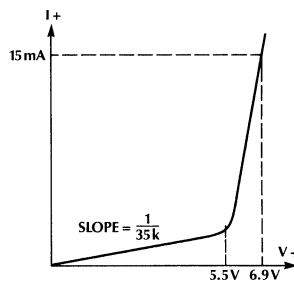
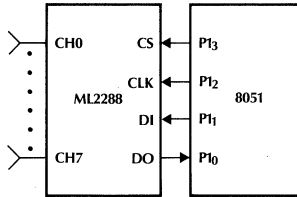


Figure 9. I-V Characteristic of the Shunt Regulator

2.0 APPLICATIONS

8051 Interface and Controlling Software

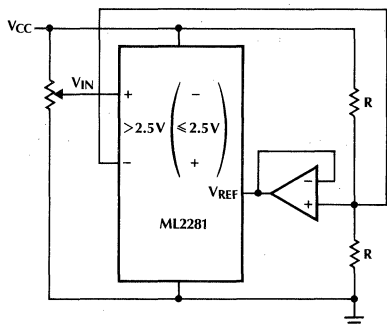


	Mnemonic		Instruction
START	ANL	P1, #0F7H	;SELECT A/D (CS = 0)
	MOV	B, #5	;BIT COUNTER ← 5
	MOV	A, #ADDR	;A ← MUX BIT
LOOP 1:	RRC	A	;CY ← ADDRESS BIT
	JC	ONE	;TEST BIT
			;BIT = 0
ZERO:	ANL	P1, #0FEH	;DI ← 0
	SJMP	CONT	;CONTINUE
			;BIT = 1
ONE:	ORL	P1, #1	;DI ← 1
CONT:	ACALL	PULSE	;PULSE SK 0 → 1 → 0
	DJNZ	B, LOOP 1	;CONTINUE UNTIL DONE
	ACALL	PULSE	;EXTRA CLOCK FOR SYNC
	MOV	B, #8	;BIT COUNTER ← 8
LOOP 2:	ACALL	PULSE	;PULSE SK 0 → 1 → 0
	MOV	A, P1	;CY ← DO
	RRC	A	
	RRC	A	
	MOV	A, C	;A ← RESULT
	RLC	A	;A(0) ← BIT AND SHIFT
	MOV	C, A	;C ← RESULT
	DJNZ	B, LOOP 2	;CONTINUE UNTIL DONE
RETI			
			;PULSE SUBROUTINE
PULSE:	ORL	P1, #04	;SK ← 1
	NOP		;DELAY
	ANL	P1, #0FBH	;SK ← 0
	RET		

ML2281, ML2282, ML2284, ML2288

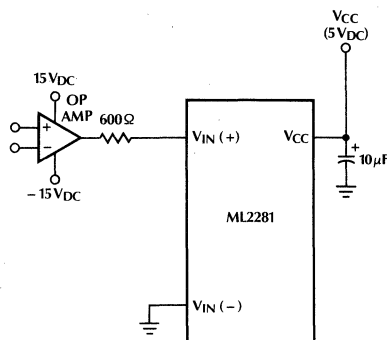
APPLICATIONS (Continued)

Obtaining 9-Bit Resolution



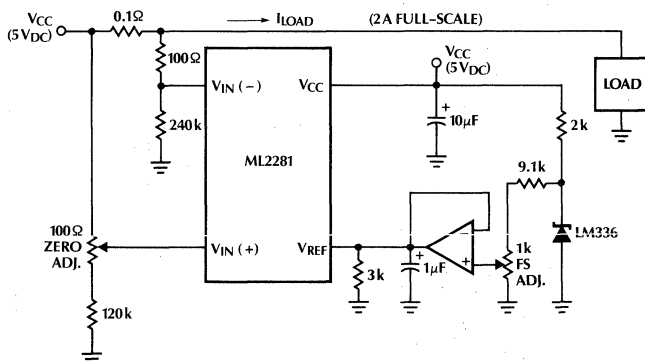
CONTROLLER PERFORMS A ROUTINE TO DETERMINE WHICH INPUT POLARITY PROVIDES A NON-ZERO OUTPUT CODE. THIS INFORMATION PROVIDES THE EXTRA BITS.

Protecting the Input

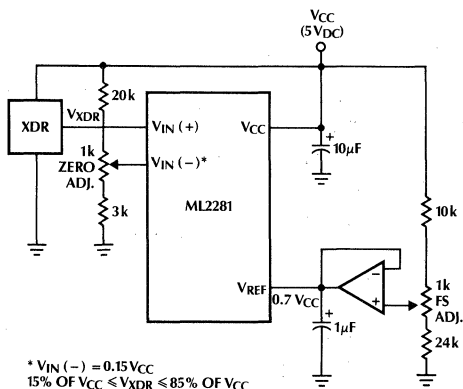


DIODE CLAMPING IS NOT NEEDED IF CURRENT IS LIMITED TO 25mA

Digitizing a Current Flow

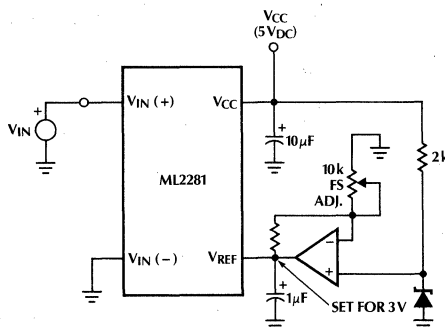


Operating with Ratiometric Transducers



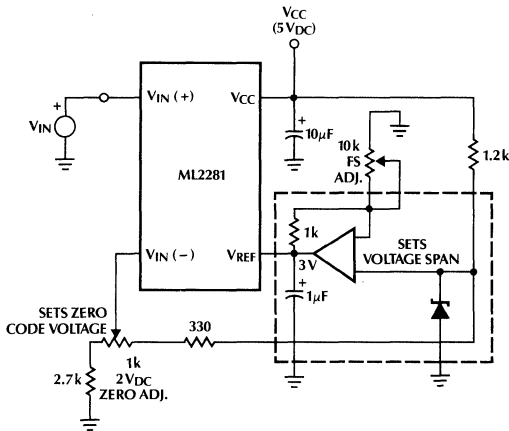
* $V_{IN(-)} = 0.15V_{CC}$
 $15\% \text{ OF } V_{CC} \leq V_{XDR} \leq 85\% \text{ OF } V_{CC}$

Span Adjust: $0V \leq V_{IN} \leq 3V$

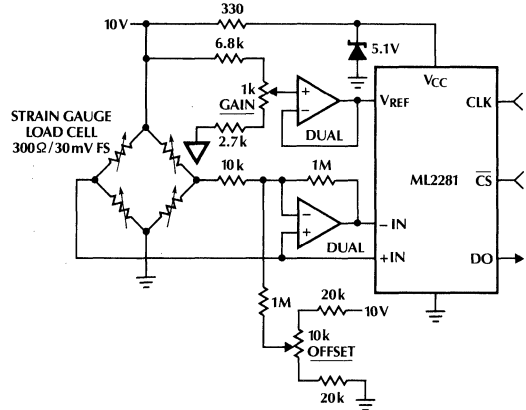


APPLICATIONS (Continued)

Zero-Shift and Span Adjust: $2V \leq V_{IN} \leq 5V$

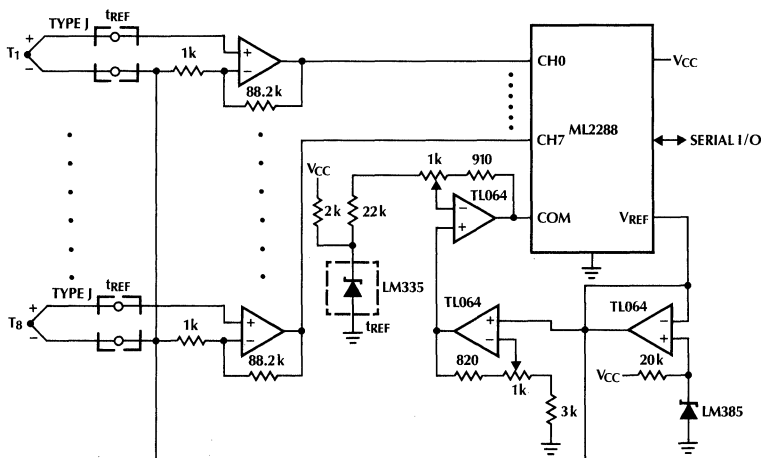


Digital Load Cell



- USES ONE MORE WIRE THAN LOAD CELL ITSELF
- TWO MINI-DIPS COULD BE MOUNTED INSIDE LOAD CELL FOR DIGITAL OUTPUT TRANSDUCER
- ELECTRONIC OFFSET AND GAIN TRIMS RELAX MECHANICAL SPECS FOR GAUGE FACTOR AND OFFSET
- LOW LEVEL CELL OUTPUT IS CONVERTED IMMEDIATELY FOR HIGH NOISE IMMUNITY

Convert 8 Thermocouples with only One Cold-Junction Compensator

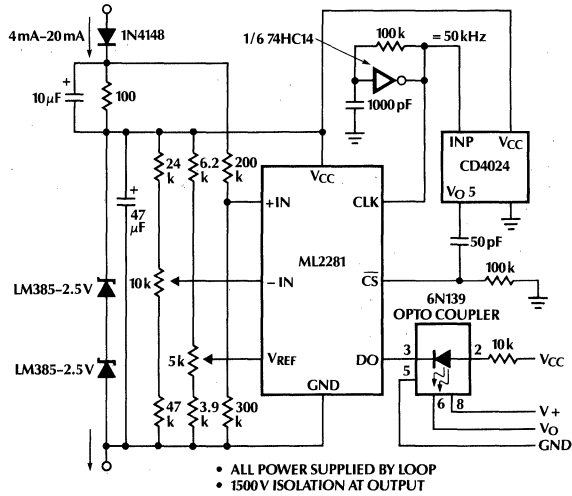


USES THE PSEUDO-DIFFERENTIAL MODE TO KEEP THE DIFFERENTIAL INPUTS CONSTANT WITH CHANGES IN REFERENCE TEMPERATURE (T_{REF}).

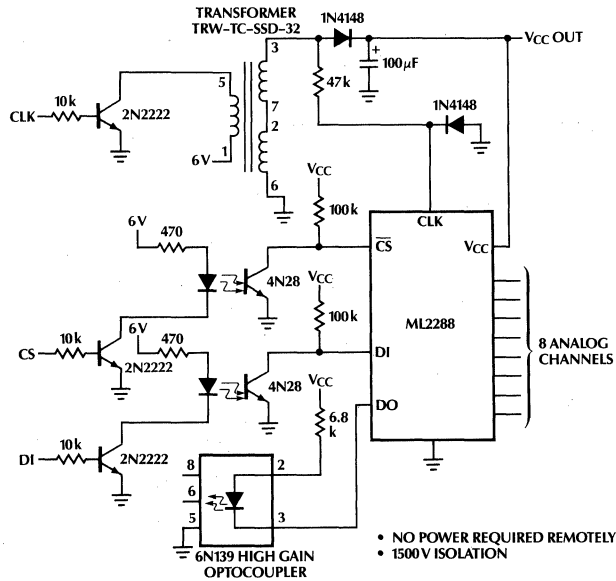
ML2281, ML2282, ML2284, ML2288

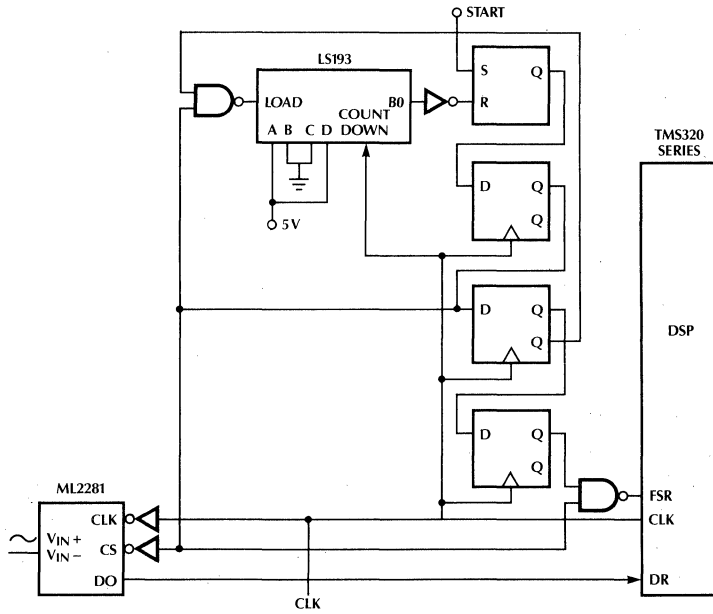
APPLICATIONS (Continued)

4mA-20mA Current Loop Converter

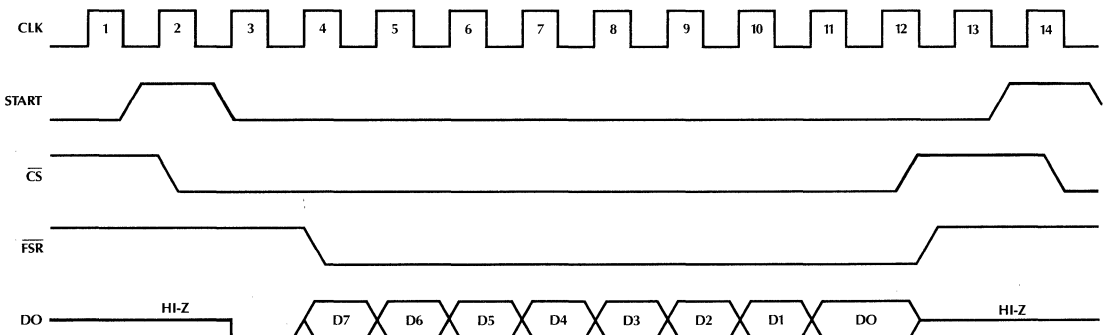


Isolated Data Converter





SAMPLING RATE 111kHz, DATA RATE 1.33MHz



Interfacing ML2281 to TMS320 Series

ML2281, ML2282, ML2284, ML2288

ORDERING INFORMATION

PART NUMBER	ALTERNATE PART NUMBER	TOTAL UNADJUSTED ERROR	TEMPERATURE RANGE	PACKAGE
SINGLE ANALOG INPUT, 8-PIN PACKAGE				
ML2281BMJ	ADC0831BJ	± 1/2 LSB	-55°C to +125°C -40°C to +85°C	HERMETIC DIP (J08)
ML2281BIJ	ADC0831BCJ			HERMETIC DIP (J08)
ML2281BCP	ADC0831BCN	± 1LSB	0°C to +70°C -40°C to +85°C	MOLDED DIP (P08)
ML2281CIJ	ADC0831CCJ			HERMETIC DIP (J08)
ML2281CCP	ADC0831CCN			MOLDED DIP (P08)
TWO ANALOG INPUTS, 8-PIN PACKAGE				
ML2282BMJ	ADC0832BJ	± 1/2 LSB	-55°C to +125°C -40°C to +85°C	HERMETIC DIP (J08)
ML2282BIJ	ADC0832BCJ			HERMETIC DIP (J08)
ML2282BCP	ADC0832BCN	± 1LSB	0°C to +70°C -40°C to +85°C	MOLDED DIP (P08)
ML2282CIJ	ADC0832CCJ			HERMETIC DIP (J08)
ML2282CCP	ADC0832CCN			MOLDED DIP (P08)
FOUR ANALOG INPUTS, 14-PIN PACKAGE				
ML2284BMJ	ADC0834BJ	± 1/2 LSB	-55°C to +125°C -40°C to +85°C	HERMETIC DIP (J14)
ML2284BIJ	ADC0834BCJ			HERMETIC DIP (J14)
ML2284BCP	ADC0834BCN	± 1LSB	0°C to +70°C -40°C to +85°C	MOLDED DIP (P14)
ML2284CIJ	ADC0834CCJ			HERMETIC DIP (J14)
ML2284CCP	ADC0834CCN			MOLDED DIP (P14)
EIGHT ANALOG INPUTS, 20-PIN PACKAGE				
ML2288BMJ	ADC0838BJ	± 1/2 LSB	-55°C to +85°C -40°C to +85°C	HERMETIC DIP (J20)
ML2288BIJ	ADC0838BCJ			HERMETIC DIP (J20)
ML2288BCP	ADC0838BCN	± 1LSB	0°C to +70°C 0°C to +70°C	MOLDED DIP (P20)
ML2288BCQ	ADC0838BCV			MOLDED PCC (Q20)
ML2288CIJ	ADC0838CCJ			HERMETIC DIP (J20)
ML2288CCP	ADC0838CCN	0°C to +70°C	0°C to +70°C	MOLDED DIP (P20)
ML2288CCQ	ADC0838CCV			MOLDED PCC (Q20)

ML2340, ML2350

Single Supply, Programmable 8-Bit D/A Converters

GENERAL DESCRIPTION

The ML2340 and ML2350 are CMOS voltage output, 8-bit D/A converters with an internal voltage reference and a μ P interface. These devices are designed to be powered by a single supply, although they can be powered from dual power supplies. The output voltage swings above zero scale (V_{ZS}) in the unipolar mode or around zero scale (V_{ZS}) in the bipolar mode, both with programmable gain. V_{ZS} can be set to any voltage from AGND to 2.25V below V_{CC} . The digital and analog grounds, DGND and AGND, are totally independent of each other. DGND can be set to any voltage from AGND to 4.5V below V_{CC} for easy interfacing to standard TTL and CMOS logic families.

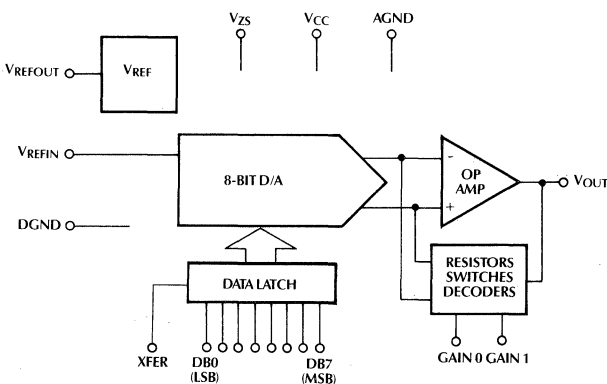
The high level of integration and versatility of the ML2340 and ML2350 makes them ideal for a wide range of applications in hard disk drives, automotive, telecom, and a variety of general purpose industrial. One specific intended application is controlling a hard disk voice coil.

The internal reference of the ML2340 provides a 2.25V or 4.50V output for use with A/D converters that use a single 5V \pm 10% power supply, while the ML2350 provide a 2.50V or 5.00V reference output.

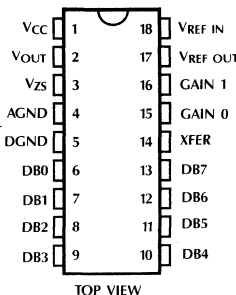
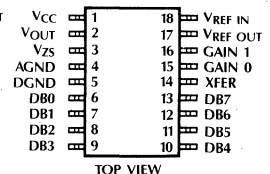
FEATURES

- Programmable output voltage gain settings of 2, 1, $\frac{1}{2}$, $\frac{1}{4}$ provide 8-, 9-, 10-, or 11-bit effective resolution around zero
- AGND to V_{CC} output voltage swing
- Bipolar or unipolar output voltage
- 4.5V to 13.2V single supply or ± 2.25 V to ± 6.5 V dual-supply operation
- Transparent latch allows microprocessor interface with 30ns setup time
- Data flow through mode
- Voltage reference output
 - ML2340 2.25V or 4.50V
 - ML2350 2.50V or 5.00V
- Nonlinearity $\pm \frac{1}{4}$ LSB or $\pm \frac{1}{2}$ LSB
- Output voltage settling time over temperature and supply voltage tolerance
 - Within 1V of V_{CC} and AGND 2.5 μ s max
 - Within 100mV of V_{CC} and AGND 5 μ s max
- TTL and CMOS compatible digital inputs
- Low supply current (5V supply) 5mA max
- 18-pin DIP or surface mount SOIC
- Operating temperature range of 0°C to +70°C, -40°C to +85°C, and -55°C to +125°C

BLOCK DIAGRAM



PIN CONNECTIONS

**ML2340
ML2350
18-Pin DIP**

**ML2340
ML2350
18-Pin SOIC**


ML2340, ML2350

PIN DESCRIPTION

PIN #	NAME	FUNCTION	PIN #	NAME	FUNCTION
1	V _{CC}	Positive supply.	8	DB2	Data input — Bit 2.
2	V _{OUT}	Voltage output of the D/A converter. V _{OUT} is referenced to V _{ZS} .	9	DB3	Data input — Bit 3.
3	V _{ZS}	Zero Scale Voltage. V _{OUT} is referenced to V _{ZS} . V _{ZS} is normally tied to AGND in the unipolar mode or to mid-supply in the bipolar mode. When the device is operated from a single power supply, V _{ZS} has a maximum current requirement of -300μA in the bipolar mode.	10	DB4	Data input — Bit 4.
4	AGND	Analog ground.	11	DB5	Data input — Bit 5.
5	DGND	Digital ground. This is the ground reference level for all digital inputs. The range is AGND < (V _{CC} - 4.5V). DGND is normally tied to system ground.	12	DB6	Data input — Bit 6.
6	DB0	Data input — Bit 0 (LSB).	13	DB7	Data input — Bit 7 (MSB).
7	DB1	Data input — Bit 1.	14	XFER	Transfer enable input. The data is transferred into the transparent latch at the high level of XFER.
			15	GAIN 0	Digital gain setting input 0.
			16	GAIN 1	Digital gain setting input 1.
			17	V _{REF OUT}	Voltage reference output. V _{REF OUT} is referenced to AGND. V _{REF OUT} is set to 2.5V and 5.0V in a low-voltage and high-voltage operation, respectively for the ML2350; 2.25V and 4.5V for the ML2340.
			18	V _{REF IN}	Voltage reference input. V _{REF IN} is referenced to AGND.

ABSOLUTE MAXIMUM RATINGS

(Note 1)

Supply Voltage V _{CC} with Respect to AGND	14.2V
DGND	-0.3V to V _{CC} + 0.3V
V _{ZS} , V _{REF IN}	-0.3V to V _{CC} + 0.3V
Logic Inputs	-0.3V to V _{CC} + 0.3V
Input Current per Pin (Note 2)	±25mA
Storage Temperature	-65°C to +150°C
Package Dissipation at T _A = 25°C (Board Mount)	875mW
Lead Temperature (Soldering 10 sec.)	
Dual-In-Line Package (Molded)	260°C
Dual-In-Line Package (Ceramic)	300°C
Molded Small Outline IC Package	
Vapor Phase (60 sec.)	215°C
Infrared (15 sec.)	220°C

OPERATING CONDITIONS

(Note 1)

Supply Voltage, V _{CC}	4.5V _{DC} to 13.2V _{DC}
Temperature Range (Note 3)	T _{MIN} ≤ T _A ≤ T _{MAX}
ML2340BMJ, ML2340CMJ	
ML2350BMJ, ML2350CMJ	-55°C to +125°C
ML2340BIJ, ML2340CIJ	
ML2350BIJ, ML2350CIJ	-40°C to +85°C
ML2340BCP, ML2340CCP	
ML2350BCP, ML2350CCP	
ML2340BCS, ML2340CCS	
ML2350BCS, ML2350CCS	0°C to +70°C

ELECTRICAL CHARACTERISTICS

Unless otherwise specified, $T_A = T_{MIN}$ to T_{MAX} , $V_{CC} - AGND = 5V \pm 10\%$ and $12V \pm 10\%$, $V_{REF IN}$ for ML2340 = 2.25V and 4.50V, for ML2350 $V_{REF IN} = 2.50V$ and $5.00V$, V_{OUT} load is $R_L = 1K$ and $C_L = 100pF$, V_{REF} load is $R_L = 1K$ and $C_L = 100pF$ and input control signals with $t_R = t_F \leq 20ns$.

PARAMETER	NOTES	CONDITIONS	ML2340XCX, ML2350XCX			ML2340XIX, ML2340XX ML2350XIX, ML2350XX			UNITS
			MIN	TYP (Note 4)	MAX	MIN	TYP (Note 4)	MAX	
Converter and Programmable Gain Amplifier									
Converter Resolution	5		8			8			Bits
Integral Linearity Error ML2340BXX, ML2350BXX ML2340CXX, ML2350CXX	5	GAIN = 2, 1, 1/2, or 1/4			$\pm 1/4$ $\pm 1/2$			$\pm 1/4$ $\pm 1/2$	LSB LSB
Differential Linearity Error ML2340BXX, ML2350BXX ML2340CXX, ML2350CXX	5	GAIN = 2, 1, 1/2, or 1/4			$\pm 1/4$ $\pm 1/2$			$\pm 1/4$ $\pm 1/2$	LSB LSB
Mode Select Unipolar Output Bipolar Output	5	V_{ZS} with respect to AGND	0 1.50		1.0 $V_{CC}-2.25$	0 1.50		1.0 $V_{CC}-2.25$	V V
Offset Error Unipolar Mode	5	Figure 1 GAIN = 1/4, 1/2, 1 GAIN = 2			± 10 ± 20			± 12 ± 24	mV mV
Bipolar Mode	5	Figure 1 GAIN = 1/4, 1/2, 1, 2			± 10 plus $\pm 2 1/2$ LSB			± 10 plus $\pm 2 1/2$ LSB	mV
Gain Error Unipolar Mode Bipolar Mode	5	Figure 1 GAIN = 1/4, 1/2, 1, 2 GAIN = 1/4, 1/2, 1, 2		± 5 ± 5	± 2 ± 2		± 5 ± 5	± 2.5 ± 2.5	%FS %FS
Reference									
$V_{REF OUT}$ Voltage ML2340BXX	5	$V_{CC} \leq 7.0V$ $T_A = 25^\circ C$ T_{MIN} to T_{MAX}	2.23 2.22	2.25	2.27 2.28	2.23 2.18	2.25	2.27 2.32	V V
		$V_{CC} \geq 8.0V$ $T_A = 25^\circ C$ T_{MIN} to T_{MAX}	4.48 4.46	4.50	4.52 4.54	4.48 4.43	4.50	4.52 4.57	V V
ML2340CXX	5	$V_{CC} \leq 7.0V$ $T_A = 25^\circ C$ T_{MIN} to T_{MAX}	2.22 2.20	2.25	2.29 2.30	2.22 2.18	2.25	2.28 2.32	V V
		$V_{CC} \geq 8.0V$ $T_A = 25^\circ C$ T_{MIN} to T_{MAX}	4.45 4.40	4.50	4.55 4.60	4.45 4.35	4.50	4.55 4.65	V V
ML2350BXX	5	$V_{CC} \leq 7.0V$ $T_A = 25^\circ C$ T_{MIN} to T_{MAX}	2.48 2.47	2.50	2.52 2.53	2.48 2.43	2.50	2.52 2.57	V V
		$V_{CC} \geq 8.0V$ $T_A = 25^\circ C$ T_{MIN} to T_{MAX}	4.98 4.96	5.00	5.02 5.04	4.98 4.90	5.00	5.02 5.10	V V
ML2350CXX	5	$V_{CC} \leq 7.0V$ $T_A = 25^\circ C$ T_{MIN} to T_{MAX}	2.45 2.44	2.50	2.55 2.58	2.46 2.42	2.50	2.55 2.59	V V
		$V_{CC} \geq 8.0V$ $T_A = 25^\circ C$ T_{MIN} to T_{MAX}	4.95 4.90	5.00	5.05 5.10	4.95 4.85	5.00	5.05 5.15	V V
Temperature Coefficient $V_{REF OUT}$				50			50		ppm/°C
V_{REF} Output Current	5		0.75		5	0.75		5	mA
$V_{REF OUT}$ Power Supply Rejection Ratio	5	100mV _{p-p} , 1kHz Sinewave on V_{CC}	-40	-60		-40	-60		dB

ML2340, ML2350

ELECTRICAL CHARACTERISTICS (Continued)

Unless otherwise specified, $T_A = T_{MIN}$ to T_{MAX} , $V_{CC} - AGND = 5V \pm 10\%$ and $12V \pm 10\%$, $V_{REF IN}$ for ML2340 = 2.25V and 4.50V; for ML2350 $V_{REF IN} = 2.50V$ and $5.00V$, V_{OUT} load is $R_L = 1K$ and $C_L = 100pF$, V_{REF} load is $R_L = 1K$ and $C_L = 100pF$ and input control signals with $t_r = t_f \leq 20ns$.

PARAMETER	NOTES	CONDITIONS	ML2340XCX, ML2350XCX			ML2340XIX, ML2340MX ML2350XIX, ML2350MX			UNITS	
			MIN	TYP (Note 4)	MAX	MIN	TYP (Note 4)	MAX		
$V_{REF IN}$ and V_ZS										
$V_{REF IN}$ Input Range	5	$V_{CC} \leq 8.75V$ $V_{CC} \geq 8.75V$	AGND+2 AGND+2		$V_{CC}-1.75$ AGND+7	AGND+2 AGND+2		$V_{CC}-1.75$ AGND+7	V V	
$V_{REF IN}$ DC Input Resistance	5		10			10			MΩ	
V_ZS Voltage Range	5, 8	$V_{CC} \leq 7.0V$	AGND		$V_{CC}-2.25$	AGND		$V_{CC}-2.25$	V	
Analog Output										
V_{OUT} Output Swing Unipolar Mode	5, 8	$R_L = 100K$	AGND+ 0.01		$V_{CC}-0.05$	AGND+ 0.01		$V_{CC}-0.05$	V	
			AGND+ 1.0		$V_{CC}-1.0$	AGND+ 1.0		$V_{CC}-1.0$	V	
	Bipolar Mode	5	$R_L = 100K$	AGND+ 0.1		$V_{CC}-0.1$	AGND+ 0.1		$V_{CC}-0.1$	V
				AGND+ 1.0		$V_{CC}-1.0$	AGND+ 1.0		$V_{CC}-1.0$	V
V_{OUT} Output Current	5	$AGND+1V < V_{OUT} < V_{CC}-1V$	-10		+10	-10		+10	mA	
Power Supply Rejection Ratio		100mV _{p-p} , 1kHz sinewave on V_{CC}		-60			-60		dB	
Digital and DC										
$V_{IN(0)}$ Logical "0" Input Voltage	5				0.8			0.8	V	
$V_{IN(1)}$ Logical "1" Input Voltage	5		2.0			2.0			V	
$I_{IN(0)}$ Logical "0" Input Current	5	$V_{IN} = DGND$	-1			-1			μA	
$I_{IN(1)}$ Logical "1" Input Current	5	$V_{IN} = V_{CC}$			1			1	μA	
Supply Current, Bipolar Mode I_{CC} , V_{CC} Current I_{AGND} , Analog Ground Current I_{VZS} , V_ZS Current	5	$V_{CC} = 5V \pm 10\%$			5.3			5.3	mA	
				-90	-300		-90	-300	mA μA	
I_{CC} , V_{CC} Current I_{AGND} , Analog Ground Current I_{VZS} , V_ZS Current	5	$V_{CC} = 12V \pm 10\%$			9.3			9.3	mA	
				-90	-300		-90	-300	mA μA	

ELECTRICAL CHARACTERISTICS (Continued)

Unless otherwise specified, $T_A = T_{MIN}$ to T_{MAX} , $V_{CC} - AGND = 5V \pm 10\%$ and $12V \pm 10\%$, $V_{REF IN}$ for ML2340 = 2.25V and 4.50V, for ML2350 $V_{REF IN} = 2.50V$ and 5.00V, V_{OUT} load is $R_L = 1K$ and $C_L = 100pF$, V_{REF} load is $R_L = 1K$ and $C_L = 100pF$ and input control signals with $t_r = t_f \leq 20ns$.

PARAMETER	NOTES	CONDITIONS	ML2340XCX, ML2350XCX			ML2340XIX, ML2340XMX ML2350XIX, ML2350XMX			UNITS
			MIN	TYP (Note 4)	MAX	MIN	TYP (Note 4)	MAX	
			Digital and DC (Continued)						
Supply Current, Unipolar Mode	5	$V_{CC} = 5V \pm 10\%$			6.0			6.0	mA
I_{CC} V_{CC} Current									
I_{AGND} , Analog Ground Current	5	$V_{CC} = 12V \pm 10\%$			-4.3			-4.3	mA
I_{VZS} , V_{ZS} Current									-1.7
I_{CC} V_{CC} Current	5	$V_{CC} = 12V \pm 10\%$			11.0			11.0	mA
I_{AGND} , Analog Ground Current									
I_{VZS} , V_{ZS} Current					-7.3			-7.3	mA
					-3.7			-3.7	mA
AC Performance									
Settling Time t_{S1}	5	Figure 2, Output Step of AGND + 1V to $V_{CC} - 1V$, $R_L = 1K$		1.2	2.5		1.2	3.0	μs
t_{S2}		Output Step of AGND + 100mV to $V_{CC} - 100mV$, $R_L = 100K$		2.5	5		2.5	6	μs
t_{S3}		Output Step of $\pm 1LSB$			1			1	μs
t_{S4} Gain Change		Change of Any Gain Setting		1.1	2.5		1.1		μs
t_{XFER} XFER Pulse Width	5	Figure 3	60			60			ns
t_{DBS} DB0-DB7 Setup Time	5	Figure 3	40			45			ns
t_{DBH} DB0-DB7 Hold Time	5	Figure 3	0			0			ns
t_{RESET} Power-On Reset Time	6				16			16	μs

- Note 1:** Absolute maximum ratings are limits beyond which the life of the integrated circuit may be impaired. All voltages unless otherwise specified are measured with respect to analog ground.
- Note 2:** When the voltage at any pin exceeds the power supply rails ($V_{IN} < AGND$ or $V_{IN} > V_{CC}$) the absolute value of current at that pin should be limited to 25mA or less.
- Note 3:** -55°C to +125°C operating temperature range devices are 100% tested at temperature extremes with worst-case test conditions. -40°C to +85°C operating temperature range devices are 100% tested with temperature limits guaranteed by 100% testing, sampling, or by correlation with worst-case test conditions.
- Note 4:** Typicals are parametric norm at 25°C.
- Note 5:** Parameter guaranteed and 100% production tested.
- Note 6:** Parameter guaranteed. Parameters not 100% tested are not in outgoing quality level calculation.
- Note 7:** Supply current and analog ground current are specified with the digital inputs stable and no load on V_{OUT} .
- Note 8:** In unipolar operation with V_{ZS} and AGND tied together, digital codes that represent an analog value of less than 100mV from AGND should be avoided.

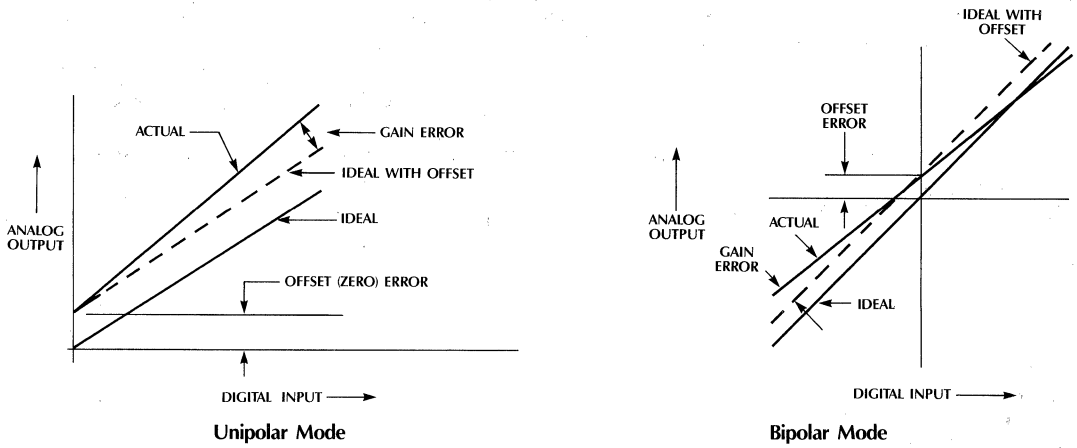


Figure 1. Gain and Offset Error

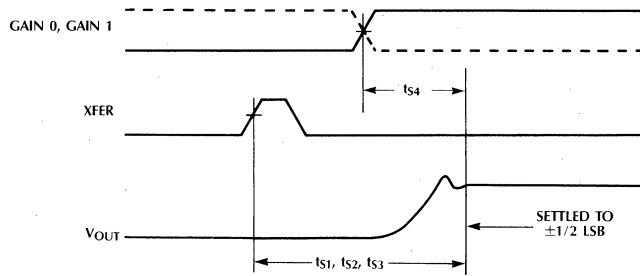


Figure 2. Settling Time

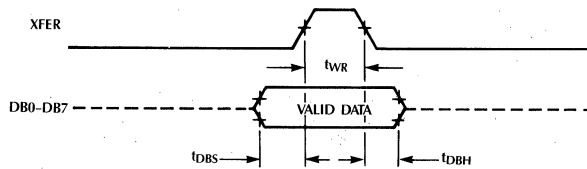


Figure 3. Single Buffered Mode

1.0 FUNCTIONAL DESCRIPTION

1.1 D/A CONVERTER

The D/A converter is implemented using an array of equal current sources that are decoded semi linearly for the four most significant bits to improve differential linearity and to reduce output glitch around major carries. See Figure 4.

The input voltage reference of the D/A converter is the difference between $V_{REF\ IN}$ and AGND. This difference voltage is converted to a reference current using an internal resistor to set up the appropriate current level

in the D/A converter. The D/A converter output current is then converted to a voltage output by an output buffer and a resistive network. The matching among the on-chip resistors preserves the gain accuracy between these conversions.

The D/A converter can be used in a multiplying mode by modulating the reference input within the specified $V_{REF\ IN}$ range.

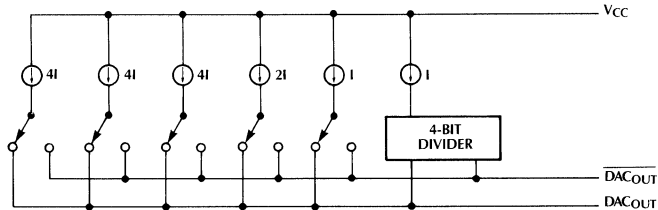


Figure 4. D/A Converter Implementation

1.2 SINGLE-SUPPLY vs. DUAL-SUPPLY OPERATION

ML2340 and ML2350 can be powered from a single supply ranging from 4.5V to 13.2V or dual supplies ranging from $\pm 2.25V$ to $\pm 6.6V$.

The internal digital and analog circuitry is powered between V_{CC} and AGND. The range of DGND is $AGND \leq DGND \leq V_{CC} - 4.5V$ with the logic thresholds set between .8V and 2.0V above DGND (standard TTL logic level). The range of V_{ZS} is $AGND \leq V_{ZS} \leq (V_{CC} - 2.25V)$.

1.3 UNIPOLAR AND BIPOLAR OUTPUT VOLTAGE SWING

The ML2340 and ML2350 can operate in either unipolar and bipolar output voltage mode. Unipolar/bipolar mode selection is determined by comparing the zero scale voltage (V_{ZS}) of these devices to a precise internal reference that is referred to AGND. V_{ZS} is ideally the voltage that will be produced at the DAC voltage output when the digital input data is set to all "0's". Unipolar mode is selected when V_{ZS} is lower than 1.00 volt, and bipolar mode is selected when V_{ZS} is greater than 1.50 volts.

1.3.1 Unipolar Output Mode

In the unipolar mode, V_{OUT} swings above V_{ZS} . Ideally the 00000000 code results in an output voltage of V_{ZS} , and the 11111111 code results in an output voltage of $V_{FS} \times 255/256$, where V_{FS} is the full-scale voltage determined by $V_{REF\ IN}$ and the gain setting.

1.3.2 Bipolar Output Mode

In the bipolar mode, V_{OUT} swings around V_{ZS} . The input data is in 2's complement binary format. Ideally, the 00000000 code results in an output voltage of V_{ZS} ; the 10000000 code results in an output voltage of $(V_{ZS} - V_{FS})$; and the 01111111 results in an output voltage of $(V_{ZS} + V_{FS} 127/128)$, where V_{FS} is the full scale output voltage determined by $V_{REF\ IN}$ and the gain setting.

1.4 OUTPUT BUFFER AND GAIN SETTING

The output buffer converts the D/A output current to a voltage output using a resistive network with proper gain setting determined by the GAIN 0 and GAIN 1 inputs. There are four possible gain settings for unipolar output voltage mode and bipolar output voltage mode as listed below:

Unipolar Output Voltage Mode

GAIN 1	GAIN 0	GAIN	Voltage Output Swing Relative to V_{ZS}
0	0	$\frac{1}{4}$	$V_{REF\ IN} \times \frac{1}{4}$
0	1	$\frac{1}{2}$	$V_{REF\ IN} \times \frac{1}{2}$
1	0	1	$V_{REF\ IN} \times 1$
1	1	2	$V_{REF\ IN} \times 2$

ML2340, ML2350

Bipolar Output Voltage Mode

GAIN 1	GAIN 0	GAIN	Voltage Output _{p-p}
0	0	1/4	$\pm V_{REF IN} \times 1/8$
0	1	1/2	$\pm V_{REF IN} \times 1/4$
1	0	1	$\pm V_{REF IN} \times 1/2$
1	1	2	$\pm V_{REF IN} \times 1$

The output buffer can source or sink as much as 10mA of current with an output voltage of at least 1V from either V_{CC} or AGND. As the output voltage approaches V_{CC} or AGND the current sourcing/sinking capability of the output buffer is reduced. The output buffer can still swing down to within 10mV of AGND and up to within 40mV of V_{CC} with a 100K load at V_{OUT} to AGND in the unipolar operation. In the bipolar operation, the output buffer swing is limited to about 100mV from either rails.

1.5 VOLTAGE REFERENCE

A bandgap voltage reference is incorporated on the ML2340 and ML2350. Two reference voltages can be produced by each device. An internal comparator monitors the power supply voltage to determine the selection of the reference voltage. A reference voltage of 2.25 volts on the ML2340 and 2.50 volts on the ML2350 is selected when the supply voltage is less than approximately 7.50 volts. Otherwise, a reference voltage of 4.50 volts and 5.00 volts is selected. To prevent the comparator from oscillating between the two selections, avoid operation with a power supply between 7.0 and 8.0 volts.

The bandgap reference is trimmed for zero Temperature Coefficient (TC) at 35°C to minimize output voltage drift over the specified operating temperature range.

The internal reference is buffered for use by the DAC and external circuits. The reference buffer will source more than 5mA of current and sink more than 1mA of current. With $V_{REF IN}$ connected to $V_{REF OUT}$, the following output voltage ranges of the DAC are obtained:

ML2340

Gain Setting	$V_{REF} = 2.25V$ with $V_{CC} \leq 7.0V$		$V_{REF} = 4.5V$ with $V_{CC} \geq 8.0V$	
	Unipolar	Bipolar	Unipolar	Bipolar
1/4	0 to 0.562V	-0.281V to +0.281V	0 to 1.125V	-0.562V to +0.562V
1/2	0 to 1.125V	-0.562V to +0.562V	0 to 2.250V	-1.125V to +1.125V
1	0 to 2.250V	-1.125V to +1.125V	0 to 4.500V	-2.250V to +2.250V
2	0 to 4.500V	-2.250V to +2.250V	0 to 9.000V	-4.500V to +4.500V

ML2350

Gain Setting	$V_{REF} = 2.50V$ with $V_{CC} \leq 7.0V$		$V_{REF} = 5.00V$ with $V_{CC} \geq 8.0V$	
	Unipolar	Bipolar	Unipolar	Bipolar
1/4	0 to 0.625V	-0.3125V to +0.3125V	0 to 1.25V	-0.625V to +0.625V
1/2	0 to 1.250V	-0.6250V to +0.6250V	0 to 2.50V	-1.250V to +1.250V
1	0 to 2.500V	-1.2500V to +1.2500V	0 to 5.00V	-2.500V to +2.500V
2	0 to 5.000V	-2.5000V to +2.5000V	0 to 10.00V	-5.000V to +5.000V

An external reference can alternatively be used on $V_{REF IN}$ to set the desired full scale voltage. The linearity of the D/A converter depends on the reference used, however. To insure integral linearity at an 8-bit level, a reference voltage of no less than 2V and no more than 7V (2.75V for operation with a low-voltage power supply) should be used.

1.6 DIGITAL INTERFACE

The digital interface of the ML2340 and ML2350 consist of a transfer input (XFER) and eight data inputs, DB0 through DB7. The digital interface operates in one of the two modes:

1.6.1 Single-Buffered Mode

Digital input data on DB0–DB7 is passed through an 8-bit transparent input latch on the rising edge of XFER. Because the outputs of the latch are connected directly to the inputs of the internal DAC, changes on the digital data while the XFER input is still active will cause an immediate change in the DAC output voltage. To hold the input data on the latch, the XFER input needs to be deactivated while the data is still stable.

1.6.2 Flow-Through Mode

In the flow-through mode, the input latch is bypassed. When XFER is set to logic "1", a change of data inputs, DB0–DB7, results in an immediate update of the output voltage.

1.7 POWER-ON-RESET

The ML2340 and ML2350 have an internal power-on-reset circuit to initialize the device when power is first applied to the device. The power-on-reset interval of typically 8 μ s begins when the supply voltage, V_{CC} reaches approximately 2.0V. During the power-on-reset interval, the transparent latch is reset to all "0's".

2.0 TYPICAL APPLICATIONS

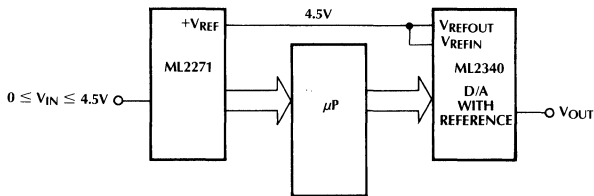


Figure 5. Using 4.50V Reference of D/A for Reference of A/D Using Single 5V $V_{CC} \pm 10\%$

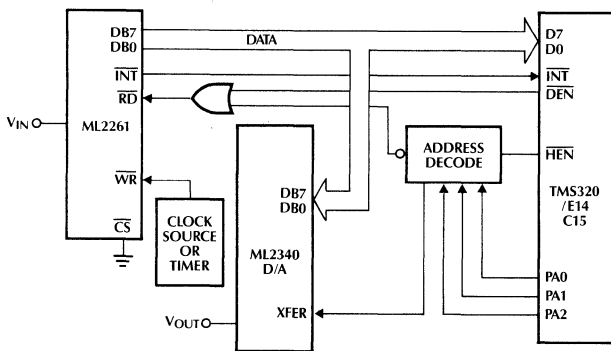


Figure 6. TMS320 Interface

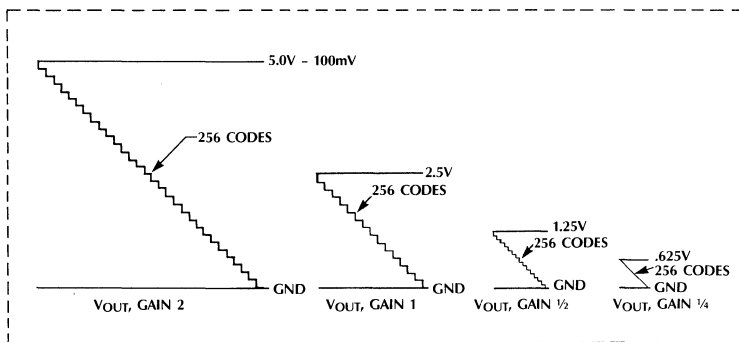
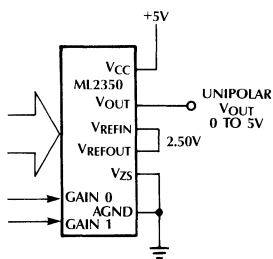


Figure 7. Single 5V Supply Unipolar V_{OUT}

2.0 TYPICAL APPLICATIONS (Continued)

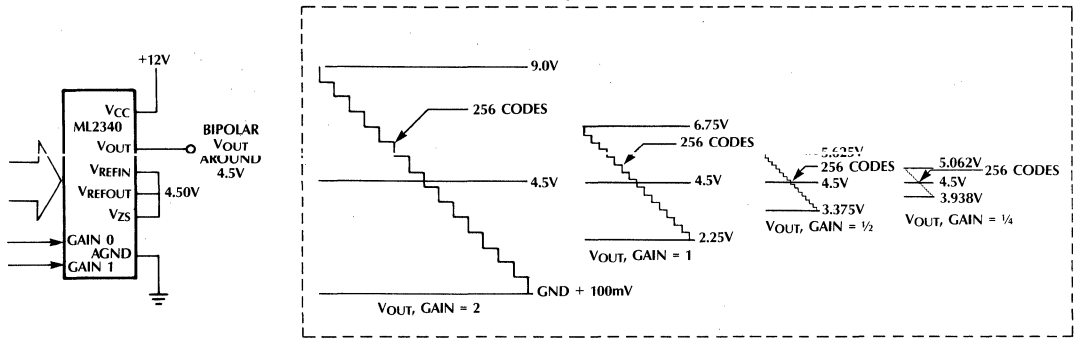


Figure 8. Single 12V Supply, Bipolar V_{OUT} with 11-Bits Resolution Around Zero

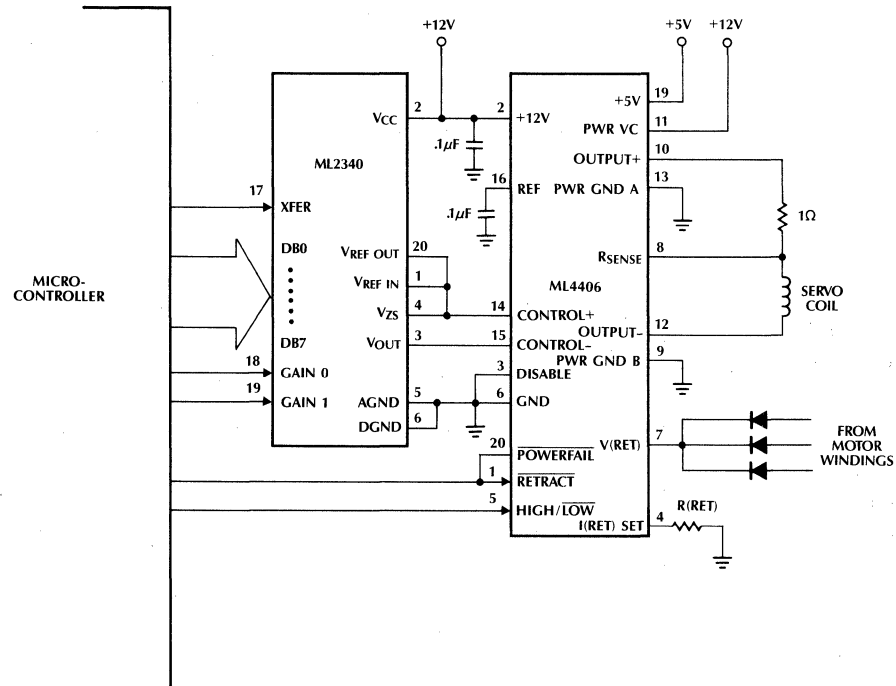


Figure 9. Hard Disc Drive Servo Coil Driver Providing 13-Bit Effective Resolution

ORDERING INFORMATION

PART NUMBER	INTEGRAL & DIFFERENTIAL NON-LINEARITY	TEMPERATURE RANGE	PACKAGE
V_{REF OUT} = 2.25V with V_{CC} = 5V			
ML2340BMJ/5 ML2340BIJ/5 ML2340BCP/5 ML2340BCQ/5	±¼ LSB	-55°C to +125°C -40°C to +85°C 0°C to +70°C	HERMETIC DIP (J18) HERMETIC DIP (J18) MOLDED DIP (P18) MOLDED PCC (Q18)
ML2340CMJ/5 ML2340CIJ/5 ML2340CCP/5 ML2340CCQ/5	±½ LSB	0°C to +70°C -55°C to +125°C -40°C to +85°C 0°C to +70°C 0°C to +70°C	HERMETIC DIP (J18) HERMETIC DIP (J18) HERMETIC DIP (J18) MOLDED DIP (P18) MOLDED PCC (Q18)
V_{REF OUT} = 2.50V with V_{CC} = 5V			
ML2350BMJ/5 ML2350BIJ/5 ML2350BCP/5 ML2350BCQ/5	±¼ LSB	-55°C to +125°C -40°C to +85°C 0°C to +70°C	HERMETIC DIP (J18) HERMETIC DIP (J18) MOLDED DIP (P18) MOLDED PCC (Q18)
ML2350CMJ/5 ML2350CIJ/5 ML2350CCP/5 ML2350CCQ/5	±½ LSB	0°C to +70°C -55°C to +125°C -40°C to +85°C 0°C to +70°C 0°C to +70°C	HERMETIC DIP (J18) HERMETIC DIP (J18) HERMETIC DIP (J18) MOLDED DIP (P18) MOLDED PCC (Q18)
V_{REF OUT} = 4.50V with V_{CC} = 12V			
ML2340BMJ/12 ML2340BIJ/12 ML2340BCP/12 ML2340BCQ/12	±¼ LSB	-55°C to +125°C -40°C to +85°C 0°C to +70°C	HERMETIC DIP (J18) HERMETIC DIP (J18) MOLDED DIP (P18) MOLDED PCC (Q18)
ML2340CMJ/12 ML2340CIJ/12 ML2340CCP/12 ML2340CCQ/12	±½ LSB	0°C to +70°C -55°C to +125°C -40°C to +85°C 0°C to +70°C 0°C to +70°C	HERMETIC DIP (J18) HERMETIC DIP (J18) HERMETIC DIP (J18) MOLDED DIP (P18) MOLDED PCC (Q18)
V_{REF OUT} = 5.00V with V_{CC} = 12V			
ML2350BMJ/12 ML2350BIJ/12 ML2350BCP/12 ML2350BCQ/12	±¼ LSB	-55°C to +125°C -40°C to +85°C 0°C to +70°C	HERMETIC DIP (J18) HERMETIC DIP (J18) MOLDED DIP (P18) MOLDED PCC (Q18)
ML2350CMJ/12 ML2350CIJ/12 ML2350CCP/12 ML2350CCQ/12	±½ LSB	0°C to +70°C -55°C to +125°C -40°C to +85°C 0°C to +70°C 0°C to +70°C	HERMETIC DIP (J18) HERMETIC DIP (J18) HERMETIC DIP (J18) MOLDED DIP (P18) MOLDED PCC (Q18)

ML2341, ML2351

Single Supply, Programmable 8-Bit D/A Converters

GENERAL DESCRIPTION

The ML2341 and ML2351 are CMOS voltage output, 8-bit D/A converters with an internal voltage reference and a μP interface. These devices are designed to be powered by a single supply, although they can be powered from dual power supplies. The output voltage swings above zero scale (V_{ZS}) in the unipolar mode or around zero scale (V_{ZS}) in the bipolar mode, both with programmable gain. V_{ZS} can be set to any voltage from AGND to 2.25V below V_{CC} . The digital and analog grounds, DGND and AGND, are totally independent of each other. DGND can be set to any voltage from AGND to 4.5V below V_{CC} for easy interfacing to standard TTL and CMOS logic families.

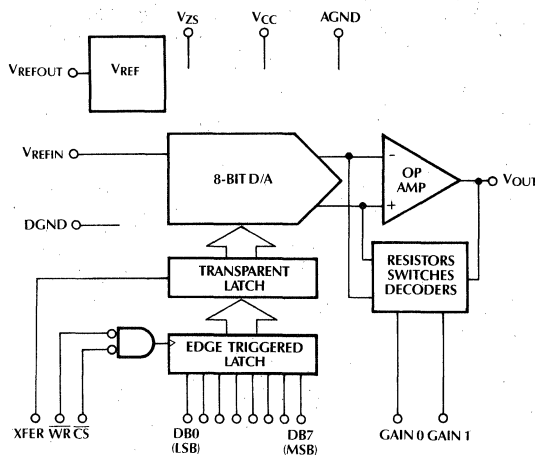
The high level of integration and versatility of the ML2341 and ML2351 makes them ideal for a wide range of applications in hard disk drives, automotive, telecom, and a variety of general purpose industrial. One specific intended application is controlling a hard disk voice coil.

The ML2341 provides a 2.25V or 4.50V reference output for use with A/D converters that use a single 5V \pm 10% power supply, while the ML2351 provides a 2.50V or 5.00V reference output.

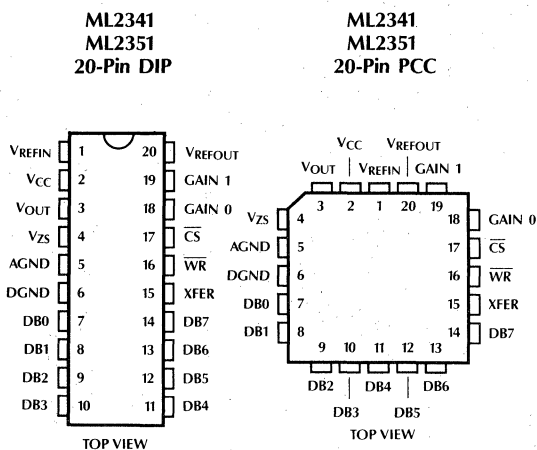
FEATURES

- Programmable output voltage gain settings of 2, 1, $\frac{1}{2}$, $\frac{1}{4}$ provide 8-, 9-, 10-, or 11-bit effective resolution around zero
- AGND to V_{CC} output voltage swing
- Bipolar or unipolar output voltage
- 4.5V to 13.2V single supply or $\pm 2.25\text{V}$ to $\pm 6.5\text{V}$ dual-supply operation
- Single- and double-buffered, edge-triggered interface with 30ns write time, 0ns hold time
- Voltage reference output
 - ML2341 2.25V or 4.50V
 - ML2351 2.50V or 5.00V
- Nonlinearity $\pm \frac{1}{4}$ LSB or $\pm \frac{1}{2}$ LSB
- Output voltage settling time over temperature and supply voltage tolerance
 - Within 1V of V_{CC} and AGND $2.5\mu\text{s}$ max
 - Within 100mV of V_{CC} and AGND $5\mu\text{s}$ max
- TTL and CMOS compatible digital inputs
- Low supply current ($V_{REF} \leq 2.5\text{V}$) 5mA max
- 20-pin DIP or PCC
- Operating temperature range of 0°C to +70°C, -40°C to +85°C, and -55°C to +125°C

BLOCK DIAGRAM



PIN CONNECTIONS



PIN DESCRIPTION

PIN #	NAME	FUNCTION	PIN #	NAME	FUNCTION
1	V _{REF IN}	Voltage reference input. V _{REF IN} is referenced to AGND.	10	DB3	Data input — Bit 3.
2	V _{CC}	Positive supply.	11	DB4	Data input — Bit 4.
3	V _{OUT}	Voltage output of the D/A converter. V _{OUT} is referenced to V _{ZS} .	12	DB5	Data input — Bit 5.
4	V _{ZS}	Zero Scale Voltage. V _{OUT} is referenced to V _{ZS} . V _{ZS} is normally tied to AGND in the unipolar mode or to mid-supply in the bipolar mode. When the device is operated from a single power supply, V _{ZS} has a maximum current requirement of -300µA in the bipolar mode.	13	DB6	Data input — Bit 6.
5	AGND	Analog ground.	14	DB7	Data input — Bit 7 (MSB).
6	DGND	Digital ground. This is the ground reference level for all digital inputs. The range is AGND < (V _{CC} - 4.5V). DGND is normally tied to system ground.	15	XFER	Transfer enable input. In the double buffered mode of operation, the data in the input latch is transferred to the D/A converter at the high level of XFER.
7	DB0	Data input — Bit 0 (LSB).	16	WR	Write enable input. While CS is low, data inputs are latched into the input latch on the rising edge of WR.
8	DB1	Data input — Bit 1.	17	CS	Chip select input. Active low input which enables latching in the data on the rising edge of WR.
9	DB2	Data input — Bit 2.	18	GAIN 0	Digital gain setting input 0.
			19	GAIN 1	Digital gain setting input 1.
			20	V _{REF OUT}	Voltage reference output. V _{REF OUT} is referenced to AGND. V _{REF OUT} is set to 2.5V and 5.0V in a low-voltage and high-voltage operation, respectively for the ML2351; 2.25V and 4.5V for the ML2341.

ABSOLUTE MAXIMUM RATINGS

(Note 1)

Supply Voltage V _{CC} with Respect to AGND	14.2V
DGND	-0.3V to V _{CC} + 0.3V
V _{ZS} , V _{REF IN}	-0.3V to V _{CC} + 0.3V
Logic Inputs	-0.3V to V _{CC} + 0.3V
Input Current per Pin (Note 2)	±25mA
Storage Temperature	-65°C to +150°C
Package Dissipation at T _A = 25°C (Board Mount)	875mW
Lead Temperature (Soldering 10 sec.)	
Dual-In-Line Package (Plastic)	260°C
Dual-In-Line Package (Ceramic)	300°C
Molded Chip Carrier Package	
Vapor Phase (60 sec.)	215°C
Infrared (15 sec.)	220°C

OPERATING CONDITIONS

(Note 1)

Supply Voltage, V _{CC}	4.5V _{DC} to 13.2V _{DC}
Temperature Range (Note 3)	T _{MIN} ≤ T _A ≤ T _{MAX}
ML2341BMJ, ML2341CMJ	
ML2351BMJ, ML2351CMJ	-55°C to +125°C
ML2341BIJ, ML2341CIJ	
ML2351BIJ, ML2351CIJ	-40°C to +85°C
ML2341BCQ, ML2341CCQ	
ML2351BCQ, ML2351CCQ	
ML2341BCP, ML2341CCP	
ML2351BCP, ML2351CCP	0°C to +70°C

ML2341, ML2351

ELECTRICAL CHARACTERISTICS

Unless otherwise specified, $T_A = T_{MIN}$ to T_{MAX} , $V_{CC} - AGND = 5V \pm 10\%$ and $12V \pm 10\%$ (Note 9), $V_{REF IN}$ for ML2341 = 2.25V and 4.50V, for ML2351 $V_{REF IN} = 2.50V$ and 5.00V, V_{OUT} load is $R_L = 1K$ and $C_L = 100pF$, V_{REF} load is $R_L = 1K$ and $C_L = 100pF$ and input control signals with $t_R = t_F \leq 20ns$.

PARAMETER	NOTES	CONDITIONS	ML2341XCX, ML2351XCX			ML2341XIX, ML2341XXM ML2351XIX, ML2351XXM			UNITS
			MIN	TYP (Note 4)	MAX	MIN	TYP (Note 4)	MAX	
Converter and Programmable Gain Amplifier									
Converter Resolution	5		8			8			Bits
Integral Linearity Error ML2341BXX, ML2351BXX ML2341CXX, ML2351CXX	5	GAIN = 2, 1, 1/2, or 1/4			$\pm 1/4$ $\pm 1/2$			$\pm 1/4$ $\pm 1/2$	LSB LSB
Differential Linearity Error ML2341BXX, ML2351BXX ML2341CXX, ML2351CXX	5	GAIN = 2, 1, 1/2, or 1/4			$\pm 1/4$ $\pm 1/2$			$\pm 1/4$ $\pm 1/2$	LSB LSB
Mode Select Unipolar Output Bipolar Output	5	V_{ZS} with respect to AGND	0 1.50		1.0 $V_{CC}-2.25$	0 1.50		1.0 $V_{CC}-2.25$	V V
Offset Error Unipolar Mode	5	Figure 1 GAIN = 1/4, 1/2, 1 GAIN = 2			± 10 ± 20			± 12 ± 24	mV mV
Bipolar Mode	5	Figure 1 GAIN = 1/4, 1/2, 1, 2			± 10 plus $\pm 2 1/2$ LSB			± 10 plus $\pm 2 1/2$ LSB	mV
Gain Error Unipolar Mode Bipolar Mode	5	Figure 1 GAIN = 1/4, 1/2, 1, 2 GAIN = 1/4, 1/2, 1, 2			± 5 ± 5	± 2 ± 2		± 5 ± 5 ± 2.5 ± 2.5	%FS %FS
Reference									
$V_{REF OUT}$ Voltage ML2341BXX/5	5	$V_{CC} = 5.0V$ $T_A = 25^\circ C$ T_{MIN} to T_{MAX}	2.23 2.22	2.25	2.27 2.28	2.23 2.18	2.25	2.27 2.32	V V
ML2341CXX/5		$V_{CC} = 5.0V$ $T_A = 25^\circ C$ T_{MIN} to T_{MAX}	2.22 2.20	2.25	2.29 2.30	2.22 2.18	2.25	2.28 2.32	V V
ML2351BXX/5	5	$V_{CC} = 5.0V$ $T_A = 25^\circ C$ T_{MIN} to T_{MAX}	2.48 2.47	2.50	2.52 2.53	2.48 2.43	2.50	2.52 2.57	V V
ML2351CXX/5		$V_{CC} = 5.0V$ $T_A = 25^\circ C$ T_{MIN} to T_{MAX}	2.45 2.44	2.50	2.55 2.58	2.46 2.42	2.50	2.55 2.59	V V
ML2341BXX/12	5	$V_{CC} = 12.0V$ $T_A = 25^\circ C$ T_{MIN} to T_{MAX}	4.48 4.46	4.50	4.52 4.54	4.48 4.43	4.50	4.52 4.57	V V
ML2341CXX/12		$V_{CC} = 12.0V$ $T_A = 25^\circ C$ T_{MIN} to T_{MAX}	4.45 4.40	4.50	4.55 4.60	4.45 4.35	4.50	4.55 4.65	V V
ML2351BXX/12	5	$V_{CC} = 12.0V$ $T_A = 25^\circ C$ T_{MIN} to T_{MAX}	4.98 4.96	5.00	5.02 5.04	4.98 4.90	5.00	5.02 5.10	V V
ML2351CXX/12		$V_{CC} = 12.0V$ $T_A = 25^\circ C$ T_{MIN} to T_{MAX}	4.95 4.90	5.00	5.05 5.10	4.95 4.85	5.00	5.05 5.15	V V
Temperature Coefficient $V_{REF OUT}$				50			50		ppm/°C
V_{REF} Output Current	5		0.75		5		0.75	5	mA
$V_{REF OUT}$ Power Supply Rejection Ratio	5	100mV _{p-p} , 1kHz Sinewave on V_{CC}	-40	-60		-40	-60		dB

ELECTRICAL CHARACTERISTICS (Continued)

Unless otherwise specified, $T_A = T_{MIN}$ to T_{MAX} , $V_{CC} - AGND = 5V \pm 10\%$ and $12V \pm 10\%$ (Note 9), $V_{REF IN}$ for ML2341 = 2.25V and 4.50V; for ML2351 $V_{REF IN} = 2.50V$ and 5.00V, V_{OUT} load is $R_L = 1K$ and $C_L = 100pF$, V_{REF} load is $R_L = 1K$ and $C_L = 100pF$ and input control signals with $t_R = t_F \leq 20ns$.

PARAMETER	NOTES	CONDITIONS	ML2341XCX, ML2351XCX			ML2341XMX, ML2351XMX			UNITS	
			MIN	TYP (Note 4)	MAX	MIN	TYP (Note 4)	MAX		
$V_{REF IN}$ and V_{ZS}										
$V_{REF IN}$ Input Range	5	$V_{CC} \leq 8.75V$ $V_{CC} \geq 8.75V$	AGND+2 AGND+2		$V_{CC}-1.75$ AGND+7	AGND+2 AGND+2		$V_{CC}-1.75$ AGND+7	V V	
$V_{REF IN}$ DC Input Resistance	5		10			10			MΩ	
V_{ZS} Voltage Range	5, 8	$V_{CC} \leq 7.0V$	AGND		$V_{CC}-2.25$	AGND		$V_{CC}-2.25$	V	
	5, 8	$V_{CC} \geq 8.0V$	AGND		$V_{CC}-3.0$	AGND		$V_{CC}-3.0$	V	
Analog Output										
V_{OUT} Output Swing Unipolar Mode	5, 8	$R_L = 100K$	AGND+ 0.01		$V_{CC}-0.5$	AGND+ 0.01		$V_{CC}-0.5$	V	
			AGND+ 1.0		$V_{CC}-1.0$	AGND+ 1.0		$V_{CC}-1.0$	V	
	Bipolar Mode	5	$R_L = 100K$	AGND+ 0.1		$V_{CC}-0.1$	AGND+ 0.1		$V_{CC}-0.1$	V
				AGND+ 1.0		$V_{CC}-1.0$	AGND+ 1.0		$V_{CC}-1.0$	V
V_{OUT} Output Current	5	$AGND+1V < V_{OUT} < V_{CC}-1V$	-10		+10	-10		+10	mA	
Power Supply Rejection Ratio		100mV _{p-p} , 1kHz sinewave on V_{CC}		-60			-60		dB	
Digital and DC										
$V_{IN(0)}$ Logical "0" Input Voltage	5				0.8			0.8	V	
$V_{IN(1)}$ Logical "1" Input Voltage	5		2.0			2.0			V	
$I_{IN(0)}$ Logical "0" Input Current	5	$V_{IN} = DGND$	-1			-1			μA	
$I_{IN(1)}$ Logical "1" Input Current	5	$V_{IN} = V_{CC}$			1			1	μA	
Supply Current, Bipolar Mode I_{CC} , V_{CC} Current I_{AGND} , Analog Ground Current I_{VZS} , V_{ZS} Current	5	$V_{CC} = 5V \pm 10\%$			5.3			5.3	mA	
					-5.0 -300		-90 -300		-5.0 -300 μA	
I_{CC} , V_{CC} Current I_{AGND} , Analog Ground Current I_{VZS} , V_{ZS} Current	5	$V_{CC} = 12V \pm 10\%$			9.3			9.3	mA	
					-9.0 -300		-90 -300		-9.0 -300 μA	

2

ML2341, ML2351

ELECTRICAL CHARACTERISTICS (Continued)

Unless otherwise specified, $T_A = T_{MIN}$ to T_{MAX} , $V_{CC} - AGND = 5V \pm 10\%$ and $12V \pm 10\%$ (Note 9), $V_{REF IN}$ for ML2341 = 2.25V and 4.50V; for ML2351 $V_{REF IN} = 2.50V$ and 5.00V, V_{OUT} load is $R_L = 1k$ and $C_L = 100pF$, V_{REF} load is $R_L = 1K$ and $C_L = 100pF$ and input control signals with $t_r = t_f \leq 20ns$.

PARAMETER	NOTES	CONDITIONS	ML2341XCX, ML2351XCX			ML2341XIX, ML2341XMX ML2351XIX, ML2351XMX			UNITS	
			MIN	TYP (Note 4)	MAX	MIN	TYP (Note 4)	MAX		
Digital and DC (Continued)										
Supply Current, Unipolar Mode	5	$V_{CC} = 5V \pm 10\%$								
I_{CC} V_{CC} Current					6.0			6.0	mA	
I_{AGND} , Analog Ground Current	5	$V_{CC} = 12V \pm 10\%$								
I_{VZS} , V_{ZS} Current					-4.3			-4.3	mA	
I_{VZS} , V_{ZS} Current										
I_{CC} V_{CC} Current					11.0			11.0	mA	
I_{AGND} , Analog Ground Current					-7.3			-7.3	mA	
I_{VZS} , V_{ZS} Current					-3.7			-3.7	mA	
AC Performance										
Settling Time	5	Figure 2, Output Step of AGND + 1V to $V_{CC} - 1V$, $R_L = 1K$								
t_{S1}				1.2	2.5		1.2	3.0	μs	
t_{S2}				Output Step of AGND + 100mV to $V_{CC} - 100mV$, $R_L = 100K$	2.5	5		2.5	6	μs
t_{S3}				Output Step of $\pm 1LSB$		1			1	μs
t_{S4} Gain Change		Change of Any Gain Setting	1.1	2.5		1.1	3	μs		
t_{WR} , WR Pulse Width	5	Figure 3	40			40		ns		
t_{XFER} , XFER Pulse Width	5	Figure 3	60			60		ns		
t_{XW} , WR1 to XFER1	6	Figure 3	30			30		ns		
t_{DBS} , DB0-DB7 Setup Time	5	Figure 3	40			45		ns		
t_{DBH} , DB0-DB7 Hold Time	5	Figure 3	0			0		ns		
t_{CSS} , CS Setup Time	5	Figure 3	50			50		ns		
t_{CSH} , CS Hold Time	5	Figure 3	0			0		ns		
t_{RESET} , Power-On Reset Time	6				16			16	μs	

Note 1: Absolute maximum ratings are limits beyond which the life of the integrated circuit may be impaired. All voltages unless otherwise specified are measured with respect to analog ground.

Note 2: When the voltage at any pin exceeds the power supply rails ($V_{IN} < AGND$ or $V_{IN} > V_{CC}$) the absolute value of current at that pin should be limited to 25mA or less.

Note 3: -55°C to +125°C operating temperature range devices are 100% tested at temperature extremes with worst-case test conditions. -40°C to +85°C operating temperature range devices are 100% tested with temperature limits guaranteed by 100% testing, sampling, or by correlation with worst-case test conditions.

Note 4: Typicals are parametric norm at 25°C.

Note 5: Parameter guaranteed and 100% production tested.

Note 6: Parameter guaranteed. Parameters not 100% tested are not in outgoing quality level calculation.

Note 7: Supply current and analog ground current are specified with the digital inputs stable and no load on V_{OUT} .

Note 8: In unipolar operation with V_{ZS} and AGND tied together, digital codes that represent an analog value of less than 100mV from AGND should be avoided.

Note 9: ML2341XXX/5 and ML2351XXX/5 are tested for 5V operation only and ML2341XXX/12 and ML2351XXX/12 are tested for 12V operation.

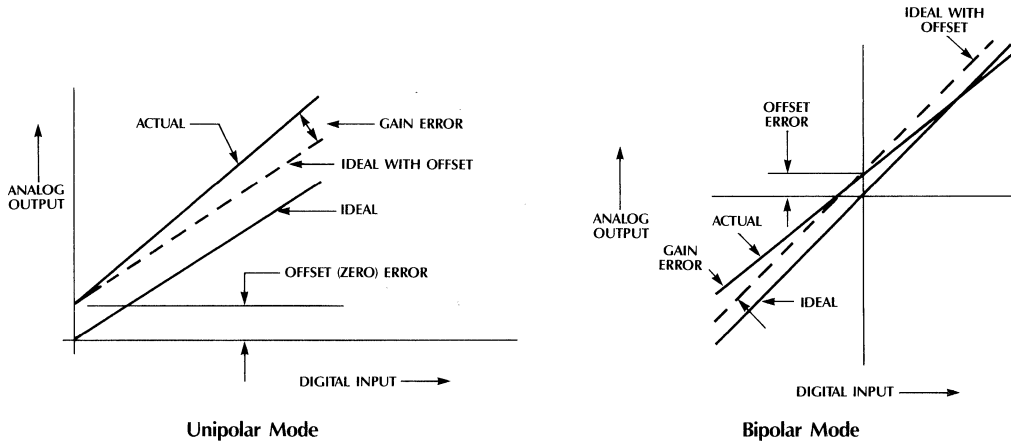


Figure 1. Gain and Offset Error

2

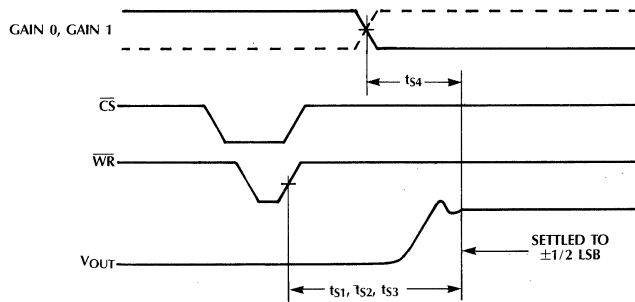


Figure 2. Settling Time

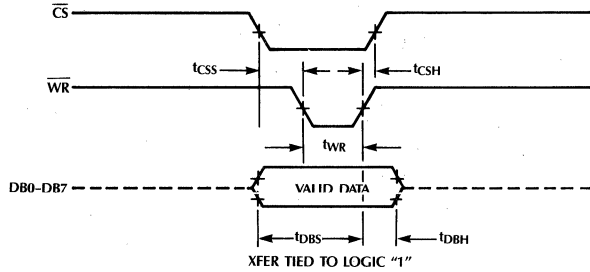


Figure 3a. Single Buffered Mode

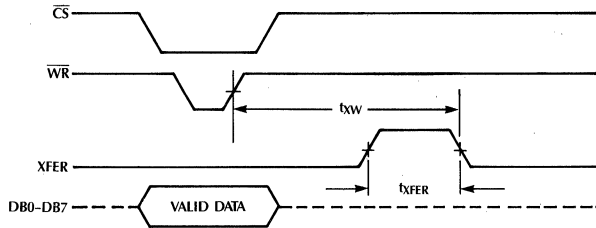


Figure 3b. Double Buffered Mode

1.0 FUNCTIONAL DESCRIPTION

1.1 D/A CONVERTER

The D/A converter is implemented using an array of equal current sources that are decoded semi linearly for the four most significant bits to improve differential linearity and to reduce output glitch around major carries. See Figure 4.

The input voltage reference of the D/A converter is the difference between $V_{REF IN}$ and AGND. This difference voltage is converted to a reference current using an internal resistor to set up the appropriate current level

in the D/A converter. The D/A converter output current is then converted to a voltage output by an output buffer and a resistive network. The matching among the on-chip resistors preserves the gain accuracy between these conversions.

The D/A converter can be used in a multiplying mode by modulating the reference input within the specified $V_{REF IN}$ range.

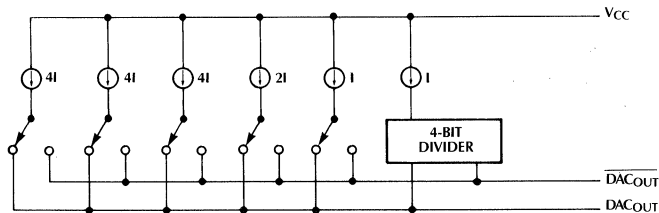


Figure 4. D/A Converter Implementation

1.2 SINGLE-SUPPLY vs. DUAL-SUPPLY OPERATION

ML2341 and ML2351 can be powered from a single supply ranging from 4.5V to 13.2V or dual supplies ranging from ±2.25V to ±6.6V.

The internal digital and analog circuitry is powered between V_{CC} and AGND. The range of DGND is AGND ≤ DGND ≤ V_{CC} - 4.5V with the logic thresholds set between .8V and 2.0V above DGND (standard TTL logic level). The range of V_{ZS} is AGND ≤ V_{ZS} ≤ (V_{CC} - 2.25V).

1.3 UNIPOLAR AND BIPOLAR OUTPUT VOLTAGE SWING

The ML2341 and ML2351 can operate in either unipolar and bipolar output voltage mode. Unipolar/bipolar mode selection is determined by comparing the zero scale voltage (V_{ZS}) of these devices to a precise internal reference that is referred to AGND. V_{ZS} is ideally the voltage that will be produced at the DAC voltage output when the digital input data is set to all "0's". Unipolar mode is selected when V_{ZS} is lower than 1.00 volt, and bipolar mode is selected when V_{ZS} is greater than 1.50 volts.

1.3.1 Unipolar Output Mode

In the unipolar mode, V_{OUT} swings above V_{ZS}. Ideally the 00000000 code results in an output voltage of V_{ZS}, and the 11111111 code results in an output voltage of V_{FS} × 255/256, where V_{FS} is the full-scale voltage determined by V_{REF IN} and the gain setting.

1.3.2 Bipolar Output Mode

In the bipolar mode, V_{OUT} swings around V_{ZS}. The input data is in 2's complement binary format. Ideally, the 00000000 code results in an output voltage of V_{ZS}; the 10000000 code results in an output voltage of (V_{ZS} - V_{FS}); and the 01111111 results in an output voltage of (V_{ZS} + V_{FS} 127/128), where V_{FS} is the full scale output voltage determined by V_{REF IN} and the gain setting.

1.4 OUTPUT BUFFER AND GAIN SETTING

The output buffer converts the D/A output current to a voltage output using a resistive network with proper gain setting determined by the GAIN 0 and GAIN 1 inputs. There are four possible gain settings for unipolar output voltage mode and bipolar output voltage mode as listed below:

Unipolar Output Voltage Mode

GAIN 1	GAIN 0	GAIN	Voltage Output Swing Relative to V _{ZS}
0	0	1/4	V _{REF IN} × 1/4
0	1	1/2	V _{REF IN} × 1/2
1	0	1	V _{REF IN} × 1
1	1	2	V _{REF IN} × 2

Bipolar Output Voltage Mode

GAIN 1	GAIN 0	GAIN	Voltage Output _{p-p}
0	0	1/4	±V _{REF IN} × 1/8
0	1	1/2	±V _{REF IN} × 1/4
1	0	1	±V _{REF IN} × 1/2
1	1	2	±V _{REF IN} × 1

The output buffer can source or sink as much as 10mA of current with an output voltage of at least 1V from either V_{CC} or AGND. As the output voltage approaches V_{CC} or AGND the current sourcing/sinking capability of the output buffer is reduced. The output buffer can still swing down to within 10mV of AGND and up to within 40mV of V_{CC} with a 100K load at V_{OUT} to AGND in the unipolar operation. In the bipolar operation, the output buffer swing is limited to about 100mV from either rails.

1.5 VOLTAGE REFERENCE

A bandgap voltage reference is incorporated on the ML2341 and ML2351. Two reference voltages can be produced by each device. An internal comparator monitors the power supply voltage to determine the selection of the reference voltage. A reference voltage of 2.25 volts on the ML2341 and 2.50 volts on the ML2351 is selected when the supply voltage is less than approximately 7.50 volts. Otherwise, a reference voltage of 4.50 volts and 5.00 volts is selected. To prevent the comparator from oscillating between the two selections, avoid operation with a power supply between 7.0 and 8.0 volts.

The bandgap reference is trimmed for zero Temperature Coefficient (TC) at 35°C to minimize output voltage drift over the specified operating temperature range.

The internal reference is buffered for use by the DAC and external circuits. The reference buffer will source more than 5mA of current and sink more than 1mA of current. With V_{REF IN} connected to V_{REF OUT}, the following output voltage ranges of the DAC are obtained:

ML2341

Gain Setting	V _{REF} = 2.25V with V _{CC} ≤ 7.0V		V _{REF} = 4.5V with V _{CC} ≥ 8.0V	
	Unipolar	Bipolar	Unipolar	Bipolar
1/4	0 to 0.562V	-0.281V to +0.281V	0 to 1.125V	-0.562V to +0.562V
1/2	0 to 1.125V	-0.562V to +0.562V	0 to 2.250V	-1.125V to +1.125V
1	0 to 2.250V	-1.125V to +1.125V	0 to 4.500V	-2.250V to +2.250V
2	0 to 4.500V	-2.250V to +2.250V	0 to 9.000V	-4.500V to +4.500V

2

ML2341, ML2351

ML2351

Gain Setting	$V_{REF} = 2.50V$ with $V_{CC} \leq 7.0V$		$V_{REF} = 5.00V$ with $V_{CC} \geq 8.0V$	
	Unipolar	Bipolar	Unipolar	Bipolar
1/4	0 to 0.625V	-0.3125V to +0.3125V	0 to 1.25V	-0.625V to +0.625V
1/2	0 to 1.250V	-0.6250V to +0.6250V	0 to 2.50V	-1.250V to +1.250V
1	0 to 2.500V	-1.2500V to +1.2500V	0 to 5.00V	-2.500V to +2.500V
2	0 to 5.000V	-2.5000V to +2.5000V	0 to 10.00V	-5.000V to +5.000V

An external reference can alternatively be used on $V_{REF IN}$ to set the desired full scale voltage. The linearity of the D/A converter depends on the reference used, however. To insure integral linearity at an 8-bit level, a reference voltage of no less than 2V and no more than 7V (2.75V for operation with a low-voltage power supply) should be used.

1.6 DIGITAL INTERFACE

The digital interface of the device consists of a chip select input, \overline{CS} , a write input, \overline{WR} , a transfer input, \overline{XFER} and eight data inputs, $DB0$ through $DB7$. The digital interface operates in one of the two modes:

1.6.1 Single-Buffered Mode

To use the ML2341 and ML2351 in the single-buffered mode, tie \overline{XFER} to logic "1". This will put the D/A latch in the transparent mode and the rising edge of \overline{WR} at low level of \overline{CS} will latch the data on $DB0$ - $DB7$ into the input latch as well as update the D/A output voltage.

1.6.2 Double-Buffered Mode

To use the devices in the double-buffered mode, timing information is applied to \overline{WR} as well as \overline{XFER} inputs. The rising edge of \overline{WR} at low level of \overline{CS} will latch the data on $DB0$ - $DB7$ into the input latch. The D/A output voltage will not be updated, however, until \overline{XFER} is brought to a high level, which transfers the data from input latch to D/A latch. Note that the D/A latch is a transparent latch controlled by the level, not edge, of the \overline{XFER} input, any write operation to the input latch while \overline{XFER} is still at a high level results in the immediate update of the D/A output voltage.

1.7 POWER-ON-RESET

The ML2341 and ML2351 have an internal power-on-reset circuit to initialize the device when power is first applied to the device. The power-on-reset interval of typically $8\mu s$ begins when the supply voltage, V_{CC} reaches approximately 2.0V. During the power-on-reset interval, both the input and data latch are reset to all "0's".

2.0 TYPICAL APPLICATIONS

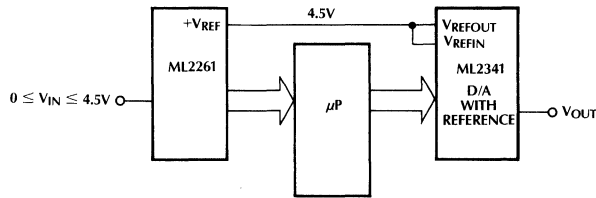


Figure 5. Using 4.50V Reference of D/A for Reference of A/D Using Single 5V V_{CC} ± 10%

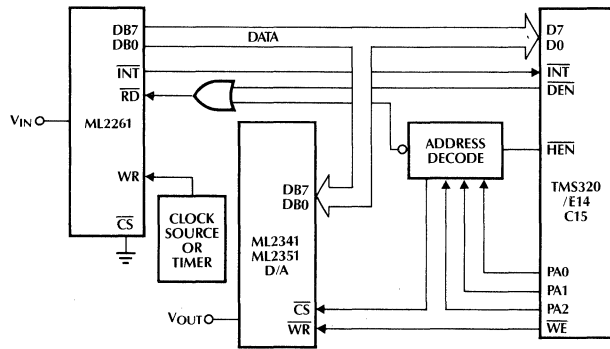


Figure 6. TMS320 Interface with D/A Output

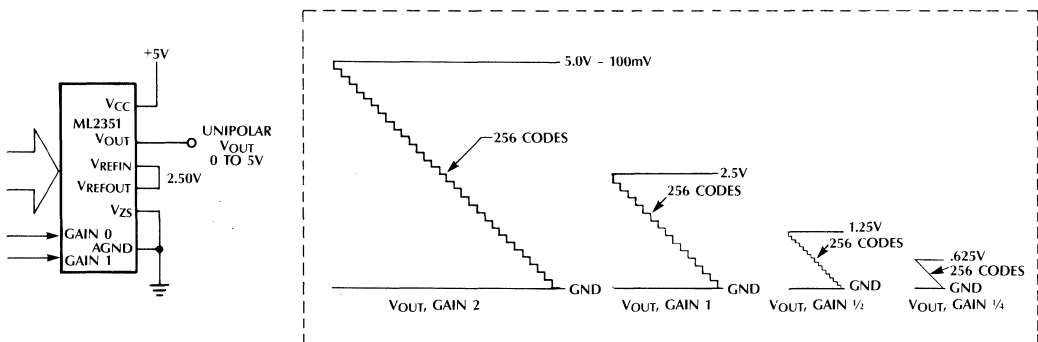


Figure 7. Single 5V Supply Unipolar V_{OUT}

2

ML2341, ML2351

2.0 TYPICAL APPLICATIONS (Continued)

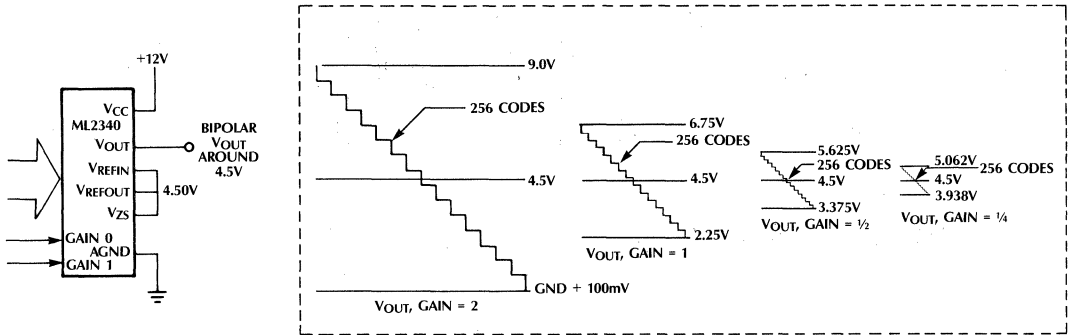


Figure 8. Single 12V Supply, Bipolar V_{OUT} with 11-Bits Resolution Around Zero

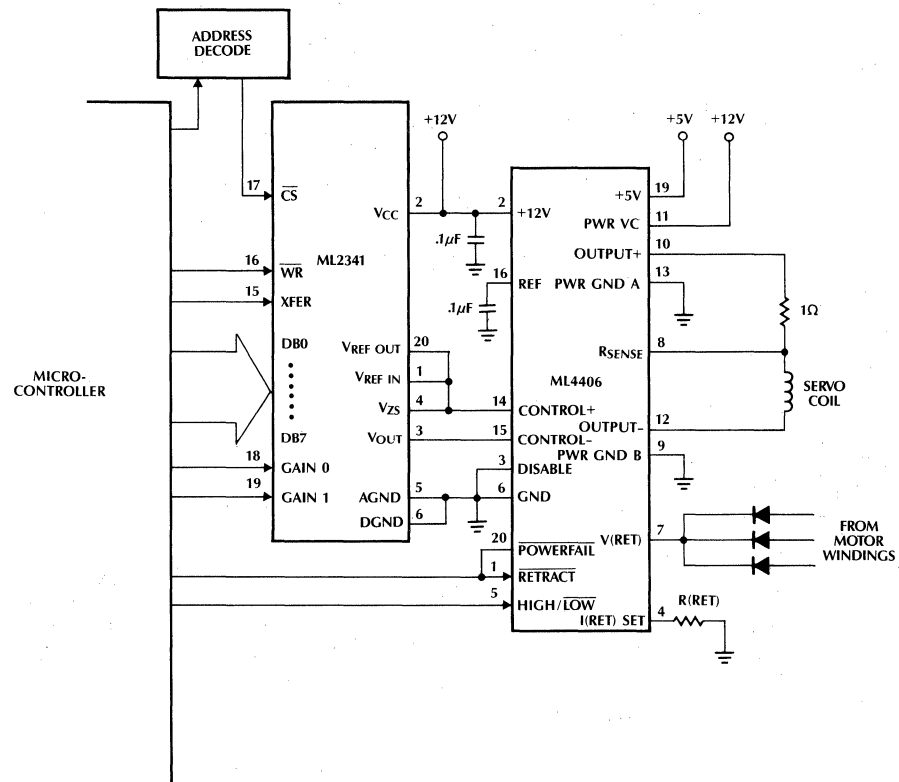


Figure 9. Hard Disc Drive Servo Coil Driver Providing 13-Bit Effective Resolution

ORDERING INFORMATION

PART NUMBER	INTEGRAL & DIFFERENTIAL NON-LINEARITY	TEMPERATURE RANGE	PACKAGE
V_{REF OUT} = 2.25V with V_{CC} = 5V			
ML2341BMJ/5 ML2341BIJ/5 ML2341BCP/5 ML2341BCQ/5	±¼ LSB	-55°C to +125°C -40°C to +85°C 0°C to +70°C 0°C to +70°C	HERMETIC DIP (J20) HERMETIC DIP (J20) MOLDED DIP (P20) MOLDED PCC (Q20)
ML2341CMJ/5 ML2341CIJ/5 ML2341CCP/5 ML2341CCQ/5	±½ LSB	-55°C to +125°C -40°C to +85°C 0°C to +70°C 0°C to +70°C	HERMETIC DIP (J20) HERMETIC DIP (J20) MOLDED DIP (P20) MOLDED PCC (Q20)
V_{REF OUT} = 2.50V with V_{CC} = 5V			
ML2351BMJ/5 ML2351BIJ/5 ML2351BCP/5 ML2351BCQ/5	±¼ LSB	-55°C to +125°C -40°C to +85°C 0°C to +70°C 0°C to +70°C	HERMETIC DIP (J20) HERMETIC DIP (J20) MOLDED DIP (P20) MOLDED PCC (Q20)
ML2351CMJ/5 ML2351CIJ/5 ML2351CCP/5 ML2351CCQ/5	±½ LSB	-55°C to +125°C -40°C to +85°C 0°C to +70°C 0°C to +70°C	HERMETIC DIP (J20) HERMETIC DIP (J20) MOLDED DIP (P20) MOLDED PCC (Q20)
V_{REF OUT} = 4.50V with V_{CC} = 12V			
ML2341BMJ/12 ML2341BIJ/12 ML2341BCP/12 ML2341BCQ/12	±¼ LSB	-55°C to +125°C -40°C to +85°C 0°C to +70°C 0°C to +70°C	HERMETIC DIP (J20) HERMETIC DIP (J20) MOLDED DIP (P20) MOLDED PCC (Q20)
ML2341CMJ/12 ML2341CIJ/12 ML2341CCP/12 ML2341CCQ/12	±½ LSB	-55°C to +125°C -40°C to +85°C 0°C to +70°C 0°C to +70°C	HERMETIC DIP (J20) HERMETIC DIP (J20) MOLDED DIP (P20) MOLDED PCC (Q20)
V_{REF OUT} = 5.00V with V_{CC} = 12V			
ML2351BMJ/12 ML2351BIJ/12 ML2351BCP/12 ML2351BCQ/12	±¼ LSB	-55°C to +125°C -40°C to +85°C 0°C to +70°C 0°C to +70°C	HERMETIC DIP (J20) HERMETIC DIP (J20) MOLDED DIP (P20) MOLDED PCC (Q20)
ML2351CMJ/12 ML2351CIJ/12 ML2351CCP/12 ML2351CCQ/12	±½ LSB	-55°C to +125°C -40°C to +85°C 0°C to +70°C 0°C to +70°C	HERMETIC DIP (J20) HERMETIC DIP (J20) MOLDED DIP (P20) MOLDED PCC (Q20)

DSP Analog I/O Peripheral

GENERAL DESCRIPTION

The ML2377 is a complete analog I/O peripheral front-end for DSP based control system. It contains a high-speed 10-bit A/D converter, a two channel simultaneous sample/hold circuit, a 6 channel input multiplexer, a 10-bit D/A converter and a 8-bit D/A converter.

The two channel simultaneous sample/hold in conjunction with the multiple channel multiplexer provided on-chip is especially well suited for disk drive applications, where minimum skew positional channel conversion and flexible calibration sensing functions are desirable.

Both input and output channel voltages are referenced to floating common points provided by the device. An additional common point is also available in the A/D input for flexibility. Bipolar conversion of ± 2 volts around the floating point is provided by the chip.

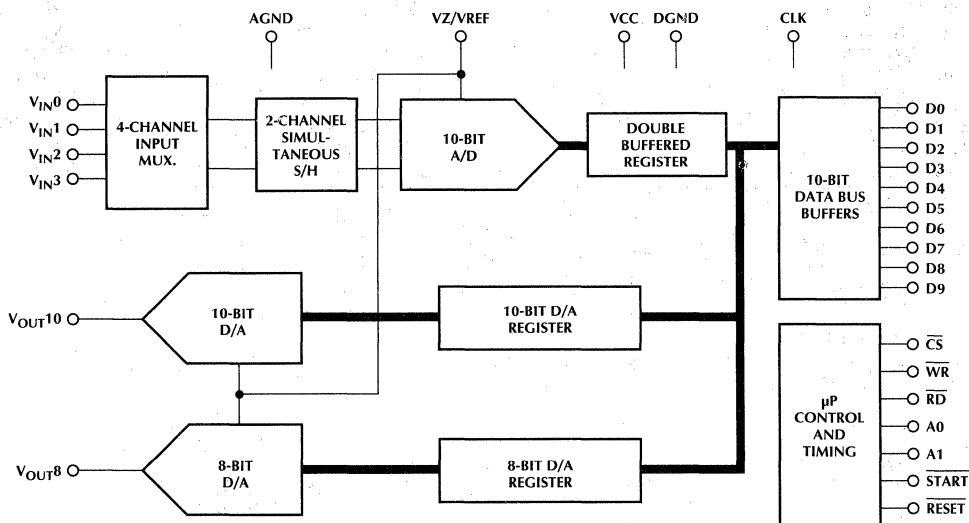
Channel multiplexing and common referencing control are provide on-chip via its easy to use microprocessing port. In addition, external control of the conversion start and MUX addressing is also available for the ML2377.

The ML2375 is a 4 channel version of the ML2377.

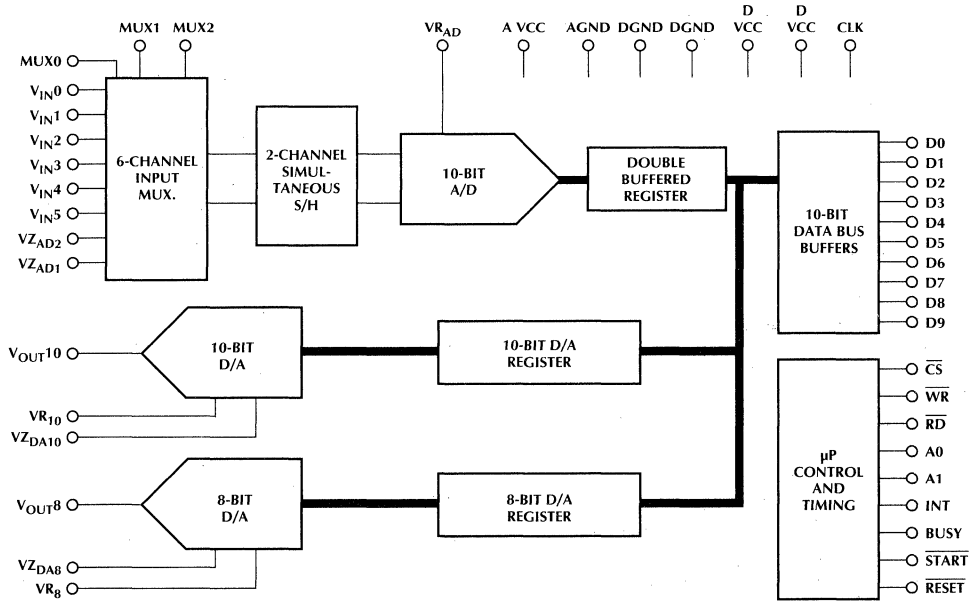
FEATURES

- 10-bit bipolar A/D resolution
- 10-bit bipolar and 8-bit bipolar D/A resolution
- 2 μ s A/D conversion time
- 2 μ s D/A settling time (ILSB, 4 μ s full scale)
- 2 channel simultaneous S/H
- 6 bipolar input channels
- A/D and D/A converters have no missing codes
- Inputs and outputs have floating commons
- ± 2 volt input/output range with 2.5V reference
- Extra floating common input for A/D
- Programmable input MUX and common
- TMS320C14 compatible microprocessor interface
- Single 5 volt power supply
- External conversion start and MUX control (ML2377)
- Additional package and bond-out options available

BLOCK DIAGRAM ML2375

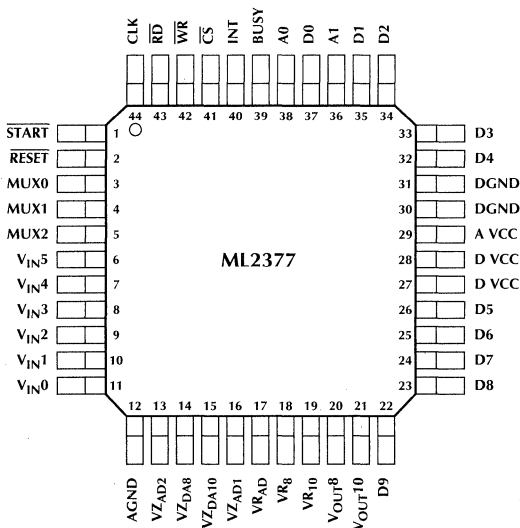


BLOCK DIAGRAM ML2377

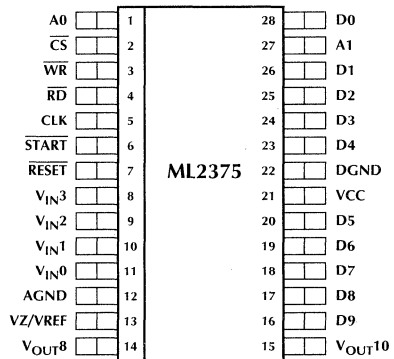


PIN CONNECTION

ML2377
44-Pin QFP (G44)



ML2375
28-Pin SSOP (R28)



ML2375, ML2377

PIN DESCRIPTION

NAME	FUNCTION	NAME	FUNCTION
VZ _{AD2}	Programmable input common	V _{OUT8}	Voltage output of the 8-bit D/A converter.
VZ _{AD1}	Default A/D input common.	V _{OUT10}	Voltage output of the 10-bit D/A converter.
VZ _{DA10}	10-bit D/A common.	D9-D0	Data I/O bit 9 through 0.
VZ _{DA8}	8-bit D/A common.	DVCC	Digital power supply. +5 volts $\pm 5\%$.
$\overline{\text{START}}$	Active low input starts A/D converter.	AVCC	Analog power supply. +5 volts $\pm 5\%$.
$\overline{\text{RESET}}$	Active low resets the IC.	DGND	Digital ground.
MUX0	Multiplexer address bit 0.	A1	Register address 1.
MUX1	Multiplexer address bit 1.	A0	Register address 0.
MUX2	Multiplexer address bit 2.	BUSY	Active high output indicates that an A/D conversion is in progress.
V _{IN5-0}	Analog input channels 5 through 0.	INT	Active high output indicating A/D conversion complete.
AGND	Analog ground.	$\overline{\text{CS}}$	Active low chip select input.
VR _{AD}	Voltage reference input for establishing \pm full scale for the A/D converter. \pm full scale value is 0.8 of the voltage on VR _{AD} , referenced to AGND.	$\overline{\text{WR}}$	Write input, active low.
VR ₈	\pm full scale value for the 8-bit D/A converter is 0.8 of the voltage reference input on VR ₈ , referenced to AGND.	$\overline{\text{RD}}$	Read input, active low.
VR ₁₀	\pm full scale value for the 10-bit D/A converter is 0.8 of the voltage reference input on VR ₁₀ , referenced to AGND.	CLK	Clock input. Clock can be generated by tying a crystal from this pin to DGND or applying a clock directly to pin.

Note: ALE (Address Latch Enable) for demultiplexing address and data information can be made available on request. Consult Micro Linear for more information.
Other pin-out options of the ML2377 are available on request.

ABSOLUTE MAXIMUM RATINGS

Supply Voltages (AV_{CC} and DV_{CC}) 6.0V
 Maximum Voltage Between AGND and DGND 1V
 Maximum Voltage Between AV_{CC} and DV_{CC} 0.3V
 Input Current per Pin ± 25 mA
 Package Dissipation @ 25°C 1W
 Lead Temperature (Soldering, 10 sec) 300°C

OPERATING CONDITIONS

Temperature Range 0°C to +70°C
 Supply Voltage (AV_{CC} and DV_{CC}) 4.5V_{DC} to 6.0V_{DC}

Absolute maximum ratings are those values beyond which the device could be permanently damaged. Absolute maximum ratings are stress ratings only and functional device operation is not implied.

ELECTRICAL CHARACTERISTICS

The following specifications apply for $AV_{CC} = DV_{CC} = +5V \pm 5\%$, AGND = DGND = 0V, T_A = Operating Temperature Range $V_{ZAD1} = V_{ZAD2} = V_{ZDA10} = V_{ZDA8} = V_{RAD} = V_{R8} = V_{R10} = 2.5V$, $T_A = T_{MIN}$ to T_{MAX} unless otherwise specified, $C_L = 50$ pF for all digital outputs, V_{OUT8} and V_{OUT10} load is $R_L = 1K$ and $C_L = 100$ pF, and input control signals with $t_R = t_F = 20$ ns, $f_{CLK} = 11$ MHz.

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
A/D Converter, S/H and Multiplexer Performance						
	Integral Linearity Error				± 3	LSB
	Differential Linearity Error				± 1	LSB
	Converter Resolution	$V_{REF} = 2.5V$	10			Bits
	Zero Error	$V_Z = 2.5V$			± 3	LSB
	Positive and Negative Full Scale Error	$V_{REF} = 2.5V$			± 3	LSB
	Input Voltage Range		0		V_{CC}	V
	DC Common Mode Error	$V_Z = 2V$, $V_{CC} - 2V$			$\pm 1/4$	LSB
	AC Common Mode Error	0 to 250kHz		TBD		LSB
	AC Power Supply Sensitivity	100mV _{P-P} 100kHz Sinewave on V_{CC}		TBD		LSB
	V_{REF} Input Resistance		2		4	k Ω
	Maximum V_{REF} Input Voltage	Referred to AGND			2.6	V
I_{ON}	On Channel Leakage Current	$0V < V_{IN} < V_{CC}$	-1		1	μA
I_{OFF}	Off Channel Leakage Current	$0V < V_{IN} < V_{CC}$	-1		1	μA
C_{ON}	On Channel Input Capacitance	$0V < V_{IN} < V_{CC}$		20		pF
C_{OFF}	Off Channel Input Capacitance	$0V < V_{IN} < V_{CC}$		10		pF
	Clock Duty Cycle		30		70	%
f_{CLK}	Input Clock Frequency		1		11	MHz
t_C	Conversion Time	Including S/H Acquisition Time		22		$1/f_{CLK}$
t_{ACQ}	Acquisition Time	Included in Conversion Time		2		$1/f_{CLK}$
SNR	Signal to Noise Ratio			TBD		dB
THD	Total Harmonic Distortion			TBD		dB
IMD	Intermodulation Distortion			TBD		dB

ML2375, ML2377

ELECTRICAL CHARACTERISTICS (continued)

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
10-Bit D/A Converter Performance						
	Integral Linearity Error	$V_{REF} = 2.5V$			± 3	LSB
	Differential Linearity Error	$V_{REF} = 2.5V$			± 1	LSB
	Settling Time	$V_{REF} = 2V$, Settling to $\pm 1/2LSB$, $\pm 1LSB$ Step 4V Step			2 4	μs μs
	Resolution	$V_{REF} = 2.5V$	10			Bits
	Zero Error				± 5	LSB
	+ and - Full Scale Error				$\pm 1\%$	FS
	Output Voltage Swing	$V_{REF} = 2.5V$	0.25		$V_{CC} - 0.25$	V
	Power Supply Sensitivity	100mV _{p-p} , 1kHz Sinewave on V_{CC}		TBD		dB
	V_{REF} Input Resistance			1		M Ω

8-Bit D/A Converter Performance

	Integral Linearity Error	$V_{REF} = 2.5V$			± 1	LSB
	Differential Linearity Error	$V_{REF} = 2.5V$			± 1	LSB
	Settling Time	$V_{REF} = 2V$, Settling to $\pm 1/2LSB$, $\pm 1LSB$ Step 4V Step			2 4	μs μs
	Resolution	$V_{REF} = 2.5V$	8			Bits
	Zero Error				± 3	LSB
	+ and - Full Scale Error				± 3	LSB
	Output Voltage Range	$V_{REF} = 2.5V$	0.25		$V_{CC} - 0.25$	V
	Power Supply Sensitivity	100mV _{p-p} , 1kHz Sinewave on V_{CC}		TBD		dB
	V_{REF} Input Resistance			1		M Ω

DC Characteristics

I_{IL}	Logic Input Current	$0 < V_{IN} < V_{CC}$, MUX0-2, \overline{START} , RESET, ALE			± 100	μA
I_{IN}	Logic Input Current	$0 < V_{IN} < V_{CC}$			± 1	μA
I_{INC}	Clock Input Current	$0 < V_{IN} < V_{CC}$			± 200	μA
V_{IH}	Logic High		2			V
V_{IL}	Logic Low				0.8	V
$V_{(CLK)H}$	Clock High	(CLK Pin)	3.7			V
$V_{(CLK)L}$	Clock LOW	(CLK Pin)			1.8	V
I_{OFF}	Output Leakage Current	$\overline{CS} = V_{IH}$, $0 < V_{OUT} < V_{CC}$			± 1	μA
V_{OL}	Output Low	$I_{OL} = 2mA$			0.4	V
V_{OH}	Output High	$I_{OH} = -1mA$	2.4			V

AC ELECTRICAL CHARACTERISTICS

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
--------	-----------	------------	-----	-----	-----	-------

Supply Current ($V_{REF} = 2.5V$, No Output Load)

	No Input Switching			10	20	mA
	Analog Ground Current	$V_{REF} = 2.5V$			10	mA

Microprocessor Interface Read Cycle Timing

t_{ST}	START Pulse Width		100			ns
t_{AD}	Address Stable to Data Valid				55	ns
t_{AR}	Address Stable Before Read		5			ns
t_{RA}	Address Hold After Read		5			ns
t_{RR}	Read Pulse Width		50			ns
t_{RD}	Data Valid from Read				50	ns
t_{DF}	Read to Data Float				20	ns
t_{RV}	Recovery Time Between Reads		25			ns

Microprocessor Interface Write Cycle Timing

t_{AW}	Address Stable Before Write		5			ns
t_{WA}	Address Hold After Write		5			ns
t_{WW}	Write Pulse Width		50			ns
t_{DS}	Data Valid Before Write		30			ns
t_{DH}	Data Hold After Write		15			ns
t_{RV}	Recovery Time Between Writes		25			ns

Microprocessor Interface Interrupt and Busy Timing

t_{CI}	Clock to Interrupt Active				50	ns
t_{RI}	Read to Interrupt Inactive				100	ns
t_{CB}	Clock to Busy Active	(ML2377 Only)			50	ns
t_{CB}	Clock to Busy Inactive	(ML2377 Only)			50	ns

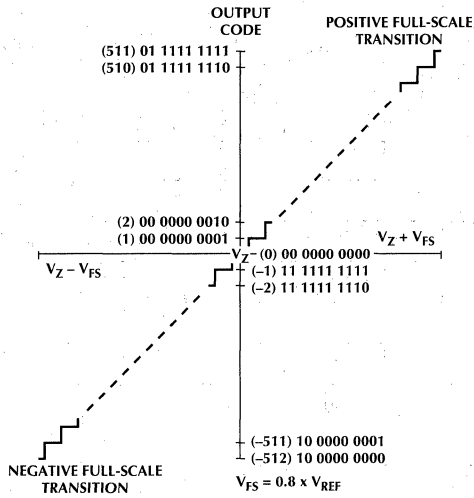


Figure 1. Transfer Characteristics

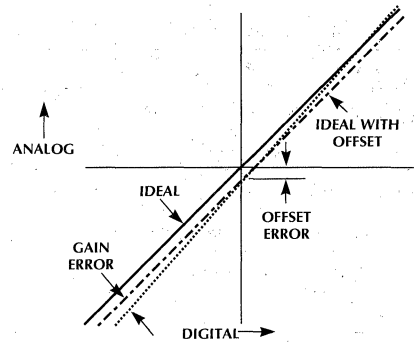


Figure 2. Gain and Offset Error

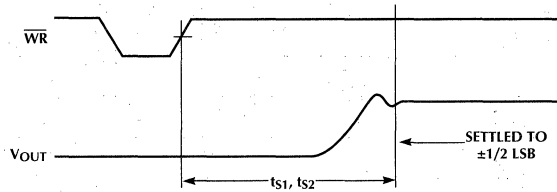


Figure 3. Settling Time

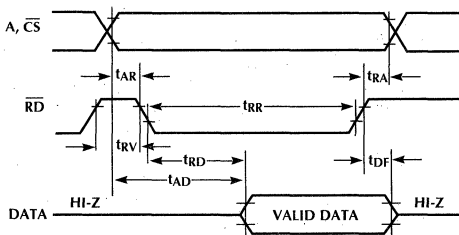


Figure 4. Read Cycle Timing

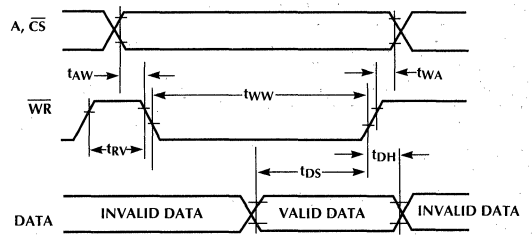


Figure 5. Write Cycle Timing

FUNCTIONAL DESCRIPTION

DEVICE TYPES

The family consists of two different devices:

- 1) The ML2375
It is a 28 pin device that contains the basic 10-bit A/D converter, 10-bit D/A converter, 8-bit D/A converter and a 4 channel multiplexer.
- 2) The ML2377
It is a 6 channel version of ML2375. In addition, the BUSY, INT and multiplexer control pins are also available to the user.

ANALOG INPUT AND OUTPUT VOLTAGES

The ML2375 and ML2377 allow the analog input and output voltages to be referenced to a common point. Thus, the input voltage swing and the offset of the ADC and the output voltage swing and the offset of the DAC can be defined by the voltage applied at the commons, $V_{Z_{AD1}}$, $V_{Z_{AD2}}$, $V_{Z_{DA10}}$ and $V_{Z_{DA8}}$.

The voltage at the common for the 10-bit D/A converter and the 8-bit D/A converter are defined by $V_{Z_{DA10}}$ and $V_{Z_{DA8}}$ respectively. For the A/D inputs, channel 0 and 1 common are defined by $V_{Z_{AD1}}$. The common of the other channels (2 to 5) may be programmed to $V_{Z_{AD2}}$ by setting the on-chip control register (see Table 3).

The peak full scale voltage is defined by the reference voltages.

$$V_{FS} = 0.8 \times V_{REF}$$

Figure 1 shows the transfer function of the ML2375 and ML2377 and the relationship between V_Z and V_{FS} .

A/D CONVERTER OPERATION

Input Multiplexer Addressing

The input multiplexer is addressed with either the MUX0 to MUX2 pins or the internal multiplexer address register. The MUX0 to MUX2 pins are not latched, and control the addressing of the multiplexer directly. If control of the multiplexer is to be done from the microprocessor addressable mux control register, then all MUX pins should be tied to a logic high. This condition, which is an illegal mux address, will then route control of the multiplexer addressing to the internal register, which is under microprocessor control.

ML2375 multiplexer can only be addressed via its on-chip multiplexer register.

Simultaneous Sample/Hold Function

The simultaneous sample/hold function is only available on channels 0 and 1. When addressing channel 0 or 1 with the external MUX pins or the internal register bits and starting a conversion, both channels will be sampled and held simultaneously. Two conversions will then proceed back to back, with the BUSY pin and status bit going active for two conversion times. The INT pin and status bit will go active after the completion of the first

conversion. A read of the data register will then clear the INT pin and status bit. After the second conversion is completed, the INT pin and status bit will go active again, indicating the need for a read of the second result from the data register. If, however, the results of the first register are not read after the first conversion and before the second conversion, the INT pin and status bit remain active until the completion of the second conversion. The results of the first conversion remain in the data register, with the results of the second conversion stored in a holding register. After completion of the second conversion, which is indicated by the BUSY status bit or pin going inactive, both results can be obtained by successive reads of the data register. The INT pin and status bit are then cleared by the act of the second data register read.

D/A CONVERTER OPERATION

The D/A converters are updated beginning on the rising edge of the WR pin. Settling time is measured from this point.

Microprocessor Interface

The ML2375 is presented as four 10-bit registers to the microprocessor. These registers are addressed via the address pins A0 and A1. The register map below describes the four registers.

Converter Registers

Table 1. A/D Converter Register

ADDRESS 00										
	D9	D8	D7	D6	D5	D4	D3	D2	D1	D0
Data Bit	9	8	7	6	5	4	3	2	1	0
	Sign	MSB								LSB

Table 2. D/A Converter Register

ADDRESS 01										
	D9	D8	D7	D6	D5	D4	D3	D2	D1	D0
D/A 10	9	8	7	6	5	4	3	2	1	0
	Sign	MSB								LSB

ADDRESS 10										
	D9	D8	D7	D6	D5	D4	D3	D2	D1	D0
D/A 8	7	6	5	4	3	2	1	0		
	Sign	MSB						LSB	0	0

ML2375, ML2377

REGISTER DESCRIPTIONS

Register Address 00

This register holds the results of the 10-bit A/D conversion results when read. The converted results are in 2's complement form, where 0 is the potential at the common pin. Reading of this register also clears the INT status bit and deasserts the INT pin on the falling edge of the RD pin.

Register Address 01

When written, it receives the 10-bit digital value for the 10-bit D/A converter. All codes are in 2's complement form, where the 0 code indicates the potential at the VZ_{DA10} or VZ/V_{REF} pin. This register can also be read, which returns the previously written value.

Register Address 10

This register receives the 8-bit digital value for the 8-bit D/A converter. This value is also in 2's complement format. When read, it returns the previously written value. 8-bit data in this 10-bit register is left justified.

Control Register

This is the control register. Functions such as A/D conversion start, multiplexer control, and A/D converter status are included.

- D9 When $VZ_5 = 1$, Channel 5 common = VZ_{AD2}
- D8 When $VZ_4 = 1$, Channel 4 common = VZ_{AD2}
- D7 When $VZ_3 = 1$, Channel 3 common = VZ_{AD2}
- D6 When $VZ_2 = 1$, Channel 2 common = VZ_{AD2}
- D5 INT. It is equivalent to the INT pin. It provides indication that a conversion is completed.
- D4 BUSY. This bit is A/D converter status bit which provides indication that a conversion is in progress. It is equivalent to the BUSY pin of the ML2377.
- D3 START. This bit has the same function as the START pin. When this bit receives a 1, it will start a conversion. After the conversion has started, this bit is cleared after 4 clock cycles.
- D2 MUX2. MUX address bit 2.
- D1 MUX1. MUX address bit 1.
- D0 MUX0. MUX address bit 0.

CHIP RESET

The chip is reset when a 0 is presented to the \overline{RESET} pin. All registers are reset to 0. Therefore, the D/A and the A/D converters are all at zero scale and the multiplexer select is addressed to channel 0. Additionally, VZ_2-VZ_5 are cleared in the control register.

Table 3. Control Register

ADDRESS 11	D9	D8	D7	D6	D5	D4	D3	D2	D1	D0
ML2375	0	0	0	0	INT	BUSY	START	0	MUX1	MUX0
ML2377	VZ_5	VZ_4	VZ_3	VZ_2	INT	BUSY	START	MUX2	MUX1	MUX0

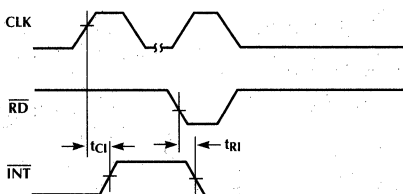


Figure 6. Interrupt Timing, ML2375, and ML2377

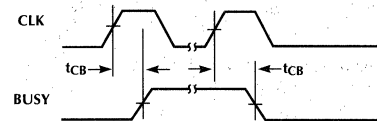


Figure 7. BUSY Timing, ML2377

ORDERING INFORMATION

PART NUMBER	PACKAGE	COMMENTS
ML2375CCP ML2375CCQ ML2375CCR	28-Pin DIP (P28) 28-Pin PCC (Q28) 28-Pin SSOP (R28)	Multiplex Address and Data Bus
ML2377CCH	44-Pin QFP (H44)	Additional Analog Inputs, Individual V_{REF} and V_{ZS}



Telecom Communications

Section 3

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Selection Guide

Gain/Attenuators

Part Number	Gain Range (dB)	Resolution (dB Steps)	Noise (dBnc @ Max Gain)	Harmonic Distortion (dB)	Digital Interface	Power Supplies (V)	Temperature Range		Package
							C	I	
ML2003	-24 to +24	0.1	0	-60	Serial, Hard Wire	±5	X	X	18-pin DIP 20-pin PCC
ML2004	-24 to +24	0.1	0	-60	Serial	±5	X	X	14-pin DIP
ML2008	-24 to +24	0.1	0	-60	8-Bit μ P	±5	X	X	18-pin DIP 20-pin PCC
ML2009	-24 to +24	0.1	0	-60	16-Bit μ P	±5	X	X	18-pin DIP 20-pin PCC

Equalizers

Part Number	Frequency Response Adjustable	Idle Channel Noise (dBnc)	Harmonic Distortion (dB)	Comment	Interface	Power Supplies (V)	Temperature Range		Package
							C	I	
ML2020	Slope, Height, Bandwidth	8	-48	60 Hz Rejection	Serial	±5	X	X	16-pin DIP 18-pin SOIC
ML2021	Slope, Height, Bandwidth	8	-48	Group Delay Optimized	Serial	±5	X	X	16-pin DIP 18-pin SOIC

Tone Detectors

Part Number	Detect Frequency (Hz)	Dynamic Range Detect (dBm)	Frequency Template (Hz)	Comment	Power Supplies (V)	Temperature Range		Package
						C	I	
ML2031	1K to 4K	-34 to +6	Detect ± 10 No Detect ± 36	Exceed Bell Pub 43004, Clock Outputs of CLK _N $\div 2, \div 8$	±5	X	X	8-pin DIP
ML2032	1K to 4K	-34 to +6	Detect ± 10 No Detect ± 36	Exceed Bell Pub 43004, Uncommitted Op Amp	±5	X	X	8-pin DIP

Programmable Sinewave Generators

Part Number	Frequency Range (Hz)	Min Resolution (Hz)	Gain Error (dB)	Harmonic Distortion (dB)	Comment	Digital Interface	Power Supplies (V)	Temperature Range		Package
								C	I	
ML2035	DC to 25K	±.75	±.1	-45	Voltage Amplitude $V_{CC}/2$	Serial	±5	X	X	8-pin DIP
ML2036	DC to 50K	±.75	±.1	-45	Adj. Voltage Amplitude, Clock Outputs of CLK _N $\div 2, \div 8$	Serial	±5	X	X	14-pin DIP 16-pin SOIC

Logarithmic Gain/Attenuator

GENERAL DESCRIPTION

The ML2003 and ML2004 are digitally controlled logarithmic gain/attenuators with a range of -24 to +24dB in 0.1dB steps.

The gain settings are selected by a 9-bit digital word. The ML2003 digital interface is either parallel or serial. The ML2004 is packaged in a 14-pin DIP with a serial interface only.

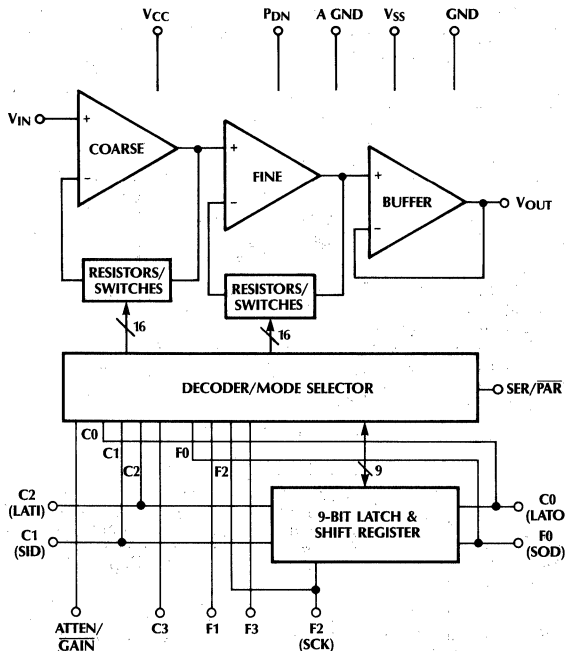
Absolute gain accuracy is 0.05dB max over supply tolerance of $\pm 10\%$ and temperature range.

These CMOS logarithmic gain/attenuators are designed for a wide variety of applications in telecom, audio, sonar, or general purpose function generation. One specific intended application is analog telephone lines.

FEATURES

- Low noise 0 dBnc max with +24dB gain
- Low harmonic distortion -60dB max
- Gain range -24 to +24dB
- Resolution 0.1dB steps
- Flat frequency response ± 0.05 dB from .3-4kHz
 ± 0.10 dB from .1-20kHz
- Low supply current 4mA max from ± 5 V supplies
- TTL/CMOS compatible digital interface
- ML2003 has pin selectable serial or parallel interface; ML2004 serial interface only
- Standard 14-pin or 18-pin 0.3" center DIP or 20-pin molded chip carrier package

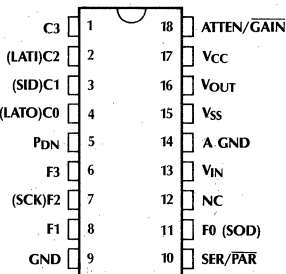
BLOCK DIAGRAM



NOTE: SERIAL MODE FUNCTIONS INDICATED BY PARENTHESES.

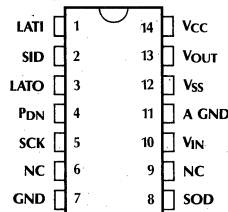
PIN CONNECTIONS

**ML2003
18-PIN DIP**



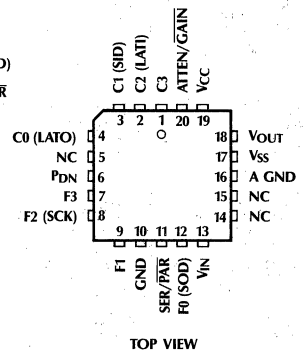
TOP VIEW

**ML2004
14-PIN DIP**



TOP VIEW

**ML2003
20-PIN PCC**



TOP VIEW

PIN DESCRIPTION

NAME	FUNCTION	NAME	FUNCTION
C3	In serial mode, pin is unused. In parallel mode, coarse gain select bit. Pin has internal pulldown resistor to GND.	F1	In serial mode, pin is unused. In parallel mode, fine gain select bit. Pin has internal pulldown resistor to GND.
(LATI) C2	In serial mode, input latch clock which loads the data from the shift register into the latch. In parallel mode, coarse gain select bit. Pin has internal pulldown resistor to GND.	GND	Digital ground. 0 volts. All digital inputs and output are referenced to this ground.
(SID) C1	In serial mode, serial data input that contains serial 9 bit data word which controls the gain setting. In parallel mode, coarse gain select bit. Pin has internal pulldown resistor to GND.	SER/ <u>PAR</u>	Serial or parallel select input. When SER/ <u>PAR</u> = 1, device is in serial mode. When SER/ <u>PAR</u> = 0, device is in parallel mode. Pin has internal pullup resistor to V _{CC} .
(LATO) C0	In serial mode, output latch clock which loads the 9 bit data word back into the shift register from the latch. In parallel mode, coarse gain select bit. Pin has internal pulldown resistor to GND.	(SOD) F0	In serial mode, serial output data which is the output of the shift register. In parallel mode, fine gain select bit. Pin has internal pulldown resistor to GND.
P _{DN}	Powerdown input. When P _{DN} = 1, device is in powerdown mode. When P _{DN} = 0, device is in normal operation. Pin has internal pulldown resistor to GND.	V _{IN}	Analog input.
F3	In serial mode, pin is unused. In parallel mode, fine gain select bit. Pin has internal pulldown resistor to GND.	AGND	Analog ground. 0 volts. Analog input and output are referenced to this ground.
(SCK) F2	In serial mode, shift register clock which shifts the serial data on SID into the shift register on rising edges and out on SOD on falling edges. In parallel mode, fine gain select bit. Pin has internal pulldown resistor to GND.	V _{SS}	Negative supply. -5 volts ±10%.
		V _{OUT}	Analog output.
		V _{CC}	Positive supply. +5 volts ±10%.
		ATTEN/ <u>GAIN</u>	In serial mode, pin is unused. In parallel mode, attenuation/gain select bit. Pin has internal pulldown resistor to GND.

3

ABSOLUTE MAXIMUM RATINGS

(Note 1)

Supply Voltage	
V _{CC}	+6.5V
V _{SS}	-6.5V
AGND with respect to GND.....	±.5V
Analog Input and Output.....	V _{SS} - 0.3V to V _{CC} + 0.3V
Digital Inputs and Outputs.....	GND - 0.3V to V _{CC} + 0.3V
Input Current Per Pin.....	±25mA
Power Dissipation.....	750mW
Storage Temperature Range.....	-65°C to +150°C
Lead Temperature (soldering, 10 sec).....	300°C

OPERATING CONDITIONS

Temperature Range (Note 2)	
ML2003CP, ML2004CP, ML2004CQ.....	0°C to +70°C
ML2003IJ, ML2004IJ.....	-40°C to +85°C
Supply Voltage	
V _{CC}	4V to 6V
V _{SS}	-4V to -6V

ML2003, ML2004

ELECTRICAL CHARACTERISTICS

Unless otherwise specified $T_A = T_{MIN}$ to T_{MAX} , $V_{CC} = 5V \pm 10\%$, $V_{SS} = -5V \pm 10\%$, Data Word: $ATTEN/GAIN = 1$, Other Bits = 0 (0dB Ideal Gain), $C_L = 100pF$, $R_L = 600\Omega$, $SCK = LATI = LATO = 0$, dBm measurements use 600Ω as reference load, digital timing measured at 1.4V.

SYMBOL	PARAMETER	NOTES	CONDITIONS	MIN	TYP NOTE 3	MAX	UNITS
Analog							
AG	Absolute gain accuracy	4	$V_{IN} = 8dBm, 1kHz$	-0.05		+0.05	dB
RG	Relative gain accuracy	4	100000001 000000000 000000001 All other gain settings All values referenced to 100000000 gain when $ATTEN/GAIN = 1$, $V_{IN} = 8dBm$ when $ATTEN/GAIN = 0$, $V_{IN} = (8dBm - \text{Ideal Gain})$ in dB	-0.05 -0.05 -0.05 -0.1		+0.05 +0.05 +0.05 +0.1	dB dB dB dB
FR	Frequency response	4	300-4000Hz 100-20,000 Hz Relative to 1kHz	-0.05 -0.1		+0.05 +0.1	dB dB
VOS	Output Offset Voltage	4	$V_{IN} = 0, +24dB$ gain			± 100	mV
ICN	Idle Channel Noise	4 5	$V_{IN} = 0, +24dB$ gain, C msg. Weighted $V_{IN} = 0, +24dB$ gain, 1kHz		-6 450	0 900	dBrc nv/ \sqrt{Hz}
HD	Harmonic Distortion	4	$V_{IN} = 8dBm, 1kHz$ Measure 2nd, 3rd harmonic relative to fundamental			-60	dB
SD	Signal to Distortion	4	$V_{IN} = 8dBm, 1kHz$. C msg. weighted	+60			dB
PSRR	Power Supply Rejection	4	200mV _{P-P} , 1kHz sine, $V_{IN} = 0$ on V_{CC} on V_{SS}		-60 -60	-40 -40	dB dB
Z_{IN}	Input Impedance, V_{IN}	4		1			Meg
V_{INR}	Input Voltage Range	4		± 3.0			V
V_{OSW}	Output Voltage Swing	4		± 3.0			V
Digital and DC							
V_{IL}	Digital Input Low Voltage	4				.8	V
V_{IH}	Digital Input High Voltage	4		2.0			V
V_{OL}	Digital Output Low Voltage	4	$I_{OL} = 2mA$.4	V
V_{OH}	Digital Output High Voltage	4	$I_{OH} = -1mA$	4.0			V
I_{NS}	Input Current, SER/PAR	4	$V_{IH} = GND$	-5		-100	μA
I_{ND}	Input Current, All Digital Inputs Except SER/PAR	4	$V_{IH} = V_{CC}$	5		100	μA
I_{CC}	V_{CC} Supply Current	4	No output load, $V_{IL} = GND$, $V_{IH} = V_{CC}$, $V_{IN} = 0$			4	mA
I_{SS}	V_{SS} Supply Current	4	No output load, $V_{IL} = GND$, $V_{IH} = V_{CC}$, $V_{IN} = 0$			-4	mA
I_{CCP}	V_{CC} Supply Current, Powerdown Mode	4	No output load, $V_{IL} = GND$, $V_{IH} = V_{CC}$.5	mA
I_{SSP}	V_{SS} Supply Current Powerdown Mode	4	No output load, $V_{IL} = GND$, $V_{IH} = V_{CC}$			-1	mA

ELECTRICAL CHARACTERISTICS (Continued)

Unless otherwise specified $T_A = T_{MIN}$ to T_{MAX} , $V_{CC} = 5V \pm 10\%$, $V_{SS} = -5V \pm 10\%$, Data Word: $ATTEN/\overline{GAIN} = 1$, Other Bits = 0 (0dB Ideal Gain), $C_L = 100pF$, $R_L = 600\Omega$, $SCK = LATI = LATO = 0$, dBm measurements use 600Ω as reference load, digital timing measured at 1.4V. $C_L = 100pF$ or SOD.

SYMBOL	PARAMETER	NOTES	CONDITIONS	MIN	TYP NOTE 3	MAX	UNITS
AC Characteristics							
t_{SET}	V_{OUT} Settling Time	4	$V_{IN} = 0.185V$. Change gain from -24 to +24dB. Measure from LATI rising edge to when V_{OUT} settles to within 0.05dB of final value.			20	μs
t_{STEP}	V_{OUT} Step Response	4	Gain = +24dB. $V_{IN} = -0.185V$ to $+0.185V$ step. Measured when V_{OUT} settles to within 0.05dB of final value.			20	μs
t_{SCK}	SCK On/Off Period	4		250			ns
t_S	SID Data Setup Time	4		50			ns
t_H	SID Data Hold Time	4		50			ns
t_D	SOD Data Delay	4		0		125	ns
t_{IPW}	LATI Pulse Width	4		50			ns
t_{OPW}	LATO Pulse Width	4		50			ns
t_{IS}, t_{OS}	LATI, LATO Setup Time	4		50			ns
t_{IH}, t_{OH}	LATI, LATO Hold Time	5		50			ns
t_{PLD}	SOD Parallel Load Delay	4		0		125	ns

Note 1: Absolute maximum ratings are limits beyond which the life of the integrated circuit may be impaired. All voltages unless otherwise specified are measured with respect to ground.

Note 2: 0°C to 70°C and -40°C to +85°C operating temperature range devices are 100% tested with temperature limits guaranteed by 100% testing, sampling, or by correlation with worst-case test conditions.

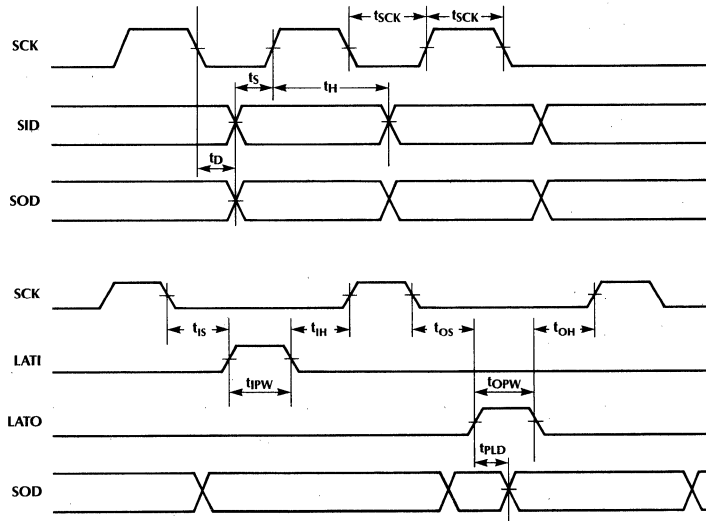
Note 3: Typicals are parametric norm at 25°C.

Note 4: Parameter guaranteed and 100% production tested.

Note 5: Parameter guaranteed. parameters not 100% tested are not in outgoing quality level calculation.

3

TIMING DIAGRAM



TIMING PARAMETERS ARE REFERENCED TO THE 1.4 VOLT MIDPOINT.

Figure 1. Serial Mode Timing Diagram

TYPICAL PERFORMANCE CURVES

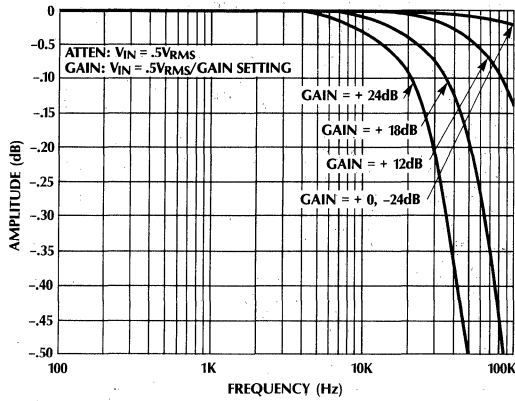


Figure 2. Amplitude vs Frequency ($V_{IN}/V_{OUT} = .5V_{RMS}$)

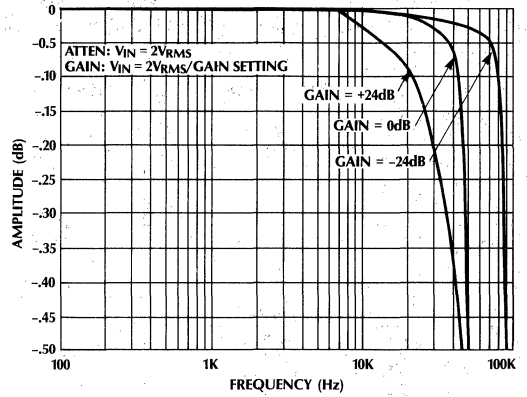


Figure 3. Amplitude vs Frequency ($V_{IN}/V_{OUT} = 2V_{RMS}$)

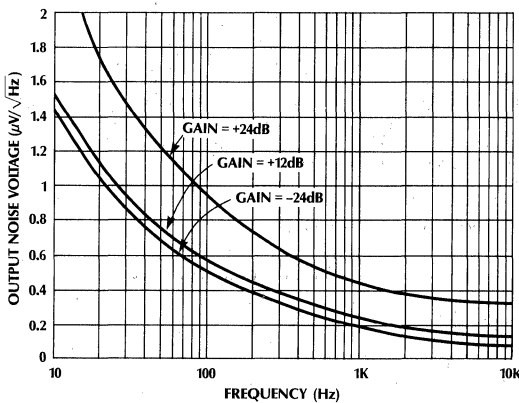


Figure 4. Output Noise Voltage vs Frequency

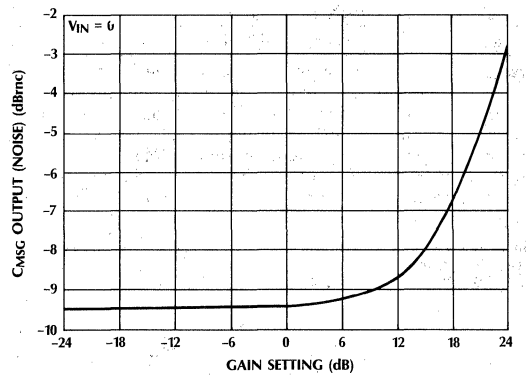


Figure 5. C_{MSG} Output Noise vs Gain Setting

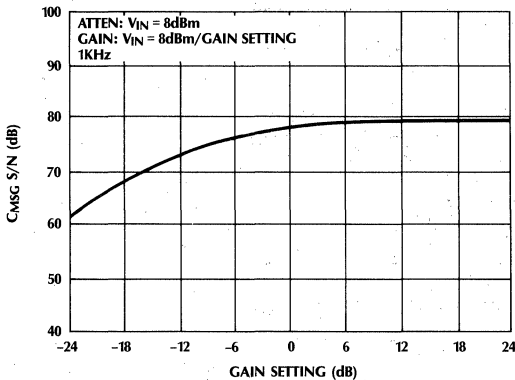


Figure 6. C_{MSG} S/N vs Gain Setting

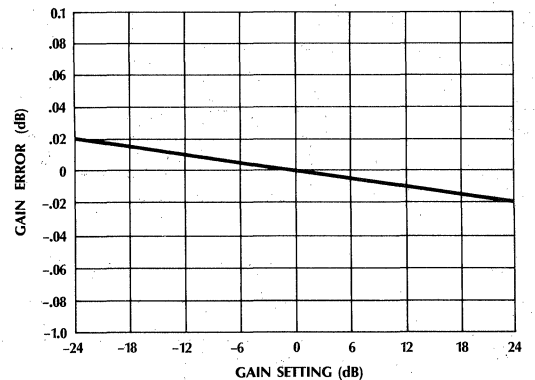
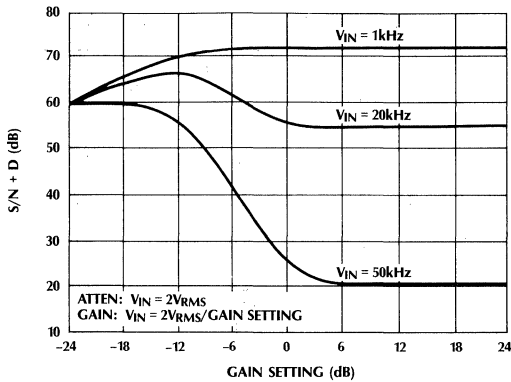
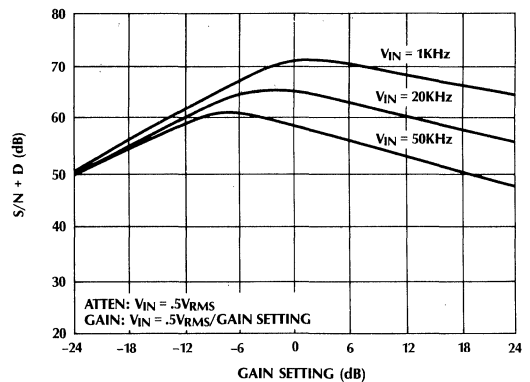


Figure 7. Gain Error vs Gain Setting

TYPICAL PERFORMANCE CURVES (Continued)

Figure 8. S/N + D vs Gain Setting ($V_{IN}/V_{OUT} = 2V_{RMS}$)Figure 9. S/N + D vs Gain Setting ($V_{IN}/V_{OUT} = .5V_{RMS}$)

1.0 FUNCTIONAL DESCRIPTION

The ML2003 consists of a coarse gain stage, a fine gain stage, an output buffer, and a serial/parallel digital interface.

1.1 Gain Stages

The analog input, V_{IN} , goes directly into the op amp input in the coarse gain stage. The coarse gain stage has a gain range of 0 to 22.5dB in 1.5dB steps.

The fine gain stage is cascaded onto the coarse section. The fine gain stage has a gain range of 0 to 1.5dB in 0.1dB steps.

In addition, both sections can be programmed for either gain or attenuation, thus doubling the effective gain range.

The logarithmic steps in each gain stage are generated by placing the input signal across a resistor string of 16 series resistors. Analog switches allow the voltage to be tapped from the resistor string at 16 points. The resistors are sized such that each output voltage is at the proper logarithmic ratio relative to the input signal at the top of the string. Attenuation is implemented by using the resistor string as a simple voltage divider, and gain is implemented by using the resistor string as a feedback resistor around an internal op amp.

1.2 Gain Settings

Since the coarse and fine gain stages are cascaded, their gains can be summed logarithmically. Thus, any gain from -24dB to +24dB in 0.1dB steps can be obtained by

combining the coarse and fine gain settings to yield the desired gain setting. The relationship between the digital select bits and the corresponding analog gain values is shown in Tables 1 and 2. Note that C3-C0 selects the coarse gain, F3-F0 selects the fine gain, and ATTEN/GAIN selects either attenuation or gain.

1.3 Output Buffer

The final analog stage is the output buffer. This amplifier has internal gain of 1 and is designed to drive 600 ohms and 100pF loads. Thus, it is suitable for driving a telephone hybrid circuit directly without any external amplifier.

1.4 Power Supplies

The digital section is powered between V_{CC} and GND, or 5 volts. The analog section is powered between V_{CC} and V_{SS} and uses AGND as the reference point, or ± 5 volts.

GND and AGND are totally isolated inside the device to minimize coupling from the digital section into the analog section. However, AGND and GND should be tied together physically near the device and ideally close to the common power supply ground connection.

Typically, the power supply rejection of V_{CC} and V_{SS} to the analog output is greater than -60dB at 1kHz. If decoupling of the power supplies is still necessary in a system, V_{CC} and V_{SS} should be decoupled with respect to AGND.

FUNCTIONAL DESCRIPTION (Continued)

Table 1. Fine Gain Settings (C3-C0 = 0)

F3	F2	F1	F0	Ideal Gain (dB)	
				ATTEN/GAIN = 1	ATTEN/GAIN = 0
0	0	0	0	.0	.0
0	0	0	1	-.1	.1
0	0	1	0	-.2	.2
0	0	1	1	-.3	.3
0	1	0	0	-.4	.4
0	1	0	1	-.5	.5
0	1	1	0	-.6	.6
0	1	1	1	-.7	.7
1	0	0	0	-.8	.8
1	0	0	1	-.9	.9
1	0	1	0	-1.0	1.0
1	0	1	1	-1.1	1.1
1	1	0	0	-1.2	1.2
1	1	0	1	-1.3	1.3
1	1	1	0	-1.4	1.4
1	1	1	1	-1.5	1.5

Table 2. Coarse Gain Settings (F3-F0 = 0)

C3	C2	C1	C0	Ideal Gain (dB)	
				ATTEN/GAIN = 1	ATTEN/GAIN = 0
0	0	0	0	.0	.0
0	0	0	1	-1.5	1.5
0	0	1	0	-3.0	3.0
0	0	1	1	-4.5	4.5
0	1	0	0	-6.0	6.0
0	1	0	1	-7.5	7.5
0	1	1	0	-9.0	9.0
0	1	1	1	-10.5	10.5
1	0	0	0	-12.0	12.0
1	0	0	1	-13.5	13.5
1	0	1	0	-15.0	15.0
1	0	1	1	-16.5	16.5
1	1	0	0	-18.0	18.0
1	1	0	1	-19.5	19.5
1	1	1	0	-21.0	21.0
1	1	1	1	-22.5	22.5

1.5 Powerdown Mode

A powerdown mode can be selected with pin P_{DN}. When P_{DN} = 1, the device is powered down. In this state, the power consumption is reduced by removing power from the analog section and forcing the analog output, V_{OUT}, to a high impedance state. While the device is in powerdown mode, the digital section is still functional and the current data word remains stored in the latch when in serial mode. When P_{DN} = 0, the device is in normal operation.

1.6 Digital Section

The ML2003 can be operated with a serial or parallel interface. The SER/PAR pin selects the desired interface. When SER/PAR = 1, the serial mode is selected. When SER/PAR = 0, the parallel mode is selected. The ML2004 digital interface is serial only.

1.6.1 Serial Mode

Serial mode is selected by setting SER/PAR pin high. The serial interface allows the gain settings to be set from a serial data word.

The timing for the serial mode is shown in Figure 10. The serial input data, SID, is loaded into a shift register on rising edges of the shift clock, SCK. The data can be parallel loaded into a latch when the input latch signal, LATI, is high. The LATI pulse must occur when SCK is low. In this way, a new data word can be loaded into the shift register without disturbing the existing data word in the latch.

The parallel outputs of the latch control the attenuation/gain setting. The order of the data word bits in the latch is shown in Figure 11. Note that bit 0 is the first bit of the data word clocked into the shift register. Tables 1 and 2 describe how the data word programs the gain.

The device also has the capability to read out the data word stored in the latch. This can be done by parallel loading the data from the latch back into the shift register when the latch signal, LATO, is high. The LATO pulse must occur when SCK is low. Then, the data word can be shifted out of the shift register serially to the output, SOD, on falling edges of the shift clock, SCK.

The loading and reading of the data word can be done continuously or in bursts. Since the shift register and latch circuitry inside the device is static, there are no minimum frequency requirements on the clocks or data pulses. However, there is coupling (typically less than 100µV) of the digital signals into the analog section. This coupling can be minimized by clocking the data bursts in during noncritical intervals or at a frequency outside the analog frequency range.

1.6.2 Parallel Mode

The parallel mode is selected by setting SER/PAR pin low. The parallel interface allows the gain settings to be set with external switches or from a parallel microprocessor interface.

In parallel mode, the shift register and latch are bypassed and connections are made directly to the gain select bits with external pins ATTEN/GAIN, C3-C0, and F3-F0. Tables 1 and 2 describe how these pins program the gain. The pins ATTEN/GAIN, C3-C0, and F3-F0 have internal pulldown resistors to GND. The typical value of these pulldown resistors is 100kΩ.

APPLICATIONS

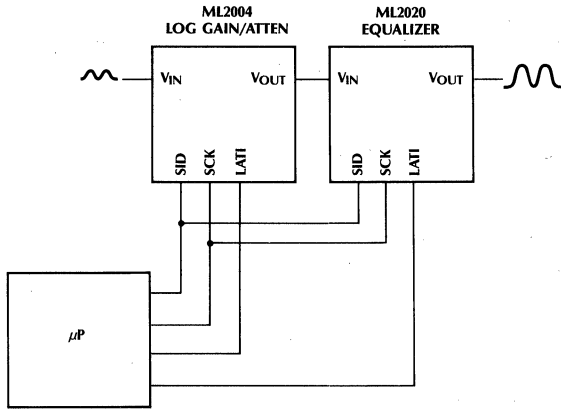


Figure 12. Typical Serial Interface

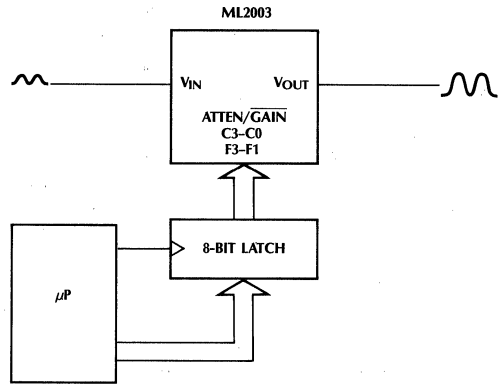


Figure 13. Typical μ P Parallel Interface

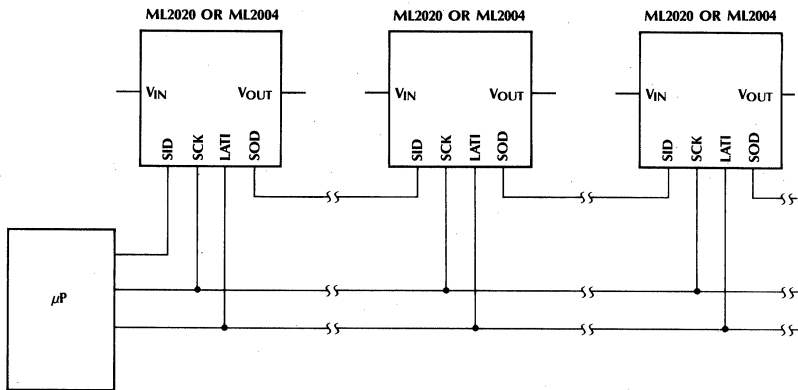


Figure 14. Controlling Multiple ML2020 and ML2004 With Only 3 Digital Lines Using One Long Data Word

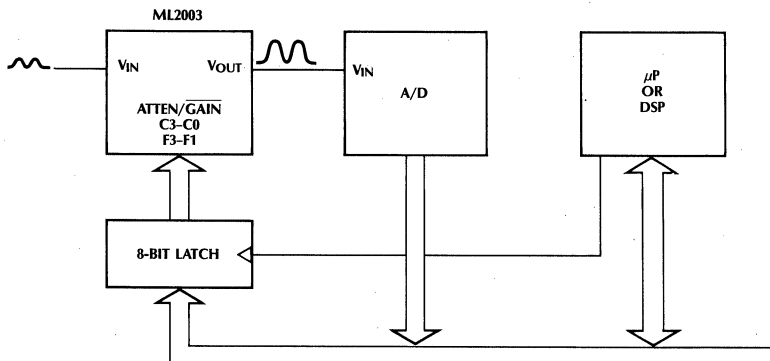


Figure 15. AGC For DSP Or Modem Front End

APPLICATIONS (Continued)

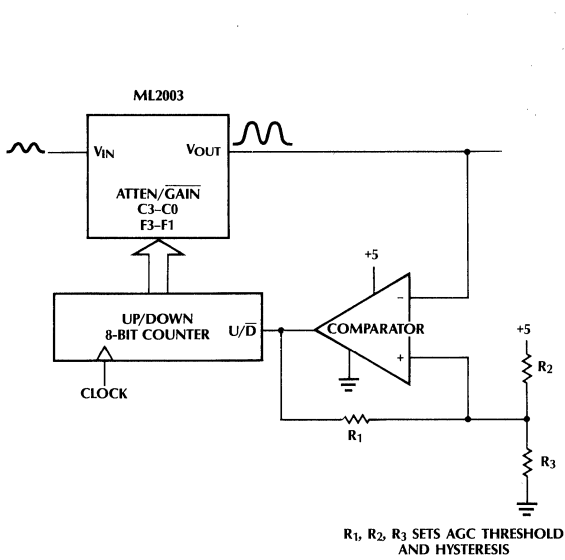


Figure 16. Analog AGC

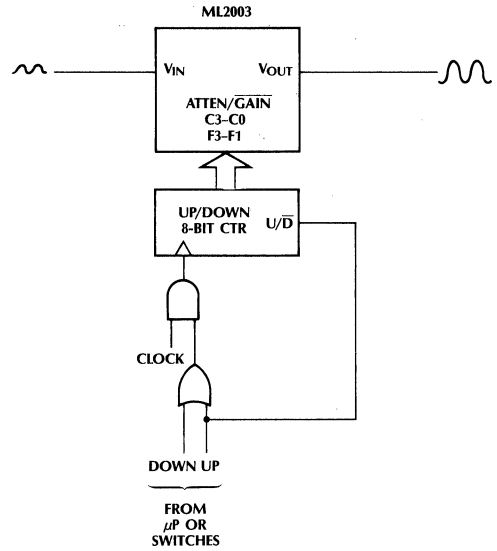


Figure 17. Digitally Controlled Volume Control

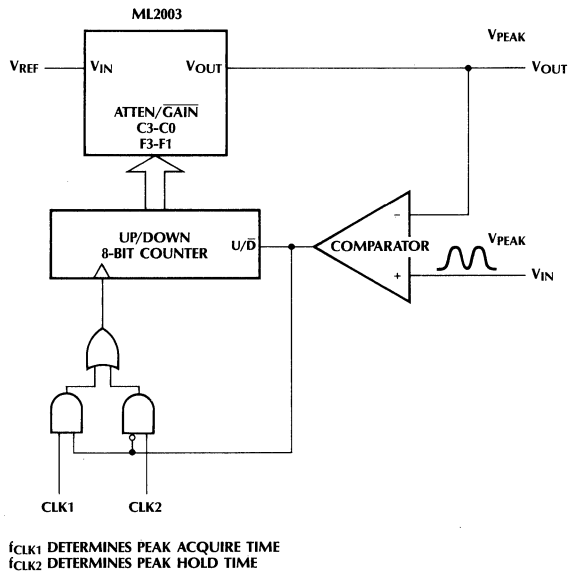


Figure 18. Precision Peak Detector ($\pm 1\%$) with Controllable Acquire and Hold Times

ML2003, ML2004

ORDERING INFORMATION

PART NUMBER	TEMP. RANGE	PACKAGE
ML2003IJ	-40°C to +85°C	HERMETIC DIP (J18)
ML2003IP	-40°C to +85°C	MOLDED DIP (P18)
ML2003IQ	-40°C to +85°C	MOLDED PCC (Q20)
ML2003CP	0°C to +70°C	MOLDED DIP (P18)
ML2003CQ	-40°C to +85°C	MOLDED PCC (Q20)
ML2004IJ	-40°C to +85°C	HERMETIC DIP (J14)
ML2004IP	-40°C to +85°C	MOLDED DIP (P14)
ML2004CP	0°C to +70°C	MOLDED DIP (P14)

μP Compatible Logarithmic Gain/Attenuator

GENERAL DESCRIPTION

The ML2008 and ML2009 are digitally controlled logarithmic gain/attenuators with a range of -24 to +24 dB in 0.1 dB steps.

Easy interface to microprocessors is provided by an input latch and control signals consisting of chip select and write.

The interface for gain setting of the ML2008 is by an 8-bit data word, while the ML2009 is designed to interface to a 16-bit data bus or with an 8-bit data bus with a single write operation by hard-wiring the gain/attenuation pin or LSB pin. The ML2008 can be power downed by the microprocessor utilizing a bit in the second write operation.

Absolute gain accuracy is 0.05 dB max over supply tolerance of ±10% and temperature range.

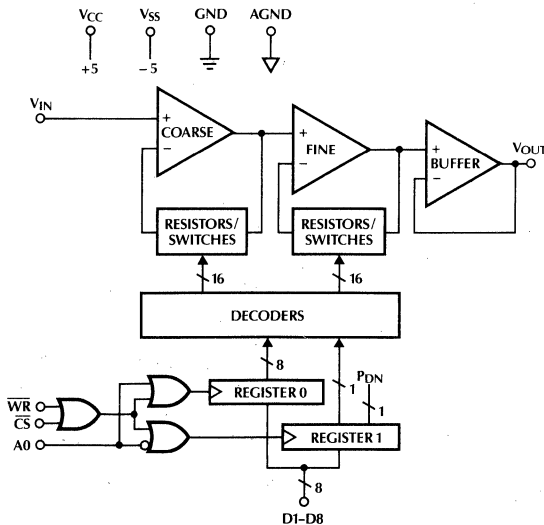
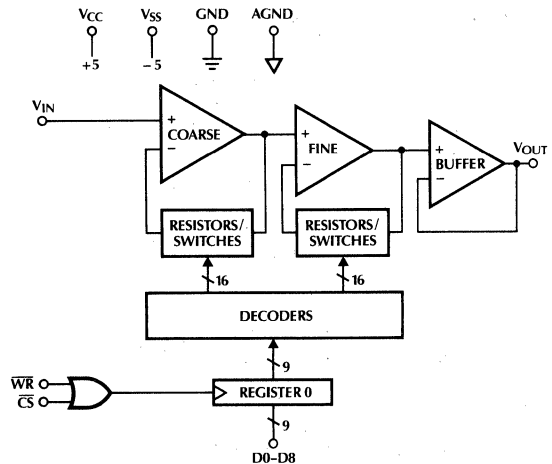
These CMOS logarithmic gain/attenuators are designed for a wide variety of applications in telecom, audio, sonar, or general purpose function generation.

FEATURES

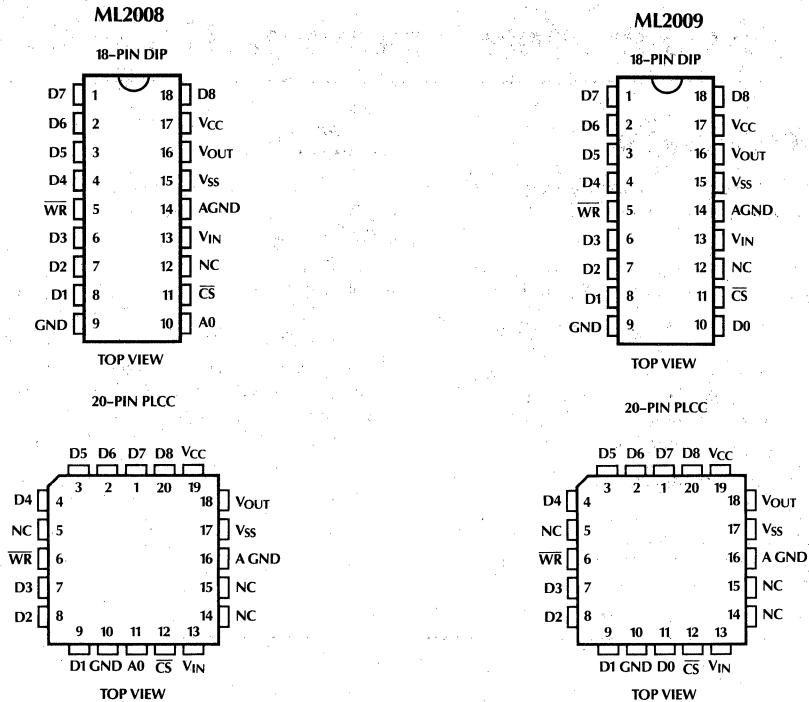
- Low noise 0dBrc max with +24dB gain
- Low harmonic distortion -60 dB max
- Gain range -24 to +24dB
- Resolution 0.1dB steps
- Flat frequency response ±0.05 dB from 0.3-4 kHz
±0.10dB from 0.1-20 kHz
- Low supply current 4mA max from ±5V supplies
- TTL/CMOS compatible digital interface
- ML2008 is designed to interface to an 8-bit data bus; ML2009 to 16-bit data bus.
- Standard 18-pin 0.3" center DIP or 20-pin molded chip carrier package

3

BLOCK DIAGRAMS

ML2008

ML2009


PIN CONNECTIONS



PIN DESCRIPTION

NAME	FUNCTION	NAME	FUNCTION
V_{SS}	Negative supply. -5 volts \pm 10%	D3	Data bit, F3
V_{CC}	Positive supply. 5 volts \pm 10%	D2	Data bit, P_{DN} , F2 ML2008; F2 ML2009
GND	Digital ground. 0 volts. All digital inputs are referenced to this ground.	D1	Data bit, F0, F1 ML2008; F1 ML2009
AGND	Analog ground. 0 volts. Analog input and output are referenced to this ground.	D0	Data bit, F0 ML2009 only
V_{IN}	Analog input	\overline{WR}	Write enable. This input latches the data bits into the registers on rising edges of \overline{WR} .
V_{OUT}	Analog output	\overline{CS}	Chip select. This input selects the device by only allowing the \overline{WR} signal to latch in data when \overline{CS} is low.
D8	Data bit, ATTN/ \overline{GAIN}	A0 (ML2008 only)	Address select. This input determines which data word is being written into the registers.
D7	Data bit, C3		
D6	Data bit, C2		
D5	Data bit, C1		
D4	Data bit, C0		

ABSOLUTE MAXIMUM RATINGS

(Note 1)

Supply Voltage	
V _{CC}	+6.5V
V _{SS}	-6.5V
AGND with Respect to GND	V _{CC} to V _{SS}
Analog Inputs and Outputs	V _{SS} -0.3V to V _{CC} +0.3V
Digital Inputs and Outputs	GND -0.3V to V _{CC} +0.3V
Input Current Per Pin	±25 mA
Power Dissipation	750 mW
Storage Temperature Range	-65°C to +150°C
Lead Temperature (Soldering 10 sec.)	300°C

OPERATING CONDITIONS

Temperature Range (Note 2)

ML2008CP, ML2009CP	0°C to +70°C
ML2008CQ, ML2009CQ	0°C to +70°C
ML2008IJ, ML2009IJ	-40°C to +85°C

Supply Voltage

V _{CC}	4V to 6V
V _{SS}	-4V to -6V

ELECTRICAL CHARACTERISTICS

Unless otherwise specified T_A = T_{MIN} to T_{MAX}, V_{CC} = 5V ± 10%, V_{SS} = -5V ± 10%, Data Word: D8 (ATTEN/ $\overline{\text{GAIN}}$) = 1, Other Bits = 0, (0dB Ideal Gain), C_L = 100pF, R_L = 600Ω, dBm measurements use 600Ω as reference load, digital timing measured at 1.4V.

SYMBOL	PARAMETER	NOTES	CONDITIONS	MIN	TYP NOTE 3	MAX	UNITS
ANALOG							
AG	Absolute Gain Accuracy	4	V _{IN} = 8dBm, 1kHz	-0.05		+0.05	dB
RG	Relative Gain Accuracy	4	100000001 000000000 000000001 All other gain settings All values referenced to 100000000 gain when D8 (ATTEN/ $\overline{\text{GAIN}}$) = 1, V _{IN} = 8dBm when D8 (ATTEN/ $\overline{\text{GAIN}}$) = 0, V _{IN} = (8dBm - Ideal Gain) in dB	-0.05 -0.05 -0.05 -0.1		+0.05 +0.05 +0.05 +0.1	dB dB dB dB
FR	Frequency Response	4	300-4000Hz 100-20,000Hz Relative to 1kHz	-0.05 -0.1		+0.05 +0.1	dB dB
V _{OS}	Output Offset Voltage	4	V _{IN} = 0, +24dB gain			±100	mV
I _{CN}	Idle Channel Noise	4 5	V _{IN} = 0, +24dB, C msg weighted V _{IN} = 0, +24dB, 1kHz		-6 450	0 900	dBrc nv/ $\sqrt{\text{Hz}}$
HD	Harmonic Distortion	4	V _{IN} = 8dBm, 1kHz Measure 2nd, 3rd, harmonic relative to fundamental			-60	dB
SD	Signal to Distortion	4	V _{IN} = 8dBm, 1kHz C msg weighted	+60			dB
PSRR	Power Supply Rejection	4	200mV _{p-p} , 1kHz sine, V _{IN} = 0 on V _{CC} on V _{SS}		-60 -60	-40 -40	dB dB
Z _{IN}	Input Impedance, V _{IN}	4		1			Meg
V _{INR}	Input Voltage Range	4		±3.0			V
V _{OSW}	Output Voltage Swing	4		±3.0			V

3

ML2008, ML2009

ELECTRICAL CHARACTERISTICS (Continued)

Unless otherwise specified $T_A = T_{MIN}$ to T_{MAX} , $V_{CC} = 5V \pm 10\%$, $V_{SS} = -5V \pm 10\%$, Data Word: D8 (ATTEN/GAIN) = 1, Other Bits = 0 (0dB Ideal Gain), $C_L = 100\text{ pF}$, $R_L = 600\Omega$, dBm measurements use 600Ω as reference load, digital timing measured at 1.4V, $C_L = 100\text{ pF}$.

SYMBOL	PARAMETER	NOTES	CONDITIONS	MIN	TYP NOTE 3	MAX	UNITS
DIGITAL AND DC							
V_{IL}	Digital Input Low Voltage	4				0.8	V
V_{IH}	Digital Input High Voltage	4		2.0			V
I_{IN}	Input Current, Low	4	$V_{IH} = \text{GND}$			-10	μA
I_{IN}	Input Current, High	4	$V_{IH} = V_{CC}$			10	μA
I_{CC}	V_{CC} Supply Current	4	No output load, $V_{IL} = \text{GND}$, $V_{IH} = V_{CC}$, $V_{IN} = 0$			4	mA
I_{SS}	V_{SS} Supply Current	4	No output load, $V_{IL} = \text{GND}$, $V_{IH} = V_{CC}$, $V_{IN} = 0$			-4	mA
I_{CCP}	V_{CC} Supply Current, ML2008 Powerdown Mode Only	4	No output load, $V_{IL} = \text{GND}$, $V_{IH} = V_{CC}$			0.5	mA
I_{SSP}	V_{SS} Supply Current, ML2008 Powerdown Mode Only	4	No output load, $V_{IL} = \text{GND}$, $V_{IH} = V_{CC}$			-0.1	mA
AC CHARACTERISTICS							
t_{SET}	V_{OUT} Settling Time	4	$V_{IN} = 0.185\text{ V}$. Change gain from -24 to $+24\text{ dB}$. Measure from \overline{WR} rising edge to when V_{OUT} settles to within 0.05 dB of final value.			20	μs
t_{STEP}	V_{OUT} Step Response	4	Gain = $+24\text{ dB}$, $V_{IN} = -3\text{ V}$ to $+3\text{ V}$ step. Measure from $V_{IN} = -3\text{ V}$ to when V_{OUT} settles to within 0.05 dB of final value.			20	μs
t_{DS}	Data Setup Time	4		50			ns
t_{DH}	Data Hold Time	4		50			ns
t_{AS}	A0 Setup Time	4		0			ns
t_{AH}	A0 Hold Time	4		0			ns
t_{CSS}	\overline{CS}^* Setup Time	4		0			ns
t_{CSH}	\overline{CS}^* Hold Time	4		0			ns
t_{PW}	\overline{WR}^* Pulse Width	4		50			ns

Note 1: Absolute maximum ratings are limits beyond which the life of the integrated circuit may be impaired. All voltages unless otherwise specified are measured with respect to ground.

Note 2: 0°C to $+70^\circ\text{C}$ and -40°C to $+85^\circ\text{C}$ operating temperature range devices are 100% tested with temperature limits guaranteed by 100% testing, sampling, or by correlation with worst-case test conditions.

Note 3: Typicals are parametric norm at 25°C .

Note 4: Parameter guaranteed and 100% production tested.

Note 5: Parameter guaranteed. Parameters not 100% tested are not in outgoing quality level calculation.

TIMING DIAGRAM

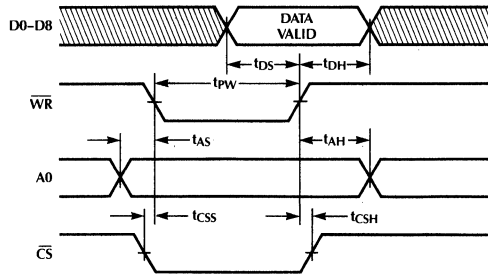


Figure 1. Timing Diagram

TYPICAL PERFORMANCE CURVES

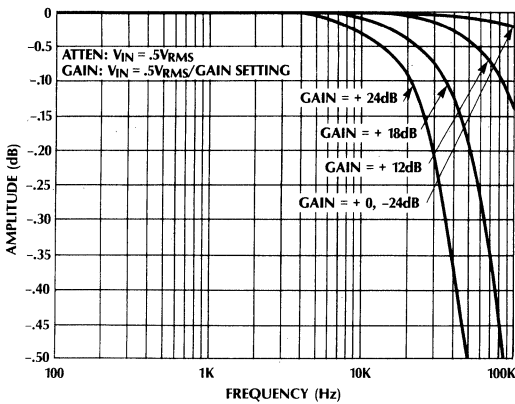


Figure 2. Amplitude vs Frequency ($V_{IN}/V_{OUT} = 0.5V_{RMS}$)

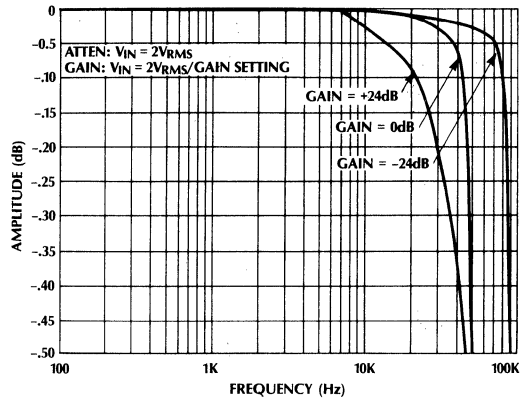


Figure 3. Amplitude vs Frequency ($V_{IN}/V_{OUT} = 2V_{RMS}$)

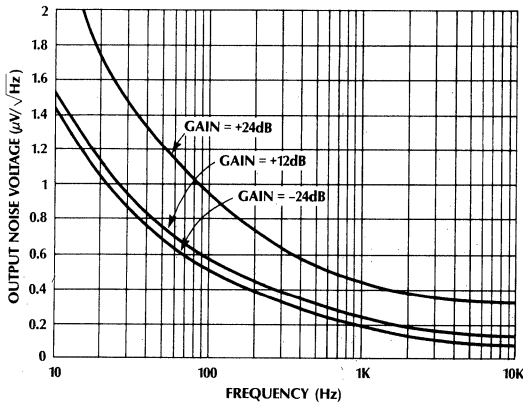


Figure 4. Output Noise Voltage vs Frequency

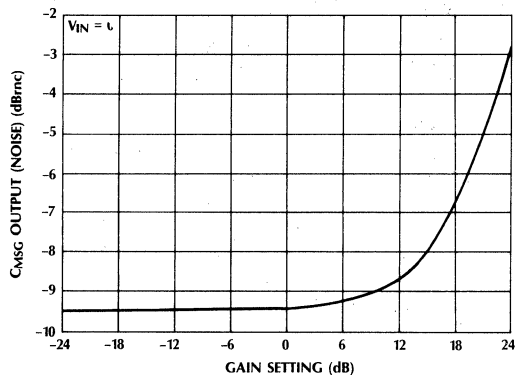


Figure 5. C_{MSC} Output Noise vs Gain Setting

TYPICAL PERFORMANCE CURVES (Continued)

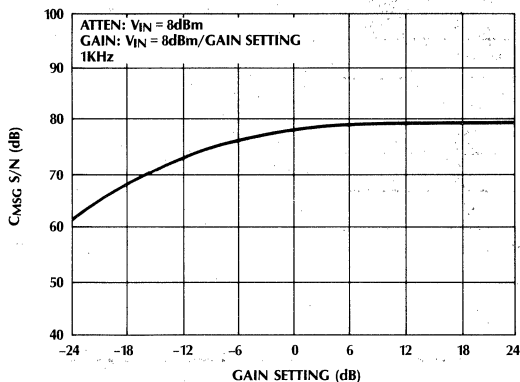


Figure 6. CMSG S/N vs Gain Setting

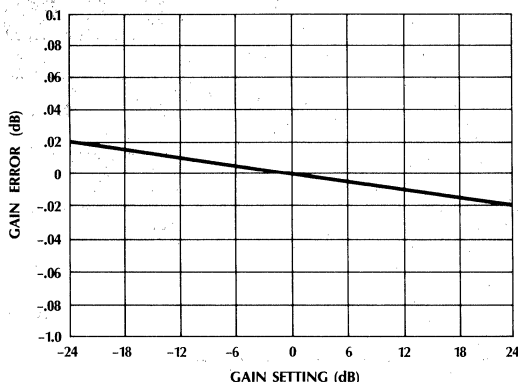


Figure 7. Gain Error vs Gain Setting

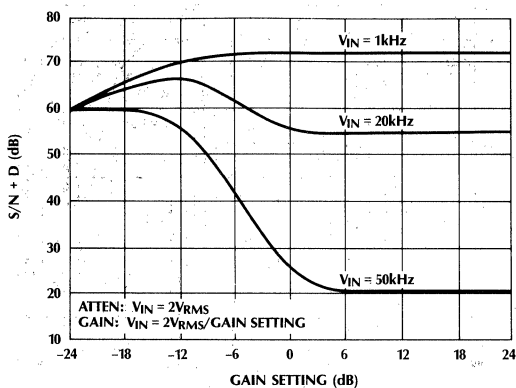


Figure 8. S/N + D vs Gain Setting ($V_{IN}/V_{OUT} = 2V_{RMS}$)

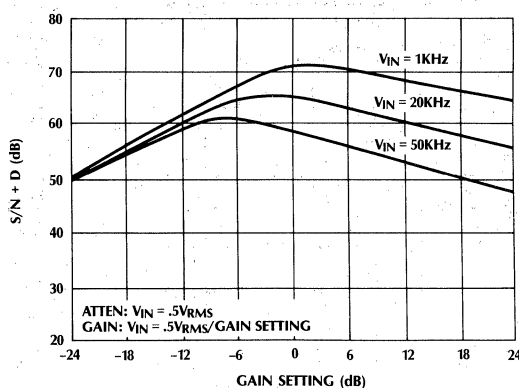


Figure 9. S/N + D vs Gain Setting ($V_{IN}/V_{OUT} = 0.5V_{RMS}$)

1.0 FUNCTIONAL DESCRIPTION

The ML2008, ML2009 consists of a coarse gain stage, a fine gain stage, an output buffer, and a μP compatible parallel digital interface.

1.1 Gain Stages

The analog input, V_{IN} , goes directly into the op amp input in the coarse gain stage. The coarse gain stage has a gain range of 0 to 22.5 dB in 1.5 dB steps.

The fine gain stage is cascaded onto the coarse section. The fine gain stage has a gain range of 0 to 1.5 dB in 0.1 dB steps.

Both stages can be programmed for either gain or attenuation, thus doubling the effective gain range.

The logarithmic steps in each gains stage are generated by placing the input signal across a resistor string of 16 series resistors. Analog switches allow the voltage to be tapped from the resistor string at 16 points. The resistors are sized such that each output voltage is at the proper logarithmic ratio relative to the input signal at the top of the string. Attenuation is implemented by using the resistor string as a simple voltage divider, and gain is implemented by using the resistor string as a feedback resistor around an internal op amp.

1.2 Gain Settings

Since the coarse and fine gain stages are cascaded, their gains can be summed logarithmically. Thus, any gain from -24 dB to $+24$ dB in 0.1 dB steps can be obtained by combining

the coarse and fine gain settings to yield the desired gain setting. The relationship between the register 0 and 1 bits and the corresponding analog gain values is shown in Tables 1 and 2. Note that C3-C0 select the coarse gain, F3-F0 select the fine gain, and ATTEN/GAIN selects either gain or attenuation.

1.3 Output Buffer

The final analog stage is the output buffer. This amplifier has internal gain of 1 and is designed to drive 600Ω, 100 pF loads. Thus, it is suitable for driving a telephone hybrid circuit directly without any external amplifier.

Table 1. Fine Gain Settings (C3 – C0 = 0)

F3	F2	F1	F0	Ideal Gain (dB)	
				ATTEN/GAIN=1	ATTEN/GAIN=0
0	0	0	0	0.0	0.0
0	0	0	1	-0.1	0.1
0	0	1	0	-0.2	0.2
0	0	1	1	-0.3	0.3
0	1	0	0	-0.4	0.4
0	1	0	1	-0.5	0.5
0	1	1	0	-0.6	0.6
0	1	1	1	-0.7	0.7
1	0	0	0	-0.8	0.8
1	0	0	1	-0.9	0.9
1	0	1	0	-1.0	1.0
1	0	1	1	-1.1	1.1
1	1	0	0	-1.2	1.2
1	1	0	1	-1.3	1.3
1	1	1	0	-1.4	1.4
1	1	1	1	-1.5	1.5

2.0 DIGITAL INTERFACE

The architecture of the digital section is shown in the preceding block diagram.

The structure of the data registers or latches is shown in Figures 10 and 11 for the ML2008 and ML2009, respectively. The registers control the attenuation/gain setting bits and with the ML2008 the power down bit.

Tables 1 and 2 describe how the data word programs the gain.

The difference between the ML2008 and ML2009 is in the register structure. The ML2008 is a 8-bit data bus version. This device has one 8-bit register and one 2-bit register to store the 9 gain setting bits and 1 powerdown bit. Two write operations are necessary to program the full 10 data bits from eight external data pins. The address pin A0 controls which register is being written into. The powerdown bit, PDN, causes the device to be placed in powerdown. When PDN=0, the device is powered down. In this state, the power consumption is reduced by removing power from the analog section and

1.4 Power Supplies

The digital section is powered between V_{CC} and GND, or 5V. The analog section is powered between V_{CC} and V_{SS} and uses AGND as the reference point, or $\pm 5V$.

GND and AGND are totally isolated inside the device to minimize coupling from the digital section into the analog section. Typically this is less than 100μV. However, AGND and GND should be tied together physically near the device and ideally close to the common power supply ground connection.

Typically, the power supply rejection of V_{CC} and V_{SS} to the analog output is greater than -60 dB at 1 kHz. If decoupling of the power supplies is still necessary in a system, V_{CC} and V_{SS} should be decoupled with respect to AGND.

Table 2. Coarse Gain Settings (F3 – F0 = 0)

C3	C2	C1	C0	Ideal Gain (dB)	
				ATTEN/GAIN=1	ATTEN/GAIN=0
0	0	0	0	0.0	0.0
0	0	0	1	-1.5	1.5
0	0	1	0	-3.0	3.0
0	0	1	1	-4.5	4.5
0	1	0	0	-6.0	6.0
0	1	0	1	-7.5	7.5
0	1	1	0	-9.0	9.0
0	1	1	1	-10.5	10.5
1	0	0	0	-12.0	12.0
1	0	0	1	-13.5	13.5
1	0	1	0	-15.0	15.0
1	0	1	1	-16.5	16.5
1	1	0	0	-18.0	18.0
1	1	0	1	-19.5	19.5
1	1	1	0	-21.0	21.0
1	1	1	1	-22.5	22.5

forcing the analog output, V_{OUT} , to a high impedance state. While the device is in powerdown, the digital section is still functional and the current data word remains stored in the registers. When PDN=0, device is in normal operation.

The ML2009 is a 9-bit data bus version. This device has one 9-bit register to store the 9 gain setting bits. The full 9 data bits can be programmed with one write operation from nine external data pins.

The internal registers or latches are edge triggered. The data is transferred from the external pins to the register output on the rising edge of \overline{WR} . The address pin, A0, controls which register the data will be written into as shown in Figures 1 and 2. The \overline{CS} control signal selects the device by allowing the \overline{WR} signal to latch in the data only when \overline{CS} is low. When \overline{CS} is high, \overline{WR} is inhibited from latching in new data into the registers.

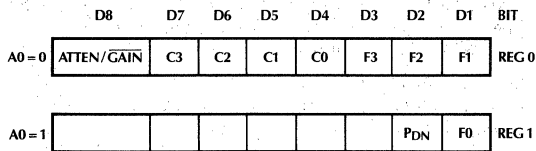


Figure 10. ML2008 Register Structure

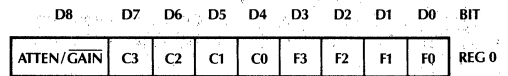


Figure 11. ML2009 Register Structure

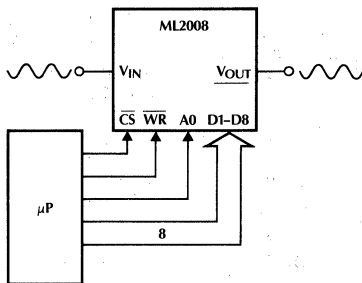


Figure 12. Typical 8-Bit μ P Interface, Double Write

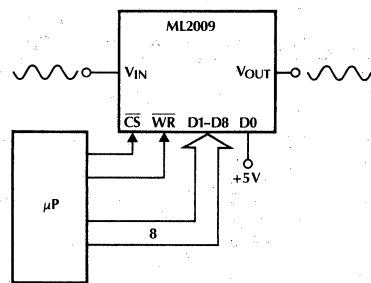


Figure 13. Typical 8-Bit μ P Interface, Single Write

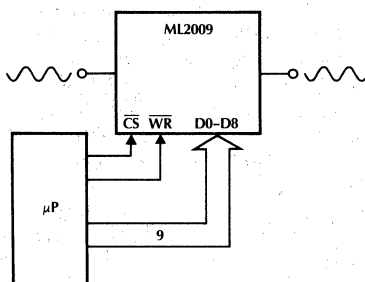


Figure 14. Typical 16-Bit μ P Interface

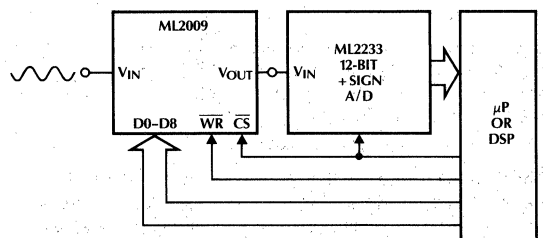


Figure 15. AGC for DSP or Modem Front End

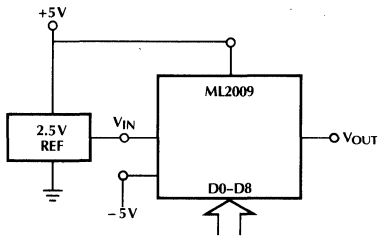


Figure 16. Operation as Logarithmic D/A Converter

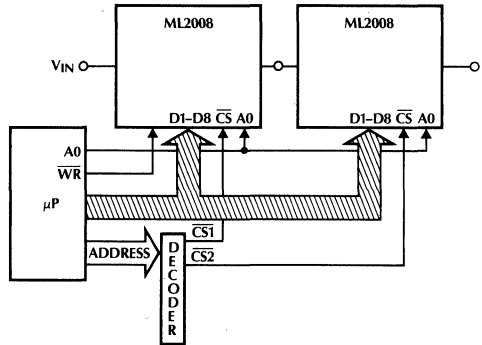


Figure 17. Controlling Multiple Gain/Attenuators

ORDERING INFORMATION

PART NUMBER	TEMP. RANGE	PACKAGE
ML2008IJ	-40°C to +85°C	HERMETIC DIP (J18)
ML2008IP	-40°C to +85°C	MOLDED DIP (P18)
ML2008IQ	-40°C to +85°C	MOLDED PCC (Q20)
ML2008CP	0°C to +70°C	MOLDED DIP (P18)
ML2008CQ	0°C to +70°C	MOLDED PCC (Q20)
ML2009IJ	-40°C to +85°C	HERMETIC DIP (J18)
ML2009IP	-40°C to +85°C	MOLDED DIP (P18)
ML2009IQ	-40°C to +85°C	MOLDED PCC (Q20)
ML2009CP	0°C to +70°C	MOLDED DIP (P18)
ML2009CQ	0°C to +70°C	MOLDED PCC (Q20)

Telephone Line Equalizer

GENERAL DESCRIPTION

The ML2020 is a monolithic analog line equalizer for telephone applications. The ML2020 consists of a switched capacitor filter that realizes a family of frequency response curves optimized for telephone line equalization.

The ML2020 consists of a continuous anti-aliasing filter, a 60 Hz rejection highpass filter section, three programmable switched capacitor equalization filters, an output smoothing filter, a 600Ω driver, and a digital section for the serial interface.

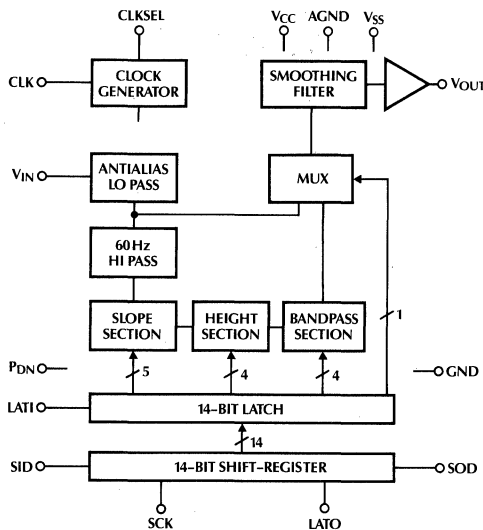
The equalization filters adjust the slope, height, and bandwidth of the frequency response. The desired frequency response is programmed by a digital 14-bit serial input data stream.

The ML2020 is implemented in a double polysilicon CMOS technology.

FEATURES

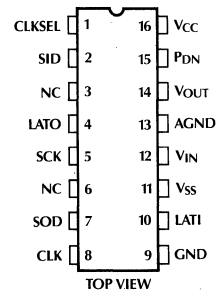
- Slope, height, and bandwidth adjustable
- 60 Hz rejection filter
- On chip anti-alias filter
- Bypass mode
- Low supply current 6 mA typical from ±5V supplies
- TTL/CMOS compatible interface
- Double buffered data latch
- Selectable master clock 1.544 or 1.536 MHz
- Synchronous or asynchronous data loading capability
- Compatible with ML2003 and ML2004 logarithmic gain/attenuator
- Standard 16-pin 0.3" center molded or hermetic dip and 18-pin SOIC
- 0°C to +70°C and -40°C to +85°C operating temperature range

BLOCK DIAGRAM

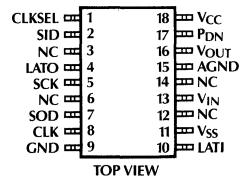


PIN CONNECTIONS

ML2020
16-PIN DIP



ML2020
18-PIN SOIC



PIN DESCRIPTION

NAME	FUNCTION	NAME	FUNCTION
CLKSEL	Clock select input. This pin selects the frequency of the CLK input. If CLK is 1.536MHz, set CLKSEL = 1. If CLK is 1.544MHz, set CLKSEL = 0. Pin has an internal pullup resistor to V _{CC} .	GND	Digital ground. 0volts. All digital inputs and output are referenced to this ground.
SID	Serial input data. Digital input that contains serial data word which controls the filter frequency response setting.	LATI	Input latch clock. Digital input which loads data from the shift register into the latch.
LATO	Output latch clock. Digital input which loads the data word back into the shift register from the latch.	V _{SS}	Negative supply. -5volts ±10%.
SCK	Shift clock. Digital input which shifts the serial data on SID into the shift register on rising edges and out onto SOD on falling edges.	V _{IN}	Analog input.
SOD	Serial output data. Digital output of the shift register.	AGND	Analog ground. 0volts. Analog input and output are referenced to this ground.
CLK	Master clock input. Digital input which generates clocks for the switched capacitor filters. Frequency can be either 1.544MHz or 1.536MHz.	V _{OUT}	Analog output.
		P _{DN}	Powerdown input. When P _{DN} = 1, device is in powerdown mode. When P _{DN} = 0, device is in normal operation. This pin has an internal pulldown resistor to GND.
		V _{CC}	Positive supply. 5volts ±10%

ABSOLUTE MAXIMUM RATINGS

(Note 1)

Supply Voltage	
V _{CC}	+6.5V
V _{SS}	-6.5V
AGND with Respect to GND	±0.5V
Analog Input and Output	V _{SS} - 0.3V to V _{CC} + 0.3V
Digital Input and Outputs	GND - 0.3V to V _{CC} + 0.3V
Input Current Per Pin	±25mA
Power Dissipation	750mW
Storage Temperature Range	-65°C to +150°C
Lead Temperature (Soldering 10 sec.)	300°C

OPERATING CONDITIONS

Temperature Range (Note 2)

ML2020CP, ML2020CS	0°C to +70°C
ML2020IJ	-40°C to +85°C
Supply Voltage	
V _{CC}	4V to 6V
V _{SS}	-4V to -6V

ELECTRICAL CHARACTERISTICS

Unless otherwise specified $T_A = T_{MIN}$ to T_{MAX} , $V_{CC} = 5V \pm 10\%$, $V_{SS} = -5V \pm 10\%$, Data Word: $\overline{BP} = 1$, Other Bits = 0, $C_L = 100\text{pF}$, $R_L = 600\Omega$, dBm measurements use 600Ω as reference load, $V_{IN} = -7\text{dBm}$, 1kHz sinusoid CLK = $1.544\text{MHz} \pm 300\text{Hz}$ and digital time measured at 1.4V

SYMBOL	PARAMETER	NOTES	CONDITIONS	MIN	TYP NOTE 3	MAX	UNITS							
ANALOG														
SR	Response, Slope Section	4	1kHz response											
			NL/L	S3	S2	S1	S0							
			0	0	0	0	1	1.4 ± 0.1	dB					
			0	0	0	1	0	2.6 ± 0.2	dB					
			0	0	1	0	0	4.7 ± 0.2	dB					
			0	1	0	0	0	7.8 ± 0.2	dB					
			0	1	1	1	1	11.4 ± 0.25	dB					
			1	0	0	0	0	0 ± 0.1	dB					
			1	0	0	1	0	0.4 ± 0.1	dB					
			1	0	1	0	0	0.9 ± 0.2	dB					
			1	1	0	0	0	1.8 ± 0.2	dB					
			1	1	1	1	1	3.7 ± 0.2	dB					
	Referenced to					6.6 ± 0.25	dB							
	0	0	0	0	0									
HR	Response, Height Section	4	3250Hz response referenced to 1kHz response with $\overline{BP} = 1$, other bits = 0											
			NL/L	H3	H2	H1	H0							
			0	0	0	0	0	0 ± 0.1	dB					
			0	0	0	0	1	0.6 ± 0.2	dB					
			0	0	0	1	0	1.2 ± 0.2	dB					
			0	0	1	0	0	2.4 ± 0.2	dB					
			0	1	0	0	0	5.8 ± 0.3	dB					
0	1	1	1	1	11.2 ± 0.3	dB								
BR	Response, Bandwidth Section (Q)	4	NL/L	B3	B2	B1	B0	H3	H2	H1	H0			
			0	0	0	0	0	1	1	1	1	1	16.1 ± 2.0	
			0	0	0	0	1	1	1	1	1	1	14.2 ± 1.5	
			0	0	0	1	0	1	1	1	1	1	12.6 ± 1.5	
			0	0	1	0	0	1	1	1	1	1	9.1 ± 1.0	
			0	1	0	0	0	1	1	1	1	1	3.6 ± 0.5	
			0	1	1	1	1	1	1	1	1	1	1.2 ± 0.35	
PK	BW Peak Frequency	4	H3 thru H0 = 1	3230	3250	3270	Hz							
AG	Absolute Gain, Flat Response	4	1 to 4kHz	-0.1	+0.1	+0.3	dB							
AGB	Absolute Gain, Bypass Mode	4	0.3 to 4kHz, $\overline{BP} = 0$	-0.1	+0.1	+0.3	dB							
ICN	Idle Channel Noise	4	$V_{IN} = 0$		3	8	dBrc							
			$V_{IN} = 0$, All Data Bits = 1		9		dBrc							

ELECTRICAL CHARACTERISTICS (Continued)

Unless otherwise specified $T_A = T_{MIN}$ to T_{MAX} , $V_{CC} = 5V \pm 10\%$, $V_{SS} = -5V \pm 10\%$, Data Word: $\overline{BP} = 1$, Other Bits = 0, $C_L = 100$ pF, $R_L = 600\Omega$, dBm measurements use 600Ω as reference load, $V_{IN} = -7$ dBm, 1kHz sinusoid CLK = 1.544MHz ± 300 Hz and digital time measured at 1.4V

SYMBOL	PARAMETER	NOTES	CONDITIONS	MIN	TYP NOTE 4	MAX	LIMIT UNITS
ANALOG							
HD	Harmonic Distortion	4	$V_{IN} = 5$ dBm, 1kHz Measure 2nd, 3rd, harmonic relative to fundamental			-48	dB
SD	Signal to Distortion	4	$V_{IN} = -12$ dBm, 1kHz C msg weighted	+48			dB
SFN	Single Frequency Noise	5	$V_{IN} = 0$, 4kHz \leq frequency \leq 150kHz			-50	dBm
PSRR	Power Supply Rejection	4	200mV _{p-p} , 1kHz sine, $V_{IN} = 0$ on V_{CC} on V_{SS}			-40 -40	dB dB
Z_{IN}	Input Impedance, V_{IN}	4		100			k Ω
V_{OS}	Output Offset Voltage	4	$V_{IN} = 0$			± 50	mV
V_{INR}	Input Voltage Range	4		± 2.0			V
V_{OSW}	Output Voltage Swing	4	$R_L = 600\Omega$	± 2.0			V
DIGITAL AND DC							
V_{IL}	Digital Input Low Voltage	4				0.8	V
V_{IH}	Digital Input High Voltage	4		2.0			V
V_{OL}	Digital Output Low Voltage	4	$I_{OL} = 2$ mA			0.4	V
V_{OH}	Digital Output High Voltage	4	$I_{OH} = -1$ mA	4.0			V
I_{LCLK}	Input Current, CLK SEL	4	$V_{IN} = 0$	5		100	μ A
I_{LPDN}	Input Current, PDN	4	$V_{IN} = V_{CC}$	-5		-100	μ A
I_L	Input Current, All Other Inputs	4	$V_{IN} = 0 - V_{CC}$			± 10	μ A
I_{CC}	V_{CC} Supply Current	4	No output load, $V_{IL} = GND$, $V_{IH} = V_{CC}$, $V_{IN} = 0$			10	mA
I_{SS}	V_{SS} Supply Current	4	No output load, $V_{IL} = GND$, $V_{IH} = V_{CC}$, $V_{IN} = 0$			-10	mA
I_{CCP}	V_{CC} Supply Current, Powerdown Mode	4	No output load, $V_{IL} = GND$, $V_{IH} = V_{CC}$			1.2	mA
I_{SSP}	V_{SS} Supply Current, Powerdown Mode	4	No output load, $V_{IL} = GND$, $V_{IH} = V_{CC}$			-1.2	mA
AC CHARACTERISTICS							
t_{DC}	Clock Duty Cycle	5		40		60	%
t_{SCK}	SCK On/Off Period	4		250			ns
t_S	SID Data Setup Time	4		50			ns
t_H	SID Data Hold Time	4		50			ns
t_D	SOD Data Delay	4		0		125	ns
t_{IPW}	LATI Pulse Width	4		50			ns
t_{OPW}	LATO Pulse Width	4		50			ns
t_{IS}, t_{OS}	LATI, LATO Setup Time	4		50			ns
t_{IH}, t_{OH}	LATI, LATO Hold Time	5		50			ns
t_{PLD}	SOD Parallel Load Delay	4		0		125	ns

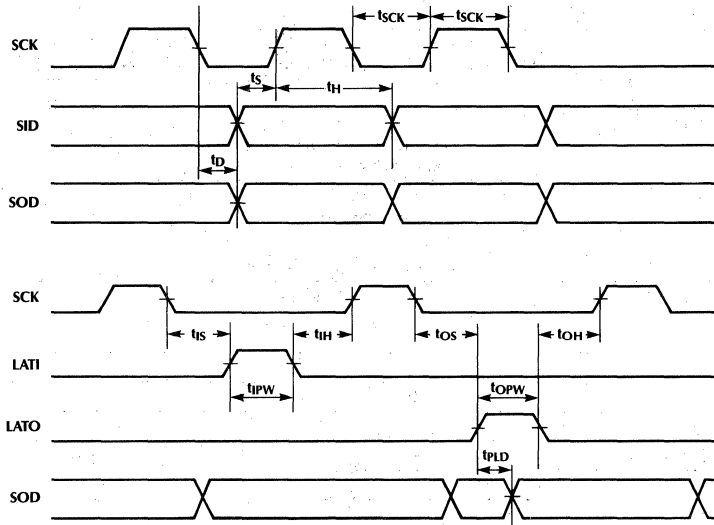
Note 1: Absolute maximum ratings are limits beyond which the life of the integrated circuit may be impaired. All voltages unless otherwise specified are measured with respect to ground.

Note 2: 0°C to +70°C and -40°C to +85°C operating temperature range devices are 100% tested with temperature limits guaranteed by 100% testing, sampling, or by correlation with worst-case test conditions.

Note 3: Typicals are parametric norm at 25°C.

Note 4: Parameter guaranteed and 100% production tested.

Note 5: Parameter guaranteed. Parameters not 100% tested are not in outgoing quality level calculation.



TIMING PARAMETERS ARE REFERENCED TO THE 1.4 VOLT MIDPOINT.

Figure 1. Serial Timing Diagram

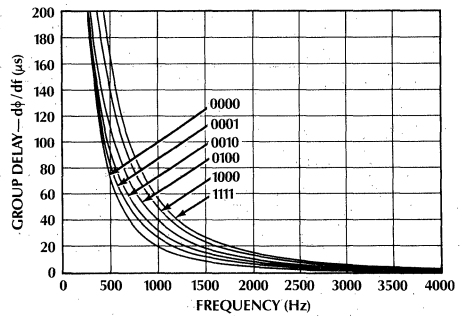
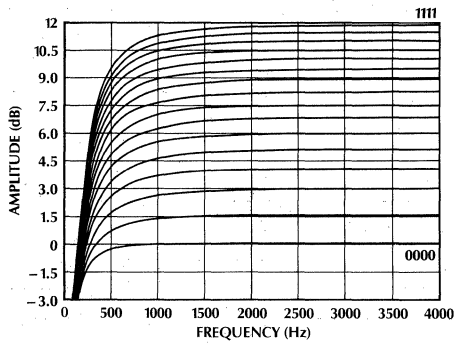


Figure 2. Typical Slope Filter Response — NL/L = 0
B3-B0, H3-H0 = 0000, S3-S0 = 0000 to 1111.

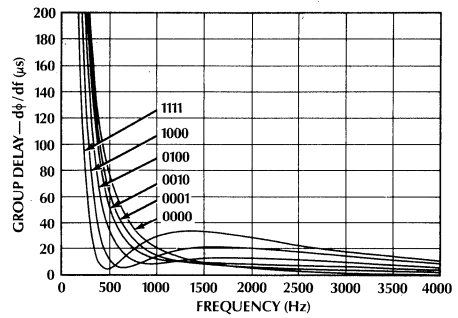
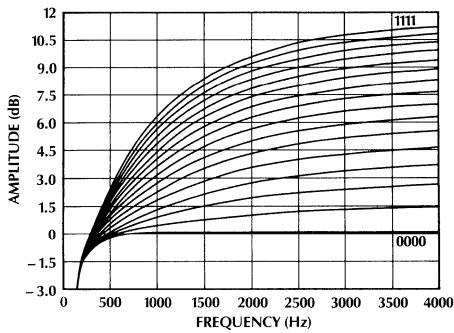


Figure 3. Typical Slope Filter Response — NL/L = 1
B3-B0, H3-H0 = 0000, S3-S0 = 0000 to 1111.

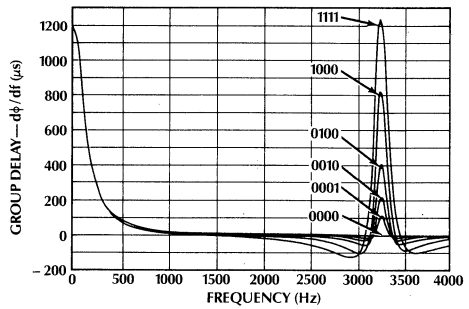
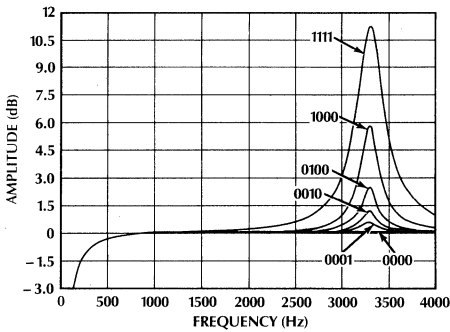


Figure 4. Typical Height Filter Response — NL/L = 0
B3-B0, S3-S0 = 0000; H3-H0 = 0000 to 1111.

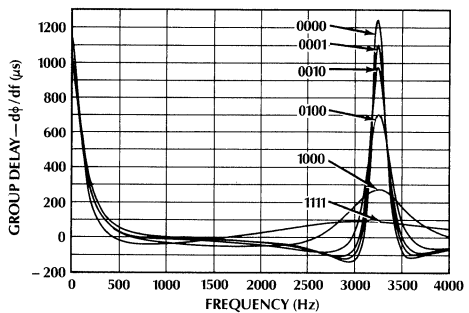
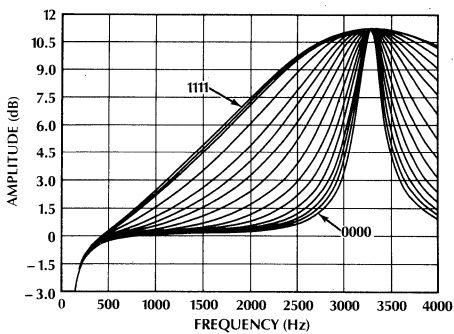


Figure 5. Typical Bandwidth Filter Response — NL/L = 0
H3-H0 = 1111; S3-S0 = 0000; B3-B0 = 0000 to 1111.

1.0 FUNCTIONAL DESCRIPTION

The ML2020 consists of a continuous anti-alias filter, a 60Hz reject highpass filter section, three programmable switched capacitor equalization filters, an output smoothing filter, an output driver, and a digital section for the serial interface.

1.1 Anti-Alias Filter

The first section is a continuous anti-alias filter. This filter is needed to prevent aliasing of high frequency signals present on the input into the passband by the sampling action of the switched capacitor filters. This section is a continuous second order lowpass filter with a typical 3dB frequency at 20kHz and 30dB of rejection at 124kHz.

1.2 60Hz Rejection Filter

The 60Hz section is a highpass switched capacitor filter designed to reject DC offsets and low frequency signals present on the input. This filter is a first order section with a typical 3dB frequency at 135Hz.

1.3 Equalization Filters

The equalizer filters follow the 60Hz highpass section. These programmable filters implement a family of frequency response curves intended to compensate for the response of telephone lines.

This filter is composed of three distinct sections: slope, height, and bandwidth.

1.3.1 Response of Slope, Height, and Bandwidth

The family of response curves generated by the slope section are shown in Figures 2 and 3. There are 4 slope select bits, S3-S0. These bits alter the slope of the highpass response under 1000Hz, and as a result, the absolute gain above 1000Hz will be unique for each setting. Table 1 gives typical 1kHz gain values for all slope settings.

Table 1. Typ. 1kHz Gain for Slope Settings

Slope Setting	Rel 1kHz Gain (dB)	
	NL/L = 1	NL/L = 0
0	0.0	Rel
1	0.4	1.4
2	0.9	2.6
3	1.4	3.7
4	1.8	4.7
5	2.3	5.5
6	2.8	6.3
7	3.4	7.2
8	3.7	7.8
9	4.2	8.4
10	4.6	9.0
11	5.0	9.5
12	5.4	10.0
13	5.8	10.5
14	6.2	11.0
15	6.6	11.4

HT, BW Bits = 0

There is an additional bit, NL/L, that also affects the highpass response of the slope filter. The slope response curves in Figure 2 are with NL/L = 0. These same response curves are shown in Figure 3 with NL/L = 1. Notice that the NL/L bit adds more droop in the highpass response below 2500Hz.

The family of response curves generated by the height section are shown in Figure 4. There are 4 height select bits, H3-H0. This section creates a peak in the response at 3250Hz and this filter controls the amount of peaking. Table 2 gives typical 1kHz gain values for all height and bandwidth settings.

Table 2. Typ. 1kHz Gain for HT and BW Settings

		Relative 1kHz Gain (dB)															
		HT Setting															
		0	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15
BW Setting	0	Rel	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0
	1	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0
	2	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0
	3	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0
	4	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0.1
	5	0	0	0	0	0	0	0	0	0	0	0	0	0	0.1	0.1	0.1
	6	0	0	0	0	0	0	0	0	0	0	0.1	0.1	0.1	0.1	0.1	0.1
	7	0	0	0	0	0	0	0	0	0.1	0.1	0.1	0.1	0.1	0.1	0.2	0.2
	8	0	0	0	0	0	0	0.1	0.1	0.1	0.1	0.1	0.1	0.2	0.2	0.3	0.3
	9	0	0	0	0	0	0.1	0.1	0.1	0.1	0.1	0.2	0.2	0.3	0.3	0.4	0.5
	10	0	0	0	0	0.1	0.1	0.1	0.1	0.1	0.2	0.3	0.3	0.4	0.5	0.6	0.7
	11	0	0	0	0.1	0.1	0.1	0.2	0.2	0.3	0.4	0.4	0.5	0.7	0.8	0.9	1.1
	12	0	0	0.1	0.1	0.1	0.2	0.2	0.3	0.4	0.5	0.7	0.8	1.0	1.1	1.4	1.6
	13	0	0	0.1	0.1	0.2	0.3	0.4	0.5	0.6	0.8	0.9	1.1	1.4	1.6	1.9	2.3
	14	0	0	0.1	0.1	0.2	0.3	0.4	0.5	0.7	0.8	1.0	1.2	1.5	1.7	2.0	2.4
	15	0	0	0.1	0.1	0.2	0.3	0.4	0.5	0.7	0.9	1.1	1.3	1.6	1.8	2.1	2.5

Slope Bits = 0

The family of response curves generated by the bandwidth section is shown in Figure 5. There are 4 bandwidth select bits, B3-B0. This section causes the response of the 3250Hz peak to be widened, and as a result, this filter controls the bandwidth of the 3250Hz peaked region.

1.3.2 Transfer Function

The transfer function for the ML2020 is shown below. This transfer function is valid for magnitude response only. The actual magnitude response from an individual device may deviate from the computed response from the transfer function by typically 0-0.2dB.

$$H(s) = \frac{-s}{s+a} \times \frac{c(s+b)}{b(s+c)} \times \frac{[s^2 + h(\omega_0/Q)s + \omega_0^2]}{[s^2 + (\omega_0/Q)s + \omega_0^2]} \times \frac{[\sin(\pi f / fc)]}{(\pi / fc)}$$

$$s = j \times 256000 \times \tan(\pi f / 128000)$$

$$a = 848.230$$

$$\omega_0 = 20463.77$$

$$fc = 128000$$

b,c : See Table 3. (slope)
 Q : See Table 4. (bandwidth)
 h : See Table 5. (height)

Table 3. Slope Response Factors (b, c)

S3-0	b	b
	NL/L = 0	NL/L = 1
0000	2.371759E+03	1.116280E+04
0001	1.985920E+03	9.345141E+03
0010	1.701779E+03	8.007156E+03
0011	1.493571E+03	7.026999E+03
0100	1.326721E+03	6.241681E+03
0101	1.196668E+03	5.629636E+03
0110	1.087277E+03	5.114881E+03
0111	9.983588E+02	4.696487E+03
1000	9.179889E+02	4.318339E+03
1001	8.537864E+02	4.016273E+03
1010	7.966049E+02	3.747249E+03
1011	7.478074E+02	3.517676E+03
1100	7.035099E+02	3.309279E+03
1101	6.651771E+02	3.128945E+03
1110	6.299477E+02	2.963214E+03
1111	5.990361E+02	2.817797E+03

S3-0	c	c
	NL/L = 0	NL/L = 1
XXXX	2.371759E+03	1.116280E+04

Table 4. Slope Response Factors (b, c)

B3-0	Q
0000	17.444906
0001	15.386148
0010	13.652451
0011	11.593677
0100	9.859960
0101	8.017864
0110	6.392453
0111	5.092080
1000	3.900003
1001	3.141338
1010	2.599369
1011	2.165724
1100	1.731965
1101	1.406509
1110	1.352248
1111	1.297981

Table 5. Height Response Factors (h)

Code	h
0000	1.000000
0001	1.071519
0010	1.148154
0011	1.230269
0100	1.318257
0101	1.445438
0110	1.603245
0111	1.757924
1000	1.949845
1001	2.137962
1010	2.317395
1011	2.540973
1100	2.786121
1101	3.019951
1110	3.311311
1111	3.672823

1.4 Smoothing Filter

The equalizer filters are followed by a continuous second order smoothing filter that removes the high frequency sample information generated by the action of the switched capacitor filters. This filter provides a continuous analog signal at the output, V_{OUT} .

1.5 Output Buffer

The final stage in the ML2020 is the output buffer. This amplifier has internal gain of 1 and is capable of driving 600Ω , 100pF loads. Thus, it is suitable for driving telephone hybrids directly without any external amplifier.

1.6 Bypass Mode

The filter sections can be bypassed by setting the bypass data bit, BP, to 0. Since the switched capacitor filters are bypassed in this mode, frequency response effects of the switched capacitor filters are eliminated. Thus, this mode offers very flat response and low noise over the 300–4000 Hz frequency range.

1.7 Filter Clock

The master clock, CLK, is used to generate the internal clocks for the switched capacitor filters. The frequency of CLK can be either 1.544 MHz or 1.536 MHz. However, the internal clock frequency must be kept at 1.536 MHz to guarantee accurate frequency response. The CLKSEL pin enables a bit swallower circuit to keep the internal clock frequency set to 1.536 MHz. When 1.544 MHz clock is used, CLKSEL should be set to logic level 0, and one bit out of every 193 bits is removed (swallowed) to reduce the internal frequency to 1.536 MHz. When 1.536 MHz clock is used, CLKSEL should be set to logic level 1, and the internal clock rate is the same as the external clock rate.

1.8 Serial Interface

The architecture of the digital section is shown in the preceding block diagram.

A timing diagram for the serial interface is shown in Figure 6. The serial input data, SID, is loaded into a shift register on rising edges of the shift clock, SCK. The data word is parallel loaded into a latch when the input latch signal, LATI, is high. The LATI pulse must occur when SCK is low. A new data word can be loaded into the shift register without disturbing the existing data word in the latch.

The parallel outputs of the latch control the filter response curves. The order of the data word bits in the latch is shown in Figure 7.

Note that bit 0 is the first bit of the data word clocked into the shift register.

The device has the capability to read out the data word stored in the latch. This is done by parallel loading the data from the latch back into the shift register when the latch signal, LATO, is high. The LATO pulse must occur when SCK is low. Then, the data word can be shifted out of the register serially to the output, SOD, on falling edges of the shift clock, SCK.

The loading and reading of the data word can be done continuously or in bursts. Since the shift register and latch

circuitry inside the device is static, there are no minimum frequency requirements on the clocks or data pulses. However, there is some coupling of the digital signals into the analog section. If this coupling is undesirable, the data can be clocked in bursts during non critical intervals, or the data rate can be done at a frequency outside the analog frequency range.

The clocks used to shift and latch data (SCK, LATI, LATO) are not related internally to the master clock and can occur asynchronous to CLK.

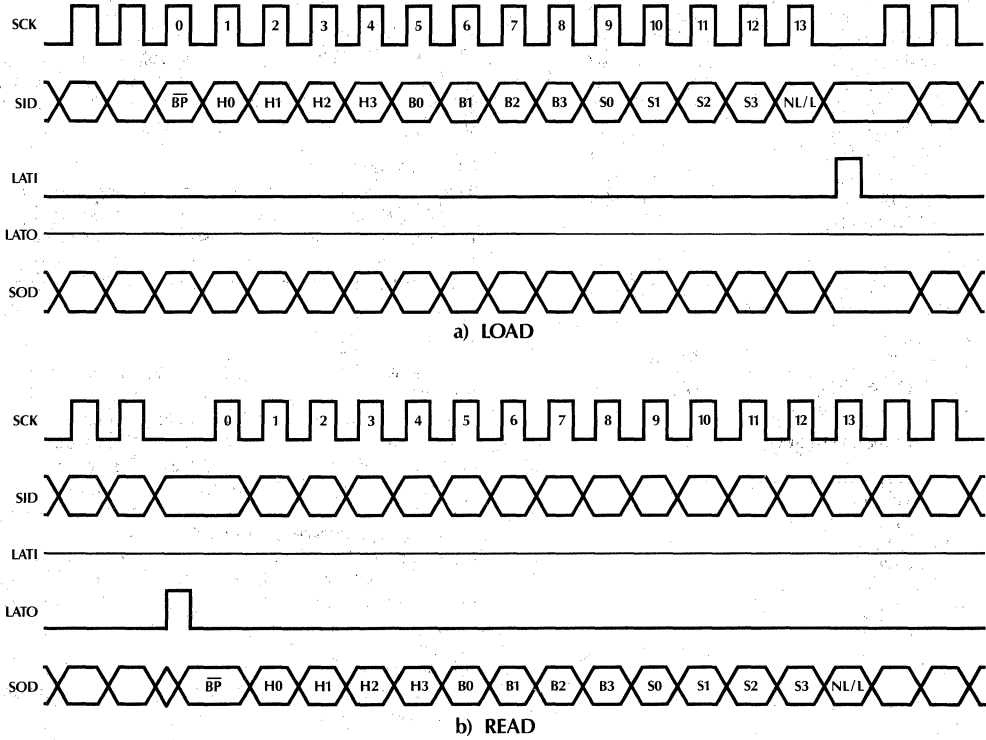


Figure 6. Serial Timing

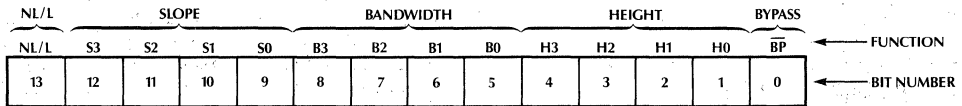


Figure 7. 14-Bit Latch

POWERDOWN MODE

A powerdown mode can be selected with pin P_{DN}. When P_{DN} = 1, the device is powered down. In this state, the power consumption is reduced by removing power from the analog section and forcing the analog output, V_{OUT}, to a high impedance state. While the device is in power down mode, the digital section is still functional and the current data word remains stored in the latch. The master clock, CLK, can be left active or removed during powerdown mode. When P_{DN} = 0, the device is in normal operation.

POWER SUPPLIES

The digital section inside the device is powered between V_{CC} and GND, or 5 volts. The analog section is powered between V_{CC} and V_{SS}, or ± 5 volts. The analog section uses AGND as the reference point.

GND and AGND are totally isolated inside the device to minimize coupling from the digital section into the analog section. Typically this is less than 100µV. However, AGND and GND should be tied together physically near the device and close to the common power supply ground connection.

The power supply rejection of V_{CC} and V_{SS} to the analog output is greater than -60dB at 1kHz, typically. If decoupling of the power supplies is still necessary in a system, V_{CC} and V_{SS} should be decoupled with respect to AGND.

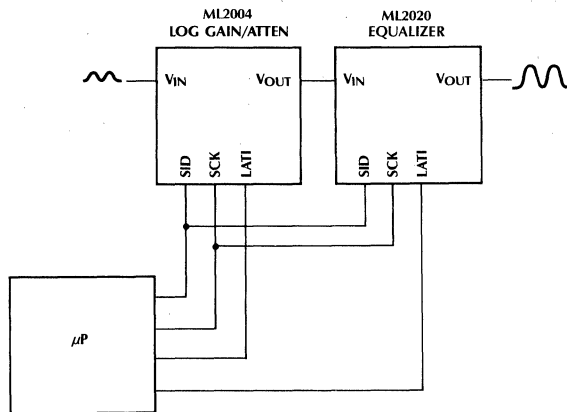


Figure 8. Typical Serial Interface

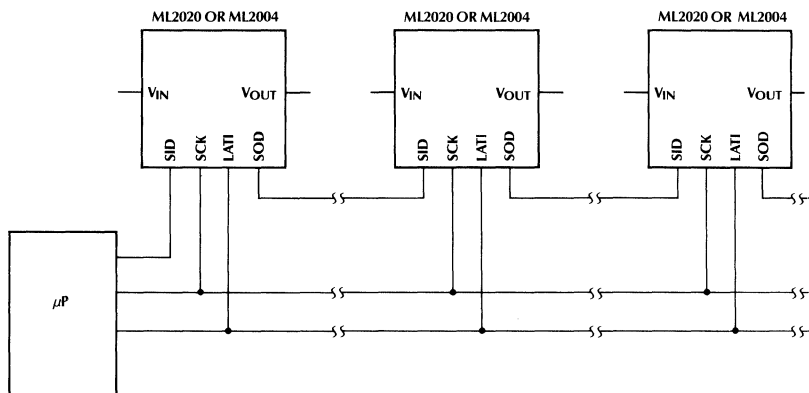


Figure 9. Controlling Multiple ML2020 and ML2004 With Only 3 Digital Lines Using One Long Data Word

ML2020

ORDERING INFORMATION

PART NUMBER	TEMP. RANGE	PACKAGE
ML2020CP	0°C to +70°C	MOLDED DIP (P16)
ML2020CS	0°C to +70°C	MOLDED SOIC (S18W)
ML2020IJ	-40°C to +85°C	HERMETIC DIP (J16)
ML2020IP	-40°C to +85°C	MOLDED DIP (P16)
ML2020IS	-40°C to +85°C	MOLDED SOIC (S18W)

Telephone Line Equalizer

GENERAL DESCRIPTION

The ML2021 is a monolithic analog line equalizer for telephone applications. The ML2021 consists of a switched capacitor filter that realizes a family of frequency response curves optimized for telephone line amplitude equalization while minimizing group delay. This ML2021 is the same function as the ML2020 telephone equalizer without the 60Hz rejection filter.

The ML2021 consists of a continuous anti-aliasing filter, three programmable switched capacitor equalization filters, an output smoothing filter, a 600Ω driver, and a digital section for the serial interface.

The equalization filters adjust the slope, height, and band-width of the frequency response. The desired frequency response is programmed by a digital 14-bit serial input data stream.

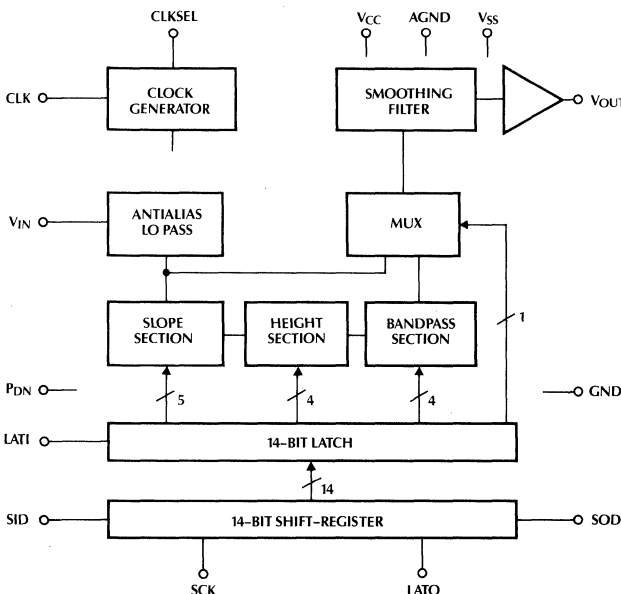
The ML2021 is implemented in a double polysilicon CMOS technology.

FEATURES

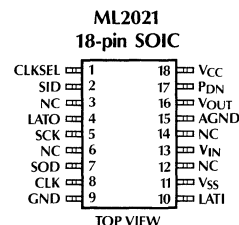
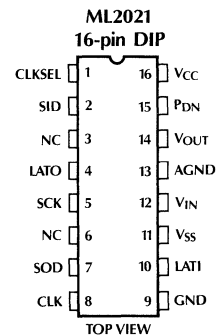
- Slope, height, and bandwidth adjustable
- Optimized group delays (500 Hz to 6.4 kHz)
- On chip anti-alias filter
- Bypass mode
- Low supply current 6mA typical from ±5V supplies
- TTL/CMOS compatible interface
- Double buffered data latch
- Selectable master clock 1.544 or 1.536 MHz
- Synchronous or asynchronous data loading capability
- Compatible with ML2003 and ML2004 logarithmic gain/attenuator
- Standard 16-pin 0.3" center molded or hermetic DIP and 18-pin SOIC
- 0°C to +70°C and -40°C to +85°C operating temperature range

3

BLOCK DIAGRAM



PIN CONNECTIONS



PIN DESCRIPTION

NAME	FUNCTION	NAME	FUNCTION
CLKSEL	Clock select input. This pin selects the frequency of the CLK input. If CLK is 1.536MHz, set CLKSEL = 1. If CLK is 1.544MHz, set CLKSEL = 0. Pin has an internal pullup resistor to V _{CC} .	GND	Digital ground. 0volts. All digital inputs and output are referenced to this ground.
SID	Serial input data. Digital input that contains serial data word which controls the filter frequency response setting.	LATI	Input latch clock. Digital input which loads data from the shift register into the latch.
LATO	Output latch clock. Digital input which loads the data word back into the shift register from the latch.	V _{SS}	Negative supply. -5 volts ±10%.
SCK	Shift clock. Digital input which shifts the serial data on SID into the shift register on rising edges and out onto SOD on falling edges.	V _{IN}	Analog input.
SOD	Serial output data. Digital output of the shift register.	AGND	Analog ground. 0volts. Analog input and output are referenced to this ground.
CLK	Master clock input. Digital input which generates clocks for the switched capacitor filters. Frequency can be either 1.544MHz or 1.536MHz.	V _{OUT}	Analog output.
		P _{DN}	Powerdown input. When P _{DN} = 1, device is in powerdown mode. When P _{DN} = 0, device is in normal operation. This pin has an internal pulldown resistor to GND.
		V _{CC}	Positive supply. 5volts ±10%

ABSOLUTE MAXIMUM RATINGS

(Note 1)

Supply Voltage	
V _{CC}	+6.5V
V _{SS}	-6.5V
AGND with Respect to GND	±0.5V
Analog Input and Output	V _{SS} - 0.3V to V _{CC} + 0.3V
Digital Input and Outputs	GND - 0.3V to V _{CC} + 0.3V
Input Current Per Pin	±25mA
Power Dissipation	750mW
Storage Temperature Range	-65°C to +150°C
Lead Temperature (Soldering 10 sec.)	300°C

OPERATING CONDITIONS

Temperature Range (Note 2)

ML2021CP, ML2021CS	0°C to +70°C
ML2021IJ	-40°C to +85°C
Supply Voltage	
V _{CC}	4V to 6V
V _{SS}	-4V to -6V

ELECTRICAL CHARACTERISTICS

Unless otherwise specified $T_A = T_{MIN}$ to T_{MAX} , $V_{CC} = 5V \pm 10\%$, $V_{SS} = -5V \pm 10\%$, Data Word: $\overline{BP} = 1$, Other Bits = 0, $C_L = 100$ pF, $R_L = 600\Omega$, dBm measurements use 600Ω as reference load, $V_{IN} = -7$ dBm, 1kHz sinusoid CLK = 1.544MHz ± 300 Hz and digital time measured at 1.4V

SYMBOL	PARAMETER	NOTES	CONDITIONS	MIN	TYP NOTE 3	MAX	UNITS	
ANALOG								
SR	Response, Slope Section	4	1kHz response					
			<u>NL/L</u> <u>S3</u> <u>S2</u> <u>S1</u> <u>S0</u>					
			0 0 0 0 1				1.4 \pm 0.1	dB
			0 0 0 1 0				2.6 \pm 0.2	dB
			0 0 1 0 0				4.7 \pm 0.2	dB
			0 1 0 0 0				7.8 \pm 0.2	dB
			0 1 1 1 1				11.4 \pm 0.25	dB
			1 0 0 0 0				0 \pm 0.1	dB
			1 0 0 0 1				0.4 \pm 0.1	dB
			1 0 0 1 0				0.9 \pm 0.2	dB
			1 0 1 0 0				1.8 \pm 0.2	dB
			1 1 0 0 0				3.7 \pm 0.2	dB
			1 1 1 1 1				6.6 \pm 0.25	dB
	Referenced to							
	0 0 0 0 0							
HR	Response, Height Section	4	3250Hz response referenced to 1kHz response with $\overline{BP} = 1$, other bits = 0					
			<u>NL/L</u> <u>H3</u> <u>H2</u> <u>H1</u> <u>H0</u>					
			0 0 0 0 0				0 \pm 0.15	dB
			0 0 0 0 1				0.5 \pm 0.2	dB
			0 0 0 1 0				1.1 \pm 0.2	dB
			0 0 1 0 0				2.3 \pm 0.2	dB
			0 1 0 0 0				5.7 \pm 0.3	dB
			0 1 1 1 1				11.1 \pm 0.3	dB
BR	Response, Bandwidth Section (Q)	4	<u>NL/L</u> <u>B3</u> <u>B2</u> <u>B1</u> <u>B0</u> <u>H3</u> <u>H2</u> <u>H1</u> <u>H0</u>					
			0 0 0 0 0 1 1 1 1				16.1 \pm 2.0	
			0 0 0 0 1 1 1 1 1				14.2 \pm 1.5	
			0 0 0 1 0 1 1 1 1				12.6 \pm 1.5	
			0 0 1 0 0 1 1 1 1				9.1 \pm 1.0	
			0 1 0 0 0 1 1 1 1				3.6 \pm 0.5	
			0 1 1 1 1 1 1 1 1				1.2 \pm 0.35	
PK	BW Peak Frequency	4	H3 thru H0 = 1	3230	3250	3270	Hz	
AG	Absolute Gain, Flat Response	4	.5 to 4kHz	-0.1	+0.1	+0.3	dB	
AGB	Absolute Gain, Bypass Mode	4	0.3 to 4kHz, $\overline{BP} = 0$	-0.1	+0.1	+0.3	dB	
ICN	Idle Channel Noise	4	$V_{IN} = 0$		3	8	dBrc	
			$V_{IN} = 0$, all data bits = 1		9		dBrc	

3

ELECTRICAL CHARACTERISTICS (Continued)

Unless otherwise specified $T_A = T_{MIN}$ to T_{MAX} , $V_{CC} = 5V \pm 10\%$, $V_{SS} = -5V \pm 10\%$, Data Word: $\overline{BP} = 1$, Other Bits = 0, $C_L = 100$ pF, $R_L = 600\Omega$, dBm measurements use 600Ω as reference load, $V_{IN} = -7$ dBm, 1kHz sinusoid CLK = $1.544\text{MHz} \pm 300\text{Hz}$ and digital time measured at 1.4V

SYMBOL	PARAMETER	NOTES	CONDITIONS	MIN	TYP NOTE 3	MAX	LIMIT UNITS
ANALOG							
HD	Harmonic Distortion	4	$V_{IN} = 5$ dBm, 1 kHz Measure 2nd, 3rd, harmonic relative to fundamental			-48	dB
SD	Signal to Distortion	4	$V_{IN} = -12$ dBm, 1 kHz C msg weighted	+48			dB
SFN	Single Frequency Noise	5	$V_{IN} = 0$, $4\text{kHz} \leq \text{frequency} \leq 150\text{kHz}$			-50	dBm
PSRR	Power Supply Rejection	4	200 mV _{p-p} , 1 kHz sine, $V_{IN} = 0$ on V_{CC} on V_{SS}			-40 -40	dB dB
Z_{IN}	Input Impedance, V_{IN}	4		100			k Ω
V_{OS}	Output Offset Voltage	4	$V_{IN} = 0$			± 50	mV
V_{INR}	Input Voltage Range	4		± 2.0			V
V_{OSW}	Output Voltage Swing	4	$R_L = 600\Omega$	± 2.0			V
DIGITAL AND DC							
V_{IL}	Digital Input Low Voltage	4				0.8	V
V_{IH}	Digital Input High Voltage	4		2.0			V
V_{OL}	Digital Output Low Voltage	4	$I_{OL} = 2$ mA			0.4	V
V_{OH}	Digital Output High Voltage	4	$I_{OH} = -1$ mA	4.0			V
I_{LCLK}	Input Current, CLK SEL	4	$V_{IN} = 0$	5		100	μ A
I_{LPDN}	Input Current, PDN	4	$V_{IN} = V_{CC}$	-5		-100	μ A
I_L	Input Current, All Other Inputs	4	$V_{IN} = 0$ to V_{CC}			± 10	μ A
I_{CC}	V_{CC} Supply Current	4	No output load, $V_{IL} = \text{GND}$, $V_{IH} = V_{CC}$, $V_{IN} = 0$			10	mA
I_{SS}	V_{SS} Supply Current	4	No output load, $V_{IL} = \text{GND}$, $V_{IH} = V_{CC}$, $V_{IN} = 0$			-10	mA
I_{CCP}	V_{CC} Supply Current, Powerdown Mode	4	No output load, $V_{IL} = \text{GND}$, $V_{IH} = V_{CC}$			1.2	mA
I_{SSP}	V_{SS} Supply Current, Powerdown Mode	4	No output load, $V_{IL} = \text{GND}$, $V_{IH} = V_{CC}$			-1.2	mA
AC CHARACTERISTICS							
t_{DC}	Clock Duty Cycle	5		40		60	%
t_{SCK}	SCK On/Off Period	4		250			ns
t_S	SID Data Setup Time	4		50			ns
t_H	SID Data Hold Time	4		50			ns
t_D	SOD Data Delay	4		0		125	ns
t_{IPW}	LATI Pulse Width	4		50			ns
t_{OPW}	LATO Pulse Width	4		50			ns
t_{IS}, t_{OS}	LATI, LATO Setup Time	4		50			ns
t_{IH}, t_{OH}	LATI, LATO Hold Time	5		50			ns
t_{PLD}	SOD Parallel Load Delay	4		0		125	ns

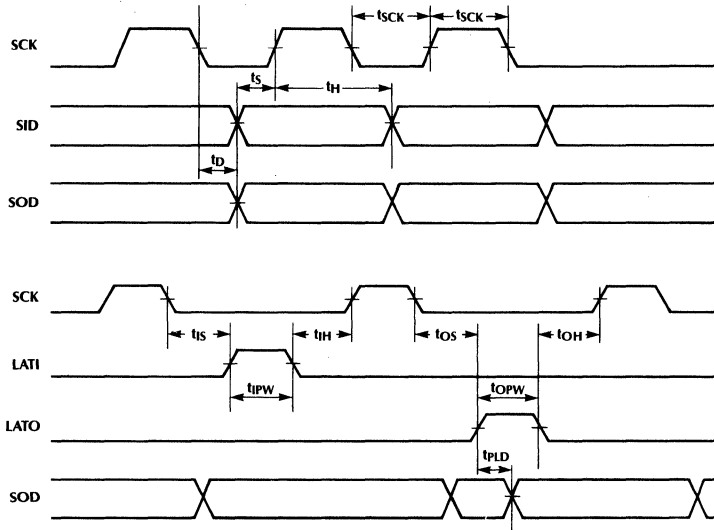
Note 1: Absolute maximum ratings are limits beyond which the life of the integrated circuit may be impaired. All voltages unless otherwise specified are measured with respect to ground.

Note 2: 0°C to $+70^\circ\text{C}$ and -40°C to $+85^\circ\text{C}$ operating temperature range devices are 100% tested with temperature limits guaranteed by 100% testing, sampling, or by correlation with worst-case test conditions.

Note 3: Typicals are parametric norm at 25°C .

Note 4: Parameter guaranteed and 100% production tested.

Note 5: Parameter guaranteed. Parameters not 100% tested are not in outgoing quality level calculation.



TIMING PARAMETERS ARE REFERENCED TO THE 1.4 VOLT MIDPOINT

Figure 1. Serial Timing Diagram

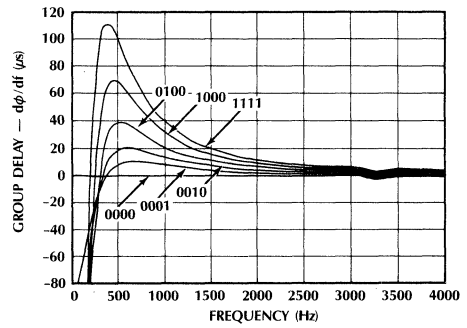
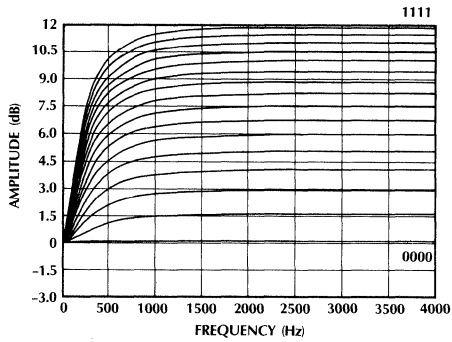


Figure 2. Typical Slope Filter Response — NL/L = 0
B3-B0, H3-H0 = 0000, S3-S0 = 0000 to 1111.

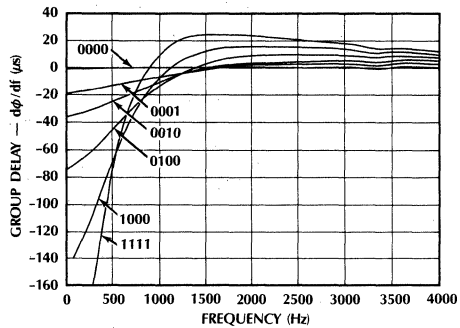
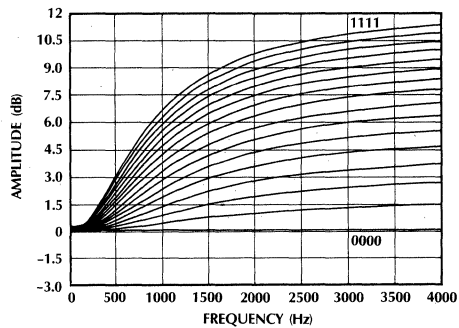


Figure 3. Typical Slope Filter Response — NL/L = 1
 B3-B0, H3-H0 = 0000, S3-S0 = 0000 to 1111.

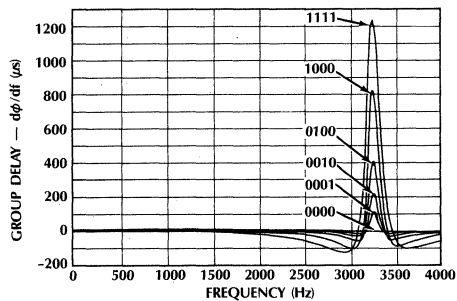
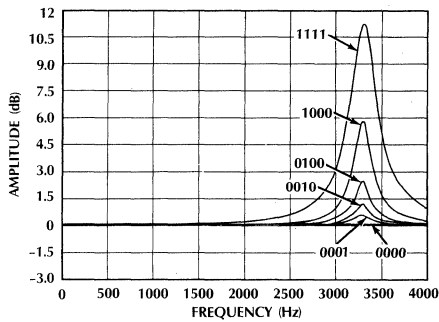


Figure 4. Typical Height Filter Response — NL/L = 0
 B3-B0, S3-S0 = 0000; H3-H0 = 0000 to 1111.

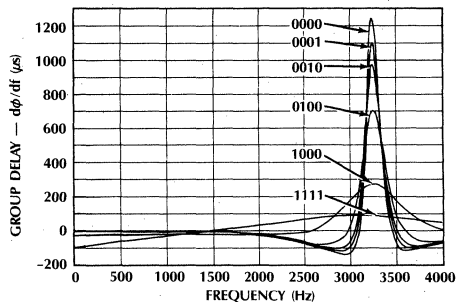
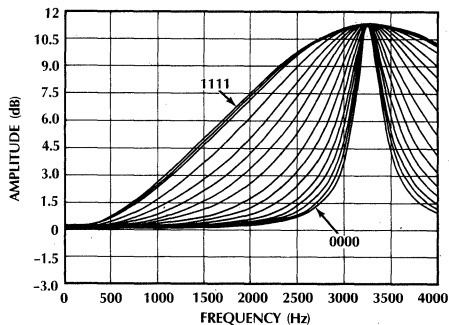


Figure 5. Typical Bandwidth Filter Response — NL/L = 0
 H3-H0 = 1111; S3-S0 = 0000; B3-B0 = 0000 to 1111.

1.0 FUNCTIONAL DESCRIPTION

The ML2021 consists of a continuous anti-alias filter, three programmable switched capacitor equalization filters, an output smoothing filter, an output driver, and a digital section for the serial interface.

1.1 ANTI-ALIAS FILTER

The first section is a continuous anti-alias filter. This filter is needed to prevent aliasing of high frequency signals present on the input into the passband by the sampling action of the switched capacitor filters. This section is a continuous second order lowpass filter with a typical 3 dB frequency at 20 kHz and 30 dB of rejection at 124 kHz.

1.2 EQUALIZATION FILTERS

The programmable filters implement a family of frequency response curves intended to compensate for the response of telephone lines.

This filter is composed of three distinct sections: slope, height, and bandwidth.

1.2.1 RESPONSE OF SLOPE, HEIGHT, AND BANDWIDTH

The family of response curves generated by the slope section are shown in Figures 2 and 3. There are 4 slope select bits, S3-S0. These bits alter the slope of the highpass response under 1000 Hz, and as a result, the absolute gain above 1000 Hz will be unique for each setting. Table 1 gives typical 1 kHz gain values for all slope settings.

Table 1. Typ. 1kHz Gain for Slope Settings

Slope Setting	Rel 1kHz Gain (dB)	
	NL/L = 1	NL/L = 0
0	0.0	Rel
1	0.4	1.4
2	0.9	2.6
3	1.4	3.7
4	1.8	4.7
5	2.3	5.5
6	2.8	6.3
7	3.4	7.2
8	3.7	7.8
9	4.2	8.4
10	4.6	9.0
11	5.0	9.5
12	5.4	10.0
13	5.8	10.5
14	6.2	11.0
15	6.6	11.4

HT, BW Bits = 0

There is an additional bit, NL/L, that also affects the highpass response of the slope filter. The slope response curves in Figure 2 are with NL/L = 0. These same response curves are shown in Figure 3 with NL/L = 1. Notice that the NL/L bit adds more droop in the highpass response below 2500 Hz.

The family of response curves generated by the height section are shown in Figure 4. There are 4 height select bits, H3-H0. This section creates a peak in the response at 3250 Hz and this filter controls the amount of peaking. Table 2 gives typical 1 kHz gain values for all height and bandwidth settings.

Table 2. Typ. 1kHz Gain for HT and BW Settings

		Relative 1kHz Gain (dB)															
		HT Setting															
BW Setting		0	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15
		0	Rel	0	0	0	0	0	0	0	0	0	0	0	0	0	0
1		0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0
2		0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0
3		0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0
4		0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0.1
5		0	0	0	0	0	0	0	0	0	0	0	0	0	0.1	0.1	0.1
6		0	0	0	0	0	0	0	0	0	0	0.1	0.1	0.1	0.1	0.1	0.1
7		0	0	0	0	0	0	0	0	0.1	0.1	0.1	0.1	0.1	0.2	0.2	0.2
8		0	0	0	0	0	0.1	0.1	0.1	0.1	0.1	0.1	0.2	0.2	0.3	0.3	0.4
9		0	0	0	0	0.1	0.1	0.1	0.1	0.1	0.2	0.2	0.3	0.3	0.4	0.5	0.6
10		0	0	0	0.1	0.1	0.1	0.1	0.1	0.2	0.3	0.3	0.4	0.5	0.6	0.7	0.8
11		0	0	0.1	0.1	0.1	0.2	0.2	0.3	0.4	0.4	0.5	0.7	0.8	0.9	1.1	
12		0	0.1	0.1	0.1	0.2	0.2	0.3	0.4	0.5	0.7	0.8	1.0	1.1	1.4	1.6	
13		0	0.1	0.1	0.2	0.3	0.4	0.5	0.6	0.8	0.9	1.1	1.4	1.6	1.9	2.3	
14		0	0.1	0.1	0.2	0.3	0.4	0.5	0.7	0.8	1.0	1.2	1.5	1.7	2.0	2.4	
15		0	0.1	0.1	0.2	0.3	0.4	0.5	0.7	0.9	1.1	1.3	1.6	1.8	2.1	2.5	

Slope Bits = 0

The family of response curves generated by the bandwidth section is shown in Figure 5. There are 4 bandwidth select bits, B3-B0. This section causes the response of the 3250 Hz peak to be widened, and as a result, this filter controls the bandwidth of the 3250 Hz peaked region.

1.2.2 TRANSFER FUNCTION

The transfer function for the ML2021 is shown below. This transfer function is valid for magnitude response only. The actual magnitude response from an individual device may deviate from the computed response from the transfer function by typically 0-0.2 dB.

$$H(s) = \frac{c(s+b)}{b(s+c)} \times \frac{[s^2 + h(\omega_o/Q)s + \omega_o^2]}{[s^2 + (\omega_o/Q)s + \omega_o^2]} \times \frac{[\sin(\pi f/c)]}{(\pi f/c)}$$

$$s = j \times 256000 \times \tan(\pi f / 128000)$$

$$\omega_o = 20463.77$$

$$f_c = 128000$$

b,c : See Table 3. (slope)

Q : See Table 4. (bandwidth)

h : See Table 5. (height)

3

Table 3. Slope Response Factors (b, c)

S3-0	b NL/L=0	b NL/L=1
0000	2.371759E+03	1.116280E+04
0001	1.985920E+03	9.345141E+03
0010	1.701779E+03	8.007156E+03
0011	1.493571E+03	7.026999E+03
0100	1.326721E+03	6.241681E+03
0101	1.196668E+03	5.629636E+03
0110	1.087277E+03	5.114881E+03
0111	9.983588E+02	4.696487E+03
1000	9.179889E+02	4.318339E+03
1001	8.537864E+02	4.016273E+03
1010	7.966049E+02	3.747249E+03
1011	7.478074E+02	3.517676E+03
1100	7.035099E+02	3.309279E+03
1101	6.651771E+02	3.128945E+03
1110	6.299477E+02	2.963214E+03
1111	5.990361E+02	2.817797E+03

S3-0	c NL/L=0	c NL/L=1
XXXX	2.371759E+03	1.116280E+04

Table 4. Slope Response Factors (b, c)

B3-0	Q
0000	17.444906
0001	15.386148
0010	13.652451
0011	11.593677
0100	9.859960
0101	8.017864
0110	6.392453
0111	5.092080
1000	3.900003
1001	3.141338
1010	2.599369
1011	2.165724
1100	1.731965
1101	1.406509
1110	1.352248
1111	1.297981

Table 5. Height Response Factors (h)

Code	h
0000	1.000000
0001	1.071519
0010	1.148154
0011	1.230269
0100	1.318257
0101	1.445438
0110	1.603245
0111	1.757924
1000	1.949845
1001	2.137962
1010	2.317395
1011	2.540973
1100	2.786121
1101	3.019951
1110	3.311311
1111	3.672823

1.2.3 GROUP DELAY

The difference between the ML2020 and ML2021 is the elimination of a 60Hz highpass filter in order to eliminate positive group delay at low frequency.

The group delay through the ML2021 can be minimized such that less than 50µs of group delay can be achieved in both unloaded and cable loaded conditions relative to 1804Hz in the frequency range of 504 to 3004Hz. Minimum group delays are dependent upon using the proper setting for slope, height, and bandwidth for a give equalization requirement.

1.3 SMOOTHING FILTER

The equalizer filters are followed by a continuous second order smoothing filter that removes the high frequency sample information generated by the action of the switched capacitor filters. This filter provides a continuous analog signal at the output, V_{OUT}.

1.4 OUTPUT BUFFER

The final stage in the ML2020 is the output buffer. This amplifier has internal gain of 1 and is capable of driving 600Ω, 100pF loads. Thus, it is suitable for driving telephone hybrids directly without any external amplifier.

1.5 BYPASS MODE

The filter sections can be bypassed by setting the bypass data bit, BP, to 0. Since the switched capacitor filters are bypassed in this mode, frequency response effects of the switched capacitor filters are eliminated. Thus, this mode offers very flat response and low noise over the 300-4000Hz frequency range.

1.6 FILTER CLOCK

The master clock, CLK, is used to generate the internal clocks for the switched capacitor filters. The frequency of CLK can be either 1.544MHz or 1.536MHz. However, the internal clock frequency must be kept at 1.536MHz to guarantee accurate frequency response. The CLKSEL pin enables a bit swallower circuit to keep the internal clock frequency set to 1.536MHz. When 1.544MHz clock is used, CLKSEL should be set to logic level 0, and one bit out of every 193 bits is removed (swallowed) to reduce the internal frequency to 1.536MHz. When 1.536MHz clock is used, CLKSEL should be set to logic level 1, and the internal clock rate is the same as the external clock rate.

1.7 SERIAL INTERFACE

The architecture of the digital section is shown in the preceding block diagram.

A timing diagram for the serial interface is shown in Figure 6. The serial input data, SID, is loaded into a shift register on rising edges of the shift clock, SCK. The data word is parallel loaded into a latch when the input latch signal, LATI, is high. The LATI pulse must occur when SCK is low. A new data word can be loaded into the shift register without disturbing the existing data word in the latch.

The parallel outputs of the latch control the filter response curves. The order of the data word bits in the latch is shown in Figure 7.

Note that bit 0 is the first bit of the data word clocked into the shift register.

The device has the capability to read out the data word stored in the latch. This is done by parallel loading the data from the latch back into the shift register when the latch signal, LATO, is high. The LATO pulse must occur when SCK is low. Then, the data word can be shifted out of the register serially to the output, SOD, on falling edges of the shift clock, SCK.

The loading and reading of the data word can be done continuously or in bursts. Since the shift register and latch

circuitry inside the device is static, there are no minimum frequency requirements on the clocks or data pulses. However, there is some coupling of the digital signals into the analog section. If this coupling is undesirable, the data can be clocked in bursts during non critical intervals, or the data rate can be done at a frequency outside the analog frequency range.

The clocks used to shift and latch data (SCK, LATI, LATO) are not related internally to the master clock and can occur asynchronous to CLK.

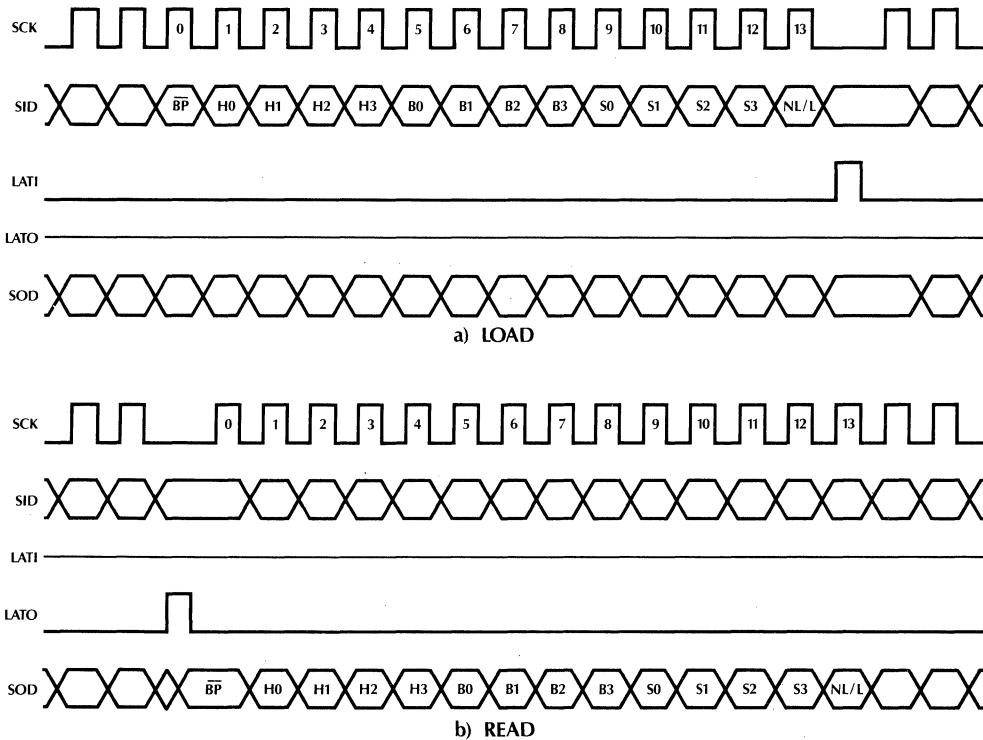


Figure 6. Serial Timing

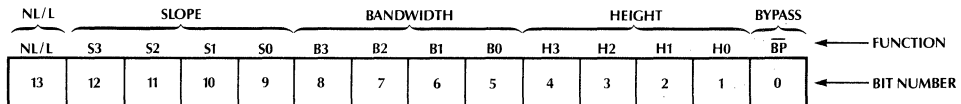


Figure 7. 14-Bit Latch

1.8 POWERDOWN MODE

A powerdown mode can be selected with pin P_{DN} . When $P_{DN}=1$, the device is powered down. In this state, the power consumption is reduced by removing power from the analog section and forcing the analog output, V_{OUT} , to a high impedance state. While the device is in power down mode, the digital section is still functional and the current data word remains stored in the latch. The master clock, CLK , can be left active or removed during powerdown mode. When $P_{DN}=0$, the device is in normal operation.

1.9 POWER SUPPLIES

The digital section inside the device is powered between V_{CC} and GND , or 5 volts. The analog section is powered between V_{CC} and V_{SS} , or ± 5 volts. The analog section uses $AGND$ as the reference point.

GND and $AGND$ are totally isolated inside the device to minimize coupling from the digital section into the analog section. Typically this is less than $100\mu V$. However, $AGND$ and GND should be tied together physically near the device and close to the common power supply ground connection.

The power supply rejection of V_{CC} and V_{SS} to the analog output is greater than $-60dB$ at $1kHz$, typically. If decoupling of the power supplies is still necessary in a system, V_{CC} and V_{SS} should be decoupled with respect to $AGND$.

2.0 APPLICATIONS

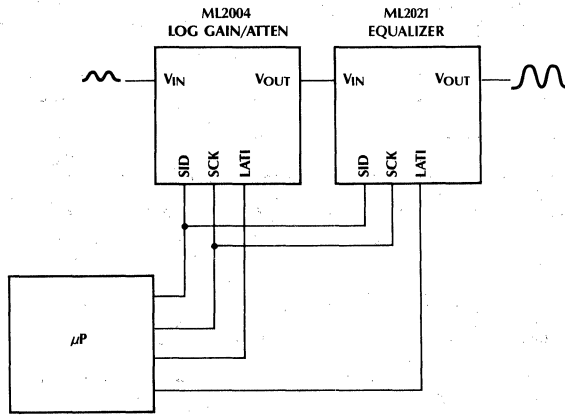


Figure 8. Typical Serial Interface

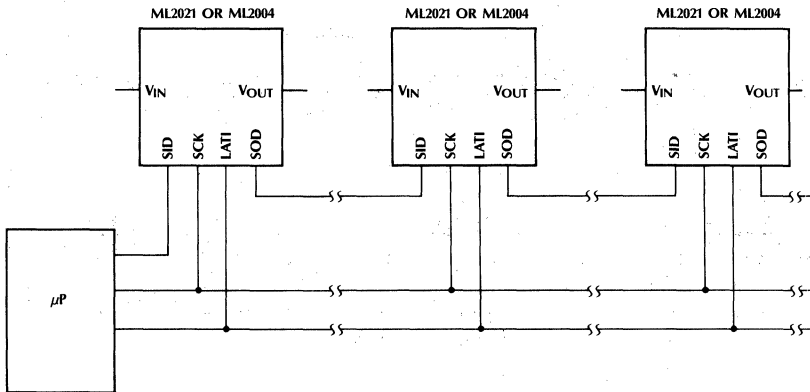
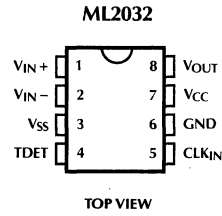
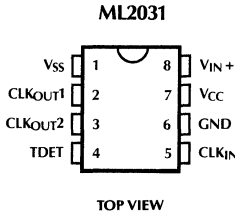


Figure 9. Controlling Multiple ML2021 and ML2004 With Only 3 Digital Lines Using One Long Data Word

ORDERING INFORMATION

PART NUMBER	TEMP. RANGE	PACKAGE
ML2021CP	0°C to +70°C	MOLDED DIP (P16)
ML2021CS	0°C to +70°C	MOLDED SOIC (S18W)
ML2021IJ	-40°C to +85°C	HERMETIC DIP (J16)
ML2021IP	-40°C to +85°C	MOLDED DIP (P16)
ML2021IS	-40°C to +85°C	MOLDED SOIC (S18W)

PIN CONNECTIONS



PIN DESCRIPTIONS

ML2031		
PIN NO.	NAME	FUNCTION
1	V _{SS}	Negative supply. $-5V \pm 10\%$
2	CLK _{OUT1}	Clock output. Digital output from oscillator divided by 2.
3	CLK _{OUT2}	Clock output. Digital output from oscillator divided by 8.
4	TDET	Tone detect output. Digital output which indicates when valid 2713 Hz tone is present on analog input.
5	CLK _{IN}	Clock input. Internal clock can be generated by tying a 12.352 MHz crystal between this pin and GND, or by applying a 12.352 MHz or 1.544 MHz clock to this pin.
6	GND	Ground. Analog and digital inputs and outputs are referenced to this point.
7	V _{CC}	Positive supply. $+5V \pm 10\%$
8	V _{IN+}	Analog input.

ML2032		
PIN NO.	NAME	FUNCTION
1	V _{IN+}	Positive Analog input. Positive input to the uncommitted op amp.
2	V _{IN-}	Negative Analog input. Negative input to the uncommitted op amp.
3	V _{SS}	Negative supply. $-5V \pm 10\%$
4	TDET	Tone detect output. Digital output which indicates when valid 2713 Hz tone is present on analog input.
5	CLK _{IN}	Clock input. Internal clock can be generated by tying a 12.352 MHz crystal between this pin and GND, or by applying a 12.352 MHz or 1.544 MHz clock to this pin.
6	GND	Ground. Analog and digital inputs and outputs are referenced to this point.
7	V _{CC}	Positive supply. $+5V \pm 10\%$
8	V _{OUT}	Analog output. Output of the uncommitted op amp.

ABSOLUTE MAXIMUM RATINGS

(Note 1)

Supply Voltage	
V_{CC}	+6.5V
V_{SS}	-6.5V
Analog Input and Output	$V_{SS} - 0.3V$ to $V_{CC} + 0.3V$
Digital Input and Outputs	-0.3V to $V_{CC} + 0.3V$
Input Current Per Pin	± 25 mA
Power Dissipation	750 mW
Storage Temperature Range	-65°C to +150°C
Lead Temperature (Soldering 10 sec.)	300°C

OPERATING CONDITIONS

Temperature Range (Note 2)

ML2031CP, ML2032CP	0°C to +70°C
ML2031IJ, ML2032IJ	-40°C to +85°C
Supply Voltage	
V_{CC}	4V to 6V
V_{SS}	-4V to -6V

ELECTRICAL CHARACTERISTICS

Unless otherwise specified $T_A = T_{MIN}$ to T_{MAX} , $V_{CC} = 5V \pm 10\%$, $V_{SS} = -5V \pm 10\%$, $CLK_{IN} = 12.352$ MHz ± 1200 Hz, or $CLK_{IN} = 1.544$ MHz ± 150 Hz, $C_L = 100$ pF, dBm measurements use 600 Ω as reference load, uncommitted op amp in unity gain configuration.

SYMBOL	PARAMETER	NOTES	CONDITIONS	MIN	TYP NOTE 3	MAX	UNITS
tone DETECT							
f_{TD}	Tone Detection Frequency	4	$V_{IN} = +6$ dBm to -34 dBm	2703		2723	Hz
f_{TR}	Tone Rejection Frequency	4		2679		2747	Hz
A_{TD}	Tone Detection Amplitude	4	$V_{IN} = 2703$ Hz to 2723 Hz	-34		+6	dBm
A_{TR}	Tone Rejection Amplitude	4		-40			dBm
SGM	Signal to Guard Margin	4	800 Hz 1400 Hz 2000 Hz 2450 Hz Signal = -13 dBm, 2713 Hz. See BELL PUB 43004 sec. 2.4 for test method	8 8 8 8		13 13 13 13	dB dB dB dB
SFI	SF Tone Immunity	5	$V_{IN} + = 2600$ Hz No tone detect			+6	dBm
t_{TD}	Tone Detect Delay	4	$V_{IN} + = -8$ dBm, 2713 Hz Figure 1	0	10	30	ms
t_{TR}	Tone Removal Delay	4	$V_{IN} + = -8$ dBm, 2713 Hz Figure 1	0	4	30	ms
OP AMP							
V_{INR}	Input Voltage Range	5		± 3			V
V_{OSW}	Output Voltage Swing	4	ML2032 Only	± 3			V
V_{OS}	Input Offset Voltage	4	ML2032 Only			± 20	mV
Z_{IN}	Input Impedance	4		1			M Ω
A_{VOL}	DC Open Loop Gain	4		1k	5k		V/V
f_{UG}	Unity Gain Frequency	5		0.5	1		MHz
I_{CN}	Noise-Input Referred	5	C msg weighted 1kHz		-9	-3 375	dB r_{nc} nV/ \sqrt{Hz}

ELECTRICAL CHARACTERISTICS (Continued)

Unless otherwise specified $T_A = T_{MIN}$ to T_{MAX} , $V_{CC} = 5V \pm 10\%$, $V_{SS} = -5V \pm 10\%$, $CLK_{IN} = 12.352MHz \pm 1200Hz$, or $CLK_{IN} = 1.544MHz \pm 150Hz$, $C_L = 100pF$, dBm measurements use 600Ω as reference load, uncommitted op amp in unity gain configuration.

SYMBOL	PARAMETER	NOTES	CONDITIONS	MIN	TYP NOTE 3	MAX	UNITS
DIGITAL AND DC							
V_{IL}	Input Low Voltage, CLK_{IN}	4				1.5	V
V_{IH}	Input High Voltage, CLK_{IN}	4		3.5			V
I_{IN}	Input Current, CLK_{IN}	4	$CLK_{IN} = 1.5V$ to $3.5V$		10	60	μA
			$CLK_{IN} = 0$ to $1.5V$; $3.5V$ to V_{CC}		150	500	μA
C_{IN}	Input Capacitance, CLK_{IN}	5			11		pF
V_{OL}	Output Low Voltage	4	$I_{OL} = -2mA$			0.4	V
V_{OH}	Output High Voltage	4	$I_{OH} = 2mA$	4.0			V
I_{CC}	V_{CC} Supply Current	4	No output load			7.5	mA
I_{SS}	V_{SS} Supply Current	4	No output load			-4.5	mA
CLOCK OUTPUT							
f_{CLK1}	CLK_{OUT1} Output Frequency	4	Figure 2	$\frac{1}{2}$		$\frac{1}{2}$	f_{CLK1}
f_{CLK2}	CLK_{OUT2} Output Frequency	4	Figure 2	$\frac{1}{8}$		$\frac{1}{8}$	f_{CLK1}
t_{1R}	CLK_{OUT1} Output Rise Time	4	Figure 2, $C_L = 50pF$	0		20	ns
t_{1F}	CLK_{OUT1} Output Fall Time	4	Figure 2, $C_L = 50pF$	0		20	ns
t_{2R}	CLK_{OUT2} Output Rise Time	4	Figure 2, $C_L = 50pF$	0		20	ns
t_{2F}	CLK_{OUT2} Output Fall Time	4	Figure 2, $C_L = 50pF$	0		20	ns

Note 1: Absolute maximum ratings are limits beyond which the life of the integrated circuit may be impaired. All voltages unless otherwise specified are measured with respect to ground.

Note 2: $0^\circ C$ to $+70^\circ C$ and $-40^\circ C$ to $+85^\circ C$ operating temperature range devices are 100% tested with temperature limits guaranteed by 100% testing, sampling, or by correlation with worst-case test conditions.

Note 3: Typicals are parametric norm at $25^\circ C$.

Note 4: Parameter guaranteed and 100% production tested.

Note 5: Parameter guaranteed. Parameters not 100% tested are not in outgoing quality level calculation.

TIMING DIAGRAMS

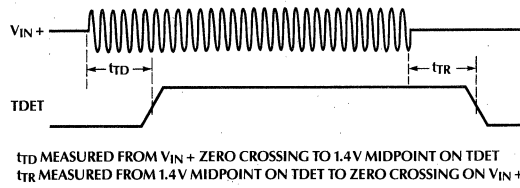


Figure 1. Tone Detect Timing

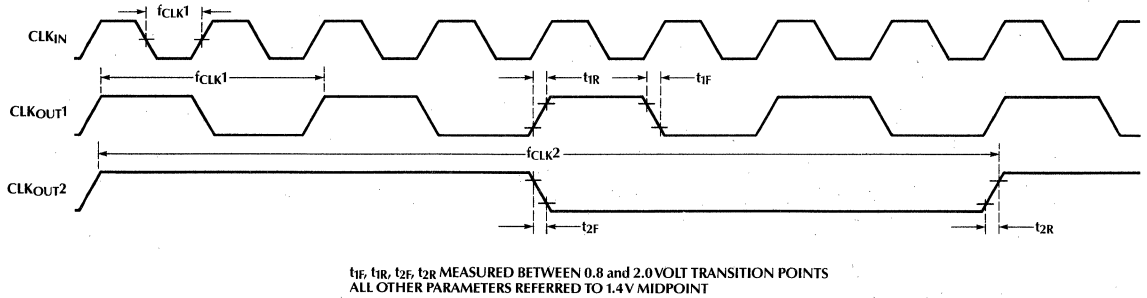


Figure 2. Digital Clock Output Timing

TYPICAL PERFORMANCE CURVE

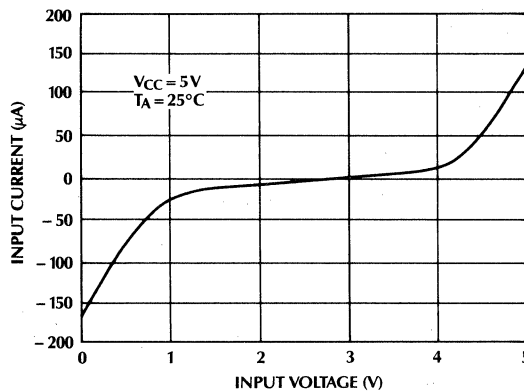


Figure 3. CLK_{IN} Input Current vs. Input Voltage

1.0 FUNCTIONAL DESCRIPTION

The ML2031 has a divide by 2 and divide by 8 clock output to drive external devices. The ML2032 has an uncommitted op amp. Refer to the block diagram.

1.1 Uncommitted Op Amp

The ML2032 features an uncommitted op amp. The ML2031 has the op amp connected in the unity gain configuration (V_{IN} – internally tied to V_{OUT}).

The uncommitted op amp is a general purpose amplifier that can be used to interface the device with the analog telephone line. It has a high impedance input, a 0.5MHz unity gain bandwidth, will drive a 1k, 100pF load, and the input and output can swing within 1.5V of the supplies.

1.2 Anti-Alias Filter

The anti-alias filter is a continuous second order low pass designed to prevent high frequency signals at the input from being aliased into the passband by the sampling action of the switched capacitor filters. The typical 3dB corner frequency is 25 kHz and the typical rejection at 124 kHz is – 30dB.

1.3 60Hz Reject Filter

The 60Hz reject filter is a switched capacitor second order high pass designed to reject 60Hz line interference on the analog input. The typical 3dB corner frequency is 300Hz and the typical rejection at 60Hz is – 24dB.

1.4 Tone Detector

The tone detector is a monolithic block designed to indicate when a valid 2713Hz tone is present on the analog input. A tone is valid if the following criteria are met:

1. 2713 Hz tone satisfies amplitude vs. frequency tone detector template shown in *Figure 4*.
2. The non-2713Hz out of band energy present on the input is sufficiently small enough compared to the 2713Hz tone (signal to guard margin).

The tone detector consists of 2713Hz bandpass and notch filters, tone and guard peak detectors, tone and guard comparators, reference, and digital output buffer.

The analog signal first goes through the 2713Hz bandpass and notch switched capacitor filters. The bandpass filter outputs any 2713Hz signal (tone), and the notch filter outputs any non-2713Hz signals (guard) in the range of 300–4500Hz, respectively.

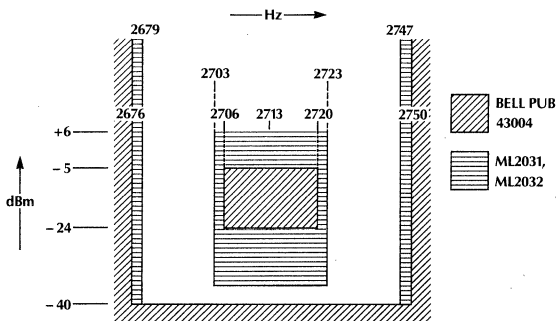


Figure 4. Tone Detector Template

The tone and guard signals then go to peak detectors which output a DC voltage proportional to the 2713Hz and non-2713Hz energy present on the analog input.

The tone comparator compares the tone energy to a fixed reference value to determine if it meets the amplitude requirements for tone detection shown in *Figure 4*.

The guard comparator compares the tone energy to the guard energy to determine if the signal to guard margin is met.

If both comparators indicate that a 2713Hz tone and no out of band energy exists, the TDET output goes high indicating valid tone detection. If the signal comparator indicates insufficient signal energy or the guard comparator indicates too much out of band energy, then the TDET output stays low indicating invalid tone output.

1.5 Crystal Oscillator/Clock Generator

The crystal oscillator/clock generator generates the necessary internal clocks from either an external clock or an external crystal.

If an external clock input is used to drive CLK_{IN} , the input frequency can either be 12.352MHz or 1.544MHz in order to meet the frequency template. The device has an internal frequency sense circuit that can sense the difference between 12.352MHz and 1.544MHz and makes the necessary changes in the clock generator to accommodate either frequency at the input.

If a crystal is used, a 12.352MHz crystal must be connected between CLK_{IN} and GND. This unique 1-pin crystal oscillator does not generally require any external capacitors or other external components to meet the frequency template. The crystal should be physically placed as close as possible to the CLK_{IN} pin to minimize stray inductances and capacitances.

The crystal must have the following characteristics:

1. Parallel resonant type
2. Frequency: 12.352000MHz
3. Tolerance: $\pm 0.005\%$ @ 25°C
4. Less than 0.005% variation over desired temperature range
5. Maximum equivalent series resistance of 15 Ω at a drive level of 1 μ W to 200 μ W
6. Maximum equivalent series resistance of 30 Ω at drive levels of 10 nW to 1 μ W
7. Typical load capacitance: 18 pF
8. Maximum case capacitance: 5 pF

The frequency of oscillation will be a function of the crystal parameters and board capacitance. If the final oscillation frequency is different than the ideal 12.352MHz, the template frequencies will change according to the formulas outlined in section 1.6. If the crystal meets the above recommended parameters and typical PC board capacitance from CLK_{IN} to GND is 2pF, then the device will meet the template specifications. Crystals that meet these requirements are M-tron 3709-010 12.352 for 0°C to +70°C and 3709-020 12.352 for –40°C to +85°C operation.

1.0 FUNCTIONAL DESCRIPTION (Continued)

The ML2031 has two clock outputs that can be used to drive other external devices. The CLK_{OUT1} output is a buffered output from the oscillator divided by 2. The CLK_{OUT2} output is a buffered output from the oscillator divided by 8. If a 12.352MHz clock or crystal is used, CLK_{OUT1}=6.176MHz and CLK_{OUT2}=1.544MHz.

1.6 Detecting Tones from 1000 Hz to 4000 Hz

The tone detector frequency template shown in Figure 5 is proportional to the frequency of CLK_{IN}. Thus, the device can be set to a center frequency (other than 2713 Hz) by adjusting CLK_{IN} frequency.

The external clock frequency, fCLK_{IN}, needed to produce a given center frequency, can be calculated by:

$$f_{CLK_{IN}} = f_C \times 4552.893$$

once fCLK_{IN} has been determined, the other template frequency points shown in Figure 5 can be calculated by:

$$f_{DL} = f_{CLK_{IN}} \times 2.18831 \times 10^{-4}$$

$$f_{DU} = f_{CLK_{IN}} \times 2.20450 \times 10^{-4}$$

$$f_{RL} = f_{CLK_{IN}} \times 2.16888 \times 10^{-4}$$

$$f_{RU} = f_{CLK_{IN}} \times 2.22393 \times 10^{-4}$$

The above formulas are valid for center frequencies with the range of 1000 Hz to 4000 Hz. The internal divide by 8 circuitry may be bypassed by applying a clock that is one eighth of the above calculated values.

When the required CLK_{IN} frequency calculated above is less than 6MHz, the internal frequency sense circuit may be-

come enabled causing the detection of an erroneous center frequency. In this case, the divide by 8 function cannot be used and only the lower clock frequency may be used. For example, for a 1004 Hz tone detector, the clock frequency applied must be 571 kHz.

1.7 Power Supplies

The analog circuits in the device run from +5 to -5 (V_{CC} to V_{SS}) and are referenced to GND.

The digital circuits in the device run from +5 to 0 (V_{CC} to GND).

It is recommended that the power supplies to the device be bypassed by placing decoupling capacitors from V_{CC} to GND and V_{SS} to GND as physically close to the device as possible.

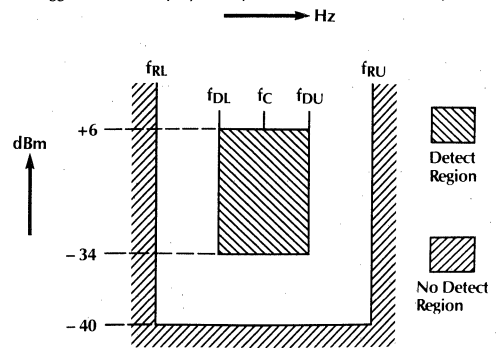


Figure 5. Tone Detector Template

2.0 APPLICATIONS

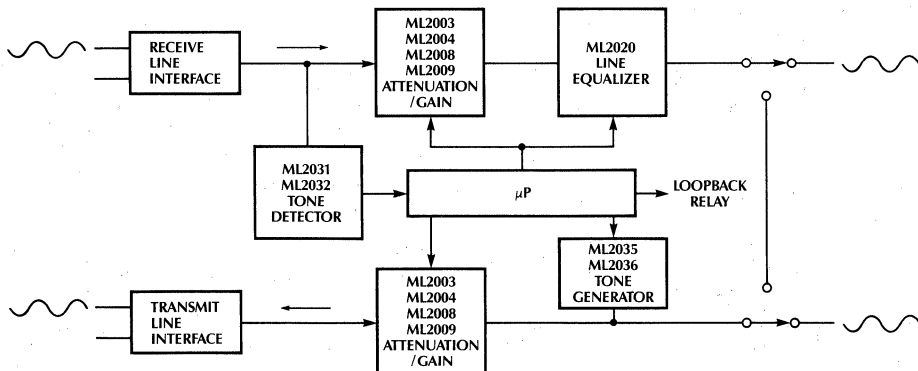


Figure 6. 4-Wire Termination Equipment

ORDERING INFORMATION

PART NUMBER	TEMP. RANGE	PACKAGE
ML2031CP	0°C to +70°C	MOLDED DIP (P08)
ML2031IP	-40°C to +85°C	MOLDED DIP (P08)
ML2032CP	0°C to +70°C	MOLDED DIP (P08)
ML2032IP	-40°C to +85°C	MOLDED DIP (P08)

Programmable Sinewave Generator

GENERAL DESCRIPTION

The frequency of these monolithic sinewave generators is programmable for the ML2035 from DC to 25kHz and for the ML2036 from DC to 50kHz. No external components are required.

The frequency of the sinewave output is derived from either an external crystal or clock input, thus providing a stable and accurate frequency reference. The frequency is programmed by a 16-bit serial data word.

The ML2035 is packaged in an 8-pin DIP and has a V_{OUT} amplitude of $\pm V_{CC}/2$.

The ML2036 provides for a V_{OUT} amplitude of either $\pm V_{REF}$ or $\pm V_{REF}/2$. Also included with the ML2036 is an inhibit input which allows the sinewave output to be held at zero volts after completing the last half cycle of the sinewave preventing steps in voltage. Two pins of the ML2036 are clock outputs designed to drive other devices with one half or one eighth of the clock input frequency.

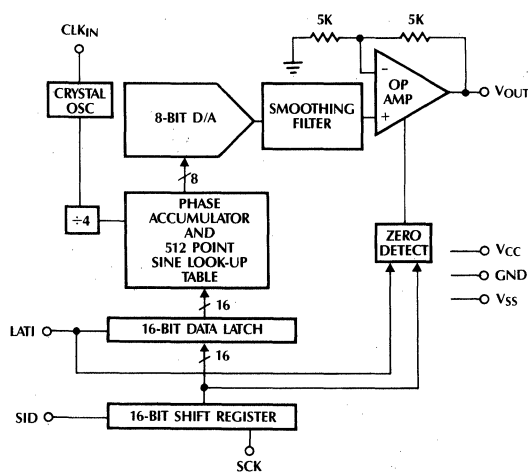
The ML2035 and ML2036 are intended for telecommunications and modem applications that need low cost and accurate generation of precise test tones, call progress tones, and signaling tones.

FEATURES

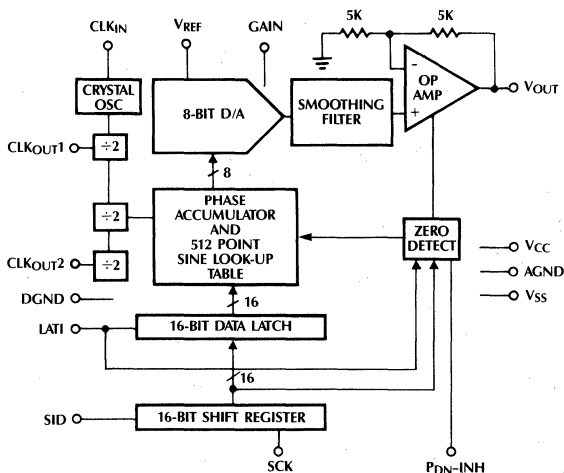
- Programmable frequency DC to 50kHz
- Frequency resolution with $f_{CLKIN} = 12\text{MHz}$ ($\pm 75\text{ Hz}$) 1.5Hz
- Absolute gain error $\pm 1\text{dB max}$
- Harmonic distortion -45dB max
- Output voltage amplitude of $\pm V_{REF}$ or $\pm V_{REF}/2$
- On chip crystal oscillator 3 to 12MHz
- ML2036 has clock outputs of 1/2 or 1/8 of the input clock frequency
- No external components required
- μP compatible serial interface
- Double buffered data latch
- Synchronous or asynchronous data loading capability
- Power dissipation 50mW max from $\pm 5\text{V}$ supplies
- Compatible with ML2031 and ML2032 tone detector, and ML2004 logarithmic gain/attenuator
- TTL/CMOS compatible inputs
- ML2035 package 8-pin DIP; ML2036 14-pin DIP or 16-pin SOIC
- 0°C to $+70^\circ\text{C}$ or -40°C to $+85^\circ\text{C}$ operating temperature range

BLOCK DIAGRAMS

ML2035

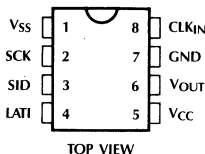


ML2036



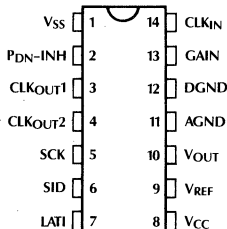
PIN CONNECTIONS

ML2035
8-Pin DIP



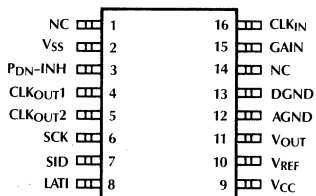
TOP VIEW

ML2036
14-Pin DIP



TOP VIEW

ML2036
16-Pin SOIC



TOP VIEW

PIN DESCRIPTIONS

ML2035

PIN NO.	NAME	FUNCTION
1	V _{SS}	Negative supply. -5V ± 10%.
2	SCK	Serial clock. Digital input which clocks in serial data on rising edges.
3	SID	Serial data. Serial input data which programs the frequency of V _{OUT} .
4	LATI	Serial latch. Digital input which latches serial data into the internal data latch on falling edges.
5	V _{CC}	Positive supply. +5V ± 10%.
6	V _{OUT}	Analog output. V _{OUT} swing is ±V _{CC} /2.
7	GND	Ground. 0 volts. All inputs and outputs referenced to this point.
8	CLK _{IN}	Clock input. Internal clock can be generated by tying a 3 to 12MHz crystal from this pin to GND or applying a clock directly to the pin.

ML2036

PIN NO.	DIP	SOIC	NAME	FUNCTION
1	2		V _{SS}	Negative supply. -5V ± 10%.
2	3		P _{DN} -INH	Three level input. Controls inhibit mode and power down mode. Current source pull up to V _{CC} .
3	4		CLK _{OUT1}	Clock output. Digital output from internal clock generator that can drive other devices. f _{CLKOUT1} = f _{CLKIN} /2.
4	5		CLK _{OUT2}	Clock output. Digital output from internal clock generator that can drive other devices. f _{CLKOUT2} = f _{CLKIN} /8.
5	6		SCK	Serial clock. Digital input which clocks in serial data on rising edges.
6	7		SID	Serial data. Serial input data which programs the frequency of V _{OUT} .
7	8		LATI	Serial latch. Digital input which latches serial data into the internal data latch on falling edges.
8	9		V _{CC}	Positive supply. +5V ± 10%.
9	10		V _{REF}	Reference input. The voltage on this pin determines the peak-peak swing on V _{OUT} . V _{REF} can be tied to V _{CC} .
10	11		V _{OUT}	Analog output.
11	12		AGND	Analog ground. 0 volts. Analog inputs and outputs referenced to this point.
12	13		DGND	Digital ground. 0 volts. Digital inputs and outputs referenced to this point.
13	15		GAIN	Sets V _{OUT} peak amplitude to V _{REF} or V _{REF} /2. Current source pull down to DGND.
14	16		CLK _{IN}	Clock input. Internal clock can be generated by tying a 3 to 12MHz crystal from this pin to DGND or applying a clock directly to the pin.

3

ML2035, ML2036

ABSOLUTE MAXIMUM RATINGS

(Note 1)

Supply Voltage	
V _{CC}	+6.5V
V _{SS}	-6.5V
Analog Input and Output	V _{SS} - 0.3V to V _{CC} + 0.3V
AGND Voltage	V _{SS} to V _{CC}
Digital Inputs and Outputs	-0.3V to V _{CC} + 0.3V
Input Current per Pin	±25mA
Power Dissipation	750mW
Storage Temperature Range	-65°C to +150°C

Lead Temperature (Soldering 10 sec)	
Dual-In-Line Package (Molded)	260°C
Dual-In-Line Package (Ceramic)	300°C
Molded Small Outline IC Package	
Vapor Phase (60 sec)	215°C
Infrared (15 sec)	220°C

OPERATING CONDITIONS

Temperature Range (Note 2)	
ML2035CP, ML2036CP, ML2036CS	0°C to +70°C
ML2035IJ, ML2036IJ	-40°C to +85°C

ML2035 ELECTRICAL CHARACTERISTICS

Unless otherwise specified, T_A = T_{MIN} to T_{MAX}, V_{CC} = 5V ± 10%, V_{SS} = -5V ± 10%, CLK_{IN} = 12.352MHz, V_{OUT} load C_L = 100pF and R_L = 1k, all digital timing measured at 1.4V midpoint, and input control signals from 10% to 90% of V_{CC} with t_R = t_F = 20ns.

SYMBOL	PARAMETER	NOTES	CONDITIONS	MIN	TYP (Note 3)	MAX	UNITS
Sinewave Generator							
HD	Harmonic Distortion	4	2nd or 3rd Harmonic Relative to Fundamental	f _{OUT} = 20Hz to 5kHz		-45	dB
				f _{OUT} = 5kHz to 25kHz		-40	dB
SND	Signal to Noise + Distortion	4	200Hz ≤ f _{OUT} ≤ 3400Hz, noise measured 200Hz to 4kHz			-45	dB
			20Hz ≤ f _{OUT} ≤ 25kHz, noise measured 20Hz to 75kHz			-40	dB
ICN	Output Idle Channel Noise	4	Power Down Mode, Cmsg weighted		-20	0	dB _{rnc}
			Power Down Mode, 1kHz		50		nV/√Hz
PSRR	Power Supply Rejection Ratio	5	200mV _{P-P} , 0-10kHz sine, measured on V _{OUT}	V _{CC}		-40	dB
				V _{SS}		-40	dB
V _{OS}	V _{OUT} Offset Voltage	4				±75	mV
V _{PK}	V _{OUT} Peak Voltage				±V _{CC} /2		V
V _{GN}	V _{OUT} Gain Error	4	Relative to V _{CC}	f _{OUT} = 20Hz to 5kHz		±1	dB
				f _{OUT} = 5kHz to 25kHz		±3	dB

Digital and DC

V _{IL,CLK}	Input Low Voltage, CLK _{IN}	4				1.5	V
V _{IH,CLK}	Input High Voltage, CLK _{IN}	4		3.5			V
I _{IN,CLK}	Input Current, CLK _{IN}	4	CLK _{IN} = 1.5V to 3.5V		10	60	μA
			CLK _{IN} = 0 to 1.5V; 3.5V to V _{CC}			250	μA
C _{IN,CLK}	Input Capacitance, CLK _{IN}	5			12		pF
V _{IL}	Input Low Voltage	4				.8	V
V _{IH}	Input High Voltage	4		2.0			V
I _{IL}	Input Low Current	4	V _{IN} = 0V	-1			μA
I _{IH}	Input High Current	4	V _{IN} = V _{CC}			1	μA
C _{IN}	Digital Input Capacitance				5		pF
V _{OL}	Output Low Voltage	4	I _{OL} = -2mA			0.4	V
V _{OH}	Output High Voltage	4	I _{OH} = 2mA	4.0			V

ML2035 ELECTRICAL CHARACTERISTICS (Continued)

Unless otherwise specified, $T_A = T_{MIN}$ to T_{MAX} , $V_{CC} = 5V \pm 10\%$, $V_{SS} = -5V \pm 10\%$, $CLK_{IN} = 12.352MHz$, V_{OUT} load $C_L = 100pF$ and $R_L = 1k$, all digital timing measured at 1.4V midpoint, and input control signals from 10% to 90% of V_{CC} with $t_R = t_F = 20ns$.

SYMBOL	PARAMETER	NOTES	CONDITIONS	MIN	TYP (Note 3)	MAX	UNITS
Digital and DC (Continued)							
I_{CC}	V_{CC} Supply Current	4	No output load, $V_{CC} = 5.5V$			5.5	mA
I_{SS}	V_{SS} Supply Current	4	No output load, $V_{SS} = -5.5V$, $V_{CC} = 5.5V$			-3.5	mA
I_{CCI}	V_{CC} Supply Current, Power Down Mode	4	No Output Load, Power Down Mode			2.0	mA
I_{SSI}	V_{SS} Supply Current, Power Down Mode	4	No Output Load, Power Down Mode			-100	μA
Digital Timing							
t_{CKI}	CLK_{IN} On/Off Period	4	$t_R = t_F = 10ns$, 2.5V midpoint	30			ns
t_{SCK}	SCK On/Off Period	4		100			ns
t_{DS}	SID DATA Setup Time	4		50			ns
t_{DH}	SID DATA Hold Time	4		50			ns
t_{LPW}	LATI Pulse Width	4		50			ns
t_{LH}	LATI Hold Time	4		50			ns
t_{LS}	LATI Setup Time	5		50			ns

ML2035, ML2036

ML2036 ELECTRICAL CHARACTERISTICS

Unless otherwise specified, $T_A = T_{MIN}$ to T_{MAX} , $V_{CC} = 5V \pm 10\%$, $V_{SS} = -5V \pm 10\%$, $AGND = DGND = 0V$, $V_{REF} = 2.5V$ to V_{CC} and $CLK_{IN} = 12.352MHz$, V_{OUT} load $C_L = 100pF$ and $R_L = 1k$, all digital timing measured at 1.4V midpoint, and input control signals from 10% to 90% of V_{CC} with $t_r = t_f = 20ns$.

SYMBOL	PARAMETER	NOTES	CONDITIONS	MIN	TYP (Note 3)	MAX	UNITS	
Sinewave Generator								
HD	Harmonic Distortion	4, 6	2nd or 3rd harmonic relative to fundamental	$f_{OUT} = 20Hz$ to 5kHz		-45	dB	
				$f_{OUT} = 5kHz$ to 50kHz		-40	dB	
SND	Signal to Noise + Distortion	4, 6	200Hz $\leq f_{OUT} \leq 3400Hz$, noise measured 200Hz to 4kHz			-45	dB	
			20Hz $\leq f_{OUT} \leq 50kHz$, noise measured 20Hz to 150kHz			-40	dB	
ICN	Output Idle Channel Noise	4	Power down mode, Cmsg weighted			-20	0	dBrc
			Power down mode, 1kHz			50		nV/ \sqrt{Hz}
			Inhibit mode, 1kHz			500		nV/ \sqrt{Hz}
PSRR	Power Supply Rejection Ratio	5	200mV _{p-p} , 0 to 10kHz sine, measured on V_{OUT}	V_{CC}		-40	dB	
				V_{SS}		-40	dB	
V _{OS}	V_{OUT} Offset Voltage	4, 7				$\pm 25 + (\pm 10 \times V_{OUTP-P})$	mV	
V _{PK}	V_{OUT} Peak Voltage	6	GAIN = V_{CC}			$\pm V_{REF}$	V	
			GAIN = DGND			$\pm V_{REF}/2$	V	
V _{SW}	V_{OUT} Swing	5	GAIN = V_{CC}		$V_{SS}+1.5V$	$V_{CC}-1.5V$	V	
V _{GN}	V_{OUT} Gain Error	4, 6	$f_{OUT} = 20Hz$ to 5kHz			± 1	dB	
			$f_{OUT} = 5kHz$ to 50kHz			± 3	dB	
R _{REF}	Reference Input Resistance	4		2.5	12		M Ω	
Digital and DC								
V _{IL,CLK}	Input Low Voltage, CLK_{IN}	4				1.5	V	
V _{IH,CLK}	Input High Voltage, CLK_{IN}	4		3.5			V	
I _{IN,CLK}	Input Current, CLK_{IN}	4	$CLK_{IN} = 1.5V$ to 3.5V			10	60	μA
			$CLK_{IN} = 0$ to 1.5V; 3.5V to V_{CC}				250	μA
C _{IN,CLK}	Input Capacitance, CLK_{IN}	5			12		pF	
V _{IL}	Input Low Voltage	4	LATI, SID, GAIN, SCK			.8	V	
V _{IH}	Input High Voltage	4	LATI, SID, GAIN, SCL		2.0		V	
I _{IL}	Input Low Current	4	$V_{IN} = 0V$, LATI, SID, GAIN, SCK		-1		μA	
I _{IH}	Input High Current	4	$V_{IN} = V_{CC}$, LATI, SID, P _{DN} -INH, SCK			1	μA	
I _{IL, P_{DN}}	Input Low Current	4	P _{DN} -INH, $V_{IN} = 0V$		-70	-20	-5	μA
I _{IH, G}	Input High Current	4	GAIN, $V_{IN} = V_{CC}$		5	20	70	μA
V _{I1}	Input Logic Low P _{DN} -INH	4		DGND-.5		.8	V	
V _{I2}	Inhibit State Voltage P _{DN} -INH	4				$V_{SS}+.5$	V	
V _{I3}	Input Logic High P _{DN} -INH	4		2.0			V	
C _{IN}	Digital Input Capacitance				5		pF	
V _{OL}	Output Low Voltage	4	$I_{OL} = -2mA$			0.4	V	
V _{OH}	Output High Voltage	4	$I_{OH} = 2mA$		4.0		V	

ML2036 ELECTRICAL CHARACTERISTICS (Continued)

Unless otherwise specified, $T_A = T_{MIN}$ to T_{MAX} , $V_{CC} = 5V \pm 10\%$, $V_{SS} = -5V \pm 10\%$, $AGND = DGND = 0V$, $V_{REF} = 2.5V$ to V_{CC} , and $CLK_{IN} = 12.352MHz$, V_{OUT} load $C_L = 100pF$ and $R_L = 1k$, all digital timing measured at 1.4V midpoint, and input control signals from 10% to 90% of V_{CC} with $t_R = t_F = 20ns$.

SYMBOL	PARAMETER	NOTES	CONDITIONS	MIN	TYP (Note 3)	MAX	UNITS
Digital and DC (Continued)							
I_{CC}	V_{CC} Supply Current	4	No output load, $V_{CC} = V_{REF} = 5.5V$			5.5	mA
I_{SS}	V_{SS} Supply Current	4	No output load, $V_{SS} = -5.5V$, $V_{CC} = V_{REF} = 5.5V$			-3.5	mA
I_{CCI}	V_{CC} Supply Current, Power Down Mode		No output load, power down mode			2.0	mA
I_{SSI}	V_{SS} Supply Current, Power Down Mode		No output load, power down mode			-100	μA
Digital Timing							
t_{CKI}	CLK_{IN} On/Off Period	4	$t_R = t_F = 10ns$, 2.5V midpoint	30			ns
t_{SCK}	SCK On/Off Period	4		100			ns
t_{DS}	SID DATA Setup Time	4		50			ns
t_{DH}	SID DATA Hold Time	4		50			ns
t_{LPW}	LATI Pulse Width	4		50			ns
t_{LH}	LATI Hold Time	4		50			ns
t_{LS}	LATI Setup Time	5		50			ns
Clock Output							
f_{CLK1}	CLK_{OUT1} Output Frequency	4	Figure 2	$\frac{1}{2}$		$\frac{1}{2}$	f_{CLKIN}
f_{CLK2}	CLK_{OUT2} Output Frequency	4	Figure 2	$\frac{1}{8}$		$\frac{1}{8}$	f_{CLKIN}
t_{1R}, t_{2R}	CLK_{OUT1}, CLK_{OUT2} Output Rise Time	5	$C_L = 40pF$, 10% and 90% transition point	0		20	ns
		4	$C_L = 100pF$, 0.8V and 2.0V transition point	0		20	ns
t_{1F}, t_{2F}	CLK_{OUT1}, CLK_{OUT2} Output Fall Time	5	$C_L = 40pF$, 10% and 90% transition point	0		20	ns
		4	$C_L = 100pF$, 0.8V and 2.0V transition point	0		20	ns

Note 1: Absolute maximum ratings are limits beyond which the life of the integrated circuit may be impaired. All voltages unless otherwise specified are measured with respect to ground.

Note 2: 0°C to +70°C and -40°C to +85°C operating temperature range devices are 100% tested with temperature limits guaranteed by 100% testing, sampling, or by correlation with worst-case test conditions.

Note 3: Typicals are parametric norm at 25°C.

Note 4: Parameter guaranteed and 100% production tested.

Note 5: Parameter guaranteed. Parameters not 100% tested are not in outgoing quality level calculation.

Note 6: Maximum peak-to-peak voltage for output sinewave is $V_{OUT-P} \leq (125kV \times Hz) / f_{OUT}$. For example at 50kHz output the maximum guaranteed voltage swing is 2.5V_{p-p}.

Note 7: Offset voltage is a function of the peak-to-peak output voltage, for example if $V_{OUT-P} = 2.5V$, $V_{OS} = \pm 50mV$ max.

TIMING DIAGRAMS

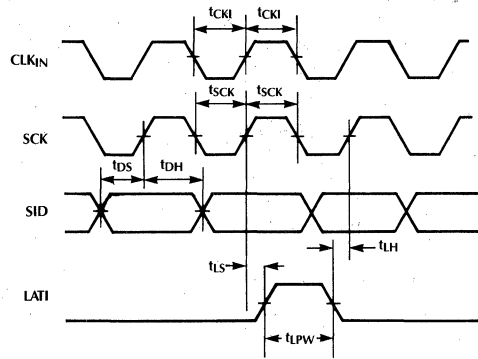
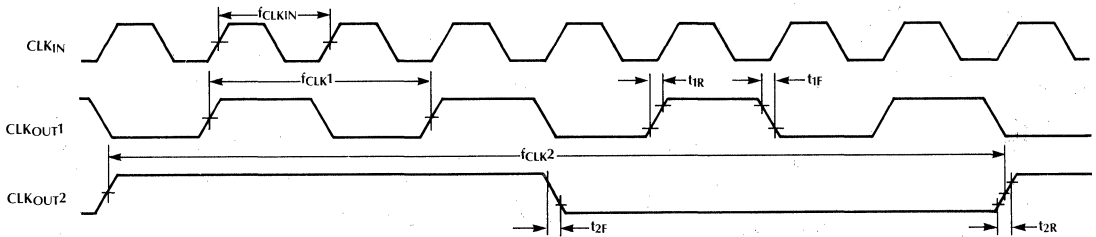


Figure 1. Serial Interface Timing



f_{CLK} PARAMETERS REFERRED TO 1.4V MIDPOINT

Figure 2. ML2036 Digital Clock Output Timing

TYPICAL PERFORMANCE

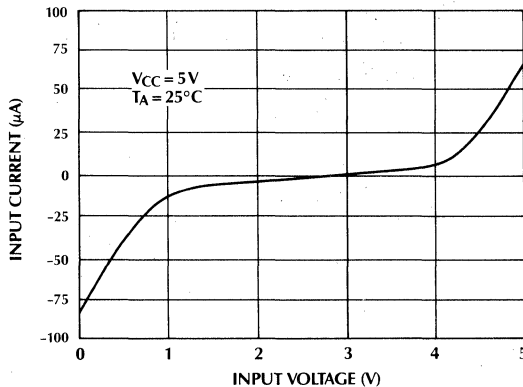


Figure 3. CLK_{IN} Input Current vs. Input Voltage

1.0 FUNCTIONAL DESCRIPTION

The ML2035 and ML2036 are composed of a programmable frequency generator, sinewave generator, crystal oscillator, and serial digital interface. The ML2035

and ML2036 frequency and sinewave generator functional block diagram is shown in figure 4.

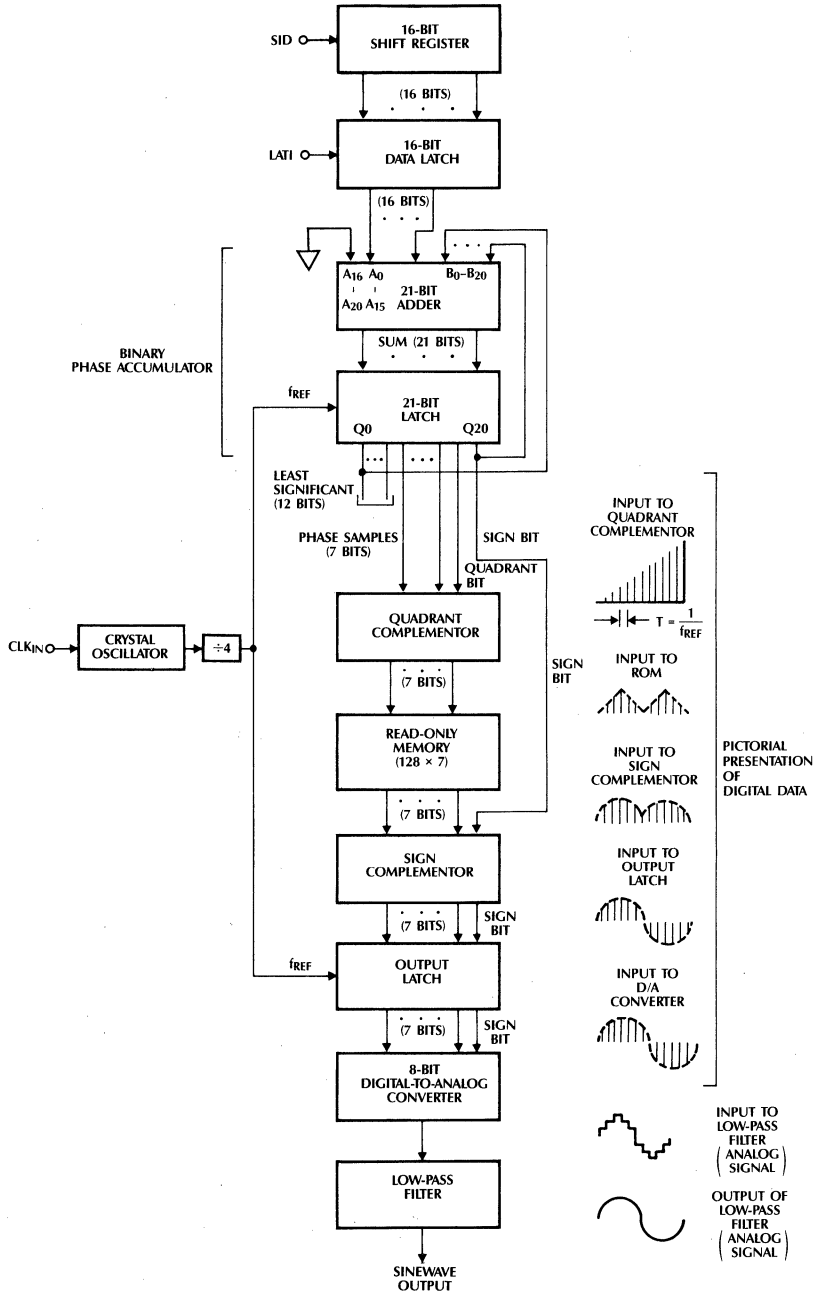


Figure 4. Frequency and Sinewave Generator Functional Block Diagram

ML2035, ML2036

1.1 Programmable Frequency Generator

The programmable frequency generator produces a digital output whose frequency is determined by a 16-bit digital word.

The frequency generator is composed of a phase accumulator which is clocked at $f_{CLKIN}/4$. The value stored in the data latch is added to the phase accumulator every 4 cycles of CLK_{IN} . The frequency of the analog output is equal to the rate at which the accumulator overflows and is given by the following equation:

$$f_{OUT} = \frac{f_{CLKIN} \times (D15-D0)_{DEC}}{2^{23}}$$

The frequency resolution and the minimum frequency are the same and is given by the following equation:

$$\Delta f_{MIN} = \frac{f_{CLKIN}}{2^{23}}$$

When $f_{CLKIN} = 12.352\text{MHz}$, $\Delta f_{MIN} = 1.5\text{Hz}$ ($\pm 0.75\text{Hz}$). Lower frequencies are obtained by using a lower clock.

Due to the phase quantization nature of the frequency generator spurious tones can be present in the output in the range of -55dB relative to fundamental. The energy from these tones is included in the signal to noise + distortion specification. The frequency of these tones can be very close to the fundamental, therefore it is not practical to filter them out.

1.2 Sinewave Generator

The sinewave generator is composed of a sine look-up table, a DAC, and an output smoothing filter. The sine look-up table is addressed by the phase accumulator. The DAC is driven by the output of the look-up table and generates a staircase representation of a sinewave.

The output smoothing filter "smooths" the analog output by removing the high frequency sampling components. The resultant voltage on V_{OUT} is a sinusoid with all distortion components at least 45dB below the fundamental.

The ML2035 provides a peak sinewave voltage of $\pm V_{CC}/2$. The ML2036 has a V_{REF} input that can be tied to V_{CC} or generated from an external voltage. With the gain input equal to a logic "1" the sinewave peak voltage is equal to $\pm V_{REF}$; with the gain input equal to a logic "0" the peak voltage is $\pm V_{REF}/2$. The sinewave output is referenced to AGND for the ML2036 and GND for the ML2035.

The analog section is designed to operate over a range from DC to 50kHz. Due to slew rate limitations, the peak-to-peak output voltage must be limited to $V_{OUTP-P} \leq (125\text{kV} \times \text{Hz})/f_{OUT}$. For example on the ML2036 an output at 50kHz must be limited to $2.5V_{P-P}$. Since the ML2035 peak-to-peak output voltage is equal to V_{CC} the maximum output frequency must be limited to 25kHz for $V_{CC} = 5\text{V}$. V_{OUT} can drive $1\text{k}\Omega$, 100pF loads and swing to within 1.5V of V_{CC} and V_{SS} , provided the slew rate limitations mentioned above are not exceeded.

The output offset voltage, V_{OS} , is a function of the peak-to-peak output voltage and is specified as $25\text{mV} + (\pm 10 \times V_{OUTP-P})$ max. For example if $V_{OUTP-P} = 2.5\text{V}$, then $V_{OS} = 50\text{mV}$ max.

1.3 Crystal Oscillator

The crystal oscillator generates an accurate reference clock for the programmable frequency generator.

The internal clock can be generated with a crystal or external clock.

If a crystal is used, it must be placed between CLK_{IN} and DGND of the ML2036 or GND of the ML2035. An on chip crystal oscillator will then generate the internal clock. No other external capacitors or components are required. The crystal should be a parallel resonant type with a frequency between 3MHz to 12.4MHz. It should be placed physically as close as possible to the CLK_{IN} and DGND (GND).

An external clock can drive CLK_{IN} directly if desired. The frequency of this clock can be anything from 0 to 12MHz.

The crystal must have the following characteristics:

1. Parallel resonant type
2. Frequency: 3MHz to 12.4MHz
3. Maximum equivalent series resistance of 15Ω at a drive level of $1\mu\text{W}$ to $200\mu\text{W}$
4. Maximum equivalent series resistance of 30Ω at drive levels of 10nW to $1\mu\text{W}$
5. Typical load capacitance: 18pF
6. Maximum case capacitance: 7pF

The frequency of oscillation will be a function of the crystal parameters and board capacitance. Crystals that meet these requirements at 12.352000MHz are M-tron 3709-010 12.352 for 0°C to $+70^\circ\text{C}$ and 3709-020 12.352 for -40°C to $+85^\circ\text{C}$ operation.

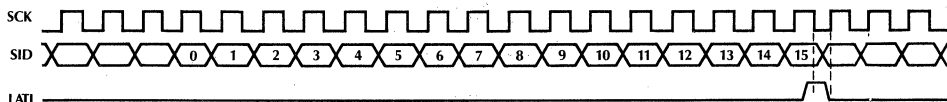


Figure 5. Serial Interface Timing

The ML2036 has two clock outputs that can be used to drive other external devices. The CLK_{OUT1} output is a buffered output from the oscillator divided by 2. The CLK_{OUT2} output is a buffered output from the oscillator divided by 8.

1.4 Serial Digital Interface

The digital interface consists of a shift register and data latch. The serial 16-bit data word on SID is clocked into a 16-bit shift register on rising edges of the serial shift clock, SCK. The LSB should be shifted in first and the MSB last as shown in figure 5. The data that has been shifted into the shift register is loaded into a 16-bit data latch on the falling edge of LATI. To insure that true data is loaded into the data latch from the shift register, LATI falling edge should occur when SCK is low, as shown in figure 1. LATI should be low while shifting data into the shift register to avoid inadvertently entering the power down mode as described in paragraph 1.5. Note that all data is entered and latched on edges, not levels, of SCK and LATI.

1.5 Inhibit and Power Down Modes

1.5.1 ML2035 Power Down Mode

The power down mode of the ML2035 can be selected by entering all zeros in the shift register and applying a logic "1" to LATI. A zero data detect circuit detects when all bits in the shift register are zero's. In this state, the power consumption is reduced to 11.5mW max, and V_{OUT} goes to 0V as shown in figure 6 and appears as 10k to analog ground. The master clock, CLK_{IN}, can be left active or removed during power down mode.

1.5.2 ML2036 Inhibit and Power Down Modes

The ML2036 has an inhibit mode and a power down mode which are controlled by the three-level P_{DN}-INH input as described in table 1. When a logic "1", V_{I3}, is applied to the P_{DN}-INH pin, the power down mode is entered in the same way as described for the ML2035. Also, the ML2036 will be placed in the power down mode by applying a logic "0" to the P_{DN}-INH pin.

If V_{SS} to V_{SS} + .5V, V_{I2}, is applied to the P_{DN}-INH pin, the inhibit mode is entered by shifting all zero's into the shift register and applying a logic "1" to the LATI pin. Once the inhibit mode is entered V_{OUT} will complete the last half cycle of the sinewave and then be held at approximately V_{OS}, such that no voltage step occurs, as shown in figure 6.

P _{DN} -INH MODE	P _{DN} -INH PIN	DATA IN SHIFT REG.	LATI	SINEWAVE OUTPUT
P _{DN} ⁽¹⁾	V _{I1} , Logic "0"	X	X	V _{OUT} = 0V (10K to AGND)
Inhibit	V _{I2} , Inhibit State Voltage, V _{SS} to V _{SS} + .5V	All 0's	Logic "1"	V _{OUT} goes to approximately V _{OS} at the next V _{OS} crossing. See figure 6.
P _{DN} ⁽¹⁾	V _{I3} , Logic "1"	All 0's	Logic "1"	V _{OUT} = 0V (10K to AGND)

Note 1: In the power down mode, the oscillator, CLK_{OUT1} and CLK_{OUT2}, shift register, and data latch are all functional.

Table 1. Three Level P_{DN}-INH Function

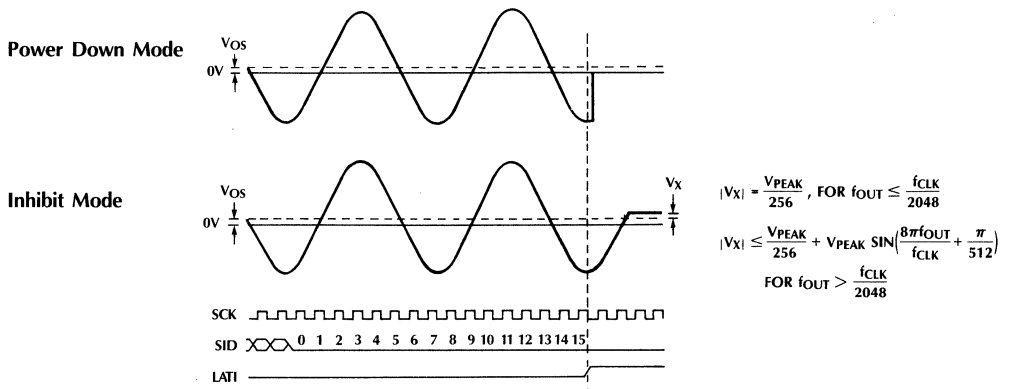


Figure 6. Power Down and Inhibit Mode

ML2035, ML2036

1.6 Power Supplies

The analog circuitry in the device are powered from +5V to -5V (V_{CC} to V_{SS}) and are referenced to AGND.

The digital circuits in the device are powered from +5V to 0V (V_{CC} to DGND).

For the ML2036, it is recommended that AGND and DGND be connected together close to the device and have a good connection back to the power source.

It is recommended that the power supplies to the device should be bypassed by placing decoupling capacitors from V_{CC} to AGND (GND for ML2035) and V_{SS} to AGND (GND for ML2035) as physically close to the device as possible.

2.0 TYPICAL APPLICATIONS

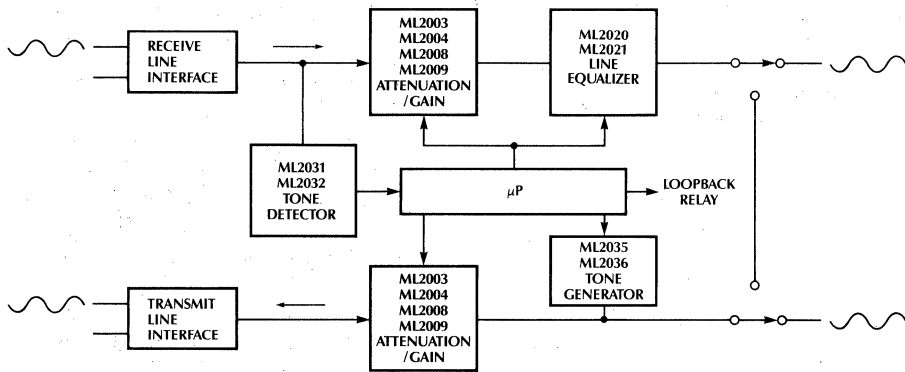


Figure 7. 4-Wire Termination Equipment

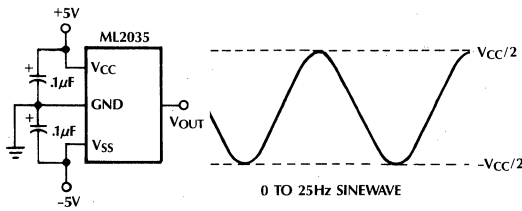


Figure 8. Sinewave Ratiometric to $\pm V_{CC}/2$

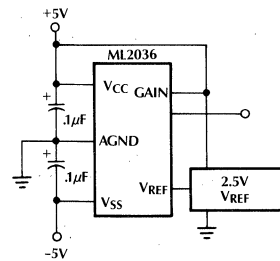


Figure 9. Sinewave with $\pm 2.5V_{P-P}$ ($5V_{P-P}$)

ORDERING INFORMATION

PART NUMBER	TEMP. RANGE	PACKAGE
ML2035IJ	-40°C to +85°C	HERMETIC DIP (J08)
ML2035IP	-40°C to +85°C	MOLDED DIP (P08)
ML2035CP	0°C to +70°C	MOLDED DIP (P08)
ML2036IJ	-40°C to +85°C	HERMETIC DIP (J14)
ML2036IP	-40°C to +85°C	MOLDED DIP (P14)
ML2036IS	-40°C to +85°C	MOLDED SOIC (S16W)
ML2036CP	0°C to +70°C	MOLDED DIP (P14)
ML2036CS	0°C to +70°C	MOLDED SOIC (S16W)

Universal Dual Filter

GENERAL DESCRIPTION

The ML2110 consists of two independent switched capacitor filters that perform second order filter functions such as lowpass, bandpass, highpass, notch and allpass. All filter configurations including Butterworth, Bessel, Cauer, and Chebyshev can be formed.

The center frequency of these filters is tuned by an external clock or the external clock and resistor ratio.

The ML2110 frequency range is specified to 30kHz with $\pm 2.25V$ (single 5V operation) to $\pm 5.5V$ power supplies.

For higher frequency operation the ML2111 is specified up to 150kHz operation.

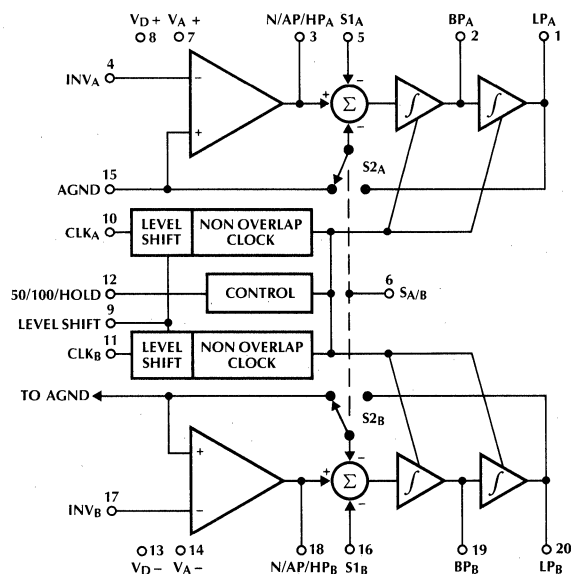
These filters are ideal where center frequency accuracy and high Q_s are needed.

The ML2110 is a pin compatible superior replacement for MF10, LMF100, and LTC1060 filters.

FEATURES

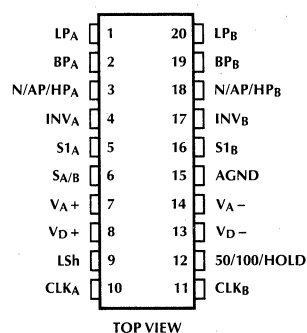
- Specified to 30kHz
- Center frequency $\times Q$ product $\leq 2MHz$
- Separate highpass, notch, allpass, bandpass, and lowpass outputs
- Center frequency accuracy $\pm 0.3\%$ or $\pm 0.8\%$ max
- Q accuracy $\pm 3\%$ or $\pm 6\%$ max
- Clock inputs TTL or CMOS compatible with duty cycle 40% to 60%
- Single 5V ($\pm 2.25V$) or $\pm 5V$ supply operation
- $0^\circ C$ to $70^\circ C$, $-40^\circ C$ to $+85^\circ C$, $-55^\circ C$ to $+125^\circ C$ operating temperature range
- Standard 0.3" 20-pin DIP or 20-pin small outline (SOIC) package

BLOCK DIAGRAM

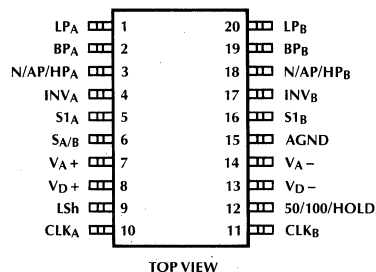


PIN CONNECTIONS

ML2110
20-PIN DIP



ML2110
20-PIN SOIC



PIN DESCRIPTION

PIN NO.	NAME	FUNCTION
1	LP _A	Lowpass output for biquad A.
2	BP _A	Bandpass output for biquad A.
3	N/AP/HP _A	Notch/allpass/highpass output for biquad A.
4	INV _A	Inverting input of the summing op amp for biquad A.
5	S1 _A	Auxiliary signal input pin used in modes 1a, 1d, 4, 5, and 6b.
6	S _{A/B}	Controls S2 input function.
7	V _{A+}	Positive analog supply.
8	V _{D+}	Positive digital supply.
9	LSh	Reference point for clock input levels. Logic threshold typically 1.4V above LSh voltage.
10	CLK _A	Clock input for biquad A.
11	CLK _B	Clock input for biquad B.

PIN NO.	NAME	FUNCTION
12	50/100/HOLD	Input pin to control the clock to center frequency ratio of 50:1 or 100:1, or stops the clock to hold the last sample of the bandpass or lowpass outputs.
13	V _{D-}	Negative digital supply.
14	V _{A-}	Negative analog supply.
15	AGND	Analog ground.
16	S1 _B	Auxiliary signal input used in modes 1a, 1d, 4, 5, and 6b.
17	INV _B	Inverting input of the summing op amp for biquad B.
18	N/AP/HP _B	Notch/allpass/highpass output for biquad B.
19	BP _B	Bandpass output for biquad B.
20	LP _B	Lowpass output for biquad B.

3

ABSOLUTE MAXIMUM RATINGS
(Note 1)

Supply Voltage	
V _{A+} , V _{D+} - V _{A-} , V _{D-}	13V
V _{A+} , V _{D+} to LSh	13V
Inputs	V _{A+} , V _{D+} + 0.3V to V _{A-} , V _{D-} - 0.3V
Outputs	V _{A+} , V _{D+} + 0.3V to V _{A-} , V _{D-} - 0.3V
V _{A+} to V _{D+}	± 0.3V
Power Dissipation	750mW
Storage Temperature Range	-65°C to 150°C
Lead Temperature (soldering, 10 sec)	300°C

OPERATING CONDITIONS

Temperature Range (Note 2)	
ML2110BCP, ML2110CCP	0°C to 70°C
ML2110BCS, ML2110CCS	-40°C to +85°C
ML2110BIJ, ML2110CIJ	-55°C to +125°C
ML2110BMJ, ML2110CMJ	± 2.25V to ± 6.0V
Supply Voltage Range	

ELECTRICAL CHARACTERISTICS

Unless otherwise specified T_A = T_{MIN} to T_{MAX}. V_{A+} = V_{D+} = 5V ± 10%, V_{A-} = V_{D-} = -5V ± 10%, C_L = 25pF, V_{IN} = 2.5V_{PK} (1.767 V_{RMS}) Clock Duty Cycle 40% to 60%.

PARAMETER	NOTES	CONDITIONS	ML2110B			ML2110C			UNITS
			MIN	TYP NOTE 3	MAX	MIN	TYP NOTE 3	MAX	
Filter									
f _o , Center Frequency Maximum	5, 6	Figure 16 (Mode 1) Q ≤ 50, Q Accuracy ≤ ± 20% Q ≤ 20, Q Accuracy ≤ ± 10%			20 30			20 30	kHz kHz
f _o , Center Frequency Minimum	5, 6	Figure 16 (Mode 1) Q ≤ 50, Q Accuracy ≤ ± 30% Q ≤ 20, Q Accuracy ≤ ± 15%	25 25			25 25			Hz Hz
f _o , Temperature Coefficient		f _{CLK} < 1MHz		-10			-10		ppm/°C
Clock to Center Frequency Ratio		Q = 10 Figure 16 (Mode 1)							
	4	50:1, f _{CLK} = 250kHz	49.85	50.0	50.15	49.60	50.0	50.40	
	4	100:1, f _{CLK} = 500kHz	100.0	100.3	100.6	99.50	100.3	101.1	

ML2110

ELECTRICAL CHARACTERISTICS (Continued)

Unless otherwise specified $T_A = T_{MIN}$ to T_{MAX} , $V_{A+} = V_{D+} = 5V \pm 10\%$, $V_{A-} = V_{D-} = -5V \pm 10\%$, $C_L = 25pF$, $V_{IN} = 2.5V_{PK}$ (1.767 V_{RMS}) Clock Duty Cycle 40% to 60%.

PARAMETER	NOTES	CONDITIONS	ML2110B			ML2110C			UNITS
			MIN	TYP NOTE 3	MAX	MIN	TYP NOTE 3	MAX	
Filter (Continued)									
Clock Frequency	5	$Q \leq 20$, Q Accuracy $\leq \pm 15\%$	2.5k		1.5M	2.5k		1.5M	Hz
Clock Feedthrough	5	$f_{CLK} \leq 1MHz$		10	20		10	20	mV(p-p)
Q Accuracy	4	$f_o = 5kHz$, $Q = 10$ Figure 16 (Mode 1)	50:1		± 3			± 6	%
			100:1		± 4		± 8	%	
Q Temperature Coefficient		$f_{CLK} < 500kHz$, $Q = 10$		20			20		ppm/°C
DC Offset V _{OS 2,3}	4	50:1, $f_{CLK} = 250kHz$ S _{A/B} High		7	40		7	60	mV
	4	S _{A/B} Low		7	40		7	60	mV
DC Offset V _{OS 2,3}	4	100:1, $f_{CLK} = 500kHz$ S _{A/B} High		14	60		14	100	mV
	4	S _{A/B} Low		14	60		14	100	mV
Gain Accuracy DC Lowpass	4	R1 = 20k, R2 = 2k, R3 = 20k		0.01	2		0.01	2	%
Bandpass at f_o	4	100:1, $f_o = 5kHz$, $Q = 10$		1	4		1	8	%
DC Notch Output	5			0.02	2		0.02	2	%
Noise	7	Figure 16 (Mode 1) Q = 1, R1 = R2 = R3 = 2k Bandpass, 5kHz, 50:1 5kHz, 100:1 Lowpass, 5kHz, 50:1 5kHz, 100:1 Notch, 5kHz, 50:1 5kHz, 100:1		80		80		μV_{RMS}	
				100		100		μV_{RMS}	
				105		105		μV_{RMS}	
				130		130		μV_{RMS}	
				80		80		μV_{RMS}	
				100		100		μV_{RMS}	
		Figure 16 (Mode 1) Q = 10, R1 = R3 = 20k, R2 = 2k Bandpass, 5kHz, 50:1 5kHz, 100:1 Lowpass, 5kHz, 50:1 (R1 = 2k) 5kHz, 100:1 Notch, 5kHz, 50:1 (R1 = 2k) 5kHz, 100:1		256		256		μV_{RMS}	
				315		315		μV_{RMS}	
				262		262		μV_{RMS}	
				320		320		μV_{RMS}	
				33		33		μV_{RMS}	
				38		38		μV_{RMS}	
Crosstalk		$f_{CLK} = 250kHz$, $f_o = 5kHz$		-70			-70		dB
Filter, $V_{A+} = V_{D+} = 2.25V$, $V_{A-} = V_{D-} = -2.25V$, $V_{IN} = 0.707 V_{PK}$ (0.5V_{RMS}) (NOTE 8)									
f_o , Center Frequency Maximum	5	Figure 16 (Mode 1) Q ≤ 50 , Q Accuracy $\leq \pm 25\%$ Q ≤ 20 , Q Accuracy $\leq \pm 12\%$			20			20	kHz
					30		30	kHz	
f_o , Center Frequency Minimum	5	Figure 16 (Mode 1) Q ≤ 50 , Q Accuracy $\leq \pm 30\%$ Q ≤ 20 , Q Accuracy $\leq \pm 15\%$	25			25			Hz
Clock to Center Frequency Ratio	4	Q = 10 Figure 16 (Mode 1) 50:1, $f_{CLK} = 250kHz$	49.85	50.0	50.15	49.60	50.0	50.40	
			5	100:1, $f_{CLK} = 500kHz$	100.0	100.3	100.6	99.50	100.3
Clock Frequency	5	Q ≤ 20 , Q Accuracy $\leq \pm 15\%$	2.5k		1.5M	2.5k		1.5M	Hz
Q Accuracy	4	$f_{CLK} = 250kHz$, $Q = 10$ Figure 16 (Mode 1)	50:1		± 4			± 8	%
			100:1		± 3		± 6	%	

ELECTRICAL CHARACTERISTICS (Continued)

Unless otherwise specified $T_A = T_{MIN}$ to T_{MAX} , $V_A + = V_{D+} = 5V \pm 10\%$, $V_A - = V_{D-} = -5V \pm 10\%$, $C_L = 25pF$, $V_{IN} = 2.5V_{PK}$ (1.767 V_{RMS}) Clock Duty Cycle 40% to 60%.

PARAMETER	NOTES	CONDITIONS	ML2110B			ML2110C			UNITS
			MIN	TYP NOTE 3	MAX	MIN	TYP NOTE 3	MAX	
Filter, $V_A + = V_{D+} = 2.25V$, $V_A - = V_{D-} = -2.25V$, $V_{IN} = 0.707 V_{PK}$ (0.5 V_{RMS}) (Continued)									
Noise	7	Figure 16 (Mode 1) $Q = 1, R1 = R2 = R3 = 2k$							
		Bandpass, 5kHz, 50:1		80			80		μV_{RMS}
		5kHz, 100:1		100			100		μV_{RMS}
		Lowpass, 5kHz, 50:1		105			105		μV_{RMS}
		5kHz, 100:1		130			130		μV_{RMS}
		Notch, 5kHz, 50:1		80			80		μV_{RMS}
		5kHz, 100:1		100			100		μV_{RMS}
		Figure 16 (Mode 1) $Q = 10, R1 = R3 = 20k, R2 = 2k$							
		Bandpass, 5kHz, 50:1		256			256		μV_{RMS}
		5kHz, 100:1		315			315		μV_{RMS}
		Lowpass, 5kHz, 50:1		262			262		μV_{RMS}
		($R1 = 2k$) 5kHz, 100:1		320			320		μV_{RMS}
		Notch, 5kHz, 50:1		33			33		μV_{RMS}
		($R1 = 2k$) 5kHz, 100:1		38			38		μV_{RMS}

Operational Amplifiers and Power Supply

V_{OS} DC Offset	4			2	15		2	15	mV
DC Open Loop Gain		$R_L = 1k$		95			95		dB
Gain Bandwidth Product				2.4			2.4		MHz
Slew Rate				2.0			2.0		V/ μs
Output Voltage Swing (Clipping Level)	5	$R_L = 2k$, $ V $ from V_{A+} or V_{A-}		.5	1.2		.5	1.2	V
Output Short Circuit Current		Source		50			50		mA
		Sink		25			25		mA

Power Supply And Clock

Supply Current ($I_{A+} + I_{D+}$)	4	$f_{CLK} = 250kHz$		13	22		13	22	mA
($I_{A-} + I_{D-}$)				12	21		12	21	mA
I_{LSH}				0.5	1		0.5	1	mA
V_{CLK} Input Threshold	4	Low High	2.0		0.8		2.0	0.8	V
CLK_A , CLK_B Pulse Width	5	CLK High or CLK Low	250				250		ns

Note 1: Absolute maximum ratings are limits beyond which the life of the integrated circuit may be impaired. All voltages unless otherwise specified are measured with respect to ground.

Note 2: $-55^{\circ}C$ to $+125^{\circ}C$ operating temperature range devices are 100% tested at temperature extremes with worst-case test conditions. $0^{\circ}C$ to $70^{\circ}C$ and $-40^{\circ}C$ to $+85^{\circ}C$ operating temperature range devices are 100% tested with temperature limits guaranteed by 100% testing, sampling, or by correlation with worst-case test conditions.

Note 3: Typicals are parametric norm at $25^{\circ}C$.

Note 4: Parameter guaranteed and 100% production tested.

Note 5: Parameter guaranteed. Parameters not 100% tested are not in outgoing quality level calculation.

Note 6: Center frequency is defined as the peak of the bandpass output.

Note 7: The noise is measured with the HP8903A audio analyzer with a bandwidth of 30kHz which is 6 times the f_o at 50:1 or at 100:1.

Note 8: For $T_A = -55^{\circ}C$ to $+125^{\circ}C$; $V_{A+} = V_{D+} = 2.375V$, $V_{A-} = V_{D-} = -2.375V$

TYPICAL PERFORMANCE CURVES

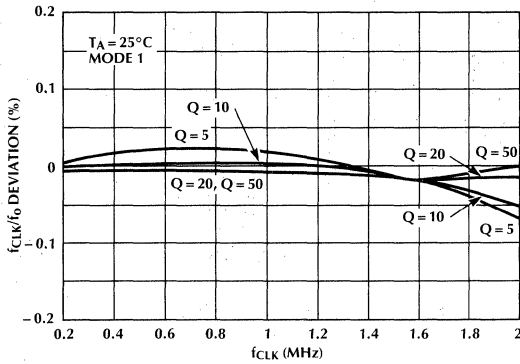


Figure 1. f_{CLK}/f_0 vs. f_{CLK} (100:1, 50:1 at $V_S = \pm 2.5V$ or $V_S = \pm 5V$)

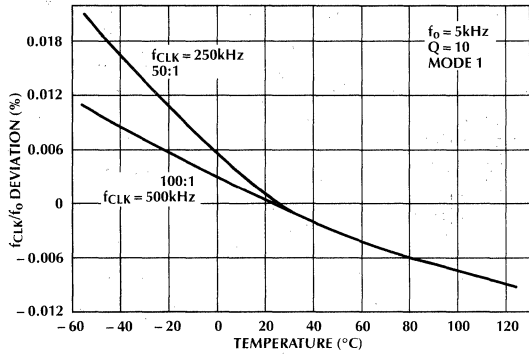
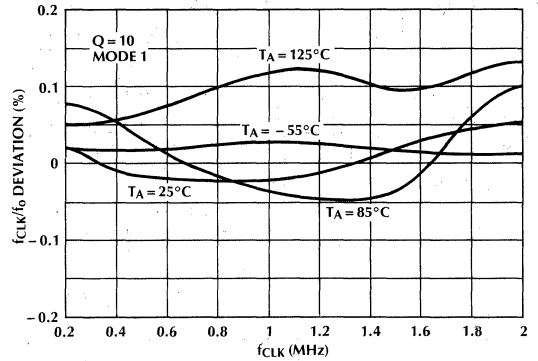


Figure 2A. f_{CLK}/f_0 Deviation vs. Temperature ($V_S = \pm 5V$)

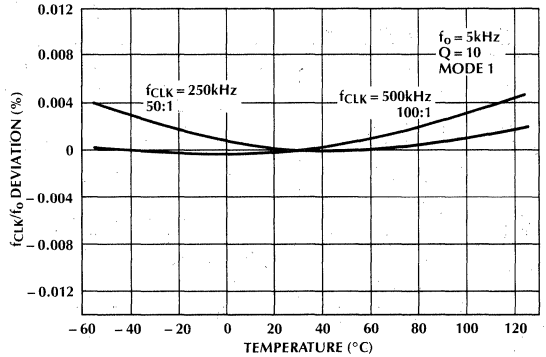


Figure 2B. f_{CLK}/f_0 Deviation vs. Temperature ($V_S = \pm 2.5V$)

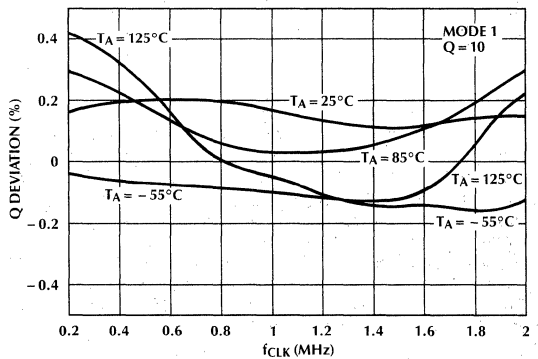
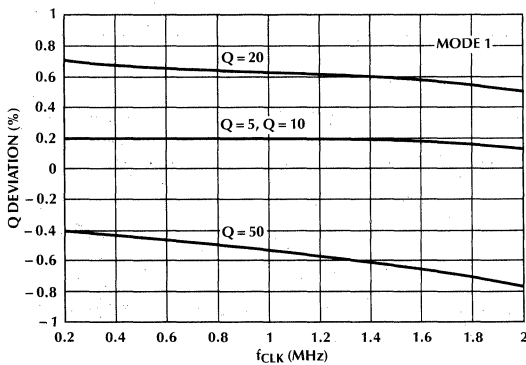


Figure 3. Q Error vs. f_{CLK} (100:1, 50:1 at $V_S = \pm 2.5V$ or $V_S = \pm 5V$)

TYPICAL PERFORMANCE CURVES (Continued)

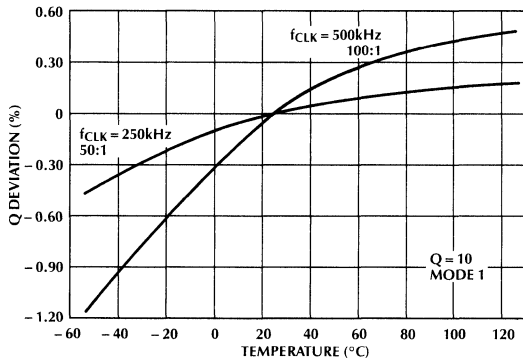


Figure 4A. Q Deviation vs. Temperature ($V_S = \pm 5V$)

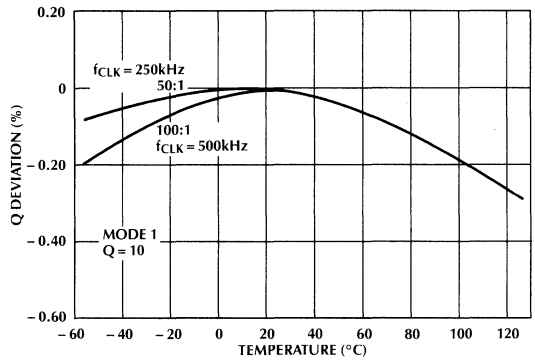


Figure 4B. Q Deviation vs. Temperature ($V_S = \pm 2.5V$)

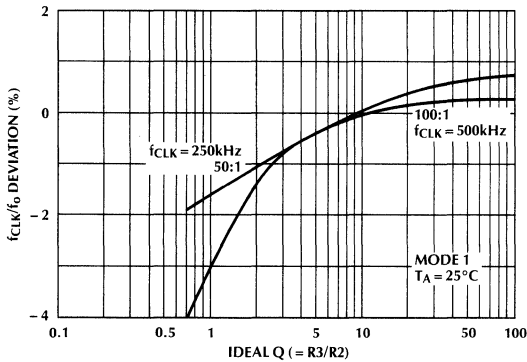


Figure 5A. f_{CLK}/f_0 Deviation vs. Q ($V_S = \pm 5V$)

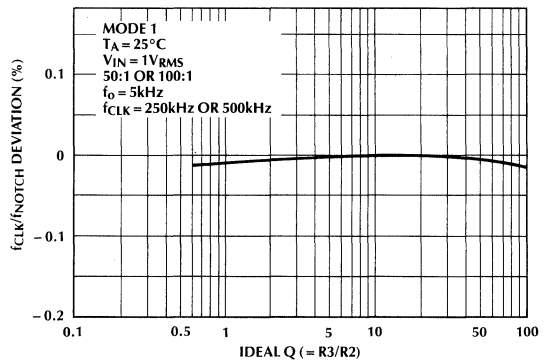


Figure 5B. f_{CLK}/f_{NOTCH} Deviation vs. Q ($V_S = \pm 5V$)

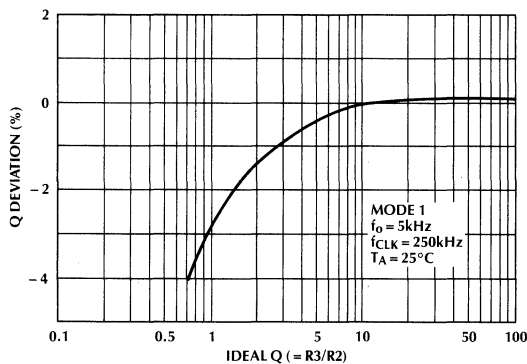


Figure 6A. Q Deviation vs. Q (50:1, $V_S = \pm 5V$)

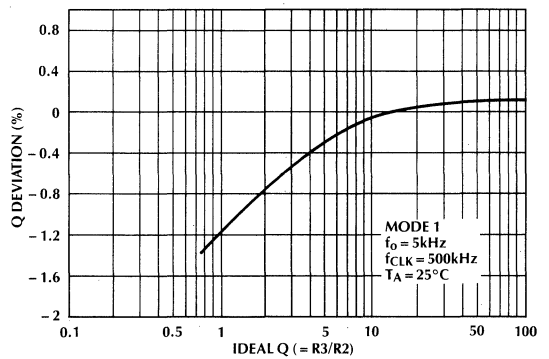


Figure 6B. Q Deviation vs. Q (100:1, $V_S = \pm 5V$)

TYPICAL PERFORMANCE CURVES (Continued)

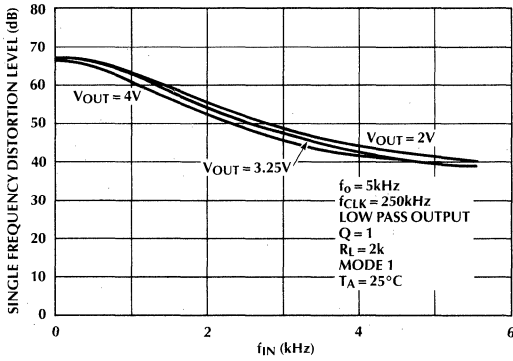


Figure 7A. Distortion vs. f_{IN} (50:1, $V_S = \pm 5V$)

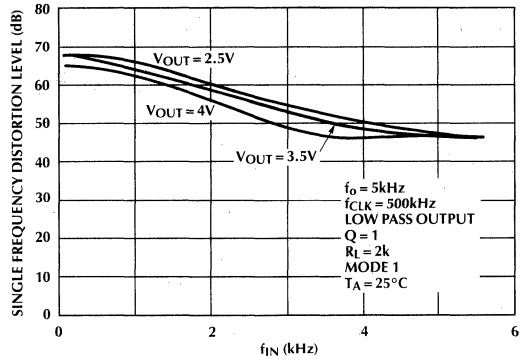


Figure 7B. Distortion vs. f_{IN} (100:1, $V_S = \pm 5V$)

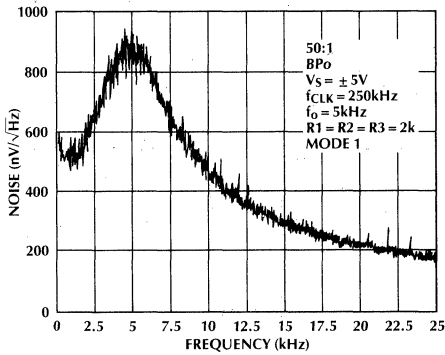


Figure 8A. Noise Spectrum Density ($Q = 1$)

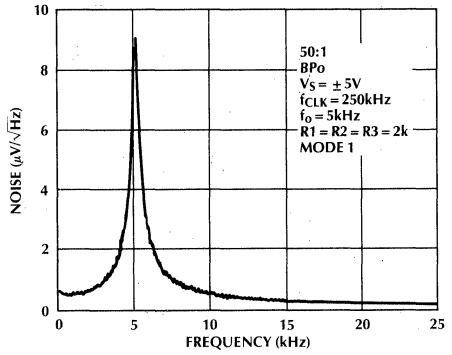


Figure 8B. Noise Spectrum Density ($Q = 10$)

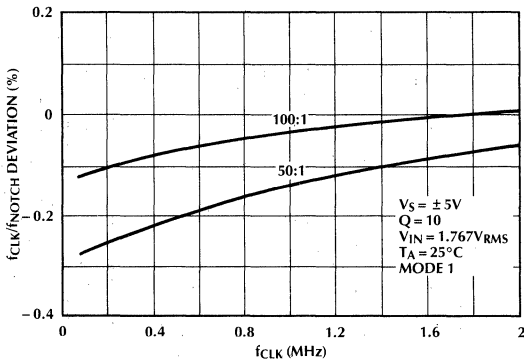


Figure 9. f_{CLK}/f_{NOTCH} vs. f_{CLK}

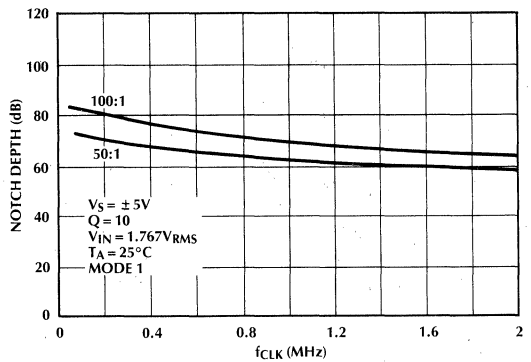


Figure 10. Notch Depth vs f_{CLK}

TYPICAL PERFORMANCE CURVES (Continued)

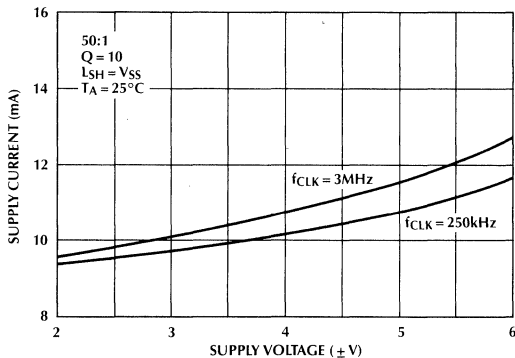


Figure 11. Supply Current vs. Supply Voltage

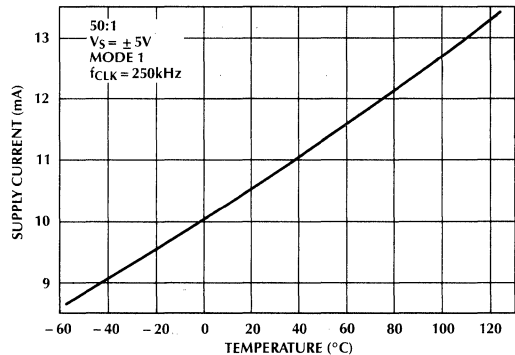


Figure 12. Supply Current vs. Temperature

FUNCTIONAL DESCRIPTION

POWER SUPPLIES

The analog (V_{A+}) and digital (V_{D+}) supply voltage pins, in most cases, are tied together and bypassed to AGND with a $0.1\mu\text{F}$ and a $0.01\mu\text{F}$ disc ceramic capacitor. If high digital noise exists, the supply pins can be bypassed separately. The ML2110 positive analog and positive digital supply pins are internally connected by the IC substrate and should be biased from the same DC source.

The ML2110 operates with a single supply from $5\text{V} \pm 10\%$ and with split supplies from $\pm 4.5\text{V}$ to $\pm 6\text{V}$ supplies.

CLOCK INPUT PINS AND LEVEL SHIFT

With dual supplies equal to or higher than $\pm 4.0\text{V}$, the level shift (LSH) pin 9 can be connected to the same potential as the AGND or V_{A-} pin. With single supply operation, the negative supply pins and the LSH pin should be tied to the system ground. The AGND, pin 15, should be biased at $1/2$ supplies. Under these conditions, the clock levels are TTL or CMOS. The input clock pins (10, 11) share the same level shift pin.

50/100/HOLD (Pin 12)

By tying pin 12 to (V_{A+} , V_{D+}) the filter operates in the 50:1 mode. By tying pin 12 to $1/2$ of the voltage supplies (AGND potential), the ML2110 operates in the 100:1 mode. The range of pin 12 with total supply voltage of $+5\text{V}$ is $2.5 \pm 0.5\text{V}$; $+10\text{V}$ is $5\text{V} \pm 0.5\text{V}$. When pin 12 is tied to the negative supply pin, the filter operation is stopped and the bandpass and lowpass outputs act as an S/H circuit holding the last sample.

S1_A, S1_B, (Pins 5 and 16)

These are voltage signal input pins and should be driven with a source impedance below $5\text{k}\Omega$. The S1_A, S1_B pins can be used to alter the clock to center frequency ratio (f_{CLK}/f_0) of the filter (see modes 1b, 1c, 2a, 2b) or to feedforward the input signal for allpass filter configurations (see modes 4 and 5). When these pins are not used, they should be tied to the AGND pin.

S_{A/B} (Pin 6)

When S_{A/B} is high, the S2 negative input of the voltage summer is tied to the lowpass output. When the S_{A/B} pin is connected to the negative supply, the S2 input switches to ground.

AGND (Pin 15)

AGND is connected to the system ground for dual supply operation. When operating with a single positive supply, the analog ground pin should be tied to $1/2$ of the supply and bypassed with a $0.1\mu\text{F}$ capacitor. The positive inputs of the internal op amps and the reference point of the internal switches are connected to the AGND pin.

f_{CLK}/f_0 RATIO

The ML2110 is a sampled data filter and approximates continuous time filters. The filter deviates from its ideal continuous filter model when the (f_{CLK}/f_0) ratio decreases and when the Qs are low.

$f_0 \times Q$ PRODUCT RATIO

The $f_0 \times Q$ product of the ML2110 depends on the clock frequency and the mode of operation. For clock frequencies below 1MHz , in mode 1 and its derivatives, the $f_0 \times Q$ product is mainly limited by the desired f_0 and Q accuracy. For the same clock frequency and for the same Q value the $f_0 \times Q$ product can be further increased if the clock to center frequency ratio is lowered below 50:1.

Mode 3, Figure 24, and the modes of operation where R4 is finite, are "slower" than the basic mode 1. The resistor R4 places the input op amp inside the resonant loop. The finite GBW of this op amp creates an additional phase shift and enhances the Q value at high clock frequencies.

OUTPUT NOISE

The wideband RMS noise of the ML2110 outputs is nearly independent from the clock frequency provided that the clock itself does not become part of the noise. The noise at the BP and LP outputs increases for high Qs.

FILTER FUNCTION DEFINITIONS

Each filter of the ML2110 with an external clock and resistors approximates 2nd order filter functions. These are tabulated below in the frequency domain.

- Bandpass function:** available at the bandpass output pins (2, 19), Figure 13.

$$G(s) = H_{OBP} \frac{s\omega_0/Q}{s^2 + (s\omega_0/Q) + \omega_0^2}$$

H_{OBP} = Gain at $\omega = \omega_0$

$f_0 = \omega_0/2\pi$; f_0 is the center frequency of the complex pole pair. f_0 is measured as the peak frequency of the bandpass output.

Q = Quality factor of the complex pole pair. It is the ratio of f_0 to the -3dB bandwidth of the 2nd order bandpass function. The Q is always measured at the filter BP output.

- Lowpass function:** available at the LP output pins (1, 20), Figure 14.

$$G(s) = H_{OLP} \frac{\omega_0^2}{s^2 + s(\omega_0/Q) + \omega_0^2}$$

H_{OLP} = DC gain of the LP output.

- Highpass function:** available only in mode 3 at the output pins (3, 18), Figure 15.

$$G(s) = H_{OHP} \frac{s^2}{s^2 + s(\omega_0/Q) + \omega_0^2}$$

H_{OHP} = gain of the HP output for $f \rightarrow \frac{f_{CLK}}{2}$

- Notch function:** available at pins 3 (18) for several modes of operation.

$$G(s) = (H_{ON2}) \frac{(s^2 + \omega_n^2)}{s^2 + s(\omega_0/Q) + \omega_0^2}$$

H_{ON2} = gain of the notch output for $f \rightarrow \frac{f_{CLK}}{2}$

H_{ON1} = gain of the notch output for $f \rightarrow 0$

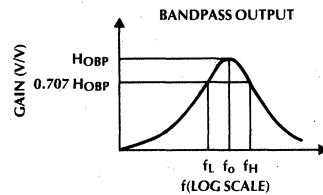
$f_n = \omega_n/2\pi$; f_n is the frequency of the notch occurrence.

- Allpass function:** available at pins 3(18) for mode 4, 4a.

$$G(s) = H_{OAP} \frac{[s^2 - s(\omega_0/Q) + \omega_0^2]}{s^2 + s(\omega_0/Q) + \omega_0^2}$$

H_{OAP} = gain of the allpass output for $0 < f < \frac{f_{CLK}}{2}$

For allpass functions, the center frequency and the Q of the numerator complex zero pair is the same as the denominator. Under these conditions, the magnitude response is a straight line. In mode 5, the center frequency f_z of the numerator complex zero pair, is different than f_0 . For high numerator Q 's, the magnitude response will have a notch at f_z .

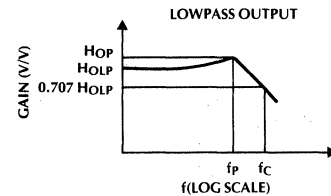


$$Q = \frac{f_0}{f_H - f_L}; f_0 = \sqrt{f_L f_H}$$

$$f_L = f_0 \left(\frac{-1}{2Q} + \sqrt{\left(\frac{1}{2Q}\right)^2 + 1} \right)$$

$$f_H = f_0 \left(\frac{1}{2Q} + \sqrt{\left(\frac{1}{2Q}\right)^2 + 1} \right)$$

Figure 13

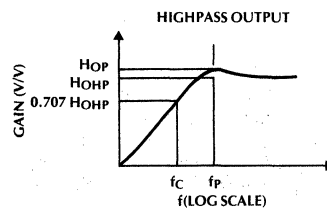


$$f_c = f_0 \times \sqrt{\left(1 - \frac{1}{2Q^2}\right) + \sqrt{\left(1 - \frac{1}{2Q^2}\right)^2 + 1}}$$

$$f_p = f_0 \sqrt{1 - \frac{1}{2Q^2}}$$

$$H_{OP} = H_{OLP} \times \frac{1}{Q \sqrt{1 - \frac{1}{4Q^2}}}$$

Figure 14



$$f_c = f_0 \times \left[\sqrt{\left(1 - \frac{1}{2Q^2}\right) + \sqrt{\left(1 - \frac{1}{2Q^2}\right)^2 + 1}} \right]^{-1}$$

$$f_p = f_0 \times \left[\sqrt{1 - \frac{1}{2Q^2}} \right]^{-1}$$

$$H_{OP} = H_{OHP} \times \frac{1}{Q \sqrt{1 - \frac{1}{4Q^2}}}$$

Figure 15

OPERATION MODES

Table 1. 1st Order Functions

MODE	PIN 2 (19)	PIN 3 (18)	f_c	f_z
6a	LP	HP	$\frac{f_{CLK}}{100(50)} \times \frac{R2}{R3}$	
6b	LP	LP	$\frac{f_{CLK}}{100(50)} \times \frac{R2}{R3}$	
7	LP	AP	$\frac{f_{CLK}}{100(50)} \times \frac{R2}{R3}$	$\frac{f_{CLK}}{100(50)} \times \frac{R2}{R3}$

Table 2. 2nd Order Functions

MODE	PIN 1 (20)	PIN 2 (19)	PIN 3 (18)	f_o	f_N
1	LP	BP	Notch	$\frac{f_{CLK}}{100(50)}$	f_o
1a	LP	BP	BP	$\frac{f_{CLK}}{100(50)}$	
1b	LP	BP	Notch	$\frac{f_{CLK}}{100(50)} \times \sqrt{1 + \frac{R6}{R5 + R6}}$	$\frac{f_{CLK}}{100(50)} \times \sqrt{1 + \frac{R6}{R5 + R6}}$
1c	LP	BP	Notch	$\frac{f_{CLK}}{100(50)} \times \sqrt{\frac{R6}{R5 + R6}}$	$\frac{f_{CLK}}{100(50)} \times \sqrt{\frac{R6}{R5 + R6}}$
1d	LP	BP		$\frac{f_{CLK}}{100(50)}$	
2	LP	BP	Notch	$\frac{f_{CLK}}{100(50)} \times \sqrt{1 + \frac{R2}{R4}}$	$\frac{f_{CLK}}{100(50)}$
2a	LP	BP	Notch	$\frac{f_{CLK}}{100(50)} \times \sqrt{1 + \frac{R2}{R4} + \frac{R6}{R5 + R6}}$	$\frac{f_{CLK}}{100(50)} \times \sqrt{1 + \frac{R6}{R5 + R6}}$
2b	LP	BP	Notch	$\frac{f_{CLK}}{100(50)} \times \sqrt{\frac{R2}{R4} + \frac{R6}{R5 + R6}}$	$\frac{f_{CLK}}{100(50)} \times \sqrt{\frac{R6}{R5 + R6}}$
3	LP	BP	HP	$\frac{f_{CLK}}{100(50)} \times \sqrt{\frac{R2}{R4}}$	
3a	LP	BP	Notch	$\frac{f_{CLK}}{100(50)} \times \sqrt{\frac{R2}{R4}}$	$\frac{f_{CLK}}{100(50)} \times \sqrt{\frac{R_h}{R_l}}$
4	LP	BP	AP	$\frac{f_{CLK}}{100(50)}$	
4a	LP	BP	AP	$\frac{f_{CLK}}{100(50)} \times \sqrt{\frac{R2}{R4}}$	
5	LP	BP	C.Z	$\frac{f_{CLK}}{100(50)} \times \sqrt{1 + \frac{R2}{R4}}$	$\frac{f_{CLK}}{100(50)} \times \sqrt{1 - \frac{R1}{R4}}$

ML2110

OPERATION MODES (Continued)

There are basically three modes of operation: mode 1, mode 2, mode 3. In the mode 1, Figure 16, the input amplifier is outside the resonant loop. Because of this, mode 1 and its derivatives (mode 1a, 1b, 1c, 1d) are faster than modes 2 and 3.

Mode 1a, Figure 17, represents the most simple hook-up of the ML2110. Mode 1a is useful when voltage gain at the bandpass output is required. The bandpass voltage gain, however, is equal to the value of Q, and a second order, clock tunable, BP resonator can be achieved with only 2 resistors. The filter center frequency directly depends on the external clock frequency. For high order filters, mode 1a is not practical as it requires several clock frequencies to tune the overall filter response.

Mode 1, Figure 16, provides a clock tunable notch. Mode 1 is a practical configuration for second order clock tunable bandpass/notch filters. In mode 1, a bandpass output with a very high Q, together with unity gain, can be obtained with the dynamics of the remaining notch and lowpass outputs.

Modes 1b and 1c, Figures 18, 19 are similar. They both produce a notch with a frequency which is always equal to the filter center frequency. The notch and the center frequency can be adjusted with an external resistor ratio.

The clock to center frequency ratio range is:

$$\frac{500}{1} \geq \frac{f_{CLK}}{f_0} \geq \frac{100}{1} \text{ or } \frac{50}{1}; \text{ mode 1b}$$

$$\frac{100}{1} \text{ or } \frac{50}{1} \geq \frac{f_{CLK}}{f_0} \geq \frac{100}{\sqrt{2}} \text{ or } \frac{50}{\sqrt{2}}; \text{ mode 1c}$$

The input impedance of the S1 pin is clock dependent, and in general R5 should not be larger than 5k. Mode 1b can be used to increase the clock to center frequency ratio beyond 100:1. For this mode, the limit for the (f_{CLK}/f_0) ratio is 500:1. Beyond this, the filter will exhibit large output offsets. Mode 1d, Figure 20, is the fastest mode of operation: In the 50:1 mode center frequencies beyond 20kHz can easily be achieved.

Modes 2, 2a, and 2b have a notch output which frequency, f_n , can be tuned independently from the center frequency, f_0 . For all cases, however, $f_n < f_0$. These modes are useful when cascading second order functions to create an overall elliptic highpass, bandpass or notch response. The input amplifier and its feedback resistors (R2/R4) are now part of the resonant loop. Because of this, mode 2 and its derivatives are slower than mode 1's.

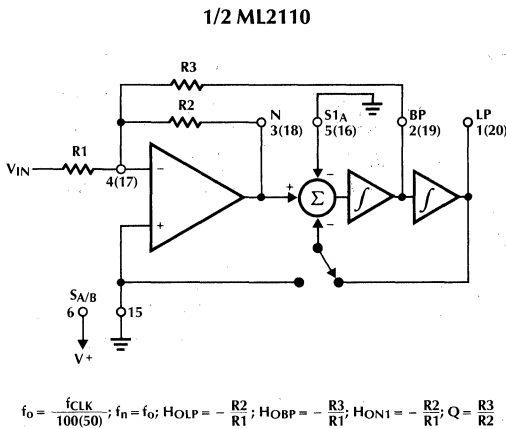


Figure 16. Mode 1: 2nd Order Filter Providing Notch, Bandpass, Lowpass

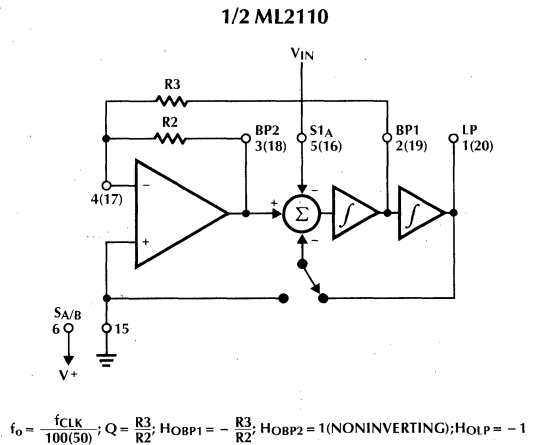


Figure 17. Mode 1a: 2nd Order Filter Providing Bandpass, Lowpass

OPERATION MODES (Continued)

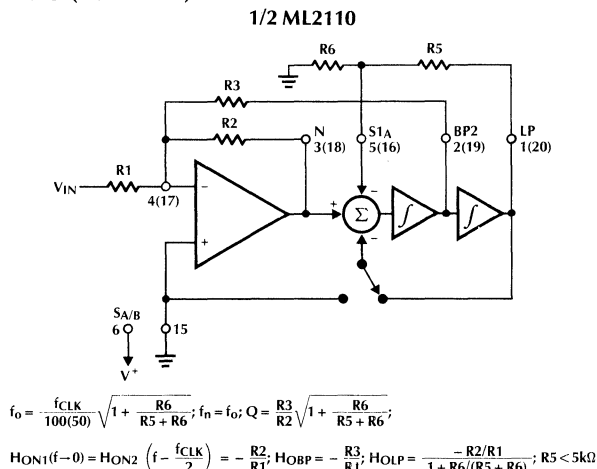


Figure 18. Mode 1b: 2nd Order Filter Providing Notch, Bandpass, Lowpass

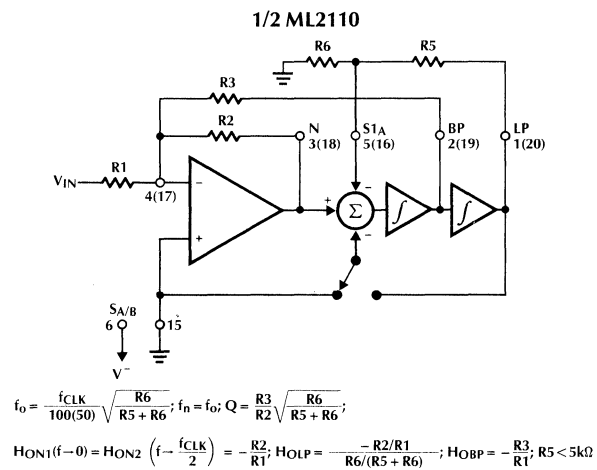


Figure 19. Mode 1c: 2nd Order Filter Providing Notch, Bandpass, Lowpass

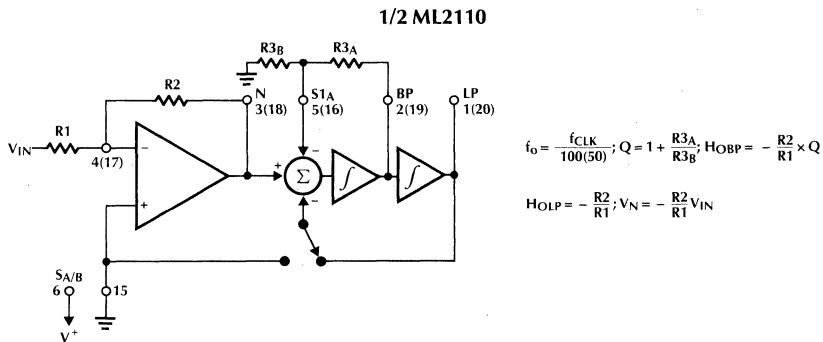
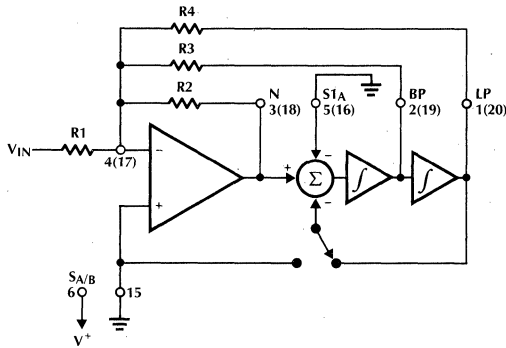


Figure 20. Mode 1d: 2nd Order Filter Providing Bandpass and Lowpass for Qs Greater Than or Equal to 1.

OPERATION MODES (Continued)

1/2 ML2110

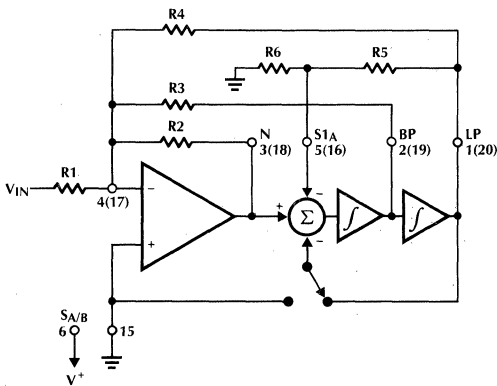


$$f_o = \frac{f_{CLK}}{100(50)} \sqrt{1 + \frac{R_2}{R_4}}; f_n = \frac{f_{CLK}}{100(50)}; Q = \frac{R_3}{R_2} \sqrt{1 + \frac{R_2}{R_4}}; H_{OLP} = \frac{-R_2/R_1}{1 + (R_2/R_4)}$$

$$H_{OBP} = -R_3/R_1; H_{ON1}(f=0) = \frac{-R_2/R_1}{1 + (R_2/R_4)}; H_{ON2} \left(f = \frac{f_{CLK}}{2} \right) = -R_2/R_1$$

Figure 21. Mode 2: 2nd Order Filter Providing Notch, Bandpass, Lowpass

1/2 ML2110



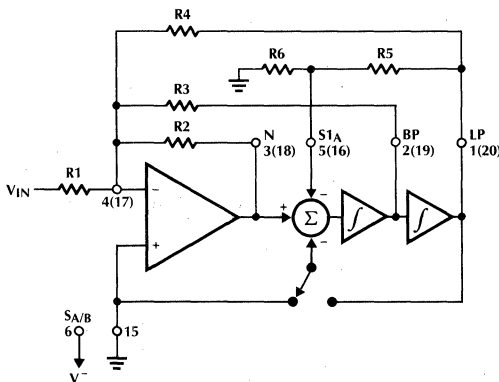
$$f_o = \frac{f_{CLK}}{100(50)} \sqrt{1 + \frac{R_2}{R_4} + \frac{R_6}{R_5 + R_6}}; f_n = \frac{f_{CLK}}{100(50)} \sqrt{1 + \frac{R_6}{R_5 + R_6}}; Q = \frac{R_3}{R_2} \sqrt{1 + \frac{R_2}{R_4} + \frac{R_6}{R_5 + R_6}}$$

$$H_{ON1}(f=0) = -\frac{R_2}{R_1} \left[\frac{1 + R_6/(R_5 + R_6)}{1 + (R_2/R_4) + [R_6/(R_5 + R_6)]} \right]; H_{ON2} \left(f = \frac{f_{CLK}}{2} \right) = -R_2/R_1$$

$$H_{OBP} = -R_3/R_1; H_{OLP} = \frac{-R_2/R_1}{1 + (R_2/R_4) + [R_6/(R_5 + R_6)]}$$

Figure 22. Mode 2a: 2nd Order Filter Providing Notch, Bandpass, Lowpass

1/2 ML2110



$$f_o = \frac{f_{CLK}}{100(50)} \sqrt{\frac{R_2}{R_4} + \frac{R_6}{R_5 + R_6}}; f_n = \frac{f_{CLK}}{100(50)} \sqrt{\frac{R_6}{R_5 + R_6}}; Q = \frac{R_3}{R_2} \sqrt{\frac{R_2}{R_4} + \frac{R_6}{R_5 + R_6}}$$

$$H_{ON1}(f=0) = -\frac{R_2}{R_1} \left[\frac{R_6/(R_5 + R_6)}{(R_2/R_4) + [R_6/(R_5 + R_6)]} \right]; H_{ON2} \left(f = \frac{f_{CLK}}{2} \right) = -R_2/R_1$$

$$H_{OBP} = -R_3/R_1; H_{OLP} = \frac{-R_2/R_1}{(R_2/R_4) + [R_6/(R_5 + R_6)]}$$

Figure 23. Mode 2b: 2nd Order Filter Providing Notch, Bandpass, Lowpass

OPERATION MODES (Continued)

In mode 3, Figure 24, a single resistor ratio (R_2/R_4) can tune the center frequency below or above the $f_{CLK}/100$ (or $f_{CLK}/50$) ratio. Mode 3 is a state variable configuration since it provides a highpass, bandpass, lowpass output through progressive integration; notches are obtained by summing the highpass and lowpass outputs (mode 3a, Figure 25). The notch frequency can be tuned below or above the center frequency through the resistor ratio (R_h/R_l). Because of this, modes 3 and 3a are the most versatile and useful modes for cascading second order sections to obtain high order elliptic filters. Figure 33, shows the 2 sections connected in mode 3a to obtain a clock tunable 4th order sharp elliptic bandpass filter. The first notch is created by summing directly the HP and LP outputs of the first section into the inverting input of the second section op amp. The individual Q's are 29.6 and the filter maintains its shape and performance up to 20kHz center frequency, as shown in Figure 34. For this circuit an external op amp is required to obtain the 2nd notch. The dynamics of Figure 34 show that the amplitude response at each output pin does not exceed 0dB. The gain in the passband

depends on the ratio of $(R_g/R_{h2}) \times (R_{22}/R_{h1}) \times (R_{21}/R_{l1})$. Any gain value can be obtained by acting on the (R_g/R_{h2}) ratio of the external op amp, the remaining ratios are adjusted for optimum dynamics of the output nodes. The external op amp of Figure 33 is not always required. In Figure 35, one section in mode 3a is cascaded with the other section in mode 2b to obtain a 4th order, 1dB ripple, elliptic bandreject filter. The clock to center frequency ratio is adjusted to 200:1; this is done in order to better approximate a linear R,C notch filter. The amplitude response of the filter is shown in Figure 36 with up to 1MHz clock frequency. The 0dB bandwidth to the stop bandwidth ratio is 8/1. When the filter is centered at 1kHz, it should theoretically have a 44dB rejection with a 50Hz stop bandwidth. For a more narrow filter than the above, the unused BP output of the mode 2b section, Figure 35, has a gain exceeding unity which limits the dynamic range of the overall filter. For very selective bandpass/bandreject filters, the mode 3a approach as in Figure 25, yields better dynamic range since the external op amp helps to optimize the dynamics of the output nodes of the ML2110.

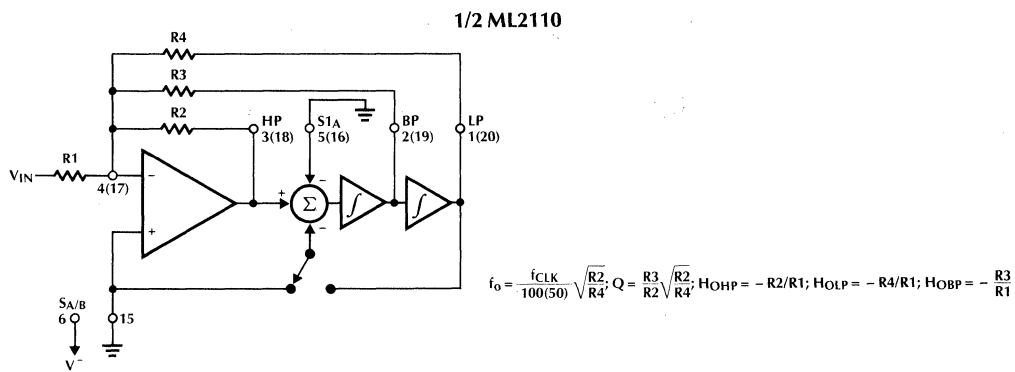


Figure 24. Mode 3: 2nd Order Filter Providing Highpass, Bandpass, Lowpass

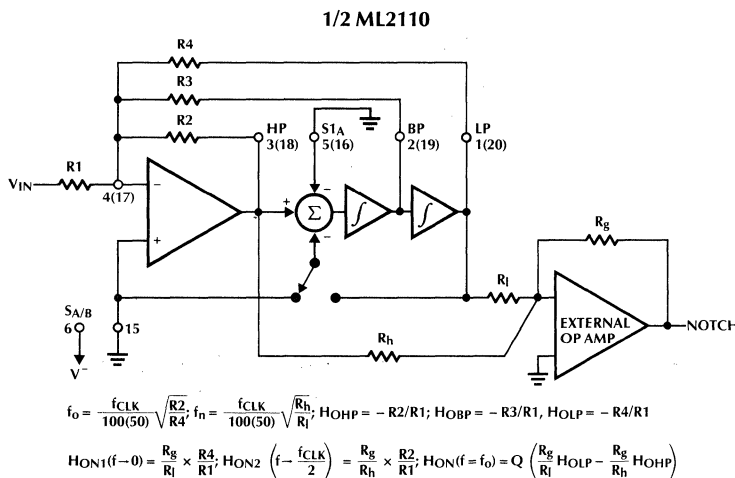
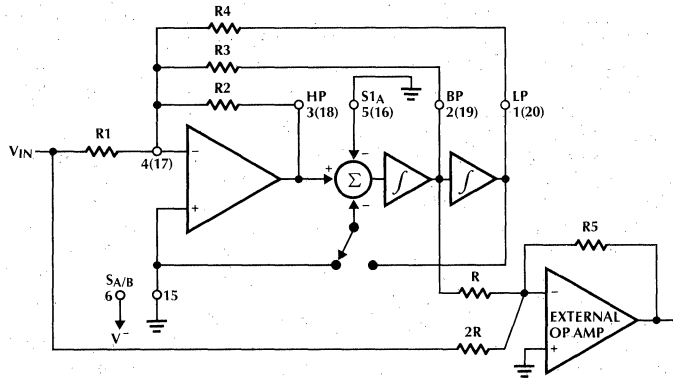


Figure 25. Mode 3a: 2nd Order Filter Providing Highpass, Bandpass, Lowpass, Notch

OPERATION MODES (Continued)

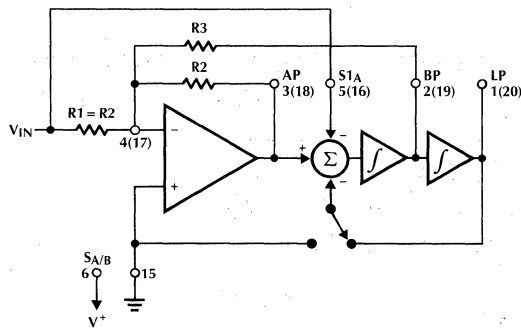
1/2 ML2110



$$f_0 = -\frac{f_{CLK}}{100(50)} \sqrt{\frac{R_2}{R_4}}; Q = \frac{R_3}{R_2} \sqrt{\frac{R_2}{R_4}}; HOAP = \frac{R_5}{2R}; HOHP = -\frac{R_2}{R_1}; HOBP = -\frac{R_3}{R_1}; HOLP = -\frac{R_4}{R_1}$$

Figure 26. Mode 4a: 2nd Order Filter Providing Highpass, Bandpass, Lowpass, Allpass

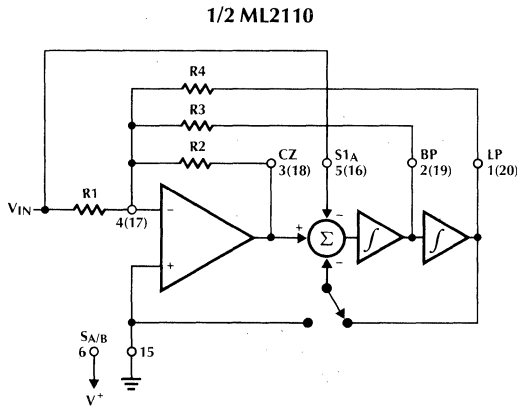
1/2 ML2110



$$f_0 = \frac{f_{CLK}}{100(50)}; Q = \frac{R_3}{R_2}; HOAP = -\frac{R_2}{R_1}; HOLP = -2; HOBP = -2 \left(\frac{R_3}{R_2} \right)$$

Figure 27. Mode 4: 2nd Order Filter Providing Allpass, Bandpass, Lowpass

OPERATION MODES (Continued)

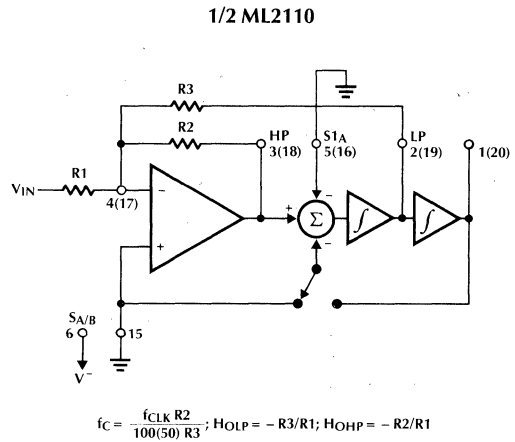


$$f_0 = \frac{f_{CLK}}{100(50)} \sqrt{1 + \frac{R_2}{R_4}}, f_z = \frac{f_{CLK}}{100(50)} \sqrt{1 - \frac{R_1}{R_4}}, Q = \frac{R_3}{R_2} \sqrt{1 + \frac{R_2}{R_4}}$$

$$Q_z = \frac{R_3}{R_1} \sqrt{1 - \frac{R_1}{R_4}}, HO_z(f=0) = \frac{(R_4/R_1) - 1}{(R_4/R_2) + 1}, HO_z\left(f = \frac{f_{CLK}}{2}\right) = \frac{R_2}{R_1}$$

$$HO_{BP} = \frac{R_3}{R_2} \left(1 + \frac{R_2}{R_1}\right); HO_{LP} = \frac{1 + (R_2/R_1)}{1 + (R_2/R_4)}$$

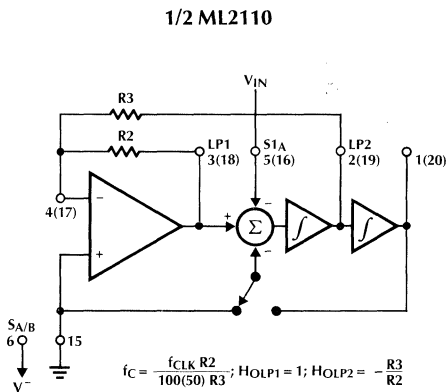
Figure 28. Mode 5: 2nd Order Filter Providing Numerator Complex Zeros, Bandpass, Lowpass



$$f_c = \frac{f_{CLK} R_2}{100(50) R_3}; HO_{LP} = -R_3/R_1; HO_{HP} = -R_2/R_1$$

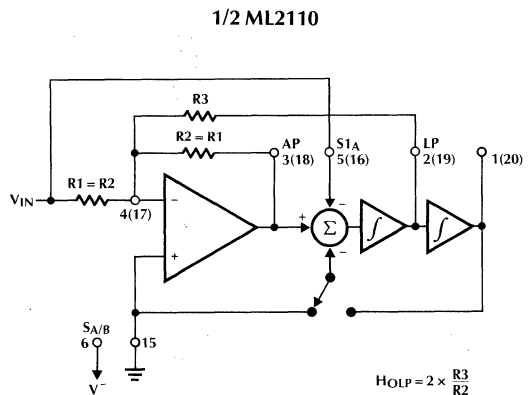
Figure 29. Mode 6a: 1st Order Filter Providing Highpass, Lowpass

3



$$f_c = \frac{f_{CLK} R_2}{100(50) R_3}; HO_{LP1} = 1; HO_{LP2} = -\frac{R_3}{R_2}$$

Figure 30. Mode 6b: 1st Order Filter Providing Lowpass



$$HO_{LP} = 2 \times \frac{R_3}{R_2}$$

$$f_p = \frac{f_{CLK} R_2}{100(50) R_3}; f_z = \frac{f_{CLK} R_2}{100(50) R_3}; GAIN AT OUTPUT = 1 \text{ FOR } 0 \leq f \leq \frac{f_{CLK}}{2}$$

Figure 31. Mode 7: 1st Order Filter Providing Allpass, Lowpass

OPERATION MODES (Continued)

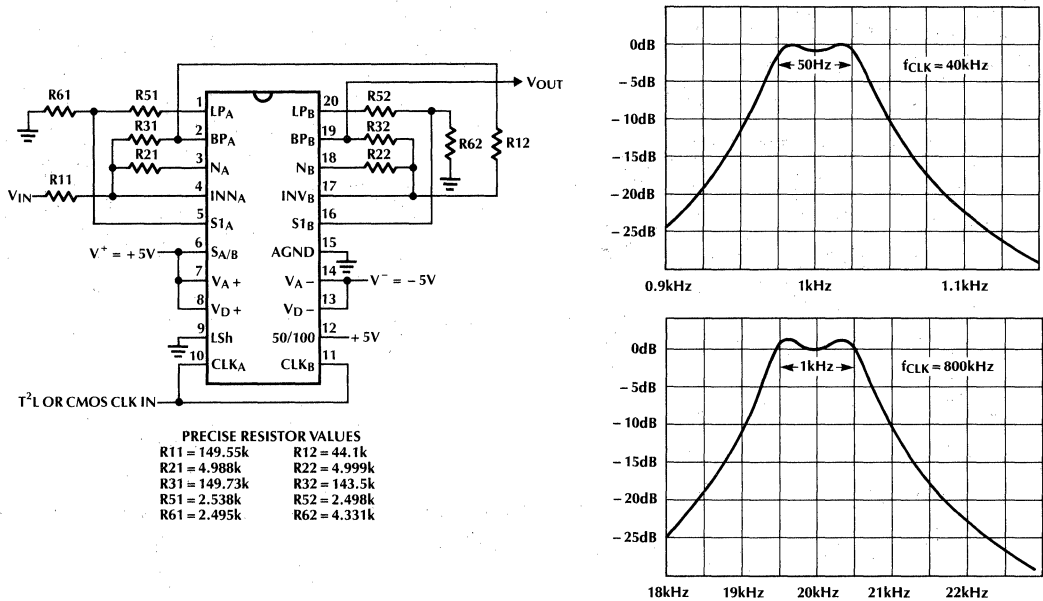
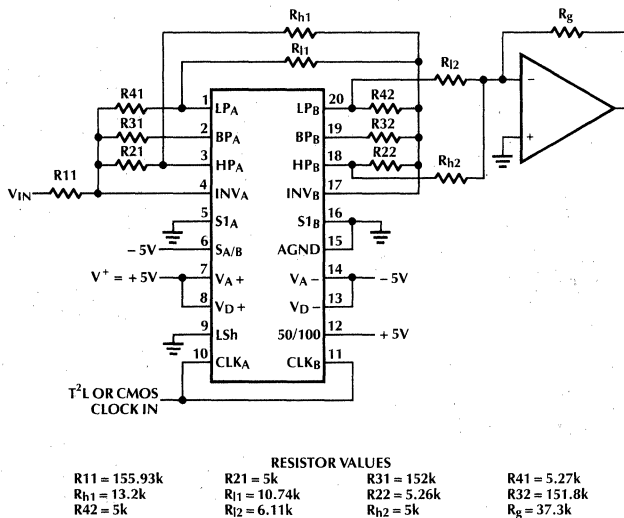


Figure 32. Cascading the 2 sections connected in mode 1b to obtain a clock tunable 4th order 1dB ripple bandpass Chebyshev filter with (center frequency)/(Ripple Bw) = 20/1.



NOTE: FOR CLOCK FREQUENCIES ABOVE 700kHz A 12pF CAPACITOR ACROSS R41 AND A 20pF CAPACITOR ACROSS R42 WERE USED TO PREVENT THE PASSBAND RIPPLE FROM ANY ADDITIONAL PEAKING.

Figure 33. Combining mode 3 with mode 3a to make the 4th order BP filter of Figure 34 with improved dynamics. The gain at each node is ≤ 0 dB for all input frequencies.

OPERATION MODES (Continued)

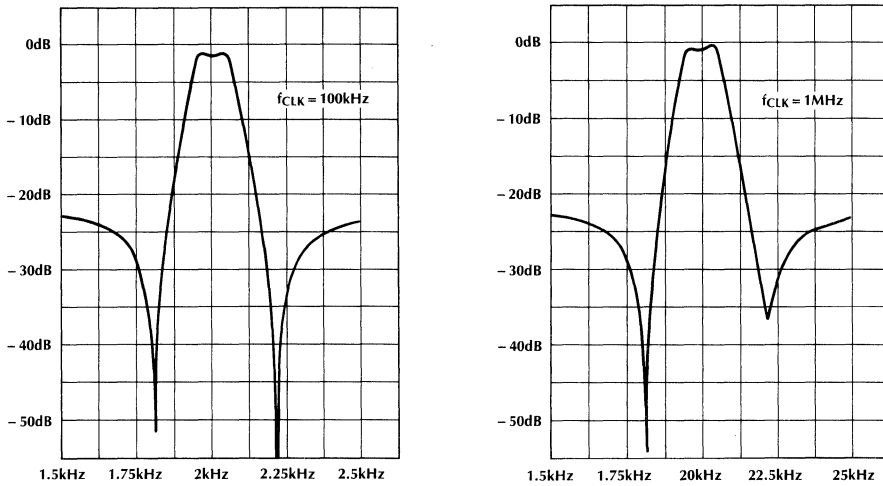


Figure 34. The BP filter of Figure 33, when swept from a 2kHz to 20kHz center frequency.

3

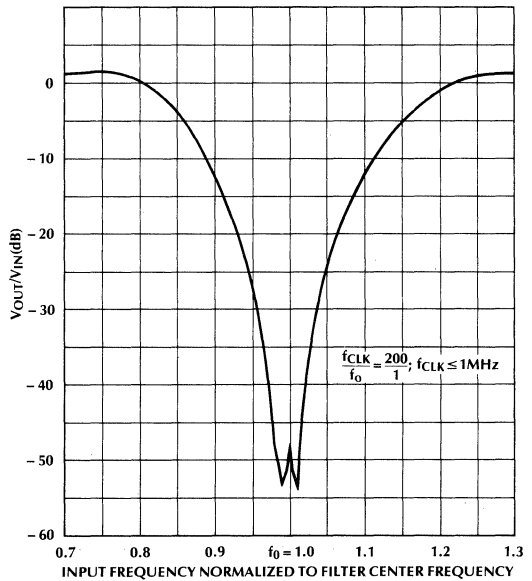
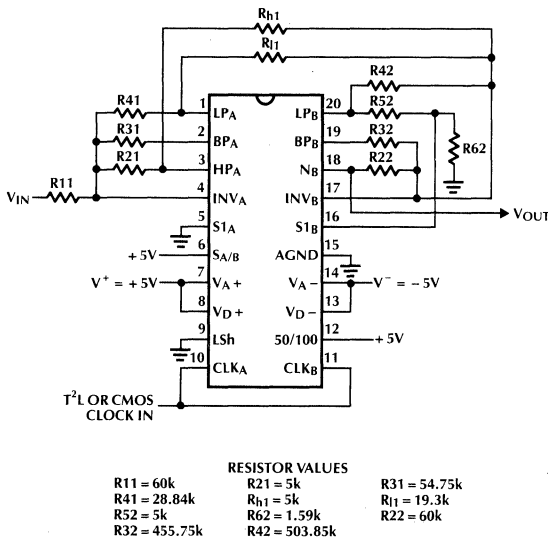


Figure 35. Combining mode 3 with mode 2b to create a 4th order BP elliptic filter with 1dB ripple and a ratio of 0db to stop bandwidth equal to 8/1.

Figure 36. Amplitude Response of the Notch Filter of Figure 35.

OFFSETS

Switched capacitor integrators generally exhibit higher input offsets than discrete R,C integrators.

These offsets are mainly the charge injection of the CMOS switches into the integrating capacitors. The internal op amp offsets also add to the overall offset budget.

Figure 37 shows half of the ML2110 filter with its equivalent input offsets V_{OS1} , V_{OS2} , V_{OS3} .

The DC offset at the filter bandpass output is always equal to V_{OS3} . The DC offsets at the remaining two outputs

(Notch and LP) depend on the mode of operation and external resistor ratios. Table 3 illustrates this.

It is important to know the value of the DC output offsets, especially when the filter handles input signals with large dynamic range. As a rule of thumb, the output DC offsets increase when:

1. The Q's decrease
2. The ratio (f_{CLK}/f_o) increases beyond 100:1. This is done by decreasing either the ($R2/R4$) or the $R6/(R5 + R6)$ resistor ratios.

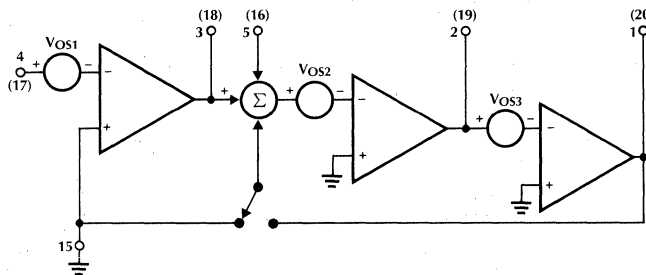


Figure 37. Equivalent Input Offsets of 1/2 ML2110 Filter

Table 3

MODE	V_{OSN} PIN 3 (18)	V_{OSBP} PIN 2 (19)	V_{OSLP} PIN 1 (20)
1,4	$V_{OS1}[(1/Q)+1+ H_{OLP}]-V_{OS3}/Q$	V_{OS3}	$V_{OSN}-V_{OS2}$
1a	$V_{OS1}[1+(1/Q)]-V_{OS3}/Q$	V_{OS3}	$V_{OSN}-V_{OS2}$
1b	$V_{OS1}[(1/Q)+1+R2/R1]-V_{OS3}/Q$	V_{OS3}	$\sim(V_{OSN}-V_{OS2})(1+R5/R6)$
1c	$V_{OS1}[(1/Q)+1+R2/R1]-V_{OS3}/Q$	V_{OS3}	$\sim(V_{OSN}-V_{OS2})\frac{(R5+R6)}{(R5+2R6)}$
1d	$V_{OS1}[1+R2/R1]$	V_{OS3}	$V_{OSN}-V_{OS2}-V_{OS3}/Q$
2, 5	$[V_{OS1}(1+R2/R1+R2/R3+R2/R4)-V_{OS3}(R2/R3)] \times [R4/(R2+R4)]+V_{OS2}[R2/(R2+R4)]$	V_{OS3}	$V_{OSN}-V_{OS2}$
2a	$[V_{OS1}(1+R2/R1+R2/R3+R2/R4)-V_{OS3}(R2/R3)] \times \left[\frac{R4(1+k)}{R2+R4(1+k)} \right] + V_{OS2} \left[\frac{R2}{R2+R4(1+k)} \right]; k = \frac{R6}{R5+R6}$	V_{OS3}	$\sim(V_{OSN}-V_{OS2})\frac{(R5+R6)}{(R5+2R6)}$
2b	$[V_{OS1}(1+R2/R1+R2/R3+R2/R4)-V_{OS3}(R2/R3)] \times \left[\frac{R4k}{R2+R4k} \right] + V_{OS2} \left[\frac{R2}{R2+R4k} \right]; k = \frac{R6}{R5+R6}$	V_{OS3}	$\sim(V_{OSN}-V_{OS2})(1+R5/R6)$
3, 4a	V_{OS2}	V_{OS3}	$V_{OS1} \left[1 + \frac{R4}{R1} + \frac{R4}{R2} + \frac{R4}{R3} \right] - V_{OS2} \left(\frac{R4}{R2} \right) - V_{OS3} \left(\frac{R4}{R3} \right)$

ORDERING INFORMATION

PART NUMBER	TEMP. RANGE	PACKAGE
ML2110BCP	0°C to +70°C	MOLDED DIP (P20)
ML2110CCP	0°C to +70°C	MOLDED DIP (P20)
ML2110BCS	0°C to +70°C	MOLDED SOIC (S20W)
ML2110CCS	0°C to +70°C	MOLDED SOIC (S20W)
ML2110BJ	-40°C to +85°C	HERMETIC DIP (J20)
ML2110CIJ	-40°C to +85°C	HERMETIC DIP (J20)

PART NUMBER	TEMP. RANGE	PACKAGE
ML2110BMJ	-55°C to +125°C	HERMETIC DIP (J20)
ML2110CMJ	-55°C to +125°C	HERMETIC DIP (J20)
ML2110BIP	-40°C to +85°C	MOLDED DIP (P20)
ML2110CIP	-40°C to +85°C	MOLDED DIP (P20)
ML2110BIS	-40°C to +85°C	MOLDED SOIC (S20W)
ML2110CIS	-40°C to +85°C	MOLDED SOIC (S20W)

Universal Hi-Frequency Dual Filter

GENERAL DESCRIPTION

The ML2111 consists of two independent switched capacitor filters that operate up to 150kHz. These filters perform second order functions, such as lowpass, bandpass, highpass, notch and allpass. All filter configurations, including Butterworth, Bessel, Cauer and Chebyshev can be formed.

The center frequency of these filters are tuned by an external clock or the external clock and a resistor ratio.

The ML2111 frequency range up to 150kHz is specified with $\pm 5.0V \pm 10\%$ power supplies. Using a single $5.0V \pm 10\%$ power supply the frequency range is up to 100kHz.

These filters are ideal where center frequency accuracy and high Qs are needed.

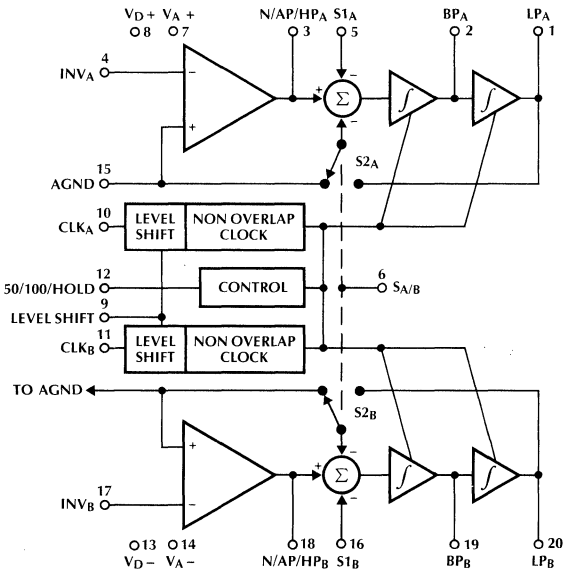
The ML2111 is a pin compatible superior replacement for MF10, LMF100, and LTC1060 filters.

FEATURES

- Guaranteed frequency range to 150kHz
- Center frequency \times Q product $\leq 5\text{MHz}$
- Separate highpass, notch, allpass, bandpass, and lowpass outputs
- Center frequency accuracy $\pm 0.4\%$ or $\pm 0.8\%$ max
- Q accuracy $\pm 4\%$, or $\pm 8\%$ max
- Clock inputs TTL or CMOS compatible
- Single $5V (\pm 2.25V)$ or $\pm 5V \pm 10\%$ supply operation guaranteed
- 0°C to $+70^\circ\text{C}$, -40°C to $+85^\circ\text{C}$, -55°C to $+125^\circ\text{C}$ operating temperature range
- Standard 0.3" 20-pin DIP or 20-pin small outline (SOIC) package

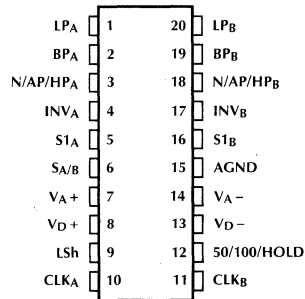
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BLOCK DIAGRAM



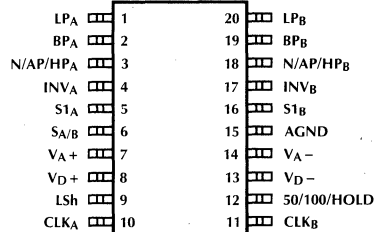
PIN CONNECTIONS

ML2111
20-PIN DIP



TOP VIEW

ML2111
20-PIN SOIC



TOP VIEW

ML2111

PIN DESCRIPTION

PIN NO.	NAME	FUNCTION	PIN NO.	NAME	FUNCTION
1	LP _A	Lowpass output for biquad A.	12	50/100/HOLD	Input pin to control the clock to center frequency ratio of 50:1 or 100:1, or stops the clock to hold the last sample of the bandpass or lowpass outputs.
2	BP _A	Bandpass output for biquad A.	13	V _{D-}	Negative digital supply.
3	N/AP/HP _A	Notch/allpass/highpass output for biquad A.	14	V _{A-}	Negative analog supply.
4	INV _A	Inverting input of the summing op amp for biquad A.	15	AGND	Analog ground.
5	S1 _A	Auxiliary signal input pin used in modes 1a, 1d, 4, 5, and 6b.	16	S1 _B	Auxiliary signal input used in modes 1a, 1d, 4, 5, and 6b.
6	S _{A/B}	Controls S2 input function.	17	INV _B	Inverting input of the summing op amp for biquad B.
7	V _{A+}	Positive analog supply.	18	N/AP/HP _B	Notch/allpass/highpass output for biquad B.
8	V _{D+}	Positive digital supply.	19	BP _B	Bandpass output for biquad B.
9	LS _H	Reference point for clock input levels. Logic threshold typically 1.4V above LS _H voltage.	20	LP _B	Lowpass output for biquad B.
10	CLK _A	Clock input for biquad A.			
11	CLK _B	Clock input for biquad B.			

ABSOLUTE MAXIMUM RATINGS

(Note 1)

Supply Voltage

V _{A+} , V _{D+} - V _{A-} , V _{D-}	13V
V _{A+} , V _{D+} to LS _H	13V
Inputs	V _{A+} , V _{D+} + 0.3V to V _{A-} , V _{D-} - 0.3V
Outputs	V _{A+} , V _{D+} + 0.3V to V _{A-} , V _{D-} - 0.3V
V _{A+} to V _{D+}	±0.3V
Power Dissipation	750mW
Storage Temperature Range	-65°C to 150°C
Lead Temperature (soldering, 10 sec)	300°C

OPERATING CONDITIONS

Temperature Range (Note 2)

ML2111BCP, ML2111CCP	0°C to 70°C
ML2111BCS, ML2111CCS	-40°C to +85°C
ML2111BIJ, ML2111CIJ	-55°C to +125°C
ML2111BMJ, ML2111CMJ	±2.25V to ±6.0V

ELECTRICAL CHARACTERISTICS

Unless otherwise specified T_A = T_{MIN} to T_{MAX}, V_{A+} = V_{D+} = 5V ± 10%, V_{A-} = V_{D-} = -5V ± 10%, C_L = 25pF, V_{IN} = 1.41V_{PK} (1.00V_{RMS}), Clock Duty Cycle 45% to 55%.

PARAMETER	NOTES	CONDITIONS	ML2111B			ML2111C			UNITS
			MIN	TYP NOTE 3	MAX	MIN	TYP NOTE 3	MAX	
Filter									
f _o , Center Frequency Maximum	5, 6	Figure 15 (Mode 1) V _{IN} = 1V _{PK} (.707V _{RMS}) Q ≤ 50, Q Accuracy ≤ ±25% Q ≤ 20, Q Accuracy ≤ ±15%			100 150			100 150	kHz kHz
f _o , Center Frequency Minimum	5, 6	Figure 15 (Mode 1) Q ≤ 50, Q Accuracy ≤ ±30% Q ≤ 20, Q Accuracy ≤ ±15%	25 25			25 25			Hz Hz
f _o , Temperature Coefficient		f _{CLK} < 5MHz		-10			-10		ppm/°C
Clock to Center Frequency Ratio		Q = 10 Figure 15 (Mode 1)							
	4 4	50:1, f _{CLK} = 5MHz 100:1, f _{CLK} = 5MHz	49.65 99.60	49.85 100	50.05 100.40	49.45 99.20	49.85 100	50.25 100.80	

ELECTRICAL CHARACTERISTICS (Continued)

Unless otherwise specified $T_A = T_{MIN}$ to T_{MAX} , $V_{A+} = V_{D+} = 5V \pm 10\%$, $V_{A-} = V_{D-} = -5V \pm 10\%$, $C_L = 25pF$, $V_{IN} = 1.41V_{PK}$ (1.00 V_{RMS}), Clock Duty Cycle 50% (Note 8).

PARAMETER	NOTES	CONDITIONS	ML2111B			ML2111C			UNITS
			MIN	TYP NOTE 3	MAX	MIN	TYP NOTE 3	MAX	
Filter (Continued)									
Clock Frequency	5	$Q \leq 20$, Q Accuracy $\leq \pm 15\%$	2.5k		7.5M	2.5k		7.5M	Hz
Clock Feedthrough	5	$f_{CLK} \leq 5MHz$		10	20		10	20	mV(p-p)
Q Accuracy	4	$f_{CLK} = 5MHz$, Q = 10 Figure 15 (Mode 1)	50:1 100:1		± 3 ± 4			± 5 ± 8	%
Q Temperature Coefficient	5	$f_{CLK} < 5MHz$, Q = 10		20			20		ppm/°C
DC Offset V _{OS2,3}	4	50:1, $f_{CLK} = 5MHz$ S _{A/B} High		7	40		7	60	mV
V _{OS2,3}	4	S _{A/B} Low		7	40		7	60	mV
DC Offset V _{OS2,3}	4	100:1, $f_{CLK} = 5MHz$ S _{A/B} High		14	60		14	100	mV
V _{OS2,3}	4	S _{A/B} Low		14	60		14	100	mV
Gain Accuracy DC Lowpass	4	R1 = 20k, R2 = 2k, R3 = 20k		0.01	2		0.01	2	%
Bandpass at f_0	4	100:1, $f_0 = 50kHz$, Q = 10		1	4		1	6	%
DC Notch Output	5			0.02	2		0.02	2	%
Noise	7	Figure 15 (Mode 1) Q = 1, R1 = R2 = R3 = 2k							
Bandpass, 100kHz, 50:1				103			103		μV_{RMS}
50kHz, 100:1				121			121		μV_{RMS}
Lowpass, 100kHz, 50:1				120			120		μV_{RMS}
50kHz, 100:1				150			150		μV_{RMS}
Notch, 100kHz, 50:1				115			115		μV_{RMS}
50kHz, 100:1				135			135		μV_{RMS}
Figure 15 (Mode 1) Q = 10, R1 = R3 = 20k, R2 = 2k									
Bandpass, 100kHz, 50:1				262			262		μV_{RMS}
50kHz, 100:1				333			333		μV_{RMS}
Lowpass, 100kHz, 50:1				268			268		μV_{RMS}
(R1 = 2k) 50kHz, 100:1				342			342		μV_{RMS}
Notch, 100kHz, 50:1				64			64		μV_{RMS}
(R1 = 2k) 50kHz, 100:1				72			72		μV_{RMS}
Crosstalk		$f_{CLK} = 5MHz$, $f_0 = 100kHz$		-50			-50		dB
Filter, $V_{A+} = V_{D+} = 2.25V$, $V_{A-} = V_{D-} = -2.25V$, $V_{IN} = 0.707 V_{PK}$ (0.5V_{RMS}) (NOTE 9)									
f_0 , Center Frequency Maximum	5	Figure 15 (Mode 1) $Q \leq 50$, Q Accuracy $\leq \pm 30\%$ $Q \leq 20$, Q Accuracy $\leq \pm 15\%$			75 100			75 100	kHz kHz
f_0 , Center Frequency Minimum	5	Figure 15 (Mode 1) $Q \leq 50$, Q Accuracy $\leq \pm 30\%$ $Q \leq 20$, Q Accuracy $\leq \pm 15\%$	25 25			25 25			Hz Hz
Clock to Center Frequency Ratio		Q = 10							
	4	Figure 15 (Mode 1) 50:1, $f_{CLK} = 2.5MHz$	49.65	49.85	50.05	49.45	49.85	50.25	
	5	100:1, $f_{CLK} = 2.5MHz$	99.60	100	100.40	99.20	100	100.80	
Clock Frequency	5	$Q \leq 20$, Q Accuracy $\leq \pm 15\%$	2.5k		5M	2.5k		5M	Hz
Q Accuracy	4	$f_{CLK} = 2.5MHz$, Q = 10 Figure 15 (Mode 1)	50:1 100:1		± 3			± 6	%

ELECTRICAL CHARACTERISTICS (Continued)

Unless otherwise specified $T_A = T_{MIN}$ to T_{MAX} , $V_{A+} = V_{D+} = 5V \pm 10\%$, $V_{A-} = V_{D-} = -5V \pm 10\%$, $C_L = 25pF$, $V_{IN} = 1.41V_{PK}$ (1.00 V_{RMS}), Clock Duty Cycle 50% (Note 8).

PARAMETER	NOTES	CONDITIONS	ML2111B			ML2111C			UNITS	
			MIN	TYP NOTE 3	MAX	MIN	TYP NOTE 3	MAX		
Filter, $V_{A+} = V_{D+} = 2.25V$, $V_{A-} = V_{D-} = -2.25V$, $V_{IN} = 0.707 V_{PK}$ (0.5 V_{RMS}) (Continued)										
Noise	7	Figure 15 (Mode 1) $Q = 1, R1 = R2 = R3 = 2k$								
		Bandpass, 100kHz, 50:1		105			105		μV_{RMS}	
		50kHz, 100:1		123			123		μV_{RMS}	
		Lowpass, 100kHz, 50:1		122			122		μV_{RMS}	
		50kHz, 100:1		152			152		μV_{RMS}	
		Notch, 100kHz, 50:1		117			117		μV_{RMS}	
		50kHz, 100:1		138			138		μV_{RMS}	
		Figure 15 (Mode 1) $Q = 10, R1 = R3 = 20k, R2 = 2k$								
		Bandpass, 100kHz, 50:1		265			265		μV_{RMS}	
		50kHz, 100:1		335			335		μV_{RMS}	
		Lowpass, 100kHz, 50:1		270			270		μV_{RMS}	
		($R1 = 2k$) 50kHz, 100:1		245			245		μV_{RMS}	
		Notch, 100kHz, 50:1		65			65		μV_{RMS}	
		($R1 = 2k$) 50kHz, 100:1		73			73		μV_{RMS}	
Operational Amplifiers and Power Supply										
V_{OS} DC Offset	4			2		15		2	15	mV
DC Open Loop Gain		$R_L = 1k$		95				95		dB
Gain Bandwidth Product				2.4				2.4		MHz
Slew Rate				2.0				2.0		V/ μs
Output Voltage Swing (Clipping Level)	5	$R_L = 2k$, V from V_{A+} or V_{A-}		0.5		1.2		0.5	1.2	V
Output Short Circuit Current		Source		50				50		mA
		Sink		25				25		mA
Power Supply And Clock										
Supply Current ($I_{A+} + I_{D+}$)	4	$f_{CLK} = 5MHz$		13		22		13	22	mA
($I_{A-} + I_{D-}$)				12		21		12	21	mA
I_{LSH}				0.5		1		0.5	1	mA
V_{CLK} Input Threshold	4	$f_{CLK} = 5MHz$	Low			0.6			0.6	V
			High	3.0			3.0			V
	5	$f_{CLK} < 2.5MHz$	Low			0.8			0.8	V
			High	2.0			2.0			V
CLKA, CLKB Pulse Width	5, 8	CLK High or	$ V_{D+} - V_{D-} \geq 4.5V$	100			100			ns
		CLK Low	$ V_{D+} - V_{D-} \geq 9.0V$	66			66			ns

Note 1: Absolute maximum ratings are limits beyond which the life of the integrated circuit may be impaired. All voltages unless otherwise specified are measured with respect to ground.

Note 2: $-55^{\circ}C$ to $+125^{\circ}C$ operating temperature range devices are 100% tested at temperature extremes with worst-case test conditions. $0^{\circ}C$ to $70^{\circ}C$, and $-40^{\circ}C$ to $+85^{\circ}C$ operating temperature range devices are 100% tested with temperature limits guaranteed by 100% testing, sampling, or by correlation with worst-case test conditions.

Note 3: Typicals are parametric norm at $25^{\circ}C$.

Note 4: Parameter guaranteed and 100% production tested.

Note 5: Parameter guaranteed. Parameters not 100% tested are not in outgoing quality level calculation.

Note 6: Center frequency is defined as the peak of the bandpass output.

Note 7: The noise is measured with the HP8903A audio analyzer with a bandwidth of 750kHz which is 7.5 times the f_0 at 50:1 and 15 times the f_0 at 100:1.

Note 8: For best performance with $f_{CLK} > 2.5MHz$ use a 50% duty cycle.

Note 9: For $T_A = -55^{\circ}C$ to $+125^{\circ}C$; $V_{A+} = V_{D+} = 2.375V$, $V_{A-} = V_{D-} = -2.375V$

FUNCTIONAL DESCRIPTION

POWER SUPPLIES

The analog (V_A+) and digital (V_D+) supply voltage pins should be tied together and bypassed to AGND with at least a $0.1\mu\text{F}$ and a $0.01\mu\text{F}$ disc ceramic capacitor. If high digital noise exists, the supply pins can be bypassed separately. The ML2111 positive analog and positive digital supply pins are internally connected by the IC substrate and should be biased from the same DC source. The ML2111 negative analog and negative digital supply are not connected internally, however they should be biased from the same DC source and bypassed with at least a $0.1\mu\text{F}$ and a $0.01\mu\text{F}$ disc ceramic capacitor.

The ML2111 operates with a single supply from 4V to 12V and with split supplies from $\pm 2.0\text{V}$ to $\pm 6\text{V}$.

CLOCK INPUT PINS AND LEVEL SHIFT

With dual supplies equal to or higher than $\pm 4.0\text{V}$, the level shift (LSH) pin 9 can be connected to the same potential as the AGND or V_A- pin. With single supply operation, the negative supply pins and the LSH pin should be tied to the system ground. The AGND, pin 15, should be biased at 1/2 supplies. Under these conditions, the clock levels are TTL or CMOS. The input clock pins (10, 11) share the same level shift pin.

50/100/HOLD (Pin 12)

By tying pin 12 to (V_A+ , V_D+) the filter operates in the 50:1 mode. By tying pin 12 to 1/2 of the voltage supplies (AGND potential), the ML2111 operates in the 100:1 mode. The range of pin 12 without affecting the 100:1 filter operation with total supply voltage of +5V is $2.5 \pm 0.5\text{V}$; +10V is $5\text{V} \pm 0.5\text{V}$. When pin 12 is tied to the negative supply pin, the filter operation is stopped and the bandpass and lowpass outputs act as an S/H circuit holding the last sample.

S1_A, S1_B, (Pins 5 and 16)

These are the auxiliary voltage signal input pins always connected to one of the negative inputs of the voltage summer (the other negative input switches between LPo and AGND according to control pin $S_{A/B}$ (pin 6)). The positive input of the voltage summer is always connected to N/AP/HP pin of the corresponding section. They should be driven with a source impedance below 5k for $f_{\text{CLK}} < 2.5\text{MHz}$ and 1k to 2k for $f_{\text{CLK}} > 2.5\text{MHz}$. The S1_A, S1_B pins can be used to alter the clock to center frequency ratio (f_{CLK}/f_o) of the filter (see modes 1b, 1c, 2a, 2b) or to feedforward the input signal for allpass filter configurations (see modes 4 and 5). They can

also be used, as in mode 1d to avoid the finite phase shift through the input amplifier, hence allowing higher operating frequencies. When these pins are not used, they should be tied to the AGND pin.

S_{A/B} (Pin 6)

When $S_{A/B}$ is high, the S2 negative input of the voltage summer is tied to the lowpass output. When the $S_{A/B}$ pin is connected to the negative supply, the S2 input switches to ground.

AGND (Pin 15)

AGND is connected to the system ground for dual supply operation. When operating with a single positive supply, the analog ground pin should be tied to 1/2 of the supply and bypassed with a $0.1\mu\text{F}$ and a $0.01\mu\text{F}$ disc ceramic capacitor. The positive inputs of the internal op amps and the reference point of the internal switches are connected to the AGND pin.

f_{CLK}/f_o RATIO

The ML2111 is a sampled data filter and approximates continuous time filters. The filter deviates from its ideal continuous filter model when the (f_{CLK}/f_o) ratio decreases and when the Q_s are low.

$f_o \times Q$ PRODUCT RATIO

The $f_o \times Q$ product of the ML2111 depends on the clock frequency and the mode of operation. For clock frequencies below 5MHz, in mode 1 and its derivatives, the $f_o \times Q$ product is mainly limited by the desired f_o and Q accuracy. For the same clock frequency and for the same Q value the $f_o \times Q$ product can be further increased if the clock to center frequency ratio is lowered below 50:1.

Mode 3, Figure 23, and the modes of operation where R4 is finite, are "slower" than the basic mode 1. The resistor R4 places the input op amp inside the resonant loop. The finite GBW of this op amp creates an additional phase shift and enhances the Q value at high clock frequencies.

OUTPUT NOISE

The wideband RMS noise of the ML2111 outputs is nearly independent from the clock frequency provided that the clock itself does not become part of the noise. The noise at the BP and LP outputs increases for high Q_s .

TYPICAL PERFORMANCE CURVES

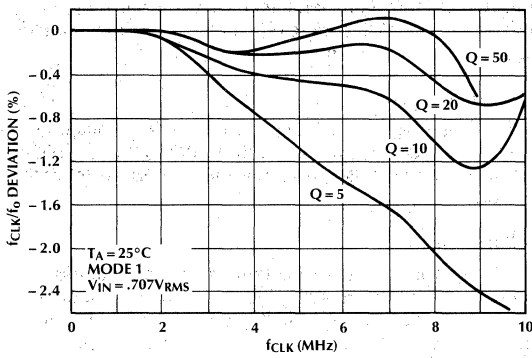


Figure 1A. f_{CLK}/f_0 vs. f_{CLK} (50:1, $V_S = \pm 5V$)

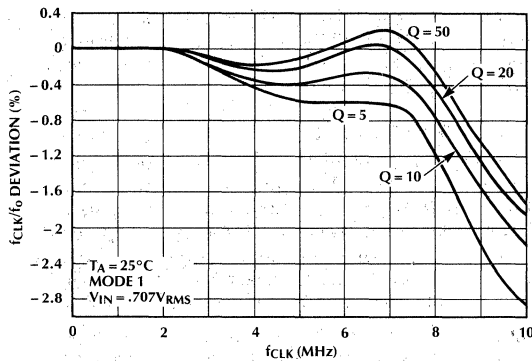
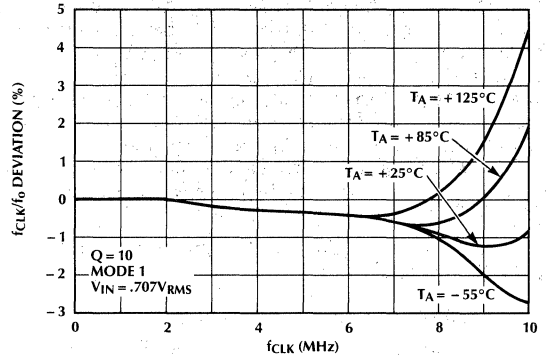


Figure 1B. f_{CLK}/f_0 vs. f_{CLK} (100:1, $V_S = \pm 5V$)

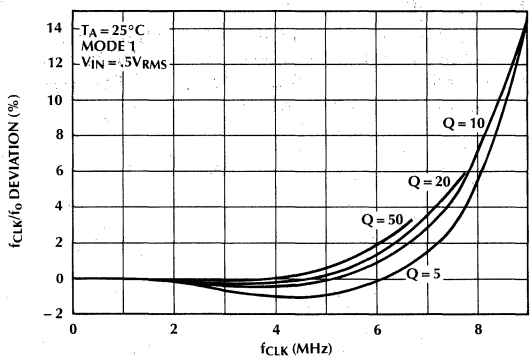
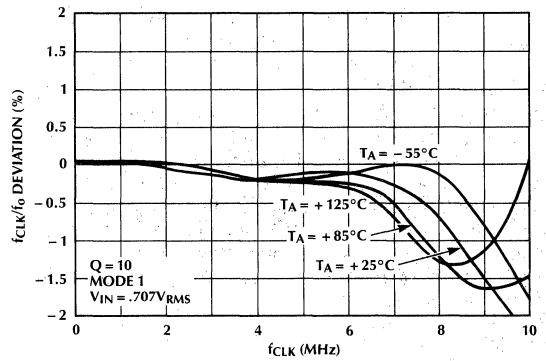
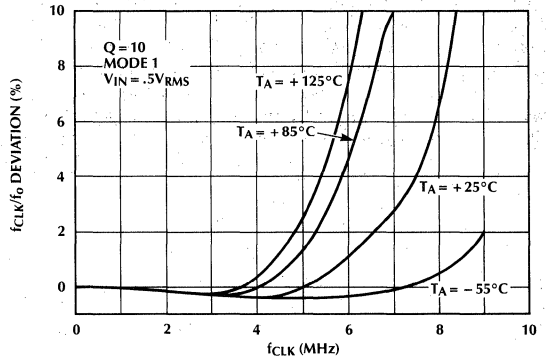


Figure 1C. f_{CLK}/f_0 vs. f_{CLK} (50:1, $V_S = \pm 2.5V$)



TYPICAL PERFORMANCE CURVES (Continued)

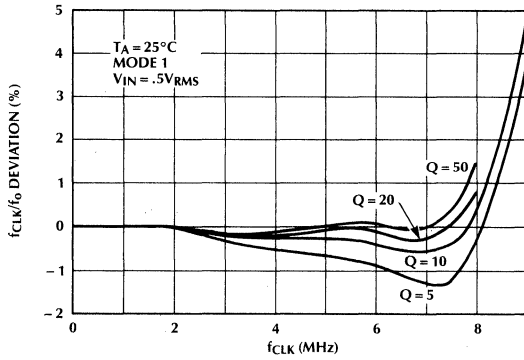


Figure 1D. f_{CLK}/f_0 vs. f_{CLK} (100:1, $V_S = \pm 2.5V$)

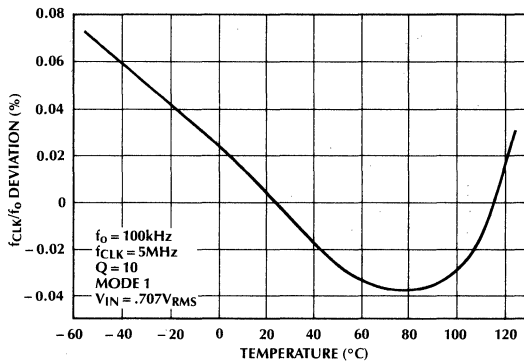
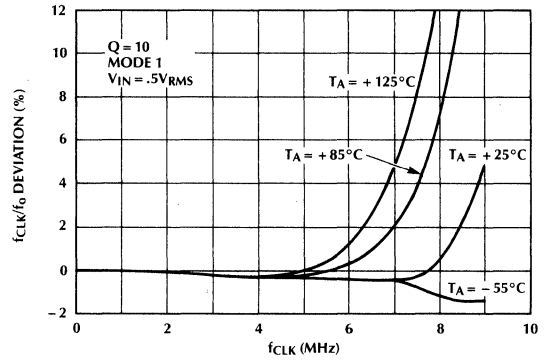


Figure 2A. f_{CLK}/f_0 Deviation vs. Temperature (50:1, $V_S = \pm 5V$)

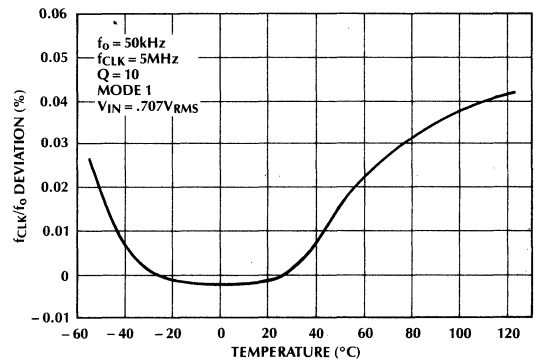


Figure 2B. f_{CLK}/f_0 Deviation vs. Temperature (100:1, $V_S = \pm 5V$)

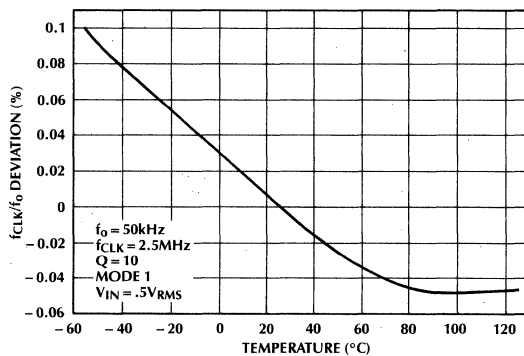


Figure 2C. f_{CLK}/f_0 Deviation vs. Temperature (50:1, $V_S = \pm 2.5V$)

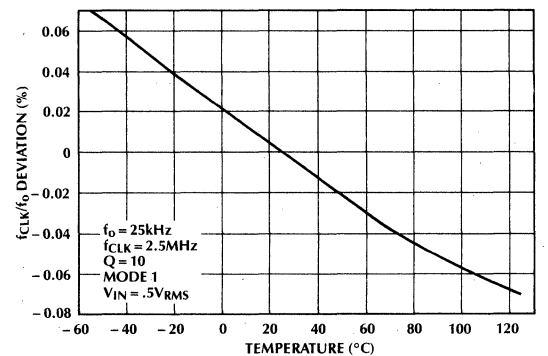


Figure 2D. f_{CLK}/f_0 Deviation vs. Temperature (100:1, $V_S = \pm 2.5V$)

3

TYPICAL PERFORMANCE CURVES (Continued)

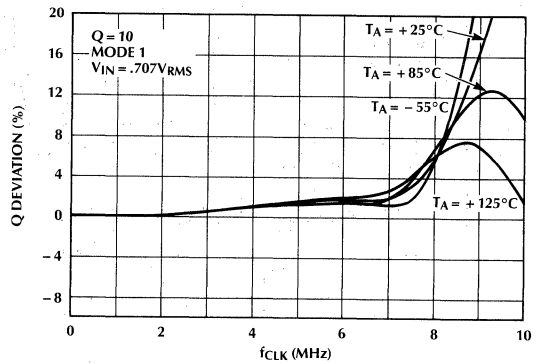
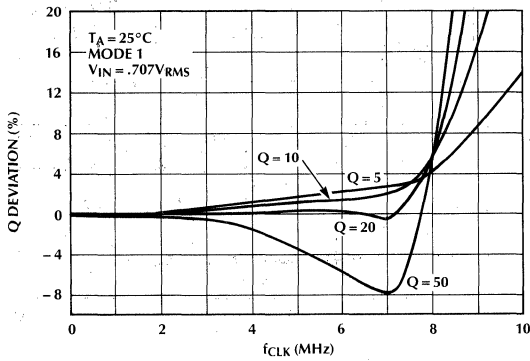


Figure 2E. Q Error vs. f_{CLK} (50:1, $V_S = \pm 5V$)

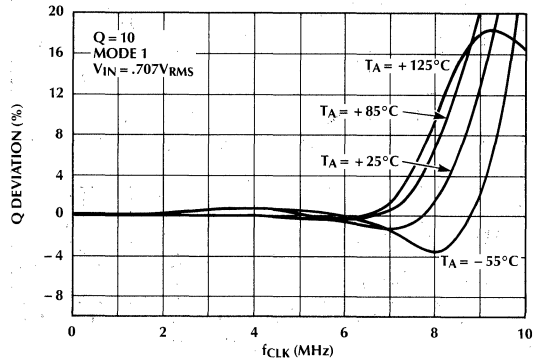
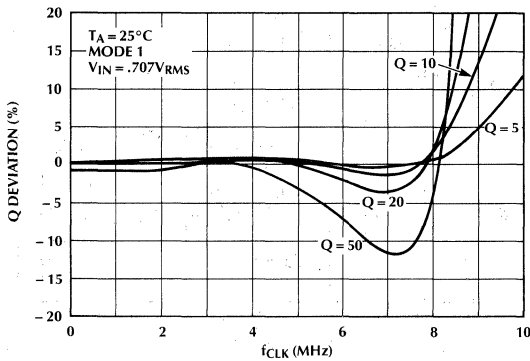


Figure 2F. Q Error vs. f_{CLK} (100:1, $V_S = \pm 5V$)

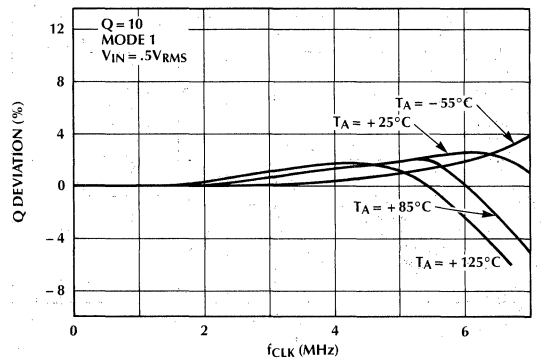
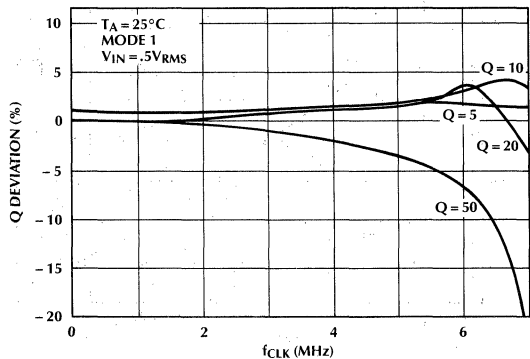


Figure 2G. Q Error vs. f_{CLK} (50:1, $V_S = \pm 2.5V$)

TYPICAL PERFORMANCE CURVES (Continued)

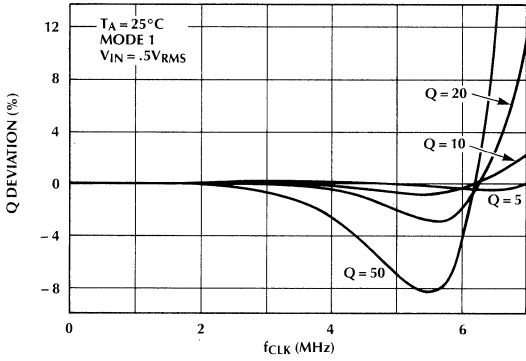


Figure 2H. Q Error vs. f_{CLK} (100:1, $V_S = \pm 2.5V$)

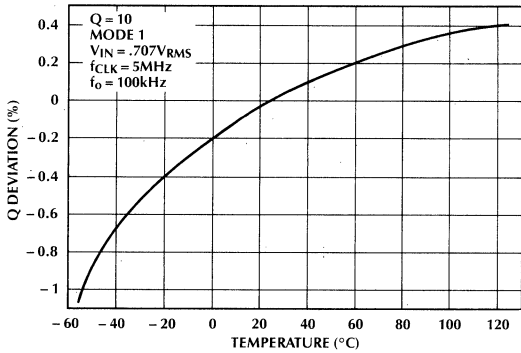
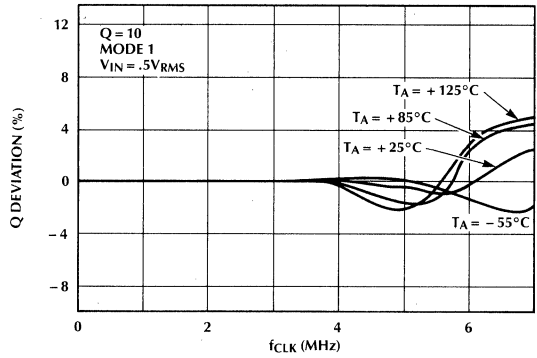


Figure 3A. Q Deviation vs. Temperature (50:1, $V_S = \pm 5V$)

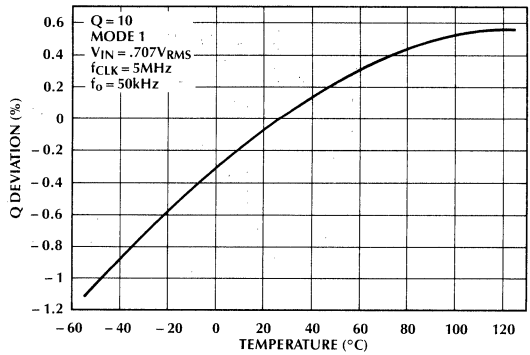


Figure 3B. Q Deviation vs. Temperature (100:1, $V_S = \pm 5V$)

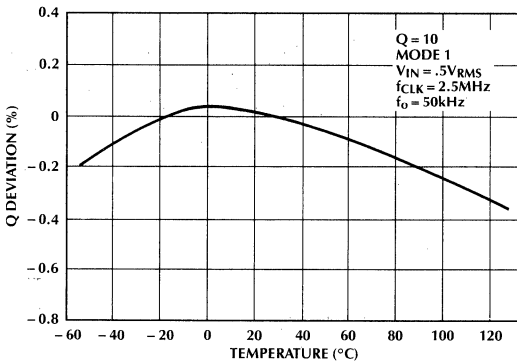


Figure 3C. Q Deviation vs. Temperature (50:1, $V_S = \pm 2.5V$)

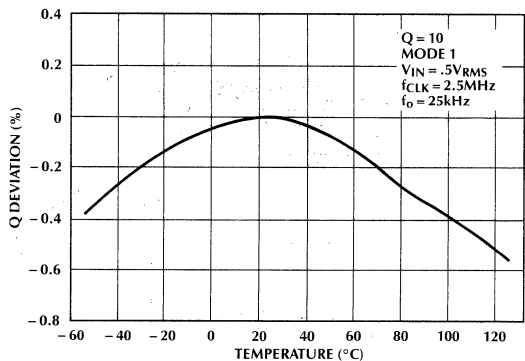


Figure 3D. Q Deviation vs. Temperature (100:1, $V_S = \pm 2.5V$)

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TYPICAL PERFORMANCE CURVES (Continued)

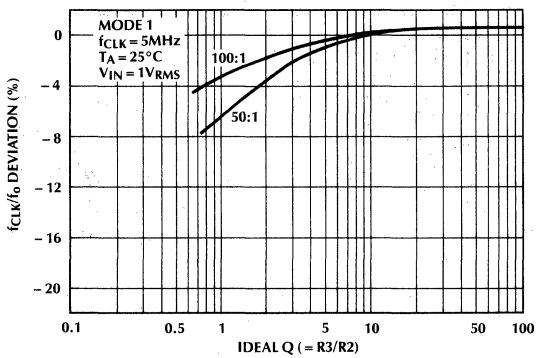


Figure 4A. f_{CLK}/f_0 Deviation vs. Q ($V_S = \pm 5V$)

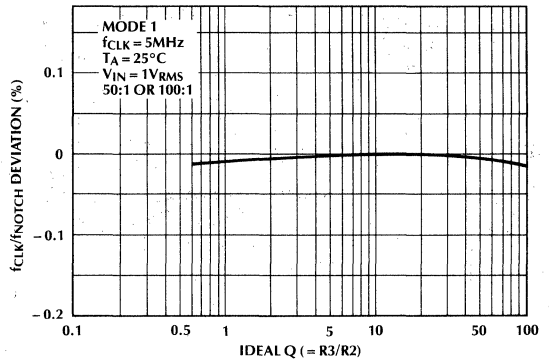


Figure 4B. f_{CLK}/f_{NOTCH} Deviation vs. Q ($V_S = \pm 5V$)

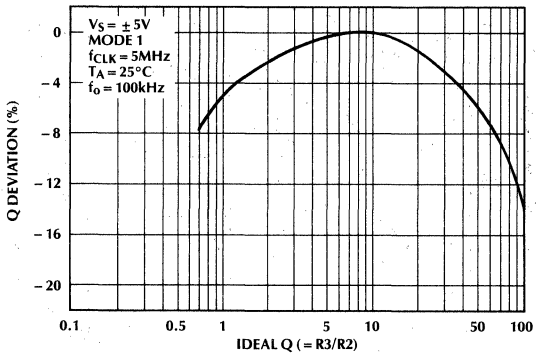


Figure 5A. Q Deviation vs. Q (50:1, $V_S = \pm 5V$)

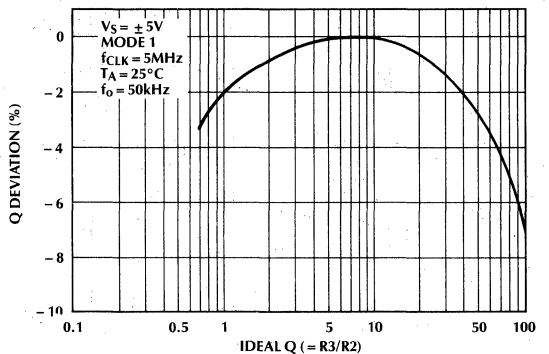


Figure 5B. Q Deviation vs. Q (100:1, $V_S = \pm 5V$)

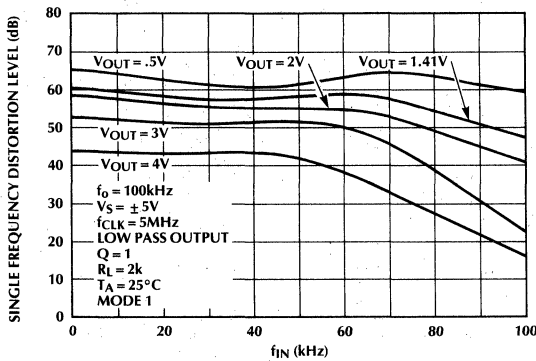


Figure 6A. Distortion vs. f_{IN} (50:1, $V_S = \pm 5V$)

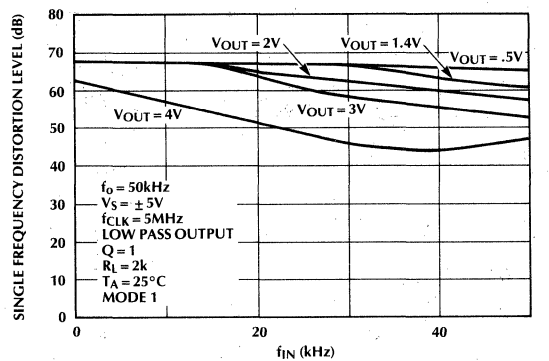


Figure 6B. Distortion vs. f_{IN} (100:1, $V_S = \pm 5V$)

TYPICAL PERFORMANCE CURVES (Continued)

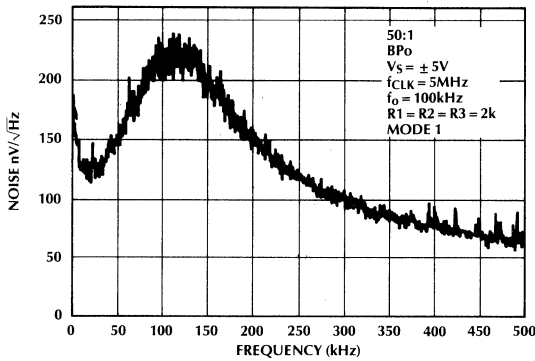


Figure 7A. Noise Spectrum Density (Q = 1)

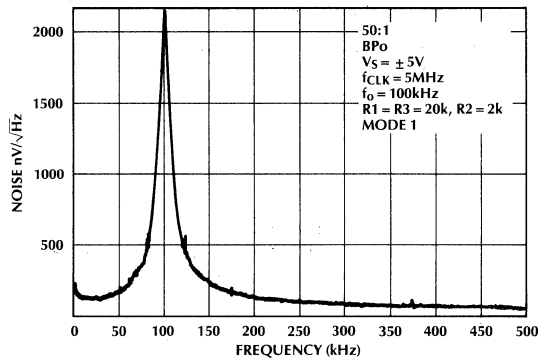


Figure 7B. Noise Spectrum Density (Q = 10)

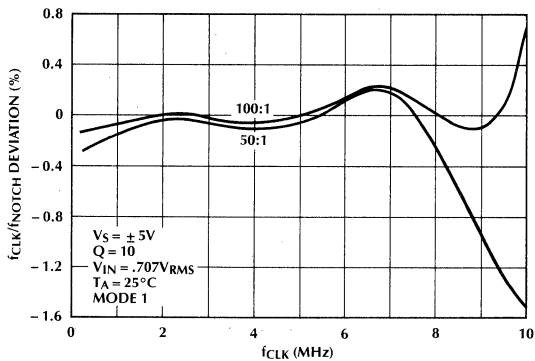


Figure 8. f_{CLK}/f_{NOTCH} vs. f_{CLK}

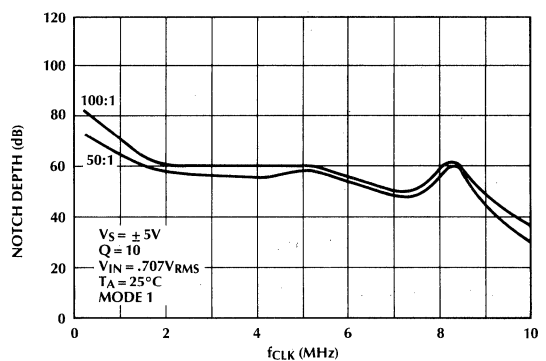


Figure 9. Notch Depth vs. f_{CLK}

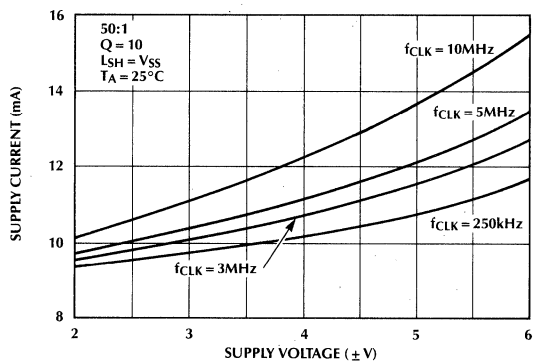


Figure 10. Supply Current vs. Supply Voltage

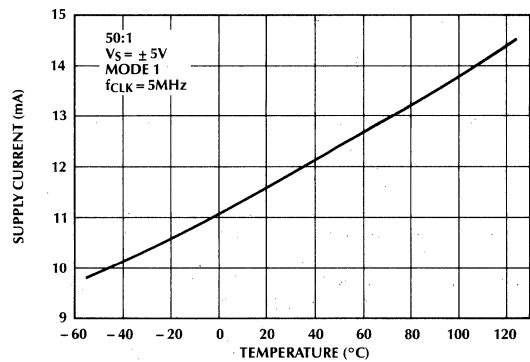


Figure 11. Supply Current vs. Temperature

3

FILTER FUNCTION DEFINITIONS

Each filter of the ML2111 with an external clock and resistors approximates 2nd order filter functions. These are tabulated below in the frequency domain.

- Bandpass function:** available at the bandpass output pins (2, 19), Figure 12.

$$G(s) = H_{OBP} \frac{s\omega_o/Q}{s^2 + (s\omega_o/Q) + \omega_o^2}$$

H_{OBP} = Gain at $\omega = \omega_o$

$f_o = \omega_o/2\pi$; f_o is the center frequency of the complex pole pair. f_o is measured as the peak frequency of the bandpass output.

Q = Quality factor of the complex pole pair. It is the ratio of f_o to the -3dB bandwidth of the 2nd order bandpass function. The Q is always measured at the filter BP output.

- Lowpass function:** available at the LP output pins (1, 20), Figure 13.

$$G(s) = H_{OLP} \frac{\omega_o^2}{s^2 + s(\omega_o/Q) + \omega_o^2}$$

H_{OLP} = DC gain of the LP output.

- Highpass function:** available only in mode 3 at the output pins (3, 18), Figure 14.

$$G(s) = H_{OHP} \frac{s^2}{s^2 + s(\omega_o/Q) + \omega_o^2}$$

H_{OHP} = gain of the HP output for $f \rightarrow \frac{f_{CLK}}{2}$

- Notch function:** available at pins 3 (18) for several modes of operation.

$$G(s) = (H_{ON2}) \frac{(s^2 + \omega_n^2)}{s^2 + s(\omega_o/Q) + \omega_o^2}$$

H_{ON2} = gain of the notch output for $f \rightarrow \frac{f_{CLK}}{2}$

H_{ON1} = gain of the notch output for $f \rightarrow 0$

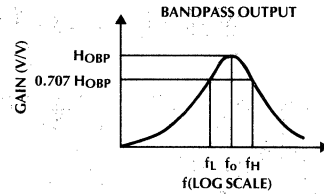
$f_n = \omega_n/2\pi$; f_n is the frequency of the notch occurrence.

- Allpass function:** available at pins 3(18) for mode 4, 4a.

$$G(s) = H_{OAP} \frac{[s^2 - s(\omega_o/Q) + \omega_o^2]}{s^2 + s(\omega_o/Q) + \omega_o^2}$$

H_{OAP} = gain of the allpass output for $0 < f < \frac{f_{CLK}}{2}$

For allpass functions, the center frequency and the Q of the numerator complex zero pair is the same as the denominator. Under these conditions, the magnitude response is a straight line. In mode 5, the center frequency f_z , of the numerator complex zero pair, is different than f_o . For high numerator Q 's, the magnitude response will have a notch at f_z .

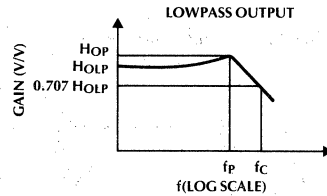


$$Q = \frac{f_o}{f_H - f_L}; f_o = \sqrt{f_L f_H}$$

$$f_L = f_o \left(\frac{-1}{2Q} + \sqrt{\left(\frac{1}{2Q}\right)^2 + 1} \right)$$

$$f_H = f_o \left(\frac{1}{2Q} + \sqrt{\left(\frac{1}{2Q}\right)^2 + 1} \right)$$

Figure 12

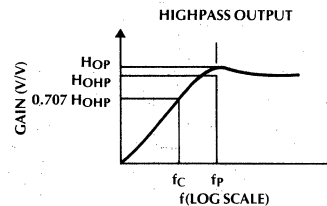


$$f_c = f_o \times \sqrt{\left(1 - \frac{1}{2Q^2}\right) + \sqrt{\left(1 - \frac{1}{2Q^2}\right)^2 + 1}}$$

$$f_p = f_o \sqrt{1 - \frac{1}{2Q^2}}$$

$$H_{OP} = H_{OLP} \times \frac{1}{Q \sqrt{1 - \frac{1}{4Q^2}}}$$

Figure 13



$$f_c = f_o \times \left[\sqrt{\left(1 - \frac{1}{2Q^2}\right) + \sqrt{\left(1 - \frac{1}{2Q^2}\right)^2 + 1}} \right]^{-1}$$

$$f_p = f_o \times \left[\sqrt{1 - \frac{1}{2Q^2}} \right]^{-1}$$

$$H_{OP} = H_{OHP} \times \frac{1}{Q \sqrt{1 - \frac{1}{4Q^2}}}$$

Figure 14

OPERATION MODES

Table 1. 1st Order Functions

MODE	PIN 2 (19)	PIN 3 (18)	f_c	f_z
6a	LP	HP	$\frac{f_{CLK}}{100(50)} \times \frac{R2}{R3}$	
6b	LP	LP	$\frac{f_{CLK}}{100(50)} \times \frac{R2}{R3}$	
7	LP	AP	$\frac{f_{CLK}}{100(50)} \times \frac{R2}{R3}$	$\frac{f_{CLK}}{100(50)} \times \frac{R2}{R3}$

Table 2. 2nd Order Functions

MODE	PIN 1 (20)	PIN 2 (19)	PIN 3 (18)	f_o	f_n
1	LP	BP	Notch	$\frac{f_{CLK}}{100(50)}$	f_o
1a	LP	BP	BP	$\frac{f_{CLK}}{100(50)}$	
1b	LP	BP	Notch	$\frac{f_{CLK}}{100(50)} \times \sqrt{1 + \frac{R6}{R5 + R6}}$	$\frac{f_{CLK}}{100(50)} \times \sqrt{1 + \frac{R6}{R5 + R6}}$
1c	LP	BP	Notch	$\frac{f_{CLK}}{100(50)} \times \sqrt{\frac{R6}{R5 + R6}}$	$\frac{f_{CLK}}{100(50)} \times \sqrt{\frac{R6}{R5 + R6}}$
1d	LP	BP		$\frac{f_{CLK}}{100(50)}$	
2	LP	BP	Notch	$\frac{f_{CLK}}{100(50)} \times \sqrt{1 + \frac{R2}{R4}}$	$\frac{f_{CLK}}{100(50)}$
2a	LP	BP	Notch	$\frac{f_{CLK}}{100(50)} \times \sqrt{1 + \frac{R2}{R4} + \frac{R6}{R5 + R6}}$	$\frac{f_{CLK}}{100(50)} \times \sqrt{1 + \frac{R6}{R5 + R6}}$
2b	LP	BP	Notch	$\frac{f_{CLK}}{100(50)} \times \sqrt{\frac{R2}{R4} + \frac{R6}{R5 + R6}}$	$\frac{f_{CLK}}{100(50)} \times \sqrt{\frac{R6}{R5 + R6}}$
3	LP	BP	HP	$\frac{f_{CLK}}{100(50)} \times \sqrt{\frac{R2}{R4}}$	
3a	LP	BP	Notch	$\frac{f_{CLK}}{100(50)} \times \sqrt{\frac{R2}{R4}}$	$\frac{f_{CLK}}{100(50)} \times \sqrt{\frac{R_h}{R_l}}$
4	LP	BP	AP	$\frac{f_{CLK}}{100(50)}$	
4a	LP	BP	AP	$\frac{f_{CLK}}{100(50)} \times \sqrt{\frac{R2}{R4}}$	
5	LP	BP	C.Z	$\frac{f_{CLK}}{100(50)} \times \sqrt{1 + \frac{R2}{R4}}$	$\frac{f_{CLK}}{100(50)} \times \sqrt{1 - \frac{R1}{R4}}$

3

OPERATION MODES (Continued)

There are basically three modes of operation: mode 1, mode 2, mode 3. In the mode 1, Figure 15, the input amplifier is outside the resonant loop. Because of this, mode 1 and its derivatives (mode 1a, 1b, 1c, 1d) are faster than modes 2 and 3. The table below gives an approximation of the frequency range for each mode.

Mode 1*	High Frequency Mode
1, 1a, 1d 1b, 1c	f ₀ up to 150kHz; Q up to 20** f ₀ up to 100kHz; Q up to 30
Mode 2	Flexible for Notches
2, 2a, 2b	f ₀ up to 30kHz; Q up to 30
Mode 3	Most Flexible/ Low Component Count
3, 3a	f ₀ up to 30kHz; Q up to 30

* Q and f₀ have an inverse relationship. This table is only an approximation. Actual performance depends on board layout and stray capacitance.

** 15% of less Q deviation. Higher Q's can be realized with greater deviation.

Mode 1a, Figure 16, represents the most simple hook-up of the ML2111. Mode 1a is useful when voltage gain at the bandpass output is required. The bandpass voltage gain, however, is equal to the value of Q, and a second order, clock tunable, BP resonator can be achieved with only 2 resistors. The filter center frequency directly depends on the external clock frequency. For high order filters, mode 1a is not practical as it requires several clock frequencies to tune the overall filter response.

Mode 1a is a good choice when Butterworth filters are desired since they have poles in a circle with the same f₀. Figure 31 shows an example of a 4th order 100kHz low-pass Butterworth filter clocked at 5MHz.

A monotonic passband response with a smooth transition band results, showing the circuit's low sensitivity, even though 1% resistors are used which result in an approximate value of Q.

Mode 1, Figure 15, provides a clock tunable notch. Mode 1 is a practical configuration for second order clock tunable bandpass/notch filters. In mode 1, a bandpass output with a very high Q, together with unity gain, can be obtained with the dynamics of the remaining notch and lowpass outputs. Figure 32 is an example of a 4th order bandpass filter implemented by cascading 2 sections each with a Q of 10. This figure shows the amplitude response when f_{CLK} = 7.5MHz resulting in a center frequency of 150kHz and a Q of 15.5.

Modes 1b and 1c, Figures 17, 18 are similar. They both produce a notch with a frequency which is always equal to the filter center frequency. The notch and the center frequency can be adjusted with an external resistor ratio.

The clock to center frequency ratio range is:

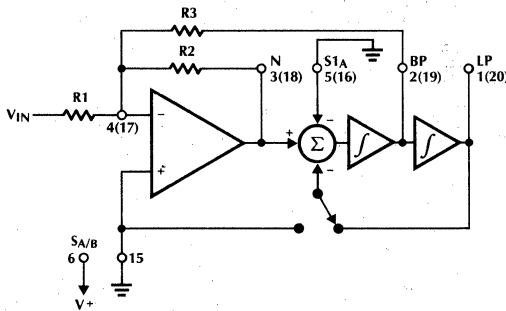
$$\frac{500}{1} \geq \frac{f_{CLK}}{f_0} \geq \frac{100}{1} \text{ or } \frac{50}{1}; \text{ mode 1c}$$

$$\frac{100}{1} \text{ or } \frac{50}{1} \geq \frac{f_{CLK}}{f_0} \geq \frac{100}{\sqrt{2}} \text{ or } \frac{50}{\sqrt{2}}; \text{ mode 1b}$$

The input impedance of the S1 pin is clock dependent, and in general R5 should not be larger than 5k for f_{CLK} < 2.5MHz and 1k to 2k for f_{CLK} > 2.5MHz. Mode 1b can be used to increase the clock to center frequency ratio beyond 100:1. For this mode, the limit for the (f_{CLK}/f₀) ratio is 500:1. Beyond this, the filter will exhibit large output offsets. Mode 1d, Figure 19, is the fastest mode of operation: In the 50:1 mode center frequencies beyond 150kHz can easily be achieved. Figure 33 is an example using mode 1d of a 4th order filter where each section has a Q of 1 independent of resistor ratios. In this mode the input amplifier is outside the damping (Q) loop. Therefore, its finite bandwidth does not degrade the response at high frequency. This allows the amplifier to be used as an anti-aliasing and continuous smoothing filter by placing a capacitor across R2.

Modes 2, 2a, and 2b have a notch output which frequency, f_n, can be tuned independently from the center frequency, f₀. For all cases, however, f_n < f₀. These modes are useful when cascading second order functions to create an overall elliptic highpass, bandpass or notch response. The input amplifier and its feedback resistors (R2/R4) are now part of the resonant loop. Because of this, mode 2 and its derivatives are slower than mode 1's.

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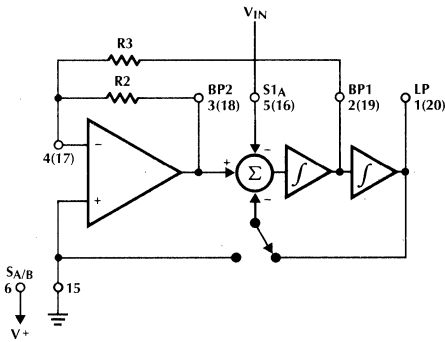
$$f_0 = \frac{f_{CLK}}{100(50)}; f_n = f_0; H_{OLP} = -\frac{R2}{R1}; H_{OBP} = -\frac{R3}{R1}; H_{ON1} = -\frac{R2}{R1}; Q = \frac{R3}{R2}$$

NOTE: ELECTRICAL TESTING IS PERFORMED WITH FOLLOWING RESISTOR VALUES:
Q = 1 WITH R1 = R2 = R3 = 2k
Q = 10 WITH R1 = R3 = 20k, R2 = 2k

Figure 15. Mode 1: 2nd Order Filter Providing Notch, Bandpass, Lowpass

OPERATION MODES (Continued)

1/2 ML2111

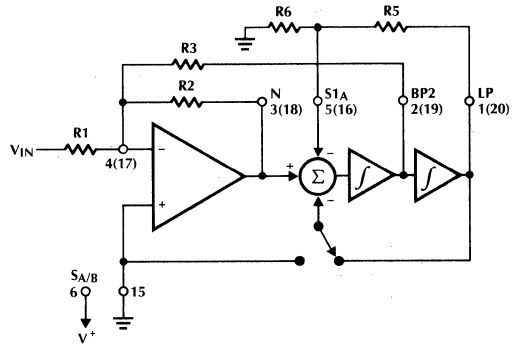


$$f_0 = \frac{f_{CLK}}{100(50)}; Q = \frac{R_3}{R_2}; H_{OBP1} = -\frac{R_3}{R_2}$$

$$H_{OBP2} = 1(\text{NONINVERTING}); H_{OLP} = -1$$

Figure 16. Mode 1a: 2nd Order Filter Providing Bandpass, Lowpass

1/2 ML2111

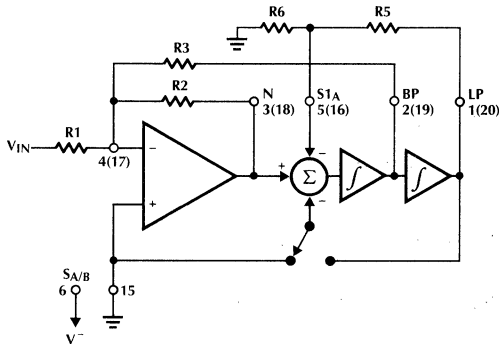


$$f_0 = \frac{f_{CLK}}{100(50)} \sqrt{1 + \frac{R_6}{R_5 + R_6}}; f_n = f_0; Q = \frac{R_3}{R_2} \sqrt{1 + \frac{R_6}{R_5 + R_6}}$$

$$H_{ON1}(f=0) = H_{ON2} \left(f - \frac{f_{CLK}}{2} \right) = -\frac{R_2}{R_1}; H_{OBP} = -\frac{R_3}{R_1}; H_{OLP} = \frac{-R_2/R_1}{1 + R_6/(R_5 + R_6)}; R_5 < 5k\Omega$$

Figure 17. Mode 1b: 2nd Order Filter Providing Notch, Bandpass, Lowpass

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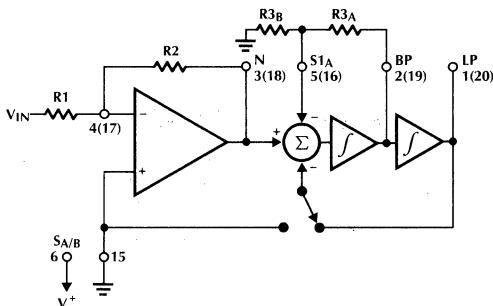


$$f_0 = \frac{f_{CLK}}{100(50)} \sqrt{\frac{R_6}{R_5 + R_6}}; f_n = f_0; Q = \frac{R_3}{R_2} \sqrt{\frac{R_6}{R_5 + R_6}}$$

$$H_{ON1}(f=0) = H_{ON2} \left(f - \frac{f_{CLK}}{2} \right) = -\frac{R_2}{R_1}; H_{OLP} = \frac{-R_2/R_1}{R_6/(R_5 + R_6)}; H_{OBP} = -\frac{R_3}{R_1}; R_5 < 5k\Omega$$

Figure 18. Mode 1c: 2nd Order Filter Providing Notch, Bandpass, Lowpass

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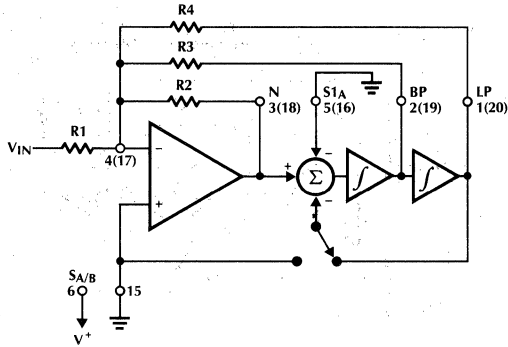
$$f_0 = \frac{f_{CLK}}{100(50)}; Q = 1 + \frac{R_{3A}}{R_{3B}}; H_{OBP} = -\frac{R_2}{R_1} \times Q$$

$$H_{OLP} = -\frac{R_2}{R_1}; V_N = -\frac{R_2}{R_1} V_{IN}$$

Figure 19. Mode 1d: 2nd Order Filter Providing Bandpass and Lowpass for Qs Greater or Equal to 1

OPERATION MODES (Continued)

1/2 ML2111

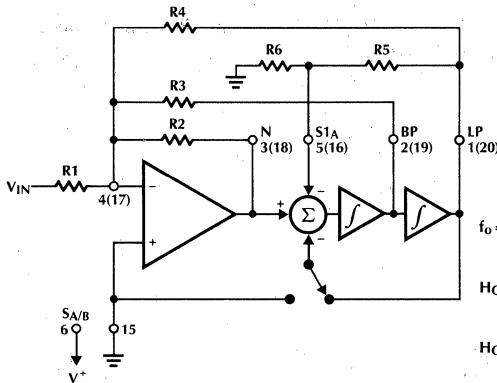


$$f_o = \frac{f_{CLK}}{100(50)} \sqrt{1 + \frac{R2}{R4}}; f_n = \frac{f_{CLK}}{100(50)}; Q = \frac{R3}{R2} \sqrt{1 + \frac{R2}{R4}}; H_{OLP} = \frac{-R2/R1}{1 + (R2/R4)}$$

$$H_{OBP} = -R3/R1; H_{ON1}(f=0) = \frac{-R2/R1}{1 + (R2/R4)}; H_{ON2} \left(f = \frac{f_{CLK}}{2} \right) = -R2/R1$$

Figure 20. Mode 2: 2nd Order Filter Providing Notch, Bandpass, Lowpass

1/2 ML2111



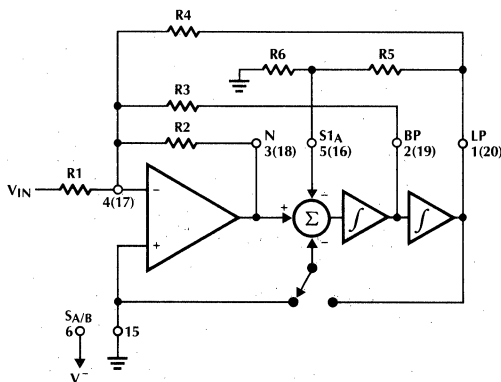
$$f_o = \frac{f_{CLK}}{100(50)} \sqrt{1 + \frac{R2}{R4} + \frac{R6}{R5 + R6}}; f_n = \frac{f_{CLK}}{100(50)} \sqrt{1 + \frac{R6}{R5 + R6}}; Q = \frac{R3}{R2} \sqrt{1 + \frac{R2}{R4} + \frac{R6}{R5 + R6}}$$

$$H_{ON1}(f=0) = -\frac{R2}{R1} \left[\frac{1 + R6/(R5 + R6)}{1 + (R2/R4) + [R6/(R5 + R6)]} \right]; H_{ON2} \left(f = \frac{f_{CLK}}{2} \right) = -R2/R1$$

$$H_{OBP} = -R3/R1; H_{OLP} = \frac{-R2/R1}{1 + (R2/R4) + [R6/(R5 + R6)]}$$

Figure 21. Mode 2a: 2nd Order Filter Providing Notch, Bandpass, Lowpass

1/2 ML2111



$$f_o = \frac{f_{CLK}}{100(50)} \sqrt{\frac{R2}{R4} + \frac{R6}{R5 + R6}}; f_n = \frac{f_{CLK}}{100(50)} \sqrt{\frac{R6}{R5 + R6}}; Q = \frac{R3}{R2} \sqrt{\frac{R2}{R4} + \frac{R6}{R5 + R6}}$$

$$H_{ON1}(f=0) = -\frac{R2}{R1} \left[\frac{R6/(R5 + R6)}{(R2/R4) + [R6/(R5 + R6)]} \right]; H_{ON2} \left(f = \frac{f_{CLK}}{2} \right) = -R2/R1$$

$$H_{OBP} = -R3/R1; H_{OLP} = \frac{-R2/R1}{(R2/R4) + [R6/(R5 + R6)]}$$

Figure 22. Mode 2b: 2nd Order Filter Providing Notch, Bandpass, Lowpass

OPERATION MODES (Continued)

In mode 3, Figure 23, a single resistor ratio ($R2/R4$) can tune the center frequency below or above the $f_{CLK}/100$ (or $f_{CLK}/50$) ratio. Mode 3 is a state variable configuration since it provides a highpass, bandpass, lowpass output through progressive integration; notches are obtained by summing the highpass and lowpass outputs (modes 3a, Figure 24). The

notch frequency can be tuned below or above the center frequency through the resistor ratio (R_h/R_l). Because of this, modes 3 and 3a are the most versatile and useful modes for cascading second order sections to obtain high order elliptic filters with frequencies up to 30kHz.

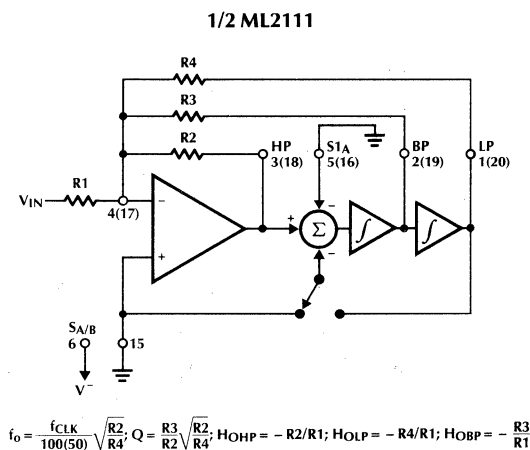


Figure 23. Mode 3: 2nd Order Filter Providing Highpass, Bandpass, Lowpass

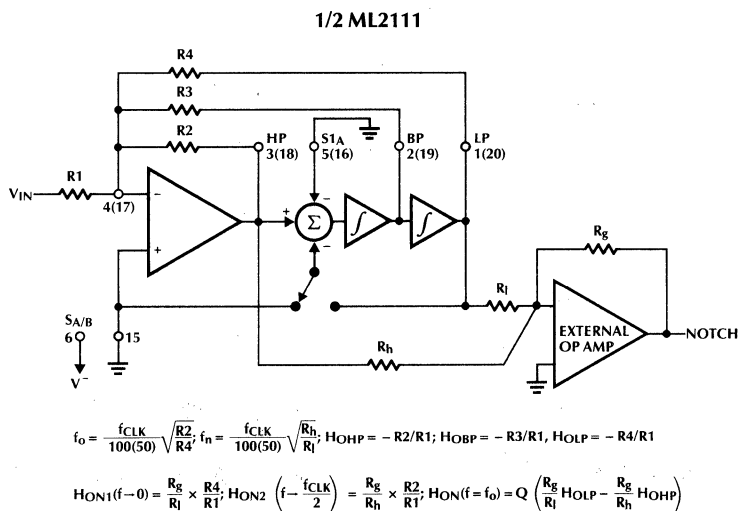
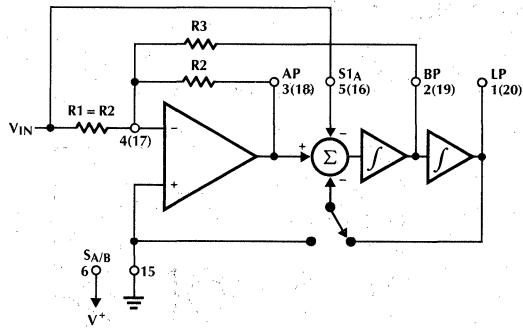


Figure 24. Mode 3a: 2nd Order Filter Providing Highpass, Bandpass, Lowpass, Notch

OPERATION MODES (Continued)

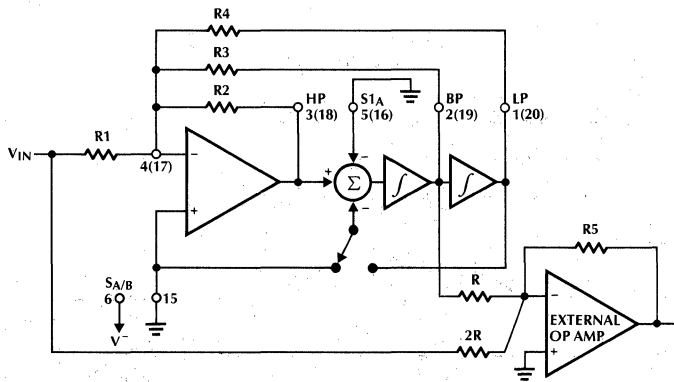
1/2 ML2111



$$f_0 = \frac{f_{CLK}}{100(50)}; Q = \frac{R_3}{R_2}; H_{OAP} = -\frac{R_2}{R_1}; H_{OLP} = -2; H_{OBP} = -2 \left(\frac{R_3}{R_2}\right)$$

Figure 25. Mode 4: 2nd Order Filter Providing Allpass, Bandpass, Lowpass

1/2 ML2111

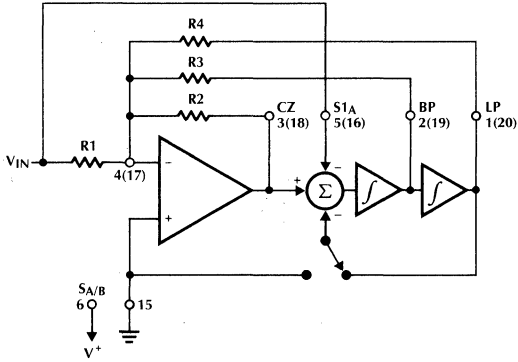


$$f_0 = \frac{f_{CLK}}{100(50)} \sqrt{\frac{R_2}{R_4}}; Q = \frac{R_3}{R_2} \sqrt{\frac{R_2}{R_4}}; H_{OAP} = \frac{R_5}{2R}; H_{OHP} = -\frac{R_2}{R_1}; H_{OBP} = -\frac{R_3}{R_1}; H_{OLP} = -\frac{R_4}{R_1}$$

Figure 26. Mode 4a: 2nd Order Filter Providing Highpass, Bandpass, Lowpass, Allpass

OPERATION MODES (Continued)

1/2 ML2111



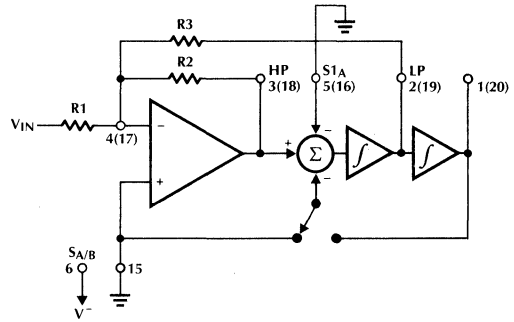
$$f_o = \frac{f_{CLK}}{100(50)} \sqrt{1 + \frac{R_2}{R_4}}; f_z = \frac{f_{CLK}}{100(50)} \sqrt{1 - \frac{R_1}{R_4}}; Q = \frac{R_3}{R_2} \sqrt{1 + \frac{R_2}{R_4}}$$

$$QZ = \frac{R_3}{R_1} \sqrt{1 - \frac{R_1}{R_4}}; HOZ(f=0) = \frac{(R_3/R_1) - 1}{(R_4/R_2) + 1}; HOZ(f = \frac{f_{CLK}}{2}) = \frac{R_2}{R_1}$$

$$HOBP = \frac{R_3}{R_2} \left(1 + \frac{R_2}{R_1}\right); HOLP = \frac{1 + (R_2/R_1)}{1 + (R_2/R_4)}$$

Figure 27. Mode 5: 2nd Order Filter Providing Numerator Complex Zeros, Bandpass, Lowpass

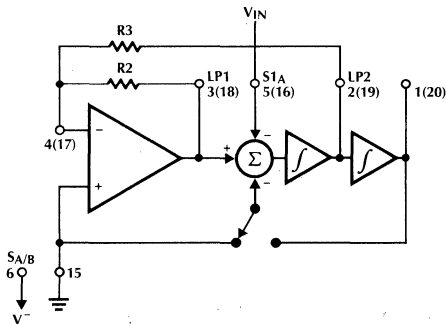
1/2 ML2111



$$f_c = \frac{f_{CLK} R_2}{100(50) R_3}; HOLP = -R_3/R_1; HOHP = -R_2/R_1$$

Figure 28. Mode 6a: 1st Order Filter Providing Highpass, Lowpass

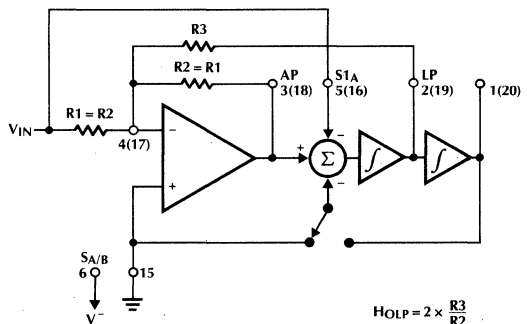
1/2 ML2111



$$f_c = \frac{f_{CLK} R_2}{100(50) R_3}; HOLP1 = 1; HOLP2 = -\frac{R_3}{R_2}$$

Figure 29. Mode 6b: 1st Order Filter Providing Lowpass

1/2 ML2111



$$f_p = \frac{f_{CLK} R_2}{100(50) R_3}; f_z = \frac{f_{CLK} R_2}{100(50) R_3}; \text{GAIN AT OUTPUT} = 1 \text{ FOR } 0 \leq f \leq \frac{f_{CLK}}{2}$$

Figure 30. Mode 7: 1st Order Filter Providing Allpass, Lowpass

OPERATION MODES (Continued)

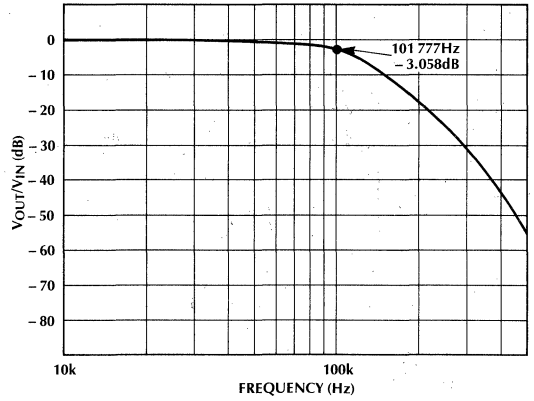
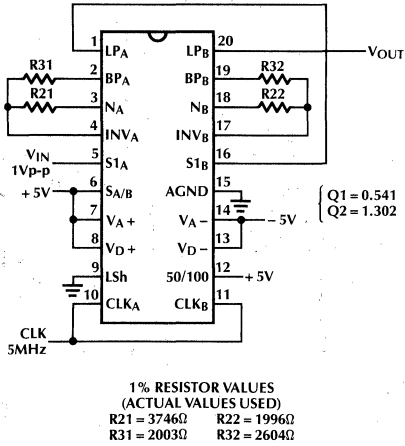


Figure 31. 4th Order, 100kHz Lowpass Butterworth Filter Obtained by Cascading 2 Sections in Mode 1a

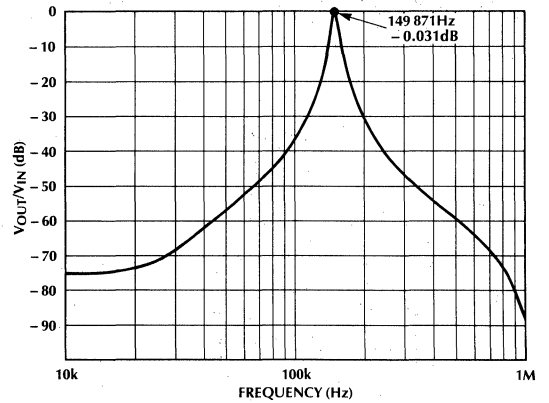
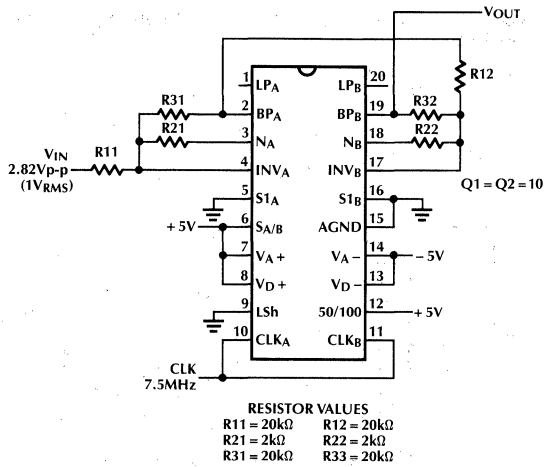


Figure 32. Cascading Two Sections in Mode 1, Each With $Q = 10$ Results in a Bandpass Filter with $Q = 15.5$ and $f_o = 150\text{kHz}$ ($f_{CLK} = 7.5\text{MHz}$)

OPERATION MODES (Continued)

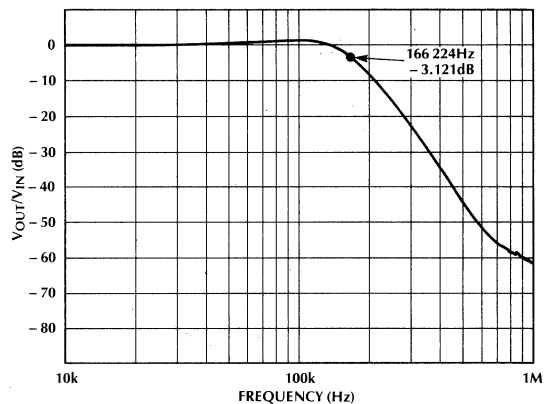
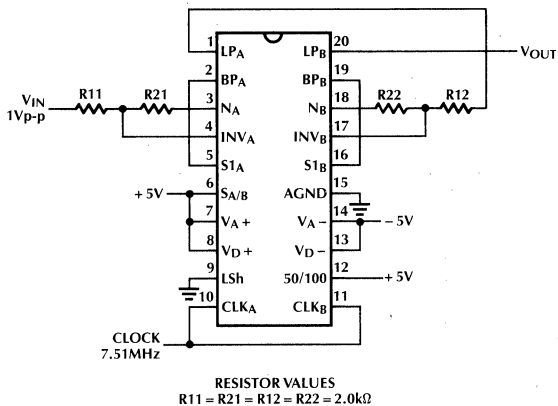


Figure 33. Cascading Two Sections in Mode 1d, Each With $Q = 1$ (Independent of Resistor Ratios) Creates a Sharper 4th Order Lowpass Filter

3

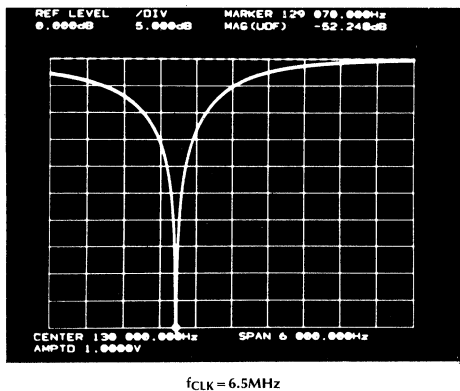
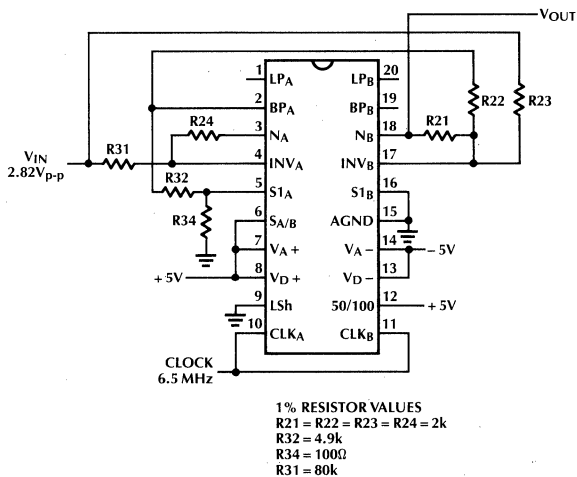


Figure 34. Notch Filter with $Q = 50$ and $f_o = 130kHz$. This Circuit Uses Side A's Biquad in Mode 1d and the Side B Op Amp to Create a Notch Whose Depth is Controlled by R31. The Notch is Created by Subtracting the Bandpass from V_{IN} . The Bandpass of Side A is Subtracted Using the Op Amp of Side B.

ML2111

OFFSETS

Switched capacitor integrators generally exhibit higher input offsets than discrete R, C integrators.

These offsets are mainly the charge injection of the CMOS switches into the integrating capacitors. The internal op amp offsets also add to the overall budget.

Figure 35 shows half of the ML2111 filter with its equivalent input offsets V_{OS1} , V_{OS2} , V_{OS3} .

The DC offset at the filter bandpass output is always equal to V_{OS3} . The DC offsets at the remaining two outputs

(Notch and LP) depend on the mode of operation and external resistor ratios. Table 3 illustrates this.

It is important to know the value of the DC output offsets, especially when the filter handles input signals with large dynamic range. As a rule of thumb, the output DC offsets increase when:

1. The Q's decrease
2. The ratio (f_{CLK}/f_o) increases beyond 100:1. This is done by decreasing either the ($R2/R4$) or the ($R6/(R5 + R6)$) resistor ratios.

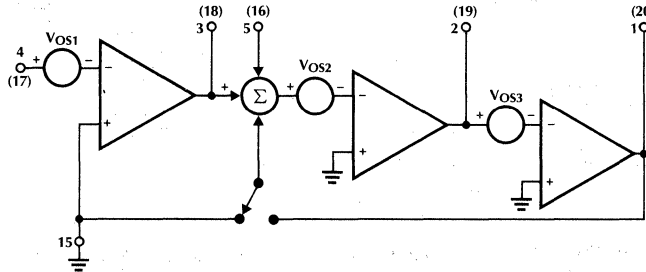


Figure 35. Equivalent Input Offsets of 1/2 ML2111 Filter

Table 3

MODE	V_{OSN} PIN 3 (18)	V_{OSBP} PIN 2 (19)	V_{OSLP} PIN 1 (20)
1,4	$V_{OS1}[(1/Q) + 1 + H_{OLP}] - V_{OS3}/Q$	V_{OS3}	$V_{OSN} - V_{OS2}$
1a	$V_{OS1}[1 + (1/Q)] - V_{OS3}/Q$	V_{OS3}	$V_{OSN} - V_{OS2}$
1b	$V_{OS1}[(1/Q) + 1 + R2/R1] - V_{OS3}/Q$	V_{OS3}	$\sim (V_{OSN} - V_{OS2})(1 + R5/R6)$
1c	$V_{OS1}[(1/Q) + 1 + R2/R1] - V_{OS3}/Q$	V_{OS3}	$\sim (V_{OSN} - V_{OS2}) \frac{(R5 + R6)}{(R5 + 2R6)}$
1d	$V_{OS1}[1 + R2/R1]$	V_{OS3}	$V_{OSN} - V_{OS2} - V_{OS3}/Q$
2, 5	$[V_{OS1}(1 + R2/R1 + R2/R3 + R2/R4) - V_{OS3}(R2/R3)] \times [R4/(R2 + R4)] + V_{OS2}[R2/(R2 + R4)]$	V_{OS3}	$V_{OSN} - V_{OS2}$
2a	$[V_{OS1}(1 + R2/R1 + R2/R3 + R2/R4) - V_{OS3}(R2/R3)] \times \left[\frac{R4(1+k)}{R2 + R4(1+k)} \right] + V_{OS2} \left[\frac{R2}{R2 + R4(1+k)} \right]; k = \frac{R6}{R5 + R6}$	V_{OS3}	$\sim (V_{OSN} - V_{OS2}) \frac{(R5 + R6)}{(R5 + 2R6)}$
2b	$[V_{OS1}(1 + R2/R1 + R2/R3 + R2/R4) - V_{OS3}(R2/R3)] \times \left[\frac{R4k}{R2 + R4k} \right] + V_{OS2} \left[\frac{R2}{R2 + R4k} \right]; k = \frac{R6}{R5 + R6}$	V_{OS3}	$\sim (V_{OSN} - V_{OS2})(1 + R5/R6)$
3, 4a	V_{OS2}	V_{OS3}	$V_{OS1} \left[1 + \frac{R4}{R1} + \frac{R4}{R2} + \frac{R4}{R3} \right] - V_{OS2} \left(\frac{R4}{R2} \right) - V_{OS3} \left(\frac{R4}{R3} \right)$

ORDERING INFORMATION

PART NUMBER	TEMP. RANGE	PACKAGE
ML2111BCP	0°C to +70°C	MOLDED DIP (P20)
ML2111CCP	0°C to +70°C	MOLDED DIP (P20)
ML2111BCS	0°C to +70°C	MOLDED SOIC (S20W)
ML2111CCS	0°C to +70°C	MOLDED SOIC (S20W)
ML2111BJJ	-40°C to +85°C	HERMETIC DIP (J20)
ML2111CIJ	-40°C to +85°C	HERMETIC DIP (J20)

PART NUMBER	TEMP. RANGE	PACKAGE
ML2111BMJ	-55°C to +125°C	HERMETIC DIP (J20)
ML2111CMJ	-55°C to +125°C	HERMETIC DIP (J20)
ML2111BJP	-40°C to +85°C	MOLDED DIP (P20)
ML2111CIP	-40°C to +85°C	MOLDED DIP (P20)
ML2111BIS	-40°C to +85°C	MOLDED SOIC (S20W)
ML2111CIS	-40°C to +85°C	MOLDED SOIC (S20W)

Data Communications

Section 4

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10BASE-T Tranceivers

Part Number	Applications	LED Outputs	Autopolarity	Package Types
ML4652	Internal MAU External MAU	6	No	24 Pin Skinny DIP 28 Pin PLCC
ML4658	Internal MAU External MAU	6	Yes	24 Pin Skinny DIP 28 Pin PLCC
ML4654	HUB MAU	5	Yes	20 Pin Skinny DIP 28 Pin PLCC
ML2652	PC Card	7	Yes	44 Pin PLCC 44 Pin TQFP

Fiber Optic Quantizers and LED Drivers

Part Number	I/O Types	Bandwidth	Package Types
ML4621	TTL, ECL	50 MHz	24 Pin Skinny DIP 28 Pin PLCC
ML4622	TTL, ECL	40 MHz	16 Pin Skinny DIP 16 Pin Narrow SOIC
ML4624	TTL, ECL	40 MHz	24 Pin Skinny DIP 28 Pin PLCC
ML6622	ECL	150 MHz	16 Pin DIP 16 Pin Narrow SOIC
ML4632	TTL, ECL	20 MHz	14 Pin DIP 16 Pin SOIC
ML6632	ECL, TTL	150 MHz	8 Pin DIP 8 Pin SOIC

AUI Multiplexers

Part Number	Number of Channels	Cascadeable	Package Type
ML4642	2	Yes	28 Pin SSOP

Token Ring Physical Interface

Part Number	Application	Data Rates	Cable Type	Package Type
ML6682	Station Concentrator	4 Mbps 16 Mbps	UTP STP	44 Pin TQFP

SONET/FDDI

Part Number	Bandwidth	Cable Type	Package
ML6622 Quantizer	150 MHz	Fiber	16 Pin DIP 16 Pin Narrow SOIC
ML6632 LED Driver	150 MHz	Fiber	8 Pin DIP 8 Pin SOIC
ML6671 MLT-3 Transceiver	80 MHz	Shielded Twisted Pair Data Grade Twisted Pair Category 5	28 Pin SSOP

Evaluation Kits

Part Number	Kit Includes	I/O	LED Outputs
ML4661EVAL ML4662EVAL ML4663EVAL	PC Board ML Samples HFBR2416 HFBR1414 Users Guide AUI Connectors Transformers	AUI Connector Fiber Optic ST Connector	Transmit Receive Jabber Collision Link Detect Power

10Base-T Physical Interface Chip

GENERAL DESCRIPTION

The ML2652, 10BASE-T Physical Interface Chip, is a complete physical interface for twisted pair and AUI Ethernet applications. It combines a 10BASE-T MAU, Manchester Encoder/Decoder, and Twisted Pair Interface filters in one monolithic IC. A complete DTE interface for twisted pair Ethernet can be implemented by combining the ML2652, an Ethernet controller, and transformers.

The ML2652 can automatically select between an AUI and twisted pair interface based on Link Pulses. The AUISEL status LED reflects which port is active. Receive Polarity can also be automatically corrected. Seven LED outputs provide complete status at the physical link. Link Test and Receive Polarity can be enabled or disabled through the LED outputs.

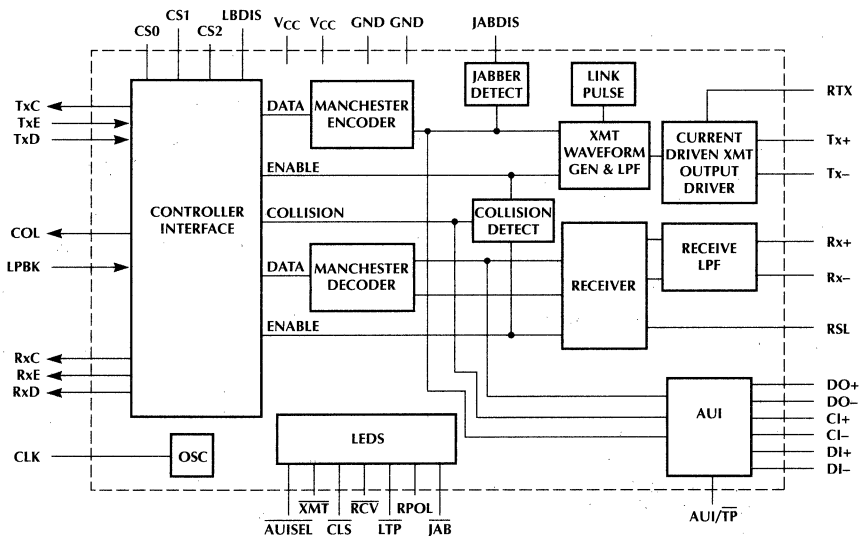
The unique transmitter design uses a waveform generator and low pass filter to meet the 10BASE-T transmitter requirements without the need for an external filter. The differential current driven output reduces common mode which in turn results in very low EMI and RFI noise.

The ML2652 is implemented in a low power double polysilicon CMOS technology.

FEATURES

- Complete physical interface solution
- Conforms to IEEE 802.3i-1990 (10Base-T)
- On-chip transmit and receive filters
- AUI and Twisted Pair interface
- Automatic AUI/Twisted Pair selection
- Power down mode
- Pins selectable controller interface-
 - Intel 82586, 82596
 - NSC DP8390
 - Seqq 8003, 8005
 - AMD 7990
 - Fujitsu MB86950
- Automatic polarity correction
- Pin selectable receive squelch levels
- Status pins for: polarity, link detect, receive & transmit activity, collision, jabber, AUI selection
- Single supply 5V \pm 5%
- 44 pin PCC package

BLOCK DIAGRAM



ML2652

PIN DESCRIPTION

NAME	FUNCTION
VCC	Positive supply. +5V
GND	Ground. 0 volts. All inputs and outputs referenced to this point.
CLK	Clock input. There must be either a 20 MHz crystal or a 20 MHz clock between this pin and GND.
Tx+	Transmit positive twisted pair output. This output is a current source that drives the twisted pair cable through a pulse transformer.
Tx-	Transmit negative twisted pair output. This output is a current source that drives the twisted pair cable through a pulse transformer.
Rx+	Receive positive twisted pair input. This input receives data from the twisted pair cable through a pulse transformer.
Rx-	Receive negative twisted pair input. This input receives data from the twisted pair cable through a pulse transformer.
DO+	AUI positive transmit output. AUI transmit data output to optional external transceiver.
DO-	AUI negative transmit output. AUI transmit data output to optional external transceiver.
DI+	AUI positive receive data input from optional external transceiver.
DI-	AUI negative receive data input from optional external transceiver.
CI+	AUI positive collision input from optional external transceiver.
CI-	AUI negative collision input from optional external transceiver.
RTX	Transmit current set. An external resistor between this pin and GND programs the absolute value of output current on Tx±.
TxC	Transmit clock output. Digital output which clocks the transmit data (TxD) into the device from the controller.
TxD	Transmit data input. Digital input which contains transmit data from the controller.
TxE	Transmit enable input. Digital input from the controller that indicates when the transmit data (TxD) is valid.
COL	Collision output. Digital output to the controller which indicates when a collision condition is present.
RxC	Receive clock output. Digital output which clocks receive data (RxD) from the device into the controller.

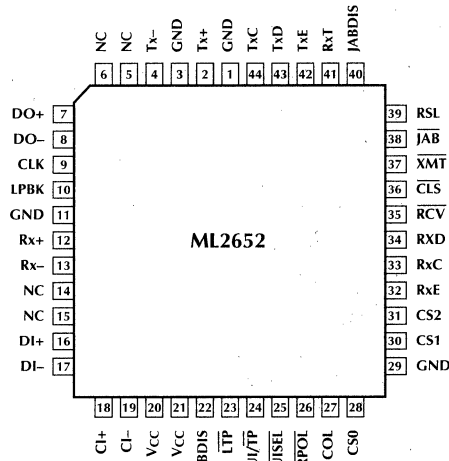
NAME	FUNCTION
RxD	Receive data output. Digital output which contains receive data sent to the controller.
RxE	Receive data valid. Digital output to the controller that indicates when the receive data (RxD) is valid.
LPBK	Local loopback. Digital input from the controller which forces the device to loopback transmit data without sending it on the media.
LBKDIS	Disables AUI loopback and collision detection. Allows loopback through R145 for diagnostic purposes. High = Loopback Disabled, Low = normal operation.
CS0	Controller selection input. Digital input which selects one of 5 standard controller timing interfaces. This pin has an internal pulldown resistor to GND.
CS1	Controller select input. Digital input which selects one of five standard controller timing interfaces. This pin has an internal pulldown resistor to GND.
CS2	Controller select input. Digital input which selects one of 5 standard controller timing interfaces. This pin has an internal pulldown resistor to GND.
RSL	Receive squelch level select input. Pin has internal pullup resistor to VCC. RSL = High Receive squelch level = 10Base-T RSL = Low Receive squelch level = extended distance
XMT	Transmit status output. Digital output which indicates data transmission on Tx+ and Tx-. Pin is open drain output with resistor pullup and is capable of driving an LED.
RCV	Receive status output. Digital output which indicates unsquelched data reception on Rx+ and Rx-. Pin is an open drain output with resistor pullup and is capable of driving an LED.
CLS	Collision status output. Digital output which indicates that collision condition has been detected. Pin is an open drain output with resistor pullup and is capable of driving an LED.
LTP	Link test pass output/input. This pin consists of an open drain output transistor with a resistor pullup that serves both as a link test pass output and a link test disable input. When used as an output, this pin is capable of driving an LED. LTP = High, link test failed LTP = Low, link test pass LTP = GND, link test disabled

PIN DESCRIPTION (Continued)

NAME	FUNCTION
AUI/TP	AUI/twisted pair interface select input. AUI/TP = High, AUI selected AUI/TP = Low, TP selected
RPOL	Receive polarity status output/input. This pin consists of an open drain output transistor with a resistor pullup that serves both as a receive polarity status output and as an automatic polarity correction input. When used as an output, this pin is capable of driving an LED. RPOL = High, receive polarity reversed RPOL = Low, receive polarity correct RPOL = GND, auto polarity correction enabled

NAME	FUNCTION
JAB	Jabber detect output. Digital output which indicates that the jabber condition has been detected. Pin is an open drain output with resistor pullup and is capable of driving a LED. JAB = High, normal JAB = Low, jabber detected
AUISEL	AUI/TP port output status AUISEL = High, TP port selected AUISEL = Low, AUI port selected
JABDIS	Jabber disable input JABDIS = High, jabber disabled JABDIS = Low, normal operation
NC	No connect. Leave this pin open circuit.

PIN CONNECTION



ML2652

ABSOLUTE MAXIMUM RATINGS

Absolute maximum ratings are limits beyond which the life of the integrated circuit may be impaired. All voltages unless otherwise specified are measured with respect to GND. (Note 1)

VCC supply voltage +6.5 volts
 All inputs and outputs -0.3v to VCC + .3v
 Input current per pin ± 25 mA
 Power dissipation 0.75 Watt
 Storage temperature range -65°C to +150°C
 Lead temperature (soldering, 10 sec) 300°C

ELECTRICAL CHARACTERISTICS

Unless otherwise specified $T_A = 0^\circ\text{C}$ to 70°C , $V_{CC} = 5V \pm 5\%$. Note 2 & 3.

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
VIL	Digital input low voltage	All except CLK CLK			.8 1.5	V V
VIH	Digital input high voltage	All except CLK CLK	2.0 3.5			V V
III	Digital input low current	VIN=GND TxD, TxE, AUI/ $\overline{\text{TP}}$ VIN=GND LPBK, CS2-0, LBDIS, JABDIS VIN=GND RSL VIN=GND LTP, RPOL, VIN=GND CLK	-10 -125	-25 -250	-1 -1 -50 -500 -250	μA μA μA μA μA
IIH	Digital input high current	VIN=VCC TxD, TxE, AUI/ $\overline{\text{TP}}$ VIN=VCC LPBK, CS2-0, LBDIS, JABDIS VIN=VCC RSL VIN=VCC LTP, RPOL VIN=VCC CLK	10	25	1 50 1 1 250	μA μA μA μA μA
CIN	Digital input capacitance	All except CLK CLK		5 10		pF pF
VOL	Digital output low voltage	IOL=-2mA TxC, COL, RxC, RxD, RxE IOL=-20mA XMT, RCV, CLS, LTP, RPOL, JAB			.4 .6	V V
VOH	Digital output high voltage	IOH=2mA TxC, COL, RxC, RxD, RxE IOL=10uA XMT, RCV, CLS, LTP, RPOL, JAB	4.0 2.4			V V
ICC	VCC supply current	TX transmission No transmission Powerdown mode			140 105 .1	mA mA mA
TOV	Tx \pm differential output voltage		2.2	2.5	2.8	Vp
THD	Tx \pm harmonic distortion	TxD=all ones	-27			dB
TCM	Tx \pm common mode output voltage				± 50	mVp
TCMR	Tx \pm common mode rejection	VCM=15vp, 10.1 MHz sine		± 100		mVp
TOVI	Tx \pm differential output voltage during idle				± 50	mVp

ELECTRICAL CHARACTERISTICS (Continued)

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
TOIA	Tx± output current accuracy	RTX=10K		50		mA
TRO	Tx± output resistance			1		Mohm
TCO	Tx± output capacitance			10		pF
RRI	Receive input resistance		5K	10K		ohms
RCI	Receive input capacitance			10		pF
RSON	Receive squelch on level	RSL=1 RSL=0	300 200		585 390	mVp mVp
RSOFF	Receive squelch off level	RSL=1 RSL=0	200 133		390 260	mVp mVp
DOV	DO± differential output voltage		± 550		± 1170	mV
DOVI	DO± differential output voltage during idle				± 40	mV
DOUS	DO± differential output voltage return to 0 undershoot				-100	mV
DOCMAC	DO± common mode AC output voltage				± 40	mV
DOCMD	DO± common mode DC output voltage			VCC *.5		V
DIRI	DI/CI input resistance		5K	10K		ohms
DICI	DI/CI input capacitance			10		pF
DIBV	DI/CI input bias voltage	DI/CI floating		VCC *.5		V
DISON	DI/CI squelch on level		-250		-375	mVp
DISOF	DI/CI squelch off level		-175		-275	mVp
t1	TxC on time		45		55	ns
t2	TxC off time		45		55	ns
t3	TxC period			100		ns
t4	TxE setup time		25			ns
t5	TxE hold time		0			ns
t6	TxD setup time		25			ns
t7	TxD hold time		0			ns
t8	Transmit propagation delay	Tx± DO±		60	200 200	ns ns
t9	Start of Idle Pulse Width	Tx± DO±	225		350	ns
t10	SOI pulse width to within 40mV of final value	Tx± DO±			4500 8000	ns ns

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ELECTRICAL CHARACTERISTICS (Continued)

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
t11	Transmit output jitter	Tx± DO±			± 8.0 ± .5	ns ns
t12	Transmit output rise and fall time	Tx± , 10–90%		5		ns
t13	TxE to $\overline{\text{XMT}}$ assert				200	ns
t14	$\overline{\text{XMT}}$ blinker pulse period		95		105	ms
t15	$\overline{\text{XMT}}$ duty cycle		45		55	%
t20	Start of receive packet to RxE assert	Rx± DI±			500 200	ns ns
t21	Start of receive packet to RxC active	Rx+ DI+			1600 1300	ns ns
t22	RxC on time		45		900	ns
t23	RxC off time		45		55	ns
t24	RxD valid before RxC		45			ns
t25	RxD valid after RxC		35			ns
t26	RxE assert to $\overline{\text{RCV}}$ assert				200	ns
t27	$\overline{\text{RCV}}$ blinker pulse period		95		105	ms
t28	$\overline{\text{RCV}}$ duty cycle		45		55	%
t29	Receive input jitter	Preamble Data			± 12 ± 18	ns ns
t30	Receive propagation delay	Tx± DI±			160 160	ns ns
t31	RxC to RxE assert		30		60	ns
t32	RxC to RxE deassert		20		40	ns
t33	RxE deassert to RxC switchover		100		200	ns
t34	Minimum SOI pulse width required for receive detection	Tx± DI±	180 180			ns ns
t40	Jabber activation delay- TxE assert to Tx± disable		20		150	ms
t41	Tx± disable to $\overline{\text{JAB}}$ assert			100		ns
t42	Jabber reset time – TxE deassert to $\overline{\text{JAB}}$ deassert		250		750	ms
t43	Tx± disable to COL assert			50		ns
t44	Tx± disable to $\overline{\text{CLS}}$ assert			50		ns
t45	$\overline{\text{JAB}}$ deassert to COL deassert				50	ns
t46	$\overline{\text{JAB}}$ deassert to $\overline{\text{CLS}}$ deassert				50	ns
t51	Transmit link pulse period		8		24	ms

ELECTRICAL CHARACTERISTICS (Continued)

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
t52	Minimum link pulse period required for receive detection		2		7	ms
t52	Maximum link pulse period required for receive detection		25		150	ms
t53	Receive link pulse no detect to $\overline{\text{LTP}}$ deassert		50		150	ms
t54	Receive link pulse detect to $\overline{\text{LTP}}$ assert		2			Link Pulse
t55	$\text{AUI}/\overline{\text{TP}}$ to $\overline{\text{AUISEL}}$ delay				200	ns
t60	TxE deassert to COL assert		.9	1.0	1.1	μs
t61	COL pulse Width		.9	1.0	1.1	μs
t70	Start of $\overline{\text{RCV}}$ packet during transmission to COL assert	Rx \pm DI \pm			500 200	ns ns
t71	Start of $\overline{\text{RCV}}$ packet during transmission to $\overline{\text{CLS}}$ assert	Rx \pm DI \pm			500 200	ns ns
t72	Start of $\overline{\text{RCV}}$ packet during transmission to RxD switching from Tx to Rx data	Rx \pm DI \pm			600 300	ns ns
t73	End of $\overline{\text{RCV}}$ packet during transmission to RxD switching from Rx to Tx data	Rx \pm DI \pm			350 350	ns ns
t74	End of $\overline{\text{RCV}}$ packet during transmission to COL deassert	Rx \pm DI \pm			250 250	ns ns
t75	$\overline{\text{CLS}}$ blinker pulse period		95		105	ms
t76	$\overline{\text{CLS}}$ duty cycle		45		55	%
t77	Transmission start during reception to COL assert	Tx \pm DO \pm			200 200	ns ns
t78	Transmission start during reception to $\overline{\text{CLS}}$ assert	Tx \pm DO \pm			200 200	ns ns
t79	CI \pm period		80		120	ns
t80	CI \pm duty cycle		40		60	%
t81	First valid negative CI \pm data transition to COL assert				100	ns
t82	First valid negative CI \pm data transition to $\overline{\text{CLS}}$ assert				100	ns
t83	Last CI \pm positive data transition to COL deassert		160		200	ns
t84	External clock input jitter				50	ps

Note 1: Absolute maximum ratings are limits beyond which the life of the integrated circuit may be impaired. All voltages unless otherwise specified are measured with respect to ground.

Note 2: Limits are guaranteed by 100% testing, sampling, or correlation with worst-case test conditions.

Note 3: Low Duty cycle pulse testing is performed at T_A .

TIMING DIAGRAMS

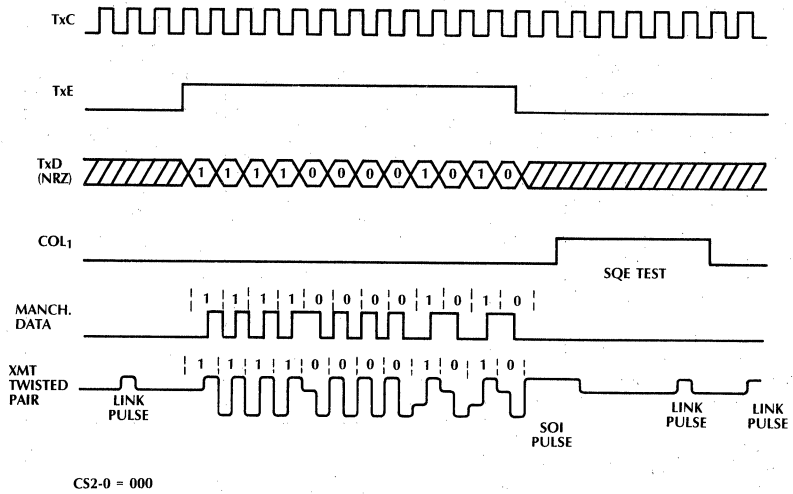


Figure 1. Transmit System Timing

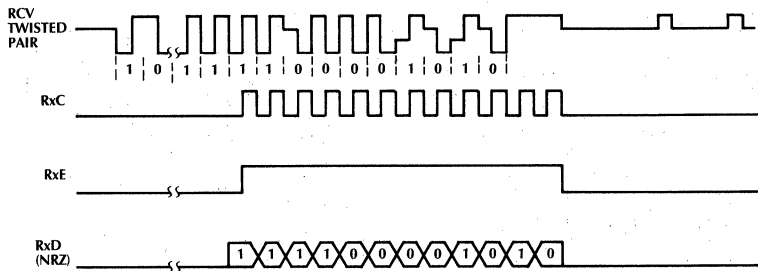


Figure 2. Receive Timing

TIMING DIAGRAMS

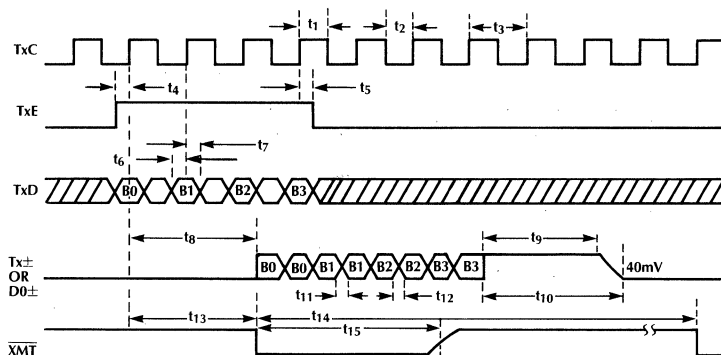
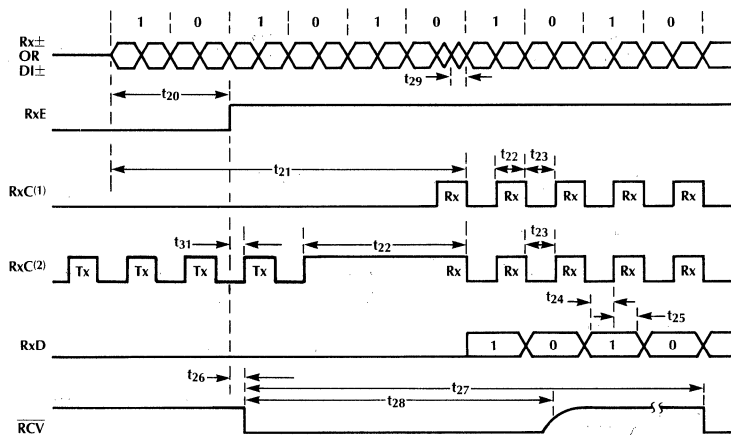


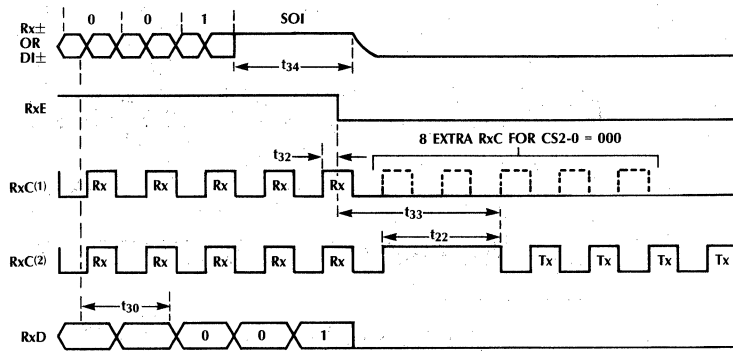
Figure 3. Transmit Timing



NOTE:
 1. RxC IS NOT CONTINUOUS DURING IDLE
 2. RxC IS CONTINUOUS DURING IDLE

Figure 4. Receive Timing – Start of Frame

TIMING DIAGRAMS



NOTE:
 1. RxC IS NOT CONTINUOUS DURING IDLE - 8 EXTRA CLOCKS ADDED FOR CS2-0 = 000
 2. RxC IS CONTINUOUS DURING IDLE

Figure 5. Receive Timing – End of Frame

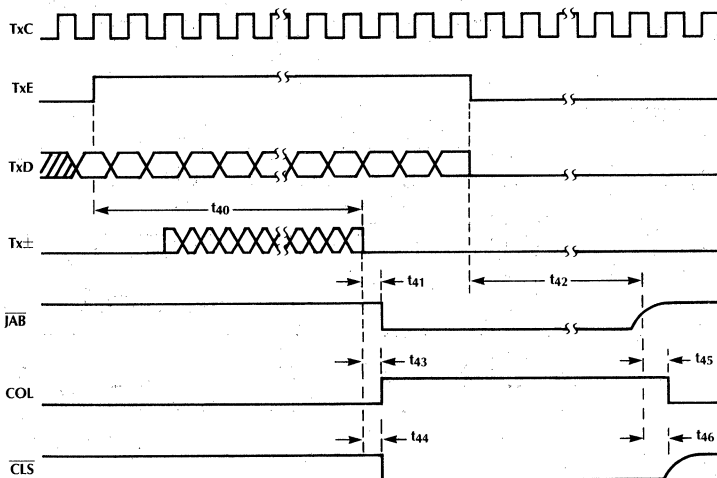


Figure 6. Jabber Timing

TIMING DIAGRAMS

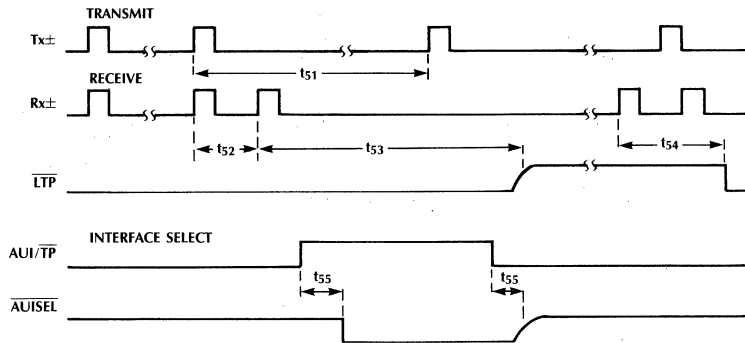


Figure 7. Link Pulse Timing

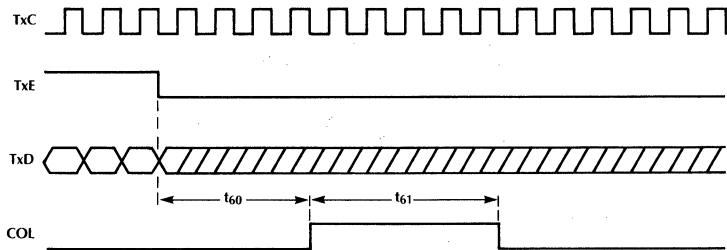


Figure 8. SQE Test Timing

TIMING DIAGRAMS

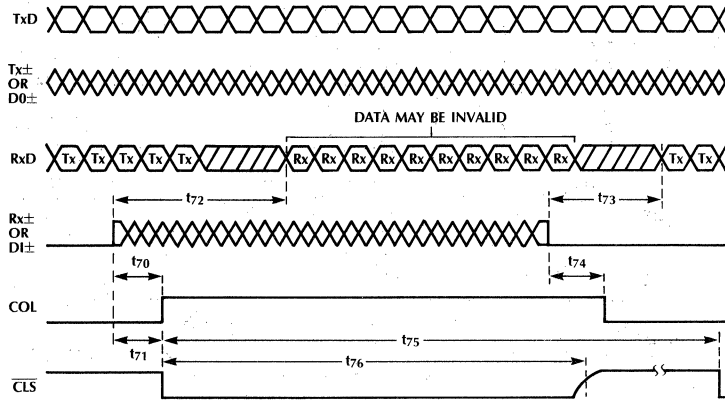


Figure 9. Collision Timing Reception During Transmission

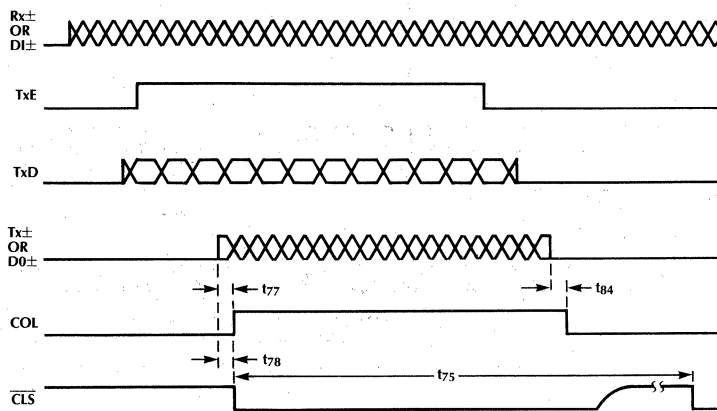


Figure 10. Collision Timing Transmission During Reception

TIMING DIAGRAMS

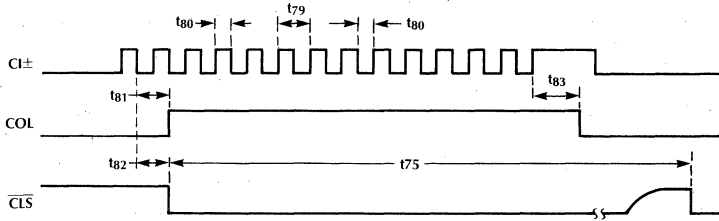


Figure 11. Cl± Collision

APPLICATION CIRCUIT

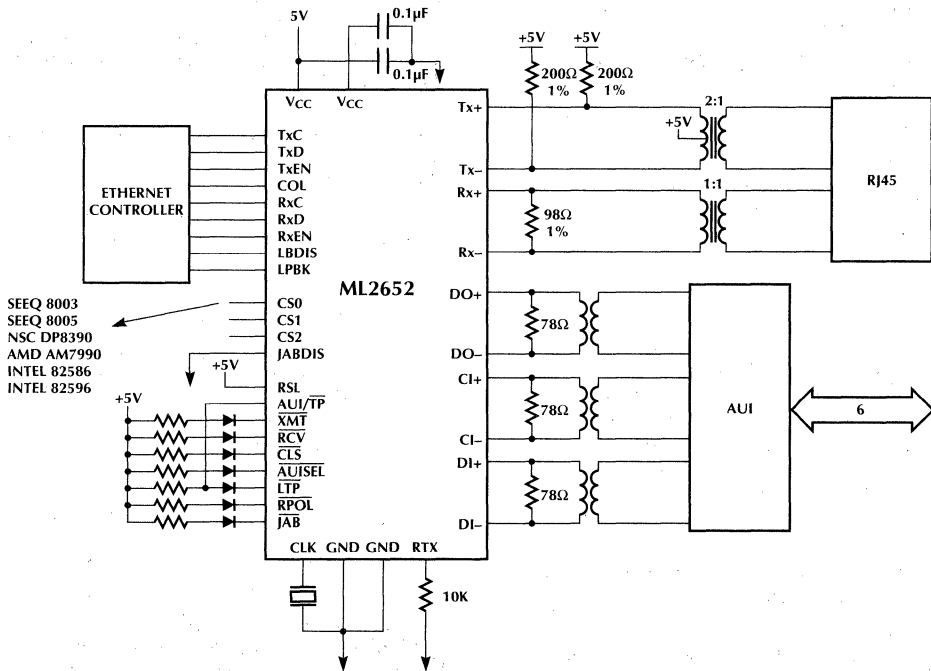


Figure 12

FUNCTIONAL DESCRIPTION

GENERAL

The ML2652 is composed of a transmitter section, receive section and some miscellaneous functions.

The transmit section consists of the manchester encoder, AUI, jabber detect, link pulse generator, start of idle (SOI) pulse generator, waveform generator, and line driver. The purpose of the transmit section is to take data from the controller, encode it, and transmit it over either the AUI or twisted pair interface. In addition, the transmit section generates link pulses, start of idle pulses, and checks for jabber condition. The transmitter keeps the data jitter to a maximum of ± 8.0 ns, and the maximum delay through the transmission section is less than 2 bits, or 200ns.

The receive section consists of the manchester decoder, collision detect, AUI, receive LPF, receive comparators, receive squelch, automatic polarity detect/correct, start of idle (SOI) detect, and link pulse detect. The purpose of the receive section is to take data from either the twisted pair cable or AUI, decode it, then send the data to the controller via the controller interface. In addition, the receive section detects and automatically corrects for reverse polarity, detects link pulses, detects start of idle pulses, and implements an intelligent receive squelch algorithm. The receive section can successfully lock onto an incoming data that contains ± 18 ns of jitter in less than 1.6 μ s.

The miscellaneous functions are the controller interface, single pin crystal oscillator, AUI, loopback modes, test mode, and powerdown mode.

The following text describes each of these blocks and functions in more detail. Refer to the block diagram.

TRANSMISSION

The transmit data (NRZ) is first clocked into the device through the controller interface. The device can be digitally programmed to accommodate any one of five standard Ethernet controllers as described in Controller section.

Then the NRZ data is encoded by the manchester encoder as shown in transmit timing diagram in Figure 1.

The manchester encoded data then goes to either the AUI or twisted pair interface. The selection of the appropriate interface is automatic. If the AUI is selected, the manchester encoded data is transmitted out differentially on the DO+ and DO- pins, and the twisted pair line driver is disabled. If the twisted pair interface is selected, the manchester encoded data is transmitted out differentially on Tx+ and Tx- pins, and the transmit AUI is disabled.

Refer to the AUI section for details on how the AUI and automatic interface selection is accomplished.

Assuming that the twisted pair interface is selected, the Manchester encoded data then goes to the transmit waveform generator. The transmit waveform generator

takes the digital Manchester encoded data and generates a waveform. When this waveform is passed through the cable model in the 10BASE-T standard (figure 14-7 IEEE Std 802.3i-1990) it meets the voltage template (figure 14-9 IEEE Std 802.3i-1990).

The transmit waveform generator is composed of a 16 x 4 bit ROM, 4 bit DAC, 3rd order LPF, and clock generator. The DAC is used to synthesize a stair-step representation of a signal that will meet the required output template. The ROM stores the digital representation of the output signal and provides a digital input to the DAC. The ROM is addressed by a 16 phase clock generator that is locked to the transmit clock TxC. The high frequency content present in the output of the DAC is removed by a 3rd order continuous LPF which smooths the output.

Since the output is synthesized by a DAC, the DAC can be reprogrammed with metal mask to provide any arbitrary output waveform, thus offering the ability to offer custom output waveforms and templates.

The transmit line driver takes the output of the waveform generator and converts this voltage to a differential output current on Tx+ and Tx- pins. When one transmit output (either Tx+ or Tx-) is sinking current, the other output is high impedance, and vice versa. In this way, a differential output voltage is developed by sinking this output current through two external 200 ohm terminating resistor and a 2:1 transformer as shown in Figure 12.

Setting the external terminating resistors to 200 ohms as shown in Figure 12 will implement a 100 ohm terminating impedance when looking back through the transformer. If other terminating impedances are required (such as 150 ohm), the terminating resistor values can be adjusted accordingly as long as the output current stays within the minimum and maximum limits (30-70mA).

The absolute value of the output current, and subsequently the output voltage level, is set by an external resistor between RTX and GND. If RTX = 10k ohms and Tx \pm is terminated as shown in Figure 12, the output level is ± 2.5 V which meets 802.3i-1990 differential output voltage requirements. If a different output current/voltage level is desired, the level can be changed by changing the value of RTX according to the following formula:

$$\begin{aligned} \text{RTX} &= K \cdot V_b / I_{out} \\ &= 125 \cdot 4v / 50mA \\ \text{RTX} &= 10k\Omega \end{aligned}$$

When data is being transmitted (and there is no collision or link pulse fail condition), the transmit data is looped back to the receive path, and the Manchester decoder will lock onto the transmit data stream.

After data transmission is completed, the transmitter sends a start of idle (SOI) pulse to signal the end of a packet. During the idle period, Tx+ and Tx- are held low. Occasionally, link pulses are transmitted during the idle period.

The $\overline{\text{XMT}}$ pin is an output that indicates transmit activity. The pin consists of an open drain output with an internal pull-up resistor and can drive an LED from VCC or another digital input. In order to make an LED visible, $\overline{\text{XMT}}$ has an internal blinker circuit that generates a 100ms blink (50ms high, 50ms low) that is triggered when a transmission starts. At the completion of the 100ms blink period, if a transmission is in progress, another 100ms blink is generated.

RECEPTION

The twisted pair receive data is typically transformer coupled and terminated with an external resistor as shown in Figure 12.

The output of the transformer is then applied to the device input pins Rx+ and Rx-. The input is differential, and the common mode input voltage is biased to VCC/2 by two internal 10K bias resistors from Rx+, Rx- to VCC/2.

The Rx+ and Rx- inputs then go to the receive filter. The receive filter is a continuous 3rd order LPF and has the following characteristics:

- | | |
|------------------------------|--------------|
| 1. 3 dB cut-off frequency | 15 MHz |
| 2. Insertion Loss (5–10 MHz) | 1.0 dB |
| 3. 30 MHz attenuation | 17.5 dB min. |

The output of the filter goes to the receive comparators. There are two receive comparators inside the chip, threshold and zero crossing. The threshold comparator determines if the receive data is valid by checking the input signal level against a predetermined positive and negative squelch level. Once the threshold comparator determines that valid data is being received, the zero crossing comparator senses zero crossings to determine data transitions. Both comparators are fast enough to respond to 12ns pulse widths with minimum squelch overdrive.

The receive squelch circuit determines when data on incoming Rx+, Rx- is valid. The receive squelch is considered "on" when the data is deemed to be invalid, and the receive squelch is considered "off" when data is determined to be valid.

The input signal must meet the following criteria in order to turn receive squelch off and be recognized as valid data:

1. The input signal must exceed the receive squelch on level. When this occurs, a 400ns squelch interval timer is started.
2. During the 400ns squelch interval, the input signal must go from one squelch threshold to the opposite polarity squelch threshold in less than 127ns.
3. During the 400ns squelch interval, the input signal has to make less than 9 squelch threshold to opposite polarity squelch threshold crossings.

When the receive squelch is turned off, the receive squelch off level is reduced to 2/3 of receive squelch on level.

The receive squelch will be turned back on if either the incoming data peaks go below the receive squelch off level for 400ns or the start of idle (SOI) pulse is detected.

The receive squelch on level can be digitally programmed for one of two possible levels by using the RSL pin. When RSL = 1, the squelch on level complies with the IEEE 802.3i-1990 specification. When RSL = 0, the receive squelch on level is lowered in order to accommodate greater receive attenuation and consequently longer twisted pair cable lengths. The receive squelch on level can be programmed as follows:

RSL	Receive Squelch On Level		
	Application	Min	Typ Max
1	10BASE-T	300	585mV
0	Long Distance	200	390mV

The $\overline{\text{RCV}}$ pin is an output that indicates receive activity. The pin consists of an open drain output with an internal pull-up resistor and can drive an LED from VCC or another digital input. In order to make an LED visible, $\overline{\text{RCV}}$ has an internal blinker circuit that generates a 100ms blink (50ms high, 50ms low) that is triggered when reception starts. At the completion of the 100ms blink period, if reception is in progress, another 100ms blink is generated.

The manchester decoder receives data from either the twisted pair interface (as described above) or the AUI (described in AUI section).

The manchester decoder is responsible for recovering clock and data from the incoming receive bit stream. Clock and data recovery is accomplished by a digital PLL which can lock on the incoming bit stream in less than 1.6 μ s.

The clock (RxC) and NRZ data (RxD) are then output to the external world via the controller interface.

SOI

A start of idle (SOI) pulse is sent at the end of transmission in order to signal to all receivers that transmission has ended and the idle period begins. Thus, the transmit section has an SOI generator and the receive section has an SOI detector.

The transmit SOI pulse generator inserts an SOI pulse at the end of each transmission. The SOI pulse is typically a 250ns positive pulse inserted after the last positive data transition. Depending on the data pattern, the positive data transition could occur either in the middle or at the end of the last bit cell. So the actual width of the transmitted SOI pulse can vary from 250–300ns, typically.

The receive SOI detector senses the SOI pulse using the zero crossing comparator. When the SOI pulse is detected, the receiver signals to the controller that receive data is no longer valid and turns the receive squelch on.

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LINK PULSE

During the idle period, link pulses are sent by the transmitter and detected by the receiver so that the integrity of the twisted pair link can be continuously monitored. Thus, the transmit section has a link pulse generator, and the receiver has a link pulse detector.

The transmit link pulse generator transmits a 100ns wide positive pulse (Tx+ high, Tx- low) every $16 \pm 8\text{ms}$.

IEEE 802.3i-1990 Section 14 requires the link pulse to be shaped to meet a template when passed or not passed through the twisted pair line model. The transmit waveform generator takes the link pulse and generates the waveform on TX± when passed or not passed through the twisted pair line model.

The receiver monitors the receive input to determine if the link pulses are present. When the device is in the link pulse pass state, normal packet transmission and reception can occur. All link pulses less than 2-7ms apart are ignored while in the link pass state. If no link pulses or receive packets are detected for a period of 50-150ms, the device goes into the link pulse fail state.

When the device is in the link pulse fail state, reception is inhibited and the transmitter is placed in the idle state (no data transmission but link pulses are still transmitted). In order for the device to exit the link pulse fail state, one complete packet or 4 consecutive link pulses must be detected, and transmit and receive must be idle.

Consecutive link pulses are defined as pulses that occur within 25-150ms of each other. If the link pulses occur 2-7ms apart in the link fail state, the device ignores the link pulses and resets the number of consecutive link pulses to zero. After the link pulse fail state is exited, transmission and reception can be resumed.

Link pulse status is indicated by the LTP pin. LTP is a dual function input/output pin that acts both as an active low link test pass output and a link test disable input. The pin consists of an open drain output with an internal pull-up resistor. If the pin is tied to GND, the pin acts as an input and the link test function is disabled. If the pin is not tied to GND, the pin acts as an active low link test pass output and can drive an LED from VCC or another digital output. Thus, the LED is lit when the link test is passing.

JABBER

The transmit section contains a jabber detect circuit. Jabber is a fault condition characterized by a babbling transmitter. The ML2652 detects jabber when a transmission packet exceeds 20-150ms in length. If jabber detect occurs, the transmit output is disabled, the collision signal COL is sent over the controller interface, and the JAB pin is pulled low. The device remains in the jabber detect state until there is at least 250-750ms of continuous non-transmission. Note that link pulses continue to be transmitted even when the device is in the jabber condition.

The jabber detection circuitry can be disabled with the JABDIS pin for testing and diagnostic purposes. Disabling

jabber means that a jabber condition is never recognized, even when it occurs. JABDIS is an active high jabber disable input and has an internal pull-down resistor to GND.

COLLISION

Collision occurs whenever the DTE card is transmitting and receiving data simultaneously. However, the collision circuit on the ML2652 operates differently depending on whether twisted pair interface or AUI is being used.

When the twisted pair interface is used, collision occurs whenever the device is transmitting and receiving data simultaneously, that is when both RxE and TxE are active. The collision state is indicated by COL and CLS pins. COL is used to signal collision to the controller. CLS is an active low open drain output. CLS is activated during Jabber, but not during SQE test while COL is activated during both.

When the AUI is used, collision is no longer detected from simultaneous transmission and reception, but the collision state is determined when a collision signal is present on the AUI collision inputs, CI+ and CI-. A 10 MHz square wave has to be applied to this input in order for the device to signal the collision state on COL and CLS.

The CLS pin is an output that indicates collision activity. The pin consists of an open drain output with an internal pull-up resistor and can drive an LED from VCC or another digital input. In order to make an LED visible, CLS has an internal blinker circuit that generates a 100ms blink (50ms high, 50ms low) that is triggered when a collision starts. At the completion of the 100ms blink period, if collision is in progress, another 100ms blink is generated.

SQE TEST

When the twisted pair interface is used, the device tests the collision circuitry at the end of each transmission by sending a 1µs collision pulse over the COL pin. This is known as SQE (signal quality error) test and is shown in the transmit timing diagram in Figure 1. The SQE test is disabled if the device is in jabber detect state or link pulse fail condition.

When AUI is used, the SQE test pulse is generated by an external MAU and the external MAU sends the SQE test pulse to the ML2652 via the collision inputs, CI+ and CI-. The ML2652 then relays the collision signal to the controller via the COL and CLS output pins.

RECEIVE POLARITY DETECT AND AUTO CORRECTION

The ML2652 contains an auto-polarity circuit that detects the polarity of the receive twisted pair leads, Rx+ and Rx- and internally reverses the leads if their polarity is incorrect.

When the device is powered up, it is assumed that the polarity is correct and no polarity correction occurs. Then receive polarity is continuously monitored by checking the polarity of the SOI and link pulses since they are always positive pulses. If either 2 consecutive SOI or 4

consecutive link pulses have incorrect $Rx\pm$ polarity, then the auto-polarity circuit internally reverses the $Rx+$ and $Rx-$ connections.

RPOL is a dual function input/output pin that acts both as an active low receive polarity status output and an auto-polarity circuit enable input. The pin consists of an open drain output with an internal pull-up resistor. If the pin is tied to GND, the pin acts as an input and the receive auto-polarity circuit is enabled. If the pin is not tied to GND, the pin acts as an active low receive polarity status output and can drive an LED from VCC or another digital output.

AUI

The ML2652 can be used with an external MAU via the Attachment Unit Interface (AUI). When the AUI is used, the internal MAU functions and twisted pair interface are disabled, and the device only uses the manchester encoder and decoder functions, as shown in the block diagram. The AUI consists of three differential signal pairs: DI, DO, and CI. The function of each pair is described below.

The DO+ and DO- are differential outputs to the external MAU which contain the transmit data output from the Manchester encoder. The DO+ and DO- output drivers are capable of driving 50 meters of 78 ohm cable with less than 5ns rise and fall time and less than ± 0.5 ns of jitter. In addition, at the end of transmission, the AUI output driver inserts a 200ns minimum pulse and meets the turnoff and idle characteristics specified in IEEE 802.3-1988. An external 78 ohm resistor across DO+ and DO- is required as shown in Figure 12 to develop the proper output levels from the internal current sources. The DO+ and DO- outputs can be coupled to an external MAU with either capacitors or a transformer. The ML2652 meets all AUI transmitter specifications outlined in IEEE 802.3-1988 Section 7.

DI+ and DI- are inputs from the external MAU which contain the receive data that goes to the manchester decoder.

The DI+ and DI- inputs contain an AUI DI squelch circuit which determines when incoming data on DI+ and DI- is valid. The DI squelch is considered "on" when the data is deemed to be invalid, and the DI squelch is considered "off" when data is determined to be valid.

The input signal on DI+ and DI- must meet the following criteria in order to turn receive squelch off and be recognized as valid data:

1. The input signal must exceed the negative AUI DI squelch on level.
2. The input signal must exceed the negative AUI DI squelch on level for more than 20ns.

When the DI squelch is turned off, the DI squelch off level is reduced to 2/3 of the DI squelch on level.

The DI squelch circuit will be turned back on if the idle period is detected by no DI squelch level transitions for more than 180ns.

An external 78 ohm termination resistor is needed across DI+ and DI- as shown in Figure 12. The DI+ and DI- inputs can be coupled from an external MAU into the ML2652 with either capacitors or a transformer. The ML2652 meets all AUI receiver specifications outlined in IEEE 802.3-1988 Section 7.

CI+ and CI- are inputs from the external MAU which contain the 10 MHz $\pm 15\%$ collision signal as defined in IEEE 802.3-1988 Section 7. The CI+ and CI- inputs contain the same squelch circuit used on the DI inputs described in previous paragraphs in this section.

An external 78 ohm termination resistor is needed across CI+ and CI- as shown in Figure 12. The CI+ and CI- inputs can be coupled from an external MAU into the ML2652 with either capacitors (shown in Figure 12) or a transformer. The ML2652 meets all AUI receiver specifications outlined in IEEE 802.3-1988 Section 7.

The ML2652 contains an AUI/ \overline{TP} select input pin which controls whether the AUI or twisted pair interface is to be used for data transmission and reception. When AUI/ \overline{TP} Switching = High, the AUI is used for data transmission and reception. When AUI/ \overline{TP} Switching = Low, the twisted pair interface is used for data transmission and reception.

The AUISEL pin is a digital status output that indicates which interface has been selected for data transfer, either twisted pair or AUI. The pin consists of an open drain output with an internal pull-up resistor and can drive an LED from VCC or another digital input. AUISEL = High indicates that the twisted pair interface has been selected. AUISEL = Low indicates that the AUI interface has been selected.

The ML2652 has the capability to automatically select between the twisted pair interface and AUI. This automatic interface selection is accomplished by tying the LTP output pin to the AUI/ \overline{TP} input pin. When these two pins are connected together, if valid link pulses are detected, it is assumed that the twisted pair interface is being used. This causes \overline{LTP} output to go low, thus forcing AUI/ \overline{TP} low, and thus enabling the twisted pair interface. If no valid link pulses are detected, it is assumed that the twisted pair interface is not being used, thus causing \overline{LTP} to go high, thus forcing AUI/ \overline{TP} high, thus enabling the AUI interface. If valid link pulses reappear, the device will automatically disable the AUI and enable the twisted pair interface. The algorithm for determining valid link pulses is described in the Link Pulse section.

LOOPBACK

There are two pins on the ML2652 which provide loopback diagnostic features, LBKDIS and LPBK.

LPBK provides a loopback through the manchester encoder/decoder, but not through the on-chip 10BASE-T MAU. No data will go out on either the AUI port or the twisted pair port in this mode. This same function is found on many discrete manchester encoder/decoders.

ML2652

IEEE 802.3 MAUs normally loop the transmit data (DO+) when transmitting with no collisions. When using an external transceiver through the ML2652's AUI port, the controller can first check the local loopback by setting LPBK. If it passes this test it can then check the AUI cable and external MAU by doing the normal MAU loopback.

LBKDIS disables the on-chip 10BASE-T MAU loopback. This allows the user to attach a special RJ-45 connector that loops back the transmit twisted pair to the receive twisted pair. This test checks the entire interface out through the connector. When LBKDIS is activated (High), collision detection is disabled so that the ML2652 can transmit and receive in full duplex without collisions.

CONTROLLER INTERFACE

The ML2652 has a flexible and programmable digital interface which enables it to directly interface to Ethernet controllers manufactured by Intel, AMD, National, Seeg, and Fujitsu.

The controller interface consists of seven pins. TxC, TxD, and TxE are the transmit clock output, transmit data input, and transmit data enable input, respectively. RxC, RxD, and RxE are the receive clock output, receive data output, and receive data enable output, respectively. COL is the collision detect output.

All the standard Ethernet controllers use a similar controller interface but differ in the polarity of COL, LPBK, TxE and RxE, differ in what edge of TxC and RxC that clocks in the data, differ on whether the RxC clock needs to be continuous or not during idle, and differ on whether polarity of RxD during idle. In order to accommodate the different controller interface definitions, the controller select pins, CS2-0, modify these signals according to the Table 1.

Table 1. Controller Select Pin Definitions

CS2-0	TxC	TxE	RxC	RxE	COL	LPBK	Idl RxC	Idl RxD	Controller	
000	r	h	r	h	h	h	m	l	NSC	DP8390
001	f	l	f	l	l	l	n	hi	Intel	82586/96
010	r	h	r	h	h	h	n	hi	AMD	AM7990
011	f	h	r	h	h	l	c	lo	Seeg	8003/5
100	f	h	f	h	l	h	c	hi	Fujitsu	MB86950
101	—	—	—	—	—	—	—	—	—	—
110	—	—	—	—	—	—	—	—	—	—
111	—	—	—	—	—	—	—	—	PDN mode	—

r = rising edge clocks data
f = falling edge clocks data

h = active high
l = active low

c = RxC required continuously
n = RxC only during RxD transmission
m = RxC only during RxD transmission + 5 extra RxC cycles

POWERDOWN

The device can be placed in the power down mode with the controller select pins CS2-0 as described in Table 1. When in powerdown mode, the current consumption is reduced to less than 100µA and all device functions are disabled.

CRYSTAL OSCILLATOR

The ML2652 requires an accurate 20 MHz reference for internal clock generation. This can be achieved by connecting an external crystal or an external clock between the CLK and GND pins.

If an external clock is used, it must have a frequency of 20 MHz ±0.01% and have high and low levels of 3.5 and 1.5 volts.

If a crystal is used, the crystal should be placed physically as close as possible to the CLK and GND pins, especially CLK. No other external capacitors or components are required. The crystal should have the following characteristics:

1. Parallel resonant type
2. Frequency: 20 MHz
3. Tolerance: ±0.005% @ 25°C
4. Less than 0.005% frequency drift across temperature.
5. Maximum equiv. series resistance:
15 ohms @ 1-200µW
30 ohms @ 0.01-1µW
6. Typical load capacitance: 20pF
7. Maximum case capacitance: 5pF

ORDERING INFORMATION

PART NUMBER	TEMPERATURE RANGE	PACKAGE
ML2652CQ	0°C TO +70°C	44-Pin Molded Leaded PCC (Q44)

Data Quantizer

GENERAL DESCRIPTION

The ML4621 Data Quantizers is a low noise, wideband, bipolar monolithic ICs designed specifically for signal recovery applications in fiberoptic receiver systems. It contains a two stage wideband limiting amplifier which is capable of accepting an input signal as low as 2mV with a 55dB dynamic range. This high level of sensitivity is achieved by using a DC restoration feedback loop which nulls any offset voltage produced in the limiting amplifier.

The output stage is a high speed comparator circuit with both TTL and ECL outputs. An enable pin is included for added control.

The Minimum Signal Discriminator circuit provides a Link Monitor function with a user selectable reference voltage. This circuit monitors the peaks of the input signal and provides a logic level output indicating when the input falls below an acceptable level. This output can be used to disable the Quantizer and/or drive an LED, providing a visible link status.

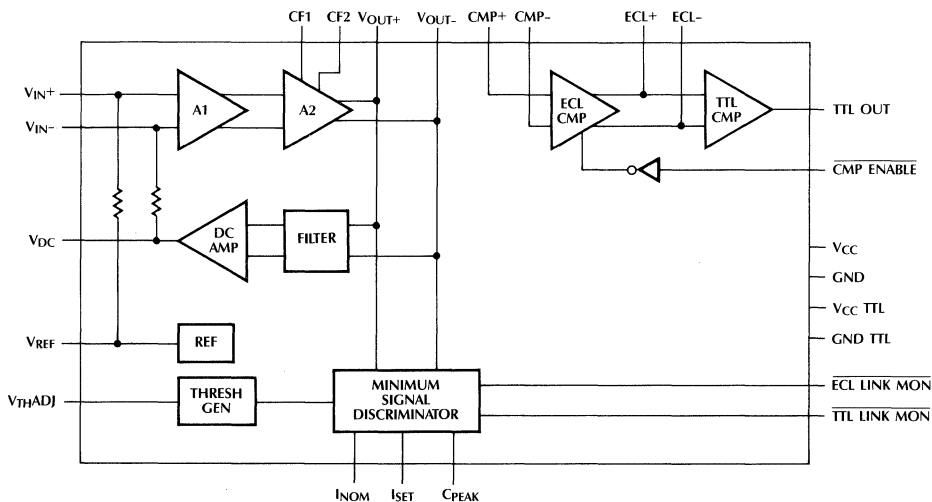
FEATURES

- 50MHz minimum bandwidth for data rates of up to 100MBd
- Can be powered by either +5V providing TTL level outputs or -5.2V providing ECL levels
- Low noise design:
 - 25 μ V RMS over 50MHz noise bandwidth
- Adjustable Link Monitor function
- Wide 55dB input dynamic range
- 10ns minimum input pulse
- Available in a 24-pin Skinny DIP and 28-pin PLCC

APPLICATIONS

- IEEE 802.3 FOIRL Receiver
- IEEE 802.5 4 and 16 Mbps Fiber Optic Token Ring
- IEEE 802.4 Fiber Optic Token Bus
- Fiber Optic Data Communications and Telecommunications Receivers

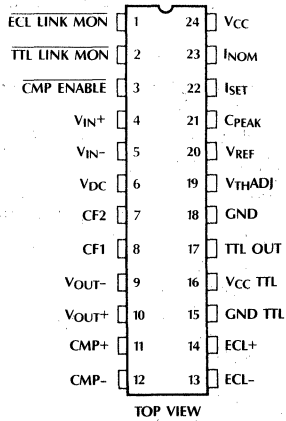
BLOCK DIAGRAM



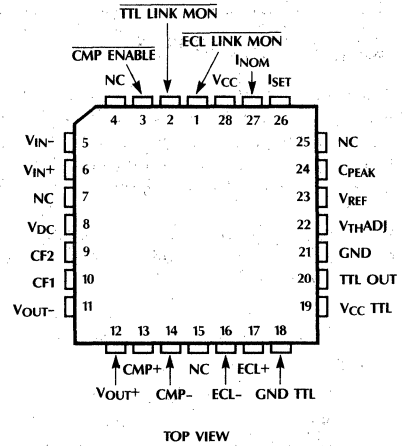
ML4621

PIN CONNECTIONS

ML4621
24-Pin Skinny DIP



ML4621
28-Pin PCC



PIN DESCRIPTION

NAME	FUNCTION	NAME	FUNCTION
$\overline{\text{ECL LINK MON}}$	ECL Link Monitor output. Signal is low when the $V_{\text{IN}+}$, $V_{\text{IN}-}$ inputs exceed the minimum threshold, which is set by a voltage on the V_{THADJ} pin. Signal is high when the input signal level is below the threshold.	V_{DC}	An external capacitor on this pin integrates an error signal which nulls the offset of the input amplifier. If the DC feedback loop is not being used, this pin should be connected to V_{REF} .
$\overline{\text{TTL LINK MON}}$	TTL Link Monitor output. Same logic function as $\overline{\text{ECL LINK MON}}$. Capable of driving a 10mA LED indicator. This pin normally tied to CMP ENABLE .	CF2	A capacitor from this pin to ground controls the maximum bandwidth of the amplifier to accommodate lower operating frequencies.
$\overline{\text{CMP ENABLE}}$	A low voltage at this TTL input pin enables both the ECL and the TTL outputs. A high TTL voltage disables the comparator output with ECL+ high, ECL- low, and TTL OUT high.	CF1	The capacitor on this pin should match the one on CF2.
$V_{\text{IN}-}$	This input pin should be capacitively coupled to the input source or to ground. (The input resistance is approximately 8k Ω .)	$V_{\text{OUT}-}$	The negative output of the amplifier, which is normally tied to CMP- .
$V_{\text{IN}+}$	This input pin should be capacitively coupled to the input source or to ground. (The input resistance is approximately 8k Ω .)	$V_{\text{OUT}+}$	The positive output of the amplifier, which is normally tied to CMP+ .
CMP-	This comparator input pin is an open base configuration which relies on the DC bias of the amplifier output to establish the proper DC operating voltage. This voltage should be reestablished if filtering is implemented between $V_{\text{OUT}-}$ and CMP- .	CMP+	This comparator input pin is an open base configuration which relies on the DC bias of the amplifier output to establish the proper DC operating voltage. This voltage should be reestablished if filtering is implemented between $V_{\text{OUT}+}$ and CMP+ .
ECL-	The ECL comparator negative output.	GND	Negative supply. Connect to -5.2V for ECL operation, or to ground for TTL operation.
ECL+	The ECL comparator positive output.	V_{THADJ}	This input pin sets the minimum amplitude of the input signal required to cause the link monitors to go low.
GND TTL	The negative supply for the TTL comparator stage. If the TTL output is not necessary, connect GND TTL and $V_{\text{CC TTL}}$ to V_{CC} .	V_{REF}	A 2.5V reference with respect to GND.
$V_{\text{CC TTL}}$	The positive supply for the TTL comparator stage. If the TTL output is not necessary, connect GND TTL and $V_{\text{CC TTL}}$ to V_{CC} .	C_{PEAK}	A capacitor from this pin to ground determines the Link Monitor response time.
TTL OUT	TTL data output. (Totem pole type output stage.)	I_{SET}	Current into an internal diode connected between this pin and GND is turned around and pulled from C_{PEAK} . This pin is normally connected to I_{NOM} .
		I_{NOM}	Sets a current of approx. 125 μA when connected to I_{SET} .
		V_{CC}	Positive supply. Connect to ground for ECL operation, or to 5V for TTL operation.

ML4621

ABSOLUTE MAXIMUM RATINGS

V_{CC} - GND	-0.3 to +70
V_{CC} TTL - GND TTL	-0.3 to +70
Inputs/Output GND	-0.3 to V_{CC} +0.3
Storage Temperature Range	-65°C to +150°C
Lead Temperature (Soldering 10 sec.)	+260°C

Absolute maximum ratings are those values beyond which the device could be permanently damaged. Absolute maximum ratings are stress ratings only and functional device operation is not implied.

ML4621 ELECTRICAL CHARACTERISTICS

Over recommended operating conditions of $T_A = 0^\circ\text{C}$ to 70°C , $V_{CC} = 5\text{V} \pm 5\%$, GND = 0V unless otherwise noted.

SYMBOL	PARAMETER	MIN	TYP	MAX	UNITS	CONDITIONS
I_{CC1}	V_{CC} Supply Current		65	100	mA	V_{CC} TTL = GND TTL = V_{CC}
I_{CC2}	V_{CC} Supply Current (TTL Out Enabled)		70	110	mA	V_{CC} TTL = V_{CC} GND TTL = GND
$I_{V_{REF}}$	V_{REF} Output Current	-5.0		0.5	mA	
V_{REF}	Reference Voltage	2.40	2.55	2.65	V	
A_V	Amplifier Gain A1 A2		75		V/V	$V_{IN} = 5\text{mV}$
V_{IN}	Input Signal Range	2		1400	mV _{P-P}	
V_{THADJ} Range	External Voltage at V_{THADJ} to set V_{TH}	1		2.5	V	
V_{OS}	Input Offset		3		mV	$V_{DC} = V_{REF}$ (DC loop inactive)
E_N	Input Referred Noise		25		μV	50MHz BW
BW	3dB Bandwidth	50	65		MHz	
V_{IN} PW	Min Input Pulsewidth		10		ns	
R_{IN}	Input Resistance		8		k Ω	V_{IN+} , V_{IN-}
t_{PD} AMP	Amplifier Propagation Delay	4		8	ns	From V_{IN+} , V_{IN-} to V_{OUT+} , V_{OUT-} $V_{IN} = 10\text{mV}_{P-P}$
t_{PD} ECL	ECL Comparator Propagation Delay	4		8	ns	From CMP+, CMP- to ECL+, ECL- $V_{IN} = 10\text{mV}_{P-P}$
t_{PD} TTL	TTL Comparator Propagation Delay	4		8	ns	From ECL+, ECL- to TTL OUT $V_{IN} = 10\text{mV}_{P-P}$
$R_{V_{THADJ}}$	Input Resistance of V_{THADJ}		6.8		k Ω	
$I_{V_{OUT}}$	Output Current of V_{OUT+} and V_{OUT-}			3	mA	
I_{CMP}	Leakage Current of CMP+ and CMP-		25		μA	
$V_{CM_{CMP}}$	Common Mode Range of CMP+ and CMP-	GND + 2.0		$V_{CC} - 1.0$	V	
ECL V_{OH}	Output High Voltage at ECL+, ECL-	3.90		4.30	V	With 200 Ω load tied to $V_{CC} - 2\text{V}$ $T_A = 25^\circ\text{C}$
ECL V_{OL}	Output Low Voltage at ECL+, ECL-	3.11		3.38	V	With 200 Ω load tied to $V_{CC} - 2\text{V}$ $T_A = 25^\circ\text{C}$
A_V ECL	ECL CMP Gain		100		V/V	
TTL V_{OH}		2.4			V	V_{CC} TTL = 5V, $I_{OH} = -50\mu\text{A}$
TTL V_{OL}				0.4	V	V_{CC} TTL = 5V, $I_{OL} = 2\text{mA}$
TTL V_{IH}		2.0			V	
TTL V_{IL}				0.8	V	
TTL I_{IH}		-50		50	μA	$V_{IH} = 2.4\text{V}$
TTL I_{IL}		-1.6		0	mA	$V_{IH} = 0.4\text{V}$
I_{NOM}			125		μA	$I_{NOM} = I_{SET}$

FUNCTIONAL DESCRIPTION

AMPLIFIER

The Quantizer has a two stage limiting amplifier with an input common mode range of (GND + 1.8V) to (V_{CC} - 1.5V). Maximum sensitivity is achieved through the use of a DC restoration feedback loop and AC coupling the input. When AC coupled, the input DC bias voltage is set by an on-chip network at about 1.9V. These coupling capacitors, in conjunction with the input impedance of the amplifier, establish a high pass filter with a 3dB corner frequency, f_L , at

$$f_L = \frac{1}{2\pi \cdot 8000 \cdot C} \quad (1)$$

Since the amplifier has a differential input, two capacitors of equal value are required. If the signal driving the input is single ended, one of the coupling capacitors can be tied to V_{CC} as shown in figure 1. The high corner frequency can also be adjusted by attaching capacitors to CF1 and CF2. The equation for adjusting this corner is

$$f_H = \frac{1}{2\pi \cdot 425 \cdot C} \quad (2)$$

Although the input is AC coupled, the offset voltage *within* the amplifier will be present at the amplifier's output. This is represented by V_{O5} in figure 2. In order to reduce this error a DC feedback loop is incorporated. This negative feedback loop nulls the offset voltage, forcing V_{O5} to be zero. An external capacitor at V_{DC} is used to store the offset voltage. Although the value of this capacitor is non-critical, the pole it creates can effect the stability of the feedback loop. To avoid stability problems using the ML4621, the

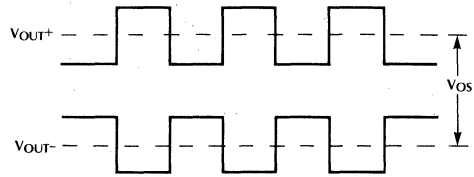


Figure 2.

value of this capacitor should be at least 100 times smaller than the input coupling capacitors.

On the ML4621, the output of the amplifier is isolated from the comparator and made available to the user. This allows the user to add circuitry between the amplifier and the comparator for wave shaping and other signal conditioning as desired.

COMPARATOR

Two types of comparators are employed in the output section of these Quantizers. The high speed ECL comparator is used to provide the ECL level outputs and in turn drives the TTL comparator. The enable pin, CMP ENABLE, is provided to control the ECL comparator. When CMP ENABLE is low the comparators function normally. When it's high, it forces ECL+ high, ECL- low, and TTL OUT high. The CMP ENABLE pin can be controlled with TTL level signals when the Quantizer is powered by 5V and ground.

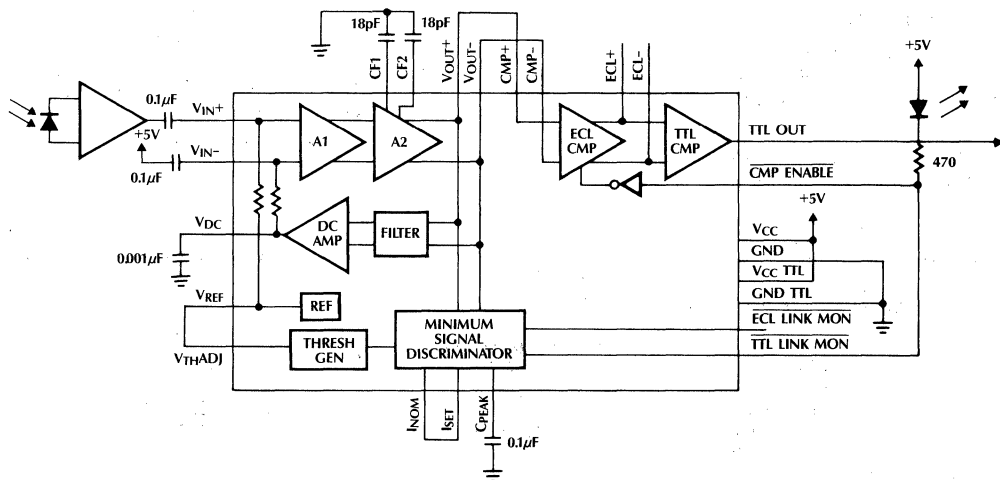


Figure 1. The ML4621 Configured for 20MHz Bandwidth with TTL Output

ML4621

LINK MONITOR

This function is implemented by the Minimum Signal Discriminator and the Threshold Generator circuits. The purpose of this function is to monitor the input signal and provide a status signal indicating when the input falls below a preset voltage level. This is done by peak detecting the output of the amplifier section and comparing this level with the voltage at V_{THADJ} .

The equation which determines the droop rate of the peak detector is

$$\frac{dV}{dt} = \frac{I_{SET}}{C} \quad (3)$$

In this equation C is the peak capacitor at C_{PEAK} . On the ML4621 the droop rate of the peak detector can be adjusted two ways:

- 1) By adjusting the value of the peak capacitor at C_{PEAK} .
- 2) By adjusting the charge current into the peak capacitor at I_{SET} .

The charge current, I_{SET} , can be controlled externally by connecting a resistor, R_{EXT} , between I_{SET} and V_{CC} . I_{SET} will then be

$$I_{SET} = \frac{V_{CC} - 0.7}{R_{EXT} + 1700} \quad (4)$$

For convenience, an on-chip current source of $125\mu A$ is available by connecting I_{NOM} to I_{SET} .

The Threshold Generator level shifts the reference voltage at V_{THADJ} through a circuit which has a temperature coefficient matching that of the limiting amplifier. The relationship between V_{THADJ} and V_{TH} (the minimum peak voltage at the input which will trigger the Link Monitor) is:

$$V_{THADJ} = 600V_{TH} + 0.7 \quad (5)$$

The on-chip reference voltage, V_{REF} , can be tied directly to V_{THADJ} to set the threshold level. This will set the minimum input signal on the ML4621 at about 3mV (peak).

A lower threshold level can be set by dividing down V_{REF} with a resistor string, as in figure 3.

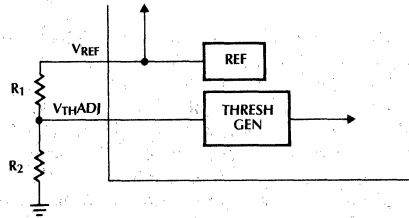


Figure 3.

Since the ML4621 has a relatively low input impedance of 6.8K and is offset by one diode drop, the equation which accounts for the load and offset is:

$$V_{THADJ} = \frac{R_2(6800V_{REF} + 0.7R_1)}{6800(R_1 + R_2) + R_1R_2} \quad (8)$$

THRESHOLD ADJUSTMENT EXAMPLE

If you are using the ML4621 and you want the Link Monitor to trigger when the received optical power goes below $1\mu W$ ($-30dBm$), you first need to calculate the resultant voltage at V_{IN+} and V_{IN-} . If you are using the Hewlett-Packard HFBR-24X6 Fiberoptic Receiver with a responsivity of $8mV/\mu W$, the peak-to-peak voltage would be:

$$1\mu W \times 8mV/\mu W = 8mV_{P-P} \quad (9)$$

So the Link Monitor should trigger at some point slightly lower than 4mV peak, say 3mV. Setting V_{TH} in equation 5 to 3mV and solving for V_{THADJ} yields:

$$V_{THADJ} = 600(.003) + 0.7 = 2.5V$$

This is a convenient value since the reference voltage supplied by the Quantizer, V_{REF} , is 2.5V.

The Link Monitor has about 0.4mV (peak) hysteresis built-in. More hysteresis can be induced by connecting a resistor between TTL LINK MON and V_{THADJ} creating a positive feedback loop.

Refer to Micro Linear's Application Note 6 for more detail.

ORDERING INFORMATION

PART NUMBER	TEMPERATURE RANGE	PACKAGE
ML4621CP	0°C to +70°C	Molded DIP (P24N)
ML4621CQ	0°C to +70°C	MOLDED PCC (Q28)

ML4622, ML4624

Fiber Optic Data Quantizer

GENERAL DESCRIPTION

The ML4622 and ML4624 Data Quantizers are low noise, wideband, bipolar monolithic ICs designed specifically for signal recovery applications in fiberoptic receiver systems. They contain a wideband limiting amplifier which is capable of accepting an input signal as low as $2mV_{p-p}$ with a 55dB dynamic range. This high level of sensitivity is achieved by using a DC restoration feedback loop which nulls any offset voltage produced in the limiting amplifier.

The output stage is a high speed comparator circuit with both TTL and ECL outputs. An enable pin is included for added control.

The Link Detect circuit provides a Link Monitor function with a user selectable reference voltage. This circuit monitors the peaks of the input signal and provides a logic level output indicating when the input falls below an acceptable level. This output can be used to disable the Quantizer and/or drive an LED, providing a visible link status.

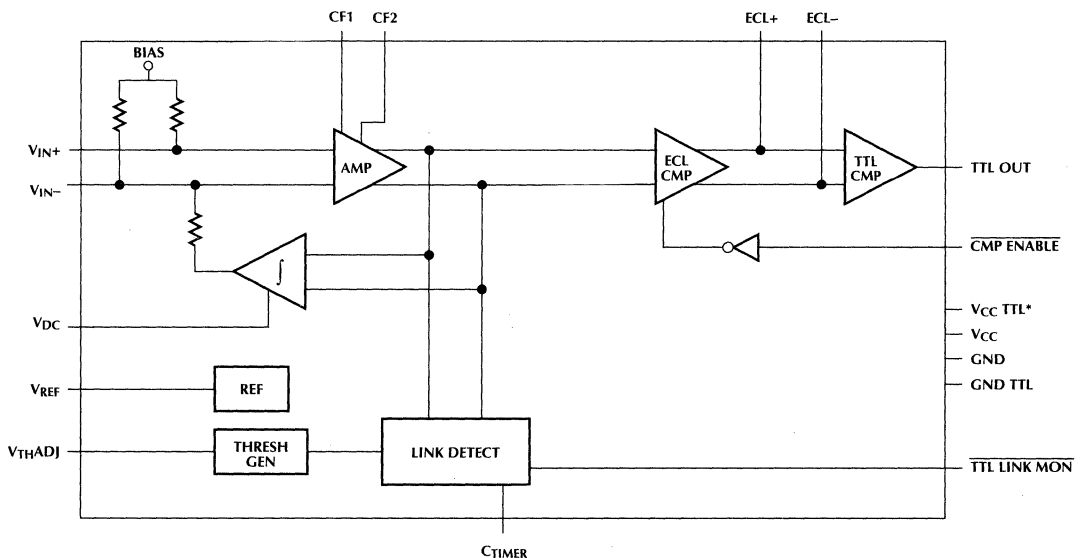
FEATURES

- Data rates up to 40MHz or 80MBd
- Can be powered by either +5V providing TTL or raised ECL level outputs or -5.2V providing ECL levels
- Low noise design: 25 μ V RMS over bandwidth
- Adjustable Link Monitor function with hysteresis
- Wide 55dB input dynamic range
- Low power design
- Available in 16-pin SOIC (Narrow) or DIP (ML4622), 24-pin Skinny DIP (ML4624) and 28-pin PCC (ML4624)
- ML4624 is pin compatible with the ML4621

APPLICATIONS

- IEEE 802.3 FOIRL, 10BASE-F Receiver
- IEEE 802.5 4 and 16 Mbps Fiber Optic Token Ring
- Fiber Optic Data Communications and Telecommunications Receivers

ML4622/ML4624 BLOCK DIAGRAM



*ML4624 ONLY

PIN DESCRIPTION

NAME	FUNCTION	NAME	FUNCTION
$\overline{\text{TTL LINK MON}}$	TTL Link Monitor output. Signal is low when the $V_{\text{IN}+}$, $V_{\text{IN}-}$ inputs exceed the minimum threshold, which is set by a voltage on the V_{THADJ} pin. Signal is high when the input signal level is below the threshold. Capable of driving a 10mA LED indicator. This pin can be tied to CMP ENABLE .	$V_{\text{CC TTL}}$	The positive supply for the TTL comparator stage. If the TTL output is not necessary, connect $V_{\text{CC TTL}}$ to V_{CC} . (ML4624 only)
CMP ENABLE	A low voltage at this TTL input pin enables both the ECL and the TTL outputs. A high TTL voltage disables the comparator output with ECL+ high, ECL- low, and TTL OUT high.	TTL OUT	TTL data output.
$V_{\text{IN}-}$	This input pin should be capacitively coupled to the input source or to filtered ground (note 5). (The input resistance is approximately 1.6K Ω .)	V_{DC}	An external capacitor on this pin integrates an error signal which nulls the offset of the input amplifier. If the DC feedback loop is not being used, this pin should be connected to V_{REF} .
$V_{\text{IN}+}$	This input pin should be capacitively coupled to the input source or to filtered ground (note 5). (The input resistance is approximately 1.6K Ω .)	CF2	A capacitor from this pin to CF1 controls the maximum bandwidth of the amplifier.
ECL-	The ECL comparator negative output. Has internal pull down resistor. External pull downs are not required unless driving a large capacitive load.	CF1	Connect to CF2 through a capacitor.
ECL+	The ECL comparator positive output. Has internal pull down resistor. External pull downs are not required unless driving a large capacitive load.	GND	Negative supply. Connect to -5.2V for ECL operation, or to ground for TTL or raised ECL operation.
GND TTL	The negative supply for the TTL comparator stage. If the TTL output is not necessary, connect GND TTL to V_{CC} .	V_{THADJ}	This input pin sets the link monitor threshold.
		V_{REF}	A 2.5V reference with respect to GND.
		C_{Timer}	A capacitor from this pin to V_{CC} determines the Link Monitor response time.
		V_{CC}	Positive supply. Connect to ground for negative ECL operation, or to 5V for TTL or raised ECL operation.

ML4622, ML4624

ABSOLUTE MAXIMUM RATINGS

(Note 1)

$V_{CC} - GND$	-0.3 to +7.0
$V_{CC} \text{ TTL} - GND \text{ TTL}$	-0.3 to +7.0
Inputs/Outputs GND	-0.3 to $V_{CC} + 0.3$
Storage Temperature Range	-65°C to +150°C
Lead Temperature (Soldering 10 sec.)	+260°C

ML4622, ML4624 ELECTRICAL CHARACTERISTICS (Note 2 and 3)

Over recommended operating conditions of $T_A = 0^\circ\text{C}$ to 70°C for commercial temperature range, $T_A = -40^\circ\text{C}$ to $+85^\circ\text{C}$ for industrial temperature range, $V_{CC} = 5\text{V} \pm 10\%$, $GND = 0\text{V}$ unless otherwise noted.

SYMBOL	PARAMETER	MIN	TYP	MAX	UNITS	CONDITIONS
I_{CC1}	V_{CC} Supply Current (TTL Output Disabled)		35	45	mA	$GND \text{ TTL} = V_{CC}$
I_{CC2}	V_{CC} Supply Current (TTL Output Enabled)		55	70	mA	$GND \text{ TTL} = GND$
V_{REF}	Reference Voltage	2.40	2.50	2.60	V	
I_{VREF}	V_{REF} Output Source Current			5	mA	
A_V	Amplifier Gain		100		V/V	
V_{IN}	Input Signal Range	2		1600	mV _{p-p}	
V_{THADJ} Range	External Voltage at V_{THADJ} to set V_{TH}	0.5		2.7	V	
V_{OS}	Input Offset		3		mV	$V_{DC} = V_{REF}$ (DC loop inactive)
E_N	Input Referred Noise		25		μV	50MHz BW
BW	3dB Bandwidth		45		MHz	
R_{IN}	Input Resistance	1	1.6	2.5	k Ω	V_{IN+}, V_{IN-}
I_{VTHADJ}	Input Bias Current of V_{THADJ}	-200	10	+200	μA	
t_{pDRTL}	Propagation Delay		15		ns	From V_{IN+}, V_{IN-} to TTL Out $V_{IN} = 10\text{mV}_{p-p}$
t_{pDECL}	Propagation Delay		11		ns	From V_{IN+}, V_{IN-} to ECL+, ECL- $V_{IN} = 10\text{mV}_{p-p}$
TTL V_{OH}		2.4			V	$V_{CC} \text{ TTL} = 5\text{V}, I_{OH} = -50\mu\text{A}$
TTL V_{OL}				0.55	V	$V_{CC} \text{ TTL} = 5\text{V}, I_{OL} = 2\text{mA}$
TTL V_{IH}		2.0			V	
TTL V_{IL}				0.8	V	
TTL I_{IH}		-50		50	μA	$V_{IH} = 2.4\text{V}$
TTL I_{IL}		-1.6		0	mA	$V_{IH} = 0.4\text{V}$

ML4622, ML4624 ELECTRICAL CHARACTERISTICS (Continued)

Over recommended operating conditions of $T_A = 0^\circ\text{C}$ to 70°C for commercial temperature range, $T_A = -40^\circ\text{C}$ to $+85^\circ\text{C}$ for industrial temperature range, $V_{CC} = 5\text{V} \pm 10\%$, $\text{GND} = 0\text{V}$ unless otherwise noted.

SYMBOL	PARAMETER	MIN	TYP	MAX	UNITS	CONDITIONS
V_{TH}	Input Threshold Voltage ML4622 ML4624	4	5	6	mV _{P-P}	$V_{THADJ} = V_{REF}$ (note 4) $V_{THADJ} = V_{REF}$ (note 4)
		5	6	7	mV _{P-P}	
Hysteresis			20		%	
V_{CM}	Common mode voltage on V_{IN+} , V_{IN-}		1.65		V	
ECL_{VOH}	Output High Voltage at $ECL+$, $ECL-$	$V_{CC} - 1.06$		$V_{CC} - 0.7$ $V_{CC} - 0.6$	(note 5)	With 200 Ω load tied to $V_{CC} - 2\text{V}$
ECL_{VOL}	Output Low Voltage at $ECL+$, $ECL-$	$V_{CC} - 1.89$		$V_{CC} - 1.62$ $V_{CC} - 1.56$	(note 5)	With 200 Ω load tied to $V_{CC} - 2\text{V}$

Note 1: Absolute maximum ratings are limits beyond which the life of the integrated circuit may be impaired. All voltages unless otherwise specified are measured with respect to ground.

Note 2: Limits are guaranteed by 100% testing, sampling, or correlation with worst-case test conditions.

Note 3: Low Duty Cycle pulse testing is performed at T_A .

Note 4: DC Tested — Threshold for switching TTL LINK MON from High (off) to Low (on).

Note 5: Industrial temperature range specification.

ML4622, ML4624

FUNCTIONAL DESCRIPTION

AMPLIFIER

The ML4622, ML4624 have an adjustable Bandwidth limiting amplifier. Maximum sensitivity is achieved through the use of a DC restoration feedback loop and AC coupling the input. When AC coupled, the input DC bias voltage is set by an on-chip network at about 1.7V. These coupling capacitors, in conjunction with the input impedance of the amplifier, establish a high pass filter with a 3dB corner frequency, f_L , at

$$f_L = \frac{1}{2\pi 1600 C} \quad (1)$$

Since the amplifier has a differential input, two capacitors of equal value are required. If the signal driving the input is single ended, one of the coupling capacitors can be tied to V_{CC} as shown in figure 1.

CF1 and CF2 create a low pass filter with the corner frequency determined by the following equation

$$f_H = \frac{1}{2\pi 800 (C + 4pF)} \quad (2)$$

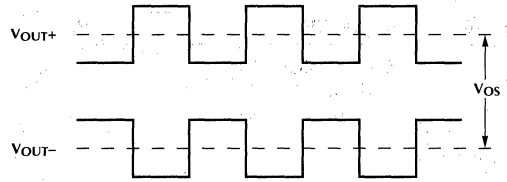
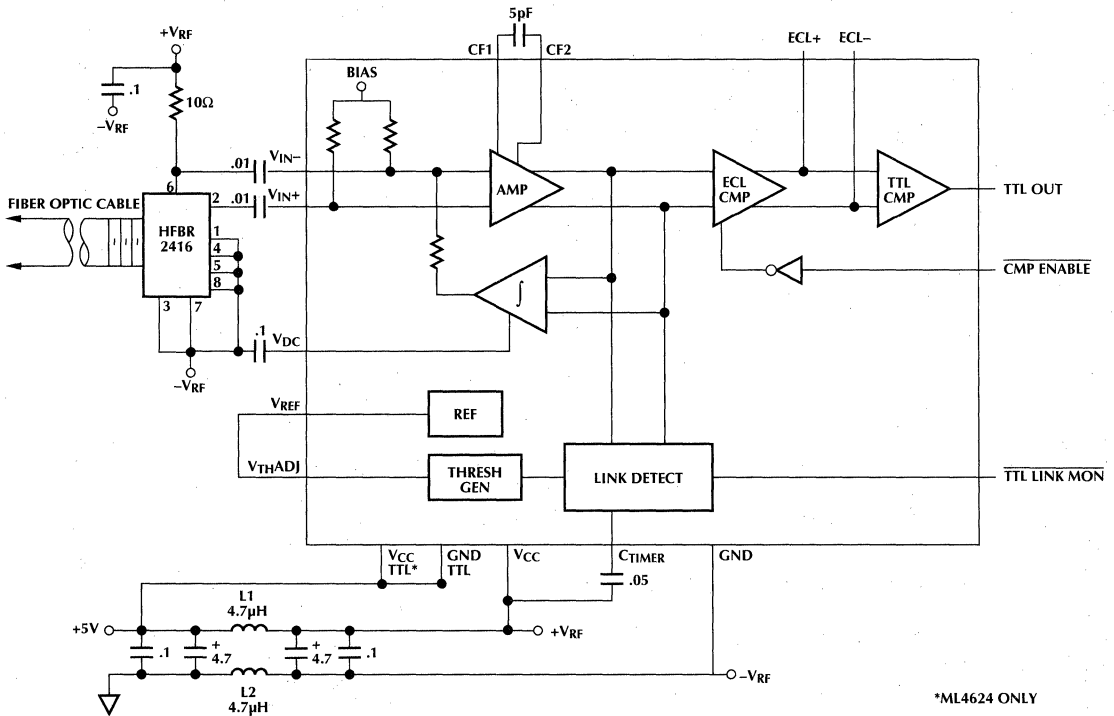


Figure 2.

The above equation applies when a single capacitor is tied between CF1 and CF2. When using two capacitors of equal value (Cap1 from CF1 to V_{CC} , Cap2 from CF2 to V_{CC}) the value derived for C should be doubled.

Although the input is AC coupled, the offset voltage *within* the amplifier will be present at the amplifier's output. This is represented by V_{OS} in figure 2. In order to reduce this error a DC feedback loop is incorporated. This negative feedback loop nulls the offset voltage, forcing V_{OS} to be zero. Although the capacitor on V_{DC} is non-



Note: If TTL OUT is used, tie GND TTL to unfiltered ground and remove L1. If TTL OUT and ECL outputs are both used, add 3K pull-down resistors at ECL outputs.

Figure 1. The ML4622, ML4624 Configured for 20MHz Bandwidth

critical, the pole it creates can effect the stability of the feedback loop. To avoid stability problems, the value of this capacitor should be at least 10 times larger than the input coupling capacitors.

COMPARATOR

Two types of comparators are employed in the output section of these Quantizers. The high speed ECL comparator is used to provide the ECL level outputs and in turn drives the TTL comparator. The enable pin, CMP ENABLE, is provided to control the ECL comparator. When CMP ENABLE is low the comparators function normally. When it's high, it forces ECL+ high, ECL- low, and TTL OUT high. The CMP ENABLE pin can be controlled with TTL level signals when the Quantizer is powered by 5V and ground.

LINK DETECT CIRCUIT

The Link Detect circuit monitors the input signal and provides a status signal indicating when the input falls below a preset voltage level. When the input falls below the preset voltage level, the TTL LINK MON output changes from active (low) to inactive (high). This signal can be fed to the ML4662 10BASE-FL transceiver or a similar type of function to indicate a Low Light Condition. This output can also be used to disable the output data by tying it to the CMP ENABLE input.

In many fiber optic systems, including Ethernet and Token Ring, a bit error rate is given at a minimum power level. For example, in a 10Base-FL receiver there must be less than 1×10^{-9} bit errors at a receive power level of -32.5dBm average. Designers of these systems must insure that the bit error rate is lower than the specification at the given minimum power level. One procedure to determine the sensitivity of a receiver is to start at the lowest optical power level and gradually increase the optical power until the BER is met. In this case the Link Detect circuit must not disable the receiver (i.e. CMP ENABLE should be tied to Ground). Once the sensitivity of the receiver is determined, the Link Detector circuit can be set just above the power level that meets the BER specification. This way the receiver will shut off before the BER is exceeded.

The ML4622 and ML4624 quantizers have greater Link Detect sensitivity, noise immunity, and accuracy than their predecessor the ML4621.

The threshold generator shifts the reference voltage at V_{THADJ} through a circuit which has a temperature coefficient matching that of the limiting amplifier. The relationship between the V_{THADJ} and the V_{TH} (the peak to peak input threshold) is:

$$V_{THADJ} = 417 V_{TH} \text{ (ML4624)} \quad (3)$$

$$V_{THADJ} = 500 V_{TH} \text{ (ML4622)}$$

In most cases, including 10Base-FL, 10Base-FB and Token-Ring, V_{THADJ} can be tied directly to V_{REF} . However if greater sensitivity is required the circuit in figure 3 can be used to adjust the V_{THADJ} voltage. Even if V_{REF} is tied to V_{THADJ} , it is a good idea to layout a board with these two resistors available. This will allow potential future adjustments without board revisions.

The response time of the Link Detect circuit is set by the C_{Timer} pin. Starting from the link off state (i.e., TTL LINK MON is high), the link can be switched on if the input exceeds the set threshold for a time given by:

$$T = \frac{C_{TIMER} \times 0.7V}{700\mu A} \quad (4)$$

To switch the link from on to off, the above time will be doubled.

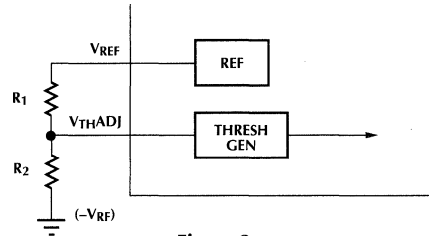


Figure 3.

ML4622, ML4624

BURST MODE

In some fiber optic links, the idle signal is DC, or of a frequency that is substantially different from the data. For these links, a faster response time of the DC loop and the Link Monitor is required.

The ML4622 and ML4624 has been designed to accommodate these two requirements. The input coupling capacitors can be relatively small and still maintain stability. With smaller input coupling capacitors and V_{DC} capacitor a faster DC loop response time can be achieved. The Link Monitor is also enhanced to have a faster response time.

ORDERING INFORMATION

PART NUMBER	TEMPERATURE RANGE	PACKAGE
ML4622CP	0°C to +70°C	Molded DIP (P16)
ML4622CS	0°C to +70°C	Molded SOIC (Narrow) (S16N)
ML4622IS	-40°C to +85°C	Molded SOIC (Narrow) (S16N)
ML4624CP	0°C to +70°C	Molded DIP (P24N)
ML4624CQ	0°C to +70°C	Molded PCC (Q28)

PIN DESCRIPTION

NAME	DESCRIPTION	NAME	DESCRIPTION
ECLN	Negative ECL data input. Tie to VBIAS for single ended ECL operation or when ECLP is used as an enable. Tie to ground during TTL only operation.	VCC	Positive power supply. +5 volts.
VBIAS	BIAS voltage for single ended ECL operation.	IOFF	Connect a resistor from this pin to VCC to increase the off current to the LED, i.e. 4.3kΩ for 1mA. With this pin open, the default IOFF current is between 0.5–1.0mA
LED	Fiber optic LED drive pin. Connect the LED between this pin and VCC.	VREF	A constant 1.2V reference output used to set up DRV.
GND	Negative power supply. This pin should be tied to the grounded side of RTSET to improve output accuracy and avoid a ground loop.	DRV	A DC input that sets the positive swing on RTSET and the high level output current to the LED.
RTSET	Output current programming pin. Connect a resistor of value V_{DRV}/I_{LED} from this pin to ground to set the high LED output current.	PTAT	Proportional to Absolute Temperature. A 1.0V reference at 25°C that moves proportional to absolute temperature, also used to set up DRV. (See figure 1)
RPK	Peaking circuit bias pin. Connect a resistor of value V_{DRV}/I_{PEAK} from this pin to ground when using the peaking circuit. Leave open circuited when peaking is not used.	TTL	TTL data input. Can also be used as an enable during ECL operation. TTL = High (enabled), TTL = Low (disabled).
PEAK	Peaking circuit output pin. When using peaking, connect this pin to VCC through a resistor of value RRPK. Then connect a capacitor from this pin to the LED cathode. When peaking is not used, open circuit RPK.	ECLP	Positive ECL data input controls signal to the LED. Tie to VBIAS during TTL only operation or use as an enable.

ABSOLUTE MAXIMUM RATINGS

VCC -0.3V to 6V
 Input Pin Voltages -0.3V to VCC +0.3V
 LED Output Current 120mA

PEAK DC Output Current 120mA
 Storage Temperature -65°C to +150°C
 Lead Temperature (Soldering 10 sec) 260°C

ELECTRICAL CHARACTERISTICS

Over the recommended operating conditions of $T_A = 0^\circ\text{C}$ to 70°C , $V_{CC} = 5V \pm 5\%$, unless otherwise specified.

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
I_{CC}	Supply Current	LED off		25	35	mA
VREF	VREF Voltage	No Load	1.14	1.20	1.26	V
V_{PTAT}	PTAT Voltage	No Load, $T_A = 25^\circ\text{C}$ $T_A = 85^\circ\text{C}$	0.9 1.08	1.0 1.2	1.1 1.32	V V
V_{OS}	Driver Offset	VDRV = 1.2V, RTSET = 20 Ω			50	mV
I_{LEDH} I_{LEDL}	LED Current Accuracy High Low	VDRV = VREF, RTSET = 20 Ω $I_{OFF} = \text{open}$	54 0.5	60 0.7	66 1.0	mA mA
t_R	Rise Time	VDRV = VREF, RTSET = 20 Ω		4.5		ns
t_F	Fall Time	VDRV = VREF, RTSET = 20 Ω		4.5		ns
t_{PLH} t_{PHL}	Propagation Delay Low to High High to Low	VDRV = VREF, RTSET = 20 Ω TTL and ECL		10.0 10.0		ns ns
t_{PWD}	Pulse Width Distortion	VDRV = VREF, RTSET = 20 Ω		1.0	2.0	ns
V_{PK}	Peaking Voltage	$R_{RPK} = 20\Omega$, $C_{PK} = 100\text{pF}$, $R_{PEAK} = 20\Omega$	1.08	1.2	1.32	V
V_{PKTR}	Peaking Rise Time	$R_{RPK} = 20\Omega$, $C_{PK} = 100\text{pF}$, $R_{PEAK} = 20\Omega$		4.5		ns
V_{PKTF}	Peaking Fall Time	$R_{RPK} = 20\Omega$, $C_{PK} = 100\text{pF}$, $R_{PEAK} = 20\Omega$		4.5		ns
I_{ECL}	ECL Input Current				20	μA
I_{TTL}	TTL Input Current				100	μA
V_{DO}	Dropout Voltage between pin 5 and 3		1.5			V
I_{OFF}	Additional LED Off Current	$V_{CC} = 5V$, $R_{OFF} = 4.3\text{k}\Omega$	0.8	1.0	1.2	mA
VBIAS	ECL BIAS Voltage	$V_{CC} = 5V$, $T_A = 25^\circ\text{C}$		3.8		V

Note 1: Absolute maximum ratings are limits beyond which the life of the integrated circuit may be impaired. All voltages unless otherwise specified are measured with respect to ground.

Note 2: Limits are guaranteed by 100% testing, sampling, or correlation with worst-case test conditions.

Note 3: Low Duty cycle pulse testing is performed at T_A .

FUNCTIONAL DESCRIPTION

The ML4632 accepts ECL and TTL input signals and generates a high speed, high accuracy output current which is independent of supply voltage variations. The output current is programmable from 20mA to 100mA. A temperature coefficient can be programmed into the output current and a peaking circuit can be added with a few external components.

The input of the LED driver accepts both ECL and TTL signals. The ECL input stage is a standard NPN differential pair with a common mode range of between 3V and 4.5V with a +5V supply. A bias voltage VBIAS is available for biasing either ECL input for single-ended operation. The TTL input has a standard switching range of between 0.8V and 2.0V. These inputs are ANDed so that the extra input can be used as an enable.

Output current to the LED is set by connecting the appropriate resistance from RTSET with ground. With the VREF and DRV pins tied together, the high level output voltage at RTSET will be 1.2V. The current in the external resistor will be equal to the current through the LED. The output current with RTSET set to 20Ω will be

$$I_{LED} (HIGH) = 1.2V/R_{TSET} = 1.2V/20\Omega = 60mA.$$

The low level output current is set internally by a resistor at approximately 0.7mA. This current prebiases the LED and results in faster optical rise times. The value of this current can be increased by connecting a resistor from the IOFF pin to VCC. The additional current will be equal to $(VCC - 0.7V)/R_{IOFF}$.

The voltage input at the DRV pin appears across the RTSET pin when the LED is turned on. The current in RTSET is directed through the LED. Therefore the voltage set at DRV along with the RTSET resistor sets the current through the LED.

A temperature coefficient of between 0ppm/°C and 3300ppm/°C can be programmed into the high level output current to compensate for the drop in LED optical output power at high temperatures. This is accomplished by driving the DRV pin from a resistor divider between the VREF and PTAT pins.

When DRV is tied directly to PTAT, the peak voltage at RTSET will be 1.0V at 25°C and have a 3300ppm/°C temperature coefficient. At 85°C, PTAT is 1.2V and equal to VREF. An arbitrary temperature coefficient less than 3300 ppm/°C can be set by using a resistor divider between PTAT and VREF to set the voltage at DRV, as shown in figure 1.

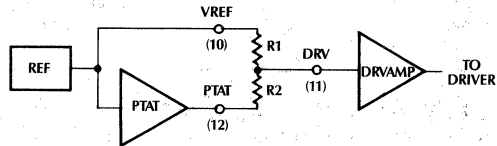


Figure 1. Current for Programming Output Temperature Coefficient

In this configuration the temperature coefficient is

$$TC_{ILED} = (3300\text{ppm}/^\circ\text{C}) \frac{R1}{R1 + R2}, \text{ and}$$

$$I_{LED} (HIGH) = \frac{1V + 0.2V \left(\frac{R2}{R1 + R2} \right)}{RTSET}$$

The output current will be a linear function of temperature. A plot of I_{LED} versus temperature for several values of the programming resistance, R1 and R2, in figure 2.

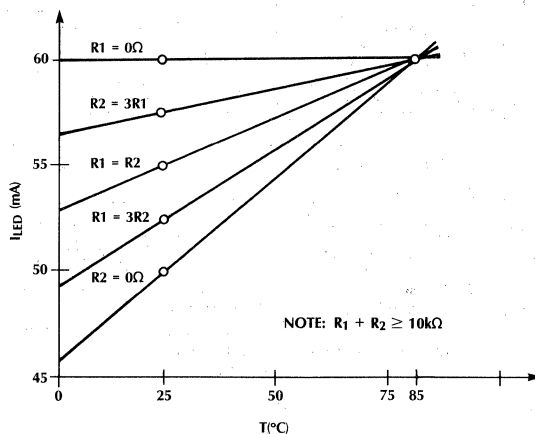


Figure 2. I_{LED} vs T , $R_{TSET} = 20\Omega$

The ML4632 output stage conducts full load current only when the LED is on, and even then power dissipation in the part is low because most of the +5V

supply voltage is dropped across the LED and external resistor R_{TSET} . Even with a low power design, the LED driver junction temperature will rise above ambient due to quiescent power dissipation and won't exactly match the LED junction temperature since it is also self-heating. Therefore, the effectiveness of a temperature compensated design will be related to component power dissipations, thermal conductance of the PC board and packaging, and the proximity of the LED driver to the LED.

The ML4632 also provides for peaking of the LED output current. Peaking is used to counteract the effects of the LED junction capacitance. By creating a controlled overshoot and undershoot in the output current waveform, charge is transferred to and from the LED capacitance on the rising and falling edges of the output, speeding up rise and fall times.

To provide peaking current, a second output stage is biased up with a resistor from RPK to ground and another from PEAK to VCC. When these bias resistors are set equal to each other, a pulse will be generated across the R_{PEAK} resistor with a magnitude equal to the voltage on the DVR pin. A coupling capacitor transfers the peaking current from the PEAK pin to the LED on the rising and falling edges of the output current waveform.

A typical application is shown in figure 3. When the resistors R_{RPK} and R_{PEAK} are both set to 20Ω , a pulse will be generated at the PEAK pin of magnitude 1.2V and equivalent resistance 20Ω (assuming $V_{DRV} = 1.2V$).

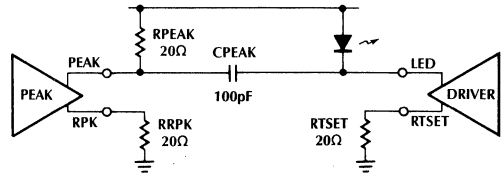
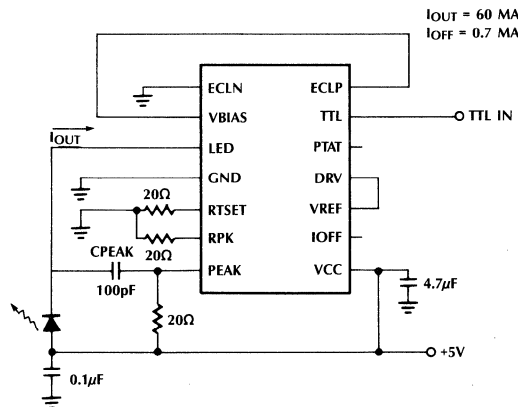


Figure 3. Application of the Peaking Circuit

The peaking current is coupled through the 100 pF capacitor, C_{PEAK} , which will transfer 120 pC of charge to and from the LED on each cycle of output current. The peaking circuit shown provides approximately a 70% overshoot current into a 0Ω LED impedance. Peaking currents will be slightly lower for real LED's.



NOTE: THE LED, PEAK & VCC TRACES SHOULD BE VERY SHORT AND SHIELDED WITH A GND PLANE TO REDUCE RINGING AND OVERSHOOT AT THE LED.

TTL DRIVEN IMPLEMENTATION
(No Temp. Comp)

ML4632

ORDERING INFORMATION

PART NUMBER	TEMPERATURE RANGE	PACKAGE
ML4632CP	0°C to +70°C	14-Pin Molded DIP (P14)
ML4632CS	0°C to +70°C	16-Pin SOIC (S16W)

AUI Multiplexer

GENERAL DESCRIPTION

The ML4642 AUI Multiplexer contains all the necessary drivers/receivers and control logic to implement a 2 port MAU when used in conjunction with a transceiver chip which has a standard 802.3 AUI interface. In addition, the ML4642 is capable of operating in stand-alone mode where it interconnects two DTEs in the absence of a network MAU. Several ML4642s can be cascaded together to implement a 4 or 8 port MAU or stand-alone device.

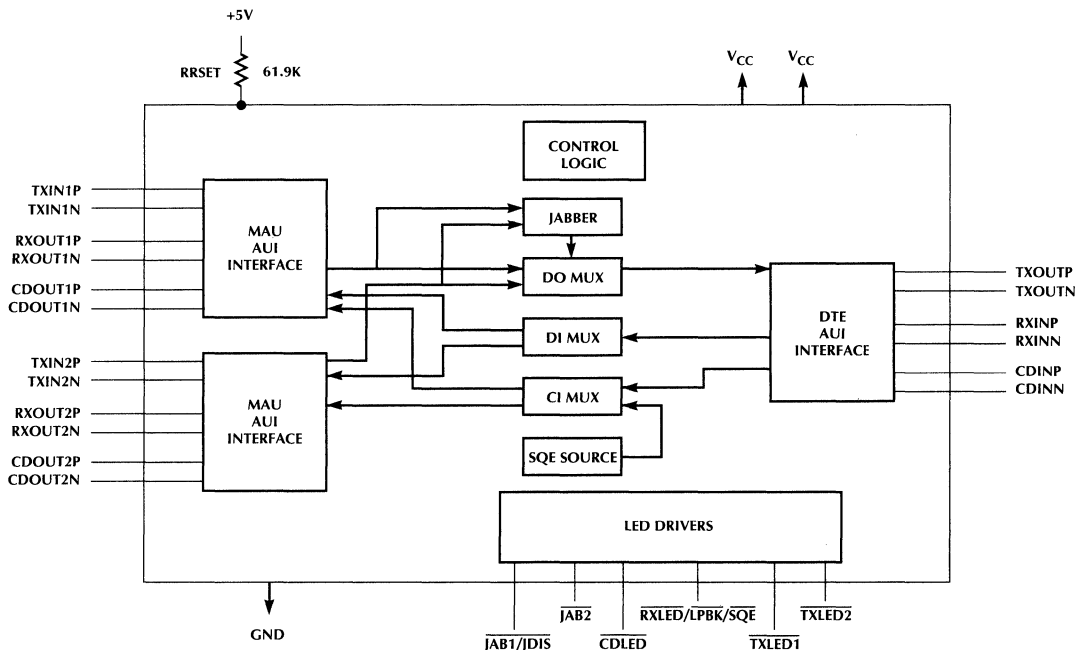
Logic within the ML4642 detects collisions resulting from multiple DTEs transmitting simultaneously. In addition, collision signals received from a transceiver attached at the MAU port are propagated to both of the DTE ports. Jabbering DTEs are prevented from loading down the network by internal jabber timers which disable babbling ports.

Squelch circuitry on port receivers prevent noise on the cables from being erroneously interpreted as valid data. Transmit, receive, collision, and jabber LED drivers indicate network activity and faults. The ML4642 is available in a 28 pin SSOP package.

FEATURES

- IEEE 802.3 compliant AUI interfaces assure compatibility with any AUI ready devices.
- No crystal or clock input.
- On-chip Jabber logic, Collision Detection, and SQE test with enable/disable option.
- Selectable Loopback, Jabber, and SQE Test allows cascading of multiple chips to increase DTE port fan-out.
- Six network status LED outputs.
- 28 pin SSOP packaging
- Semi-standard options available

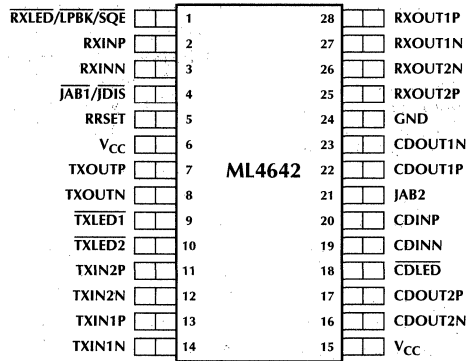
BLOCK DIAGRAM



ML4642

PIN CONNECTIONS

ML4642
28-PIN SSOP (R28)



TOP VIEW

PIN DESCRIPTION

PIN NO.	NAME	FUNCTION	DESCRIPTION
1	RXLED/LPBK/SQE	I/O	Active low receive LED driver for MAU port. If tied to ground, this pin enables internal loopback of the active TXIN pair to the RXOUT pairs and enables SQE test. If tied to 0.6 volts internal loopback is enable but SQE test is disabled.
2	RXINP	Input	Receive signal pair for MAU port.
3	RXINN	Onput	Receive signal pair for MAU port.
4	JAB1/DIS	I/O	Active low jabber LED driver for DTE port 1. If tied to ground, the jabber function is disabled at TXIN1 and TXIN2.
5	RRSET	Input	Bias setting external resistor, 61.9K Ω .
6	V _{CC}	Power	+5 volt power supply
7	TXOUTP	Output	Transmit signal pair for MAU port.
8	TXOUTN	Output	Transmit signal pair for MAU port.
9	TXLED1	Output	Active low transmit LED driver for DTE AUI port 1.
10	TXLED2	Output	Active low transmit LED driver for DTE AUI port 2.
11	TXIN2P	Input	Transmit signal pair for DTE port 2.
12	TXIN2N	Input	Transmit signal pair for DTE port 2.
13	TXIN1P	Input	Transmit signal pair for DTE port 1.
14	TXIN1N	Input	Transmit signal pair for DTE port 1.
15	V _{CC}	Power	+5 volt power supply
16	CDOUT2N	Output	Collision signal pair for DTE port 2.
17	CDOUT2P	Output	Collision signal pair for DTE port 2.
18	CDLED	Output	Active low collision LED driver.
19	CDINN	Input	Collision signal pair for MAU port.
20	CDINP	Input	Collision signal pair for MAU port.
21	JAB2	Output	Active low jabber LED driver for DTE port 2.
22	CDOUT1P	Output	Collision signal pair for DTE port 1.
23	CDOUT1N	Output	Collision signal pair for DTE port 1.
24	GND	Ground	GND.
25	RXOUT2P	Output	Receive signal pair for DTE port 2.
26	RXOUT2N	Output	Receive signal pair for DTE port 2.
27	RXOUT1N	Output	Receive signal pair for DTE port 1.
28	RXOUT1P	Output	Receive signal pair for DTE port 1.

ML4642

ABSOLUTE MAXIMUM RATINGS (Note 1)

Power Supply Voltage Range V_{CC} -0.3 to +6.0V
 Input Current RRSET, JAB1/JABD, JAB2, CDLED,
 RxLED/LPBK/SQE, TxLED1, TxLED2 60mA
 Storage Temperature -65°C to +150°C
 Lead Temperature (Soldering 10 seconds) 260°C

OPERATING CONDITIONS (Note 2)

Supply Voltage (V_{CC}) 5V \pm 10%
 LED on Current 10mA
 RRSET 61.9K Ω \pm 1%

ML4642 ELECTRICAL CHARACTERISTICS

Unless otherwise specified $T_A = 0^\circ\text{C}$ to 70°C (Note 3), $V_{CC} = 5\text{V} \pm 10\%$.

PARAMETER	CONDITIONS	MIN	TYP.	MAX	UNITS
Power Supply Current I_{CC} (Note 4)	$V_{CC} = 5\text{V}$		60	90	mA
LED Drivers: V_{OL}	$R_L = 510\Omega$ for $\overline{\text{CDLED}}$, $\overline{\text{TXLED1,2}}$, JAB2 $R_L = 270\Omega$ for JAB1/JDIS, $\overline{\text{RxLED/LPBK/SQE}}$ (Note 5)			0.8	V
Transmit Squelch Voltage Level (Tx+, Tx-)		-300	-250	-200	mV
Differential Output Voltage		± 550		± 1200	mV
Common Mode Output Voltage			4.0		V
Differential Output Voltage Imbalance			2	± 40	mV
$\overline{\text{RxLED/LPBK/SQE}}$	$\overline{\text{SQE}}$ Enabled/Loopback Enabled $\overline{\text{SQE}}$ Disabled/Loopback Enabled	0.4	0.6	0.3 0.8	V

Note 1: Absolute maximum ratings are limits beyond which the life of the integrated circuit may be impaired. All voltages unless otherwise specified are measured with respect to ground.

Note 2: Limits are guaranteed by 100% testing, sampling, or correlation with worst-case test conditions.

Note 3: Low Duty cycle pulse testing is performed at T_A .

Note 4: This does not include the current from the AUI pull down resistors or the LED output pins.

Note 5: LED drivers can sink up to 20mA, but V_{OL} will be higher.

ML4642 ELECTRICAL CHARACTERISTICS (Continued)

AC ELECTRICAL CHARACTERISTICS

SYMBOLS	PARAMETER	MIN	TYP.	MAX	UNITS
TRANSMIT					
t _{TXNPW}	Transmit Turn-On Pulse Width		20		ns
t _{TXFPW}	Transmit Turn-Off Pulse Width		180		ns
t _{XODY}	Transmitter Turn-On Delay		30		ns
t _{TXLP}	Transmit Loopback Startup Delay		40		ns
t _{TXSDY}	Transmit Steady State Prop. Delay		15		ns
t _{TXJ}	Transmitter Jitter		1		ns
RECEIVE					
t _{RXODY}	Receive Turn-On Delay		20		ns
t _{RXSDY}	Receive Steady State Prop. Delay		15		ns
t _{RXJ}	Receiver Jitter		1		ns
t _{AR}	Differential Output Rise Time 20% to 80% (Rx+/-, COL+/-)		3		ns
t _{AF}	Differential Output Fall Time 20% to 80% (Rx+/-, COL+/-)		3		ns
COLLISION					
t _{CPSQE}	Collision Present to SQE Assert	0		200	ns
t _{SQEXR}	Time for SQE to Deactivate after a collision	200		500	ns
t _{CLF}	Collision Frequency	8.5	10	11.5	MHz
t _{CLPDC}	Collision Pulse Duty Cycle	40	50	60	%
t _{SQEDY}	SQE Test Delay (Tx Inactive to SQE)	0.6	1.1	1.6	μs
t _{SQETD}	SQE Test Duration	0.5	1.0	1.5	μs
t _{SQEB}	SQE Blank Period	4		7	μsec
JABBER, LINK TEST AND LED TIMING					
t _{JAD}	Jabber Activation Delay	7	13.5	20	ms
t _{JRT}	Jabber Reset Unjab Time	250	450	750	ms
t _{JSQE}	Delay from Outputs Disabled to Collision Oscillator On		100		ns
t _{LEDT}	\overline{CDLED} , \overline{RxLED} , $\overline{TxLED1}$, $\overline{TxLED2}$ On Time	20	50	300	ms

TIMING DIAGRAMS

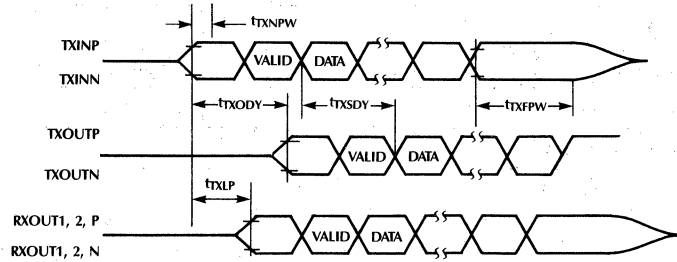


Figure 1. Transmit and Loopback Timing

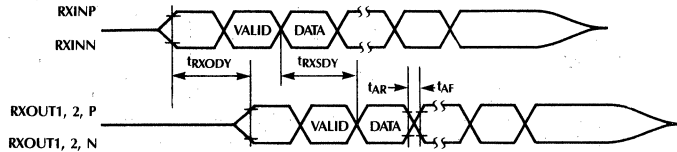


Figure 2. Receive Timing

TIMING DIAGRAMS (Continued)

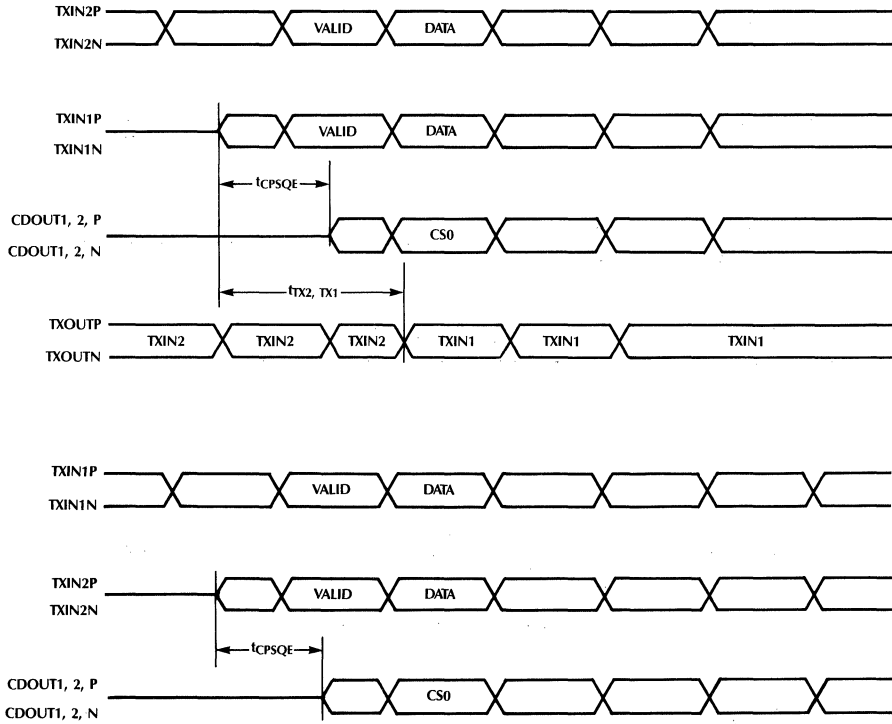


Figure 3. Collision Timing

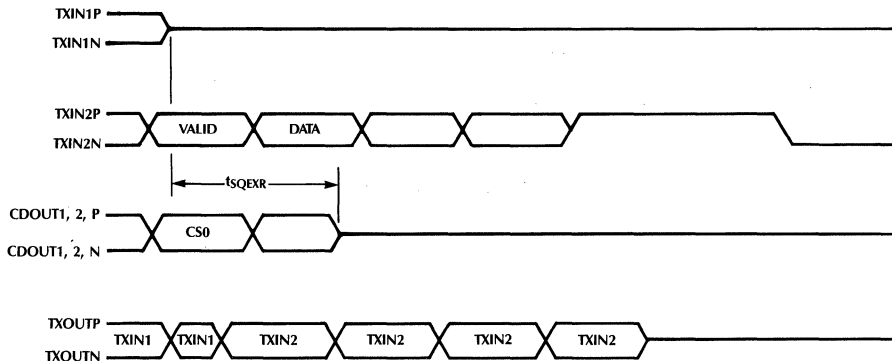


Figure 4. Collision Timing

TIMING DIAGRAMS (Continued)

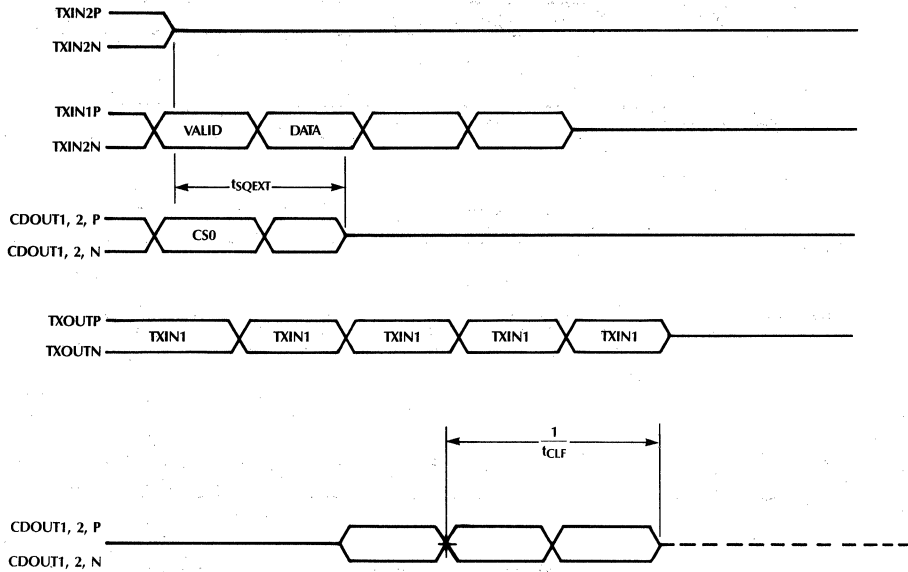


Figure 5. Collision Timing

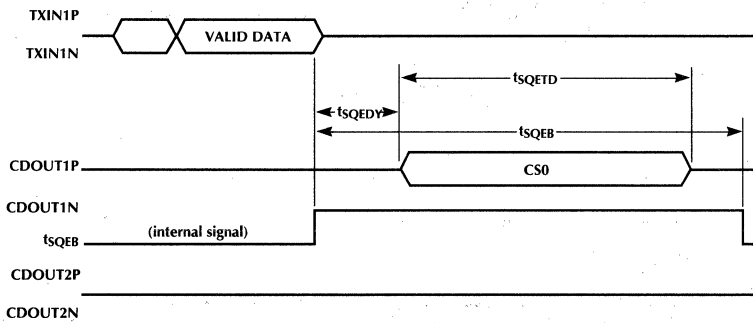


Figure 6. SQE Timing

TIMING DIAGRAMS (Continued)

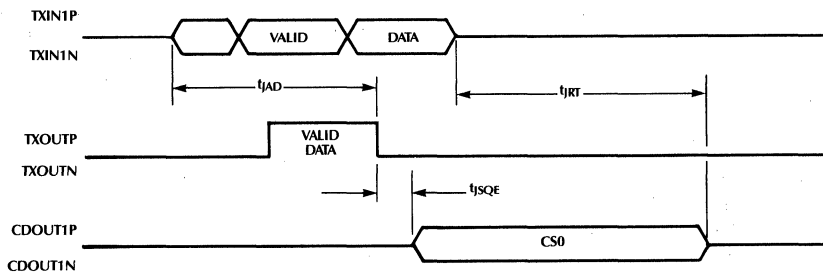


Figure 7. Jabber Timing

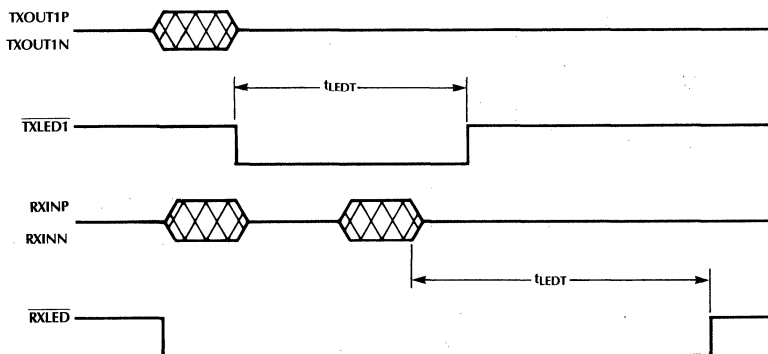


Figure 8. LED Timing

FUNCTIONAL DESCRIPTION

Figure 9 is a block diagram of a Two Port Multiplexer using the ML4642 chip. All AUI interfaces are shown AC coupled as they would be in an AUI multiplexer which does not include the MAU circuitry on the same board.

TRANSMISSION

The transmit function consists of detecting data on either of the TXIN differential receivers (TXIN1 or TXIN2) and transmitting this data out the TXOUT differential driver at the MAU port as well as both RXOUT1 and RXOUT2 drivers of the DTE ports. (Note: the looping back of data received at a TXIN pair to the RXOUT pairs is discussed in the Loopback section.)

Before data will be transmitted to the TXOUT and RXOUT pins from the TXIN pins it must meet the unsquelch requirements of the TXIN receiver circuitry. The squelch circuitry prevents any noise on the TXIN wires from being

misinterpreted as data and transmitted to the TXOUT and RXOUT pins. The squelch circuit rejects signals with pulse widths less than typically 20ns and voltage levels more positive than -250mV. Once the TXIN receiver is unsquelched it remains so until reception of the input idle signal, which is detected when the TXIN signal is more positive than -170mV for longer than 180ns.

RECEPTION

The receive function consists of detecting data at the RXIN differential receiver of the MAU port transmitting this data to both DTE port RXOUT pairs.

Before data will be transmitted to the RXOUT pins of the DTE ports it must meet the unsquelch requirements for the RXIN receiver circuitry. The squelch circuitry at the RXIN differential receiver input performs the same function as that of the TXIN squelch circuitry using the same noise rejection criteria.

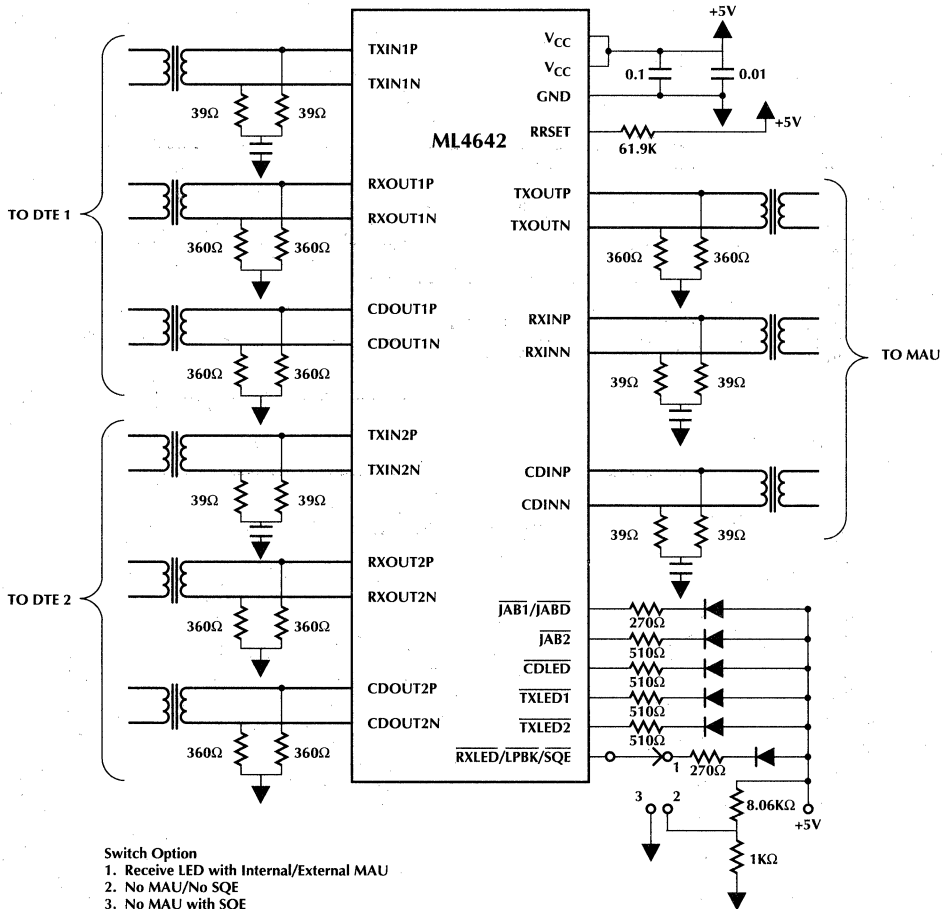


Figure 9. Two Port AUI Multiplexer

COLLISION

There are two conditions that constitute a collision from the point of view of the ML4642:

- a) If data is received at the TXIN inputs of both DTE ports simultaneously a **local collision** occurs within the ML4642.
- b) If the CDIN input is active at any time other than the inter-packet gap window allowed for the SQE Test function described below.

In either of the above circumstances it is necessary for the ML4642 to drive the CDOUT pairs on both DTE ports with the collision signal. The collision signal consists of a 10 MHz \pm 15% square wave matching the AUI specifications and capable of driving a 78 Ω load. The collision signal shall turn on within 2 bit times of the origination of the collision condition and shall turn off within 2–5 bit times after the collision condition subsides.

During a collision condition there are two sources for data to be transmitted to TXOUT, TXIN1 and TXIN2. The highest priority source for data to be transmitted to TXOUT is the TXIN1 receiver.

For example if TXIN2 begins transmission then TXIN1 turns on, the collision oscillator will turn on and TXOUT will switch from TXIN2 to TXIN1. If the collision ends by TXIN1 turning off first, TXOUT will switch from TXIN1 to TXIN2, and 2–5 bit times later the collision oscillator will turn off.

The MAU port's CDIN receiver contains squelch circuitry to prevent noise from causing the erroneous detection of a collision signal. A signal on the CDIN pair will not be considered active until it exceeds the same squelch requirements as those of the TXIN receivers.

LOOPBACK

The loopback function allows the ML4642 to emulate a coaxial transceiver by propagating the TXIN data back out the RXOUT pair of the same DTE port that is sourcing the data as well as the RXOUT pair of the idle DTE port. This allows the Ethernet controller sending the data to monitor its transmit packets and detect network faults.

The loopback function is enabled at both DTE ports when the RXLED pin is tied to ground, or 0.6 volts.

SQE TEST FUNCTION

The Signal Quality Error (SQE) Test function allows the DTE to determine whether or not the collision detection circuitry is functional. After each transmission, during the inter-packet gap time, the collision signal will be activated on the CDOUT pair of the same port as the TXIN pair which received the packet, for typically 1 μ s. The SQE function will not be activated on DTE ports of the ML4642 which are in the Jabber state. The SQE function is enabled on both DTE ports when the RXLED/LPBK/SQE pin is grounded.

JABBER

The jabber function prevents a babbling transmitter from loading down the network. Within the ML4642 is a jabber timer on each TXIN receiver. Each timer starts at the beginning of a received packet and resets at the end of each packet. If a packet lasts longer than 7 to 20ms the jabber logic disables its corresponding TXIN receiver (thus preventing its data from being retransmitted) and generates a collision signal on the babbling port's CDOUT pair. When the TXIN pair finally goes idle, a second timer measures 0.5 seconds of idle on TXIN prior to re-enabling the receiver and turning off the collision signal. If the TXIN pair becomes active again before the 0.5 seconds has expired, the timer is reset and measures another 0.5 seconds of idle time.

The jabber function can be disabled on both ports by tying the JAB1/JABD pin to ground.

LED DRIVERS

The ML4642 has six LED driver pins. Each DTE port has a transmit LED and a jabber LED and the MAU port has a receive LED. Additionally, there is a collision LED which indicates the presence of a collision condition. All LED drivers are active low 10mA current sources.

The TXLED, RXLED, and CDLED outputs have 50ms pulse stretchers on them to enable the LEDs to be visible. The JLED outputs do not have pulse stretchers on them because their conditions occur long enough for the LEDs to be visible.

Two of the ML4642 LED outputs serve as configuration pins as well. RXLED/LPBK/SQE and JAB1/JDIS may be tied through a resistor to V_{CC} , tied through a resistor and a LED to V_{CC} or grounded. Additionally RXLED/LPBK/SQE may be tied to a specific voltage. When these pins are grounded or tied to a 0.6 Volts they become configuration inputs. Otherwise when tied high they become status outputs.

CASCADING THE ML4642 FOR 4 AND 8 PORT DESIGNS

The configurability of such functions as loopback, jabber, and SQE allows ease of cascading multiple ML4642 chips for larger fan-out designs. Figure 10 shows a four port AUI Multiplexer design. For a type 0 configuration both jabber and transmit LEDs are available on a per port basis for status. The RXLED/LPBK/SQE pins are tied through a resistor to 5 volts, and CDLED is wire OR'ED with the other chip for one collision detect status LED per system. There is also only one receive LED status output which is displayed in a type 2 configuration. This particular pin in a type 2 configuration offers three options. In option 1, when tied to +5 volts through a resistor and an LED, an internal or external MAU will be connected. For stand-alone operation without an internal or external MAU a loopback is required. Option 2 allows loopback with no SQE test while option 3 provides loopback with an SQE test.

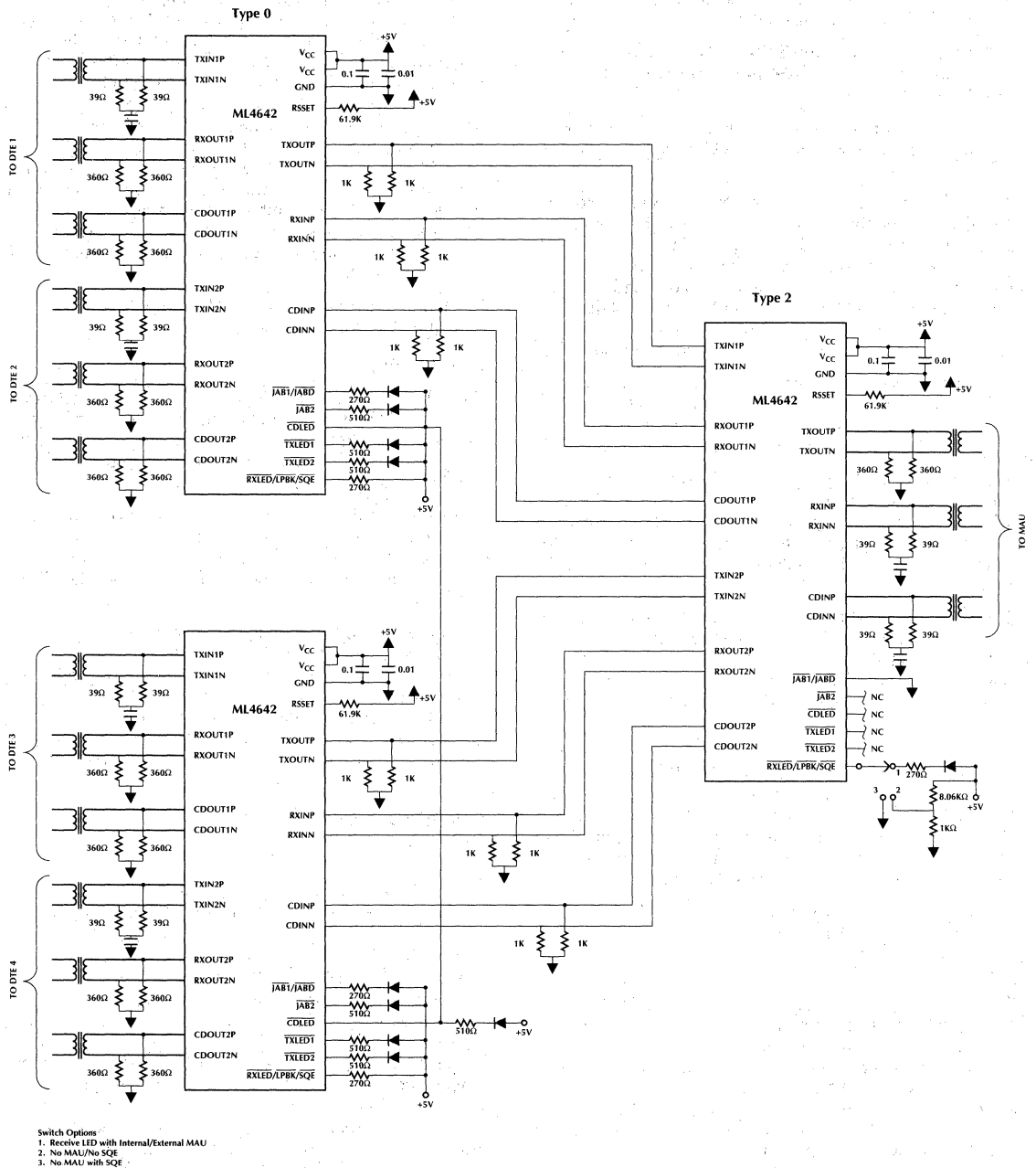


Figure 10. Four Port AUI Multiplexer

An eight port design is accomplished in the same way as shown in the block diagram in Figure 11. In an eight port design Type 0 and Type 2 configuration remain the same as in a four port design. Type 1 however only differs from Type 2 by tying RXLED/LPBK/SQE through a resistor to +5 volts. Table 1 summarizes all of the different LED configurations.

SQE TEST WHEN CASCADING

As mentioned before, after each transmission during the interpacket gap time the collision signal will be activated on the CDOUT pair of the same port as the TXIN pair which received the packet. When cascading ML4642s to implement 4 or 8 port designs, the path is remembered and followed to achieve this function. The paths that did not carry the transmit data blocks CDOUT for 4-7 μ sec after transmission to guarantee that only the port that transmitted will see SQE test.

TABLE 1. LED Configurations for 2, 4, and 8 Port Designs

	JAB1/JABD	RXLED/LPBK/SQE	JAB2	CDLED	TXLED1	TXLED2
Two Port AUI Mux	LED	GND, 0.6V, LED	LED	LED	LED	LED
Type 0	LED	270 Ω to +5V	LED	WIRE'ORED	LED	LED
Type 1	GND	270 Ω to +5V	NC	NC	NC	NC
Type 2	GND	GND, 0.6V, LED	NC	NC	NC	NC

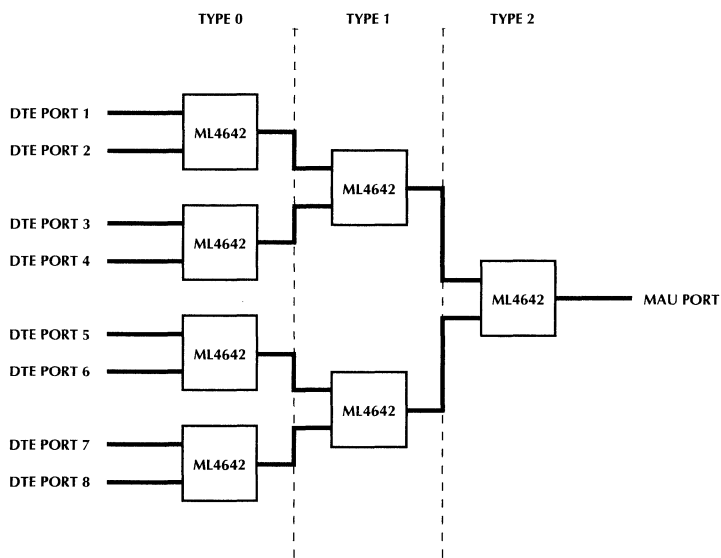


Figure 11. Eight Port AUI Multiplexer

ML4642

ORDERING INFORMATION

PART NUMBER	TEMPERATURE RANGE	PACKAGE
ML4642CR	0°C to 70°C	28-Pin SSOP (R28)

ML4652, ML4658

10BASE-T Transceiver

GENERAL DESCRIPTION

The ML4652/ML4658 10BASE-T Transceivers are single chip cable line driver/receivers that provides all of the functionality required to implement both an internal and external IEEE 802.3 10BASE-T MAU. These parts offer a standard IEEE 802.3 AU interface that allows them to directly connect to industry standard manchester encoder/decoder chips or an AUI cable.

These parts require a minimal number of external components, and are compliant to the IEEE 802.3 10BASE-T standard. The differential current driven transmitter offers superior performance because of its highly symmetrical switching. This results in low RFI noise and low jitter.

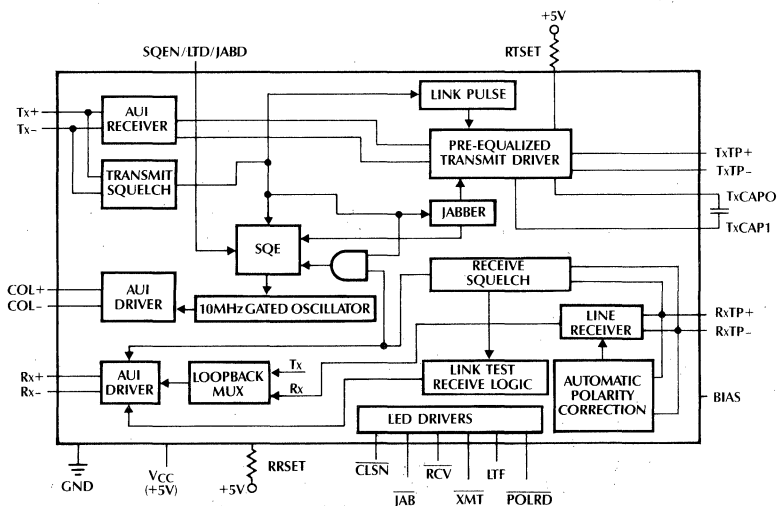
The Transceiver easily interfaces to 100Ω unshielded twisted pair cable, 150Ω shielded twisted pair cable, or a range of other characteristic impedances by simply changing one external resistor. Jabber, Link Test, and SQE Test are fully integrated onto the chip with enable/disable options. A polarity detection status pin, which can drive an LED, is provided for receive data, and the ML4658 offers automatic polarity correction.

The ML4652 and ML4658 are available in 24 pin skinny DIP as well as a 28 pin PLCC.

FEATURES

- Complete implementation of IEEE 802.3 10BASE-T Medium Attachment Unit (MAU)
- Incorporates an AU interface for use in an external MAU or internal MAU
- Single +5 volt supply $\pm 10\%$
- No crystal or clock input
- Current Driven Output for low RFI noise and low jitter
- Capable of driving 100Ω unshielded twisted pair cable or 150Ω shielded twisted pair cable
- Polarity detect status pin capable of driving an LED
- Automatic Polarity Correction on the ML4658
- On-chip Jabber logic, Link Test, and SQE test with enable/disable option
- ML4652 and ML4658 provide six network status LED output pins
- ML4652 and ML4658 are available in a 24 pin skinny DIP or 28 pin PLCC
- Semi-standard option using Micro Linear's FB3651 LAN Transceiver Tile Array

BLOCK DIAGRAM



PIN DESCRIPTION (DIP)

PIN NO.	NAME	FUNCTION
1	CLSN	Indicates that a collision is taking place. Active low LED driver, open collector. Event is extended 100ms for visibility.
2	COL+	Gated 10MHz signal used to indicate a collision, SQE test, or jabber. Balanced differential line driver outputs that meet AU interface specifications. AC or DC coupled.
3	COL-	
4	SQEN/LTD/JABD	SQE Test Enable, Link Test Disabled, Jabber Disabled. This input uses four voltage levels to configure the chip as shown in Table 1.

Table 1. SQEN/LTD/JABD Pin Configuration

Pin	SQE Test	Link Test	Jabber
0V (GND)	Disabled	Enabled	Enabled
1.2V BIAS	Disabled	Disabled	Disabled
5V (V _{CC})	Enabled	Disabled	Enabled
	Enabled	Enabled	Enabled

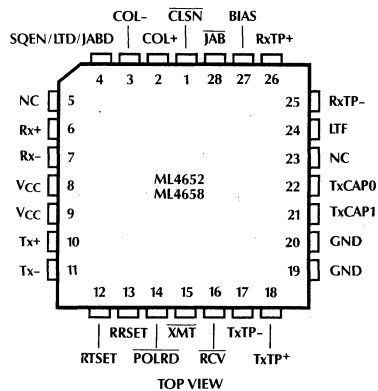
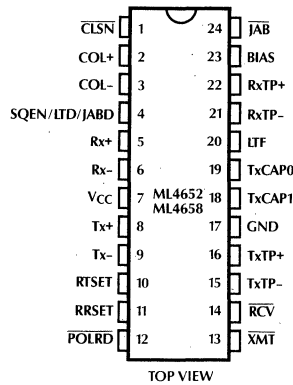
When link test is disabled, no link pulses are transmitted, and the transmitter and receiver will not be disabled as a result of a loss of receive link pulses. When jabber is disabled the transmitter can transmit continuously without interruption, and the collision oscillator will not be activated.

5	Rx+	Manchester encoded receive data output to the local device. Balanced differential line driver outputs that meet AU interface specifications. AC or DC coupled.
6	Rx-	
7	V _{CC}	+5 Volt power input.
8	Tx+	Balanced differential line receiver inputs that meet AU interface specifications. These inputs may be AC or DC coupled. When AC coupled, the BIAS pin is used to set the common mode voltage. Signals meeting the transmitter squelch input requirements are pre-equalized and output on TxTP+ and TxTP-.
9	Tx-	
10	RTSET	When using 100Ω unshielded twisted pair, a 220Ω resistor is tied between this pin and V _{CC} . When using 150Ω shielded twisted pair, a 330Ω resistor is tied between this pin and V _{CC} .
11	RRSET	A 1% 61.9KΩ resistor tied from this pin to V _{CC} is used for internal biasing.
12	POLRD	Receive Polarity status. Active low LED Driver, open collector output. Indicates the polarity of the receive twisted pair regardless of auto polarity correction. When this pin is high, the receive polarity is correct, and when this pin is low the receive polarity is reversed.
13	XMT	Indicates that transmission is taking place on the TxTP+, TxTP- pair. Active low LED driver, open collector. It is extended 100ms for visibility.
14	RCV	Indicates that the transceiver has unsquelched and is receiving data from the twisted pair. Active low LED driver, open collector. It is extended 100ms for visibility.
15	TxTP-	Pre-equalized differential balanced current driven output. These outputs are connected to a balanced transmit output filter which drives the twisted pair cable through pulse transformers. The output current is set with an external resistor connected to RTSET. allowing the chip to drive 100Ω unshielded twisted pair, 150Ω shielded twisted pair cables or a range of other characteristic impedances.
16	TxTP+	
17	GND	Ground reference.
18	TxCAP1	An external capacitor of 330pF is tied between these two pins to set the pulse width for the pre-equalization on the transmitter. If these two pins are shorted together, no pre-equalization occurs.
19	TxCAP0	
20	LTF	Link Test Fail. Active high. Normally this pin is low, indicating that the link is operational. If the link goes down resulting from the absence of link pulses or frames being received, the chip will go into the Link Test Fail state and bring LTF high. In the Link Test Fail state, both the transmitter and receiver are disabled, however link pulses are still sent. A station that only has access to the AUI can detect a Link Test Fail by the absence of loopback. This pin is low when the Link Test is disabled. Open collector LED output.

PIN DESCRIPTION (DIP) (Continued)

PIN NO.	NAME	FUNCTION
21	RxTP-	Twisted Pair receive data input. When this signal exceeds the receive squelch requirements the receive data is buffered and sent to the Rx+/- outputs.
22	RxTP+	
23	BIAS	Bias voltage, output. Used to bias the receive twisted pair inputs as well as the Tx+/- inputs when they are AC coupled.
24	$\overline{\text{JAB}}$	Open collector TTL output capable of driving an LED. When in the Jabber state, this pin will be low and the transmitter will be disabled. In the Jabber "OK" state this pin will be high.

PIN CONNECTION



ML4652, ML4658

ABSOLUTE MAXIMUM RATINGS

(Note 1)

Power Supply Voltage Range	
V_{CC}	-0.3 to 6V
Input Voltage Range	
Digital Inputs (SQEN, LTD)	-0.3 to V_{CC}
Tx+, Tx-, RxTP+, RxTP-	-0.3 to V_{CC}
Input Current	
RRSET, RTSET, JAB, CLSN, XMT, RCV, LTF	60mA
Output Current	
TxTP+, TxTP-	80mA
Storage Temperature	-65°C to +150°C
Lead Temperature (Soldering 10 seconds)	260°C

OPERATING CONDITIONS

(Note 2)

Supply Voltage (V_{CC})	5V \pm 10%
LED on Current	10mA
RRSET	61.9K Ω \pm 1%
RTSET	220 Ω \pm 1%
TxCAP	330pF

ELECTRICAL CHARACTERISTICS

Unless otherwise specified $T_A = 0^\circ\text{C}$ to 70°C (Note 3), $V_{CC} = 5\text{V} \pm 10\%$

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
Power Supply Current I_{CC} (Note 4)	$V_{CC} = 5\text{V}$			140	mA
LED Drivers: V_{OL}	$R_L = 510\Omega$ (Note 5)			0.8	V
Transmit Peak Output Current	RTSET = 220 Ω		42 (Note 6)		mA
Transmit Squelch Voltage Level (Tx+, Tx-)			-170		mV
Differential Input Voltage (RxTP+, RxTP-)		± 0.300		± 3.1	V
Receiver Input Resistance		10			K Ω
SQEN/LTD/JABD Input Resistance			12		K Ω
Receive Squelch Voltage Level (RxTP+, RxTP-)		300	450	585	mV-p
Differential Output Voltage (Rx+/-, COL+/-)		± 550		± 1200	mV
Common Mode Output Voltage (Rx+/-, COL+/-)			4.0		V
Differential Output Voltage Imbalance (Rx+/-, COL+/-)			2	± 40	mV
BIAS Voltage			3.2		V
SQEN/LTD/JABD	SQE TEST disabled All disabled Link Test disabled All Enabled	1.1 BIAS-0.15 $V_{CC}-0.05\text{V}$.3 1.4 BIAS+0.15	V

Note 1: Absolute maximum ratings are limits beyond which the life of the integrated circuit may be impaired. All voltages unless otherwise specified are measured with respect to ground.

Note 2: Limits are guaranteed by 100% testing, sampling, or correlation with worst-case test conditions.

Note 3: Low Duty cycle pulse testing is performed at T_A .

Note 4: This does not include the current from the AUI pull down resistors, the transmit pins TxTP+ and TxTP- or the LED output pins.

Note 5: LED drivers can sink up to 20mA, but V_{OL} will be higher.

Note 6: This current will result in a 2.5V peak output voltage on unshielded twisted pair cable when connected through an external filter and transformer as shown in Figure 12.

ELECTRICAL CHARACTERISTICS (Continued)**AC ELECTRICAL CHARACTERISTICS**

SYMBOL	PARAMETER	MIN	TYP	MAX	UNITS
Transmit					
t _{TXNPW}	Transmit Turn-On Pulse Width		20		ns
t _{TXFPW}	Transmit Turn-Off Pulse Width		180		ns
t _{TXLP}	Transmit Loopback Startup Delay			200	ns
t _{TXODY}	Transmitter Turn-On Delay			200	ns
t _{TXSDY}	Transmit Steady State Prop. Delay		15	100	ns
t _{TXJ}	Transmitter Jitter		±2	±3.5	ns
Receive					
t _{RXODY}	Receive Turn-On Delay if Transmit is Idle		420	500	ns
t _{RXTDY}	Receive Turn-On Delay if Transmit is Active		650	800	ns
t _{RXFX}	Last Bit Received to Start Slow Decay Output	230	800		ns
t _{RXSDY}	Receive Steady State Prop. Delay		15	100	ns
t _{RXJ}	Receiver Jitter		±0.7	±1.5	ns
t _{AR}	Differential Output Rise Time 20% to 80% (Rx+/-, COL+/-)		3		ns
t _{AF}	Differential Output Fall Time 20% to 80% (Rx+/-, COL+/-)		3		ns
Collision					
t _{CPSQE}	Collision Present to SQE Assert	0		900	ns
t _{TXRX}	Time for Loopback to Switch from Tx to RxTP During a Collision	0		900	ns
t _{SQEXR}	Time for SQE to Deactivate Given That RxTP Goes Idle and TxTP Continues	0		900	ns
t _{SQEXT}	Time for SQE to Deactivate Given That TxTP Goes Idle and RxTP Continues	0		900	ns
t _{CLF}	Collision Frequency	8.5	10	11.5	MHz
t _{CLPDC}	Collision Pulse Duty Cycle	40	50	60	%
t _{SQEDY}	SQE Test Delay (Tx Inactive to SQE)	0.6	1.1	1.6	μs
t _{SQETD}	SQE Test Duration	0.5	1.0	1.5	μs
Jabber, Link Test and LED Timing					
t _{JAD}	Jabber Activation Delay	20	70	150	ms
t _{JRT}	Jabber Reset Unjab Time	250	450	750	ms
t _{JSQE}	Delay from Outputs Disabled to Collision Oscillator On		100		ns
t _{LLT}	Link Loss Time	50	95	150	ms
t _{LTN}	Link Test Pulse Receive Minimum Time	2	4.2	7	ms
t _{LTX}	Link Test Pulse Receive Maximum Time	25	70	150	ms
t _{TLR}	Link Test Pulse Repetition Rate	8	16	24	ms
t _{LTPW}	Link Test Pulse Width	85	100	200	ns
t _{LEDT}	XMT, RCV, CLSN On Time	30	100	300	ms

TIMING DIAGRAMS

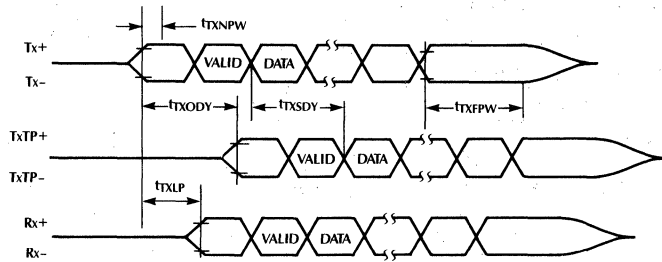


Figure 1. Transmit and Loopback Timing

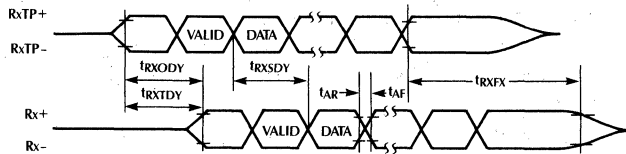


Figure 2. Receive Timing

TIMING DIAGRAMS (Continued)

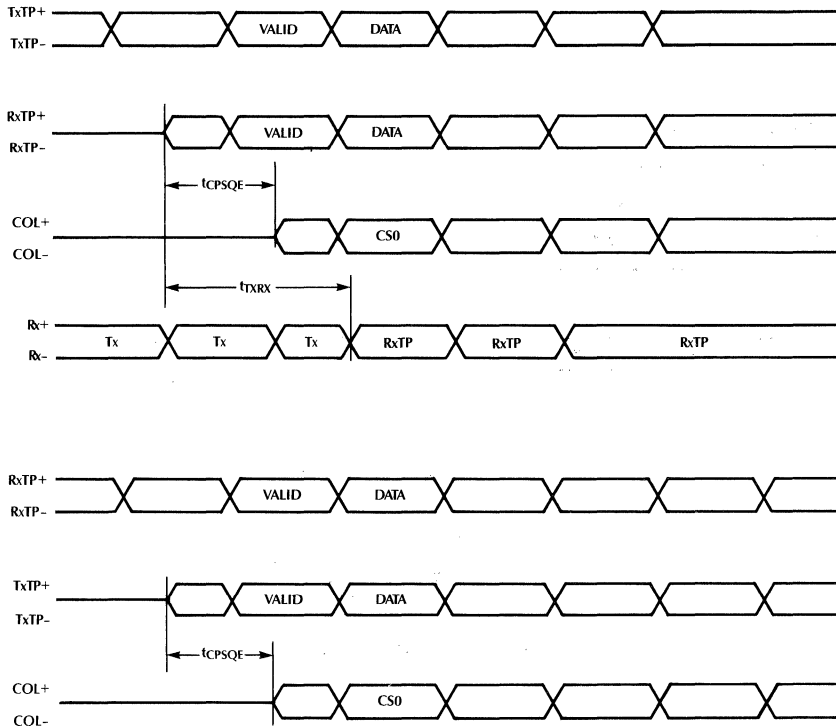


Figure 3. Collision Timing

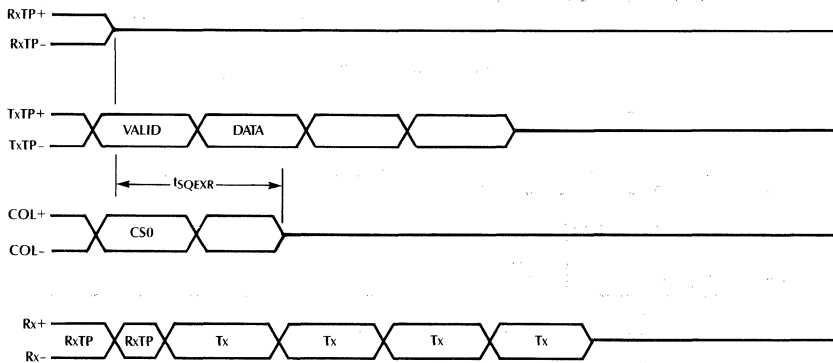


Figure 4. Collision Timing

TIMING DIAGRAMS (Continued)

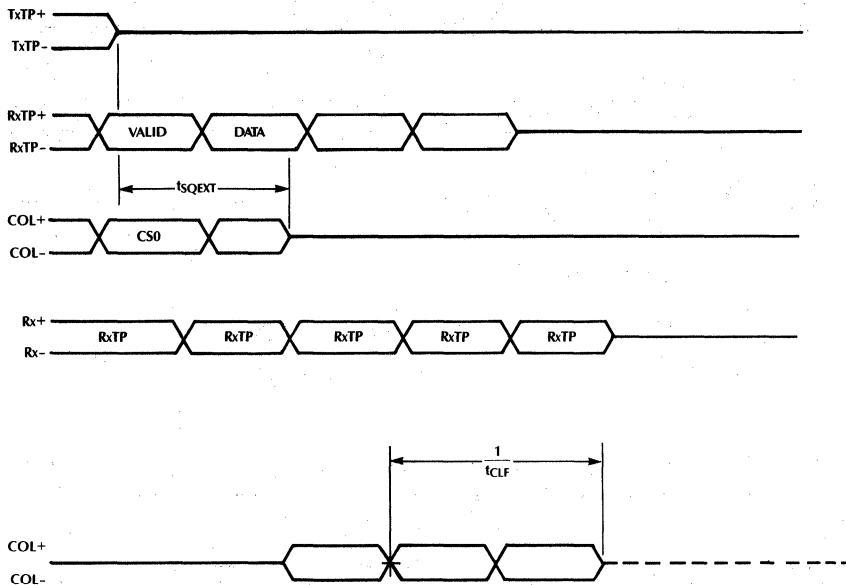


Figure 5. Collision Timing

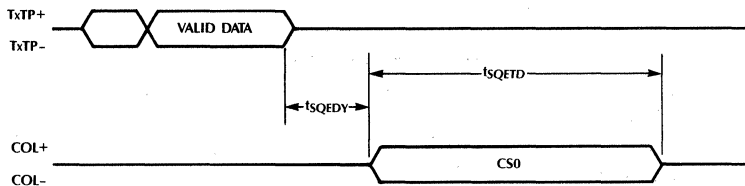


Figure 6. SQE Timing

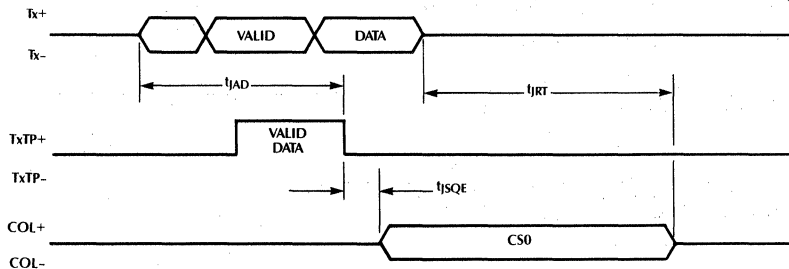


Figure 7. Jabber Timing

TIMING DIAGRAMS (Continued)

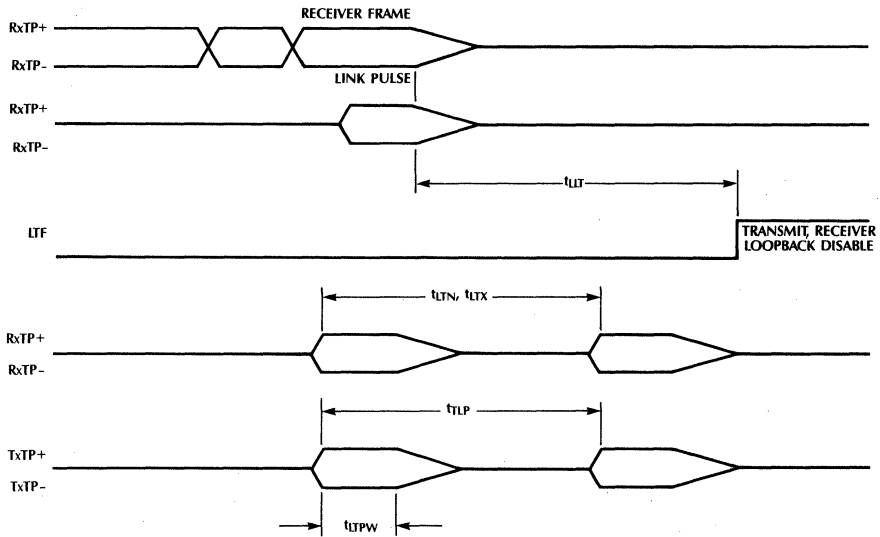


Figure 8. Link Pulse Timing

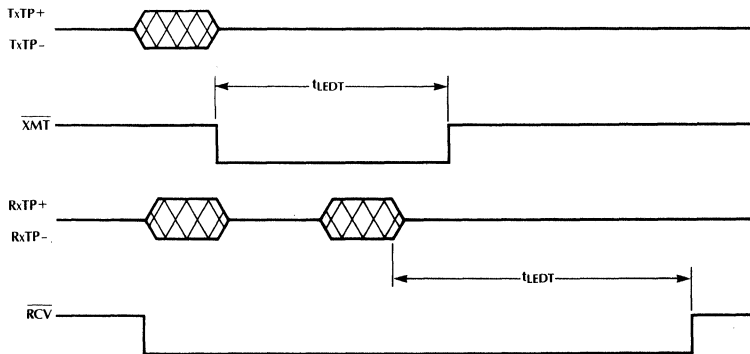


Figure 9. LED Timing

SYSTEM DESCRIPTION

Figure 10 shows a typical block diagram of an external 10BASE-T transceiver interface. On one side of the transceiver is the AU interface and the other is the twisted pair. The AU interface is AC coupled when used in an external transceiver or can be AC or DC coupled when used in an internal transceiver. The AU interface for an external transceiver includes isolation transformers, some biasing resistors, and a voltage converter for power.

The twisted pair side of the transceiver requires external transmit and receive filters, isolation transformers, and terminating resistors. These components can be obtained in a single hybrid package from suppliers listed in figure 12. The transmitter sends pre-equalized data through the transmit filters onto the twisted pair. The pre-equalized data uses a standard two step output waveform that lowers the amplitude of the 5MHz component so that at the receiving end both the 5MHz and 10MHz components have the same amplitude. The external transmit filter smooths the edges of the signal before passing it onto the twisted pair.

The receive pair side of the transceiver accepts the data after it passes through the isolation transformer and the receive low pass filter. Since this is an AC coupled input, the BIAS pin is used to set the proper common mode voltage for the receive inputs. A pair of 50Ω resistors correctly terminate the receive pair and provide a common mode for the Bias voltage connection point.

AU INTERFACE

The AU interface consists of 3 pair of signals, DO, CI and DI as shown in Figure 10. The DO pair contains transmit data from the DTE which is received by the transceiver and sent out onto the twisted pair. The DI pair contains valid data that has been either received from the twisted pair or looped back from the DO and output through the DI pair to the DTE. The CI pair indicates whether a transmit based collision has occurred. It is an output that oscillates at 10MHz. CI pair is also used for Jabber and SQE Test.

The transceiver may be AC or DC coupled depending on the application. For the AC coupled interface, the DO input must be DC biased (shifted up in voltage) for the proper common mode input voltage. The BIAS pin serves this purpose. When DC coupled, the manchester encoder/decoder transmit output pair provides this common mode voltage and the Bias pin is not connected.

The two 39Ω 1% resistors tied to the Tx+ and Tx- pins serve two purposes. They provide a point to connect the common mode bias voltage, and they provide the proper matching termination for the AU1 cable. The CI and DI pair, which are output drivers from the transceiver to the AU1 cable, require 360Ω pull down resistors when terminated with a 78Ω load. However on a DTE card, CI and DI do not need 78Ω terminating resistors. This also means that the pull down resistors on CI and DI can be 1KΩ or greater depending upon the particular manchester encoder/decoder chip used.

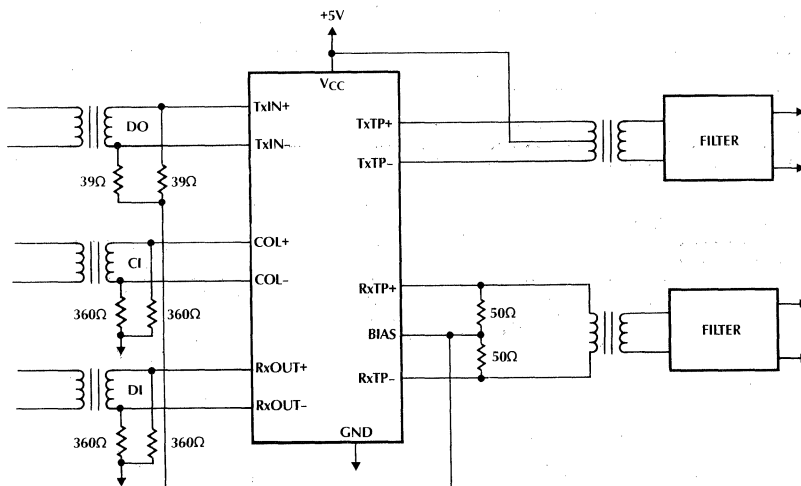


Figure 10. System Block Diagram

The AUI drivers are capable of driving the full 50 meters of cable length and have a rise and fall time of typically 3ns. The rise and fall times match to within 1ns. In the idle state, the outputs go to the same voltage to prevent DC standing current in the isolation transformers.

TRANSMISSION

The transmit function consists of detecting the presence of data from the AUI DO input (Tx+, Tx-) and driving that data onto the transmit twisted pair (TxTP+, TxTP-). A positive signal on the Tx+ lead relative to the Tx- lead of the DO circuit will result in a positive signal on the TxTP+ lead of the chip with respect to the TxTP- lead.

Before data will be transmitted onto the twisted pair from the AU interface, it must exceed the squelch requirements for the DO pair. The Tx squelch circuit serves the function of preventing any noise from being transmitted onto the twisted pair. This circuit rejects signals with pulse widths less than typically 20ns and voltage levels more positive than -175mV. Once the Tx squelch circuit has unsquelched, it looks for the start of idle signal to turn on the squelch circuit again. The transmitter turns on the squelch again when it receives an input signal at Tx+/- that is more positive than -175mV for more than approximately 180ns.

At the start of a packet transmission, no more than 2 bits are received from the DO circuit and not transmitted onto the twisted pair. The difference between start-up delays (bit loss plus steady-state propagation delay) for any two packets that are separated by 9.6 μ s or less will not exceed 200ns.

The output stage of the transmitter is a current mode switch which develops the output voltage by driving current through the terminating resistor and the output filter. The transmitter employs a center tap 2:1 transformer where the center tap is tied to V_{CC} (+5V). While one pin of the transmit pair (TxTP+, TxTP-) is pulled low, the other pin floats. The output pins to the twisted pair wires, TxTP+ and TxTP-, can drive a 100 Ω , 150 Ω load, or a variety of impedances that are characteristic of the twisted pair wire. RTSET selects the current into the TxTP+, TxTP- pins. This current along with the characteristic impedance of the cable determines the output voltage.

Once the characteristic impedance of the twisted pair is determined, one must select the appropriate RTSET resistor as well as match the terminating impedances of the transmit and receive filter. The RTSET resistor can be selected as follows:

$$RTSET = (R_L/100) * 220\Omega$$

where R_L is the characteristic impedance of the twisted pair cable.

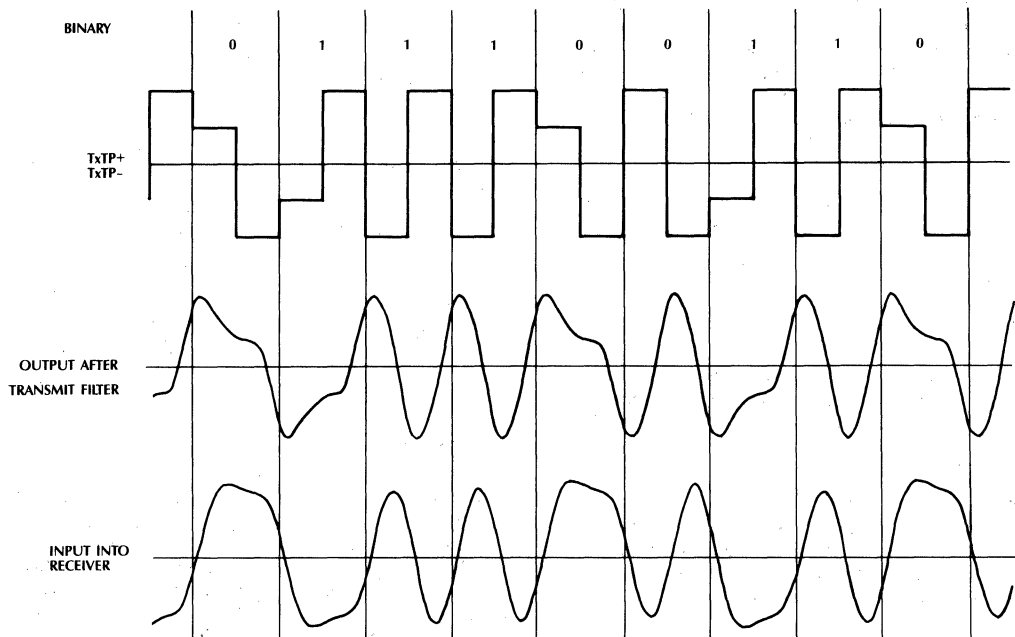


Figure 11. Transmit Pre-Equalization Waveform

ML4652, ML4658

The transmitter incorporates a pre-equalization circuit for driving the twisted pair line. Pre-equalization compensates for the amplitude and phase distortion introduced by the twisted pair cable. The twisted pair line will attenuate the 10MHz signal more than the 5MHz signal. Therefore pre-equalization insures that both the 5 and 10MHz components will be roughly the same amplitude at the far end receiver.

The pre-equalization circuit reduces the current output when a 5MHz bit is being transmitted. After 50ns of a 5MHz bit, the current level is reduced to approximately 2/3 of its peak for the remaining 50ns. Figure 11 illustrates the pre-equalization.

An on-chip one-shot determines the pulse width of the pre-equalized transmit signal. This requires an external capacitor connected to pins TxCAP0 and TxCAP1. The proper value for this one-shot is 330pF. Pre-equalization can be disabled by shorting TxCAP0 and TxCAP1 together.

The transmitter enters the idle state when it detects start of idle on Tx+ and Tx- input pins. The transmitter maintains a minimum differential output voltage of at least 450mV for 250ns after the last low to high transition. The driver differential output voltage will then be within 50mV of 0V within 45 bit times.

RECEPTION

The twisted pair receive data is transformer coupled and low pass filtered before it is fed into the input pins RxTP+/- . The input is differential with the common mode voltage set by the chip's Bias pin. At the start of packet reception from the twisted pair link, no more than 5 bits are received from the twisted pair cable and not transmitted onto the DI circuit. The first bit sent on the DI circuit may contain phase violations or invalid data, but all subsequent bits are valid.

The receive squelch will reject the following differential signals on the RxTP+ and RxTP- inputs:

1. All signals that produce a peak magnitude less than 300mV.
2. All continuous sinusoidal signals of amplitude less than $6.2V_{P-P}$ and frequency less than 2MHz.
3. All single sinusoidal cycles of amplitude less than $6.2V_{P-P}$ and either polarity, where the frequency is between 2MHz and 15MHz. For a period of 4 BT before and after this single cycle, the signal will conform to (1) above.
4. All sinusoidal cycles gated by a 100ns pulse gate of amplitude less than $6.2V_{P-P}$ and either polarity, where the sinusoidal frequency is between 2MHz and 30MHz. The off time of the pulse gate on the sinusoidal signal shall be at least 400ns.

The first three receive squelch criteria are required to conform to the 10BASE-T standard. The fourth receive squelch criteria exceeds the 10BASE-T requirements and enhances the performance of the receiver. The fourth squelch criteria prevents a false unsquelch caused by cross talk or noise typically found coupling from the phone lines onto the receive twisted pair.

When the receive squelch is on during idle, the input voltage must exceed approximately $\pm 450mV$ peak several times before unsquelch occurs. If the transmitter is inactive, the receiver has up to 5 bit times to unsquelch and output the receive data on the Rx+, Rx- pair. If the transmitter is active, the receive squelch extends the time it takes to determine whether to unsquelch. If the receiver unsquelches while the transmitter is active, a collision will result. Therefore the receive squelch uses the additional time to insure that a collision will not be reported as a result of a false receive squelch.

After the receiver is unsquelched, the detection threshold is lowered to 275mV. Upon passing the receive squelch requirements the receive data propagates into the multiplexer and eventually passes to the Rx+ and Rx- outputs of the AU interface. The addition of jitter through the receive section is no more than $\pm 1.5ns$.

While in the unsquelch state, the receive squelch circuit looks for the start of idle signal at the end of the packet. When start of idle is detected, receive squelch is turned on again. The proper start of idle occurs when the input signal remains above 300mV for 160ns. Nevertheless, if no transitions occur for 160ns, receive squelch is still turned on.

COLLISION

Whenever the receiver and the transmitter are active at the same time the chip will activate the collision output. The collision output is a differential square wave matching the AUI specifications and capable of driving a 78Ω load. The frequency of the square wave is $10MHz \pm 15\%$ with a 60/40 to 40/60 duty cycle. The collision oscillation turns on no more than 9 bit times after the collision condition begins, and turns off no more than 9 bit times after the collision condition is removed. The collision oscillator also is activated during SQE Test and Jabber.

LOOPBACK

The loopback function emulates a coax Ethernet transceiver where the transmit data sent by the DTE is looped back over the AUI receive pair. Many LAN controllers report the status of the carrier sense for each packet transmitted. The software can use this loopback information to determine whether a MAU is connected to the DTE by checking the status of carrier sense after each packet transmission.

When data is received by the chip while transmitting, a collision condition exists. This will cause the collision oscillator to turn on within 9 bit times. The data on the DI AUI pair (Rx+, Rx-) changes from Tx+, Tx- to RxTP+, RxTP-, when entering the collision state. During a collision, if the receive data (RxTP+, RxTP-) drops out before the transmit data (Tx+, Tx-), Rx+, Rx- will switch back to Tx+, Tx-.

SQE TEST FUNCTION (SIGNAL QUALITY ERROR)

The SQE test function allows the DTE to determine whether the collision detect circuitry is functional. After each transmission, during the inter-packet gap time, the collision oscillator will be activated for typically 1 μ s. The SQE test will not be activated if the chip is in the link fail state, or the jabber state.

For SQE to operate, the SQEN pin must be tied to V_{CC} or BIAS. The SQE test can be disabled by tying the SQEN pin to 1.2V or ground. This allows the chip to be interfaced to a repeater.

JABBER FUNCTION

The Jabber function prevents a babbling transmitter from bringing down the network. Within the transceiver is a Jabber timer that starts at the beginning of each transmission and resets at the end of each transmission. If the transmission lasts longer than 20ms the jabber logic disables the transmitter, and turns on the collision oscillator COL+, COL-. When Tx+ and Tx- finally go idle, a second timer measures 0.5 seconds of idle on Tx+ and Tx- before re-enabling the transmitter and turning off the collision oscillator. If transmission starts up again before 0.5 seconds has expired, the timer is reset and measures another 0.5 seconds of idle time.

Even though the transmitter is disabled during jabber, Link Pulses are still transmitted if the Link Test is enabled.

Jabber can be disabled by placing 1.2V on the SQEN/LTD/JABD pin. This is useful for measuring jitter performance on the transmitter.

LINK TEST FUNCTION

Transmission — Whenever data is not being delivered to the twisted pair link, the idle signal is applied. The idle signal is a sequence of Link Pulses separated by a 16ms period of silence. The idle signal starts with a period of silence after a packet transmission ends. The link test pulse is a single high pulse with the same amplitude requirements as the data signal.

Reception — The transceiver monitors the receive twisted pair input for packet and link pulse activity. If neither a packet nor a link test pulse is received for 50 to 150ms, the transceiver enters the Link Test Fail state and inhibits transmission and reception. Link pulses received with the wrong polarity will be ignored and cause the chip to go into link test fail.

A DTE can determine that the transceiver is in Link Test Fail one of two ways: it can monitor the LTF pin if the transceiver is internal, or it can monitor loopback. If the MAU is on-board the LTF pin can be sampled to determine that the transceiver is in the link fail state. If the MAU is external the DTE can monitor carrier sense during transmission. A loss of carrier sense is an indication of Link Test Fail State, since in Link Test Fail, loopback is disabled. Note that jabber also disables loopback but with Jabber the collision signal will be on.

When a packet, or two consecutive link test pulses is received from the twisted pair input, the transceiver will exit the Link Test Fail state upon transmit and receive data being idle, and re-enable transmission and reception.

Link test pulses that do not occur within at most 25 to 150ms of each other are not considered consecutive. In addition, detected pulses that occur within a time between 2 to 7ms of a previous pulse will be considered as noise by the link test circuitry.

POLARITY CIRCUITRY

The ML4652 offers polarity detection, while the ML4658 offers automatic polarity correction. The ML4652 and ML4658 are pin for pin compatible. The POLRD pin is used to report the status of the receive pair polarity. This pin reflects the true status of the receive polarity regardless of whether the part has autopolarity correction or not.

Polarity Detection — ML4652 — The internal circuitry uses the start of idle signal to determine the receive polarity. With the correct receive polarity, the Start of Idle signal (the end of the frame) will remain above 300mV for more than 160ns. If the polarity is reversed, the Start of Idle signal will end with a negative voltage.

The POLRD status pin is updated only when two consecutive frames are received with the same Start of Idle polarity. In the case where the part is powered up with the receive polarity reversed and no frames are received, the part will go into link test fail without reflecting a reverse polarity condition. Without autopolarity correction, the part will remain in link test fail unless a frame is received or the correct polarity link pulses are received.

ML4652, ML4658

Automatic Polarity Correction — ML4658 — In the link OK state, receive polarity is updated when two consecutive frames are received with the same Start of Idle polarity. In the Link Test Fail state the part will use either the Start of Idle signal or link pulses to correct the receive polarity.

In the case where the part is powered up with the receive polarity reversed and no frames are received, the part will go into Link Test Fail. After two link pulses are received with the same polarity, the part will exit Link Test Fail and correct the receive polarity. The POLRD pin will continue to reflect the true polarity of the receive pair.

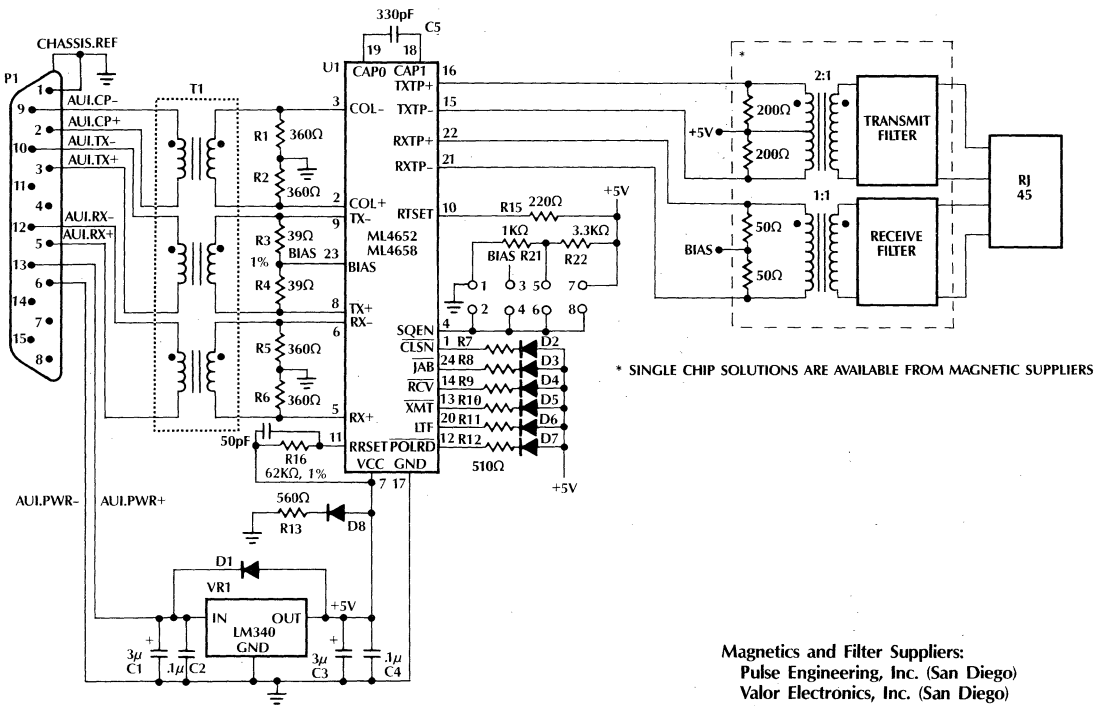
LED DRIVERS

The ML4652, ML4658 have six LED drivers for transmit, receive, collision, Link Test Fail, reverse polarity, and jabber. The LEDs are normally off except for LTF which is normally on and active high. The LEDs are tied to their respective pins through a 510Ω resistor to 5 Volts.

The $\overline{\text{XMT}}$, $\overline{\text{RCV}}$ and $\overline{\text{CLSN}}$ pins have pulse stretchers on them which enables the LEDs to be visible. When transmission or reception occurs, the LED XMT, RCV or CLSN status pins will activate low for 100ms. If another transmit, receive or collision condition occurs during the first 100ms, the LED timer will reset and begin timing again for 100ms. The LEDs will remain on for consecutive frames. The JAB, POLRD, and LTF LEDs do not have pulse stretchers on them since their conditions occur long enough for the eye to see.

SEMI-STANDARD OPTION

The ML4652 and ML4658 are designed using Micro Linear's Bipolar Tile Array technology. They use a special Tile Array, the FB3651, that was designed for Data Communications applications. As a result these parts are customizable, and can be modified to suit a specific customer application. Please contact your local representative or Micro Linear for more information on semi-standard options.



* SINGLE CHIP SOLUTIONS ARE AVAILABLE FROM MAGNETIC SUPPLIERS

Magnetics and Filter Suppliers:
 Pulse Engineering, Inc. (San Diego)
 Valor Electronics, Inc. (San Diego)
 Fil-Mag (San Diego)

Figure 12. External MAU

ML4652, ML4658

APPLICATION: INTERNAL MAU

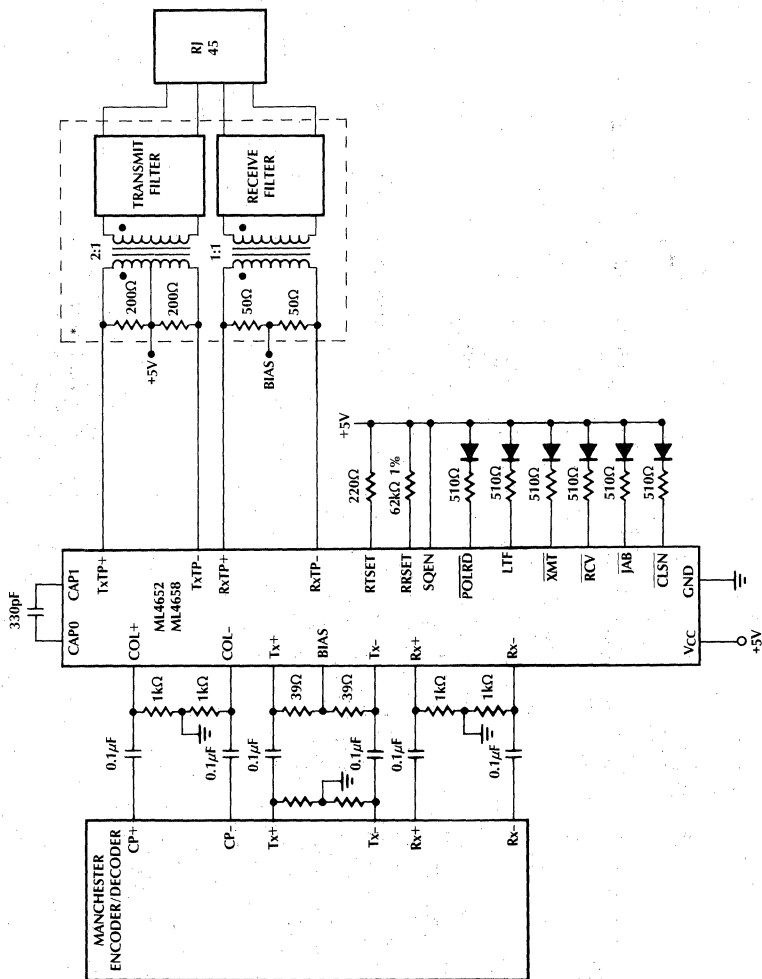


Figure 13. Internal MAU

ORDERING INFORMATION

ORDERING NUMBER	PACKAGE	PIN COUNT	AUTO-POLARITY
ML4652CP	Skinny DIP (P24N)	24 pins	No
ML4658CP	Skinny DIP (P24N)	24 pins	Yes
ML4652CQ	PLCC (Q28)	28 pins	No
ML4658CQ	PLCC (Q28)	28 pins	Yes

10BASE-T Transceiver for Multi-Port Repeaters

GENERAL DESCRIPTION

The ML4654 10BASE-T Transceiver is a single chip cable line driver/receiver that provides all of the functionality required to implement an internal 10BASE-T Transceiver for a Multi-Port Repeater. The ML4654 provides a TTL interface well suited for Multi-Port Repeater control logic.

The ML4654 uses a minimal number of external components, and fully conforms to the IEEE 802.3 10BASE-T standard. The transmitter offers a current driven output that is less sensitive to power supply variation and noise. It offers superior performance because of its highly symmetrical switching which results in low RFI noise and low jitter.

The Transceiver easily interfaces to 100Ω unshielded twisted pair cable, 150Ω shielded twisted pair cable, or a range of other characteristic impedances by changing one external resistor. Jabber and Link Test Function are fully integrated into the chip with enable/disable options. An autopolarity circuit detects the polarity of the receive pair and automatically corrects it if necessary. A polarity status pin that can drive an LED reflects the true polarity of the receive pair.

The ML4654 is available in a 20 pin skinny DIP as well as a surface mount 28 pin PLCC. The ML4654 is

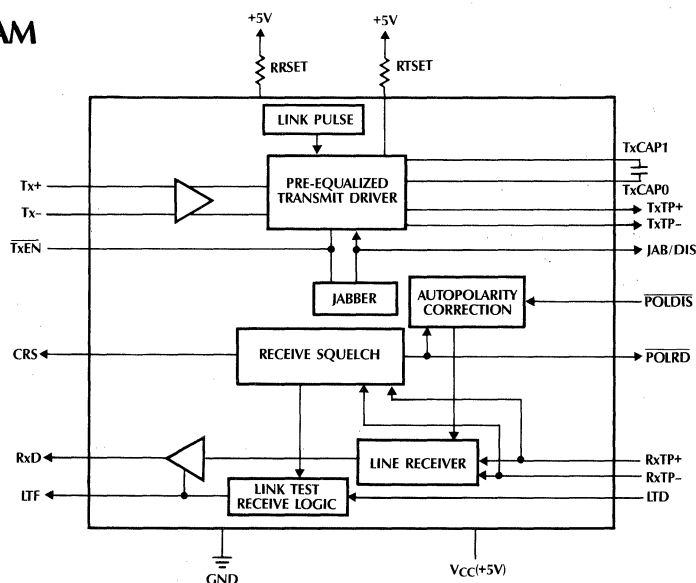
designed using Micro Linear's Bipolar Tile Array technology. It uses a special Tile Array designed for Data Communications applications. Semi-Standard options are available to suit a particular customer application.

FEATURES

- Complete implementation of IEEE 802.3 10BASE-T internal Medium Attachment Unit (MAU)
- TTL interface for direct connection to Multi-Port Repeater control logic
- Automatic polarity correction with a status pin to reflect the true receive polarity
- Single +5 volt supply $\pm 10\%$
- No clock or crystal required
- Capable of driving 100Ω unshielded twisted pair cable or 150Ω shielded twisted pair cable
- Fully integrated Link Test logic, with Link Test Fail Status pin and enable/disable option
- On-chip Jabber logic, with enable/disable option
- Available in a 20 pin skinny DIP or 28 pin PLCC
- Semi-standard option using Micro Linear's FB3651 LAN Transceiver Tile Array

4

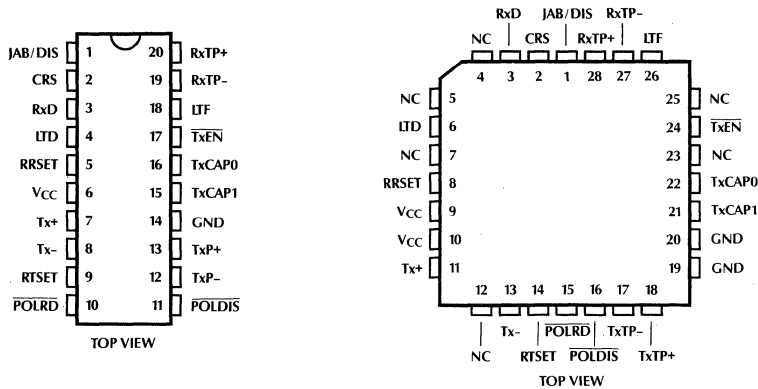
BLOCK DIAGRAM



PIN DESCRIPTION

PIN NO.	NAME	FUNCTION	PIN NO.	NAME	FUNCTION
1	JAB/DIS	Jabber Status Output/Jabber Disable. When this pin is tied to ground, the Jabber function is disabled and the transmitter is allowed to transmit indefinitely. This pin has an internal pullup so that when tied to a TTL input it will be low in the unjab state and high in the jab state. When in the jab state, the transmitter will be disabled.	11	POLDIS	Automatic Polarity Correction Disable. When grounded or left to float this pin will disable automatic polarity correction. The POLRD status pin continues to reflect the status of the receive polarity, even when automatic polarity is disabled. When this pin is tied high, automatic polarity correction is enabled.
2	CRS	Carrier Sense. Indicates valid receive data from the twisted pair. TTL output active high.	12	TxTP-	Pre-equalized differential balanced output driver. These outputs are connected to terminating resistors, a transformer and a balanced transmit output filter. The output current is set with an external resistor connected to RTSET allowing the chip to drive 100Ω unshielded twisted pair, 150Ω shielded twisted pair cables or a range of other characteristic impedances.
3	RxD	Receive data output to the local device. TTL levels.	13	TxTP+	
4	LTD	Link Test Disable. When tied high or left to float, link test is disabled. When Link Test is disabled no link pulses are transmitted, and the transmitter and receiver will not be disabled as a result of a loss of receive link pulses. When this pin is grounded, link pulses will be transmitting during idle, and the link test receive logic is enabled.	14	GND	Ground reference.
5	RRSET	A 1% 61.9KΩ resistor tied from this pin to V _{CC} is used for biasing internal nodes.	15	TxCAP1	An external capacitor of 330pF is tied between these two pins to set the pulse width for the pre-equalization on the transmitter. If these two pins are shorted together, no pre-equalization occurs.
6	V _{CC}	+5 Volt power input.	16	TxCAP0	
7	Tx+	Differential transmit data pair input from the local device, with TTL levels.	17	TxEn	
8	Tx-		When this pin is low the transmitter is enabled and transmitting the data received from the Tx+/- input pair. TTL input-active low.		
9	RTSET	When using 100Ω unshielded twisted pair cable, tie a 220Ω resistor between this pin and V _{CC} . When using 150Ω shielded twisted pair cable, tie a 330Ω resistor between this pin and V _{CC} .	18	LTF	Link Test Fail. Active high. Normally this pin is low, indicating that the link is operational. If the link goes down resulting from the absence of link pulses and frames being received, the chip will go into the Link Test Fail state and bring LTF high. In the Link Test Fail state, both the transmitter and receiver are disabled, however link pulses are still sent. This pin is low when Link Test is disabled.
10	POLRD	Polarity Reversal Detection. This pin reflects the true receive polarity status regardless of the state of the autopolarity logic. A low indicates that RxTP+ and RxTP- are reversed. Open collector TTL output.	19	RxTP-	Twisted Pair Receive Data Input. When this signal exceeds the receive squelch requirements the receive data is buffered and sent to the RxD output pin.
			20	RxTP+	

PIN CONNECTION



ABSOLUTE MAXIMUM RATINGS

(Note 1)

Power Supply Voltage Range	
V _{CC}	-0.3 to 6V
Input Voltage Range	-0.3 to V _{CC}
Input Current	60mA
Output Current	
TxTP+, TxTP-	80mA
Storage Temperature (T _j)	135°C
Lead Temperature (Soldering 10 seconds)	300°C

OPERATING CONDITIONS

(Note 2)

Supply Voltage (V _{CC})	5V ± 10%
LED on Current	10mA
RRSET	61.9KΩ ± 1%
RTSET	220Ω or 330Ω ± 1%
TxCAP	330pF

ELECTRICAL CHARACTERISTICS

Unless otherwise specified T_A = 0°C to 70°C (Note 3), V_{CC} = 5V ± 10%

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
Power Supply Current I _{CC} (Note 4)	V _{CC} = 5V			120	mA
TTL Inputs: V _{IL} V _{IH} (LTD, TxEN)		2		.8	V V
TTL Outputs: V _{OL} V _{OH} (CRS, RxD, LTF)	I _{OL} = 1.6mA I _{OH} = -400μA	2.4		.4	V V
LED Drivers: V _{OL} (JAB/DIS, POLRD)	R _L = 510Ω (Note 5)			0.8	V
Transmit Peak Output Current	RTSET = 220Ω		42 (Note 6)		mA
Differential Input Voltage (RxTP+, RxTP-)		±0.300		±3.1	V
Receiver Input Resistance		10			KΩ
Receive Squelch Voltage Level (RxTP+, RxTP-)		300	450	585	mVp

Note 1: Absolute maximum ratings are limits beyond which the life of the integrated circuit may be impaired. All voltages unless otherwise specified are measured with respect to ground.

Note 2: Limits are guaranteed by 100% testing, sampling, or correlation with worst-case test conditions.

Note 3: Low Duty cycle pulse testing is performed at T_A.

Note 4: This does not include the current supplied into the transmit pins TxTP+ and TxTP-.

Note 5: LED drivers can sink up to 20mA, but V_{OL} will be higher.

Note 6: This current will result in a 2.5V peak output voltage on unshielded twisted pair cable when connected through an external filter and transformer as shown in Figure 5.

ML4654

ELECTRICAL CHARACTERISTICS (Continued)

AC ELECTRICAL CHARACTERISTICS

SYMBOL	PARAMETER	MIN	TYP	MAX	UNITS
Transmit					
t_{TXEN}	Transmit Enable to Data Out			100	ns
t_{TXDIS}	Transmit Disable to Start Slow Decay			50	ns
t_{TXSDY}	Transmit Steady State Prop. Delay		15	100	ns
t_{TXJ}	Transmitter Jitter			± 2	ns
Receive					
t_{RXOCR}	Valid Receive Data to CRS Turn-On			500	ns
t_{RXTCR}	Valid Receive Data to CRS Turn-On if Transmit is Active			800	ns
t_{RXSDY}	Receive Steady State Prop. Delay		15	100	ns
t_{RXFCR}	Receive Turn-Off to CRS Inactive	150	230	300	ns
t_{RXJ}	Receiver Jitter			± 1.5	ns
t_{AR}	Rx Output Rise Time 20% to 80%		4		ns
t_{AF}	Rx Output Fall Time 20% to 80%		4		ns
Jabber, Link Test and LED Timing					
t_{JAD}	Jabber Activation Delay	20	70	150	ms
t_{JRT}	Jabber Reset Unjab Time	250	450	750	ms
t_{JJAB}	Delay from Outputs Disabled to JAB/DIS Active		-20		ns
t_{LLT}	Link Loss Time	50	95	150	ms
t_{LTN}	Link Test Pulse Minimum Time	2	4.2	7	ms
t_{LTX}	Link Test Pulse Maximum Time	25	70	150	ms
t_{LTPW}	Link Test Pulse Width	85	100	130	ns
t_{LTP}	Link Pulse Repetition Rate	8	16	24	ms

TIMING DIAGRAMS

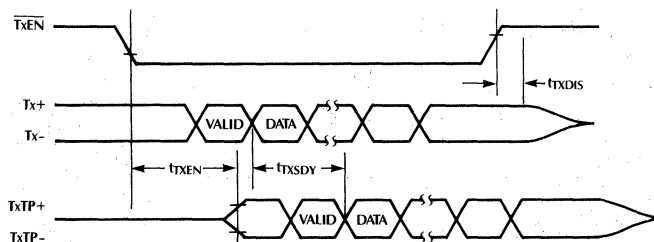


Figure 1. Transmit Timing

TIMING DIAGRAMS (Continued)

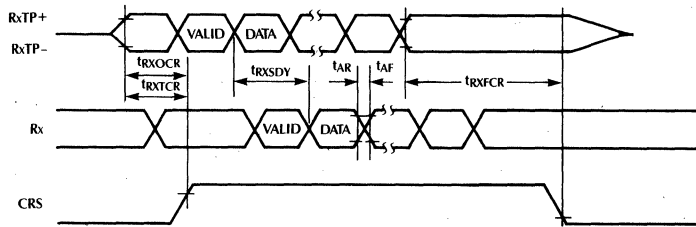


Figure 2. Receive Timing

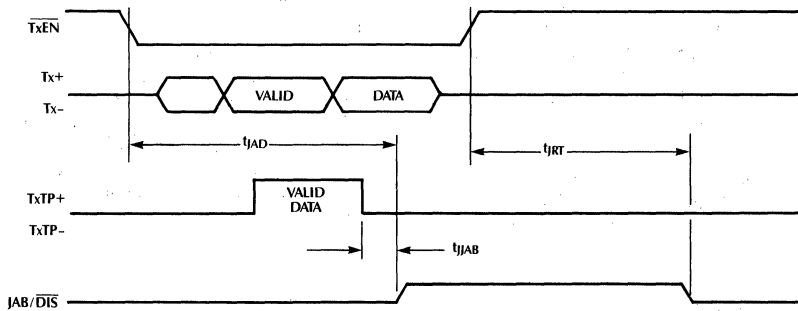


Figure 3. Jabber Timing

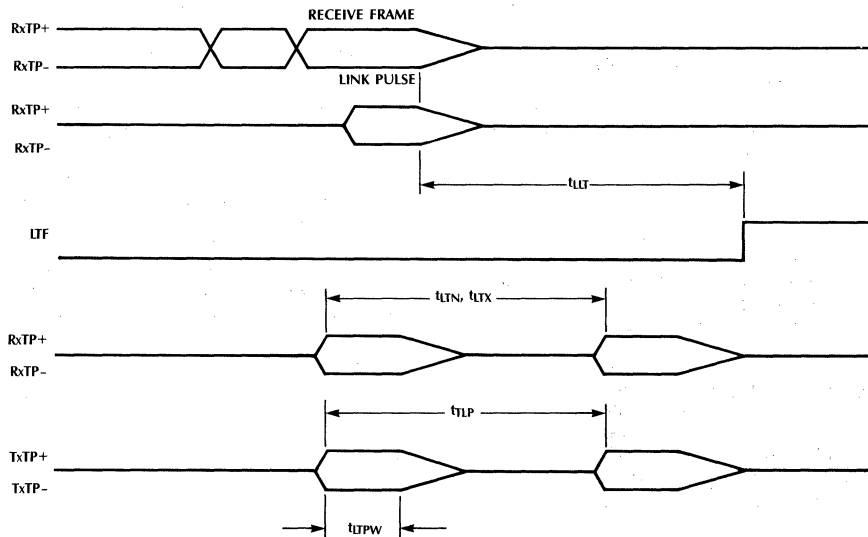


Figure 4. Link Pulse Timing

FUNCTIONAL DESCRIPTION

Figure 5 shows a typical block diagram of the ML4654 in an internal 10BASE-T transceiver interface. On one side of the transceiver is the local controller interface and the other is the twisted pair. The twisted pair side of the transceiver requires external transmit and receive filters, isolation transformer, and termination resistors.

The transmitter sends pre-equalized data through the transmit filters onto the twisted pair. The pre-equalized data uses a standard two step output waveform that lowers the amplitude of the 5MHz component so that at the receiving end both the 5MHz and 10MHz components have the same amplitude. The external transmit filter smooths the edges of the transmitter's output before passing it onto the twisted pair. Figure 6 illustrates the transmit output waveforms at different stages of the system.

The receive pair side of the transceiver accepts the data after it passes through the isolation transformer and the receive low pass filter. Since this is an AC coupled input, an internal DC bias is used to set the proper common mode voltage for the receive inputs.

LOCAL INTERFACE

The local interface consists of transmit, and receive signals which all use TTL levels. The transmit input signals entail a pair of true differential TTL transmit data pins, and an enable signal.

Once the transmitter is enabled, the output on TxTP+, TxTP- is determined by the transmit input pair Tx+, Tx-. The transmit input pair is a true differential TTL input that determines the switching point based on both inputs. Driving this input single ended is also possible by letting Tx- float. After the last bit is transmitted, Tx+ should be held high and Tx- held low for two bit times before TxEn goes high.

During reception the carrier sense pin (CRS) is activated asynchronously to receive data. Receive data is output through the receive data output pin (RxD). At the end of the packet, CRS goes inactive two bit times after the last low to high transition on RxD.

TRANSMISSION

The transmit function consists of enabling the transmitter with TxEn and driving the data onto the transmit twisted pair (Tx+, Tx-). A positive signal on the Tx+ lead relative to the Tx- lead results in a positive signal on the TxTP+ lead of the chip with respect to the TxTP- lead.

At the start of a packet transmission, no more than 1 bit is received from the Tx+, Tx- circuit and not transmitted onto the twisted pair. The difference between start-up delays (bit loss plus steady-state propagation delay) for any two packets that are separated by 9.6µs or less will not exceed 200ns.

The output stage of the transmitter is a current mode switch which develops the output voltage by driving current through the terminating resistor. The transmitter employs a center tap 2:1 transformer where the center tap is tied to V_{CC} (+5V). While one pin of the transmit pair is pulled low, the other pin floats.

The output pins to the twisted pair wires, TxTP+ and TxTP-, drive a 100Ω load, 150Ω load, or a variety of impedances that are characteristic of the twisted pair wire. To select the correct drive current for a characteristic impedance of the twisted pair wire, one must select the appropriate RTSET resistor. The RTSET resistor can be determined as follows:

$$RTSET = (R_L/100) * 220\Omega$$

where R_L is the characteristic impedance of the twisted pair cable.

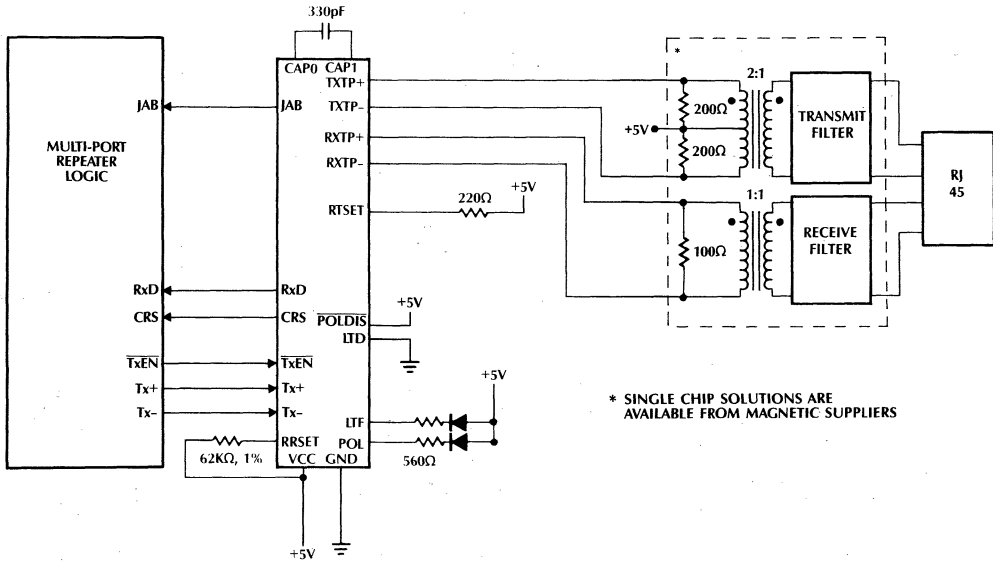
The transmitter incorporates a pre-equalization circuit for driving the twisted pair line. Equalization of the transmit signal is needed to decrease the voltage of the 5MHz component of the Manchester encoded signal. The twisted pair line will decrease the voltage of the 10MHz signal more than the 5MHz signal. Therefore the pre-equalization insures that both the 5 and 10MHz components will have the same amplitude at the far end receiver.

The pre-equalization circuit reduces the output current when a 5MHz bit is being transmitted. After 50ns of a 5MHz bit, the current level is reduced to approximately 2/3 of its peak for the remaining 50ns. Figure 6 illustrates the pre-equalization.

An on-chip one-shot determines the pulse width of the pre-equalized transmit signal. This requires an external capacitor connected to pins TxCAP0 and TxCAP1. The proper value for this capacitor is 330pf. Pre-equalization can be disabled by shorting TxCAP0 and TxCAP1 together.

The transmitter enters the idle state when it is disabled by TxEn. The Tx+ pin should remain high and the Tx- pin should remain low or float for two bit times before the TxEn signal goes high. When this happens, the transmitter maintains a minimum differential output voltage of at least 450mV for two bit times after the last low to high transition. The driver's differential output voltage will then be within 40mV of 0V within 80 bit times. In addition the current into the load will be limited in magnitude to 4mA within 80 bit times.

Figure 5. System Block Diagram



Magnetics and Filter Supplies:
 Pulse Engineering, Inc. (San Diego)
 Valor Electronics, Inc. (San Diego)
 Coilcraft (Cary, Illinois)
 Fil-Mag (San Diego)
 Bel Fuse (Jersey City)
 TDK (Torrance, CA)

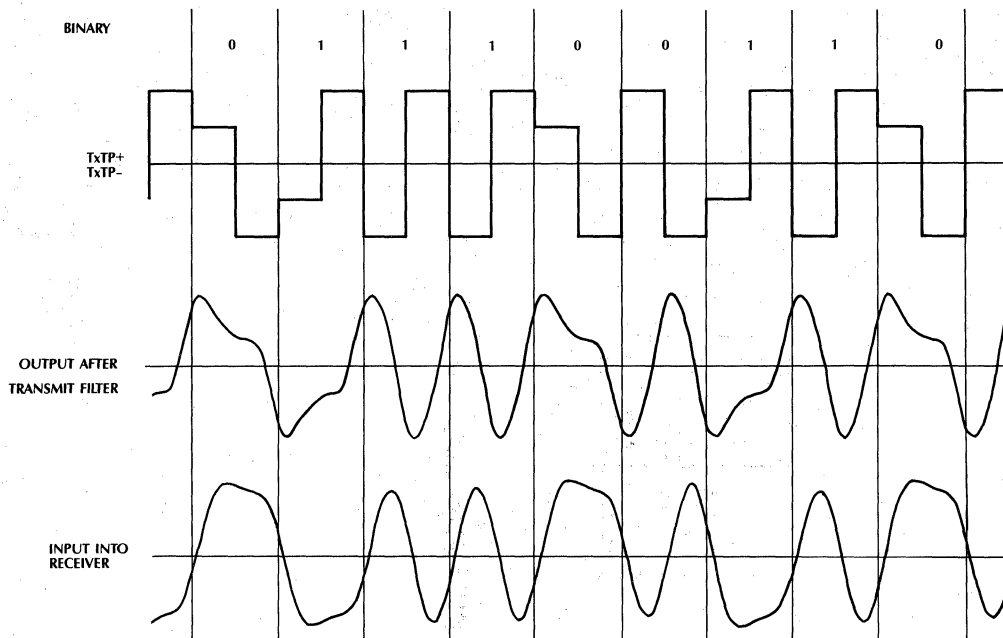


Figure 6. Transmit Output Waveforms

RECEPTION

Before the twisted pair receive data is input into the transceiver it is transformer coupled and low pass filtered. The RxTP+/- input is differential with the common mode voltage set internally at approximately halfway between V_{CC} and GND. At the start of a packet reception from the twisted pair link, no more than 5 bits are received from the twisted pair cable and not transmitted onto the Rx pin. The first bit sent to Rx may contain phase violations or invalid data, but all subsequent bits are valid.

The receive squelch will reject the following differential signals on the RxTP+ and RxTP- inputs:

1. All signals that produce a peak magnitude less than 300mV.
2. All continuous sinusoidal signals of amplitude less than $6.2V_{P-P}$ and frequency less than 2MHz.
3. All single sinusoidal cycles of amplitude less than $6.2V_{P-P}$ and either polarity, where the frequency is between 2MHz and 15MHz. For a period of 4 BT before and after this single cycle, the signal will conform to (1) above.
4. All sinusoidal cycles gated by a 100ns pulse gate of amplitude less than $6.2V_{P-P}$ and either polarity, where the sinusoidal frequency is between 2MHz and 30MHz. The off time of the pulse gate on the sinusoidal signal shall be at least 400ns.

The first three receive squelch criteria are required to conform to the 10BASE-T standard. The fourth receive squelch criteria exceeds the 10BASE-T requirements. It enhances the receiver's performance without compromising on conformance to the standard. The additional squelch criteria prevents a false unsquelch from occurring due to cross talk or noise typically coming from the telephone system twisted pair wires.

When the receive squelch is on, the input voltage must exceed $\pm 450mV$ peak several times before unsquelch occurs. If the transmitter is inactive, the receiver has up to 5 bit times to unsquelch and output the receive data on the Rx+, Rx- pair. If the transmitter is active, the receive squelch extends the time it takes to determine whether to unsquelch. If the receiver unsquelches while the transmitter is active, a collision will result. Therefore the receive squelch uses the additional time to insure that a collision will not be reported as a result of a false receive squelch.

After the receiver is unsquelched, the data detection threshold is lowered to 275mV. Upon passing the receive squelch requirements the receive data propagates to the Rx TTL output. This TTL output has been bolstered to reduce jitter. The addition of jitter through the receive section is no more than $\pm 1.5ns$.

While in the unsquelch state, the receive squelch circuit looks for the start of idle signal at the end of the packet. When start of idle is detected, receive squelch is turned on again and the carrier sense pin goes inactive. The proper start of idle occurs when the input signal remains above 300mV for 160ns. Nevertheless if no transitions occur for 160ns, receive squelch is still turned on. In this case however, the polarity may be reversed. A reverse polarity condition will be registered into the autopolarity circuit if the start of idle signal is negative. It will take several reverse polarity start of idle signals and/or reverse polarity link pulses to actually change the polarity on the receive circuit. (See Automatic Polarity Reversal section for more detail)

JABBER FUNCTION REQUIREMENTS

The Jabber function prevents a babbling transmitter from bringing down the network. Within the transceiver is a Jabber timer that starts at the beginning of each transmission and resets at the end of each transmission. If the transmission last longer than 20ms the jabber logic disables the transmitter, and activates the JAB/DIS pin, signaling the controller of the jabber condition. When Tx+ and Tx- finally go idle, a second timer measures 0.5 seconds of idle on Tx+ and Tx- before re-enabling the transmitter and deactivating the JAB/DIS pin. If transmission starts up again before the 0.5 seconds has expired, the timer is reset and measures another 0.5 seconds of idle time. Even though the transmitter is disabled during jabber, Link Pulses are still transmitted.

The Jabber function can be disabled by tying the JAB/DIS pin to ground. This forces the ML4654 into the Unjab state allowing indefinite transmission.

LINK TEST FUNCTION

Transmission — Whenever data is not being delivered to the twisted pair link, the idle signal is used. The idle signal is a sequence of link pulses separated by 16ms of silence. The idle signal starts with a 16ms period of silence after a packet transmission ends. The link test pulse is a single high pulse which meets the amplitude requirements for a pulse of duration BT.

Reception — The transceiver monitors its twisted pair input for packet and link pulse activity. If neither a packet nor a link test pulse is received for 50 to 150ms, the transceiver enters the Link Test Fail state and inhibits transmission and reception. The Hub Controller can determine that the transceiver is in the Link Test Fail state by monitoring the LTF pin. If LTF is low, the link is operational. But if LTF goes high, the ML4654 has entered the Link Test Fail state as a result of a loss of both Link Pulses and Receive Frames.

When a packet, or two consecutive link test pulses is received, the transceiver will exit the link test fail state. Exiting the link test fail state may be deferred if either TxEn is high or the receive squelch is off indicating receive data activity. After the link test fail state is exited, transmission and reception are re-enabled.

Link test pulses that do not occur within at most 25 to 150ms of each other are not considered consecutive. In addition, a Link Test Pulse that occurs within a time between 2 to 7ms of a previous Link Test Pulse will be considered as noise by the link test circuitry. In the Link Test Fail state, such pulses reset the counted number of consecutive link test pulses to zero.

AUTOMATIC POLARITY REVERSAL

This circuit determines the polarity for the receive pair only, and decides whether the polarity should be reversed. After 240ms of consistent reverse polarity information, the POLRD pin will change states and the polarity on the receive circuit will switch. The polarity on the receive pair RxTP+, RxTP- is determined using both Link Pulses and the Start of Idle signal at the end of a receive packet. When the Start of Idle signal is negative, it is treated as a reverse polarity indication. When a Link Pulse begins with a negative transition it is treated as a reverse polarity indication. When both Link Pulses and/or Start of Idle signals consistently indicate a reverse polarity condition for 240ms, the polarity on the receiver will be reversed.

The $\overline{\text{POLRD}}$ pin will reflect the true polarity on the receive pair regardless of the automatic correction circuit. For example if the polarity on the receive pair is reversed, after 240ms the $\overline{\text{POLRD}}$ pin will go low and the data on the RxD pin will have the correct polarity. This condition will remain as long as the polarity stays reversed. If the reverse polarity is then corrected, after 240ms the $\overline{\text{POLRD}}$ pin will go high and the RxD pin will have the correct polarity.

The $\overline{\text{POLDIS}}$ pin will disable the automatic polarity correction but have no affect on the POLRD pin. Therefore when $\overline{\text{POLDIS}}$ pin is tied low and the polarity is reversed on the receive pair, after 240ms the $\overline{\text{POLRD}}$ pin will go low, but the RxD will continue to pass the data on in the reverse polarity condition.

If the ML4654 is powered up with the RxTP+/- polarity reversed, and no data is received, it will go into link test fail. After 240ms of reverse polarity information, the auto-polarity circuit will reverse the polarity. The link test circuitry will then receive two correct polarity link pulses, and exit the link test fail state.

ML4654

ORDERING INFORMATION

ORDERING NUMBER	PACKAGE	PIN COUNT
ML4654CP ML4654CQ	Skinny DIP PLCC	20 pins (P20) 28 pins (Q28)

FOIRL Transceiver

GENERAL DESCRIPTION

The ML4661 FOIRL transceiver combined with the ML4621 or ML4622 fiber optic quantizers provides all of the functionality required to implement both an internal and external IEEE 802.3 FOIRL MAU. The ML4661 offers a standard IEEE 802.3 AU interface that allows it to be directly connected to industry standard manchester encoder/decoder chips or an AUI cable.

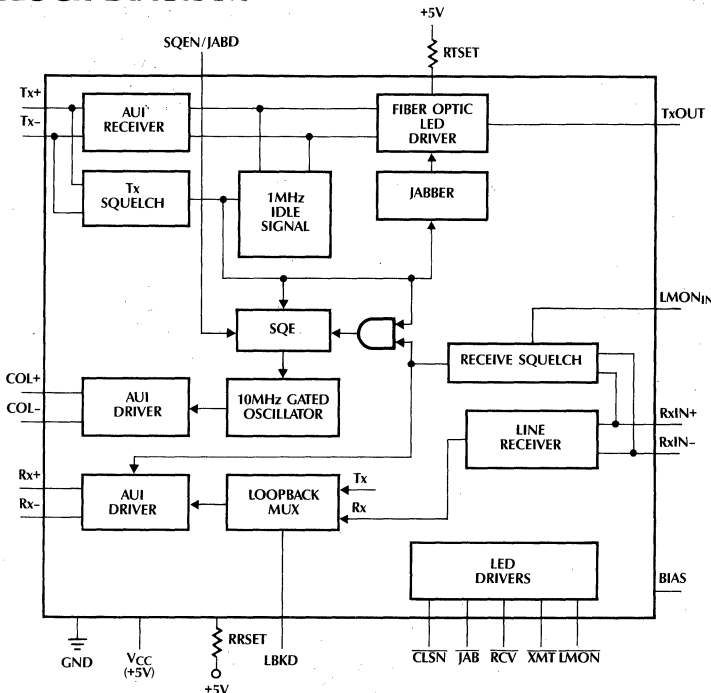
The ML4661 provides a highly integrated solution that requires a minimal number of external components, and is compliant to the IEEE 802.3 FOIRL standard. The transmitter offers a current driven output that directly drives a fiber optic LED transmitter. Jabber, 1MHz idle signal, and SQE Test are fully integrated onto the chip.

The receiver accepts an ECL level input coming from the ML4621 or ML4622 fiber optic quantizers. The 1MHz idle signal is removed and the AUI output is activated when the receive squelch criteria is exceeded. A Link Monitor function is also provided for low light detection.

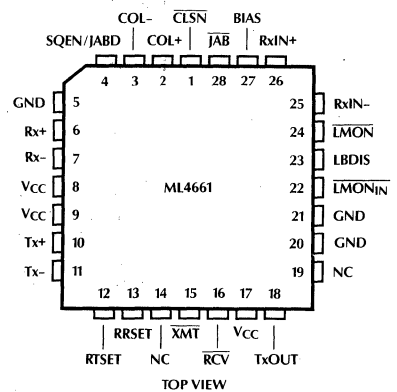
FEATURES

- Combined with the ML4621 or ML4622, offers a complete implementation of an FOIRL Medium Attachment Unit (MAU)
- Incorporates an AU interface for use in an external MAU or an internal MAU
- Single +5 volt supply $\pm 10\%$
- No crystal or clock required
- On-chip Jabber, 1MHz idle, and SQE Test with enable/disable option
- Five network status LED outputs
- Available in a 28-pin PLCC package
- Semi-standard option using Micro Linear's FB3651 LAN Transceiver Tile Array

BLOCK DIAGRAM



PIN CONNECTION



ML4661

PIN DESCRIPTION

PIN #	NAME	FUNCTION	PIN #	NAME	FUNCTION
1	CLSN	Indicates that a collision is taking place. Active low LED driver, open collector. Event is extended with internal timer for visibility.	17	V _{CC}	+5 volt supply.
2	COL+	Gated 10MHz oscillation used to indicate a collision, SQE test, or jabber. Balanced differential line driver outputs that meet AUI specifications.	18	TxOUT	Fiber optic LED driver output.
3	COL-		19	NC	No Connection.
4	SQEN/JABD	SQE Test Enable, Jabber Disable. When tied low, SQE test is disabled, when tied high SQE test is enabled. When tied to BIAS both SQE test and Jabber are disabled.	20	GND	Ground Reference.
5	GND	Ground Reference.	21	GND	Ground Reference.
6	Rx+	Manchester encoded receive data output to the local device. Balanced differential line driver outputs.	22	$\overline{\text{LMON}}_{\text{IN}}$	Link Monitor Input from the ML4621 or ML4622. This input must be low (active) for the receiver to unquench.
7	Rx-		23	LBDIS	Loopback Disable. When this pin is tied to V _{CC} , the AUI transmit pair data is not looped back to the AUI receive pair, and collision is disabled. When this pin is tied to GND (normal operation), the AUI transmit pair data is looped back to the AUI receiver pair.
8	V _{CC}	+5 Volt power input.	24	$\overline{\text{LMON}}$	Link Monitor LED status output. This pin is pulled low when $\overline{\text{LMON}}_{\text{IN}}$ input is low and there are transitions on RxIN \pm indicating an idle signal or active data. If either $\overline{\text{LMON}}_{\text{IN}}$ goes high or transitions cease on RxIN \pm , $\overline{\text{LMON}}$ will go high. Active low LED driver, open collector.
9	V _{CC}	Balanced differential line receiver inputs that meet AUI specifications. These inputs may be transformer, AC or DC coupled. When transformer or AC coupled, the BIAS pin is used to set the common mode voltage.	25	RxIN-	Fiber Optic receive pair. This ECL level signal is received from the ML4621 or ML4622 fiber optic quantizer. When this signal exceeds the receive squelch requirements, and the $\overline{\text{LMON}}_{\text{IN}}$ input is low, the receive data is buffered and sent to the AUI receive outputs.
10	Tx+		26	RxIN+	
11	Tx-		12	RTSET	Sets the current driven out of the transmitter.
13	RRSET	A 1% 61.9 K Ω resistor tied from this pin to V _{CC} sets the biasing currents for internal nodes.	14	NC	No Connection.
14	NC	Indicates that transmission is taking place. Active low LED driver, open collector. Event is extended with internal timer for visibility.	15	$\overline{\text{XMT}}$	Indicates that transmission is taking place. Active low LED driver, open collector. Event is extended with internal timer for visibility.
15	$\overline{\text{XMT}}$		16	$\overline{\text{RCV}}$	
16	$\overline{\text{RCV}}$		17	V _{CC}	+5 volt supply.
			18	TxOUT	Fiber optic LED driver output.
			19	NC	No Connection.
			20	GND	Ground Reference.
			21	GND	Ground Reference.
			22	$\overline{\text{LMON}}_{\text{IN}}$	Link Monitor Input from the ML4621 or ML4622. This input must be low (active) for the receiver to unquench.
			23	LBDIS	Loopback Disable. When this pin is tied to V _{CC} , the AUI transmit pair data is not looped back to the AUI receive pair, and collision is disabled. When this pin is tied to GND (normal operation), the AUI transmit pair data is looped back to the AUI receiver pair.
			24	$\overline{\text{LMON}}$	Link Monitor LED status output. This pin is pulled low when $\overline{\text{LMON}}_{\text{IN}}$ input is low and there are transitions on RxIN \pm indicating an idle signal or active data. If either $\overline{\text{LMON}}_{\text{IN}}$ goes high or transitions cease on RxIN \pm , $\overline{\text{LMON}}$ will go high. Active low LED driver, open collector.
			25	RxIN-	Fiber Optic receive pair. This ECL level signal is received from the ML4621 or ML4622 fiber optic quantizer. When this signal exceeds the receive squelch requirements, and the $\overline{\text{LMON}}_{\text{IN}}$ input is low, the receive data is buffered and sent to the AUI receive outputs.
			26	RxIN+	
			27	BIAS	BIAS output voltage for the AUI Tx+, Tx- inputs when they are AC coupled.
			28	$\overline{\text{JAB}}$	Jabber network status LED. When in the Jabber state, this pin will be low and the transmitter will be disabled. In the Jabber "OK" state this pin will be high. Open collector TTL output.

ABSOLUTE MAXIMUM RATINGS

(Note 1)

Power Supply Voltage Range
 V_{CC} -0.3 to 6V

Input Voltage Range
 Digital Inputs (SQEN, LMON_{IN}, LBDIS) -0.3 to V_{CC}
 $Tx+$, $Tx-$, $RxIN+$, $RxIN-$ -0.3 to V_{CC}

Input Current
 $RRSET$, $RTSET$, JAB , $CLSN$, XMT , RCV , $LMON$ 60mA

Output Current
 $TxOUT$ 100mA

Storage Temperature -65° C to +150°C
 Lead Temperature (Soldering 10 seconds) 260°C

OPERATING CONDITIONS

(Note 2)

Supply Voltage (V_{CC}) 5V ± 10%
 LED on Current 10mA
 $RRSET$ 61.9KΩ ± 1%
 $RTSET$ 162Ω ± 1%

ELECTRICAL CHARACTERISTICS

Unless otherwise specified, $T_A = -30^{\circ}C$ to $85^{\circ}C$, $V_{CC} = 5V \pm 10\%$ (Note 3)

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
Power Supply Current I_{CC} : Idle While Transmitting	$V_{CC} = 5V$, $RTSET = 162\Omega$ (Note 4)			170 200	mA
LED Drivers: V_{OL}	$R_L = 510\Omega$ (Note 5)			0.8	V
Transmit Peak Output Current	$RTSET = 162\Omega$		57		mA
Transmit Squelch Voltage Level ($Tx+$, $Tx-$)		-300	-250	-200	mV
Common Mode Input Voltage ($Tx\pm$, $RxIN\pm$)		2		$V_{CC} - 0.5$	V
Receive Squelch Voltage Level ($RxIN+$, $RxIN-$)			±175		mV-p
Differential Output Voltage ($Rx\pm$, $COL\pm$)		±550		±1200	mV
Common Mode Output Voltage ($Rx\pm$, $COL\pm$)			4.0		V
Differential Output Voltage Imbalance ($Rx\pm$, $COL\pm$)				±40	mV
BIAS Voltage			3.2		V
SQE/JABD	SQE Test Disable Jabber Disable Both Enabled	BIAS - .15 $V_{CC} - 0.05$.3 BIAS + .15	V
LBDIS Threshold	Disabled Enabled	$V_{CC} - 0.10$		1	V

4

AC ELECTRICAL CHARACTERISTICS

SYMBOL	PARAMETER	MIN	TYP	MAX	UNITS
Transmit					
t _{TXNPW}	Transmit Turn-On Pulse Width		20		ns
t _{TXFPW}	Transmit Turn-Off Pulse Width from Data to Idle	400		2100	ns
t _{TXLP}	Transmit Loopback Startup Delay			500	ns
t _{TXODY}	Transmitter Turn-On Delay			100	ns
t _{TXIDF}	Transmit Idle Frequency	0.85		1.25	MHz
t _{TXDC}	Transmit Idle Duty Cycle	45		55	%
t _{TXSDY}	Transmit Steady State Propagation Delay		15	50	ns
t _{TXJ}	Transmitter Jitter into 31Ω Load			±1.5	ns
Receive					
t _{RXSFT}	Receive Squelch Frequency Threshold	1.3		4	MHz
t _{RXODY}	Receive Turn-On Delay			350	ns
t _{RXFX}	Last Bit Received to Slow Decay Output	230	800		ns
t _{RXSDY}	Receive Steady State Propagation Delay		15	50	ns
t _{RXJ}	Receiver Jitter			±1.5	ns
t _{AR}	Differential Output Rise Time 20% to 80% (R _{x±} , COL±)		4		ns
t _{AF}	Differential Output Fall Time 20% to 80% (R _{x±} , COL±)		4		ns
Collision					
t _{CPSQE}	Collision Present to SQE Assert	0		400	ns
t _{SQEXR}	Time for SQE to Deactivate After Collision	450		1,000	ns
t _{CLF}	Collision Frequency	8.5		11.5	MHz
t _{CLPDC}	Collision Pulse Duty Cycle	40	50	60	%
t _{SQEDY}	SQE Test Delay (Tx Inactive to SQE)	0.6		1.6	μs
t _{SQETD}	SQE Test Duration	0.5	1.0	1.5	μs
Jabber and LED Timing					
t _{JAD}	Jabber Activation Delay	20	70	150	ms
t _{JRT}	Jabber Reset Unjab Time	250	450	750	ms
t _{JSQE}	Delay from Outputs Disabled to Collision Oscillator On		100		ns
t _{LEDTRC}	R _{CV} , C _{LSN} On Time	10		70	ms
t _{LEDTT}	X _{MIT} On Time	5		30	ms
t _{LLPH}	Low Light Present to LMON High	3		10	μs
t _{LLCL}	Low Light Clear to LMON Low	1		6	ms

Note 1: Absolute maximum ratings are limits beyond which the life of the integrated circuit may be impaired. All voltages unless otherwise specified are measured with respect to ground.

Note 2: Limits are guaranteed by 100% testing, sampling, or correlation with worst-case test conditions.

Note 3: Low Duty Cycle pulse testing is performed at T_A.

Note 4: This does not include the current from the AUI pull down resistors, or LED status outputs.

Note 5: LED drivers can sink up to 20mA, but V_{OL} will be higher.

SYSTEM DESCRIPTION

Figure 1 shows a typical block diagram of the ML4661 in an internal or external FOIRL MAU. On one side of the transceiver is the AU interface and the other is the fiber optic interface. The AU interface is AC coupled when used in an external transceiver or can be AC or DC coupled when used in an internal transceiver. The AU interface for an external transceiver includes isolation transformers, some biasing resistors, and a voltage regulator for power.

The fiber optic side of the transceiver requires an external fiber optic transmitter, fiber optic receiver, and the ML4621 or ML4622 fiber optic quantizers. The transmitter uses a current driven output that directly drives the fiber optic transmitter. The receive side of the transceiver accepts the data after passing through the fiber optic receiver and the ML4621/ML4622 fiber optic quantizer.

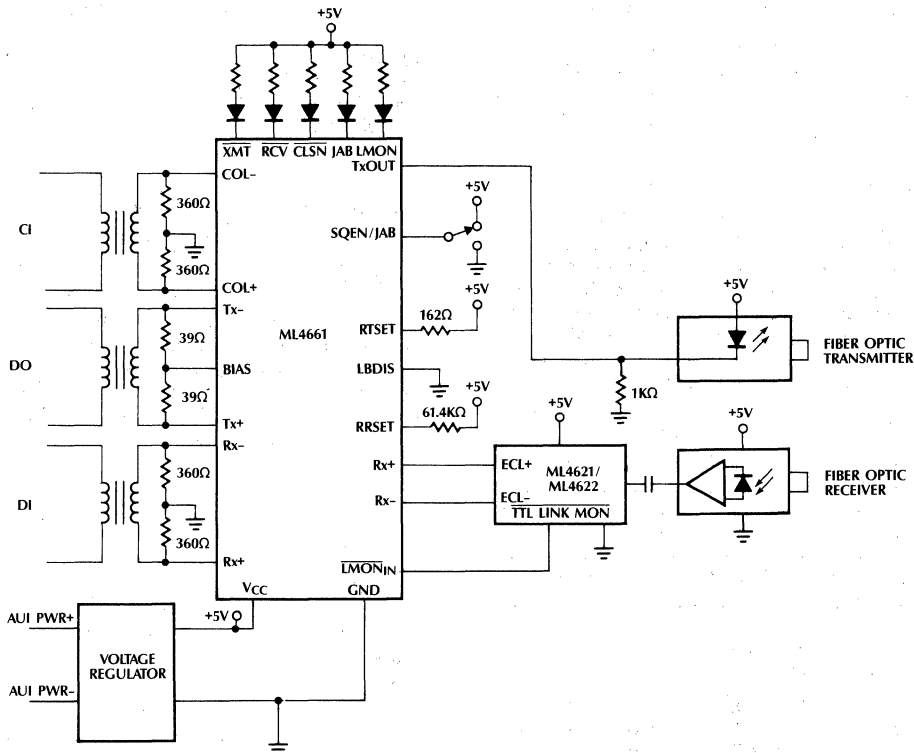


Figure 1. FOIRL System Block Diagram

AUI INTERFACE

The AUI interface consists of 3 pairs of signals, DO, CI and DI as shown in figure 1. The DO pair contains transmit data from the DTE which is received by the transceiver and sent out onto the fiber optic cable. The DI pair contains valid data that has been either received from the fiber optic cable or looped back from the DO and output through the DI pair to the DTE. The CI pair indicates whether a collision has occurred. It is an output that oscillates at 10MHz if a collision jabber or SQE Test has taken place, otherwise it remains idle.

When the transceiver is external, these three pairs are AC coupled through isolation transformers, while an internal transceiver may be AC or DC coupled. For the AC coupled interface, DO which is an input must be DC biased (shifted up in voltage) for the proper common mode input voltage. The BIAS pin serves this purpose. When DC coupled, transmit output pair coming from the serial interface provides this common mode voltage and the BIAS pin is not connected.

The two 39 Ω 1% resistors tied to the Tx+ and Tx- pins serve two purposes. First they provide a point to connect the common mode bias voltage as discussed above, and they provide the proper matching termination for the AUI cable. The CI and DI pair, which are output from the transceiver to the AUI cable, require 360 Ω pull down resistors when terminated with a 78 Ω load. However on a DTE card, CI and DI do not need 78 Ω terminating resistors. This also means that the pull down resistors on CI and DI can be 1K Ω or greater depending upon the particular manchester encoder/decoder chip used. Using higher value pull down resistors as in a DTE card will save power.

The AUI drivers are capable of driving the full 50 meters of cable length and have a rise and fall time of typically 3ns. In the idle state, the outputs go to the same voltage to prevent DC standing current in the isolation transformers.

TRANSMISSION

The transmit function consists of detecting the presence of data from the AUI DO input (Tx+, Tx-) and driving that data onto the fiber optic LED transmitter. A positive signal on the Tx+ lead relative to the Tx- lead of the DO circuit will result in no current, hence the fiber optic LED is in a low light condition. When Tx+ is more negative than Tx-, the ML4661 will sink current into the chip and the LED will light up.

Before data will be transmitted onto the fiber optic cable from the AUI interface, it must exceed the squelch requirements for the DO pair. The Tx squelch circuit serves the function of preventing any noise from being transmitted onto the fiber. This circuit rejects signals with pulse widths less than typically 20ns (negative going), or with levels less than -250mV. Once Tx squelch circuit has unsquelched, it looks for the start of idle signal to turn on the squelch circuit again. The transmitter turns on the squelch again when it receives an input signal at TxIN \pm that is more positive than -250mV for more than approximately 180ns.

At the start of a packet transmission, no more than 1 bit is received from the DO circuit and not transmitted onto the fiber optic cable. The difference between start-up delays (bit loss plus steady-state propagation delay) for any two packets that are separated by 9.6 μ s or less will not exceed 200ns.

The output stage of the transmitter is a current mode switch which develops the output light by sinking current through the LED into the TxOUT pin. Once the current requirement for the LED is determined, the RTSET resistor is selected. The following equation is used to select the correct RTSET resistor:

$$RTSET = \left(\frac{42mA}{I_{OUT}} \right) 220\Omega$$

The transmitter enters the idle state when it detects start of idle on Tx+ and Tx- input pins. After detecting the start of idle the transmitter switches to a 1MHz output idle signal.

RECEPTION

The input to the transceiver comes from the ECL outputs of the ML4621 or ML4622. At this point it is a clean digital ECL signal. At the start of packet reception no more than 3.5 bits are received from the fiber cable and not transmitted onto the DI circuit. The receive squelch will reject frequencies lower than 1.3MHz and input voltage less than $\pm 175mV$. The receive squelch will also reject any receive input if the LMON_{IN} pin is high.

While in the unsquelch state, the receive squelch circuit looks for the start of idle signal at the end of the packet. Start of idle occurs when the input signal remains idle for more than 160ns. When start of idle is detected, the receive squelch circuit returns to the squelch state and the start of idle signal is output on the DI circuit (Rx+, Rx-).

COLLISION

Whenever the receiver and the transmitter are active at the same time the chip will activate the collision output. The collision output is a differential square wave matching the AUI specifications and capable of driving a 78 Ω load. The frequency of the square wave is 10MHz \pm 15% with a 60/40 to 40/60 duty cycle. The collision oscillation turns on no more than 4.5 bit times after the collision condition begins, and turns off between 4.5 and 7 bit times after the collision condition is removed. The collision oscillator also is activated during SQE Test and Jabber.

LOOPBACK

The loopback function emulates an Ethernet (10BASE-5) transceiver whereby the transmit data sent by the DTE is looped back over the AUI receive pair. Some LAN controllers use this loopback information to determine whether a MAU is connected by monitoring the carrier sense while transmitting. The software can use this loopback information to determine whether a MAU is connected to the DTE by checking the status of carrier sense after each packet transmission.

When data is received by the chip while transmitting, a collision condition exists. This will cause the collision oscillator to turn on within 4.5 bit times. The data on the DI pair will remain with the DO pair until DO goes idle. At this point DI will switch to RxIN if it is still active, or DI will go idle if RxIN is idle. After a collision is detected, the collision oscillator will remain on until either DO or RxIN go idle. The exception to this is when DO starts, then RxIN starts, then DO stops, then RxIN stops. In this case the collision oscillator will remain on until RxIN goes idle according to the IEEE FOIRL standard.

Loopback can be disabled by strapping LBDIS to V_{CC}. In this mode the chip operates as a full duplex transmitter and receiver, and collision detection is disabled. A loopback through the transceiver can be accomplished by tying the fiber transmitter to the receiver.

SQE TEST FUNCTION (SIGNAL QUALITY ERROR)

The SQE test function allows the DTE to determine whether the collision detect circuitry is functional. After each transmission, during the inter packet gap time, the collision oscillator will be activated for typically 1 μ s. The SQE test will not be activated if the chip is in the low light state, or the jabber on state.

For SQE to operate, the SQEN pin must be tied to V_{CC}. This allows the MAU to be interfaced to a DTE. The SQE test can be disabled by tying the SQEN pin to ground, for a repeater interface.

JABBER FUNCTION REQUIREMENTS

The Jabber function prevents a babbling transmitter from bringing down the network. Within the transceiver is a Jabber timer that starts at the beginning of each transmission and resets at the end of each transmission. If the transmission last longer than 20ms the jabber logic disables the transmitter, and turns on the collision signal COL+, COL-. When Tx+ and Tx- finally go idle, a second timer measures 0.5 seconds of idle time before the transmitter is enabled and collision is turned off. Even though the transmitter is disabled during jabber, the 1MHz idle signal is still transmitted.

LED DRIVERS

The ML4661 has five LED drivers. The LED driver pins are active low, and the LEDs are normally off. The LEDs are tied to their respective pins through a 500 Ω resistor to 5 Volts.

The $\overline{\text{XMT}}$, $\overline{\text{RCV}}$ and $\overline{\text{CLSN}}$ pins have pulse stretchers on them which enables the LEDs to be visible. When transmission or reception occurs, the LED XMT, RCV or CLSN status pins will activate low for several milliseconds. If another transmit, receive or collision conditions occurs before the timer expires, the LED timer will reset and restart the timing. Therefore rapid events will leave the LEDs continuously on. The JAB and LMON LEDs do not have pulse stretchers on them since their conditions occur long enough for the eye to see.

LOW LIGHT CONDITION

The $\overline{\text{LMON}}$ LED output is used to indicate a low light condition. LMON is activated low when both LMON_{IN} is low and there are transitions on RxIN_{\pm} less than 3 μ s apart. If either one of these conditions do not exist, LMON will go high.

TIMING DIAGRAMS

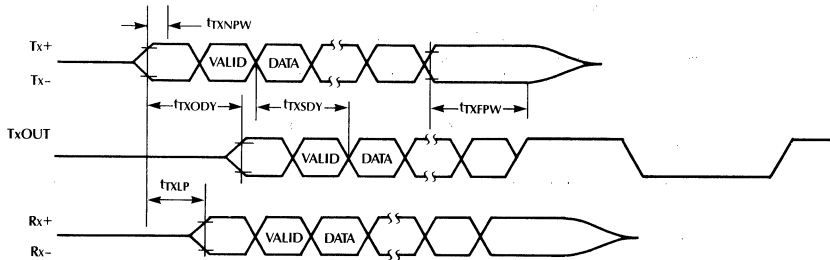


Figure 2. Transmit and Loopback Timing

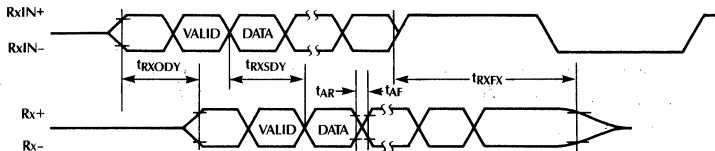


Figure 3. Receive Timing

TIMING DIAGRAMS (Continued)

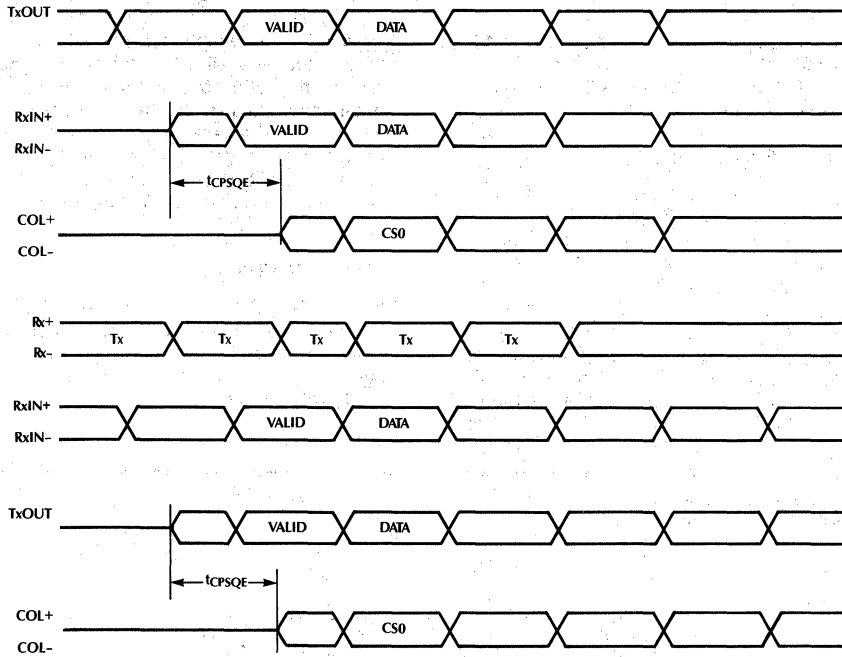
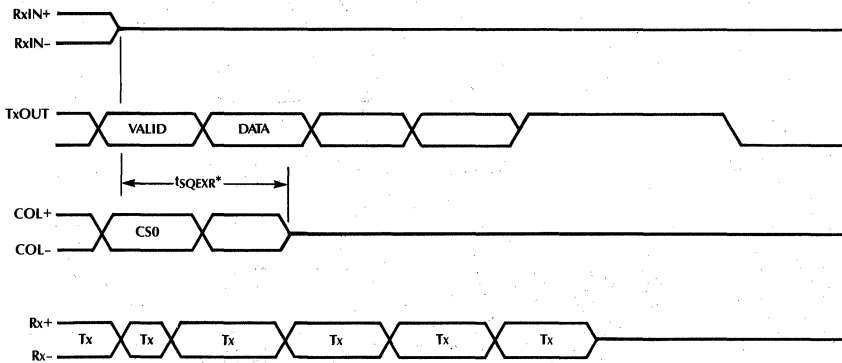


Figure 4. Collision Timing



* ASSUMES THAT TxOUT BEGAN BEFORE RxIN

Figure 5. Collision Timing

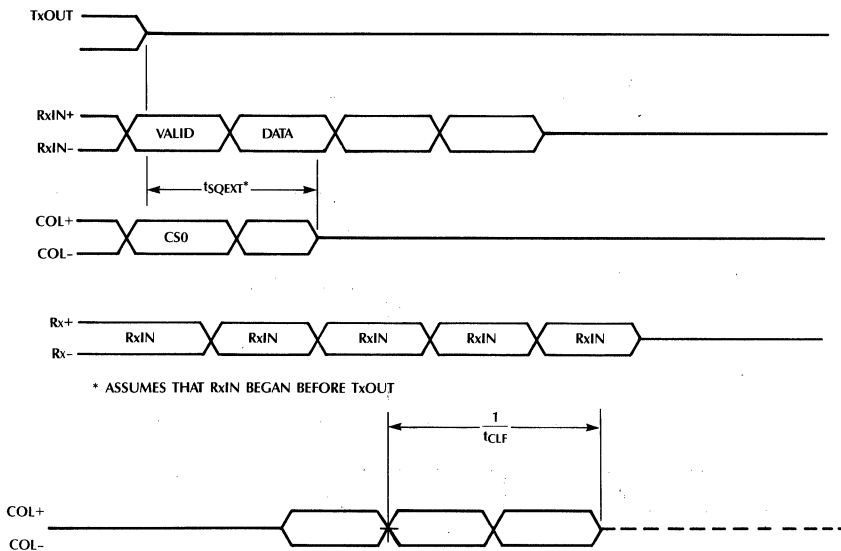


Figure 6. Collision Timing

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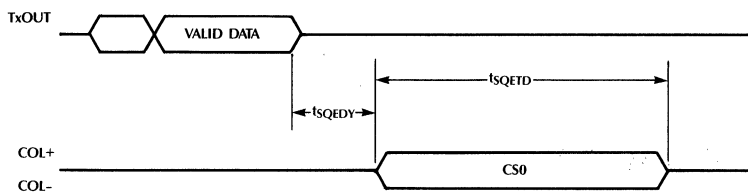


Figure 7. SQE Timing

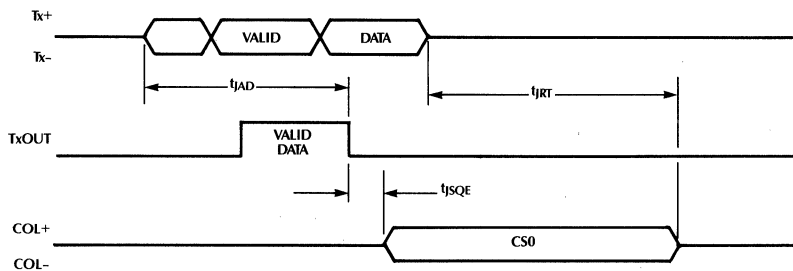


Figure 8. Jabber Timing

ML4661

TIMING DIAGRAMS (Continued)

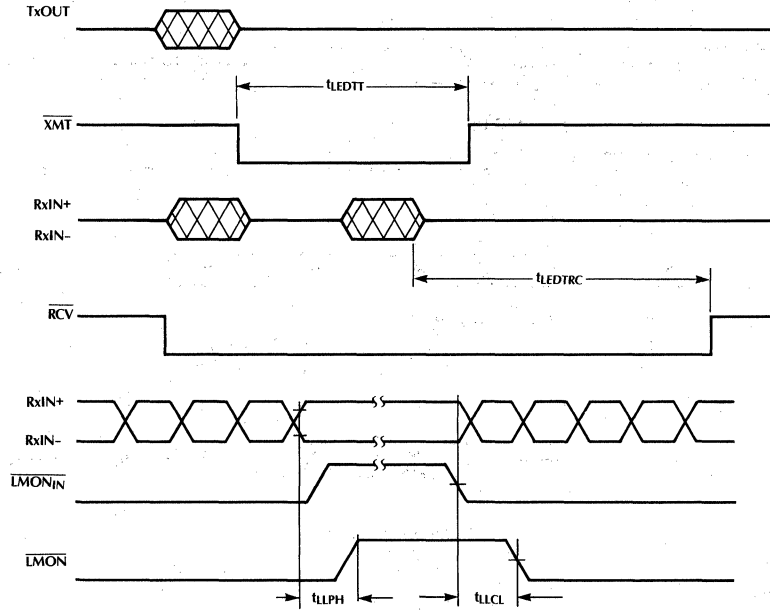


Figure 9. LED Timing

ORDERING INFORMATION

PART NUMBER	PIN COUNT	PACKAGE
ML4661CQ	28 Pins	Molded PCC (Q28)

ML4661EVAL

FOIRL Evaluation Kit

GENERAL DESCRIPTION

The ML4661EVAL is an external MAU designed to evaluate the ML4661/ML4662 FOIRL/10BASE-FL transceiver and the ML4621/ML4624 fiber optic quantizer. The board interfaces to the AUI port through the transformer and to the fiber optic cable through the HP fiber optic transmitter and receiver.

FEATURES

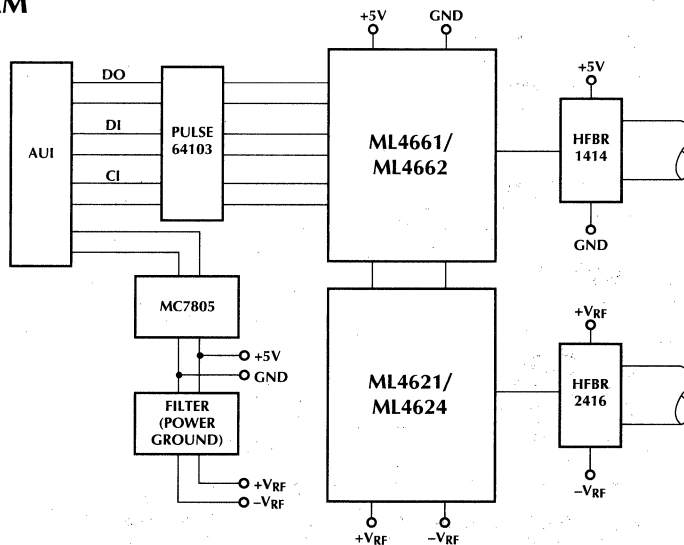
- Jumper switches to enable or disable Loop Back.
- Jumper switches to enable or disable SQE and Jabber functions.
- Capable of adjusting the receive sensitivity.
- 6 status LEDs.
- Current consumption: 260-290mA typically

KIT COMPONENTS

The ML4661EVAL kit includes the following items to help the customer speed up their design, layout and debug process.

- 1) BLANK PCB: 4 layer board with separate power and ground plane (inner layers).
- 2) COMPONENT KIT: Includes the key components as listed below. The rest of the components should be provided by the customer based on the parts list of the ML4661EVAL.
 - HFBR1414: HP fiber Optic LED transmitter.
 - HFBR2416: HP fiber optic pin diode receiver.
 - Two 28 pin sockets for the ML4661 and the ML4621.
 - ML4661CQ: FOIRL transceiver.
 - ML4621CQ: Fiber Optic quantizer.
 - PE64103: PULSE AUI coupling transformer.
 - AUI CONNECTOR: 15 pins D SUB connector.
- 3) DOCUMENTATION: Includes the following items:
 - Demo board description.
 - Block Diagram of the DEMO board.
 - Schematic of the demo board.
 - Lay out of the demo board.
 - Parts list of the ML4661EVAL.

BLOCK DIAGRAM



10BASE-FL Transceiver

GENERAL DESCRIPTION

The ML4662 10Base-FL transceiver combined with the ML4622 or ML4624 fiber optic quantizers provides all of the functionality required to implement both an internal and external IEEE 802.3 10Base-FL MAU. The ML4662 offers a standard IEEE 802.3 AU interface that allows it to be directly connected to industry standard manchester encoder/decoder chips or an AUI cable.

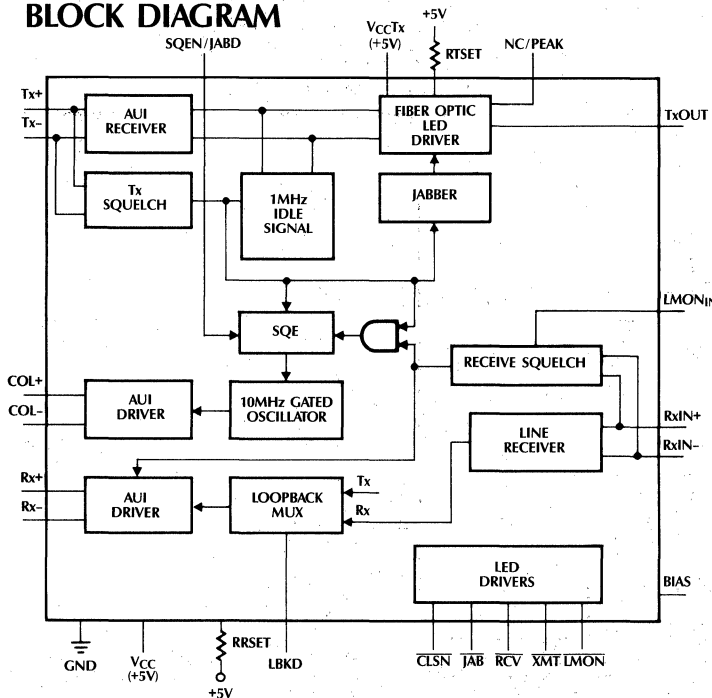
The ML4662 provides a highly integrated solution that requires a minimal number of external components, and is compliant to the IEEE 802.3 10Base-FL standard. The transmitter offers a current driven output that directly drives a fiber optic LED transmitter. Jabber, 1MHz idle signal, and SQE Test are fully integrated onto the chip.

The receiver accepts an ECL level input coming from the ML4622 or ML4624 fiber optic quantizers. The 1MHz idle signal is removed and the AUI output is activated when the receive squelch criteria is exceeded. A Link Monitor function is also provided for low light detection.

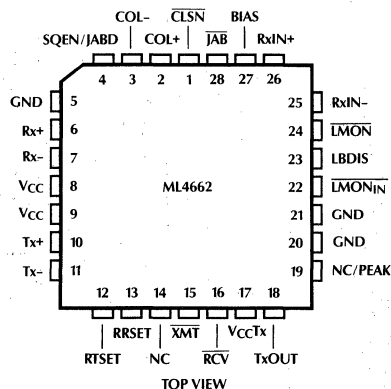
FEATURES

- Combined with the ML4622 or ML4624, offers a complete implementation of an 10Base-FL Medium Attachment Unit (MAU)
- Pin compatible with the ML4661 FOIRL Transceiver
- Incorporates an AU interface for use in an external MAU or an internal MAU
- Single +5 volt supply $\pm 10\%$
- No crystal or clock required
- On-chip Jabber, 1MHz idle, and SQE Test with enable/disable option
- Five network status LED outputs
- Available in a 28-pin PCC package
- Semi-standard option available

BLOCK DIAGRAM



PIN CONNECTION



PIN DESCRIPTION

PIN #	NAME	FUNCTION	PIN #	NAME	FUNCTION
1	CLSN	Indicates that a collision is taking place. Active low LED driver, open collector. Event is extended with internal timer for visibility.	19	NC/PEAK	Normally this pin can be left floating. (tying it to GND or V _{CC} is OK too.) Some fiber optic LEDs may need an additional peaking circuit to speed-up the rise and fall times. For this case, tie pin 19 (NC/PEAK) to pin 18 (TxOUT). When using the HP HFBR 1414, let pin 19 float. Using the peaking circuit may deteriorate optical overshoot and undershoot.
2	COL+	Gated 10MHz oscillation used to indicate a collision, SQE test, or jabber. Balanced differential line driver outputs that meet AUI specifications.	20	GND	Ground Reference.
3	COL-		21	GND	Ground Reference.
4	SQEN/JABD	SQE Test Enable, Jabber Disable. When tied low, SQE test is disabled, when tied high SQE test is enabled. When tied to BIAS both SQE test and Jabber are disabled.	22	LMON _{IN}	Link Monitor Input from the ML4622 or ML4624. This input must be low (active) for the receiver to unsquelch.
5	GND	Ground Reference.	23	LBDIS	Loopback Disable. When this pin is tied to V _{CC} , the AUI transmit pair data is not looped back to the AUI receive pair, and collision is disabled. When this pin is tied to GND (normal operation), the AUI transmit pair data is looped back to the AUI receiver pair.
6	Rx+	Manchester encoded receive data output to the local device. Balanced differential line driver outputs that meet AUI specifications.	24	LMON	Link Monitor LED status output. This pin is pulled low when LMON _{IN} input is low and there are transitions on RxIN _± indicating an idle signal or active data. If either LMON _{IN} goes high or transitions cease on RxIN _± , LMON will go high. Active low LED driver, open collector.
7	Rx-		25	RxIN-	Fiber Optic receive pair. This ECL level signal is received from the ML4622 or ML4624 fiber optic quantizer. When this signal exceeds the receive squelch requirements, and the LMON _{IN} input is low, the receive data is buffered and sent to the AUI receive outputs.
8	V _{CC}	Balanced differential line receiver inputs that meet AUI specifications. These inputs may be transformer, AC or DC coupled. When transformer or AC coupled, the BIAS pin is used to set the common mode voltage.	26	RxIN+	
9	V _{CC}		12	RTSET	Sets the current driven out of the transmitter.
10	Tx+	A 1% 61.9 KΩ resistor tied from this pin to V _{CC} sets the biasing currents for internal nodes.	14	NC	No Connection.
11	Tx-		15	XMT	Indicates that transmission is taking place. Active low LED driver, open collector. Event is extended with internal timer for visibility.
12	RTSET		16	RCV	Indicates that the transceiver is receiving a frame from the optical input. Active low LED driver, open collector. Event is extended with internal timer for visibility.
13	RRSET		17	V _{CC} Tx	+5 volt supply for LED driver.
14	NC		18	TxOUT	Fiber optic LED driver output.
15	XMT				
16	RCV				
17	V _{CC} Tx				
18	TxOUT				
			27	BIAS	BIAS output voltage for the AUI Tx+, Tx- inputs when they are AC coupled.
			28	JAB	Jabber network status LED. When in the Jabber state, this pin will be low and the transmitter will be disabled. In the Jabber "OK" state this pin will be high. Open collector TTL output.

ML4662

ABSOLUTE MAXIMUM RATINGS

(Note 1)

Power Supply Voltage Range	
V_{CC}	-0.3 to 6V
Input Voltage Range	
Digital Inputs (SQEN, LMON _{IN} , LBDIS)	-0.3 to V_{CC}
Tx+, Tx-, RxIN+, RxIN-	-0.3 to V_{CC}
Input Current	
RRSET, RTSET, JAB, CLSN, XMT, RCV, LMON	60mA
Output Current	
TxOUT	70mA
Storage Temperature	-65° C to +150°C
Lead Temperature (Soldering 10 seconds)	260°C

OPERATING CONDITIONS

(Note 2)

Supply Voltage (V_{CC})	5V ± 10%
LED on Current	10mA
RRSET	61.9KΩ ± 1%
RTSET	162Ω ± 1%

ELECTRICAL CHARACTERISTICS

Unless otherwise specified, $T_A = 0^\circ\text{C}$ to 70°C , $V_{CC} = 5\text{V} \pm 10\%$ (Note 3)

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
Power Supply Current I_{CC} While Transmitting	$V_{CC} = 5\text{V}$, RTSET = 162Ω (Note 4)			200	mA
LED Drivers: V_{OL}	$I_{OL} = 10\text{mA}$ (Note 5)			0.8	V
Transmit Peak Output Current (Note 6)	RTSET = 162Ω, $V_{CC} = V_{CC}Tx = 5\text{V} \pm 5\%$	47	52	57	mA
Transmit Squelch Voltage Level (Tx+, Tx-)		-300	-250	-200	mV
Common Mode Input Voltage (Tx±, RxIN±)		2		$V_{CC} - 0.5$	V
Differential Output Voltage (Rx±, COL±)		±550		±1200	mV
Common Mode Output Voltage (Rx±, COL±)			4.0		V
Differential Output Voltage Imbalance (Rx±, COL±)				±40	mV
BIAS Voltage			3.2		V
SQE/JABD	SQE Test Disable Jabber Disable Both Enabled	1.5 $V_{CC} - 0.5$.3 $V_{CC} - 2$	V
LBDIS Threshold	Disabled Enabled	$V_{CC} - 0.10$		1	V

AC ELECTRICAL CHARACTERISTICS

SYMBOL	PARAMETER	MIN	TYP	MAX	UNITS
Transmit					
t _{TXNPW}	Transmit Turn-On Pulse Width		20		ns
t _{TXFPW}	Transmit Turn-Off Pulse Width from Data to Idle	400		2100	ns
t _{TXLP}	Transmit Loopback Startup Delay			500	ns
t _{TXODY}	Transmitter Turn-On Delay			100	ns
t _{TXIDF}	Transmit Idle Frequency	0.85		1.25	MHz
t _{TXDC}	Transmit Idle Duty Cycle	45		55	%
t _{TXSDY}	Transmit Steady State Propagation Delay		15	50	ns
t _{TXJ}	Transmitter Jitter into 31Ω Load			±1.5	ns
Receive					
t _{RXSFT}	Receive Squelch Frequency Threshold	2.51		4.5	MHz
t _{RXODY}	Receive Turn-On Delay			250	ns
t _{RXFX}	Last Bit Received to Slow Decay Output	230	300		ns
t _{RXSDY}	Receive Steady State Propagation Delay		15	50	ns
t _{RXJ}	Receiver Jitter			±1.5	ns
t _{AR}	Differential Output Rise Time 20% to 80% (Rx±, COL±)		4		ns
t _{AF}	Differential Output Fall Time 20% to 80% (Rx±, COL±)		4		ns
Collision					
t _{CPSQE}	Collision Present to SQE Assert	0		350	ns
t _{SQEXR}	Time for SQE to Deactivate After Collision	0		700	ns
t _{CLF}	Collision Frequency	8.5		11.5	MHz
t _{CLPDC}	Collision Pulse Duty Cycle	40	50	60	%
t _{SQEDY}	SQE Test Delay (Tx Inactive to SQE)	0.6		1.6	μs
t _{SQETD}	SQE Test Duration	0.5	1.0	1.5	μs
Jabber and LED Timing					
t _{JAD}	Jabber Activation Delay	20	70	150	ms
t _{JRT}	Jabber Reset Unjab Time	250	450	750	ms
t _{JSQE}	Delay from Outputs Disabled to Collision Oscillator On		100		ns
t _{LED}	R _{CV} , CL _{SN} , X _{MIT} On Time	8	16	32	ms
t _{LLPH}	Low Light Present to L _{MON} High	3	5	10	μs
t _{LLCL}	Low Light Clear to L _{MON} Low	250		750	ms

Note 1: Absolute maximum ratings are limits beyond which the life of the integrated circuit may be impaired. All voltages unless otherwise specified are measured with respect to ground.

Note 2: Limits are guaranteed by 100% testing, sampling, or correlation with worst-case test conditions.

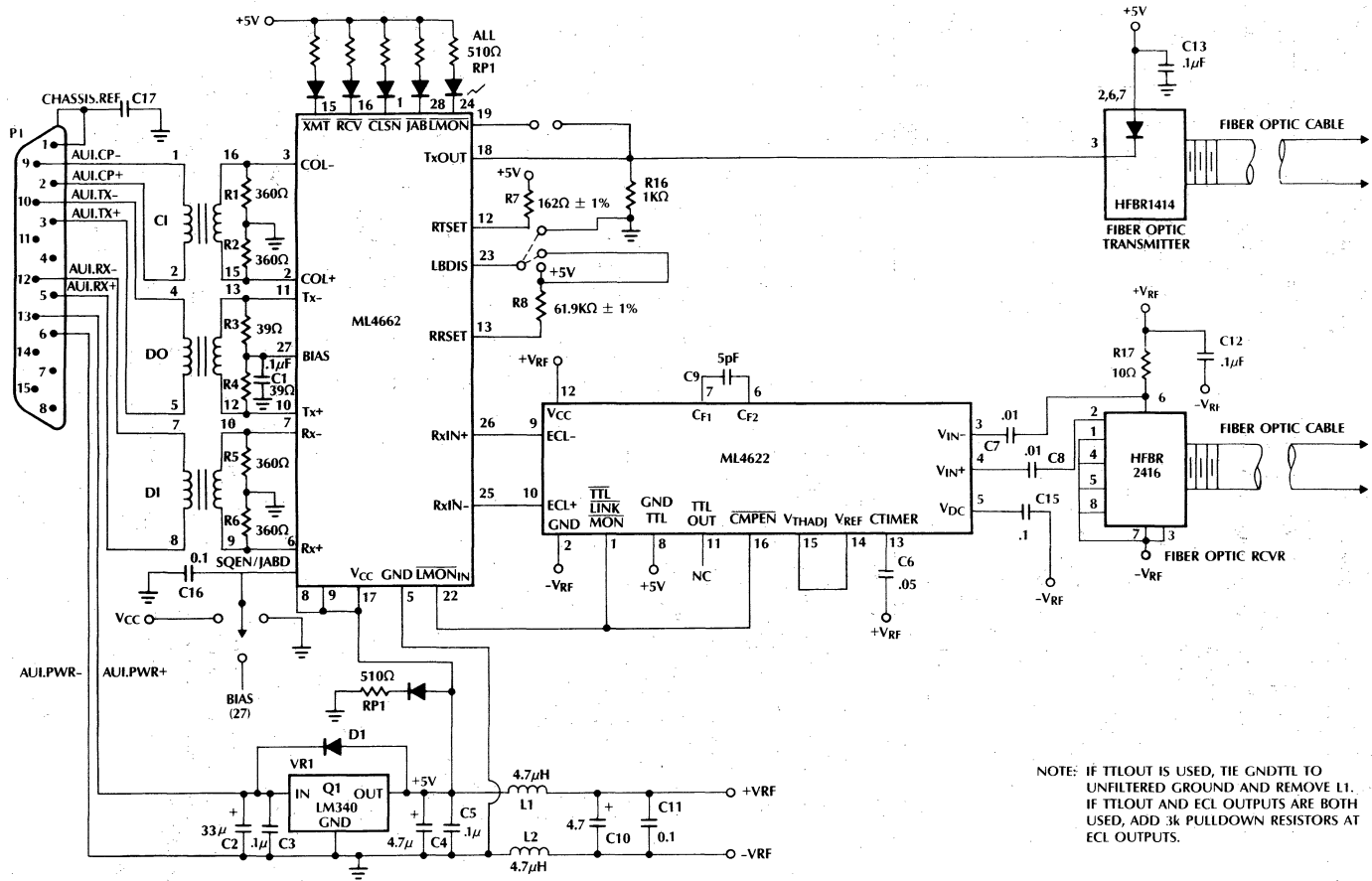
Note 3: Low Duty Cycle pulse testing is performed at T_A.

Note 4: This does not include the current from the AUI pull down resistors, or LED status outputs.

Note 5: LED drivers can sink up to 20mA, but V_{OL} will be higher.

Note 6: Does not include prebias current for fiber optic LED which would typically be 3mA.

Figure 1. ML4662 Schematic Diagram



SYSTEM DESCRIPTION

Figure 1 shows a schematic diagram of the ML4662 in an internal or external 10Base-FL MAU. On one side of the transceiver is the AU interface and the other is the fiber optic interface. The AU interface is AC coupled when used in an external transceiver or can be AC or DC coupled when used in an internal transceiver. The AU interface for an external transceiver includes isolation transformers, some biasing resistors, and a voltage regulator for power.

The fiber optic side of the transceiver requires an external fiber optic transmitter, fiber optic receiver, and the ML4622 or ML4624 fiber optic quantizers. The transmitter uses a current driven output that directly drives the fiber optic transmitter. The receive side of the transceiver accepts the data after passing through the fiber optic receiver and the ML4622/ML4624 fiber optic quantizer.

AU INTERFACE

The AUI interface consist of 3 pair of signals, DO, CI and DI as shown in figure 1. The DO pair contains transmit data from the DTE which is received by the transceiver and sent out onto the fiber optic cable. The DI pair contains valid data that has been either received from the fiber optic cable or looped back from the DO and output through the DI pair to the DTE. The CI pair indicates whether a collision has occurred. It is an output that oscillates at 10MHz if a collision Jabber or SQE Test has taken place, otherwise it remains idle.

When the transceiver is external, these three pair are AC coupled through isolation transformers, while an internal transceiver may be AC or DC coupled. For the AC coupled interface, DO which is an input must be DC biased (shifted up in voltage) for the proper common mode input voltage. The BIAS pin serves this purpose. When DC coupled, transmit output pair coming from the serial interface provides this common mode voltage and the BIAS pin is not connected.

The two 39Ω 1% resistors tied to the Tx+ and Tx- pins serve two purposes. First they provide a point to connect the common mode bias voltage as discussed above, and they provide the proper matching termination for the AUI cable. The CI and DI pair, which are output from the transceiver to the AUI cable, require 360Ω pull down resistors when terminated with a 78Ω load. However on a DTE card, CI and DI do not need 78Ω terminating resistors. This also means that the pull down resistors on CI and DI can be 1KΩ or greater depending upon the particular manchester encoder/decoder chip used. Using higher value pull down resistors as in a DTE card will save power.

The AUI drivers are capable of driving the full 50 meters of cable length and have a rise and fall time of typically 4ns. In the idle state, the outputs go to the same voltage to prevent DC standing current in the isolation transformers.

TRANSMISSION

The transmit function consists of detecting the presence of data from the AUI DO input (Tx+, Tx-) and driving that data onto the fiber optic LED transmitter. A positive signal on the Tx+ lead relative to the Tx- lead of the DO circuit will result in no current, hence the fiber optic LED is in a low light condition. When Tx+ is more negative than Tx-, the ML4662 will sink current into the chip and the LED will light up.

Before data will be transmitted onto the fiber optic cable from the AUI interface, it must exceed the squelch requirements for the DO pair. The Tx squelch circuit serves the function of preventing any noise from being transmitted onto the fiber. This circuit rejects signals with pulse widths less than typically 20ns (negative going), or with levels less than -250mV. Once Tx squelch circuit has unsquelched, it looks for the start of idle signal to turn on the squelch circuit again. The transmitter turns on the squelch again when it receives an input signal at TxIN± that is more positive than -250mV for more than approximately 180ns.

At the start of a packet transmission, no more than 1 bit is received from the DO circuit and not transmitted onto the fiber optic cable. The difference between start-up delays (bit loss plus steady-state propagation delay) for any two packets that are separated by 9.6μs or less will not exceed 200ns.

FIBER OPTIC LED DRIVER

The output stage of the transmitter is a current mode switch which develops the output light by sinking current through the LED into the TxOUT pin. Once the current requirement for the LED is determined, the RTSET resistor is selected. The following equation is used to select the correct RTSET resistor:

$$RTSET = \left(\frac{55\text{mA}}{I_{OUT}} \right) 162\Omega$$

The transmitter enters the idle state when it detects start of idle on Tx+ and Tx- input pins. After detecting the start of idle the transmitter switches to a 1MHz output idle signal.

The output current is switched through the TxOUT pin during the on cycle and the V_{CC}Tx pin during the off cycle as shown in figure 2. Since the sum of the current in these two pins is constant, V_{CC}Tx should be connected as close as possible to the V_{CC} connection for the LED.

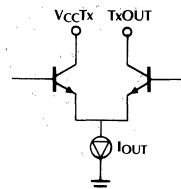


Figure 2. Fiber Optic LED Driver Structure.

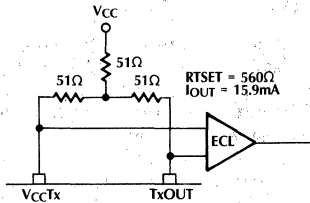


Figure 3. Converting Optical LED Driver Output to Differential ECL.

If not driving an optical LED directly, a differential output can be generated by tying resistors from V_{CCTx} and $TxOUT$ to V_{CC} as shown in figure 3. The minimum voltage on these two pins should not be less than $V_{CC} - 2V$.

RECEPTION

The input to the transceiver comes from the ECL outputs of the ML4622 or ML4624. At this point it is a clean digital ECL signal. At the start of packet reception no more than 2.5 bits are received from the fiber cable and not transmitted onto the DI circuit. The receive squelch will reject frequencies lower than 2.51MHz and any receive input if the $LMON_{IN}$ pin is high.

While in the unsquelch state, the receive squelch circuit looks for the start of idle signal at the end of the packet. Start of idle occurs when the input signal remains idle for more than 160ns. When start of idle is detected, the receive squelch circuit returns to the squelch state and the start of idle signal is output on the DI circuit ($Rx+$, $Rx-$).

COLLISION

Whenever the receiver and the transmitter are active at the same time the chip will activate the collision output, except when loopback is disabled ($LBDIS = V_{CC}$). The collision output is a differential square wave matching the AUI specifications and capable of driving a 78Ω load. The frequency of the square wave is $10MHz \pm 15\%$ with a 60/40 to 40/60 duty cycle. The collision oscillator also is activated during SQE Test and Jabber.

LOOPBACK

The loopback function emulates a 10Base-T transceiver whereby the transmit data sent by the DTE is looped back over the AUI receive pair. Some LAN controllers use this loopback information to determine whether a MAU is connected by monitoring the carrier sense while transmitting. The software can use this loopback information to determine whether a MAU is connected to the DTE by checking the status of carrier sense after each packet transmission.

When data is received by the chip while transmitting, a collision condition exists. This will cause the collision oscillator to turn on and the data on the DI pair will follow $RxIN_{\pm}$. After a collision is detected, the collision oscillator will remain on until either DO or $RxIN$ go idle.

Loopback can be disabled by strapping $LBDIS$ to V_{CC} . In this mode the chip operates as a full duplex transmitter and receiver, and collision detection is disabled. A loopback through the transceiver can be accomplished by tying the fiber transmitter to the receiver.

SQE TEST FUNCTION (SIGNAL QUALITY ERROR)

The SQE test function allows the DTE to determine whether the collision detect circuitry is functional. After each transmission, during the inter packet gap time, the collision oscillator will be activated for typically 1μs. The SQE test will not be activated if the chip is in the low light state, or the jabber on state.

For SQE to operate, the $SQEN$ pin must be tied to V_{CC} . This allows the MAU to be interfaced to a DTE. The SQE test can be disabled by tying the $SQEN$ pin to ground, for a repeater interface.

JABBER FUNCTION REQUIREMENTS

The Jabber function prevents a babbling transmitter from bringing down the network. Within the transceiver is a Jabber timer that starts at the beginning of each transmission and resets at the end of each transmission. If the transmission last longer than 20ms the jabber logic disables the transmitter, and turns on the collision signal $COL+$, $COL-$. When $Tx+$ and $Tx-$ finally go idle, a second timer measures 0.5 seconds of idle time before the transmitter is enabled and collision is turned off. Even though the transmitter is disabled during jabber, the 1MHz idle signal is still transmitted.

LED DRIVERS

The ML4662 has five LED drivers. The LED driver pins are active low, and the LEDs are normally off. The LEDs are tied to their respective pins through a 500Ω resistor to 5 Volts.

The \overline{XMT} , \overline{RCV} and \overline{CLSN} pins have pulse stretchers on them which enables the LEDs to be visible. When transmission or reception occurs, the LED \overline{XMT} , \overline{RCV} or \overline{CLSN} status pins will activate low for several milliseconds. If another transmit, receive or collision conditions occurs before the timer expires, the LED timer will reset and restart the timing. Therefore rapid events will leave the LEDs continuously on. The \overline{JAB} and \overline{LMON} LEDs do not have pulse stretchers on them since their conditions occur long enough for the eye to see.

LOW LIGHT CONDITION

The LMON LED output is used to indicate a low light condition. LMON is activated low when both LMON_{IN} is low and there are transitions on RxIN± less than 3μs apart. If either one of these conditions do not exist, LMON will go high.

DIFFERENCES BETWEEN 10BASE-FL AND FOIRL

10Base-FL is an improved version of the original FOIRL standard. The 10Base-FL standard allows backward compatibility of a 10Base-FL transceiver with a FOIRL transceiver. The main improvements incorporated into 10Base-FL are that it can attach to a DTE by adding the SQE test, and the distance has been increased from 1Km to 2Km. The other differences are much more subtle.

1. SQE Test: The FOIRL standard did not include the option of attaching a fiber transceiver to a DTE. Adding the SQE test to 10Base-FL enables a 10Base-FL transceiver to attach to a DTE. Micro Linear's ML4661 FOIRL transceiver has a SQE test, but this is beyond the scope of the FOIRL standard.

2. 0 to at Least 2Km Distance: The FOIRL standard specifies a 1Km distance while 10Base-FL specifies 2Km. The additional 1Km distance for 10Base-FL comes from an increased flux budget for the cable of 3.5dB. This 3.5dB increase came from an increase of 2.5dB sensitivity for the receiver and a 1dB improvement for the transmitter. The following table illustrates the transmit and receive power requirements for the two standards. Note: FOIRL specifies optical power in peak and 10Base-FL specifies it in average. Subtracting 3dB from peak will give the average. In the table below the FOIRL specifications were converted from peak to average power.

Transmit/Receive Average Power	MIN	MAX	Conditions
FOIRL			
Transmitter	-12dBm	-21dBm	
Receiver	-12dBm	-30dBm	BER < 10 ⁻¹⁰
10Base-FL			
Transmitter	-12dBm	-20dBm	
Receiver	-12dBm	-32.5dBm	BER < 10 ⁻⁹

3. MAU State Diagrams are Different: The state diagrams for 10Base-FL are similar to 10Base-T, while the state diagrams for FOIRL are slightly different. The differences are in the AUI loopback, and in the link integrity function.

AUI Loopback — In 10Base-FL, the DO to DI loopback is always disabled during a collision, and optical receive data is passed through to DI. For FOIRL there are some cases where loopback continues (i.e. DO looped to DI) during a collision, and others where loopback is disabled during a collision. 10Base-FL is identical to 10Base-T in this case. Please refer to the IEEE standards for greater detail.

Link Integrity — 10Base-FL adds an additional state to the Link Integrity Test function that will not allow an exit from the Low Light State until both the transmitter and receiver are idle. In FOIRL it is possible to exit from the Low Light State while still receiving data.

MAU Timing Differences — The timing differences between 10Base-FL and FOIRL relate to propagation delays, start-up delays, and collision deassert delays. The following table provides the details of these parameters.

Timing Parameter Differences	FOIRL (bit times)	10Base-FL (bit times)
ORD_input to input on DI		
Steady State Prop Delay	0.5	2
Start-up Delay	3.5	5
output on DO to OTD_output		
Steady State Prop Delay	0.5	2
Start-up Delay	3.5	5
Collision Deassert to SQE Deassert minimum	4.5	0

OTD — Optical Transmit Data
 ORD — Optical Receive Data
 DI, DO, CI — AUI Interface Signals

TIMING DIAGRAMS

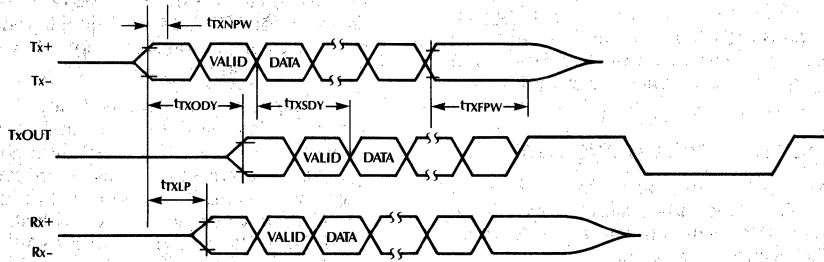


Figure 4. Transmit and Loopback Timing

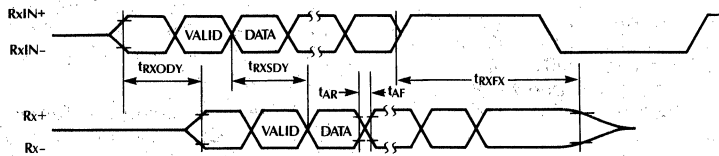


Figure 5. Receive Timing

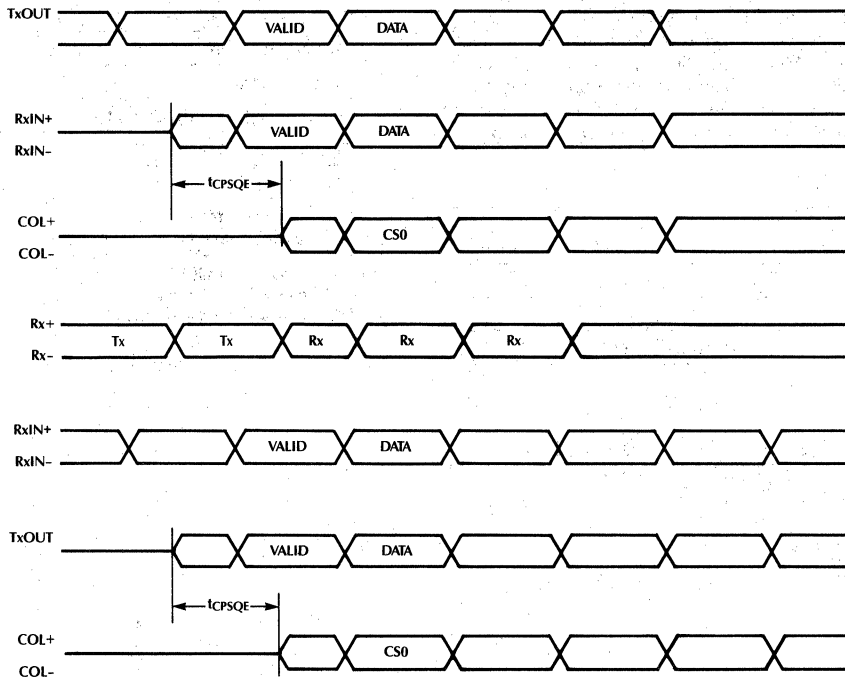


Figure 6. Collision Timing

TIMING DIAGRAMS (Continued)

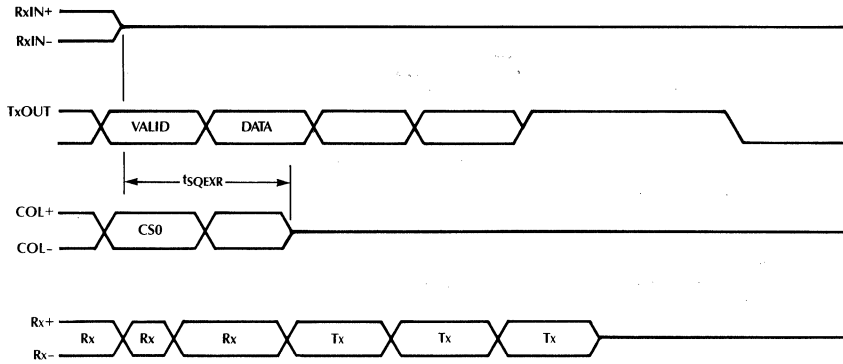


Figure 7. Collision Timing

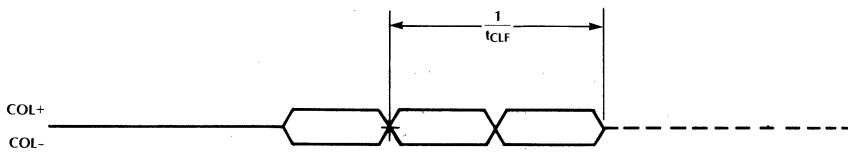
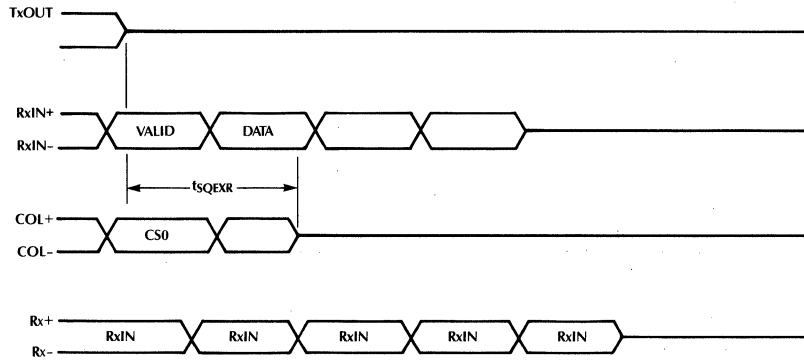


Figure 8. Collision Timing

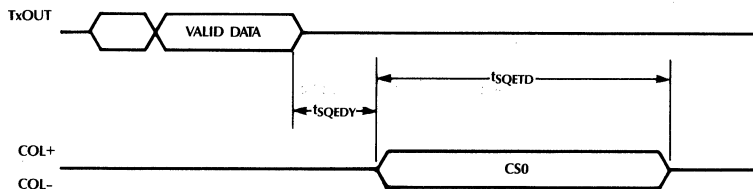


Figure 9. SQE Timing

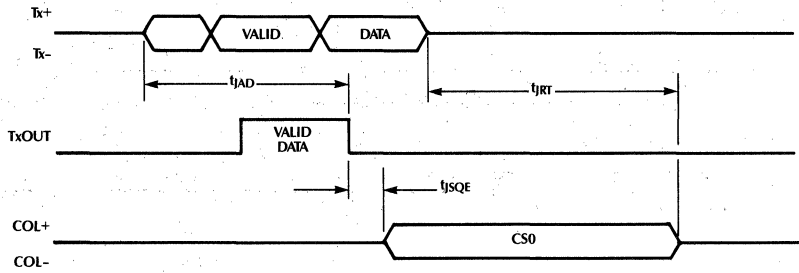


Figure 10. Jabber Timing

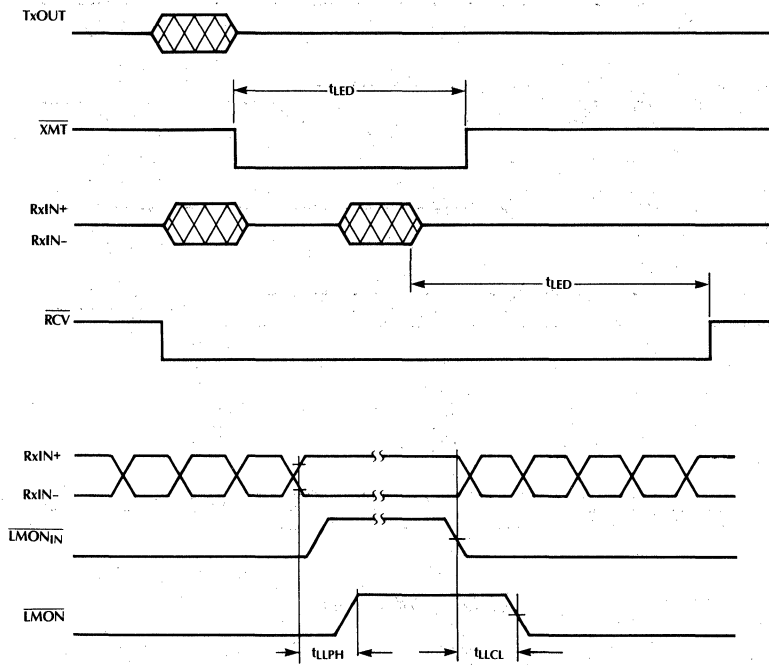


Figure 11. LED Timing

ORDERING INFORMATION

PART NUMBER	PIN COUNT	PACKAGE
ML4662CQ	28 Pins	Molded PCC (Q28)



10BASE-FL Evaluation Kit

GENERAL DESCRIPTION

The ML4662EVAL is an external MAU designed to evaluate the ML4661/ML4662 FOIRL/10BASE-FL transceiver and the ML4622 fiber optic quantizer. The board interfaces to the AUI port through the transformer and to the fiber optic cable through the HP fiber optic transmitter and receiver.

The ML4662EVAL package includes the following items to help the customer speed up their design, layout and debug process.

FEATURES

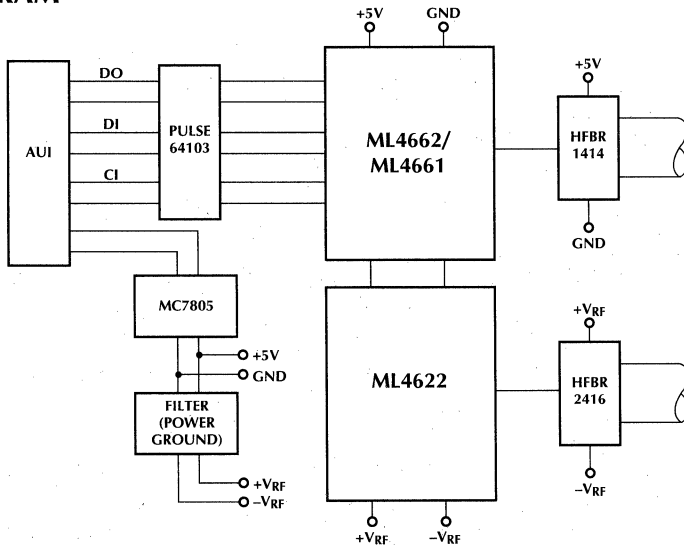
- Jumper switches to enable or disable Loop Back.
- Jumper switches to enable or disable SQE and Jabber functions.
- Capable of adjusting the receive sensitivity.
- 6 status LEDs.
- Current consumption: 260-290mA Typically

KIT COMPONENTS

The ML4661EVAL kit includes the following items to help the customer speed up their design, layout and debug process.

- 1) BLANK PCB: 4 layer board with separate power and ground plane (inner layers).
- 2) COMPONENT KIT: Includes the key components as listed below. The rest of the components should be provided by the customer based on the parts list of the ML4662EVAL.
 - HFBR1414: HP fiber Optic LED transmitter.
 - HFBR2416: HP fiber optic pin diode receiver.
 - One 28 pin sockets for the ML4661.
 - ML4662CQ: 10BASE-FL transceiver.
 - ML4622CP: Fiber Optic quantizer.
 - PE64103: PULSE AUI coupling transformer.
 - AUI CONNECTOR: 15 pins D SUB connector.
- 3) DOCUMENTATION: Includes the following items:
 - Demo board description.
 - Block Diagram of the DEMO board.
 - Schematic of the demo board.
 - Lay out of the demo board.
 - Parts list of the ML4662EVAL.

BLOCK DIAGRAM



Single Chip 10BASE-FL Transceiver

GENERAL DESCRIPTION

The ML4663 Single Chip 10BASE-FL Transceiver integrates both a ML4662 10BASE-FL Transceiver with a ML4622 Fiber Optic Data Quantizer to implement a highly integrated solution for 10BASE-FL transceivers. ML4663 offers a standard IEEE 802.3 AU interface that allows it to be directly connected to industry standard manchester encoder/decoder chips or an AUI connector.

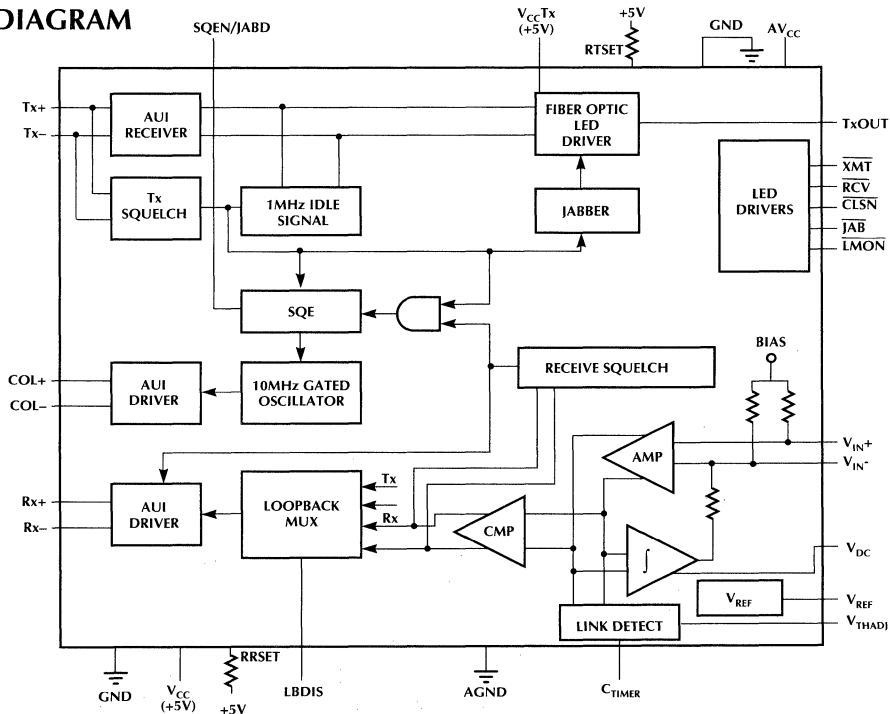
The ML4663 provides a highly integrated solution that requires a minimal number of external components, and is compliant to the IEEE 802.3 10BASE-FL standard. The transmitter offers a current drive output that directly drives a fiber optic LED transmitter. The receiver offers a highly stable fiber optic data quantizer capable of accepting input signals as low as $2mV_{p-p}$ with a 55dB dynamic range.

The transmitter automatically inserts 1MHz signal during idle time and removes this signal on reception. Low Light is continuously monitored for both activity as well as power level. Five LED status indicators monitor error conditions as well as transmissions, receptions and collisions.

FEATURES

- Single chip solution for 10BASE-FL internal or external Medium Attachment Units (MAUs)
- Incorporates an AU interface
- Highly stable data quantizer with 55dB input dynamic range
- Input sensitivity as low as $2mV_{p-p}$
- Current driven fiber optic LED driver for accurate launch power
- Single +5 volt supply
- No crystal or clock required
- Five network status LED outputs
- Available in 28 pin PCC package
- Semi-Standard option available

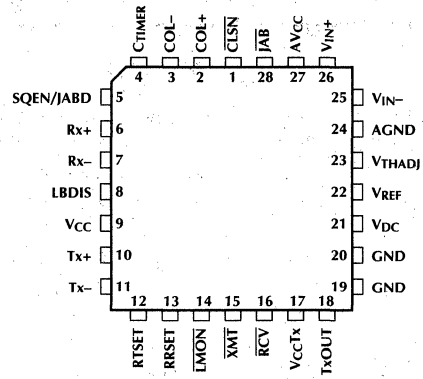
BLOCK DIAGRAM



ML4663

PIN CONNECTION

28-Pin PCC Q-28



PIN DESCRIPTION

PIN	NAME	FUNCTION	PIN	NAME	FUNCTION
1	CLSN	Indicates that a collision is taking place. Active low LED driver, open collector. Event is extended with internal timer for visibility.	15	XMT	Indicates that transmission is taking place. Active low LED driver, open collector. Event is extended with internal timer for visibility.
2	COL+	Gated 10MHz oscillation used to indicate a collision, SQE test, or jabber. Balanced differential line driver outputs that meet AUI specifications.	16	RCV	Indicates that the transceiver is receiving a frame from the optical input. Active low LED driver, open collector. Event is extended with internal timer for visibility.
3	COL-				
4	C _{TIMER}	A capacitor from this pin to V _{CC} determines the Link Monitor response time.	17	V _{CC} Tx	+5 volt supply for fiber optic LED driver.
5	SQEN/JABD	SQE Test Enable, Jabber Disable. When tied low, SQE test is disabled, when tied high SQE test is enabled. When tied to 2.0V both SQE test and Jabber are disabled.	18	TxOUT	Fiber optic LED driver output.
6	Rx+	Manchester encoded receive data output to the local device. Balanced differential line driver outputs that meet AUI specifications.	19	GND	Ground Reference.
7	Rx-		20	GND	Ground Reference.
8	LBDIS	Loopback Disable. When this pin is tied to V _{CC} , the AUI transmit pair data is not looped back to the AUI receive pair, and collision is disabled. When this pin is tied to GND (normal operation) or left floating, the AUI transmit pair data is looped back to the AUI receiver pair, except during collision.	21	V _{DC}	An external capacitor on this pin integrates an error signal which nulls the offset of the input amplifier. If the DC feedback loop is not being used, this pin should be connected to V _{REF} .
9	V _{CC}	+5 volt power input.	22	V _{REF}	A 2.5V reference with respect to GND.
10	Tx+	Balanced differential line receiver inputs that meet AUI specifications. These inputs may be transformer or capacitively coupled. The Tx input pins are internally DC biased for AC coupling.	23	V _{THADJ}	This input pin sets the link monitor threshold.
11	Tx-		24	AGND	Analog Filtered Ground.
12	RTSET	Sets the current driven output of the transmitter.	25	V _{IN-}	This input pin should be capacitively coupled to the input source or to filtered AV _{CC} . (The input resistance is approximately 1.3k Ω .)
13	RRSET	A 1% 61.9k Ω resistor tied from this pin to V _{CC} sets the biasing currents for internal nodes.	26	V _{IN+}	This input pin should be capacitively coupled to the input source or to filtered AV _{CC} . (The input resistance is approximately 1.3k Ω .)
14	LMON	Link Monitor "Low Light" LED status output. This pin is pulled low when the voltage on the V _{IN+} , V _{IN-} inputs exceed the minimum threshold set by the V _{THADJ} pin, and there are transitions on V _{IN+} , V _{IN-} indicating an idle signal or active data. If either the voltage on the V _{IN+} , V _{IN-} inputs fall below the minimum threshold or transitions cease on V _{IN+} , V _{IN-} , LMON will go high. Active low LED driver, open collector.	27	AV _{CC}	Analog Filtered +5 volts.
			28	JAB	Jabber network status LED. When in the Jabber state, this pin will be low and the transmitter will be disabled. In the Jabber "OK" state this pin will be high. Active low LED, open collector.

ML4663

ABSOLUTE MAXIMUM RATINGS

(Note 1)

Power Supply Voltage Range	
V_{CC}	-0.3 to 6V
Input Voltage Range	
Digital Inputs (SQEN, LBDIS)	-0.3 to V_{CC}
$Tx+$, $Tx-$, V_{IN+} , V_{IN-}	-0.3 to V_{CC}
Input Current	
RRSET, RTSET, JAB, CLSN, XMT, RCV, LMON	60mA
Output Current	
$TxOUT$	70mA
Storage Temperature Range	-65°C to +150°C
Lead Temperature (soldering 10 sec)	260°C

OPERATING CONDITIONS

Supply Voltage (V_{CC})	5V ± 5%
LED on Current	10mA
RRSET	61.9kΩ ± 1%
RTSET	162Ω ± 1%

ELECTRICAL CHARACTERISTICS

Unless otherwise specified, $T_A = 0^\circ\text{C}$ to 70°C , $V_{CC} = V_{CCTx} = 5V \pm 5\%$ (Note 2 and 3)

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
Power Supply Current I_{CC} : While Transmitting	$V_{CC} = 5V$, RTSET = 162Ω (Note 4)			220	mA
LED Drivers: V_{CC}	$I_{OL} = 10\text{mA}$ (Note 5)			0.8	V
Transmit Peak Output Current (Note 6)	RTSET = 162Ω	47	52	57	mA
Transmit Squelch Voltage Level ($Tx+$, $Tx-$)		-300	-250	-200	mV
Differential Output Voltage ($Rx\pm$, $COL\pm$)		±550		±1200	mV
Common Mode Output Voltage ($Rx\pm$, $COL\pm$)			4.0		V
Differential Output Voltage Imbalance ($Rx\pm$, $COL\pm$)				±40	mV
SQE/JABD	SQE Test Disable Both Disabled Both Enabled	1.5 $V_{CC} - 0.5$		0.3 $V_{CC} - 2$	V V V
LBDIS Threshold	Disabled Enabled	$V_{CC} - 0.1$		1	V V
Common Mode Voltage ($Tx+$, $Tx-$)			3.5		V
Common Mode Voltage (V_{IN+} , V_{IN-})			1.65		V
Reference Voltage		2.35	2.45	2.55	V
V_{REF} Output Source Current				5	mA
Amplifier Gain			100		V/V
Input Signal Range		2		1600	mV _{P-P}
External Voltage at V_{THADJ} to Set V_{TH}		0.5		2.7	V
Input Offset	$V_{DC} = V_{REF}$ (DC loop active)		3		mV
Input Referred Noise	50MHz BW		25		μV
Input Resistance	V_{IN+} , V_{IN-}	0.8	1.3	2.0	kΩ
Input Bias Current of V_{THADJ}		-200	10	+200	μA
Input Threshold Voltage	$V_{THADJ} = V_{REF}$ (Note 7)	5	6	7	mV _{P-P}
Hysteresis			20		%

AC ELECTRICAL CHARACTERISTICS

SYMBOL	PARAMETER	MIN	TYP	MAX	UNITS
Transmit					
t _{TXNPW}	Transmit Turn-On Pulse Width		20		ns
t _{TXFPW}	Transmit Turn-Off Pulse Width from Data to Idle	400		2100	ns
t _{TXLP}	Transmit loopback Start-up Delay			500	ns
t _{TXODY}	Transmit Turn-On Delay			100	ns
t _{TXIDF}	Transmit Idle Frequency	0.85		1.25	MHz
t _{TXDC}	Transmit Idle duty Cycle	45		55	%
t _{TXSDY}	Transmit Steady State Propagation Delay		15	50	ns
t _{TXJ}	Transmit Jitter into 31Ω Load			±1.5	ns
Receive					
t _{RXSFT}	Receive Squelch Frequency Threshold	2.51		4.5	MHz
t _{RXODY}	Receive Turn-On Delay			270	ns
t _{RXFX}	Last Bit Received to Slow Decay Output	230	300		ns
t _{RXSDY}	Receive Steady State Propagation Delay		15	50	ns
t _{RXJ}	Receive Jitter			±1.5	ns
t _{AR}	Differential Output Rise Time 20% to 80% (R _{x±} , COL _±)		4		ns
t _{AF}	Differential Output Fall Time 20% to 80% (R _{x±} , COL _±)		4		ns
Collision					
t _{CPSQE}	Collision Present to SQE Assert	0		350	ns
t _{SQEXR}	Time for SQE to Deactivate After Collision	0		700	ns
t _{CLF}	Collision Frequency	8.5		11.5	MHz
t _{CLPDC}	Collision Pulse Duty Cycle	40	50	60	%
t _{SQEDY}	SQE Test Delay (Tx Inactive to SQE)	0.6		1.6	μs
t _{SQETD}	SQE Test Duration	0.5	1.0	1.5	μs
Jabber and LED Timing					
t _{JAD}	Jabber Activation Delay	20	70	150	ms
t _{JRT}	Jabber Reset Unjab Time	250	450	750	ms
t _{JSQE}	Delay from Outputs Disabled to Collision Oscillator On		100		ns
t _{LED}	\overline{RCV} , \overline{CLSN} , \overline{XMT} On Time	8	16	32	ms
t _{LLPH}	Low Light Present to \overline{LMON} High	3	5	10	μs
t _{LLCL}	Low Light Present to \overline{LMON} Low	250		750	ms

Note 1: Absolute maximum ratings are limits beyond which the life of the integrated circuit may be impaired. All voltages unless otherwise specified are measured with respect to ground.

Note 2: Limits are guaranteed by 100% testing, sampling, or correlation with worst-case test conditions.

Note 3: Low Duty Cycle pulse testing is performed at T_A.

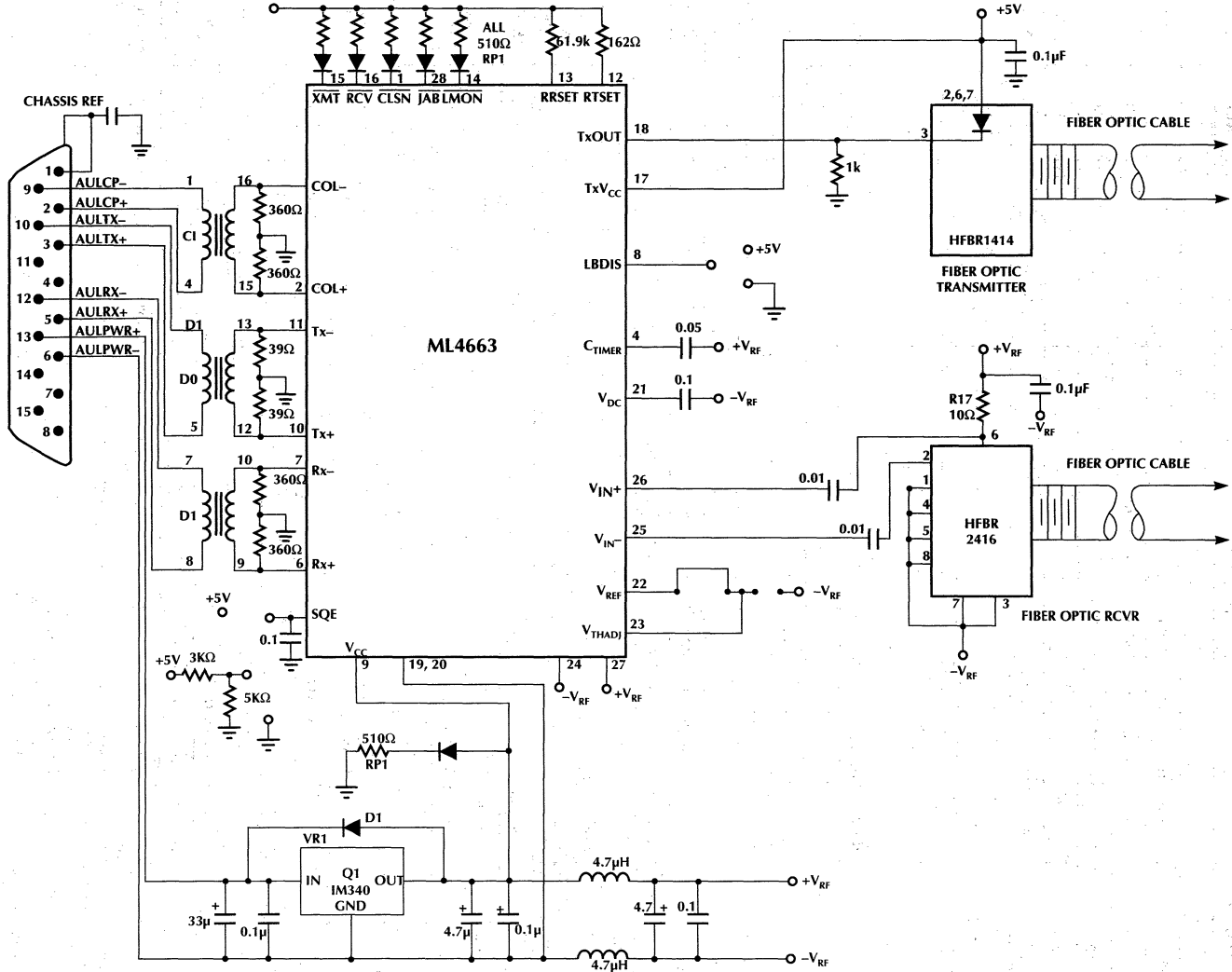
Note 4: This does not include the current from the AUI pull-down resistors, or LED status outputs.

Note 5: LED drivers can sink up to 20mA, but V_{OL} will be higher.

Note 6: Does not include pre-bias current for fiber optic LED which would typically be 3mA.

Note 7: Threshold for switching from Link Fail to Link Pass (Low Light).

Figure 1. ML4663 Schematic Diagram



SYSTEM DESCRIPTION

Figure 1 shows a schematic diagram of the ML4663 in an internal or external 10BASE-FL MAU. On one side of the transceiver is the AU interface and the other is the fiber optic interface. The AU interface is AC coupled when used in an external transceiver or an internal transceiver. The AU interface for an external transceiver includes isolation transformers, some biasing resistors, and a voltage regulator for power.

The fiber optic side of the transceiver requires an external fiber optic transmitter and fiber optic receiver. The transmitter uses a current driven output that directly drives the fiber optic transmitter. The receive side of the transceiver accepts the data after passing through a fiber optic receiver, which consists of a module containing a pin diode and a transimpedance amplifier.

AU INTERFACE

The AU interface consist of 3 pair of signals, DO, CI and DI as shown in figure 1. The DO pair contains transmit data from the DTE which is received by the transceiver and sent out onto the fiber optic cable. The DI pair contains valid data that has been either received from the fiber optic cable or looped back from the DO and output through the DI pair to the DTE. The CI pair indicates whether a collision has occurred. It is an output that oscillates at 10MHz if a collision, Jabber or SQE Test has taken place, otherwise it remains idle.

When the transceiver is external, these three pairs are AC coupled through isolation transformers, while an internal transceiver may be capacitively coupled. Tx+, Tx- is internally DC biased (shifted up in voltage) for the proper common mode input voltage.

The two 39Ω 1% resistors (or one 78Ω 1% resistor) tied to the Tx+ and Tx- pins will provide the proper termination. The CI and DI pair, which are output from the transceiver to the AUI cable, require 360Ω pull down resistors when terminated with a 78Ω load. However on a DTE card, CI and DI do not need 78Ω terminating resistors. This also means that the pull down resistors on CI and DI can be 1kΩ or greater depending upon the particular Manchester encoder/decoder chip used. Using higher value pull down resistors as in a DTE card will save power. Refer to Application Note 13 for a more detailed explanation of the AUI pull-down resistors.

The AUI drivers are capable of driving the full 50 meters of cable length and have a rise and fall time of typically 4ns. In the idle state, the outputs go to the same voltage to prevent DC standing current in the isolation transformers.

TRANSMISSION

The transmit function consists of detecting the presence of data from the AUI DO input (Tx+, Tx-) and driving that data onto the fiber optic LED transmitter. A positive signal on the Tx+ lead relative to the Tx- lead of the DO circuit will result in no current, hence the fiber optic LED is in a low light condition. When Tx+ is more negative than Tx-, the ML4663 will sink current into the chip and the fiber optic LED will light up.

Before data will be transmitted onto the fiber optic cable from the AUI interface, it must exceed the squelch requirements for the DO pair. The Tx squelch circuit serves the function of preventing any noise from being transmitted onto the fiber. This circuit rejects signals with pulse widths less than typically 20ns (negative going), or with levels less than -250mV. Once Tx squelch circuit has unsquelched, it looks for the start of idle signal to turn on the squelch circuit again. The transmitter turns on the squelch again when it receives an input signal at Tx+, Tx- that is more positive than -250mV for more than approximately 180ns.

At the start of a packet transmission, no more than 2 bits are received from the DO circuit and not transmitted onto the fiber optic cable. The difference between start-up delays (bit loss plus steady-state propagation delay) for any two packets that are separated by 9.6μs or less will not exceed 200ns.

FIBER OPTIC LED DRIVER

The output stage of the transmitter is a current mode switch which develops the output light by sinking current through the LED into the TxOUT pin. Once the current requirement for the LED is determined, the RTSET resistor is selected. The following equation is used to select the correct RTSET resistor:

$$RTSET = \left(\frac{52mA}{I_{OUT}} \right) 162\Omega$$

The transmitter enters the idle state when it detects start of idle on Tx+ and Tx- input pins. After detecting the start of idle the transmitter switches to a 1MHz output idle signal.

The output current is switched through the TxOUT pin during the on cycle and the V_{CC}Tx pin during the off cycle as shown in figure 2. Since the sum of the current in these two pins is constant, V_{CC}Tx should be connected as close as possible to the V_{CC} connection for the LED.

If not driving an optical LED directly, a differential output can be generated by tying resistors from V_{CC}Tx and TxOUT to V_{CC} as shown in figure 3. The minimum voltage on these two pins should not be less than V_{CC} - 2V.

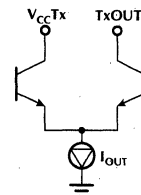


Figure 2. Fiber Optic LED Driver Structure.

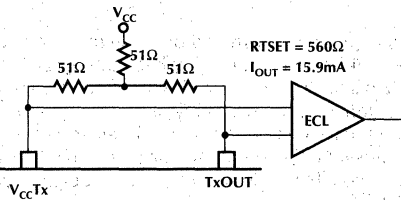


Figure 3. Converting Optical LED Driver Output to Differential ECL.

RECEPTION

The input to the transceiver comes from a fiber optic receiver as shown in figure 1. At the start of packet reception no more than 2.7 bits are received from the fiber cable and not transmitted onto the DI circuit. The receive squelch will reject frequencies lower than 2.51MHz.

While in the unsquelch state, the receive squelch circuit looks for the start of idle signal at the end of the packet. Start of idle occurs when the input signal remains idle for more than 160ns. When start of idle is detected, the receive squelch circuit returns to the squelch state and the start of idle signal is output on the DI circuit (Rx+, Rx-).

COLLISION

Whenever the receiver and the transmitter are active at the same time the chip will activate the collision output, except when loopback is disabled (LBDIS = VCC). The collision output is a differential square wave matching the AUI specifications and capable of driving a 78Ω load. The frequency of the square wave is 10MHz ± 15% with a 60/40 to 40/60 duty cycle. The collision oscillator also is activated during SQE Test and Jabber.

LOOPBACK

The loopback function emulates a 10BASE-T transceiver whereby the transmit data sent by the DTE is looped back over the AUI receive pair. Some LAN controllers use this loopback information to determine whether a MAU is connected by monitoring the carrier sense while transmitting. The software can use this loopback information to determine whether a MAU is connected to the DTE by checking the status of carrier sense after each packet transmission.

When data is received by the chip while transmitting, a collision condition exists. This will cause the collision oscillator to turn on and the data on the DI pair will follow V_{IN+}, V_{IN-}. After a collision is detected, the collision oscillator will remain on until either DO or V_{IN+}, V_{IN-} go idle.

Loopback can be disabled by strapping LBDIS to V_{CC}. In this mode the chip operates as a full duplex transmitter and receiver, and collision detection is disabled. A loopback through the transceiver can be accomplished by tying the fiber transmitter to the receiver.

SQE TEST FUNCTION (SIGNAL QUALITY ERROR)

The SQE test function allows the DTE to determine whether the collision detect circuitry is functional. After each transmission, during the inter packet gap time, the collision oscillator will be activated for typically 1μs. The SQE test will not be activated if the chip is in the low light state, or the jabber on state.

For SQE to operate, the SQEN pin must be tied to V_{CC}. This allows the MAU to be interfaced to a DTE. The SQE test can be disabled by tying the SQEN pin to ground, for a repeater interface.

JABBER FUNCTION REQUIREMENTS

The Jabber function prevents a babbling transmitter from bringing down the network. Within the transceiver is a Jabber timer that starts at the beginning of each transmission and resets at the end of each transmission. If the transmission last longer than 20ms the jabber logic disables the transmitter, and turns on the collision signal COL+, COL-. When Tx+ and Tx- finally go idle, a second timer measures 0.5 seconds of idle time before the transmitter is enabled and collision is turned off. Even though the transmitter is disabled during jabber, the 1MHz idle signal is still transmitted.

LED DRIVERS

The ML4663 has five LED drivers. The LED driver pins are active low, and the LEDs are normally off (except for LMON). The LEDs are tied to their respective pins through a 500Ω resistor to 5 Volts.

The XMT, RCV and CLSN pins have pulse stretchers on them which enables the LEDs to be visible. When transmission or reception occurs, the LED XMT, RCV or CLSN status pins will activate low for several milliseconds. If another transmit, receive or collision conditions occurs before the timer expires, the LED timer will reset and restart the timing. Therefore rapid events will leave the LEDs continuously on. The JAB and LMON LEDs do not have pulse stretchers on them since their conditions occur long enough for the eye to see.

LOW LIGHT CONDITION

The LMON LED output is used to indicate a low light condition. LMON is activated low when both the receive power exceeds the Link Monitor threshold and there are transitions on V_{IN+}, V_{IN-} less than 3μs apart. If either one of these conditions do not exist, LMON will go high.

INPUT AMPLIFIER

The V_{IN+}, V_{IN-} input signal is fed into a limiting amplifier with a gain of about 100 and input resistance of 1.3kΩ. Maximum sensitivity is achieved through the use of a DC restoration feedback loop and AC coupling the input. When AC coupled, the input DC bias voltage is set by an on-chip network at about 1.7V. These coupling capacitors, in conjunction with the input impedance of the amplifier, establish a high pass filter with 3dB corner frequency, f_L, at

$$f_L = \frac{1}{2\pi 1300C} \quad (1)$$

Since the amplifier has a differential input, two capacitors of equal value are required. If the signal driving the input is single ended, one of the coupling capacitors can be tied to AV_{CC} as shown in figure 1.

The internal amplifier has a lowpass filter built-in to band limit the input signal which in turn will improve the signal to noise ratio.

Although the input is AC coupled, the offset voltage *within* the amplifier will be present at the amplifier's output. This is represented by V_{OS} in figure 4. In order to reduce this error a DC feedback loop is incorporated. This negative feedback loop nulls the offset voltage, forcing V_{OS} to be zero. Although the capacitor on V_{DC} is non-critical, the pole it creates can effect the stability of the feedback loop. To avoid stability problems, the value of this capacitor should be at least 10 times larger than the input coupling capacitors.

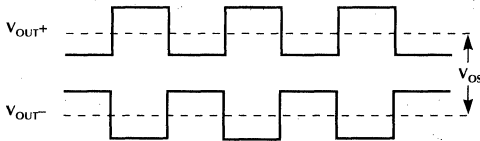


Figure 4.

The comparator is a high-speed differential zero crossing detector that slices and accurately digitizes the receive signal. The output of the comparator is fed in parallel into both the receive squelch circuit and the loopback MUX.

LINK DETECT CIRCUIT AND LOW LIGHT

The link detect circuit monitors the input signal and determines when the input falls below a preset voltage level. When the input falls below a preset voltage level, the ML4663 goes into the Low Light state. In the Low Light state the transmitter is disabled, but continues sending the 1MHz idle signal, the loopback is disabled, the receiver is disabled, and the LMON LED pin goes to high shutting off the LMON LED. To return to the Link Pass state, the optical receiver power must be 20% higher than the shut-off state. This built-in hysteresis adds stability to the Link Monitor circuit. Once the receiver power threshold is exceeded, the ML4663 waits 250ms to 750ms, then checks to see that Tx+ is idle and no data is being received before re-enabling the transmitter, receiver, loopback circuit, and lighting up the LMON LED.

The V_{THADJ} pin is used to adjust the sensitivity of the receiver. The ML4663 is capable of exceeding the 10BASE-FL specifications for sensitivity. The sensitivity is dependent on the layout of the PC board. A good low noise layout will exceed the 10BASE-FL specifications, while a poor layout will fail to meet the sensitivity and BER spec.

The threshold generator shifts the reference voltage at V_{THADJ} through a circuit which has a temperature coefficient matching that of the limiting amplifier. The relationship between the V_{THADJ} and the V_{TH} (the peak to peak input threshold) is:

$$V_{THADJ} = 408V_{TH} \quad (2)$$

In a 10BASE-FL receiver there must be less than 1×10^{-9} bit errors at a receive power level of -32.5dBm average. One procedure to determine the sensitivity of a receiver is to start at the lowest optical power level and gradually increase the optical power until the BER is met. In this case the Link Detect circuit must not disable the receiver (i.e. V_{THADJ} should be tied to Ground). Once the sensitivity of the receiver is determined, V_{THADJ} can be set just above the power level that meets the BER specification. This way the receiver will shut-off before the BER is exceeded.

For 10BASE-FL V_{THADJ} can be tied directly to V_{REF} . However if greater sensitivity is required the circuit in figure 5 can be used to adjust the V_{THADJ} voltage. Even if V_{REF} is tied to V_{THADJ} , it is a good idea to layout a board with these two resistors available. This will allow potential future adjustments without board revisions.

The response time of the Link Detect circuit is set by the C_{TIMER} pin. Starting from the link off state the link can be switched on if the input exceeds the set threshold for a time given by:

$$T = \frac{C_{TIMER} \times 0.7V}{700\mu A} \quad (3)$$

To switch the link from on to off, the above time will be doubled. A value of $0.05\mu F$ will meet to 10BASE-FL specifications.

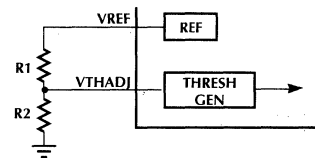


Figure 5.

DIFFERENCES BETWEEN 10BASE-FL AND FOIRL

10BASE-FL is an improved version of the original FOIRL standard. The 10BASE-FL standard allows backward compatibility of a 10BASE-FL transceiver with a FOIRL transceiver. The main improvements incorporated into 10BASE-FL are that it can attach to a DTE by adding the SQE test, and the distance has been increased from 1Km to 2Km. The other differences are much more subtle.

ML4663

1. SQE Test: The FOIRL standard did not include the option of attaching a fiber transceiver to a DTE. Adding the SQE test to 10BASE-FL enables a 10BASE-FL transceiver to attach to a DTE. Micro Linear's ML4661 FOIRL transceiver has a SQE test, but this is beyond the scope of the FOIRL standard.

2. 0 to at Least 2Km Distance: The FOIRL standard specifies a 1Km distance while 10BASE-FL specifies 2Km. The additional 1Km distance for 10BASE-FL comes from an increased flux budget for the cable of 3.5dB. This 3.5dB increase came from an increase of 2.5dB sensitivity for the receiver and a 1dB improvement for the transmitter. The following table illustrates the transmit and receive power requirements for the two standards. Note: FOIRL specifies optical power in peak and 10BASE-FL specifies it in average. Subtracting 3dB from peak will give the average. In the table below the FOIRL specifications were converted from peak to average power.

TRANSMIT/RECEIVE	MIN	MAX	CONDITIONS
FOIRL			
Transmitter	-12dBm	-21dBm	
Receiver	-12dBm	-30dBm	BER < 10 ⁻¹⁰
10BASE-FL			
Transmitter	-12dBm	-20dBm	
Receiver	-12dBm	-32.5dBm	BER < 10 ⁻⁹

3. MAU State Diagrams are Different: The state diagrams for 10BASE-FL are similar to 10BASE-T, while the state diagrams for FOIRL are slightly different. The differences are in the AUI loopback, and in the link integrity function.

AUI Loopback — In 10BASE-FL, the DO to DI loopback is always disabled during a collision, and optical receive data is passed through to DI. For FOIRL there are some cases where loopback continues (i.e. DO looped to DI) during a collision, and others where loopback is disabled during a collision. 10BASE-FL is identical to 10BASE-T in this case. Please refer to the IEEE standards for greater detail.

Link Integrity — 10BASE-FL adds an additional state to the Link Integrity Test function that will not allow an exit from the Low Light State until both the transmitter and receiver are idle. In FOIRL it is possible to exit from the Low Light State while still receiving data.

MAU Timing Differences — The timing differences between 10BASE-FL and FOIRL relate to propagation delays, start-up delays, and collision deassert delays. The following table provides the details of these parameters.

TIMING PARAMETER DIFFERENCES	FOIRL (BIT TIMES)	10BASE-FL (BIT TIMES)
ORD_input to input on DI		
Steady State Prop Delay	0.5	2
Start-Up Delay	3.5	5
output on DO to OTD_output		
Steady State Prop Delay	0.5	2
Start-Up Delay	3.5	5
Collision Deassert to SQE Deassert minimum	4.5	0

OTD — Optical Transmit Data
 ORD — Optical Receive Data
 DI, DO, CI — AUI Interface Signals

TIMING DIAGRAMS

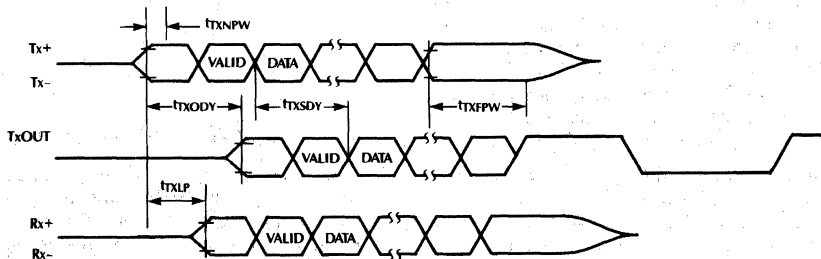


Figure 6. Transmit and Loopback Timing

TIMING DIAGRAMS

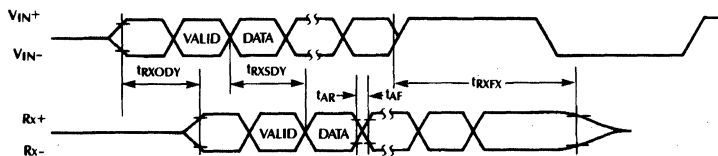


Figure 7. Receive Timing

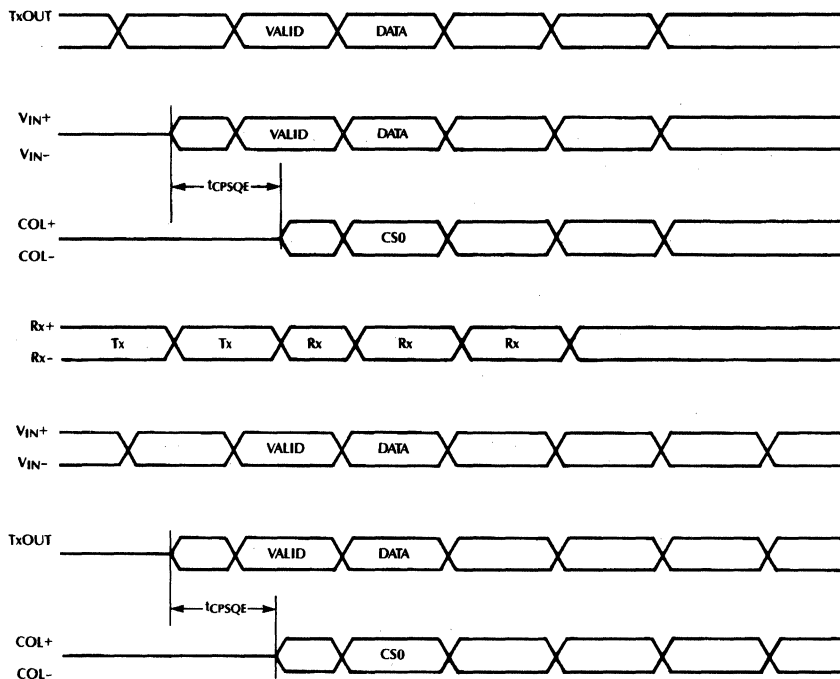


Figure 8. Collision Timing

TIMING DIAGRAMS

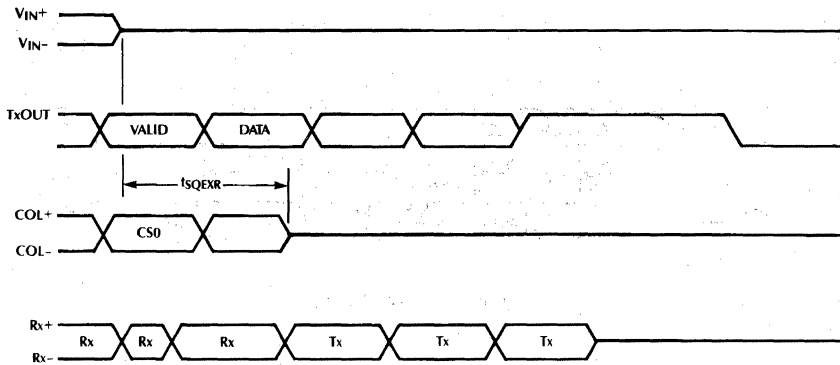


Figure 9. Collision Timing

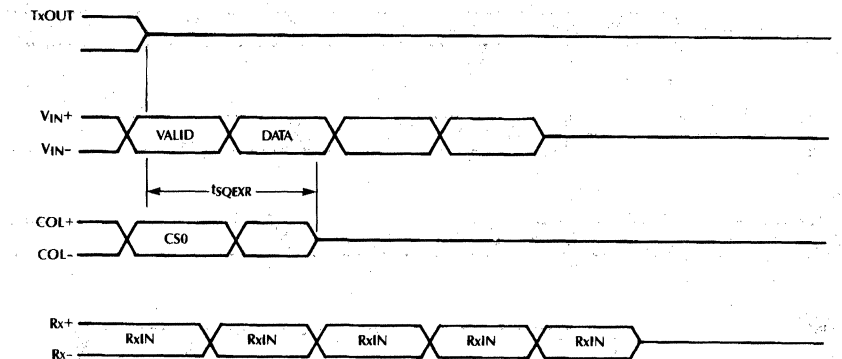


Figure 10. Collision Timing

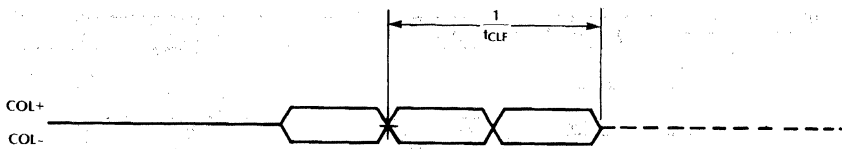


Figure 10. Collision Timing

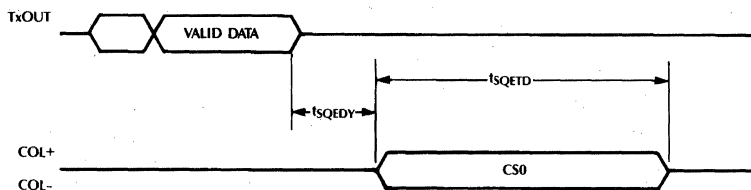


Figure 11. SQE Timing

TIMING DIAGRAMS

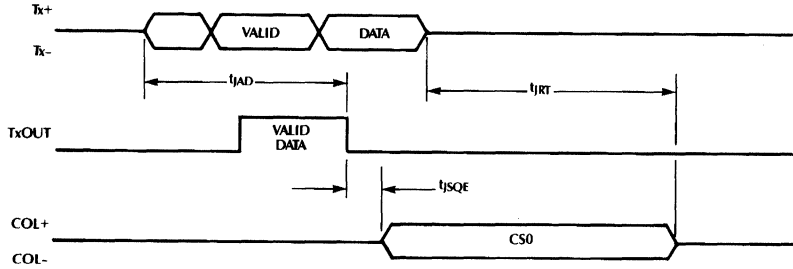


Figure 12. Jabber Timing

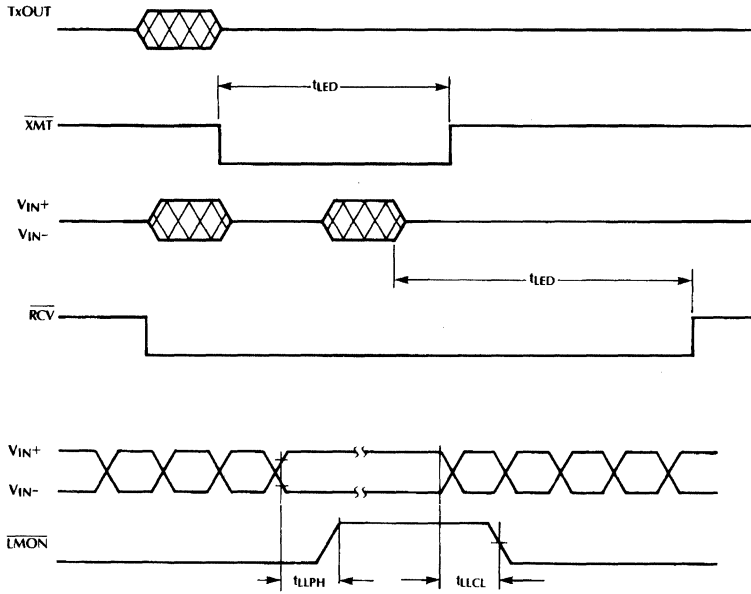


Figure 13. LED Timing

ML4663

ORDERING INFORMATION

PART NUMBER	PIN COUNT	PACKAGE
ML4663CQ	28 Pins	Molded PCC (Q28)

ML4663EVAL

10BASE-FL Evaluation Kit

GENERAL DESCRIPTION

The ML4663EVAL is an external MAU designed to evaluate the ML4663 10BASE-FL PMD chip. The board interfaces to the AUI port through the transformer and to the fiber optic cable through the HP fiber optic transmitter and receiver.

The ML4663EVAL package includes the following items to help the customer speed up their design, layout and debug process.

FEATURES

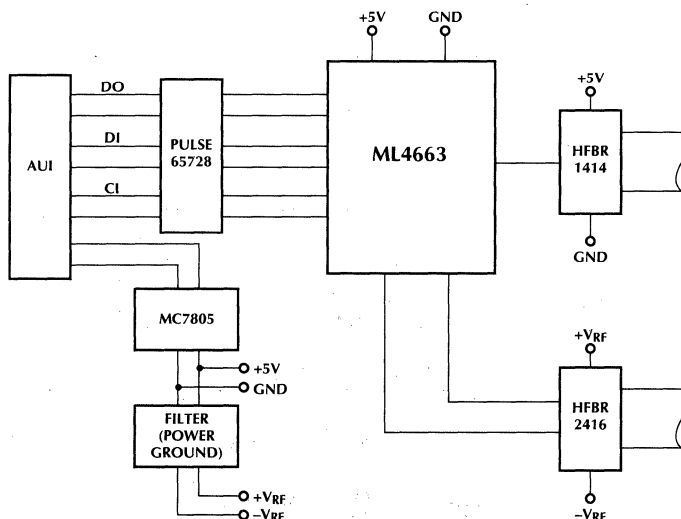
- Jumper switches to enable or disable Loop Back.
- Jumper switches to enable or disable SQE and Jabber functions.
- Capable of adjusting the receive sensitivity.
- 6 status LEDs.
- Current consumption 260-280mA typically

KIT COMPONENTS

The ML4661EVAL kit includes the following items to help the customer speed up their design, layout and debug process.

- 1) BLANK PCB: 4 layer board with separate power and ground plane (inner layers):
- 2) COMPONENT KIT: Includes the key components as listed below. The rest of the components should be provided by the customer based on the parts list of the ML4663EVAL.
 - HFBR1414: HP fiber Optic LED transmitter.
 - HFBR2416: HP fiber optic pin diode receiver.
 - One 28 pin sockets for the ML4663.
 - ML4663CQ: 10BASE-FL combo transceiver and quantizer.
 - PE65728: PULSE AUI coupling transformer.
 - AUI CONNECTOR: 15 pins D SUB connector.
- 3) DOCUMENTATION: includes the following items:
 - Demo board description.
 - Block Diagram of the DEMO board.
 - Schematic of the demo board.
 - Lay out of the demo board.
 - Parts list of the ML4663EVAL.

BLOCK DIAGRAM



High-Speed Data Quantizer

GENERAL DESCRIPTION

The ML6622 High-Speed Data Quantizer is a low noise, wide-band, BiCMOS monolithic IC designed specifically for signal recovery applications in FDDI and SONET fiber-optic receiver systems. An internal DC restoration feedback loop nulls any offset voltage produced in the input stage. The limiting amplifier contributes to a high level of sensitivity and a minimum of duty cycle distortion.

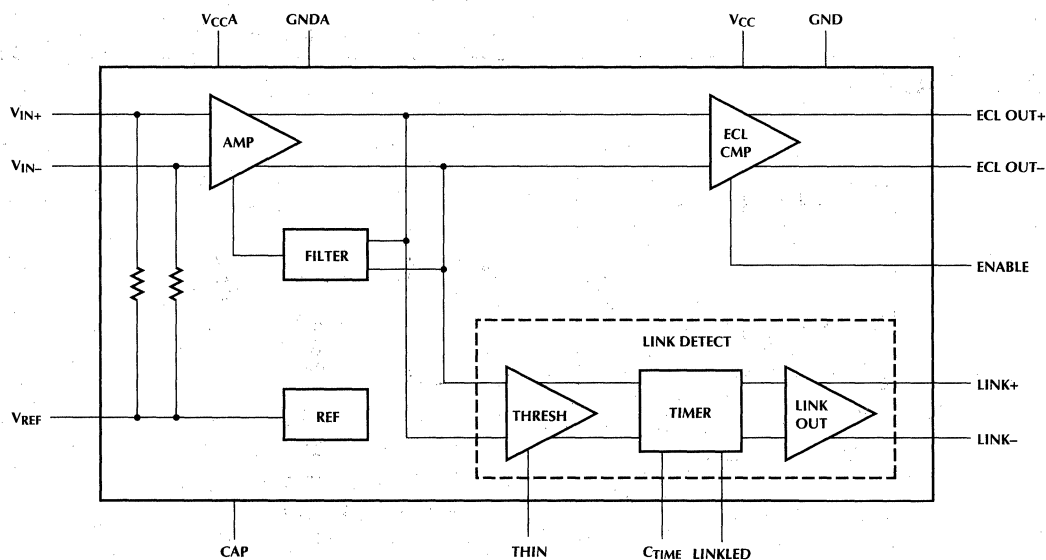
The output of the data path is a high-speed comparator with ECL outputs. An enable pin gates the comparator on or off in response to the input signal level or a system control signal.

The Link Detect circuit provides an Assert-Deassert function with a user-selectable threshold voltage. This circuit monitors the input signal and provides an ECL High output within 100ms of signal acquisition and an ECL Low output within 350ms of signal loss. The ECL discriminator output can be used to disable the comparator when the signal is below the user-selected threshold. LINKLED drives an LED for a visible indication of the link status.

FEATURES

- 150 MHz bandwidth
- Low noise design
- Adjustable Link Detect function
- Low power design: 35mA typical
- Available in 16-pin DIP and 16-pin Skinny SOIC

BLOCK DIAGRAM

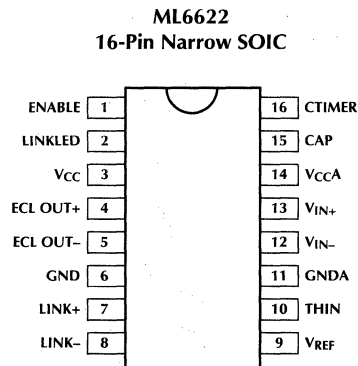
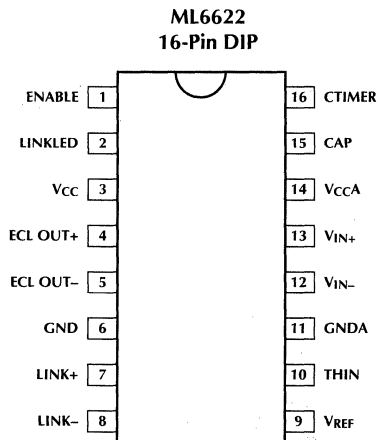


ML6622 High-Speed Data Quantizer

PIN DESCRIPTION

NAME	FUNCTION	NAME	FUNCTION
V _{CC}	Positive Power Supply for less noise sensitive nodes. +5 Volts	V _{REF}	A 2.5V reference with respect to GND. When tied to THIN, the link monitor threshold is set to a convenient value, depending on the responsiveness of the front end receive circuit.
ECL OUT+ ECL OUT-	Positive and Negative ECL Comparator outputs. 1mA internal pull downs are incorporated so that external pull downs aren't necessary for light loads.	V _{IN+}	This input pin should be capacitively coupled to the input source or to V _{CC} .
GND	Ground 0 volts. Used for less noise sensitive nodes.	V _{IN-}	This input pin should be capacitively coupled to the input source or to V _{CC} .
LINKLED	Link Detect Status output. LINKLED is an open collector active low signal. It will be active low when the input signal applied to V _{IN+} , V _{IN-} exceeds the programmed threshold level at the THIN pin. Capable of driving a 20mA LED indicator.	ENABLE	ECL input active low. When this input is tied to LINKLED the ECL Comparator output is automatically enabled and disabled by the Link Detect circuit. This input can be tied to GND for continuous enable. When the ECL Comparator is disabled, ECL OUT- goes low and ECL OUT+ goes high.
LINK -	Negative ECL Link Detect output. This output is active low when the input signal exceeds the programmed Link Detect threshold. 1mA internal pull downs are incorporated.	GNDA	This ground goes to all of the noise sensitive circuits in the chip; the input amplifier, DC restoration loop, part of the Comparator and part of the link detect circuit. In some system designs, it may be advantageous to separate GND and GNDA.
LINK +	Positive ECL Link Detect output. This output is active high when the input signal exceeds the programmed Link Detect threshold. 1mA internal pull down are incorporated.	V _{CCA}	This V _{CC} goes to all of the noise sensitive circuits as mentioned in GNDA.
C _{TIMER}	A capacitor from this pin to ground determines the Link Detect response time. To Meet FDDI specifications this capacitor should be 2,000pF. This capacitor can be removed for the fastest response time.	CAP	A capacitor is tied from this pin to V _{REF} . This capacitor sets the lower frequency rejection and helps remove internal DC offset. This capacitor should be 10 times larger than the input capacitors.
THIN	Threshold Input. A voltage applied to this input pin sets the minimum amplitude of the input signal required to cause the link detect to activate. In most cases this can be tied to V _{REF} .		

PIN CONNECTION



ML6622

ABSOLUTE MAXIMUM RATINGS

(Note 1)

$V_{CC} - GND$ -0.3 to +7.0
 $V_{CCA} - GND_A$ -0.3 to +7.0
 Inputs/Outputs GND -0.3 to $V_{CC} + 0.3$

Storage Temperature Range -65°C to +150°C
 Lead Temperature (Soldering 10 sec.) +260°C

ELECTRICAL CHARACTERISTICS

Over recommended operating conditions of $T_A = 0^\circ\text{C}$ to 70°C , $V_{CC} = 5V \pm 5\%$, $GND = 0V$, unless otherwise noted.

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
I_{CC}	V_{CC} Supply Current (No load on ECL outputs)			35	50	mA
V_{REF}	Reference Voltage		2.37	2.47	2.57	V
$I_{V_{REF}}$	V_{REF} Output Current		-1	3	+5	mA
V_{IN}	Input Signal Range		2		1600	mV _{p-p}
V_{TH} ADJ Range	External Voltage at THIN to set V_{TH}		0.5		V_{REF}	V
EN	Input-referred Voltage Noise	100 MHz BW		25		μV_{RMS}
R_{IN}	Input Resistance	V_{IN+} , V_{IN-}	500	770	1000	Ω
I_{THIN}	Input Bias Current of THIN		-100		+100	μA
$V_{OL-V_{CC}}$	ECL Output Voltage-Low	Through 50Ω to $V_{CC} - 2V$, $T_A = 25^\circ\text{C}$	-1.840	-1.730	-1.620	V
$V_{OH-V_{CC}}$	ECL Output Voltage-High	Through 50Ω to $V_{CC} - 2V$, $T_A = 25^\circ\text{C}$	-1.045	-0.963	-0.880	V
t_r	Data Output Rise Time					ns
t_f	Data Output Fall Time		0.35		1.3	ns

LINK DETECT

AS_Max	Assert Time (off to on)	$C_{TIME} = 2000\text{pF}$	0		100	μs
ANS_Max	Deassert Time (on to off)	$C_{TIME} = 2000\text{pF}$	0		350	μs
V_{TH}	Input threshold Hysteresis	THIN = V_{REF} Assert	8 1.5	10 1.7	12 2	mV dB

Note 1: Absolute maximum ratings are limits beyond which the life of the integrated circuit may be impaired. All voltages unless otherwise specified are measured with respect to ground.

Note 2: Limits are guaranteed by 100% testing, sampling, or correlation with worst-case test conditions.

Note 3: Low Duty Cycle pulse testing is performed at T_A .

FUNCTIONAL DESCRIPTION

The ML6622 High Speed Data Quantizer takes a low level analog signal from a pin diode and transimpedance amp front end and converts it into digital ECL levels for subsequent digital processing. The input signal, coming from a transimpedance amplifier, enters the chip and is immediately amplified by a two-stage video amplifier. At the output of this amplifier the signal takes two parallel paths.

The data path passes the signal into a high speed comparator and outputs raised ECL differential data on the ECL OUT± pins. The Link Detection path monitors the magnitude of the amplified input signal, compares it to a user-settable threshold, and provides the result of the comparison as a raised-ECL differential output on the Link± pins. The timer following the threshold block is used to set the Link Detect output acquire and deacquire time using a capacitor.

AMPLIFIER

The amplifier is a two stage video amplifier with a gain of approximately 55V/V. Maximum sensitivity is achieved through the use of the DC restoration feedback loop and AC coupling the input. The AC coupling input capacitors, in conjunction with the input impedance of the amplifier, establish a high pass filter with the lower 3dB point determined by the input resistance and the input coupling capacitors.

Since the amplifier has a differential input, two AC capacitors of equal value are required. If the signal driving the input is single ended, the other coupling capacitor should be tied to V_{CC}.

Another low-pass filter is created with the CAP capacitor. The lower 3dB point controlled by a capacitor tied from the CAP pin to V_{REF} as shown in the application circuit. For stability CAP should be 10 times larger than the input coupling capacitors.

$$f_L = \frac{1}{2\pi 100C}$$

Although the input is AC coupled, the offset voltage within the amplifier will be present at the amplifier's output. The removal of the dc offset in the amp helps the circuit respond to small input voltages, and reduces duty-cycle distortion. In order to reduce this error, a negative feedback loop nulls the offset voltage. An external capacitor CAP is used to store the offset voltage. This voltage is compared to V_{REF} and a difference current proportional to the result is applied to the negative side of the input stage of the AMP thereby nulling the DC offset.

COMPARATOR

A high speed ECL comparator is used for zero crossing detection. The second stage of the comparator outputs raised ECL levels. The comparator has an Enable input pin which takes an ECL level. This Enable pin is normally driven by LINKLED, which causes the output to be enabled when the link is up and disabled when the link is down. When ENABLE is low the comparator is operational. When ENABLE is high the comparator is disabled causing ECL OUT- to go low and ECL OUT+ to go high. The ENABLE pin can be tied to ground to keep the comparator permanently enabled.

LINK DETECT CIRCUIT

The Link Detection Circuit is used to accurately measure the input amplitude to determine whether it is large enough to reliably recover the input signal. Once the Bit Error Rate (BER) for the ML6622 receive circuit is determined, the link detect threshold can be set so that the Link Detect Circuit will shut off before the error rate exceeds the link requirement.

The Link Detection Circuit consists of three functional blocks; Thresh, Timer, and Link Out. Thresh detects the output of Amp and compares it to a programmable threshold input THIN. As long as the input amplitude is greater than the programmable threshold input, the Link Detect output remains active.

When the peak input drops below THIN, Thresh's output changes state and Timer delays the Link Out state change for a programmable amount of time. When using the default C_{TIME} capacitance of 2000pF, the deassert time is 350µs max and the assert time is 100µs max. These default values conform to the ANSI X3.166-1990 PMD standard for FDDI.

To improve stability, the Link Detect circuit includes 1.7dB of hysteresis.

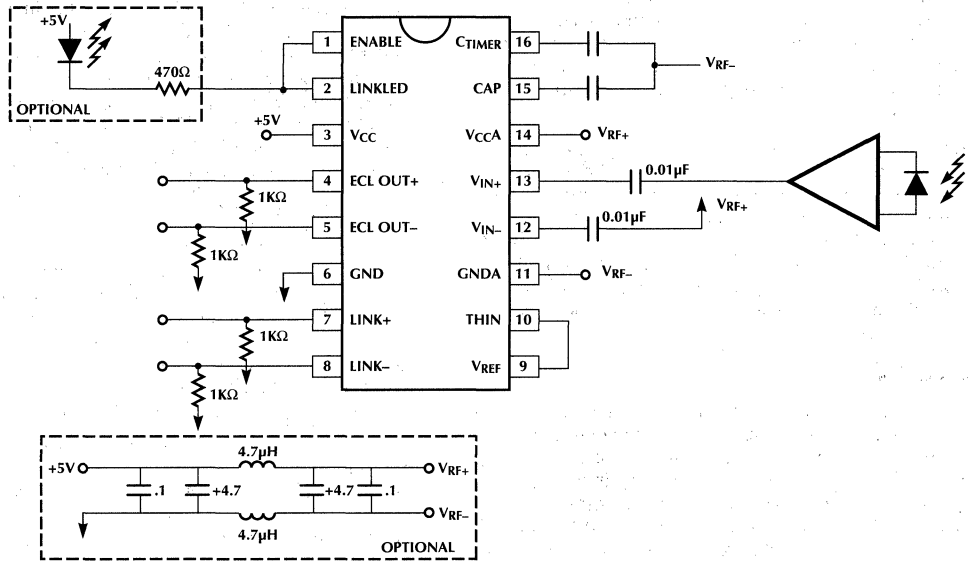
The V_{REF} output can be tied directly to THIN to set the Link Detect threshold. For greater sensitivities, V_{REF} can be divided down before applied to THIN. The formula for the threshold on the thin pin is as follows:

$$\text{Threshold (Assert)} = \frac{V_{THIN}}{500}$$

$$\text{Threshold (Deassert)} = \frac{V_{THIN}}{750}$$

ML6622

APPLICATION CIRCUIT



ORDERING INFORMATION

PART NUMBER	TEMPERATURE RANGE	PACKAGE
ML6622CP	0° to +70°C	16-Pin MOLDED DIP (P16)
ML6622CS	0° to +70°C	16-Pin MOLDED SOIC (S16N)

High Speed Fiber Optic LED Driver

GENERAL DESCRIPTION

The ML6632 is a high speed fiber optic LED driver suited for networking applications up to 200 Mbps. The part is capable of driving up to 120 mA of current through a fiber optic LED from an ECL level input signal. Its efficient output stage provides a high current that can be programmed for accurate absolute output level which insures precise launch power.

The LED driver's output stage provides a fast well matched rise and fall time through a unique class B output stage that burns supply current only when the LED is on. An optional peaking circuit is also available for 820 nm applications.

The ML6632 high speed fiber optic LED driver is implemented in BiCMOS process and is available in a 8-pin SOIC or PDIP package.

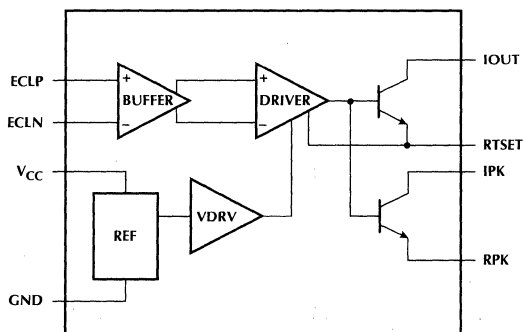
FEATURES

- Data rates up to 200 Mbps
- Current driven output for accurate launch power
- Programmable output current from 20 mA to 120 mA
- High Efficiency Output Stage
- Low EMI/RFI Noise
- ECL inputs
- Optional Peaking Circuit

APPLICATIONS

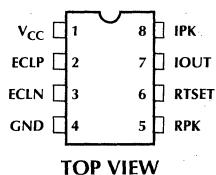
- Fiber Optic Token Ring
- FDDI
- SONET OC1 and OC3
- Fiber Optic Data Communications and Telecommunications

BLOCK DIAGRAM



PIN CONNECTION

8-Pin SOIC or PDIP



ML6632

PIN DESCRIPTION

PIN NO.	NAME	DESCRIPTION	PIN NO.	NAME	DESCRIPTION
1	V _{CC}	Positive power supply. +5 volts.	6	RTSET	Output current programming pin. Connect a resistor of value $1/I_{LED}$ from this pin to ground to set the high LED output current.
2	ECLP	Positive ECL data input controls signal to the LED.	7	IOUT	Fiber optic LED drive pin. Connect the LED between this pin and V _{CC} .
3	ECLN	Negative ECL data input.	8	IPK	Peaking circuit output pin. When using peaking, connect this pin to V _{CC} through a resistor of value R _{RPK} . Then connect a capacitor from this pin to the LED cathode. When peaking is not used, open circuit RPK.
4	GND	Negative power supply. This pin should be tied to the grounded side of RTSET to improve output accuracy and avoid a ground loop.			
5	RPK	Peaking circuit bias pin. Connect a resistor of value $1/I_{PEAK}$ from this pin to ground when using the peaking circuit. Leave open circuited when this pin is not used.			

ABSOLUTE MAXIMUM RATINGS (Note 1)

V _{CC}	-0.3V to 6V	Peak DC Output Current (IPK, IOUT)	120mA
Input Pin Voltages	-0.3V to V _{CC} +0.3V	Storage Temperature	-65°C to +150°C
LED Output Current (IPK, IOUT)	120mA	Lead Temperature (Soldering 10 sec)	260°C

ELECTRICAL CHARACTERISTICS

Over the recommended operating conditions of T_A = 0°C to 70°C, V_{CC} = 5V ± 5%, RTSET = 16.5Ω ± 1%, RPK = 16.5Ω ± 1%, unless otherwise specified. (Notes 2 and 3)

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP.	MAX	UNITS
I _{CC}	Supply Current	LED off			20	mA
I _{LEDH} I _{LEDL}	LED and Peaking Current Accuracy (IOUT, IPK) High Low		57 0.5	60 0.7	63 1.0	mA mA
t _R	Rise Time (IOUT, IPK)				2	ns
t _F	Fall Time (IOUT, IPK)				2	ns
t _{PLH} t _{PHL}	Propagation Delay (IOUT, IPK) Low to High High to Low				10 10	ns ns
t _{PWD}	Pulse Width Distortion (IOUT, IPK)				1.0	ns
I _{ECL}	ECL Input Current				20	μA
V _{DO}	Dropout Voltage between pin 6 and 7		1.5			V

Note 1: Absolute maximum ratings are limits beyond which the life of the integrated circuit may be impaired. All voltages unless otherwise specified are measured with respect to ground.

Note 2: Limits are guaranteed by 100% testing, sampling, or correlation with worst-case test conditions.

Note 3: Low Duty Cycle pulse testing is performed at T_A.

FUNCTIONAL DESCRIPTION

The ML6632 accepts ECL input signals and generates a high speed, high accuracy output current which is independent of supply voltage variations. The output current is programmable from 20mA to 120mA.

The ECL input stage is a standard NPN differential pair with a common mode range of between 1V and 4.5V with a +5V supply. With this common mode range it is possible to convert the ECL inputs into TTL. If the ECLN input is biased up to the TTL switching level, the ECLP pin can be driven by a TTL or CMOS output. Figure 1d shows a circuit implementing this technique. This circuit may degrade pulse width distortion and should be checked for acceptable performance in this configuration.

Output current to the LED is set by connecting the appropriate resistance from RTSET to ground. The high level output voltage at RTSET will be 1.0V. The current in the external resistor will be equal to the current through the LED. The output current with RTSET set to 16.5Ω will be

$$I_{LED(HIGH)} = 1.0V/R_{TSET} = 1.0V/16.5\Omega = 60mA.$$

With RTSET = 16.5Ω, the low level output current is approximately 0.7mA. This current prebiases the LED and results in faster optical rise times.

The ML6632 contain two separate, but identical current driven output stages. These two stages can be paralleled

to double the drive current or one can be used for peaking while the other is used to drive the LED. See Figure 1c.

The ML6632 provides for peaking of the LED output current. Peaking is used to counteract the effects of the LED junction capacitance. By creating a controlled overshoot and undershoot in the output current waveform, charge is transferred to and from the LED capacitance on the rising and falling edges of the output, speeding up rise and fall times.

To provide peaking current, a second output stage is biased up with a resistor from RPK to ground and another from IPK to VCC. When these bias resistors are set equal to each other, a pulse will be generated across the RPEAK resistor with a magnitude equal to the voltage on the RTSET pin. A coupling capacitor transfers the peaking current from the IPK pin to the LED on the rising and falling edges of the output current waveform.

A typical application is shown in Figure 1c. When the resistors R_{RPK} and R_{IPK} are both set to 16Ω, a pulse will be generated at the IPK pin of magnitude 1.0V and equivalent resistance 16Ω.

The peaking current is coupled through the 100 pF capacitor, C_{PEAK} , which will transfer 100 pC of charge to and from the LED on each cycle of output current. The peaking circuit shown provides approximately a 70% overshoot current into a 0Ω LED impedance. Peaking currents will be slightly lower for real LED's.

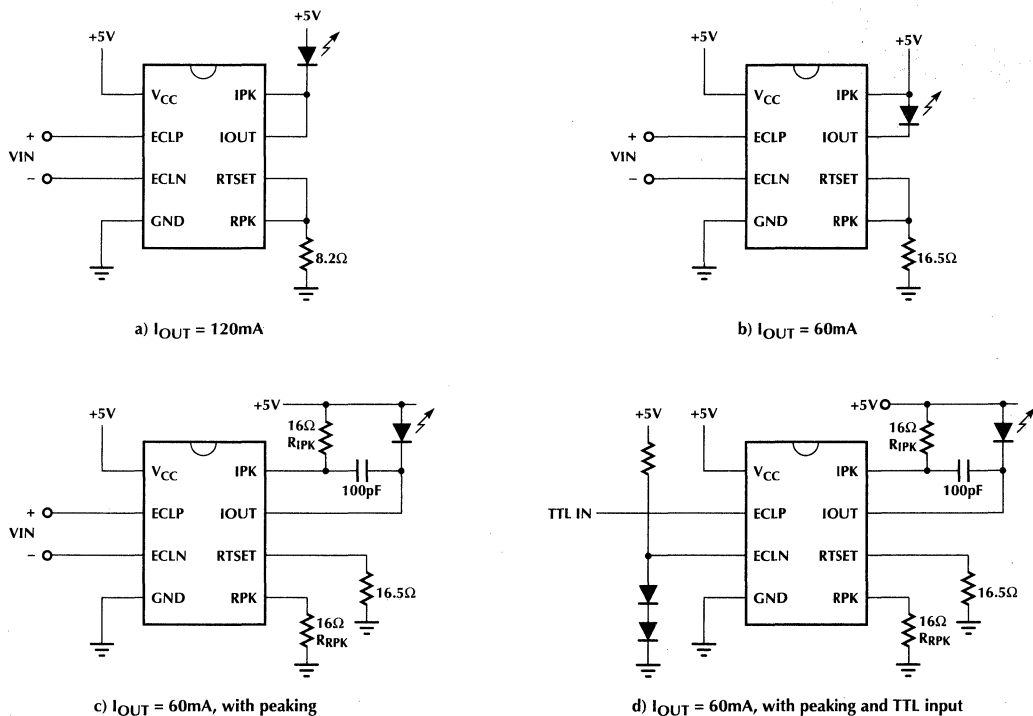


Figure 1. Typical Applications

ML6632

ORDERING INFORMATION

PART NUMBER	TEMPERATURE RANGE	PACKAGE
ML6632CP	0°C to +70°C	8-Pin Molded DIP (P08)
ML6632CS	0°C to +70°C	8-Pin SOIC (S08N)

TP-PMD MLT-3 Transceiver

GENERAL DESCRIPTION

The ML6671 is a complete monolithic transceiver for 125 Mbaud MLT-3 encoded data transmission over Category 5 Unshielded Twisted Pair and Shielded Twisted Pair cables. The adaptive equalizer in the ML6671 will accurately compensate for line losses of up to 100m of UTP. The part is internally trimmed during manufacturing and requires only an external 1% resistor for accurate equalization.

The ML6671 receive section consists of an equalizing filter with a feedback loop for controlling effective line compensation. The feedback loop contains a filter and detection block for determining the proper control signal. The ML6671 also contains a pair of data comparators with precisely controlled slicing thresholds and an MLT3 to NRZI translator. An ECL 100K compatible buffer at the output interfaces directly with existing FDDI PHY silicon from various manufacturers.

The ML6671 transmit section accepts ECL 100K compatible NRZI inputs and converts them to differential current-mode MLT-3 signals. Transmit amplitude is controlled by a single resistor.

Several additional functions are provided by the ML6671 to simplify applications. A common-mode reference is provided to set the input DC level for the equalizer and

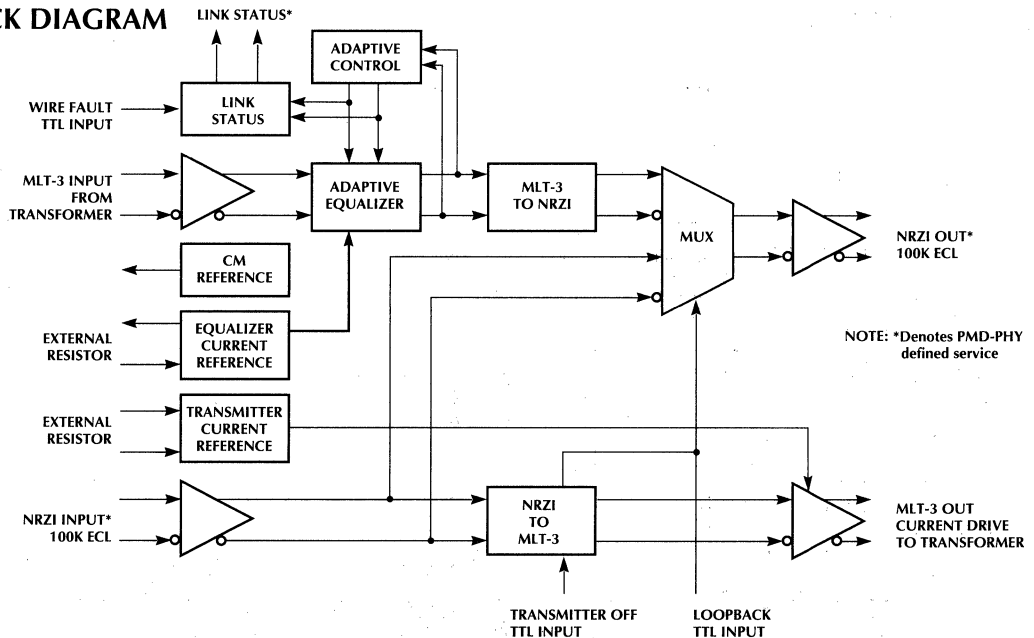
the near-end transformer winding. This terminal may be used as an AC ground for the transformer center-tap or termination resistors. A link status circuit monitors line integrity and provides a proper logic level output signal to interface with the host system.

The ML6671 is implemented in a 1.5 μ m BiCMOS process. A differential signal path throughout minimizes the effects of power supply transients and noise. A variety of package options are available to accommodate surface mount and thru-hole assembly requirements.

FEATURES

- Complies with ANSI X3T9.5 TP-PMD emerging standard
- Transmitter converts NRZI ECL signals to MLT-3 current driven outputs
- Transmitter can be externally turned off for idle generation
- Receiver includes adaptive equalizer and MLT-3 to NRZI decoder
- Operates over 100 meters of STP or category 5 UTP Twisted Pair Cable
- 1.5 μ m BiCMOS process
- 28 pin surface mount package
- Semi-standard options available

BLOCK DIAGRAM



Token Ring Physical Interface

GENERAL DESCRIPTION

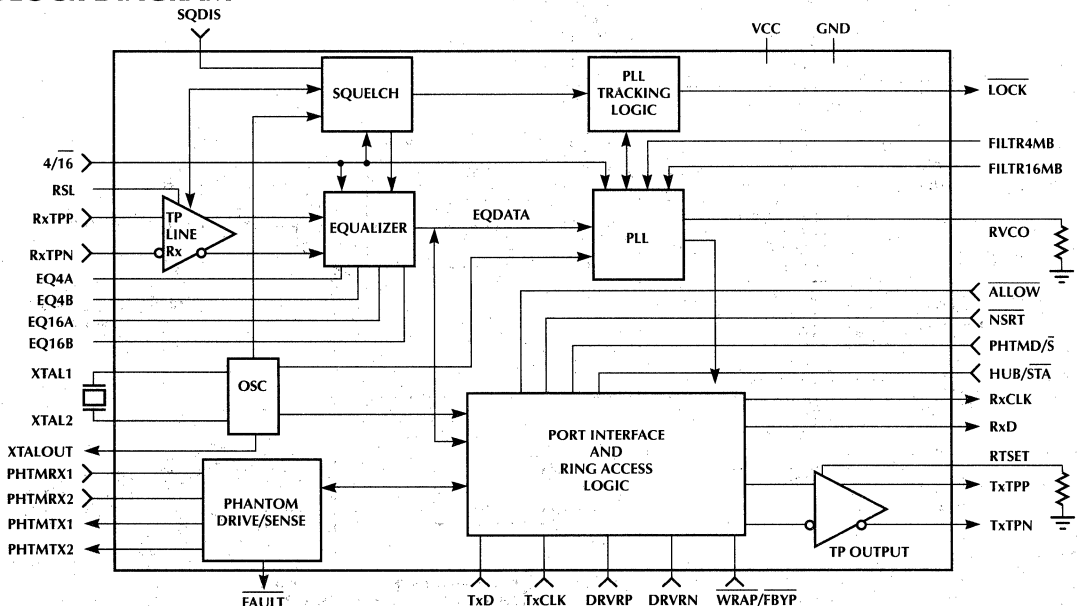
The ML6682 Token Ring Physical Interface Circuit is designed for IEEE 802.5 networks using shielded twisted pair (STP) or unshielded twisted pair (UTP) media. It may be used either as a station port front-end for a token ring MAC controller, or as a lobe or Ring In/Ring Out port in an active concentrator. It includes a receiver equalizer for suppression of inter-symbol interference jitter, a narrow bandwidth PLL with a constant-gain phase/frequency detector for enhanced clock tracking and low VCO output phase distortion, internal frequency/phase tracking select logic, and a twisted pair transmit driver and receiver. The circuit also includes phantom wire fault detection and output drivers for use with an external phantom switching mechanism.

External components are minimized by the use of internally-controlled station fault, watchdog timer, receiver pulsewidth squelch, on-chip crystal oscillator, and internal 4/16 Mbps switching logic. External pin connections select either station port or concentrator port operation and data rate (4mb/s or 16mb/s). In a concentrator application the ML6682 performs the switching function eliminating the need for relays. Isolation can be achieved optically. The circuit requires a single +5V power supply, and is fabricated in BiCMOS technology.

FEATURES

- Supports the complete interface for both a station and a concentrator.
- Supports Active Retiming and Regeneration for each lobe port and Ring In/Ring Out ports for UTP/STP Extended distance concentrators.
- Provides complete physical interface for a UTP/STP station port and full compatibility with TMS380C16/TMS380C26.
- Pin-selectable 16 and 4Mb/s data rates
- Supports fault tolerant Ring In/Ring Out Trunks
- Fault isolation at each concentrator port available for Network Management.
- Provides Phase-Locked Loop with constant gain phase detector for clock regeneration and data recovery.
- Phantom voltage drive/sense for both transmit and receive cable pairs.
- On-chip crystal oscillator can also be driven by external clock.
- On-chip receiver channel equalization switchable for both 4 and 16 Mbps
- 44-Pin package
- Advanced BiCMOS technology

BLOCK DIAGRAM



Disk Drive ICs

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Disk Drive Component Selection Guide

1. READ/WRITE AMPLIFIERS

Part Number	Numbers of Channels	Head Type	Max Input Noise (nV/VHz)	Write Current Range (mA)	Key Features	Package Options
ML117	2, 4 or 6	Ferrite	2.1	10 to 50	Write Current Disable Function	PDIP-18, 22, 28; SO-18, 24; PCC-28
ML117R	2, 4 or 6	Ferrite	2.1	10 to 50	Internal Damping Resistors	PDIP-18, 22, 28; SO-18, 24; PCC-28
ML501	6 or 8	Ferrite	1.5	10 to 50	Enhanced Write Stability	PDIP-28, 40; SO-32; PCC-28, 44
ML501R	6 or 8	Ferrite	1.5	10 to 50	ML501 with Internal Damping Resistors	PDIP-28, 40; SO-28, 32; PCC-28, 44
ML502	6 or 8	Thin Film	1.5	10 to 50	Enhanced Write Stability	PDIP-28, 40; SO-32; PCC-28, 44
ML502R	6, 7 or 8	Thin Film	1.5	10 to 50	ML502 with Internal Damping Resistors	PDIP-28, 40; SO-32; PCC-28, 44
ML511	4, 6 or 8	Ferrite	1.5	10 to 40	Improved Write Stability	SO-24; PCC-28, 44
ML511R	4, 6, 7 or 8	Ferrite	1.5	10 to 40	ML511 with Internal Damping Resistor	SO-24; PCC-28, 44
ML4415	15	Ferrite	1.5	10 to 40	Improved Write Current Stability	PCC-44
ML4416	14	Ferrite	1.5	10 to 40	Chip Select Input	PCC-44
ML4610R	2 or 4	Thin Film	0.85	5 to 35	Switchable Damping Res. (700Ω)	SO-16, 20
ML4611R	4	Thin Film	0.85	5 to 35	Switchable Damping Res. (700Ω) and Write Current Adjust	SO-24

2. READ/WRITE SIGNAL PROCESSING

Part Number	Function	Key Feature	Package Option
ML4041	Read Data Processor	Fast AGC Recovery, 1ns Pulse Pairing	PDIP-24, SO-24, PCC-28
ML4042	ML4041 with Undervoltage Detector	Fast AGC Recovery, 1ns Pulse Pairing	PDIP-28, SO-28, PCC-28
ML4417 ML4427	Zone Bit Recording IC	100 MHz VCO	SO-16, PDIP-16
ML4568	Pulse Detector with Embedded Servo	5V Only; 1ns Pulse Pairing	PCC-28
ML541	Read Data Processor	15 MBits/sec Data Rate	PDIP-24, CERDIP-24, PCC-28, SO-24

Continued on next page

Disk Drive Component Selection Guide (Continued)

2. READ/WRITE SIGNAL PROCESSING (Continued)

Part Number	Function	Key Feature	Package Option
ML6005	24 Mbps HDD Filter/Equalizer	Low Power/High Performance	SSOP-20
ML6006	36 Mbps HDD Filter/Equalizer	Low Power/High Performance	SSOP-20
ML6010	36 Mbps Read Channel Combo	Low Cost/High Integration Configurable Array	QFP-52
ML8464B	Pulse Detector	DP8464B Second Source	PDIP-24, PCC-28
ML8464C	Pulse Detector	1ns Pulse Pairing	PDIP-24, PCC-28

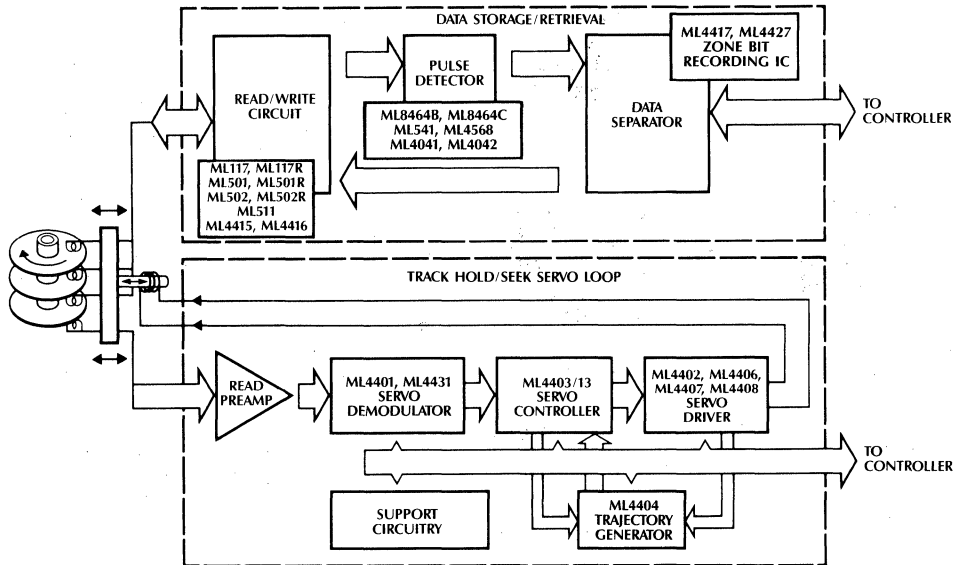
3. SERVO CONTROL ICs

Part Number	Function	Key Feature	Package Option
ML4401	Servo Demodulator	ECL Output VCO	PDIP-28, PCC-28
ML4402	Servo Driver, External Power Drive	Low Offset ($\pm 5\text{mV}$)	PDIP-20, PCC-20
ML4403	Servo Controller	On-Chip Interpolation Function	PDIP-20, PCC-20
ML4404	Analog Trajectory Generator	User-Defined Trajectory, 2 DACs	PDIP-28, PCC-28
ML4406	Servo Driver, Internal Power Drive	Internal Threshold Reference	PCC-20
ML4407	Servo Driver, Internal Power Drive	External Threshold Reference	PCC-20
ML4408	Low Voltage Drop Servo Driver	5V Only or 12V Operation	SO-24
ML4413	Servo Controller	ML4403 with Ext. Amp. Nulling	PDIP-24, PCC-28
ML4431	Servo Demodulator	Enhanced ML4401; TTL Output	PCC-32
ML4532	Servo Burst Area Detector	Includes PWM DAC	SSOP-20, PCC-20
ML4533	Servo Burst Area Detector	No PWM DAC, Reference Levels Compatible to ML A/D Converters	SO-16
ML4536	Servo Burst Area Detector	No PWM DAC, Reference Levels Compatible to Zilog μC with ADC	SO-16
ML4534	SUM/DIFF Area Detector	For Hybrid Servo	PCC-20
ML4535	Hybrid Servo Demodulator	Integration/Area Detection	PCC-32

4. SPINDLE MOTOR CONTROL ICs

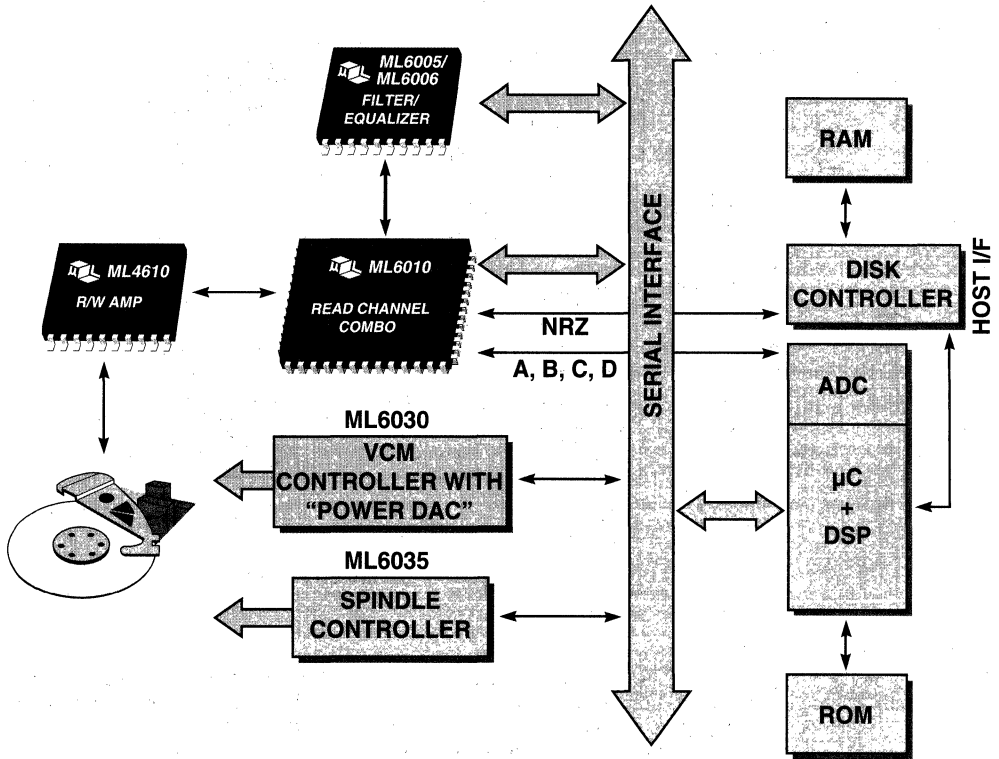
Part Number	Function	Key Feature	Package Option
ML4410	Sensorless Spindle Motor Control	Back-EMF-Commutation	PCC-28

Hard Disk Drive Data Path and Servo Control Diagram



Micro Linear provides a full set of Winchester Hard Disk Drive support chips including the data path and the head servo positioning path. Micro Linear supports both dedicated and embedded servo disk drives with read-write preamps, pulse detectors, data separators, servo demodulators, controllers and drivers, and 8- and 10-bit data converters for digital servo systems

**Next Generation
Disk Drive System**



2, 4, or 6-Channel Read/Write Circuits

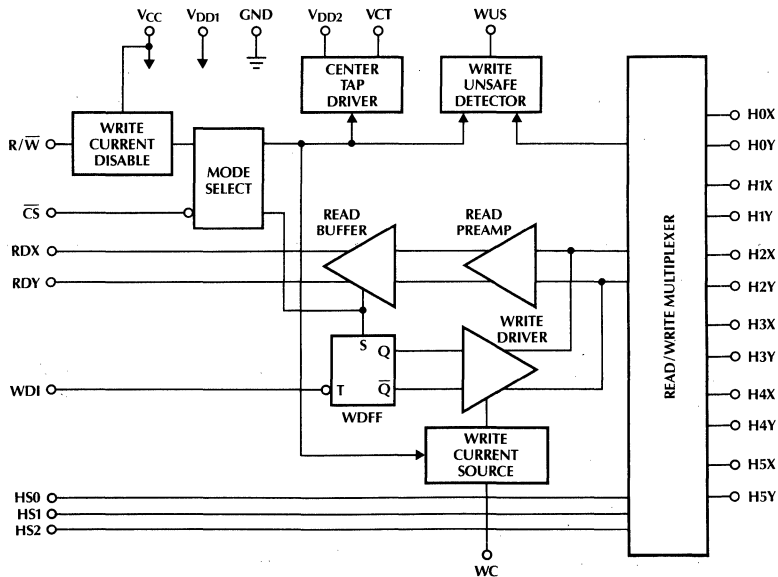
GENERAL DESCRIPTION

The ML117 devices are bipolar monolithic integrated circuits designed for use with center-tapped ferrite recording heads. They provide a low noise read path, write current control, and data protection circuitry for as many as six channels. The ML117 requires +5V and +12V power supplies and is available in 2, 4, or 6-channel versions with a variety of packages. The ML117 contains exclusive circuitry that inhibits write current during device power-up, thereby eliminating power-up "glitches" common to similar read/write circuits. The ML117R differs from the ML117 by having internal damping resistors.

FEATURES

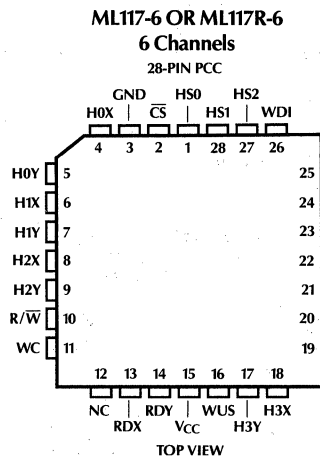
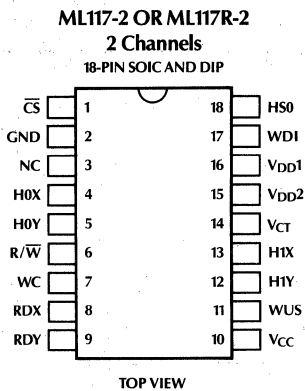
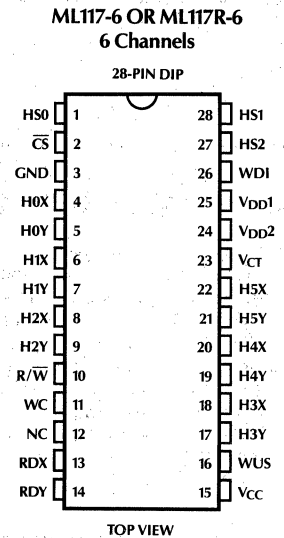
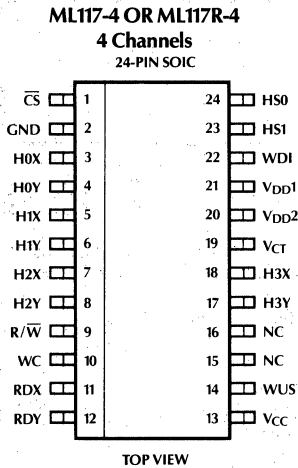
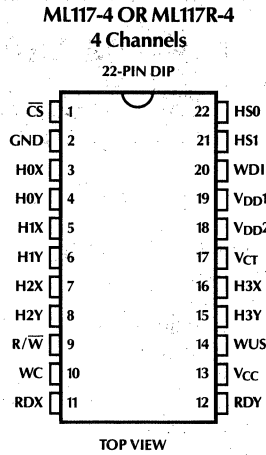
- Exclusive write current disable during power-up
- Replacement for SSI 32R117/117R
- +5V, +12V power supplies
- Single or multi-platter Winchester drives
- Designed for center-tapped ferrite heads
- Programmable write current source
- Available in 2, 4, or 6 channels
- Easily multiplexed for larger systems
- Includes write unsafe detection
- TTL compatible control signals

BLOCK DIAGRAM



ML117, ML117R

PIN CONNECTIONS



PIN DESCRIPTION

NAME	FUNCTION	NAME	FUNCTION
HS0-HS2	Head Select (six heads)	RDX, RDY	X, Y Read Data (differential read signal out)
\overline{CS}	Chip Select (low level enables chip)	WC	Write Current (used to set the write current magnitude)
R/\overline{W}	Read/Write (high level selects Read mode)	V_{CT}	Voltage Center Tap (center tap voltage source)
WUS	Write Unsafe, open collector output (high level indicates alarm)	V_{CC}	+5 volts
WDI	Write Data In (negative transition toggles head current direction)	V_{DD1}	+12 volts
H0X-H5X	X head connections	V_{DD2}	Positive supply for center tap
H0Y-H5Y	Y head connections	GND	Ground

ABSOLUTE MAXIMUM RATINGS

(Note 1)

Power Supply Voltage Range	
V _{DD1}	-0.3 to 14V _{DC}
V _{DD2}	-0.3 to 14V _{DC}
V _{CC}	-0.3 to 6V _{DC}
Input Voltage Range	
Digital Inputs (\overline{CS} , R/ \overline{W} , HS, WDI)	-0.3 to V _{CC} +0.3V _{DC}
Head Ports (H0X-H5X, H0Y-H5Y)	-0.3 to V _{DD1} +0.3V _{DC}
Write Unsafe (WUS)	-0.3 to 14V _{DC}
Write Current (I _W)	60mA
Output Current	
Read Data (RDX, RDY)	-10mA
Center Tap Current (I _{CT})	-60mA
Write Unsafe (WUS)	12mA
Storage Temperature	-65°C to 150°C
Junction Temperature (T _J)	125°C
Lead Temperature (Soldering 10 sec.)	300°C

OPERATING CONDITIONS

Supply Voltage	
V _{DD1}	12V ±10%
V _{CC}	5V ±10%
V _{DD2}	6.5 to V _{DD1}
Head Inductance (L _H)	5 to 15μH
Damping Resistor (R _D , ML117 only)	500 to 2000Ω
RCT Resistor (1/2 Watt)	130Ω ±5%
Write Current (I _W)	25 to 50mA

ELECTRICAL CHARACTERISTICS

Unless otherwise specified V_{DD1}=12V ±10%, V_{CC}=5V ±10%, 0°C ≤ T_A ≤ 70°C (Notes 2 and 3).

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
--------	-----------	------------	-----	-----	-----	-------

DC OPERATING CHARACTERISTICS

POWER SUPPLY

I _{CC}	V _{CC} Supply Current	Read or Idle Mode			25	mA
		Write Mode			30	mA
I _{DD}	V _{DD} Supply Current	Read Mode			50	mA
		Write Mode			30+I _W	mA
		Idle Mode			25	mA
P _D	Power Dissipation	Read Mode			600	mW
		Write Mode I _W =50mA, R _{CT} =130Ω			700	mW
		Write Mode I _W =50mA, R _{CT} =0Ω			1050	mW
		Idle Mode			400	mW

DIGITAL INPUTS (\overline{CS} , R/ \overline{W} , HS, WDI)

V _{IH}	High Voltage		2		V _{CC} +0.3	V _{DC}
V _{IL}	Low Voltage		-0.3		0.8	V _{DC}
I _{IH}	High Current	V _{IH} =2.0V			100	μA
I _{IL}	Low Current	V _{IL} =0.8V	-0.4			mA

WUS OUTPUT

V _{OL}	Output Low Voltage	I _{OL} =8mA (Safe)			0.5	V _{DC}
I _{OH}	Output High Current	V _{OH} =5V (Unsafe)			100	μA

CENTER TAP VOLTAGES

V _{CT}	Read Mode			4		V _{DC}
V _{CT}	Write Mode			6		V _{DC}

5

ML117, ML117R

ELECTRICAL CHARACTERISTICS (Continued)

Unless otherwise specified $V_{DD1} = 12V \pm 10\%$, $V_{CC} = 5V \pm 10\%$, $I_W = 45\text{ mA}$, $L_H = 10\mu\text{H}$, $R_D = 750\Omega$, $f_{DATA} = 5\text{ MHz}$, C_L (RDX, RDY) $\leq 20\text{ pF}$, $0^\circ\text{C} \leq T_A \leq 70^\circ\text{C}$ (Notes 2 and 3) (V_{IN} is referenced to V_{CT} for Read Mode Characteristics).

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
WRITE MODE CHARACTERISTICS						
I_{WR}	Write Current Range	$I_W = K/R_{WC}$	10		50	mA
K	Write Current Constant		133		147	V
V_{HD}	Differential Head Voltage Swing		8			V_{PK}
I_{HU}	Unselected Head Transient Current				2	mA_{PK}
C_{OD}	Differential Output Capacitance				15	pF
R_{OD}	Differential Output Resistance	ML117	10k			Ω
		ML117R	562		938	Ω
f_{WDI}	WDI Transition Frequency	WUS = Low	250			kHz
A_I	I_{WS} to Head Current Gain			20		A/A
I_L	Unselected Head Leakage	Sum of X & Y Side Leakage Current			85	μA
READ MODE CHARACTERISTICS						
A_V	Differential Voltage Gain	$V_{IN} = 1\text{ mV}_{P,P} @ 300\text{ kHz}$, R_L (RDX, RDY) = $1\text{ k}\Omega$	80		120	V/V
DR	Dynamic Range	DC Input Voltage (V_i) Where Gain Falls 10%, $V_{IN} = V_i + 0.5\text{ mV}_{P,P} @ 300\text{ kHz}$	-3		+3	mV
BW	Bandwidth (-3 dB)	$ Z_S < 5\Omega$, $V_{IN} = 1\text{ mV}_{RMS}$	30			MHz
e_{IN}	Input Noise Voltage	BW = 15 MHz, $L_H = 0$, $R_H = 0$			2.1	$\text{nV}/\sqrt{\text{Hz}}$
C_{IN}	Differential Input Capacitance				20	pF
R_{IN}	Differential Input Resistance	ML117	2k			Ω
		ML117R	390		810	Ω
I_{IN}	Input Bias Current				45	μA
CMRR	Common-Mode Rejection Ratio	$V_{CM} = V_{CT} + 100\text{ mV}_{P,P} @ f = 5\text{ MHz}$	50			dB
PSRR	Power Supply Rejection Ratio	$100\text{ mV}_{P,P} @ 5\text{ MHz}$ on V_{DD1} , V_{DD2} , or V_{CC}	45			dB
CS	Channel Separation	Unselected Channels: $V_{IN} = 100\text{ mV}_{P,P} @ 5\text{ MHz}$ and Selected Channel: $V_{IN} = 0\text{ mV}_{P,P}$	45			dB
V_{OS}	Output Offset Voltage		-480		+480	mV
V_{OCM}	Common-Mode Output Voltage	Read Mode	5		7	V
		Write or Idle Mode		4.3		V
R_{OUT}	Single-Ended Output Resistance	$f = 5\text{ MHz}$			30	Ω
I_L	Leakage Current, RDX, RDY	RDX, RDY = 6V Write or Idle Mode	-100		+100	μA
I_O	Output Current	AC Coupled Load, RDX to RDY	2			mA

ELECTRICAL CHARACTERISTICS (Continued)

Unless otherwise specified $V_{DD1} = 12V \pm 10\%$, $V_{CC} = 5V \pm 10\%$, $I_W = 45\text{ mA}$, $L_H = 10\mu\text{H}$, $R_D = 750\Omega$, $f_{DATA} = 5\text{ MHz}$, $0^\circ\text{C} \leq T_A \leq 70^\circ\text{C}$ (Notes 2 and 3).

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
SWITCHING CHARACTERISTICS						
t_{RW}	R/ \bar{W} to Write Switching Delay	To 90% of Write Current Output			1	μS
t_{WR}	R/ \bar{W} to Read Switching Delay	To 90% of 100mV, 10MHz Read Signal Envelope or to 90% Decay of Write Current			1	μS
t_{IW} or t_{IR}	$\overline{\text{CS}}$ to Select Switching Delay	To 90% of Write Current or to 90% of 100mV, 10MHz Read Signal Envelope			1	μS
t_{WI} or t_{RI}	$\overline{\text{CS}}$ to Select Switching Delay	To 90% Decay of 100mV, 10MHz Read Signal Envelope or to 90% Decay of Write Current			1	μS
t_{HS}	Head Select Switching Delay	To 90% of 100mV, 10MHz Read Signal Envelope			1	μS
t_{D1}	Safe to Unsafe Write Unsafe Delay	$I_W = 50\text{ mA}$	1.6		8	μS
t_{D2}	Unsafe to Safe Write Unsafe Delay	$I_W = 50\text{ mA}$			1	μS
t_{D3}	Head Current Prop. Delay	$L_H = 0$, $R_H = 0$ From 50% points			25	nS
t_{D3}	Head Current Asymmetry	WDI has 50% Duty Cycle and 1nS Rise/Fall Time			2	nS
	Time Head Current Rise/Fall	10% and 90% Points			20	nS

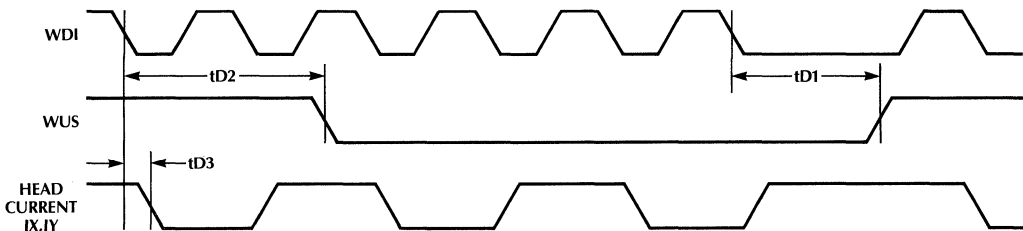
Note 1: Absolute maximum ratings are limits beyond which the life of the integrated circuit may be impaired. All voltages unless otherwise specified are measured with respect to ground.

Note 2: Limits are guaranteed by 100% testing, sampling, or correlation with worst-case test conditions.

Note 3: Maximum junction temperature (T_j) should not exceed 125°C .

5

TIMING DIAGRAMS



Write Mode Timing Diagram

FUNCTIONAL DESCRIPTION

CIRCUIT OPERATION

The ML117, ML117R functions as a write driver or as a read amplifier for the selected head. Head selection and mode control are described in *Tables 1 & 2*. Both R/W and CS have internal pull-up resistors for the prevention of an accidental write condition.

READ MODE

In the Read Mode the ML117, ML117R is configured as a low noise differential amplifier, the write current source and the write unsafe detector are deactivated, and the write data flip-flop is set. The RDX and RDY outputs are driven by emitter followers and are in phase with the "X" and "Y" head ports.

The internal write current source is deactivated for both the Read and the Chip Deselect modes which eliminates the need for external gating of the write current source.

WRITE MODE

The Write mode configures the ML117, ML117R as a current switch and activates the Write Unsafe Detector. The head current is toggled between the X- and Y-side of the recording head on the falling edges of WDI, Write Data Input. A preceding read operation initializes the Write Data Flip-Flop, WDFF, to pass current through the X-side of the head. The magnitude of the write current, given by:

$$I_W = K/R_{WC}, \text{ where } K = \text{Write Current Constant}$$

is set by the external resistor, R_{WC} , connected from pin WC to GND.

Any of the following conditions will be indicated as a high level on the Write Unsafe, WUS, open collector output.

- Head open
- Head center tap open
- WDI frequency too low
- Device in Read mode
- Device not selected
- No write current

Two negative transitions on WDI are required to clear WUS after the fault condition is removed.

Table 1.

Head Select			
HS2	HS1	HS0	HEAD
0	0	0	0
0	0	1	1
0	1	0	2
0	1	1	3
1	0	0	4
1	0	1	5
1	1	X	NONE

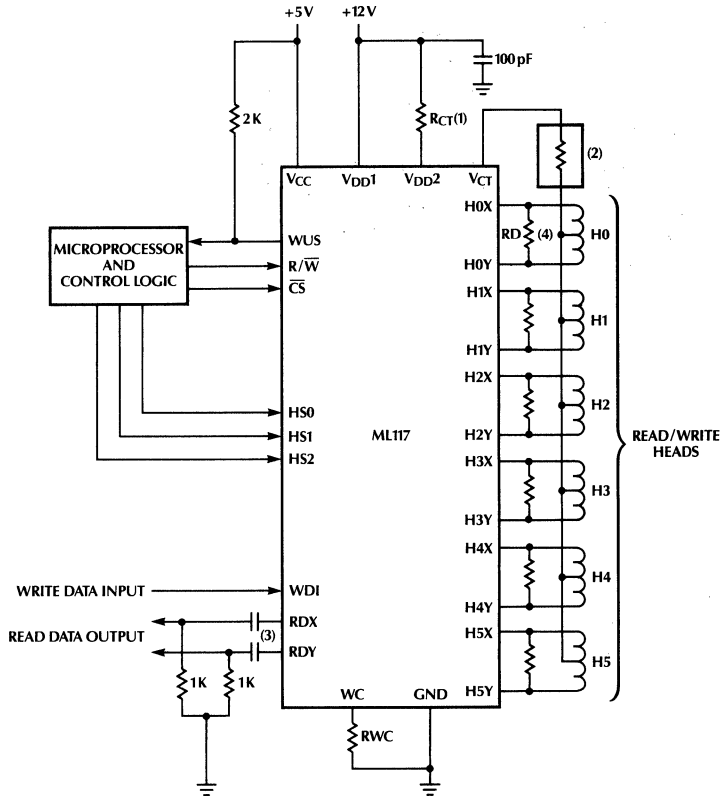
0 = Logic Level Low
 1 = Logic Level High
 X = Don't Care

Table 2.

Mode Select		
$\overline{\text{CS}}$	R/W	MODE
0	0	Write
0	1	Read
1	X	Idle

0 = Logic Level Low
 1 = Logic Level High
 X = Don't Care

TYPICAL APPLICATION



NOTES:

1. RCT is optional and is used to limit internal power dissipation (Otherwise connect V_{DD1} to V_{DD2}).
 $RCT (1/2 \text{ Watt}) = 130 (55/I_w)$ ohms
 where I_w = Write Current, in mA
2. Ferrite bead optional: used to suppress write current overshoot and ringing. Recommend Ferroxcube 3659065/4A6.
3. RDX and RDY load capacitance 20 pF maximum. RDX and RDY output current must be limited to 100 μ A.
4. Damping resistors not required on ML117R.

THERMAL CHARACTERISTICS

28-Lead PDIP	80°C/W
PCC	60°C/W
24-Lead SOIC	60°C/W
22-Lead PDIP	100°C/W
18-Lead PDIP	115°C/W
SOIC	85°C/W

ML117, ML117R

ORDERING INFORMATION

PART NUMBER	PACKAGE	NUMBER OF CHANNELS
ML117-2CP	18-Lead Molded DIP (P18)	2
ML117R-2CP	18-Lead Molded DIP (P18)	2
ML117-2CS	18-Lead Molded SOIC (S18)	2
ML117R-2CS	18-Lead Molded SOIC (S18)	2
ML117-4CP	22-Lead Molded DIP (P22)	4
ML117R-4CP	22-Lead Molded DIP (P22)	4
ML117-4CS	24-Lead Molded SOIC (S24)	4
ML117R-4CS	24-Lead Molded SOIC (S24)	4
ML117-6CP	28-Lead Molded DIP (P28)	6
ML117R-6CP	28-Lead Molded DIP (P28)	6
ML117-6CQ	28-Lead PCC (Q28)	6
ML117R-6CQ	28-Lead PCC (Q28)	6

6, 7, or 8-Channel Read/Write Circuits

GENERAL DESCRIPTION

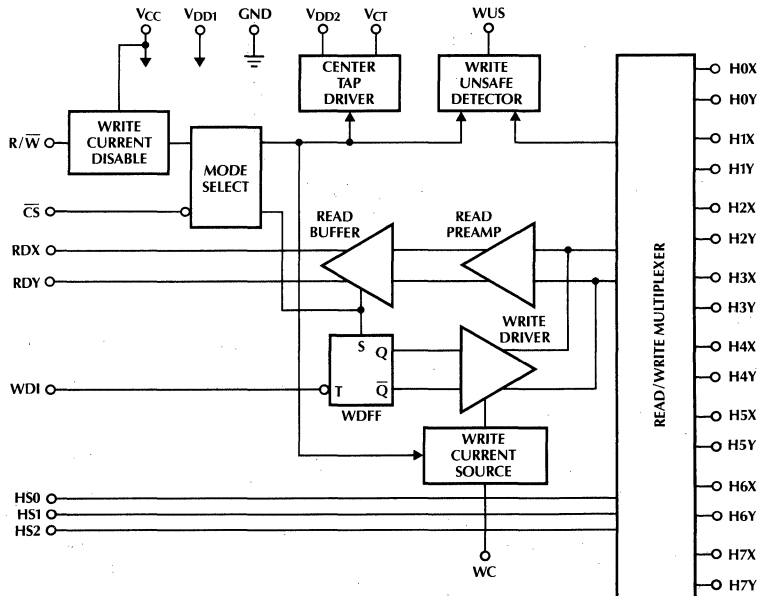
The ML501, ML502 family of devices are bipolar monolithic read/write circuits designed for use with fixed disk center-tapped recording heads. The ML501 and ML501R are designed for use with ferrite recording heads while the ML502, ML502R and ML502S are designed for thin film or composite heads. The R and S designation in the part number indicate that these parts have internal head damping resistors.

The ML501, ML502 family provides up to eight multiplexed read/write data channels. These circuits exhibit features not found in similar read/write circuits such as improved write current stability and the elimination of write current "glitches" during power-up. The exclusive ML502 is identical to the ML501 except that the write unsafe detect circuitry is designed to operate with lower head inductance.

FEATURES

- Exclusive write current disable during power-up
- Enhanced write current stability
- ML501, ML501R is replacement for SSI 32R501/501R and is designed for center-tapped ferrite heads
- ML502, ML502R, and ML502S are designed for center-tapped thin film or composite heads
- Single or multi-platter Winchester drives
- Easily multiplexed for larger systems
- Available in 6, 7 or 8 channels
- TTL compatible control signals
- Programmable write current source
- Includes write unsafe detection
- Available in a selection of packages
- +5V, +12V power supplies

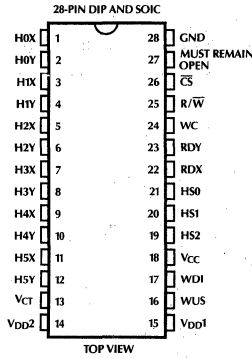
BLOCK DIAGRAM



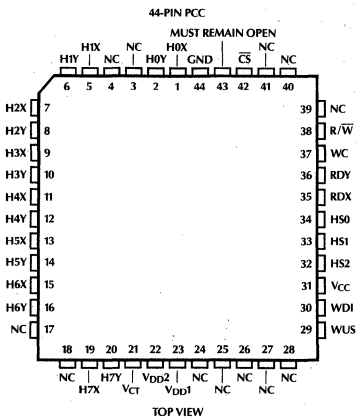
ML501, ML501R, ML502, ML502R, ML502S

PIN CONNECTIONS

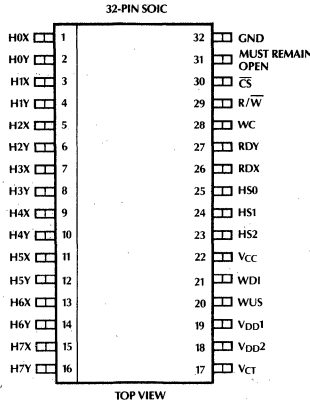
**ML501-6 OR ML501R-6
OR ML502-6 OR ML502R-6
6 Channels**



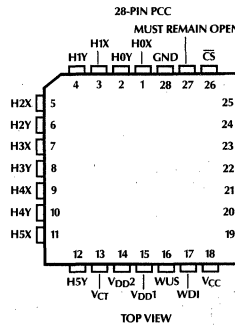
**ML501-8 OR ML501R-8
OR ML502-8 OR ML502R-8
OR ML502S-8
8 Channels**



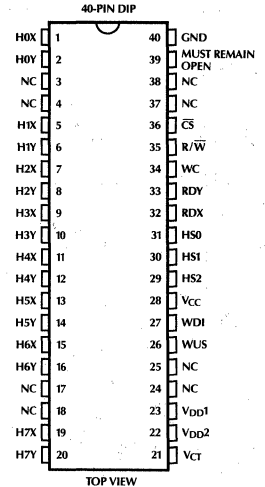
**ML501-8 OR ML501R-8
OR ML502-8 OR ML502R-8
8 Channels**



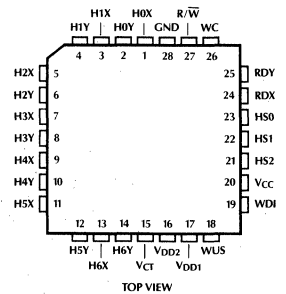
**ML501-6 OR ML501R-6
OR ML502-6 OR ML502R-6
OR ML502S-6
6 Channels**



**ML501-8 OR ML501R-8
OR ML502-8 OR ML502R-8
8 Channels**



**ML502S-7CQ
ML502R-7CQ**



PIN DESCRIPTION

NAME	FUNCTION	NAME	FUNCTION
HS0-HS2	Head Select (eight heads)	RDX, RDY	X, Y Read Data (differential read signal out)
CS	Chip Select (low level enables chip)	WC	Write Current (used to set the write current magnitude)
R/W	Read/Write (high level selects Read mode)	VCT	Voltage Center Tap (center tap voltage source)
WUS	Write Unsafe, open collector output (high level indicates an unsafe writing condition)	VCC	+5 volts
WDI	Write Data In (negative transition toggles head current direction)	VDD1	+12 volts
H0X-H7X	X head connections	VDD2	Positive supply for center tap
H0Y-H7Y	Y head connections	GND	Ground

ML501, ML501R, ML502, ML502R, ML502S

ABSOLUTE MAXIMUM RATINGS

(Note 1)

Power Supply Voltage Range	
V _{DD1}	−0.3 to 14V _{DC}
V _{DD2}	−0.3 to 14V _{DC}
V _{CC}	−0.3 to 6V _{DC}
Input Voltage Range	
Digital Inputs (\overline{CS} , R/\overline{W} , HS, WDI)	−0.3 to V _{CC} + 0.3V _{DC}
Head Ports (H0X–H7X, H0Y–H7Y)	−0.3 to V _{DD1} + 0.3V _{DC}
Write Unsafe (WUS)	−0.3 to 14V _{DC}
Write Current (I _W)	60 mA
Output Current	
Read Data (RDX, RDY)	−10 mA
Center Tap Current (I _{CT})	−60 mA
Write Unsafe (WUS)	12 mA
Storage Temperature	−65°C to 150°C
Junction Temperature (T _J)	135°C
Lead Temperature (Soldering 10 sec.)	300°C

OPERATING CONDITIONS

Supply Voltage	
V _{DD1}	12V ± 10%
V _{CC}	5V ± 10%
Head Inductance	
L _H , ML501 or ML501R only	5 to 15 μH
L _H , ML502, ML502R, ML502S only	400 to 1000 nH
Damping Resistor (R _D , ML501 only)	500 to 2000 Ω
RCT Resistor (1/2 Watt)	120 Ω ± 5%
Write Current (I _W)	22 to 50 mA

ELECTRICAL CHARACTERISTICS

Unless otherwise specified V_{DD1} = 12V ± 10%, V_{CC} = 5V ± 10%, R_{CT} = 120 Ω ± 5%, I_W = 45 mA, 0°C ≤ T_A ≤ 70°C (Notes 2 and 3).

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
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DC OPERATING CHARACTERISTICS

POWER SUPPLY

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
I _{CC}	V _{CC} Supply Current	Read or Idle Mode			25	mA
		Write Mode			25	mA
I _{DD}	V _{DD} Supply Current	Read Mode			48	mA
		Write Mode			25 + I _W	mA
		Idle Mode			20	mA
P _D	Power Dissipation	Read Mode			.770	mW
		Write Mode I _W = 50 mA			830	mW
		Write Mode I _W = 50 mA, R _{CT} = 0 Ω			1125	mW
		Idle Mode			400	mW

DIGITAL INPUTS (\overline{CS} , R/\overline{W} , HS, WDI)

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
V _{IH}	High Voltage		2			V _{DC}
V _{IL}	Low Voltage				0.8	V _{DC}
I _{IH}	High Current	V _{IH} = 2.0V			100	μA
I _{IL}	Low Current	V _{IL} = 0.8V	−0.4			mA

WUS OUTPUT

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
V _{OL}	Output Low Voltage	I _{OL} = 8 mA (Safe)			0.5	V _{DC}
I _{OH}	Output High Current	V _{OH} = 5V (Unsafe)			100	μA

CENTER TAP VOLTAGES

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
V _{CT}	Read Mode	Read Mode		4		V _{DC}
V _{CT}	Write Mode	Write Mode		6		V _{DC}

ML501, ML501R, ML502, ML502R, ML502S

ELECTRICAL CHARACTERISTICS (Continued)

Unless otherwise specified $V_{DD1} = 12V \pm 10\%$, $V_{CC} = 5V \pm 10\%$, $R_{CT} = 120\Omega \pm 5\%$, $I_W = 45\text{ mA}$, $L_H = 10\mu\text{H}$ (ML501, ML501R), $L_H = 600\text{ nH}$ (ML502, ML502R, ML502S), $R_D = 750\Omega$ (ML501), $f_{DATA} = 5\text{ MHz}$, C_L (RDX, RDY) $\leq 20\text{ pF}$, $0^\circ\text{C} \leq T_A \leq 70^\circ\text{C}$ (Notes 2 and 3) (V_{IN} is referenced to V_{CT} for Read Mode Characteristics).

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
WRITE MODE CHARACTERISTICS						
I_{WR}	Write Current Range	$I_W = K/R_{WC}$	10		50	mA
K	Write Current Constant		129		151	V
V_{HD}	Differential Head Voltage Swing		7.5			V_{PK}
I_{HU}	Unselected Head Transient Current				2	mA_{PK}
C_{OD}	Differential Output Capacitance				15	pF
R_{OD}	Differential Output Resistance	ML501, ML502	10k			Ω
		$T_J = 25^\circ\text{C}$ ML501R, ML502S/ML502R	560/180		940/300	Ω
f_{WDI}	WDI Transition Frequency	WUS = Low	250			kHz
A_I	I_{WC} to Head Current Gain			20		A/A
I_L	Unselected Head Leakage	Sum of X & Y Side Leakage Current			85	μA
READ MODE CHARACTERISTICS						
A_V	Differential Voltage Gain	$V_{IN} = 1\text{ mV}_{P,P}$ @ 300kHz, R_L (RDX, RDY) = 1k Ω	90		120	V/V
DR	Dynamic Range	DC Input Voltage (V_I) Where Gain Falls 10%, $V_{IN} = V_I + 0.5\text{ mV}_{P,P}$ @ 300kHz	-3		+3	mV
BW	Bandwidth (-3 dB)	$ Z_S < 5\Omega$, $V_{IN} = 1\text{ mV}_{P,P}$	30			MHz
e_{IN}	Input Noise Voltage	BW = 15 MHz, $L_H = 0$, $R_H = 0$			1.5	$\text{nV}/\sqrt{\text{Hz}}$
C_{IN}	Differential Input Capacitance	$f = 5\text{ MHz}$			23	pF
R_{IN}	Differential Input Resistance	$f = 5\text{ MHz}$, $T_J = 25^\circ\text{C}$ ML501, ML502	2k			Ω
		$V_{IN} = 6\text{ mV}_{P,P}$ ML501R, ML502S/ML502R	530/180		790/300	Ω
I_{IN}	Input Bias Current (1 side)				100	μA
CMRR	Common-Mode Rejection Ratio	$V_{CM} = V_{CT} + 100\text{ mV}_{P,P}$ @ $f = 5\text{ MHz}$	50			dB
PSRR	Power Supply Rejection Ratio	100 $\text{mV}_{P,P}$ @ 5 MHz on V_{DD1} , V_{DD2} , or V_{CC}	45			dB
CS	Channel Separation	Unselected Channels: $V_{IN} = 100\text{ mV}_{P,P}$ @ 5 MHz and Selected Channel: $V_{IN} = 0\text{ mV}_{P,P}$	45			dB
V_{OS}	Output Offset Voltage		-480		+480	mV
V_{OCM}	Common-Mode Output Voltage	Read Mode	5		7	V
		Write or Idle Mode		4.3		V
R_{OUT}	Single-Ended Output Resistance	$f = 5\text{ MHz}$			30	Ω
R_L	External Resistive Load (AC Coupled to Output)	Per Side to GND	100			Ω
I_L	Leakage Current, RDX, RDY	$3V < (RDX, RDY) < 8V$ Write or Idle Mode	-50		50	μA
Z_O	Center Tap Output Impedance	$0\text{ MHz} \leq f \leq 5\text{ MHz}$			150	Ω
I_O	Output Current	AC Coupled Load, RDX to RDY	2			mA

ML501, ML501R, ML502, ML502R, ML502S

ELECTRICAL CHARACTERISTICS (Continued)

Unless otherwise specified $V_{DD1} = 12V \pm 10\%$, $V_{CC} = 5V \pm 10\%$, $R_{CT} = 120\Omega \pm 5\%$, $I_W = 45mA$, $L_H = 10\mu H$ (ML501, ML501R), $L_H = 600nH$ (ML502, ML502R, ML502S), $R_D = 750\Omega$ (ML501), $f_{DATA} = 5MHz$, $0^\circ C \leq T_A \leq 70^\circ C$ (Notes 2 and 3).

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
SWITCHING CHARACTERISTICS						
t_{RW}	R/W to Write Switching Delay	To 90% of Write Current Output			600	ns
t_{WR}	R/W to Read Switching Delay	To 90% of 100mV, 10MHz Read Signal Envelope or to 90% Decay of Write Current			600	ns
t_{IW} or t_{IR}	\overline{CS} to Select Switching Delay	To 90% of Write Current or to 90% of 100mV, 10MHz Read Signal Envelope			600	ns
t_{WI} or t_{RI}	\overline{CS} to Unselect Switching Delay	To 90% Decay of 100mV, 10MHz Read Signal Envelope or to 90% Decay of Write Current			600	ns
t_{HS}	Head Select Switching Delay	To 90% of 100mV, 10MHz Read Signal Envelope			600	ns
t_{D1}	Safe to Unsafe Write Unsafe Delay	$I_W = 50mA$	1.6		8	us
t_{D2}	Unsafe to Safe Write Unsafe Delay	$I_W = 20mA$			1	us
t_{D3}	Prop. Delay Head Current	$L_H = 0$, $R_H = 0$ From 50% points		25	40	ns
t_{D3}	Asymmetry Head Current	WDI has 50% Duty Cycle and 1ns Rise/Fall Time			2	ns
	Rise/Fall Head Current	10% and 90% Points			20	ns

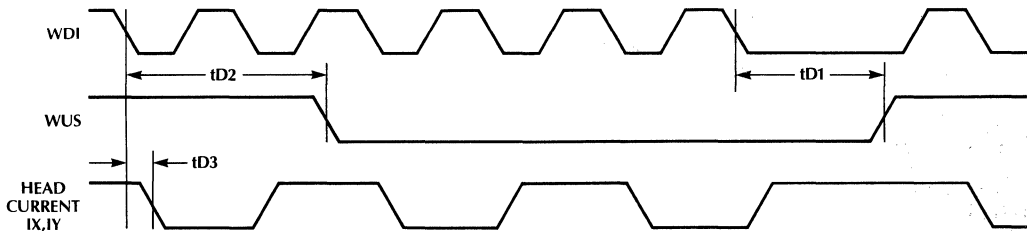
Note 1: Absolute maximum ratings are limits beyond which the life of the integrated circuit may be impaired. All voltages unless otherwise specified are measured with respect to ground.

Note 2: Limits are guaranteed by 100% testing, sampling, or correlation with worst-case test conditions.

Note 3: Maximum junction temperature (T_J) should not exceed 135°C.

5

TIMING DIAGRAM



Write Mode Timing Diagram

ML501, ML501R, ML502, ML502R, ML502S

FUNCTIONAL DESCRIPTION

CIRCUIT OPERATION

For any selected head, the ML501, ML502 functions as a read amplifier when in the Read mode, or as a write current switch when in the Write mode. Pins HS0, HS1 and HS2 determine head selection while pin R/W controls the Read/Write mode. A detected "write-unsafe" condition is indicated by pin WUS.

READ MODE

When the ML501, ML502 is in the Read Mode, it operates as a low-noise differential amplifier on the selected channel. In Read mode the write data flip-flop is set and both the write unsafe detector and the write current source are deactivated. The center tap voltage is also lowered. Pins RDX and RDY provide differential emitter follower outputs which are in phase with the X and Y head input pins.

Note that during the Read or Chip Deselect mode the internal write current is deactivated, thus making external write current gating unnecessary.

WRITE MODE

The ML501, ML502 operates as a write-current switch when in the Write mode. Write current magnitude is determined by the following relationship:

$$I_W = K/R_{WC}$$

Where: K = Write Current Constant

R_{WC} = Resistance connected between pin WC and GND.

The head current is toggled between the X and Y side of the selected head by a negative transition on WDI (Write Data Input). When switching the ML501, ML502 to write mode, the WDFF (Write Data Flip-Flop) is initialized to pass write current through the X-side of the head.

The ML501, ML502 exhibit enhanced write current stability, compared to similar read/write circuits, which reduces the problem of oscillation. This is a result of increased internal write current compensation. Also, write current "glitches" during power-up, common in similar read/write circuits, are eliminated with an exclusive write current disabling function.

The WUS (Write Unsafe) pin is an open collector output that gives a logic high level for any of the following unsafe write conditions:

- Open head
- Open head center-tap
- Too low WDI frequency
- Read mode selected
- Device not selected
- No write current

Two negative transitions on WDI are required to clear WUS after the fault condition is removed.

The ML502, ML502R, ML502S differ from the ML501, ML501R by having write unsafe detect circuitry that is designed to operate with lower amplitude write pulse voltages, which result from the lower head inductance of thin film or composite heads.

Table 1.

Head Select			
HS2	HS1	HS0	HEAD
0	0	0	0
0	0	1	1
0	1	0	2
0	1	1	3
1	0	0	4
1	0	1	5
1	1	0	6
1	1	1	7

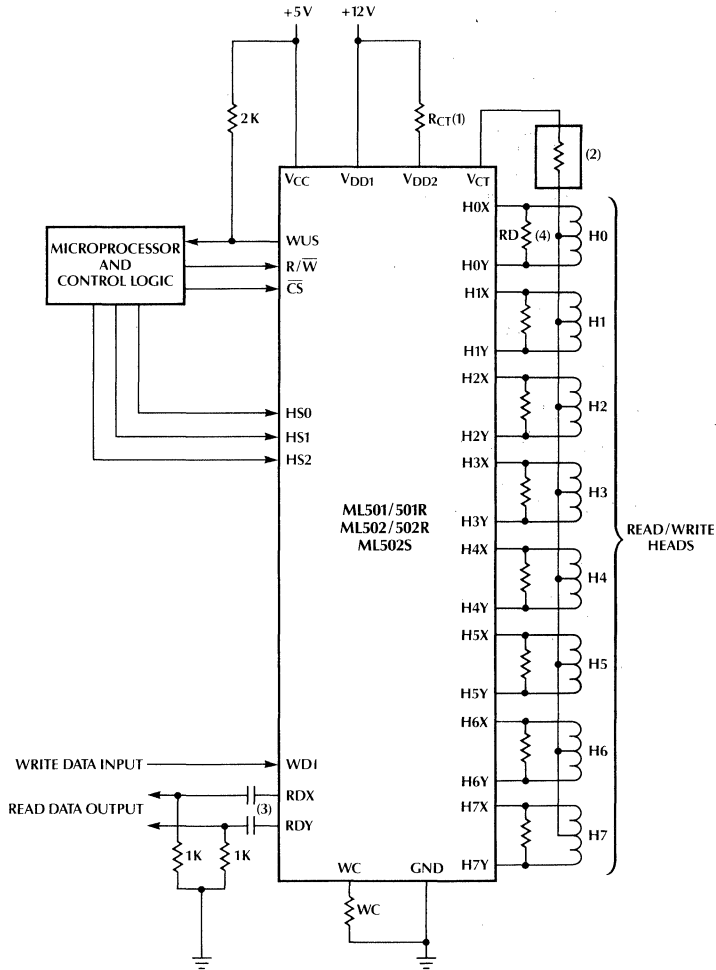
0 = Logic Level Low
 1 = Logic Level High
 X = Don't Care

Table 2.

Mode Select		
\overline{CS}	R/W	MODE
0	0	Write
0	1	Read
1	X	Idle

0 = Logic Level Low
 1 = Logic Level High
 X = Don't Care

TYPICAL APPLICATION



NOTES:

1. RCT is optional and is used to limit internal power dissipation (Otherwise connect V_{DD1} to V_{DD2}).
 $RCT (1/2 \text{ Watt}) = 120 (50 / I_w)$ ohms
 where I_w = Write Current, in mA
2. Ferrite bead optional: used to suppress write current overshoot and ringing. Recommend Ferroxcube 3659065/4A6.
3. RDX and RDY load capacitance 20 pF maximum. RDX and RDY output current must be limited to 100 μ A.
4. Damping resistors not required on ML501R or ML502R.

ML501, ML501R, ML502, ML502R, ML502S

ORDERING INFORMATION

PART NUMBER	PACKAGE	NUMBER OF CHANNELS	TRANSDUCER HEAD TYPE
ML501-6CP	28-Lead Molded DIP (P28)	6	Ferrite Heads
ML501-6CQ	28-Lead PCC (Q28)	6	
ML501-6CS	28-Lead SOIC (S28)	6	
ML501-8CP	40-Lead Molded DIP (P40)	8	
ML501-8CQ	44-Lead PCC (Q44)	8	
ML501-8CS*	32-Lead SOIC (S32)	8	
ML501R-6CP	28-Lead Molded DIP (P28)	6	Ferrite Heads
ML501R-6CQ	28-Lead PCC (Q28)	6	
ML501R-6CS	28-Lead SOIC (S28)	6	
ML501R-8CP	40-Lead Molded DIP (P40)	8	
ML501R-8CQ	44-Lead PCC (Q44)	8	
ML501R-8CS*	32-Lead SOIC (S32)	8	
ML502-6CP	28-Lead Molded DIP (P28)	6	Thin Film or Composite Heads
ML502-6CQ	28-Lead PCC (Q28)	6	
ML502-6CS	28-Lead SOIC (S28)	6	
ML502-8CP	40-Lead Molded DIP (P40)	8	
ML502-8CQ	44-Lead PCC (Q44)	8	
ML502-8CS*	32-Lead SOIC (S32)	8	
ML502R-6CP	28-Lead Molded DIP (P28)	6	Thin Film or Composite Heads
ML502R-6CQ	28-Lead PCC (Q28)	6	
ML502R-6CS	28-Lead SOIC (S28)	6	
ML502R-7CQ	28-Lead PCC (Q28)	7	
ML502R-8CP	40-Lead Molded DIP (P40)	8	
ML502R-8CQ	44-Lead PCC (Q44)	8	
ML502R-8CS*	32-Lead SOIC (S32)	8	
ML502S-6CQ	28-Lead PCC (Q28)	6	Thin Film or Composite Heads
ML502S-7CQ	28-Lead PCC (Q28)	7	
ML502S-8CQ	44-Lead PCC (Q44)	8	

* This package is available as a special order only.

ML511, ML511R-Series

4, 6, 7, or 8-Channel Ferrite Read/Write Circuits

GENERAL DESCRIPTION

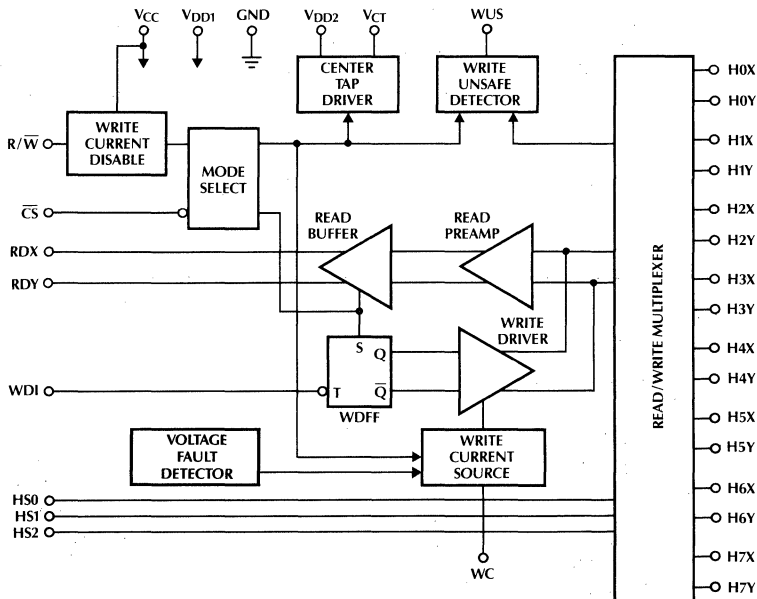
The ML511 is a bipolar monolithic read/write circuit designed for use with center-tapped ferrite recording heads. The ML511 and ML511R are performance upgrades from the ML501 and ML501R. The R designation in the part number indicates that this part has internal head damping resistors.

The ML511 provides up to eight multiplexed read/write data channels. These circuits exhibit features not found in similar read/write circuits such as improved write current stability and the elimination of write current "glitches" during power-up. The ML511 also provides a low noise read data path, and data protection circuitry for all of the channels.

FEATURES

- Enhanced write current stability
- ML511, ML511R is replacement for SSI 32R511/511R and is designed for center-tapped ferrite heads
- Single or multi-platter Winchester drives
- Easily multiplexed for larger systems
- Power supply fault protection
- 1.5 nV/√Hz maximum input noise voltage
- TTL compatible control signals
- Programmable write current source
- Includes write unsafe detection
- Available in a selection of packages
- +5V, +12V power supplies

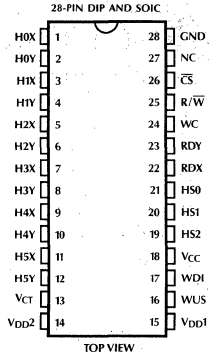
BLOCK DIAGRAM



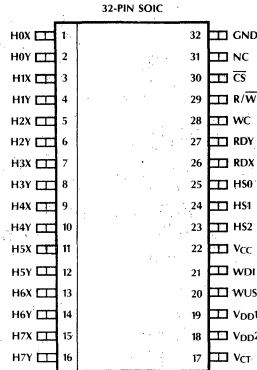
ML511, ML511R

PIN CONNECTIONS

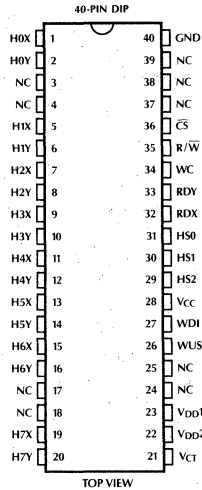
ML511-6 OR ML511R-6
6 Channels



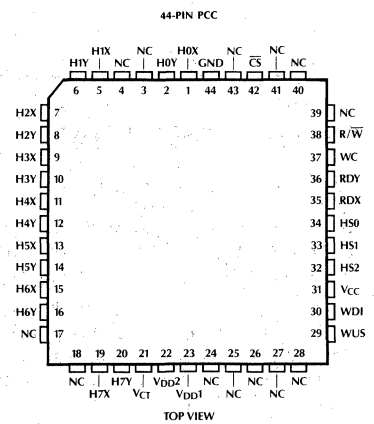
ML511-8 OR ML511R-8
8 Channels



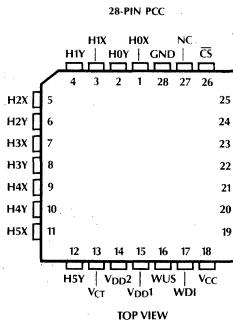
ML511-8 OR ML511R-8
8 Channels



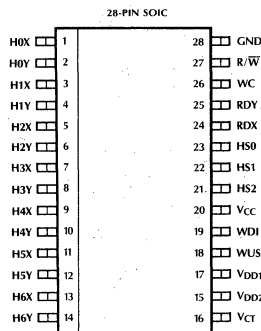
ML511-8 OR ML511R-8
8 Channels



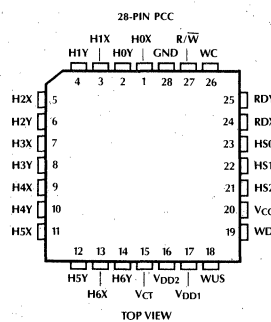
ML511-6 OR ML511R-6
6 Channels



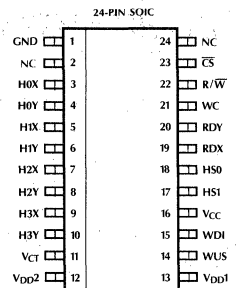
ML511R-7CS
28-Lead SOIC



ML511R-7CQ
28-Lead PCC



ML511-4 OR ML511R-4
4 Channels



PIN DESCRIPTION

NAME	FUNCTION
H50–H52	Head Select (eight heads)
CS	Chip Select (low level enables chip)
R/W	Read/Write (high level selects Read mode)
WUS	Write Unsafe, open collector output (high level indicates an unsafe writing condition)
WDI	Write Data In (negative transition toggles head current direction)
H0X–H7X	X head connections
H0Y–H7Y	Y head connections

NAME	FUNCTION
RDX, RDY	X, Y Read Data (differential read signal out)
WC	Write Current (used to set the write current magnitude)
VCT	Voltage Center Tap (center tap voltage source)
VCC	+5 volts
VDD1	+12 volts
VDD2	Positive supply for center tap
GND	Ground

ABSOLUTE MAXIMUM RATINGS

(Note 1)

Power Supply Voltage Range	
V _{DD1}	-0.3 to 14V _{DC}
V _{DD2}	-0.3 to 14V _{DC}
V _{CC}	-0.3 to 6V _{DC}
Input Voltage Range	
Digital Inputs (\overline{CS} , R/ \overline{W} , HS, WDI)	-0.3 to V _{CC} +0.3V _{DC}
Head Ports (H0X-H7X, H0Y-H7Y)	-0.3 to V _{DD1} +0.3V _{DC}
Write Unsafe (WUS)	-0.3 to 14V _{DC}
Write Current (I _W)	60 mA
Output Current	
Read Data (RDX, RDY)	-10 mA
Center Tap Current (I _{CT})	-60 mA
Write Unsafe (WUS)	12 mA
Storage Temperature	-65°C to 150°C
Junction Temperature (T _J)	135°C
Lead Temperature (Soldering 10 sec.)	300°C

OPERATING CONDITIONS

Supply Voltage	
V _{DD1}	12V ± 10%
V _{CC}	5V ± 10%
Head Inductance	
L _H , ML511 or ML511R	5 to 15 μH
Damping Resistor (R _D , ML511 only)	500 to 2000 Ω
RCT Resistor (1/4 Watt)	120 Ω ± 5%
Write Current (I _W)	10 to 40 mA

ELECTRICAL CHARACTERISTICS

Unless otherwise specified V_{DD1} = V_{DD2} = 12V ± 10%, V_{CC} = 5V ± 10%, R_{CT} = 120 Ω ± 5%, I_W = 40 mA, 0°C ≤ T_A ≤ 70°C (Notes 2 and 3).

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
DC OPERATING CHARACTERISTICS						
POWER SUPPLY						
I _{CC}	V _{CC} Supply Current	Read or Idle Mode			35	mA
		Write Mode			30	mA
I _{DD}	V _{DD} Supply Current	Read Mode			35	mA
		Write Mode			20 + I _W	mA
		Idle Mode			20	mA
P _D	Power Dissipation	Read Mode			655	mW
		Write Mode I _W = 40 mA, R _{CT} = 0 Ω			960	mW
		Idle Mode			455	mW
DIGITAL INPUTS (\overline{CS}, R/\overline{W}, HS, WDI)						
V _{IH}	High Voltage		2			V _{DC}
V _{IL}	Low Voltage				0.8	V _{DC}
I _{IH}	High Current	V _{IH} = 2.0V			100	μA
I _{IL}	Low Current	V _{IL} = 0.8V	-0.4			mA
WUS OUTPUT						
V _{OL}	Output Low Voltage	I _{OL} = 8 mA (Safe)			0.5	V _{DC}
I _{OH}	Output High Current	V _{OH} = 5V (Unsafe)			100	μA
CENTER TAP VOLTAGES						
V _{CT}	Read Mode	Read Mode		4		V _{DC}
V _{CT}	Write Mode	Write Mode		6		V _{DC}

ML511, ML511R

ELECTRICAL CHARACTERISTICS (Continued)

Unless otherwise specified $V_{DD1} = 12V \pm 10\%$, $V_{CC} = 5V \pm 10\%$, $R_{CT} = 120\Omega \pm 5\%$, $I_W = 35\text{ mA}$, $L_H = 10\mu\text{H}$, $R_D = 750\Omega$ (ML511), $f_{DATA} = 5\text{ MHz}$, C_L (RDX, RDY) $\leq 20\text{ pF}$, $0^\circ\text{C} \leq T_A \leq 70^\circ\text{C}$ (Notes 2 and 3) (V_{IN} is referenced to V_{CT} for Read Mode Characteristics).

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
WRITE MODE CHARACTERISTICS						
I_{HCW}	Head Current (per side)	Write Mode $0 \leq V_{CC} \leq 3.7V$ $0 \leq V_{DD1} \leq 8.7V$	-200		200	μA
I_{WR}	Write Current Range	$I_W = K/R_{WC}$	10		40	mA
K	Write Current Constant		2.375		2.625	
V_{HD}	Differential Head Voltage Swing		7.0			V_{PK}
I_{HU}	Unselected Head Transient Current				2	mA_{PK}
C_{OD}	Differential Output Capacitance				15	pF
R_{OD}	Differential Output Resistance	ML511	10k			Ω
		$T_J = 25^\circ\text{C}$ ML511R	600		960	Ω
f_{WDI}	WDI Transition Frequency	WUS = Low	250			kHz
A_I	I_{WC} to Head Current Gain			0.99		mA/mA
I_L	Unselected Head Leakage	Sum of X & Y Side Leakage Current			85	μA
READ MODE CHARACTERISTICS						
A_V	Differential Voltage Gain	$V_{IN} = 1\text{ mV}_{P.P.}$ @ 300kHz, R_L (RDX, RDY) = $1\text{ k}\Omega$	85		115	V/V
DR	Dynamic Range	DC Input Voltage (V_I) Where Gain Falls 10%, $V_{IN} = V_I + 0.5\text{ mV}_{P.P.}$ @ 300kHz	-3		+3	mV
BW	Bandwidth (-3 dB)	$ Z_S < 5\Omega$, $V_{IN} = 1\text{ mV}_{P.P.}$	30			MHz
e_{IN}	Input Noise Voltage	$BW = 15\text{ MHz}$, $L_H = 0$, $R_H = 0$			1.5	nV/ $\sqrt{\text{Hz}}$
C_{IN}	Differential Input Capacitance	$f = 5\text{ MHz}$			20	pF
R_{IN}	Differential Input Resistance	$f = 5\text{ MHz}$, $T_J = 25^\circ\text{C}$ ML511	2k			Ω
		$V_{IN} = 6\text{ mV}_{P.P.}$ ML511R	460		860	Ω
I_{HCR}	Head Current (per side)	Read or Idle Mode $0 \leq V_{CC} \leq 5.5V$ $0 \leq V_{DD1} \leq 13.2V$	-200		200	μA
I_{IN}	Input Bias Current (1 side)				45	μA
CMRR	Common-Mode Rejection Ratio	$V_{CM} = V_{CT} + 100\text{ mV}_{P.P.}$ @ $f = 5\text{ MHz}$	50			dB
PSRR	Power Supply Rejection Ratio	$100\text{ mV}_{P.P.}$ @ 5 MHz on V_{DD1} , V_{DD2} , or V_{CC}	45			dB
CS	Channel Separation	Unselected Channels: $V_{IN} = 100\text{ mV}_{P.P.}$ @ 5 MHz and Selected Channel: $V_{IN} = 0\text{ mV}_{P.P.}$	45			dB
V_{OS}	Output Offset Voltage	Read Mode	-460		+460	mV
		Write or Idle Mode	-20		+20	mV
V_{OCM}	Common-Mode Output Voltage	Read Mode	4.5		6.5	V
		Write or Idle Mode		5.3		V
R_{OUT}	Single-Ended Output Resistance	$f = 5\text{ MHz}$			30	Ω
I_L	Leakage Current, RDX, RDY	(RDX, RDY) = $6V$ Write or Idle Mode	-100		100	μA
I_O	Output Current	AC Coupled Load, RDX to RDY	± 2.1			mA

ELECTRICAL CHARACTERISTICS (Continued)

Unless otherwise specified $V_{DD1} = 12V \pm 10\%$, $V_{CC} = 5V \pm 10\%$, $R_{CT} = 120\Omega \pm 5\%$, $I_W = 35\text{ mA}$, $L_H = 10\mu\text{H}$, $R_D = 750\Omega$ (ML511), $f_{\text{DATA}} = 5\text{ MHz}$, $0^\circ\text{C} \leq T_A \leq 70^\circ\text{C}$ (Notes 2 and 3).

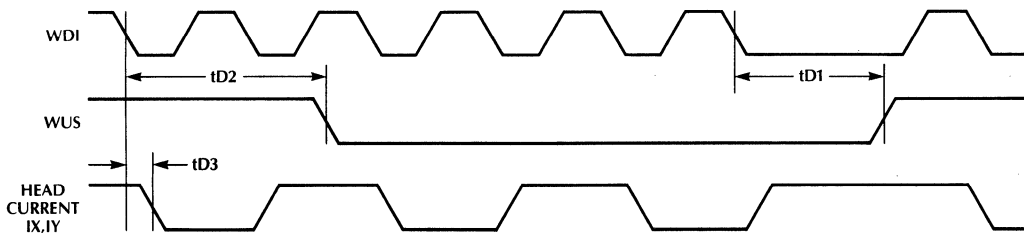
SYMBOL	PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
SWITCHING CHARACTERISTICS						
t_{RW}	R/ \bar{W} to Write Switching Delay	To 90% of Write Current Output			1	μs
t_{WR}	R/ \bar{W} to Read Switching Delay	To 90% of 100mV, 10MHz Read Signal Envelope or to 90% Decay of Write Current			1	μs
t_{IW} or t_{IR}	\bar{CS} to Select Switching Delay	To 90% of Write Current or to 90% of 100mV, 10MHz Read Signal Envelope			1	μs
t_{WI} or t_{RI}	\bar{CS} to Unselect Switching Delay	To 90% Decay of 100mV, 10MHz Read Signal Envelope or to 90% Decay of Write Current			1	μs
t_{HS}	Head Select Switching Delay	To 90% of 100mV, 10MHz Read Signal Envelope			1	μs
t_{D1}	Safe to Unsafe Write Unsafe Delay	$I_W = 35\text{ mA}$	1.6		8	us
t_{D2}	Unsafe to Safe Write Unsafe Delay	$I_W = 35\text{ mA}$			1	us
t_{D3}	Prop. Delay Head Current	$L_H = 0$, $R_H = 0$ From 50% points			25	ns
	Asymmetry Head Current	WDI has 50% Duty Cycle and 1nS Rise/Fall Time			2	ns
	Rise/Fall Head Current	10% and 90% Points			20	ns

Note 1: Absolute maximum ratings are limits beyond which the life of the integrated circuit may be impaired. All voltages unless otherwise specified are measured with respect to ground.

Note 2: Limits are guaranteed by 100% testing, sampling, or correlation with worst-case test conditions.

Note 3: Maximum junction temperature (T_j) should not exceed 135°C .

5

TIMING DIAGRAM

Write Mode Timing Diagram

FUNCTIONAL DESCRIPTION

CIRCUIT OPERATION

For any selected head, the ML511 functions as a read amplifier when in the Read mode, or as a write current switch when in the Write mode. Pins HS0, HS1 and HS2 determine head selection while pin R/W controls the Read/Write mode. A detected "write-unsafe" condition is indicated by pin WUS.

READ MODE

When the ML511 is in the Read Mode, it operates as a low-noise differential amplifier on the selected channel. In Read mode the write data flip-flop is set and both the write unsafe detector and the write current source are deactivated. The center tap voltage is also lowered. Pins RDX and RDY provide differential emitter follower outputs which are in phase with the X and Y head input pins.

Note that during the Read or Chip Deselect mode the internal write current is deactivated, thus making external write current gating unnecessary.

WRITE MODE

The ML511 operates as a write-current switch when in the Write mode. Write current magnitude is determined by the following relationship:

$$I_W = K/R_{WC}$$

Where: K = Write Current Constant

R_{WC} = Resistance connected between pin WC and GND.

The head current is toggled between the X and Y side of the selected head by a negative transition on WDI (Write Data Input). When switching the ML511 to write mode, the Wdff (Write Data Flip-Flop) is initialized to pass write current through the X-side of the head.

The ML511, ML511R exhibit enhanced write current stability, compared to similar read/write circuits, which reduces the problem of oscillation. This is a result of increased internal write current compensation. Also, write current "glitches" during power-up, common in similar read/write circuits, are eliminated with an exclusive write current disabling function.

The WUS (Write Unsafe) pin is an open collector output that gives a logic high level for any of the following unsafe write conditions:

- Open head
- Open head center-tap
- Too low WDI frequency
- Read mode selected
- Device not selected
- No write current

Two negative transitions on WDI are required to clear WUS after the fault condition is removed.

The ML511 also offers a voltage fault detection circuit that prevents write current during power-loss or power-up.

Table 1.

Head Select			
HS2	HS1	HS0	HEAD
0	0	0	0
0	0	1	1
0	1	0	2
0	1	1	3
1	0	0	4
1	0	1	5
1	1	0	6
1	1	1	7

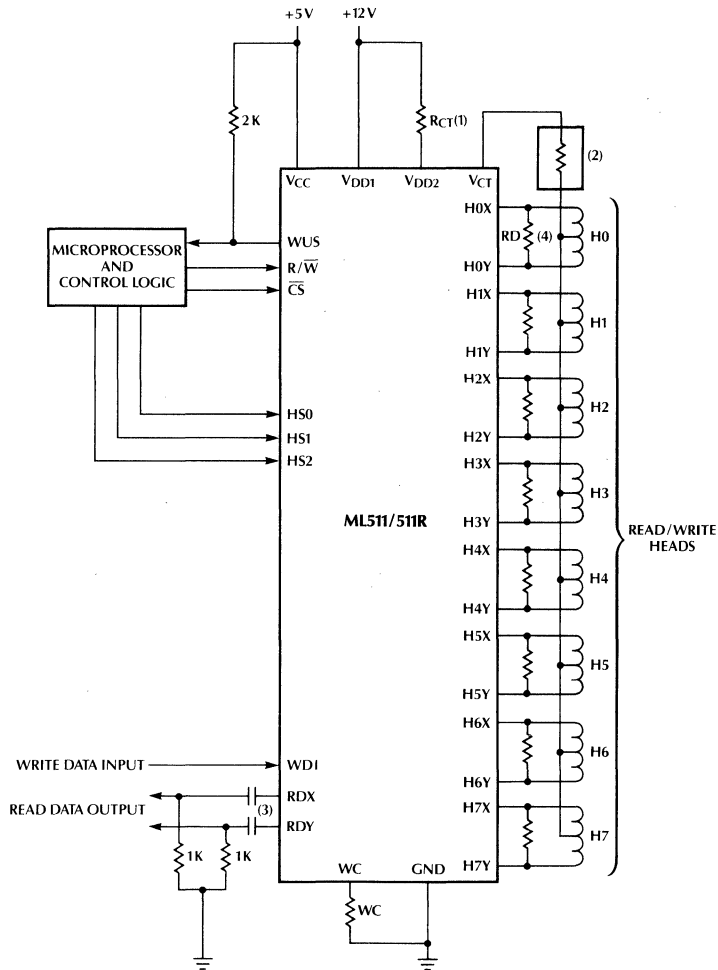
0 = Logic Level Low
 1 = Logic Level High
 X = Don't Care

Table 2.

Mode Select		
\overline{CS}	R/W	MODE
0	0	Write
0	1	Read
1	X	Idle

0 = Logic Level Low
 1 = Logic Level High
 X = Don't Care

TYPICAL APPLICATION



NOTES:

1. RCT is optional and is used to limit internal power dissipation (Otherwise connect V_{DD1} to V_{DD2}).
 $RCT (1/2 \text{ Watt}) = 120 (40/I_w)$ ohms
 where I_w = Write Current, in mA
2. Ferrite bead optional: used to suppress write current overshoot and ringing. Recommend Ferroxcube 3659065/4A6.
3. RDX and RDY load capacitance 20 pF maximum. RDX and RDY output current must be limited to 100 μ A.
4. Damping resistors not required on ML511R.

ML511, ML511R

ORDERING INFORMATION

PART NUMBER	PACKAGE	NUMBER OF CHANNELS
ML511-4CS	24-Lead SOIC (S24)	4
ML511R-4CS	24-Lead SOIC (S24)	4
ML511-6CP	28-Lead Molded DIP (P28)	6
ML511R-6CP	28-Lead Molded DIP (P28)	6
ML511-6CQ	28-Lead PCC (Q28)	6
ML511R-6CQ	28-Lead PCC (Q28)	6
ML511-6CS	28-Lead SOIC (S28)	6
ML511R-6CS	28-Lead SOIC (S28)	6
ML511R-7CS	28-Lead SOIC (S28)	7
ML511R-7CQ	28-Lead PCC (Q28)	7
ML511-8CP	40-Lead Molded DIP (P40)	8
ML511R-8CP	40-Lead Molded DIP (P40)	8
ML511-8CQ	44-Lead PCC (Q44)	8
ML511R-8CQ	44-Lead PCC (Q44)	8
ML511-8CS	32-Lead SOIC (S32)	8
ML511R-8CS	32-Lead SOIC (S32)	8

THERMAL CHARACTERISTICS

PIN COUNT	PACKAGE	θ_{ja}
24-Lead	SOIC	75°C/W
28-Lead	PDIP	55°C/W
28-Lead	PCC	65°C/W
28-Lead	SOIC	70°C/W
32-Lead	SOIC	60°C/W
44-Lead	PCC	60°C/W
40-Lead	PDIP	45°C/W

Read Data Processor

GENERAL DESCRIPTION

The ML541 is a monolithic bipolar integrated circuit for use in a disk drive system to detect analog pulse peaks generated by the recording head during a Read operation. Connected to the read/write amplifier output, it detects valid data and provides a TTL output to the data separator for further processing. It contains both analog and digital circuitry and supports the reading of MFM and RLL encoded data at rates up to 15 megabits/second.

The primary functional blocks within the device include an AGC amplifier, a level detector, a slope detector, and output logic. Operating modes Read, Write, and Hold are selectable with input logic signals. Read mode is used for pulse peak detection during a Read operation. Write mode disables the device's output during a Write operation, while Hold mode holds the AGC gain constant during recovery of embedded servo information.

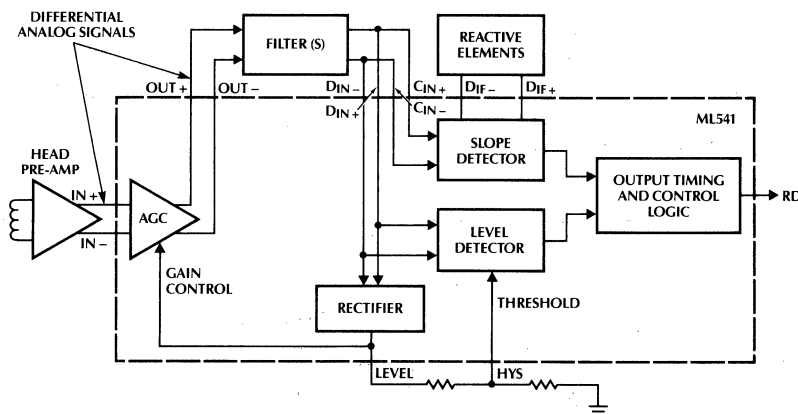
By using both level and slope detection, accurate pulse validation and peak time detection is achieved. The ML541 performance can be adjusted to fit particular needs through external component selection.

The ML541 is available both in a 24-pin PDIP and 28-pin PCC.

FEATURES

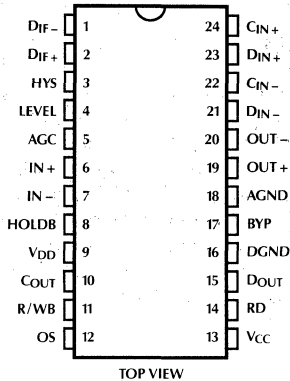
- Second source for SSI 541
- Data rates up to 15 megabits/second
- Supports MFM and RLL encoded read data
- 25MHz wide-bandwidth AGC amplifier
- Fast AGC region for fast transient recover
- Slow AGC region for minimum zero crossing distortion
- Write to read transient suppression
- Supports embedded servo decoding
- +5V, +12V power supplies

SIMPLIFIED BLOCK DIAGRAM

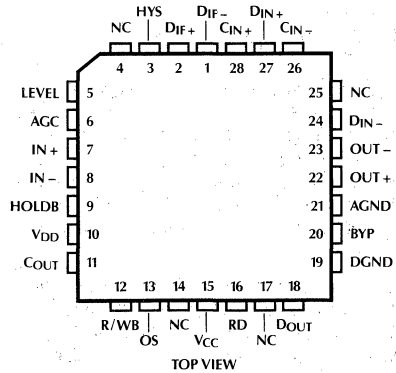


PIN CONNECTIONS

24-Pin DIP and SOIC Package



28-Pin PCC Package



PIN DESCRIPTION

NAME	FUNCTION	NAME	FUNCTION
VCC	+5V	HYS	Input for setting hysteresis level of the hysteresis comparator.
VDD	+12V	LEVEL	Provides rectified signal level for input to the hysteresis comparator.
AGND	Analog Ground.	DOUT	Buffered test point for monitoring D input of the flip-flop.
DGND	Digital Ground.	CIN+, CIN-	Analog input to the differentiator.
R/WB	TTL compatible Read/Write Control pin.	DIF+, DIF-	External differentiating network connection pins.
IN+, IN-	Analog Signal Input pins	COUT	Buffered test point for monitoring the clock input to the flip-flop.
OUT+, OUT-	AGC Amplifier Output pins	OS	Connection for read output pulse width setting capacitor COS.
BYP	The AGC timing capacitor CAGC is tied between this pin and AGND.	RD	TTL compatible read output.
HOLDB	TTL compatible pin that holds the AGC gain when pulled low.		
AGC	Reference input voltage level for the AGC circuit.		
DIN+, DIN-	Analog input to the hysteresis comparator.		

TABLE 1 MODE SELECT

R/WB	HOLDB	MODE	DESCRIPTION
1	1	READ	AGC amp section active, Digital section active.
1	0	HOLD	AGC gain constant, Digital section active.
0	X	WRITE	AGC gain maximum, Digital section inactive, Input common mode resistance reduced.

0 = Logic level low
 1 = Logic level high
 X = Don't care

ABSOLUTE MAXIMUM RATINGS

(Note 1)

Power Supply Voltage Range	
V_{CC}	-0.3 to 6V _{DC}
V_{DD}	-0.3 to 14V _{DC}
Terminal Voltage Range	
R/WB, IN+, IN-, HOLDB	-0.3V to $V_{CC} + 0.3V$
RD	-0.3V to $V_{CC} + 0.3V$ or +12mA
All others	-0.3V to $V_{DD} + 0.3V$
Storage Temperature Range	-65°C to +150°C
Junction Temperature (T_J)	+135°C
Lead Temperature (Soldering, 10sec)	260°C

OPERATING CONDITIONS

Supply Voltage	
V_{CC}	5V \pm 10%
V_{DD}	12V \pm 10%
$V_{(CIN+ - CIN-)}, V_{(DIN+ - DIN-)}$	1V _{P-P}
V_{HYS}	1.0V
C_{OS}	50 to 200pF
Typical Component Values (Refer to Typical Applications)	
C_{IN}	0.001 μ F
C_S	0.01 μ F
C_{OUT}	0.0047 μ F
R_{OUT}	400 Ω
C_{AGC1}	220pF
C_{AGC2}	2000pF
R_{AGC}	2.21k Ω
C_{LEVEL}	150pF
R_{LEVEL1}	1.54k Ω
R_{LEVEL2}	6.49k Ω
C_{OS}	50pF

ELECTRICAL CHARACTERISTICS

The following specifications apply over the recommended operating conditions of $V_{CC} = 5V \pm 10\%$, $V_{DD} = 12V \pm 10\%$, $0^\circ C < T_A < 70^\circ C$ and external components as specified under recommended operating conditions unless otherwise specified. (See Note 2.)

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP NOTE 5	MAX	UNITS
DC Characteristics						
I_{CC}	V_{CC} Supply Current	Outputs unloaded			14	mA
I_{DD}	V_{DD} Supply Current	Outputs unloaded			70	mA
P_D	Power Dissipation	Outputs unloaded, $T_A = 70^\circ C$			930	mW
Digital Inputs Characteristics (HOLDB, R/WB)						
V_{IH}	High Voltage		2			V
V_{IL}	Low Voltage				0.8	V
I_{IH}	High Current	$V_{IH} = 2.4V$			100	μA
I_{IL}	Low Current	$V_{IL} = 0.4V$	-0.4			mA
Digital Outputs Characteristics (C_{OUT}, RD)						
V_{OL}	Output Low Voltage	$I_{OL} = 4mA$			0.4	V
V_{OH}	Output High Voltage	$I_{OH} = 400\mu A$	2.4			V
WRITE AND HOLD MODE CHARACTERISTICS						
Mode Control						
t_{RW}	Read to Write Transition Time				1	μs
t_{WR}	Write to Read Transition Time	AGC settling not included, time to high input resistance	1.2		3	μs
t_{RH}	Read to Hold Transition Time				1	μs
Write Mode						
Z_{IC}	Common Mode Input Impedance (both sides)	R/WB pin = low		250		Ω

ELECTRICAL CHARACTERISTICS (Continued)

The following specifications apply over the recommended operating conditions of $V_{CC} = 5V \pm 10\%$, $V_{DD} = 12V \pm 10\%$, $0^\circ C \leq T_A \leq 70^\circ C$, IN+ and IN- AC coupled, OUT+ and OUT- differentially loaded with $>600\Omega$ and each side loaded with $<10pF$ to GND, $C_{BYP} = 2000pF$, OUT+ and OUT- AC coupled to D_{IN+} and D_{IN-} respectively, $V_{AGC} = 2.2V$ unless otherwise specified. (See Note 2.)

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP NOTE 5	MAX	UNITS
READ MODE CHARACTERISTICS						
AGC Amplifier						
R_{ID}	Differential Input Resistance	$V_{(IN+ - IN-)} = 100mV_{P,P}$ @ 2.5 MHz		5		k Ω
C_{ID}	Differential Input Capacitance	$V_{(IN+ - IN-)} = 100mV_{P,P}$ @ 2.5 MHz			10	pF
Z_{IC}	Common Mode Input Impedance (both sides)	R/WB pin high		1.8		k Ω
		R/WB pin low		0.25		k Ω
A_{VR}	Gain Range	$1V_{P,P} \leq V_{OUT\ diff} < 2.5V_{P,P}$	4		83	V/V
e_N	Input Noise Voltage	Gain set to maximum			30	nV/ \sqrt{Hz}
BW	Bandwidth	Gain set to maximum, -3 dB point	25			MHz
V_{OP}	Maximum Output Voltage Swing	Set by V_{AGC}	3			$V_{P,P}$
I_{OD}	OUT+ to OUT- Pin Current	No DC path to GND, See Note 3	± 3.2			mA
R_O	Output Resistance			20	30	Ω
C_O	Output Capacitance			12		pF
V_{IP} V_{AGC}	$(D_{IN+} - D_{IN-})$ Input Voltage Swing VS AGC Input Level	$30mV_{P,P} \leq V_{(IN+ - IN-)} \leq 550mV_{P,P}$ $1.5V \leq V_{AGC} \leq 3.75V$		0.48		$V_{P,P}/V$
V_{IP}	$(D_{IN+} - D_{IN-})$ Input Voltage Swing Variation	$30mV_{P,P} < V_{(IN+ - IN-)} < 550mV_{P,P}$ AGC Fixed, over supply and temp.			+8	%
t_D	Gain Decay Time	See Figure 1a; $V_{IN} = 300mV_{P,P}$ then $>150mV_{P,P}$ at 2.5 MHz, V_{OUT} to 90% of final value.		50		μs
t_A	Gain Attack Time	See Figure 1b; from Write to Read transition to V_{OUT} at 110% of final value, $V_{IN} = 400mV_{P,P}$ @ 2.5 MHz		4		μs
I_{AGCfc}	Fast AGC Capacitor Charge Current	$V_{(D_{IN+} - D_{IN-})} = 1.6V$, $V_{AGC} = 3.0V$		1.5		mA
I_{AGCsc}	Slow AGC Capacitor Charge Current	$V_{(D_{IN+} - D_{IN-})} = 1.6V$, Vary V_{AGC} until slow discharge begins		0.17		mA
	Fast to Slow Attack Switchover Point	$V_{(D_{IN+} - D_{IN-})}$ $V_{(D_{IN+} - D_{IN-})}$ Final		1.25		-
I_{AGCD}	AGC Capacitor Discharge Current	$V_{(D_{IN+} - D_{IN-})} = 0.0V$ Read Mode		4.5		μA
		Hold Mode		-0.2	+0.2	μA
CMRR	CMRR (Input Referred)	$V_{IN+} = V_{IN-} = 100mV_{P,P}$ @ 5MHz, gain at max.	40			dB
PSRR	PSRR (Input Referred)	V_{CC} or $V_{DD} = 100mV_{P,P}$ @ 5MHz, gain at max.	30			dB
Hysteresis Comparator						
V_{IP}	Input Signal Range				1.5	$V_{P,P}$
R_{ID}	Differential Input Resistance	$V_{(D_{IN+} - D_{IN-})} = 100mV_{P,P}$ @ 2.5 MHz	5		15	k Ω
C_{ID}	Differential Input Capacitance	$V_{(D_{IN+} - D_{IN-})} = 100mV_{P,P}$ @ 2.5 MHz			6.0	pF
Z_{IC}	Common Mode Input Impedance	(both sides)		2.0		k Ω
V_{IO}	Comparator Offset Voltage	HYS pin at -0.5V, $\leq 1.5k\Omega$ across D_{IN+} , D_{IN-}		5		mV

ELECTRICAL CHARACTERISTICS (Continued)

The following specifications apply over the recommended operating conditions of $V_{CC} = 5V \pm 10\%$, $V_{DD} = 12V \pm 10\%$, $0^\circ C \leq T_A \leq 70^\circ C$, IN+ and IN- AC coupled, OUT+ and OUT- differentially loaded with $>600\Omega$ and each side loaded with $<10pF$ to GND, $C_{BYP} = 2000pF$, OUT+ and OUT- AC coupled to D_{IN+} and D_{IN-} respectively, $V_{ACC} = 2.2V$ unless otherwise specified.

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
READ MODE CHARACTERISTICS (Continued)						
Hysteresis Comparator (Continued)						
V_{HYS} V_{HYS}	Peak Hysteresis Voltage vs HYS pin voltage (input referred)	$1V < V_{HYS} < 3V$		0.21		V/V
I_I	HYS Pin Input Current	$1V < V_{HYS} < 3V$	0		-20	μA
I_O	LEVEL Pin Max Output Current		3			mA
R_O	LEVEL Pin Output Resistance	$I_{LEVEL} = 0.5mA$		180		Ω
V_{OL}	D_{OUT} Pin Output Low Voltage	$T_A = 70^\circ C$	$V_{DD} - 4.0$		$V_{DD} - 2.5$	V
V_{OH}	D_{OUT} Pin Output High Voltage	$T_A = 70^\circ C$	$V_{DD} - 2.2$		$V_{DD} - 1.5$	V
V_{OL}	D_{OUT} Pin Output Low Voltage	$T_A = 25^\circ C$	$V_{DD} - 4.0$		$V_{DD} - 2.8$	V
V_{OH}	D_{OUT} Pin Output High Voltage	$T_A = 25^\circ C$	$V_{DD} - 2.5$		$V_{DD} - 1.6$	V
Active Differentiator						
V_{IP}	Input Signal Range				1.5	$V_{P,P}$
R_{ID}	Differential Input Resistance	$V_{(C_{IN+} - C_{IN-})} = 100mV_{P,P} @ 2.5MHz$	5		15	k Ω
C_{ID}	Differential Input Capacitance	$V_{(C_{IN+} - C_{IN-})} = 100mV_{P,P} @ 2.5MHz$			6	pF
Z_{IC}	Common Mode Input Impedance	(both sides)		2.0		k Ω
I_{OD}	D_{IF+} to D_{IF-} Pin Current	Differentiator Imped must be set so as not to clip signal at this current level	± 1.3			mA
V_{IO}	Comparator Offset Voltage	D_{IF+} , D_{IF-} AC Coupled		5		mV
V_{OL}	C_{OUT} Pin Output Low Voltage	$0 \leq I_{OH} \leq 0.5mA$		$V_{DD} - 3$		V
V_{PO}	C_{OUT} Pin Output Pulse Voltage	$0 \leq I_{OH} \leq 0.5mA$		0.4		V
PW_0	C_{OUT} Pin Output Pulse Width	$0 \leq I_{OH} \leq 0.5mA$		30		ns

The following specifications apply over the recommended operating conditions of $V_{CC} = 5V \pm 10\%$, $V_{DD} = 12V \pm 10\%$, $0^\circ C \leq T_A \leq 70^\circ C$, $V_{(C_{IN+} - C_{IN-})} = V_{(D_{IN+} - D_{IN-})} = 1.0V_{P,P}$ AC coupled sine wave at 2.5MHz, $R_{DIF} = 100\Omega$, $C_{DIF} = 65pF$, $V_{HYS} = 1.8V$, $C_{OS} = 60pF$, $4k\Omega$ to V_{CC} and $10pF$ to GND on pin RD unless otherwise specified.

Output Data Characteristics (Refer to Figure 2)

t_{D1}	D-Flip-Flop Set Up Time	Min delay from $V_{(D_{IN+} - D_{IN-})}$ exceeding threshold to $V_{(D_{IF+} - D_{IF-})}$ reaching a peak	0			ns
t_{D3}	Propagation Delay				110	ns
t_{D5}	Output Data Pulse Width	$T_A = 25^\circ C$, $V_{CC} = 5V$, $V_{DD} = 12V$		$\pm 15\%$		
t_{D5}	Output Data Pulse Width Variation	$C_{OS} = 60pF$, See Note 4	30		80	ns
$t_{D3} - t_{D4}$	Logic Skew (Pulse Pairing)				3	ns
t_R	Output Rise Time	$V_{OH} = 2.4V$			18	ns
t_F	Output Fall Time	$V_{OL} = 0.4V$			14	ns

Note 1: Absolute maximum ratings are limits beyond which the life of the integrated circuit may be impaired. All voltages unless otherwise specified are measured with respect to ground.

Note 2: Limits are guaranteed by 100% testing, sampling, or correlation with worst-case test conditions.

Note 3: AGC amplifier output current may be increased as in Figure 4.

Note 4: $t_{D5} \cong 770 (C_{OS})$, $50pF < C_{OS} < 150pF$.

Note 5: Typicals are parametric norm at $25^\circ C$

FUNCTIONAL DESCRIPTION

Operating Modes

The ML541 has three definitive operation modes which are: Read mode, Write mode and Hold mode. These modes are defined by input pins HOLDB and R/WB as shown in Table 1. Read mode, the mode used normally for pulse detection, is assumed in the following sections unless otherwise noted.

AGC Amplifier Section

The purpose of the AGC amplifier is to provide a constant read signal level for both the level and slope detectors. Full differential processing of the read signal is used to minimize noise and distortion in the analog signal. A wide gain range is required due to large signal variation when moving the recording head from an inside to outside data track or variations in media.

The differential output voltage level V_{OUT} from the AGC amp is determined by voltage V_{AGC} present at pin AGC. V_{OUT} is full wave rectified and compared against V_{AGC} to create charge/discharge current for capacitor C_{BYP} connected at pin BYP. Voltage V_{BYP} across C_{BYP} controls the gain in the AGC amplifier.

Two distinct values of I_{BYP} are possible which determine a fast and slow AGC gain response attack rate. When V_{OUT} is more than 125% of the set level a high value of I_{BYP} is sourced which provides a fast AGC attack rate. When V_{OUT} is within 100% to 125% of the set level a reduced value of I_{BYP} is sourced which provides a slower attack rate. The fast-slow gain response attack rates provides for an initial quick system response and then minimum zero crossing distortion of the analog signal once the gain is within working range. V_{AGC} should be set so that the differential input voltage V_{DIN} into the level comparator is $1V_{P-P}$ at nominal Read signal conditions. The AGC amp section gain is given by:

$$\frac{A_{V2}}{A_{V1}} = \exp \frac{V_{BYP2} - V_{BYP1}}{5.8 \times V_T}$$

Where: A_{V1} , A_{V2} are initial and final amplifier gain values corresponding to initial and final V_{BYP} values.

$$V_T = (KT)/Q = 26 \text{ mV at room temperature.}$$

The AGC amp's differential inputs must be AC coupled to the read amplifier (ML117, ML501, etc.) differential outputs. Similarly, AC coupling must be used at the AGC amp outputs.

AGC Amp During Write Mode—When the ML541 is put into write mode, the AGC amp's input impedance is lowered to allow a faster dampening of the Write to Read transient from the head pre-amp. The AGC gain is also set to maximum gain so that fast AGC attack will occur when changing back to the Read mode. Internal device timing is controlled so that settling occurs prior to Read mode activation. Minimal value input coupling capacitors should be chosen to reduce settling time, however, bandwidth requirements also need to be considered.

AGC Amp During Hold Mode—During the Hold mode, the charge/discharge current driving pin BYP is internally disconnected. AGC compensation capacitor C_{AGC} will then hold the present gain setting. The amplitude of V_{OUT} will therefore not affect the AGC gain and gain will remain constant.

Hold mode is used so that AGC gain will not be adjusted when embedded servo information is read. This prevents losing the pulse peak amplitude information needed during position decoding, or creating additional gain settling time when again reading data. Embedded servo pulses are normally taken at outputs D_{IF-} and D_{IF+} , as shown in the typical application.

External Filter Network

Filtering for the level and slope detectors can be performed with a single filter or two separate filters. If separate filters are used, care must be used to insure that time delays are matched. A multi-pole Bessel filter is recommended due to the group delay and linear phase characteristics.

Level Detector

The full wave rectified V_{OUT} is buffered and available at pin LEVEL. The level detector uses a hysteresis comparator to compare the processed read signal amplitude against a reference voltage derived from voltage V_{LEVEL} output from pin LEVEL. Using V_{LEVEL} provides a feed-forward function that allows valid level detection to be performed prior to AGC amp gain settling. The level detector hysteresis value is set in a way that will only allow relatively large read pulse peaks (negative or positive) to be detected.

Slope Detector

The slope detector uses an external reactive component network to produce a voltage signal proportional to the differential of the read signal. By using a hysteresis comparator to detect zero slope of the read signal, the time occurrence of positive or negative read pulse peak values can be determined.

An external reactive network, shown in the Typical Application, is used between the D_{IF+} and D_{IF-} pins to provide the differential function given by:

$$A_V = \frac{-2000Cs}{LCs^2 + (R + 92)Cs + 1}$$

Where: C = External capacitor (20 pF to 150 pF)

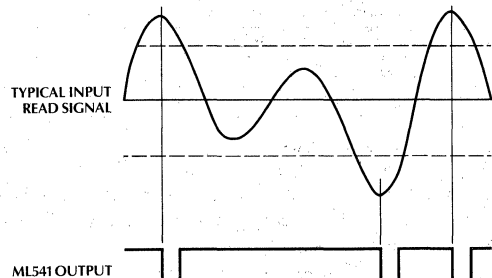
L = External inductor

R = External resistor

$s = j\omega = j2\pi f$

Output Logic

The output logic provides a negative TTL pulse at pin RD which begins at the peak of a valid read pulse, as shown below.

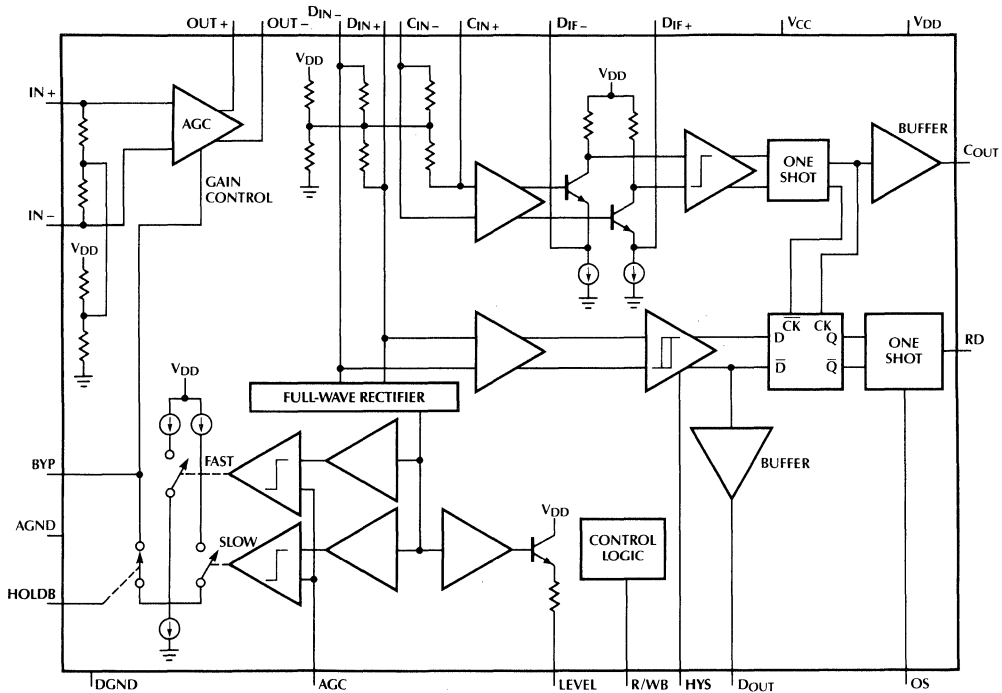


Pin R/WB must be high for the output logic to be active. The key element in the output logic is the D flip-flop. The flip-flop is clocked by the slope detector at the time of a zero crossing, which loads data from level detector. The flip-flop inputs only change state when the level detector detects a peak amplitude of a polarity opposite to the previous valid peak. Thus, through the output logic the slope detector determines output timing and the level detector determines pulse validity.

Layout Considerations

As with any high gain, wide bandwidth analog circuitry, care needs to be exercised in PC layout. Power supply and ground lines should be bypassed and well isolated from other circuitry. A ground plane is recommended, as is keeping analog lines short and well balanced to prevent interaction with nearby circuitry in the disk drive.

BLOCK DIAGRAM



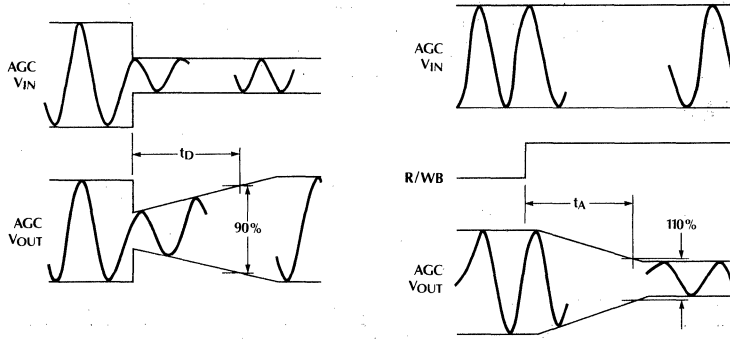


Figure 1. AGC Timing Diagram

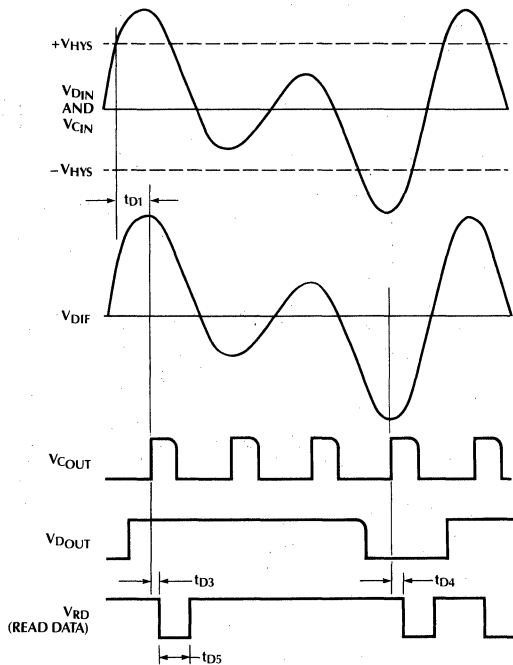


Figure 2. Output Logic Timing Diagram

TYPICAL APPLICATIONS

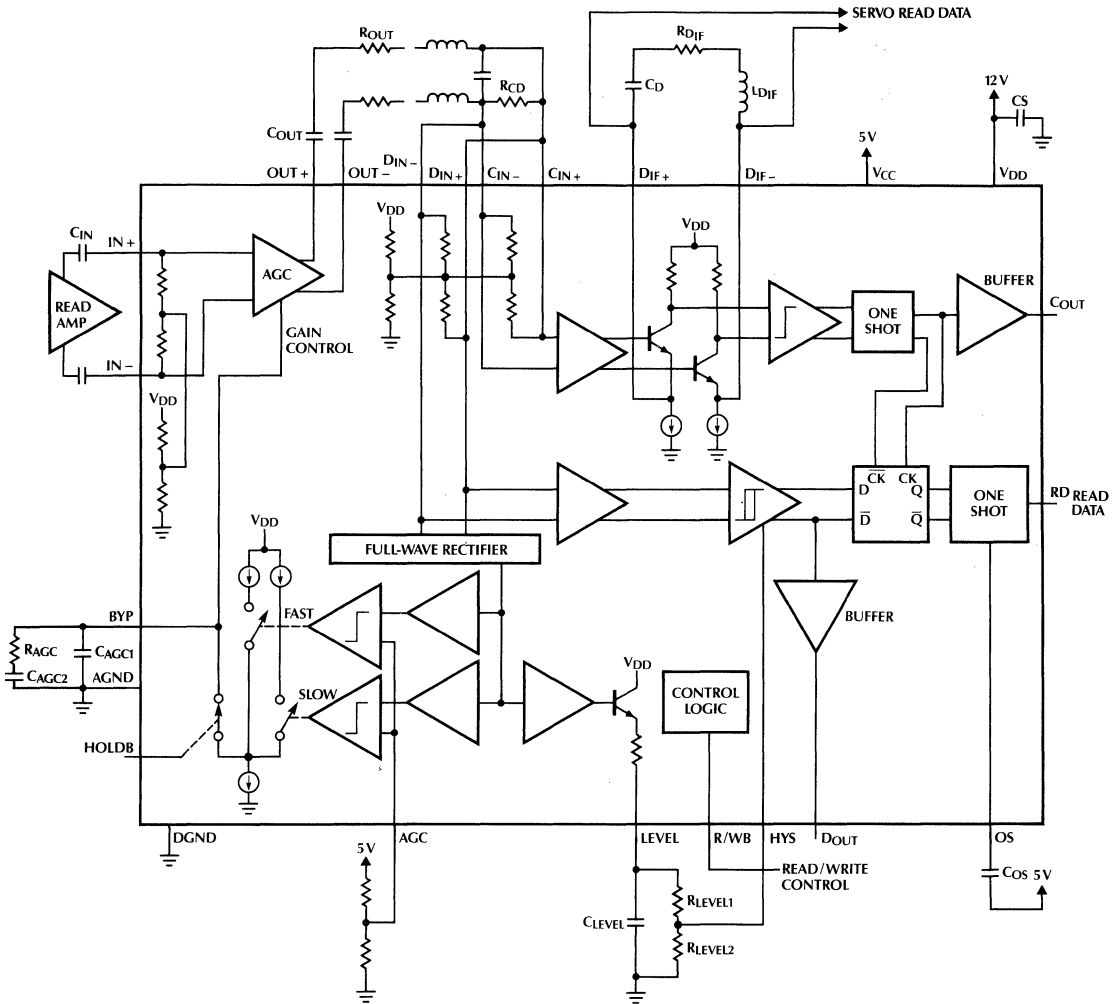


Figure 3. Typical Application Diagram

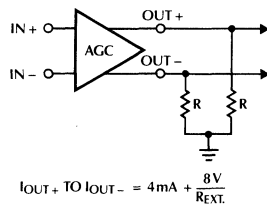


Figure 4. Modification of AGC Amplifier Output to Drive Low Impedance Filters

ML541

ORDERING INFORMATION

PART NUMBER	TEMP. RANGE	PACKAGE
ML541CP	0°C to +70°C	MOLDED DIP (P24)
ML541CJ	0°C to +70°C	HERMETIC DIP (J24)
ML541CQ	0°C to +70°C	MOLDED PCC (Q28)
ML541CS	0°C to +70°C	MOLDED SOIC (S24)

Read Data Processor

GENERAL DESCRIPTION

The ML4041, ML4042 is a monolithic bipolar integrated circuit used in disk drive systems to detect amplitude peaks generated by the recording heads during a Read operation. Connected to the read/write amplifier output, it detects valid data and provides a TTL output to the data separator. Containing both analog and digital circuitry, it supports the reading of MFM and RLL encoded data at rates up to 24 megabits/second.

Operating modes Read, Write, and Hold are selectable with input logic signals. Read mode is used for pulse peak detection during a Read operation. Write mode disables the device's output during a Write operation, while Hold mode holds the AGC gain constant during recovery of embedded servo information.

By using both level and slope detection, accurate pulse validation and peak time detection is achieved. The ML4041, ML4042 characteristics can be modified to fit particular needs through external component selection. The ML4041, ML4042 has a swift Write to Read recovery time of $2\mu\text{s}$ ($10\mu\text{s}$ max) allowing for better format efficiency with faster access times. Pulse pairing of 1ns max reduces data decoding errors by allowing tighter specs for the clock recovery circuit.

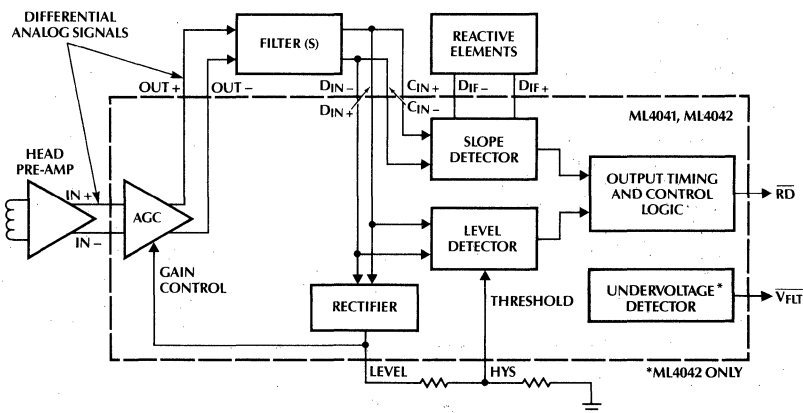
FEATURES

- Fully compatible with industry standard read data processor
- Write to Read recovery time — $2\mu\text{s}$ typical, $10\mu\text{s}$ max
- Pulse pairing — 1ns max
- Data rates up to 24 megabits/second
- Supports MFM and RLL encoded read data
- 30MHz wide-bandwidth AGC amplifier
- Fast AGC region for fast transient recover
- Slow AGC region for minimum zero crossing distortion
- +5V and +12V undervoltage fault detection (ML4042 only)
- Write to read transient suppression
- Hold pin supports embedded servo decoding

The ML4042 is identical to the ML4041 but in addition it includes a +5V and +12V undervoltage detector. The ML4041 is available in a 24-pin PDIP, 24-pin SOIC, or a 28-pin PCC, while the ML4042 is available in a 28-pin PDIP, 28-pin SOIC, or a 28-pin PCC.

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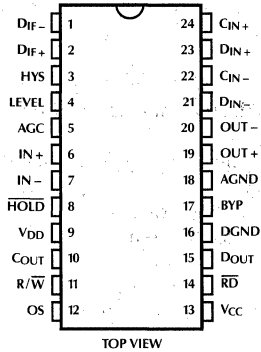
SIMPLIFIED BLOCK DIAGRAM



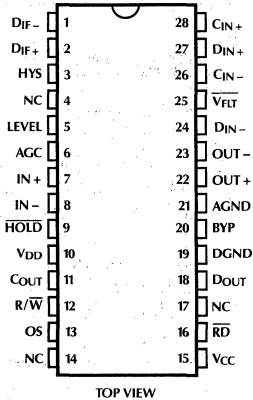
ML4041, ML4042

PIN CONNECTIONS

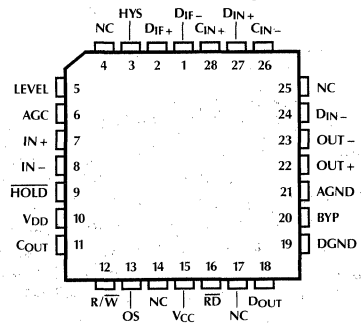
ML4041
24-Pin DIP and SOIC Package



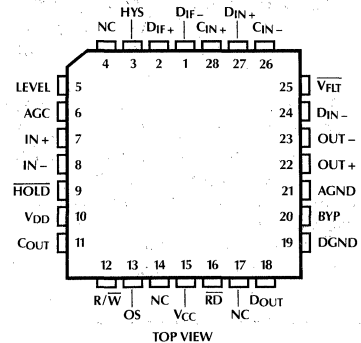
ML4042
28-Pin DIP and SOIC Package



ML4041
28-Pin PCC Package



ML4042
28-Pin PCC Package



PIN DESCRIPTION

NAME	FUNCTION	NAME	FUNCTION
V _{CC}	+5V	HYS	Input for setting hysteresis level of the hysteresis comparator.
V _{DD}	+12V	LEVEL	Provides rectified signal level for input to the hysteresis comparator.
AGND	Analog Ground.	DOUT	Buffered test point for monitoring D input of the flip-flop.
DGND	Digital Ground.	C _{IN+} , C _{IN-}	Analog input to the differentiator.
R/W	TTL compatible Read/Write Control pin.	D _{IF+} , D _{IF-}	External differentiating network connection pins.
IN+, IN-	Analog Signal Input pins	COUT	Buffered test point for monitoring the clock input to the flip-flop.
OUT+, OUT-	AGC Amplifier Output pins	OS	Connection for read output pulse width setting capacitor C _{OS} .
BYP	The AGC timing capacitor C _{AGC} is tied between this pin and AGND.	RD	TTL compatible read output.
HOLD	TTL compatible pin that holds the AGC gain when pulled low.	V _{FLT}	Undervoltage detector output, active low; ML4042 only.
AGC	Reference input voltage level for the AGC circuit.		
D _{IN+} , D _{IN-}	Analog input to the hysteresis comparator.		

TABLE 1 MODE SELECT

R/W	HOLD	MODE	DESCRIPTION
1	1	READ	AGC amp section active, Digital section active.
1	0	HOLD	AGC gain constant, Digital section active.
0	X	WRITE	AGC gain maximum, Digital section inactive, Input common mode resistance reduced.

0 = Logic level low
1 = Logic level high
X = Don't care

ABSOLUTE MAXIMUM RATINGS

(Note 1)

Power Supply Voltage Range	
V _{CC}	-0.3 to 6V _{DC}
V _{DD}	-0.3 to 14V _{DC}
Terminal Voltage Range	
R/W, IN+, IN-, HOLD	-0.3V to V _{CC} +0.3V
RD	-0.3V to V _{CC} +0.3V or +12mA
All others	-0.3V to V _{DD} +0.3V
Storage Temperature Range	-65°C to +150°C
Junction Temperature (T _j)	+135°C
Lead Temperature (Soldering, 10sec)	260°C

OPERATING CONDITIONS

Supply Voltage	
V _{CC}	5V ± 10%
V _{DD}	12V ± 10%
V _{(CIN+ - CIN-), V_(DIN+ - DIN-)}	1V _{P-P}
V _{HYS}	1.0V
C _{OS}	50 to 200pF
Typical Component Values (Refer to Typical Applications)	
C _{IN}	0.001μF
C _S	0.01μF
C _{OUT}	0.0047μF
R _{OUT}	400Ω
C _{AGC1}	220pF
C _{AGC2}	2000pF
R _{AGC}	2.21kΩ
C _{LEVEL}	150pF
R _{LEVEL1}	1.54kΩ
R _{LEVEL2}	6.49kΩ
C _{OS}	50pF

ELECTRICAL CHARACTERISTICS

The following specifications apply over the recommended operating conditions of V_{CC} = 5V ± 10%, V_{DD} = 12V ± 10%, 0°C < T_A < 70°C and external components as specified under operating conditions unless otherwise specified. (See Note 2.)

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP NOTE 4	MAX	UNITS
DC Characteristics						
I _{CC}	V _{CC} Supply Current	Outputs unloaded			14	mA
I _{DD}	V _{DD} Supply Current	Outputs unloaded			70	mA
P _D	Power Dissipation	Outputs unloaded, T _A = 70°C			930	mW
Digital Inputs Characteristics (HOLD, R/W)						
V _{IH}	High Voltage		2			V
V _{IL}	Low Voltage		-0.3		0.8	V
I _{IH}	High Current	V _{IH} = 2.4V			100	μA
I _{IL}	Low Current	V _{IL} = 0.4V	-0.4			mA
Digital Outputs Characteristics (C_{OUT}, RD)						
V _{OL}	Output Low Voltage	I _{OL} = 4mA			0.4	V
V _{OH}	Output High Voltage	I _{OH} = 400μA	2.4			V
WRITE AND HOLD MODE CHARACTERISTICS						
Mode Control						
t _{RW}	Read to Write Transition Time				1	μs
t _{WR}	Write to Read Transition Time	AGC settling not included, time to high input resistance	1.2		3	μs
t _{RH}	Read to Hold Transition Time				1	μs
Write Mode						
Z _{IC}	Common Mode Input Impedance (both sides)	R/W pin = low		250		Ω

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ML4041, ML4042

ELECTRICAL CHARACTERISTICS (Continued)

The following specifications apply over the recommended operating conditions of $V_{CC} = 5V \pm 10\%$, $V_{DD} = 12V \pm 10\%$, $0^\circ\text{C} \leq T_A \leq 70^\circ\text{C}$, $\text{IN}+$ and $\text{IN}-$ AC coupled, $\text{OUT}+$ and $\text{OUT}-$ differentially loaded with $>600\Omega$ and each side loaded with $<10\text{pF}$ to GND, $C_{\text{BYP}} = 2000\text{pF}$, $\text{OUT}+$ and $\text{OUT}-$ AC coupled to $D_{\text{IN}+}$ and $D_{\text{IN}-}$ respectively, $V_{\text{AGC}} = 2.2\text{V}$ unless otherwise specified. (See Note 2.)

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP NOTE 4	MAX	UNITS
READ MODE CHARACTERISTICS						
AGC Amplifier						
R_{ID}	Differential Input Resistance	$V_{(\text{IN}+ - \text{IN}-)} = 100\text{mV}_{\text{P-P}} @ 2.5\text{MHz}$		5		$\text{k}\Omega$
C_{ID}	Differential Input Capacitance	$V_{(\text{IN}+ - \text{IN}-)} = 100\text{mV}_{\text{P-P}} @ 2.5\text{MHz}$			10	pF
Z_{IC}	Common Mode Input Impedance (both sides)	R/\sqrt{W} pin high		1.8		$\text{k}\Omega$
		R/\sqrt{W} pin low		0.25		$\text{k}\Omega$
A_{VMAX}	Maximum Gain	$V_{\text{BYP}} = 2.6\text{V}$	83			V/V
A_{VMIN}	Minimum Gain	$V_{\text{BYP}} = 6\text{V}$	2		4	V/V
e_{N}	Input Noise Voltage	Gain set to maximum			30	$\text{nV}/\sqrt{\text{Hz}}$
BW	Bandwidth	Gain set to maximum, -3dB point	30			MHz
ΔV_{OS}	Maximum Gain and Minimum Gain AGC Amp Output Offset Voltage Difference	$V_{\text{BYP}} = 2.6\text{V}$ for maximum gain $V_{\text{BYP}} = 5.0\text{V}$ for minimum gain			700	mV
V_{BYPMAX}	Max Voltage at BYP Pin at Minimum Gain	$V_{(D_{\text{IN}+} - D_{\text{IN}-})} = 1.6\text{V}$, $V_{\text{AGC}} = 3.0\text{V}$		6.0	6.7	V
V_{OP}	Maximum Output Voltage Swing	Set by V_{AGC}	3			$V_{\text{P-P}}$
I_{OD}	$\text{OUT}+$ to $\text{OUT}-$ Pin Current	No DC path to GND, See Note 3	± 3.2			mA
R_{O}	Output Resistance			18	32	Ω
C_{O}	Output Capacitance			12		pF
V_{IP} V_{AGC}	$(D_{\text{IN}+} - D_{\text{IN}-})$ Input Voltage Swing VS AGC Input Level	$30\text{mV}_{\text{P-P}} \leq V_{(\text{IN}+ - \text{IN}-)} \leq 550\text{mV}_{\text{P-P}}$ $0.5V_{\text{P-P}} \leq V_{(D_{\text{IN}+} - D_{\text{IN}-})} \leq 1.5V_{\text{P-P}}$	0.37	0.48	0.56	$V_{\text{P-P}}/\text{V}$
V_{IP}	$(D_{\text{IN}+} - D_{\text{IN}-})$ Input Voltage Swing Variation	$30\text{mV}_{\text{P-P}} < V_{(\text{IN}+ - \text{IN}-)} < 550\text{mV}_{\text{P-P}}$ AGC Fixed, over supply and temp.			+8	%
t_{D}	Gain Decay Time	See Figure 1a; $V_{\text{IN}} = 300\text{mV}_{\text{P-P}}$ then $>150\text{mV}_{\text{P-P}}$ at 2.5MHz, V_{OUT} to 90% of final value.		50		μs
t_{A}	Gain Attack Time	See Figure 1b; from Write to Read transition to V_{OUT} at 110% of final value, $V_{\text{IN}} = 400\text{mV}_{\text{P-P}} @ 2.5\text{MHz}$		4		μs
I_{AGCfc}	Fast AGC Capacitor Charge Current	$V_{(D_{\text{IN}+} - D_{\text{IN}-})} = 1.6\text{V}$, $V_{\text{AGC}} = 3.0\text{V}$	1.3	1.5	2.0	mA
I_{AGCsc}	Slow AGC Capacitor Charge Current	$V_{(D_{\text{IN}+} - D_{\text{IN}-})} = 1.6\text{V}$, Vary V_{AGC} until slow discharge begins	0.14	0.17	0.22	mA
	Fast to Slow Attack Switchover Point	$\frac{V_{(D_{\text{IN}+} - D_{\text{IN}-})}}{V_{(D_{\text{IN}+} - D_{\text{IN}-}) \text{ Final}}}$		1.25		-
I_{AGCD}	AGC Capacitor Discharge Current	$V_{(D_{\text{IN}+} - D_{\text{IN}-})} = 0.0\text{V}$ Read Mode		4.5		μA
		Hold Mode	-0.2		+0.2	μA
CMRR	CMRR (Input Referred)	$V_{\text{IN}+} = V_{\text{IN}-} = 100\text{mV}_{\text{P-P}} @ 5\text{MHz}$, gain at maximum	40			dB
PSRR	PSRR (Input Referred)	V_{CC} or $V_{\text{DD}} = 100\text{mV}_{\text{P-P}} @ 5\text{MHz}$, gain at maximum	30			dB
TREC	Write to Read Recovery Time. Includes AGC Settling	$V_{(\text{IN}+ - \text{IN}-)} = 100\text{mV}_{\text{P-P}} @ 2.5\text{MHz}$	1.2	2	10	μs

ELECTRICAL CHARACTERISTICS (Continued)

The following specifications apply over the recommended operating conditions of $V_{CC} = 5V \pm 10\%$, $V_{DD} = 12V \pm 10\%$, $0^\circ C \leq T_A \leq 70^\circ C$, IN+ and IN- AC coupled, OUT+ and OUT- differentially loaded with $>600\Omega$ and each side loaded with $<10\text{ pF}$ to GND, $C_{BYP} = 2000\text{ pF}$, OUT+ and OUT- AC coupled to D_{IN+} and D_{IN-} respectively, $V_{AGC} = 2.2V$ unless otherwise specified. (See Note 2.)

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP NOTE 4	MAX	UNITS
READ MODE CHARACTERISTICS (Continued)						
Hysteresis Comparator						
V_{IP}	Input Signal Range				1.5	V_{P-P}
R_{ID}	Differential Input Resistance	$V_{(D_{IN+} - D_{IN-})} = 100\text{ mV}_{P-P} @ 2.5\text{ MHz}$	5		15	$k\Omega$
C_{ID}	Differential Input Capacitance	$V_{(D_{IN+} - D_{IN-})} = 100\text{ mV}_{P-P} @ 2.5\text{ MHz}$			6.0	pF
Z_{IC}	Common Mode Input Impedance	(both sides)		2.0		$k\Omega$
V_{IO}	Comparator Offset Voltage	HYS pin at $-0.5V$, $\leq 1.5\text{ k}\Omega$ across D_{IN+} , D_{IN-}		5		mV
V_{HYSP} V_{HYS}	Peak Hysteresis Voltage vs HYS pin voltage (input referred)	$1V < V_{HYS} < 3V$	0.16	0.21	0.25	V/V
I_I	HYS Pin Input Current	$1V < V_{HYS} < 3V$	0		-20	μA
I_O	LEVEL Pin Max Output Current		3			mA
R_O	LEVEL Pin Output Resistance	$I_{LEVEL} = 0.5\text{ mA}$		180		Ω
V_{OL}	D_{OUT} Pin Output Low Voltage	$T_A = 70^\circ C$	$V_{DD} - 4.0$		$V_{DD} - 2.5$	V
V_{OH}	D_{OUT} Pin Output High Voltage	$T_A = 70^\circ C$	$V_{DD} - 2.2$		$V_{DD} - 1.5$	V
V_{LEVEL} $V_{D_{IN}}$	Level Pin Output Voltage vs $V_{(D_{IN+} - D_{IN-})}$	$0.6 < V_{(D_{IN+} - D_{IN-})} < 1.3 V_{P-P}$ $10\text{ k}\Omega$ from level pin to GND	1.5		2.5	V/V_{P-P}
Active Differentiator						
V_{IP}	Input Signal Range				1.5	V_{P-P}
R_{ID}	Differential Input Resistance	$V_{(C_{IN+} - C_{IN-})} = 100\text{ mV}_{P-P} @ 2.5\text{ MHz}$	5		11	$k\Omega$
C_{ID}	Differential Input Capacitance	$V_{(C_{IN+} - C_{IN-})} = 100\text{ mV}_{P-P} @ 2.5\text{ MHz}$			6	pF
Z_{IC}	Common Mode Input Impedance	(both sides)		2.0		$k\Omega$
I_{OD}	D_{IF+} to D_{IF-} Pin Current	Differentiator Imped must be set so as not to clip signal at this current level	± 1.3			mA
V_{IO}	Comparator Offset Voltage	D_{IF+} , D_{IF-} AC Coupled		5		mV
V_{OL}	C_{OUT} Pin Output Low Voltage	$0 \leq I_{OH} \leq 0.5\text{ mA}$		$V_{DD} - 3$		V
V_{PO}	C_{OUT} Pin Output Pulse Voltage	$0 \leq I_{OH} \leq 0.5\text{ mA}$		0.4		V
PW_0	C_{OUT} Pin Output Pulse Width	$0 \leq I_{OH} \leq 0.5\text{ mA}$		30		ns
A_V	Voltage Gain From $C_{IN\pm}$ to $D_{IF\pm}$	$R_{(D_{IF+} \text{ to } D_{IF-})} = 2\text{ k}\Omega$	1.7		2.2	V/V
Undervoltage Detector (ML4042 Only)						
$V_{CC\ TH+}$	V_{CC} Fault Threshold +	$\overline{V_{FLT}}$ transition from low to high	3.8	4.2	4.5	V
$V_{CC\ TH-}$	V_{CC} Fault Threshold -	$\overline{V_{FLT}}$ transition from high to low	3.8	4.1	4.5	V
$V_{DD\ TH+}$	V_{DD} Fault Threshold +	$\overline{V_{FLT}}$ transition from low to high	9.6	10.2	10.8	V
$V_{DD\ TH-}$	V_{DD} Fault Threshold -	$\overline{V_{FLT}}$ transition from high to low	9.6	10.0	10.8	V
V_{OL}	Output Low Voltage ($\overline{V_{FLT}}$)	$I_{OL} = 1.6\text{ mA}$			0.4	V
V_{OH}	Output High Voltage ($\overline{V_{FLT}}$)	$I_{OH} = -400\mu A$	2.7			V

ML4041, ML4042

ELECTRICAL CHARACTERISTICS (Continued)

The following specifications apply over the recommended operating conditions of $V_{CC} = 5V \pm 10\%$, $V_{DD} = 12V \pm 10\%$, $0^\circ C \leq T_A \leq 70^\circ C$, $V_{(C_{IN+} - C_{IN-})} = V_{(D_{IN+} - D_{IN-})} = 1.0V_{p,p}$ AC coupled sine wave at 2.5MHz, $R_{D_{IF}} = 100\Omega$, $C_{D_{IF}} = 65pF$, $V_{HYS} = 1.8V$, $C_{OS} = 60pF$, $4k\Omega$ to V_{CC} and $10pF$ to GND on pin \overline{RD} unless otherwise specified.

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP NOTE 4	MAX	UNITS
Output Data Characteristics (Refer to Figure 2)						
t_{D1}	D-Flip-Flop Set Up Time	Min delay from $V_{(D_{IN+} - D_{IN-})}$ exceeding threshold to $V_{(D_{IF+} - D_{IF-})}$ reaching a peak	0			ns
t_{D3}	Propagation Delay				110	ns
t_{D5}	Output Data Pulse Width Variation	(See Note 5) $C_{OS} = 60pF$, $T_A = 25^\circ C$	40	50	65	ns
$t_{D3} - t_{D4}$	Logic Skew (Pulse Pairing)				1	ns
t_R	Output Rise Time	$V_{OH} = 2.4V$			18	ns
t_F	Output Fall Time	$V_{OL} = 0.4V$			14	ns

Note 1: Absolute maximum ratings are limits beyond which the life of the integrated circuit may be impaired. All voltages unless otherwise specified are measured with respect to ground.

Note 2: Limits are guaranteed by 100% testing, sampling, or correlation with worst-case test conditions.

Note 3: AGC amplifier output current may be increased as in Figure 4.

Note 4: Typicals are parametric norm at $25^\circ C$.

Note 5: $t_{D5} = 830 (C_{OS})$, $50pF < C_{OS} < 150pF$

FUNCTIONAL DESCRIPTION

Operating Modes

The ML4041, ML4042 has three definitive operation modes which are: Read mode, Write mode and Hold mode. These modes are defined by input pins \overline{HOLD} and R/W as shown in Table 1. Read mode, the mode used normally for pulse detection, is assumed in the following sections unless otherwise noted.

AGC Amplifier Section

The purpose of the AGC amplifier is to provide a constant read signal level for both the level and slope detectors. Full differential processing of the read signal is used to minimize noise and distortion in the analog signal. A wide gain range is required due to large signal variation when moving the recording head from an inside to outside data track or variations in media.

The differential output voltage level V_{OUT} from the AGC amp is determined by voltage V_{AGC} present at pin AGC. V_{OUT} is full wave rectified and compared against V_{AGC} to create charge/discharge current for capacitor C_{BYP} connected at pin BYP. Voltage V_{BYP} across C_{BYP} controls the gain in the AGC amplifier.

Two distinct values of I_{BYP} are possible which determine a fast and slow AGC gain response attack rate. When V_{OUT} is more than 125% of the set level a high value of I_{BYP} is sourced which provides a fast AGC attack rate. When V_{OUT} is within 100% to 125% of the set level a reduced value of I_{BYP} is sourced which provides a slower attack rate. The fast-slow gain response attack rates provides for an initial quick system

response and then minimum zero crossing distortion of the analog signal once the gain is within working range. V_{AGC} should be set so that the differential input voltage $V_{D_{IN}}$ into the level comparator is $1V_{p,p}$ at nominal Read signal conditions. The AGC amp section gain is given by:

$$\frac{A_{V2}}{A_{V1}} = \exp \frac{V_{BYP2} - V_{BYP1}}{5.8 \times V_T}$$

Where: A_{V1} , A_{V2} are initial and final amplifier gain values corresponding to initial and final V_{BYP} values.

$$V_T = (KT)/Q = 26mV \text{ at room temperature.}$$

The AGC amp's differential inputs must be AC coupled to the read amplifier (ML117, ML501, etc.) differential outputs. Similarly, AC coupling must be used at the AGC amp outputs.

AGC Amp During Write Mode— When the ML4041, ML4042 is put into write mode, the AGC amp's input impedance is lowered to allow a faster dampening of the Write to Read transient from the head pre-amp. The AGC gain is also set to maximum gain so that fast AGC attack will occur when changing back to the Read mode. Internal device timing is controlled so that settling occurs prior to Read mode activation. Minimal value input coupling capacitors should be chosen to reduce settling time, however, bandwidth requirements also need to be considered.

AGC Amp During Hold Mode—During the Hold mode, the charge/discharge current driving pin BYP is internally disconnected. AGC compensation capacitor C_{AGC} will then hold the present gain setting. The amplitude of V_{OUT} will therefore not affect the AGC gain and gain will remain constant.

Hold mode is used so that AGC gain will not be adjusted when embedded servo information is read. This prevents losing the pulse peak amplitude information needed during position decoding, or creating additional gain settling time when again reading data. Embedded servo pulses are normally taken at outputs D_{IF-} and D_{IF+} , as shown in the typical application.

External Filter Network

Filtering for the level and slope detectors can be performed with a single filter or two separate filters. If separate filters are used, care must be used to insure that time delays are matched. A multi-pole Bessel filter is recommended due to the group delay and linear phase characteristics.

Level Detector

The full wave rectified V_{OUT} is buffered and available at pin LEVEL. The level detector uses a hysteresis comparator to compare the processed read signal amplitude against a reference voltage derived from voltage V_{LEVEL} output from pin LEVEL. Using V_{LEVEL} provides a feed-forward function that allows valid level detection to be performed prior to AGC amp gain settling. The level detector hysteresis value is set in a way that will only allow relatively large read pulse peaks (negative or positive) to be detected.

Slope Detector

The slope detector uses an external reactive component network to produce a voltage signal proportional to the differential of the read signal. By using a hysteresis comparator to detect zero slope of the read signal, the time occurrence of positive or negative read pulse peak values can be determined.

An external reactive network, shown in the Typical Application, is used between the D_{IF+} and D_{IF-} pins to provide the differential function given by:

$$A_V = \frac{-2000Cs}{LCs^2 + (R + 92)Cs + 1}$$

Where: C = External capacitor (20 pF to 150 pF)

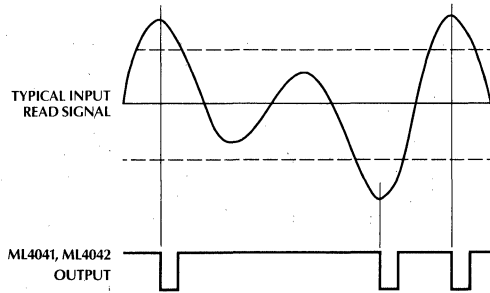
L = External inductor

R = External resistor

s = $j\omega = j2\pi f$

Output Logic

The output logic provides a negative TTL pulse at pin \overline{RD} which begins at the peak of a valid read pulse, as shown below.



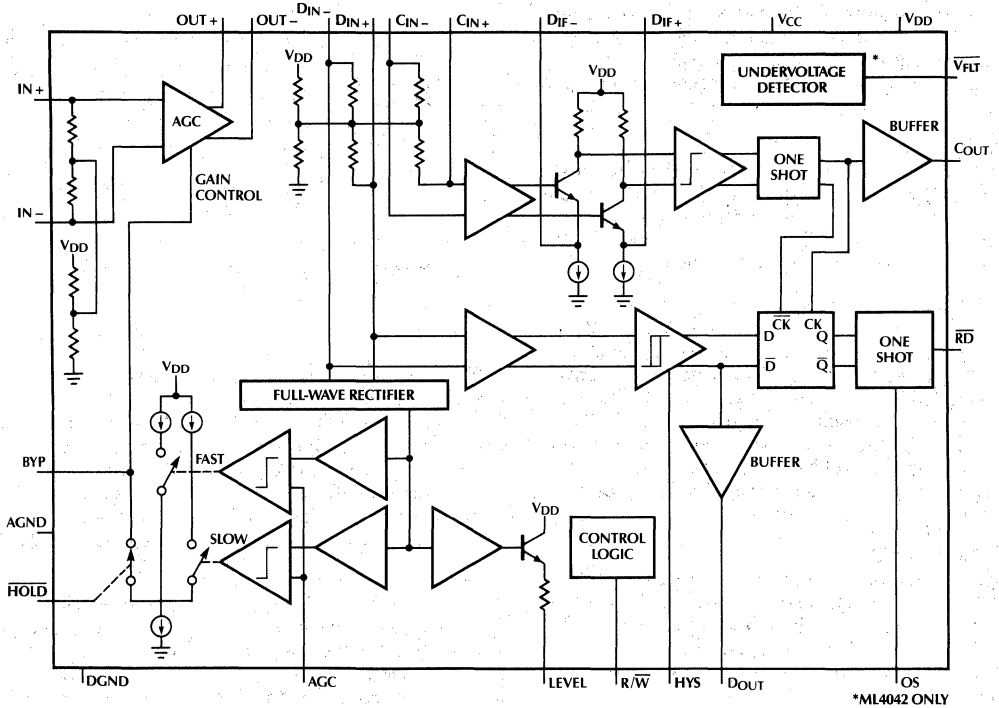
Pin R/\overline{W} must be high for the output logic to be active. The key element in the output logic is the D flip-flop. The flip-flop is clocked by the slope detector at the time of a zero crossing, which loads data from level detector. The flip-flop inputs only change state when the level detector detects a peak amplitude of a polarity opposite to the previous valid peak. Thus, through the output logic the slope detector determines output timing and the level detector determines pulse validity.

Layout Considerations

As with any high gain, wide bandwidth analog circuitry, care needs to be exercised in PC layout. Power supply and ground lines should be bypassed and well isolated from other circuitry. A ground plane is recommended, as is keeping analog lines short and well balanced to prevent interaction with nearby circuitry in the disk drive.

ML4041, ML4042

BLOCK DIAGRAM



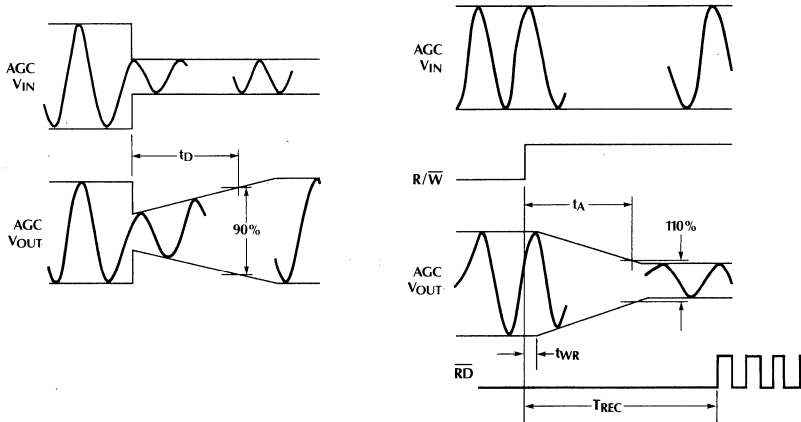


Figure 1. AGC Timing Diagram

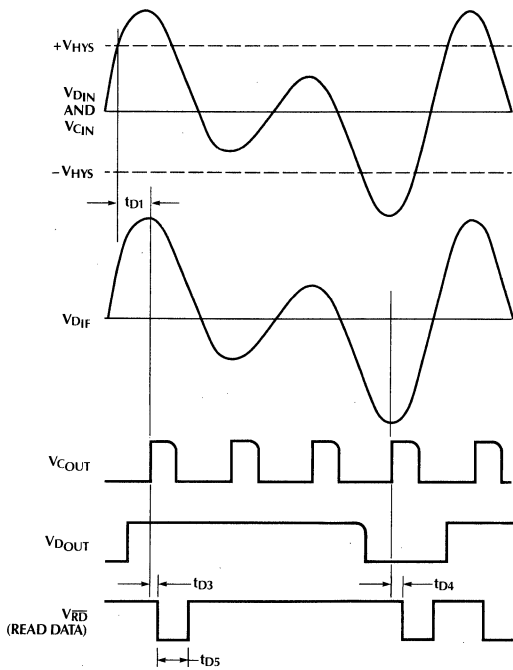


Figure 2. Output Logic Timing Diagram

TYPICAL APPLICATIONS

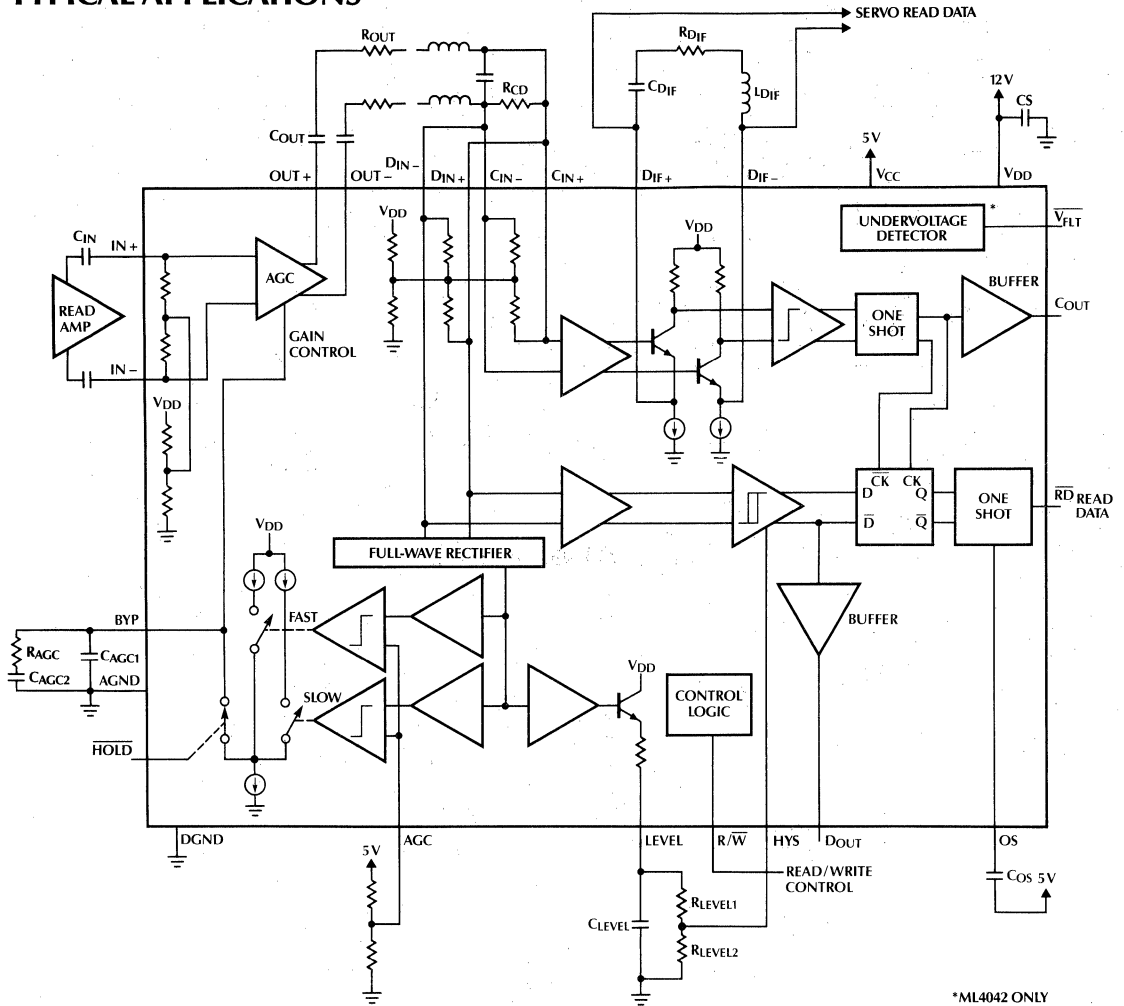
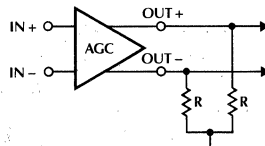


Figure 3. Typical Application Diagram



$$I_{OUT+} = I_{OUT-} = 4.2\text{mA} + \frac{8\text{V}}{R_{EXT}}$$

Figure 4. Modification of AGC Amplifier Output to Drive Low Impedance Filters

ORDERING INFORMATION

PART NUMBER	TEMP. RANGE	PACKAGE
ML4041CP	0°C to +70°C	MOLDED DIP (P24)
ML4041CQ	0°C to +70°C	MOLDED PCC (Q28)
ML4041CS	0°C to +70°C	MOLDED SOIC (S24)
ML4042CP	0°C to +70°C	MOLDED DIP (P28)
ML4042CQ	0°C to +70°C	MOLDED PCC (Q28)
ML4042CS	0°C to +70°C	MOLDED SOIC (S28)

Servo Demodulator

GENERAL DESCRIPTION

The ML4401 provides all of the analog circuitry necessary for the demodulation of di-bit servo signal information in Winchester disk drives. It interfaces to the servo head preamp and provides quadrature position signal outputs for the servo controller circuitry.

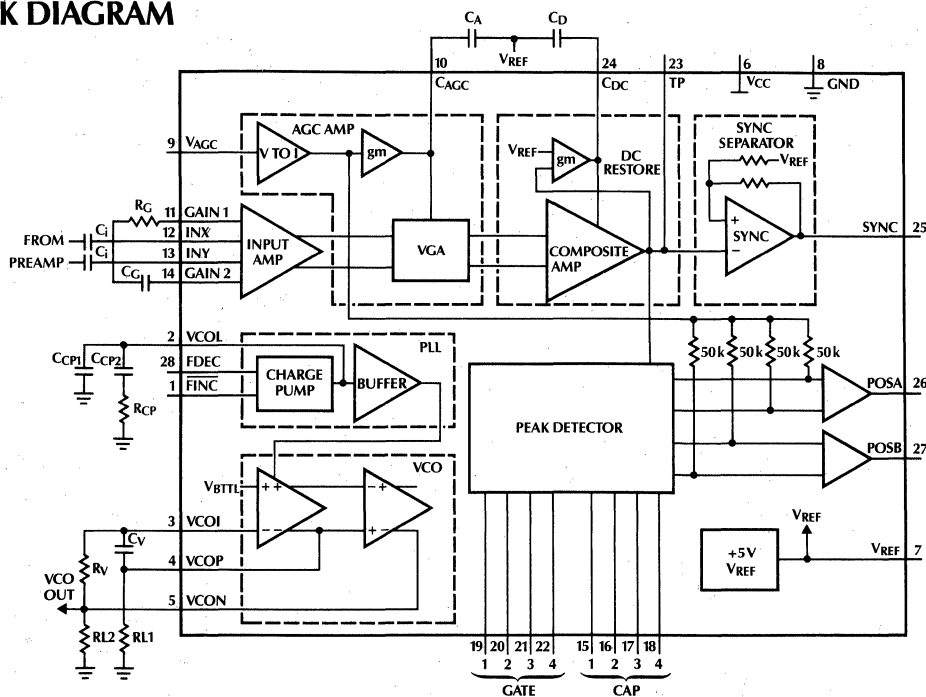
The ML4401 includes a high-performance 592-type input amplifier and differential AGC circuit. External logic is designed to meet the needs of the particular servo system utilizing the VCO and Charge Pump to create a PLL time base for Peak Detector gating. The SYNC output provides servo channel timing information for the logic.

The ML4401 when combined with the ML4402, ML4406/07/08 Servo Driver, the ML4403, ML4413 Servo Controller and the ML4404 Trajectory Generator, provides a flexible closed-loop servo control system.

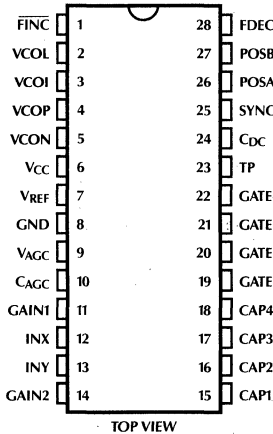
FEATURES

- Combines all analog di-bit demodulation circuitry
- Logic track-type switching can be used to minimize demodulator offset
- Exponential AGC characteristics makes AGC settling independent of input step size
- External loop compensation of analog blocks
- External digital circuitry allows flexible pattern format
- On-chip band gap voltage reference eliminates external referencing
- Operates from 12V power supply
- Compatible with Micro Linear's ML4403, ML4413 Servo Controller, ML4402, ML4406/07/08 Servo Driver and ML4404 Trajectory Generator

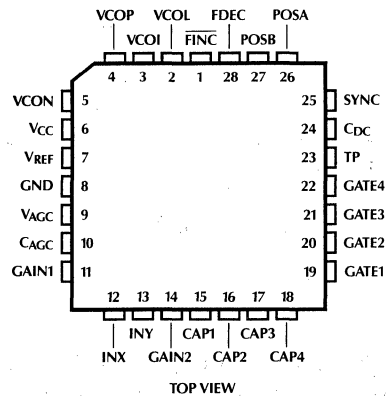
BLOCK DIAGRAM



PIN CONNECTIONS

ML4401 28-Pin DIP
(Prototypes Only)

ML4401 28-Pin PCC



PIN DESCRIPTION

PIN #	NAME	FUNCTION	PIN #	NAME	FUNCTION
1	FINC	Charge pump frequency increment input (TTL).	15	CAP1	Peak detector 1 capacitor terminal.
2	VCOL	PLL loop compensation terminal.	16	CAP2	Peak detector 2 capacitor terminal.
3	VCOI	VCO high impedance input.	17	CAP3	Peak detector 3 capacitor terminal.
4	VCOP	VCO positive output, for capacitive feedback to VCOI.	18	CAP4	Peak detector 4 capacitor terminal.
5	VCON	VCO negative output, drives resistance feedback to VCOI, also provides ECL output on ML4401 and TTL output on ML4411.	19	GATE1	Peak detector 1 gate input (TTL) high enabled, low disabled.
6	VCC	+12V supply.	20	GATE2	Peak detector 2 gate input (TTL) high enabled, low disabled.
7	VREF	Voltage reference output (+5V).	21	GATE3	Peak detector 3 gate input (TTL) high enabled, low disabled.
8	GND	Ground.	22	GATE4	Peak detector 4 gate input (TTL) high enabled, low disabled.
9	VAGC	AGC gain reference voltage input.	23	TP	Composite test point, normally left unconnected.
10	CAGC	External capacitor terminal to set AGC response.	24	CDC	External capacitor terminal to set DC restore response.
11	GAIN1	Input amplifier gain adjusting RC terminal 1.	25	SYNC	SYNC pulse output (TTL).
12	INX	X input into input amplifier.	26	POSA	Position output A.
13	INY	Y input into input amplifier.	27	POSB	Position output B.
14	GAIN2	Input amplifier gain adjusting RC terminal 2.	28	FDEC	Charge pump frequency decrement input (TTL).

ABSOLUTE MAXIMUM RATINGS

Absolute maximum ratings are those values beyond which the device could be permanently damaged. Absolute maximum ratings are stress ratings only and functional device operation is not implied.

Power Supply Voltage, V_{CC}	14V
Input Voltages:	
GAIN1, GAIN2	-0.3 to 8V
V_{ACC}	-0.3 to 7.0V
V_{AGC}	-0.3 to 5.3V
CAP1, CAP2, CAP3, CAP4	3.0 to 10V
GATE1, GATE2, GATE3, GATE4, VCOP	-0.3 to 7.5V
INX, INY, VCON, VCOI, FINC, FDEC, C_{DC}	-0.3 to $V_{CC} + 0.3V$
θ_{JA} for 28-Pin Plastic Dip	$\leq 60^{\circ}C/Watt$
θ_{JA} for 28-Pin PLCC	$60^{\circ}C/Watt$
Storage Temperature Range	$-65^{\circ}C$ to $150^{\circ}C$
Junction Temperature (T_{JMAX})	$150^{\circ}C$
Lead Temperature (Soldering, 10sec)	$260^{\circ}C$

OPERATING CONDITIONS

Temperature Range	$0^{\circ}C$ to $70^{\circ}C$
Supply Voltage (V_{CC})	$12V_{DC} \pm 10\%$
Input Coupling Capacitance (C_i)	$0.01\mu F$
Input Amp Gain Capacitance (C_G)	$0.047\mu F$
Input Amp Gain Resistance (R_G)	$1k\Omega$
AGC Response Compensation Capacitance (C_A)	$0.082\mu F$
Composite DC Restore Capacitance (C_D)	$0.01\mu F$
PLL Compensation Components:	
C_{CP1}	$0.1\mu F$
C_{CP2}	$1\mu F$
R_{CP}	910Ω
PLL Gain Components:	
R_V	1000Ω
RL1, RL2	1000Ω
Peak Detector Capacitance (CAP1 thru CAP4)	$270pF$
SYNC Output Pull-Up Resistor (to 5V)	$3.3k$
On track Base-to-Peak Voltage at pin TP	$1.75V$
V_{GA} Gain Control Voltage (at pin C_{ACC})	0.65

ELECTRICAL CHARACTERISTICS

The following specifications apply over the recommended operating conditions of $T_A = 0$ to $70^{\circ}C$, $V_{CC} = 10.8$ to $13.2V$, $V_{VACC} = 5.0V$, and external components as recommended above, unless otherwise specified (See Note 1.)

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP NOTE 2	MAX	UNITS
Power Supply						
I_{CC}	Supply Current	$V_{CC} = 12V$		81	110	mA
TTL Inputs FINC, FDEC, GATE1, GATE2, GATE3, GATE4						
V_{IH}	High Level Input Voltage		2.0			V
V_{IL}	Low Level Input Voltage				0.8	V
I_{IH}	High Level Input Current	$V_{IH} = 2.4V$	-1		30	μA
I_{IL}	Low Level Input Current	$V_{IL} = 0.4V$	-20		1	μA
SYNC Output (TTL Open Collector) See Note 3						
V_{OL}	Low Level Output Voltage	$I_{OL} = 1.6mA$	0	0.3	0.5	V
V_{THR}	Positive going input threshold			$V_{REF} + 0.9$		V
V_{THF}	Negative going input threshold			V_{REF}		V
$t_{PD} \pm$	Propagation Delay Rising, Falling	$R_L = 2k, C_L = 15pF$		50		ns
VCOP Output ML4401 ($T_A = 25^{\circ}C$)						
V_{OH}	High Level Output Voltage	$R_L = 1k\Omega$	4.0	4.3	4.6	V
V_{OL}	Low Level Output Voltage	$R_L = 1k\Omega$	2.9	3.2	3.5	V
VCOP Output ML4411						
V_{OH}	High Level Output Voltage	$I_{OH} = 50\mu A$	2.4			V
V_{OL}	Low Level Output Voltage	$I_{OL} = 1.6mA$	0		0.5	V
VCO and Charge Pump Section						
I_{BIAS}	V_{COI} Input Bias Current		0	25	50	μA
I_{CH}, I_{DIS}	V_{COL} Charge and Discharge Current		495	660	825	μA
I_{CH}/I_{DIS}	V_{COL} Charge/Discharge Ratio		0.95	1.00	1.05	$\mu A/\mu A$
I_{OFF}	V_{COL} OFF State Current	FINC = 2.0 FDEC = 0.8	0	25	50	nA

ELECTRICAL CHARACTERISTICS (Continued)

The following specifications apply over the recommended operating conditions of $T_A = 0$ to 70°C , $V_{CC} = 10.8$ to 13.2V , $V_{AGC} = 5.0\text{V}$, and external components as recommended above, unless specified (See Note 1.)

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP NOTE 2	MAX	UNITS
VCO and Charge Pump Section (Continued)						
F_{MAX}	MAX VCO Frequency to Maintain + and - 5% Control Range Note 4		30			MHz
F_{VCO}	VCO Frequency Range Note 4	$T_A = 25^\circ\text{C}$, $V_{CC} = 12$, $V_{COL} = 6\text{V}$ $C_V = 1000\text{pF}$, $R_V = 604\Omega$	9.7	10.0	10.3	MHz
K_{VCO}	VCO Voltage to Frequency Factor			2		%/V
Input AMP, AGC AMP, and DC Restore						
R_{IN}	INX, INY Differential Input Resistance		7.5	10	20	k Ω
$I_{GAIN1,2}$	GAIN1, GAIN2 Bias Current		0.66	1.0	1.20	mA
I_{BIAS}	V_{AGC} Input Bias Current		0	5	20	μA
G_{MAGC}	AGC Transconductance at C_{AGC}			370		μMHOS
R_{AGC}	Control Range of AGC Loop to Regulate Composite Amplitude to within 2% of Nominal			7/1		V/V
BW	Bandwidth from INX, INY to Composite Note 4		10	15		MHz
GMDCR	DC Restore Transconductance			200		μMHOS
Peak Detectors						
I_{CH}	Charge Current		12.7			mA
I_{DIS}	Discharge Current	$T_A = 25^\circ\text{C}$	25	45	60	μA
T_{CDIS}	Tempco of I_{DIS}			-0.17		$\mu\text{A}/^\circ\text{C}$
Voltage Reference						
V_{REF}	Reference Voltage	$T_A = 25^\circ\text{C}$	4.85	5.10	5.35	V
TC	Tempco			50		ppm/ $^\circ\text{C}$
R_{OUT}	Load Regulation			2		mV/mA
PSRR	Line Regulation			10		mV/V
I_{SINK}	Maximum SINK Current		0.8			mA
Output Amplifiers (POSA, POSB)						
V_{OS}	Input Offset	$V_{CAP1-4} = 6\text{V}$	-10	0	10	mV
A_V	Gain		1.23	1.28	1.33	V/V
A_{VA}/A_{VB}	Gain Tracking		-3	0	+3	%
V_{OUT}	Output Voltage Range		1.0		9.5	V
I_{SRC}	Output Source Current		5			mA
I_{SNK}	Output Sink Current		2			mA
SR	Slew Rate			2.5		V/ μs
BW	3dB Gain Bandwidth			3		MHz

Note 1: 0°C to 70°C operating temperature range devices are 100% tested with temperature limits guaranteed by 100% testing, sampling, or correlation with worst-case test conditions.

Note 2: Typicals are parametric norm at 25°C .

Note 3: Pin 25 is an open collector output which should not exceed 7 volts in the high state.

Note 4: This parameter is guaranteed but not 100% tested and is not used in outgoing quality level calculations.

APPLICATION HINTS

Using a nominal on-track servo signal, amplitude adjustment should be made as follows:

- Set composite signal amplitude, measured at pin TP, by adjusting voltage at pin V_{AGC} (approximately 4.7 volts). The composite signal should be set to 1.75 volts base to peak of an on-track position pulse (an off-track position pulse will be about 3.5 volts maximum).
- Adjust R_g so that the VGA is in mid-range. This is determined by measuring the voltage at pin C_{AGC} ; it should be approximately 0.9 volts. C_{AGC} voltage will vary approximately ± 0.5 volts over the AGC range.

FUNCTIONAL DESCRIPTION

Input Amplifier

The input amplifier is equivalent to a wide-band 592 type video amplifier and provides amplification and buffering to the AGC circuitry. The Inputs INX and INY, which must be AC coupled, accept the composite analog signal from the servo head differential preamplifier. Internal input termination resistors eliminate the need for external bias resistors. Prefiltering of the signal is normally desired to eliminate unwanted components. External components R_G and C_G determine the input amplifier's low frequency cutoff and gain as follows:

$$FC = \frac{1}{2\pi (R_G + 60\Omega) C_G} \quad A_V = \frac{1700}{R_G + 60\Omega}$$

Where: C_G = External series capacitance between pins GAIN1 and GAIN2

R_G = External series resistance between pins GAIN1 and GAIN2

Automatic Gain Control (AGC)

The purpose of the AGC loop is to maintain a constant peak output voltage level at outputs POSA and POSB. This peak level is established by the reference voltage applied to pin V_{AGC} .

$$V_{P-P}(\text{Composite}) = K1 \times V_{AGC} + K2$$

Where: $K1 = 0.65$
 $K2 = 0.41V$

In this closed-loop system, the peak detector output voltages are fed back and combined with the V_{AGC} voltage to provide a gain control current. The current controls the variable gain amplifier (VGA) and is compensated at pin C_{AGC} to provide control of AGC bandwidth. The bandwidth of the entire AGC loop is determined by:

$$BW = \frac{K V_{AGC}}{2\pi C_A}$$

Where: $K = 4.3 \times 10^{-4}$
 V_{AGC} = External reference voltage at pin V_{AGC}
 C_A = External capacitance at pin C_{AGC}

Optimum system stability is achieved by deriving V_{AGC} from the V_{REF} output using a resistive divider.

Composite Amplifier

The input amplifier and AGC circuit of the ML4401 operate in a differential signal mode to provide good common mode and power supply rejection. The composite amplifier converts the differential signal into a buffered single-ended signal for the peak detector circuitry. The DC base line of the composite signal is equal to V_{REF} . The bandwidth of the DC restore function is controlled by capacitor C_D at pin C_{DC} with the following relationship:

$$BW = \frac{gm}{2\pi C_D}$$

Where: $gm = 1/5 \text{ k}\Omega$
 C_D = External capacitance at pin C_{DC}

The composite signal is available at pin TP and is normally left unconnected. For short circuit protection a 425 Ω resistor is connected in series with pin TP internally.

Synchronization Pulse Separator

The SYNC pulse separator is a threshold comparator with hysteresis which passes pulses from the composite amplifier above a set threshold. It provides a buffered open collector TTL output. The SYNC output, when gated through an external one-shot, is used to control the external gate timing and PLL logic.

Peak Detector

The peak detector circuit captures the peak signal amplitude of the di-bit pulses. The gates are controlled by inputs GATE1 through GATE4. Timing is established by the external logic circuitry. The external peak detector capacitors are connected from pins CAP1 through CAP4 to ground. The peak detector discharge rate (set by CAP1-CAP4) determines the maximum track crossing rate during an access operation. The performance of this block can be enhanced by using the velocity output of the ML4403, ML4413 to create a velocity proportional discharge. The peak detector outputs are fed into internal differential amplifiers that calculate the track error signals and provide buffered outputs POSA and POSB as follows:

$$POSA = 1.25 (CAP1 - CAP2) + V_{REF}$$

$$POSB = 1.25 (CAP3 - CAP4) + V_{REF}$$

Voltage Controlled Oscillator and Charge Pump

The VCO and external phase compare logic provide a time base for peak detector gate synchronization. Inputs FINC and FDEC provide increment and decrement signals to the charge pump for changing the oscillator frequency. The FINC and FDEC inputs gate the charge pump for the duration of the pulse width. The RC timing network formed by C_V and R_V at pins VCOI, VCON, and VCOP control the oscillators center frequency. (See Typical Performance Characteristics)

R_V should be greater than 330 Ω . Too low of a value will result in excessive power dissipation. $RL1$, $RL2$ and R_V should be approximately equal, although the values of $RL1$ and $RL2$ do not require accuracy.

The VCO output should only be taken from pin VCON. Charge pump capacitor C_{CP1} is connected from pin VCOL to ground. Components R_{CP} and C_{CP2} are also connected in series from pin VCOL to ground to provide VCO loop compensation.

Internal Voltage Reference

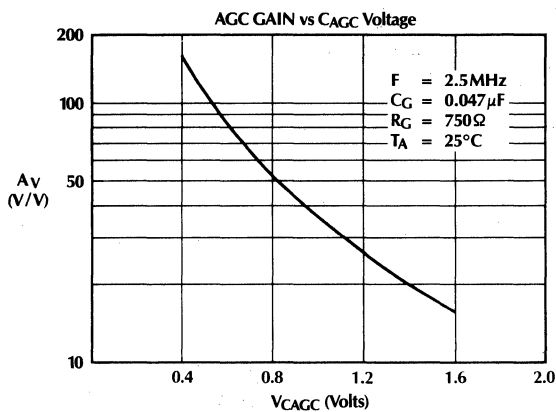
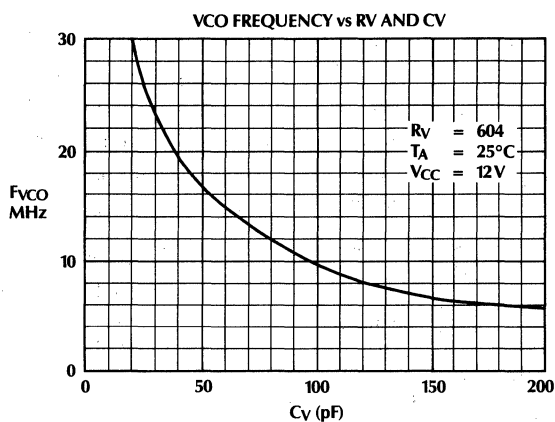
V_{REF} is an internal band-gap voltage reference. It is buffered and available at pin V_{REF} and is used by the ML4402, ML4403, ML4404 and other chips requiring a 5 volt reference.

External Logic

The external logic provided by the user typically has a complexity of about 150 to 300 equivalent gates. Complexity and architecture depends on the users di-bit pattern and control function.

Note: Stray capacitance should be considered in applying the above relationships when low capacitor values are used. Stray capacitance of the integrated circuit terminal is typically about 2 to 3 pF.

TYPICAL PERFORMANCE CHARACTERISTICS



ORDERING INFORMATION

PART NUMBER	TEMP. RANGE	PACKAGE
ML4401CP	0°C to +70°C	MOLDED DIP (P28)
ML4401YCQ	0°C to +70°C	MOLDED PCC (Q28)

GENERAL DESCRIPTION

The ML4402 Servo Driver contains all of the control circuitry necessary to drive the head positioning actuator of a hard or rigid disk drive system. It receives the error signal generated from a servo controller circuit, such as the ML4403, ML4413, and drives an external transistor bridge which controls the head positioning voice coil actuator. The ML4402 output control circuitry includes current sense inputs to provide closed-loop control of actual actuator current. By using an external power transistor bridge, flexible thermal and space management is allowed as well as transistor selection which enables a wide application range.

Included in the device is a unique disable function which permits interruption of actuator current. During a disable, the output control amplifiers are shut down which cuts off all current to the external transistor bridge. Disable can be activated by a logic high into pin DIS or by the on-board low-voltage detector. Use of the low-voltage disable function prevents actuator response to a false error signal during a power failure. The low voltage detector can monitor up to two power supplies and has user definable low voltage trigger levels.

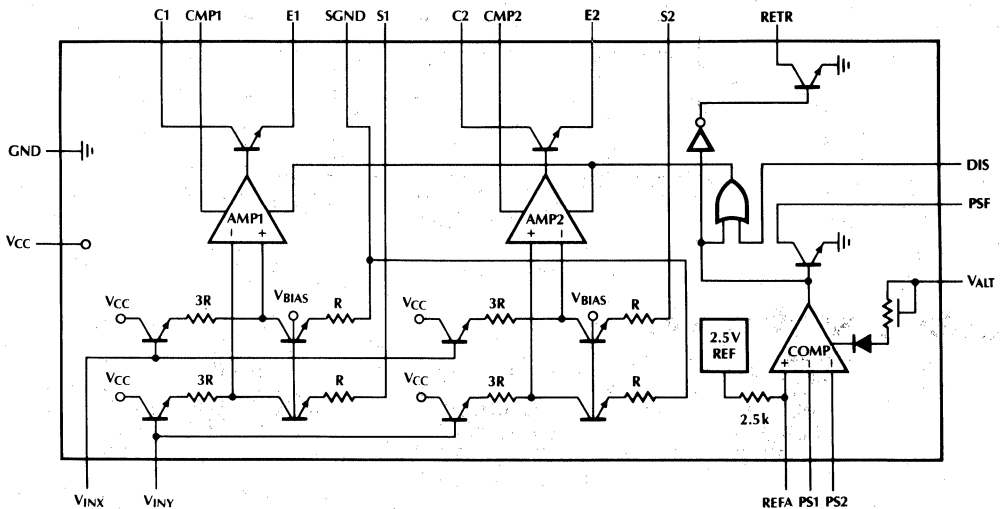
The ML4402, when combined with the ML4401/4431 Servo Demodulator, the ML4403, ML4413 Analog Servo Controller and the ML4404 Trajectory Generator, provides a flexible high-performance head positioning servo system.

FEATURES

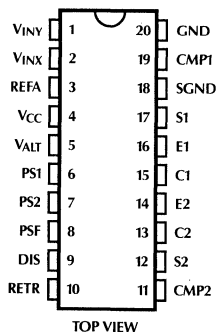
- Low differential input offset voltage
- Contains all control circuitry necessary to drive an external transistor bridge
- Differential amplifiers internally compensated
- Unique disable function interrupts actuator current
- Programmable dual supply low voltage detector
- Single +12V power supply
- Compatible with Micro Linear's ML4401/4431 Servo Demodulator, ML4403, ML4413 Servo Controller and ML4404 Trajectory Generator chips

The ML4402-1 and ML4402-2 differ in offset voltage at the differential error signal inputs which is a result of the manufacturing trim process.

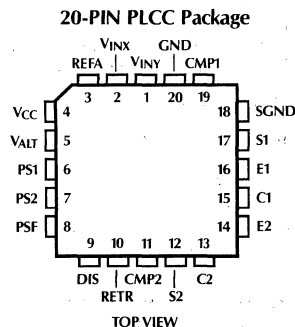
BLOCK DIAGRAM



PIN CONNECTIONS



20-PIN DIP (Prototypes Only)



PIN DESCRIPTION

PIN #	NAME	FUNCTION	PIN #	NAME	FUNCTION
1	V _{INY}	Inverting input for error voltage signal. Used as a reference voltage (analog ground) input when using a single ended output from the ML4403 Servo Controller. Obtained from the V _{REF} output of the ML4401 Servo Demodulator.	11	CMP2	Compensation node of AMP2 used to add additional compensation; the device is manufactured with approximately 27 pF of internal compensation. Bandwidth Effects: $f = \frac{gm}{2\pi(C+27\text{ pF})}$ Slew Rate Effects: $SR = \frac{20\mu\text{A}}{C+27\text{ pF}}$ Where: gm = 150 μmhos C = External Compensation Capacitor C _{CMP1} or C _{CMP2}
2	V _{INX}	Non-inverting input for error voltage signal. Used as the signal input pin when using a single ended output from the ML4403.	12	S2	Current sense input for AMP2.
3	REFA	Reference pin for low voltage comparator.	13	C2	Collector of output transistor of AMP2.
4	V _{CC}	+12 V power supply pin.	14	E2	Emitter of output transistor of AMP2.
5	V _{ALT}	Optional +5V power supply pin to keep the PSF pin operating if V _{CC} fails. With V _{ALT} at +5V, the PSF pin will go low if V _{CC} goes to zero, or too low to operate the comparator.	15	C1	Collector of output transistor of AMP1.
6	PS1	Voltage input for low voltage comparator.	16	E1	Emitter of output transistor of AMP1.
7	PS2	Voltage input for low voltage comparator.	17	S1	Current sense input for AMP1.
8	PSF	Power supply failure indication, is an open collector output of comparator. Logic low indicates PS1 and/or PS2 voltage has gone below REFA.	18	SGND	Reference ground for S1, S2 feedback.
9	DIS	Amplifier Disable pin. TTL input that disables both amplifiers with a logic high.	19	CMP1	Compensation node of AMP1, used to add additional compensation. The device is manufactured with approximately 27 pF of internal compensation. Bandwidth and Slew Rate effects are the same as the CMP2 pin.
10	RETR	Return spring output, clamped open collector output, opposite logic polarity as pin PSF. Used to drive optional safety circuitry.	20	GND	Ground.

ABSOLUTE MAXIMUM RATINGS

Absolute maximum ratings are those values beyond which the device could be permanently damaged. Absolute maximum ratings are stress ratings only and functional device operation is not implied.

Power Supply Voltage (V_{CC})	14V
Terminal Voltage Range (V_{INX} , V_{INY} , V_{ALT} , PS1, PS2, REFA, DIS)	-0.3 to $V_{CC}+0.3V$
S1, S2	7V
Terminal Input Current (CMP1, CMP2)	0.1mA
Storage Temperature Range	-65°C to +150°C
Junction Temperature (T_J)	125°C
Lead Temperature (Soldering, 10sec)	260°C

OPERATING CONDITIONS

Supply Voltage	
V_{CC}	12V ± 10%
V_{ALT}	5V ± 10%
Typical Component Values (Refer to Typical Application)	
R_{OA}	470Ω
R_{OB}	240Ω
R_{OC}	150Ω
R_{OD}	0.5Ω

ELECTRICAL CHARACTERISTICS

The following specifications apply over the recommended operating conditions of $V_{CC}=10.8V$ to $13.2V$, $V_{INY}=5V$, $T_A=0$ to $70^\circ C$, and external components as shown above unless otherwise specified (See Note 1).

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP NOTE 2	MAX	UNITS
Power Supply						
I_{CC}	V_{CC} Supply Current	Outputs unloaded, Pin REFA open		10	20	mA
I_{DD}	V_{ALT} Supply Current	$V_{CC}=GND$		150	500	μA
Amplifier Characteristics						
A_{V1}	Voltage Gain at Pin S1, $V_{S1}/(V_{INX}-V_{INY})$; Applies when $V_{INX}>V_{INY}$	$V_{INX}=5.1$ and $6V$ $V_{INY}=5.0V$	0.342	0.352	0.362	V/V
A_{V2}	Voltage Gain at Pin S2, $V_{S2}/(V_{INX}-V_{INY})$; Applies when $V_{INY}>V_{INX}$	$V_{INX}=4.9$ and $4V$ $V_{INY}=5.0V$	0.342	0.352	0.362	V/V
e_{AV}	Gain Linearity Error	$(A_{V1}-A_{V2})/0.5(A_{V1}+A_{V2})$	-2	0	2	%
V_{OS}	V_{INX} , V_{INY} Input Offset Voltage with Respect to Either Pin S1 or Pin S2	V_{OS} defined where A_{V1} or $A_{V2}>0.16$ $T_A=25^\circ C$	-10		+10	mV
$V_{OS\ DIFF}$	Differential Input Offset	$V_{OS1}-V_{OS2}$ $T_A=25^\circ C$	-5 -10		+5 +10	mV
T_{CVOS}	Offset Voltage Tempco			15		μV/°C
V_S	Voltage Swing Range of Pin S1, S2 Above Ground	V_{S1} : $V_{INX}=6.7V$ V_{S2} : $V_{INX}=3.3V$		0.5	0.65	V
I_{VR}	Input Voltage Range into V_{INX} and V_{INY}		3.3		10	V
I_{IB1}	Input Bias Current, V_{INX} and V_{INY}		0	10	75	μA
I_{IB2}	Input Bias Current, Pin S1 or S2 (sourcing)	$V_{S1}, V_{S2}=GND$	-1.6	-1.2	-0.8	mA
PSRR	Power Supply Rejection			60		dB
CMRR	Common Mode Rejection Ratio			80		dB
GBP	Gain Bandwidth Product	$C_{CMP1,2}=0$		0.83		MHz
SR	Slew Rate	$C_{CMP1,2}=0$		0.74		V/μs
Output Transistor Characteristics						
I_{OUT}	Output Current; I_{C1} , I_{E1} , I_{C2} , I_{E2}	$V_{INX}-V_{INY}=+1V$ $V_{C1}, V_{C2}=3V$; $V_{E1}, V_{E2}=0.7V$	50	100		mA

ELECTRICAL CHARACTERISTICS (Continued)

The following specifications apply over the recommended operating conditions of $V_{CC} = 10.8\text{V}$ to 13.2V , $V_{IN} = 5\text{V}$, $T_A = 0$ to 70°C , and external components as shown unless otherwise specified (See Note 1).

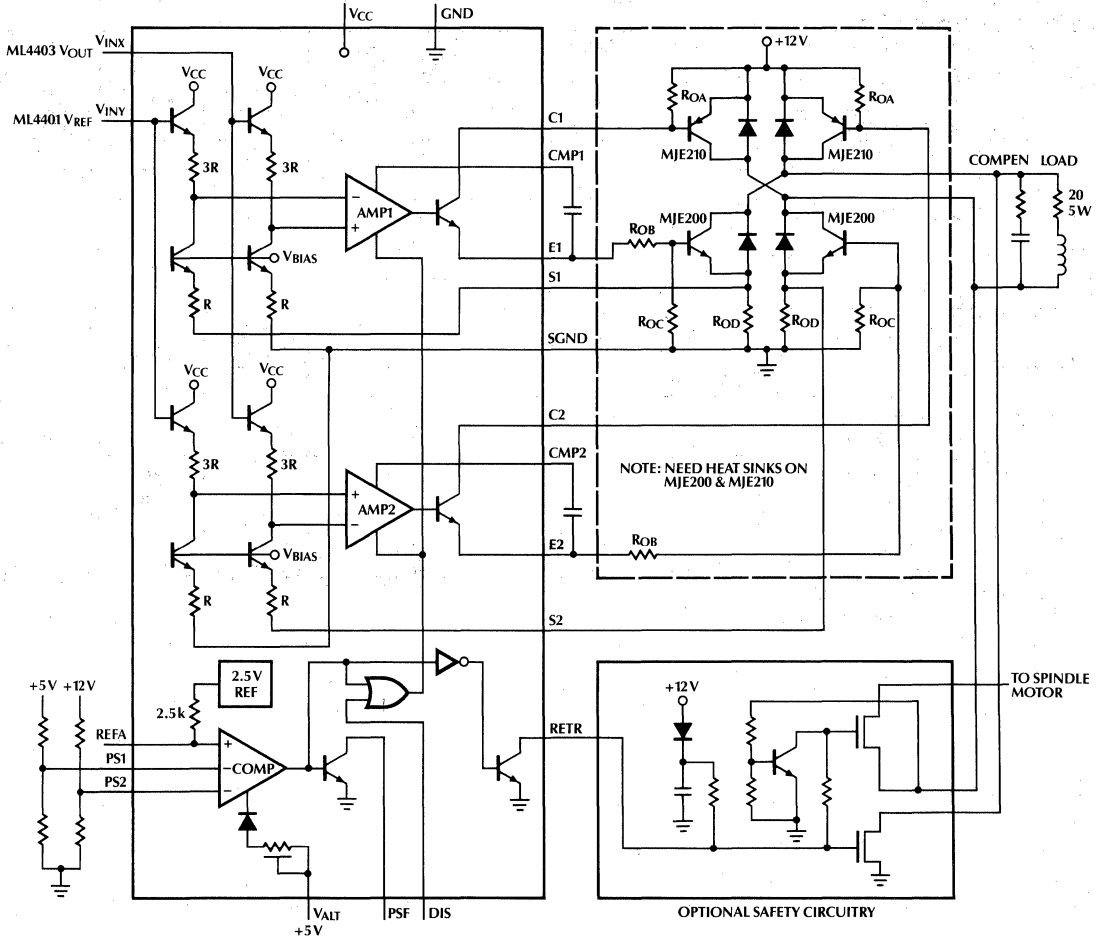
SYMBOL	PARAMETER	CONDITIONS	MIN	TYP NOTE 2	MAX	UNITS
Internal Voltage Reference (V_{REF})						
PS_{MIN}	Minimum Allowable V_{CC} Voltage	Where $V_{REF} > 2.48\text{V}$	4.75			V
V_{REF}	V_{REF} Voltage	$T_J = 25^\circ\text{C}$	2.44	2.55	2.66	V
T_{REG}	V_{REF} Thermal Stability	Over Specified Range		50		ppm/ $^\circ\text{C}$
R_{REF}	R_{REF} Resistance	(Internal Resistor from V_{REF} to Pin REFA)		2.55		k Ω
Comparator						
V_{OS}	Input Offset Voltage, any Two Inputs		-30	5	30	mV
I_{IN}	Input Bias Current		-0.5	-0.1	0	μA
V_{OL}	PSF Logic 0 Voltage	$I_{OL} = 1.6\text{mA}$	0	0.2	0.4	V
V_{OL}	RETR Logic 0 Voltage	$I_{SINK} = 3\text{mA}$	0	0.5	1	V
I_{OH}	PSF Logic 1 Leakage Current	$V_{PSF} = 5\text{V}$	-10	0.2	10	μA
I_{OH}	RETR Logic 1 Leakage Current	$V_{RETR} = 2\text{V}$	-20	0.05	20	μA
Amplifier Disable Section						
V_{IH}	DIS Logic High Voltage		2.0			V
I_{IH}	DIS Logic High Current	$V_{IH} = 2.4\text{V}$	-20		20	μA
V_{IL}	DIS Logic Low Voltage				0.8	V
I_{IL}	DIS Logic Low Current	$V_{IL} = 0.4\text{V}$	-20		20	μA

Note 1: 0°C to 70°C operating temperature range devices are 100% tested with temperature limits guaranteed by 100% testing, sampling, or by correlation with worst-case test conditions.

Note 2: Typicals are parametric norm at 25°C .

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TYPICAL APPLICATION DIAGRAM



ORDERING INFORMATION

PART NUMBER	TEMP. RANGE	PACKAGE	COMMENTS
ML4402-1CP	0°C to +70°C	MOLDED DIP (P20)	Input Offset = $\pm 5mV$
ML4402-1CQ	0°C to +70°C	MOLDED PCC (Q20)	Input Offset = $\pm 5mV$
ML4402-2CP	0°C to +70°C	MOLDED DIP (P20)	Input Offset = $\pm 10mV$
ML4402-2CQ	0°C to +70°C	MOLDED PCC (Q20)	Input Offset = $\pm 10mV$

GENERAL DESCRIPTION

The ML4403/4413 Servo Controller provides analog circuitry used in high performance trajectory and position control system for disk drive transducer heads. As a part of a head positioning servo system, this bipolar monolithic chip is designed to accept quadrature position signals and generate a servo error signal. While designed for minimum track access time, the ML4403/4413 supports a wide range of system designs.

Trajectory control functions include a track crossing detector, a velocity signal generator, and a velocity event detector. System stability and short settling time is insured by the interpolator function, which generates a ramp signal used to smooth the external position DAC output.

Position control is provided by a signal error amplifier within the device. When used with the ML4401/4431 Servo Position Demodulator, the track selection is performed by ML4401/4431 peak detector timing. This selection method eliminates track to track voltage offset problems and allows minimum track spacing. The ML4413

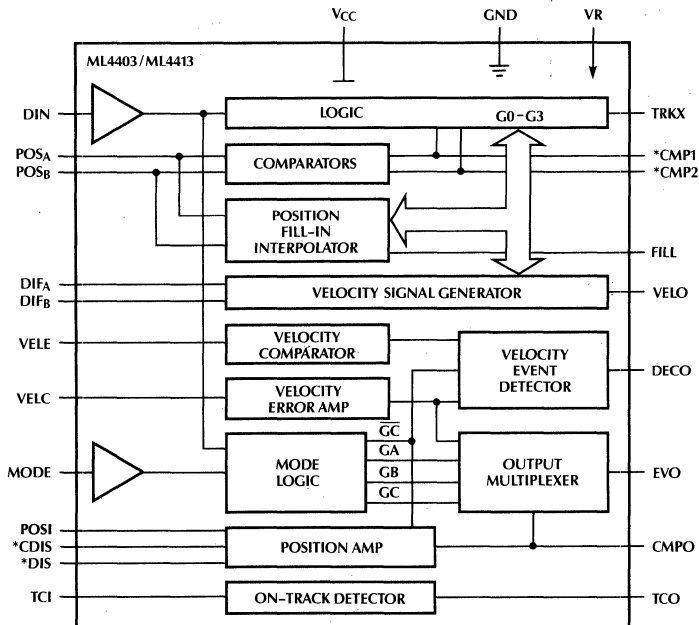
FEATURES

- Interpolate function smooths trajectory curve
- Flexible architecture allows user defined loop response
- Provides minimum track access time and maximum track density
- Single +12V power supply
- Compatible with ML4401 Servo Demodulator, ML4402, ML4406/07/08 Servo Driver and ML4404 Trajectory Generator

has a discharge function that enables zeroing of the external error amplifier compensation. This feature further reduces position settling time. An on-board on-track detector is provided which is used as a safety alarm by the controller for an off-track condition.

The ML4403/4413 Servo Controller, when combined with the ML4401 Servo Demodulator, the ML4402, ML4406/07/08 Servo Driver and the ML4404 Trajectory Generator provides a flexible closed-loop servo control system.

BLOCK DIAGRAM

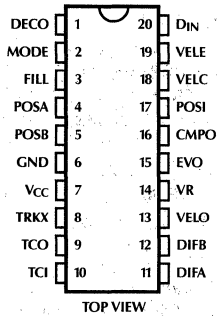


* PINS WITH THE * IN FRONT ARE USED ON THE ML4413 ONLY.

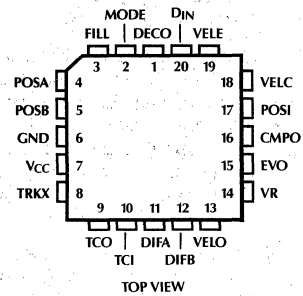
ML4403, ML4413

PIN CONNECTIONS

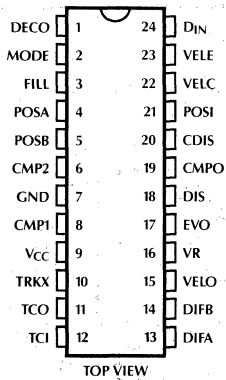
ML4403 20-PIN DIP (Prototypes only)



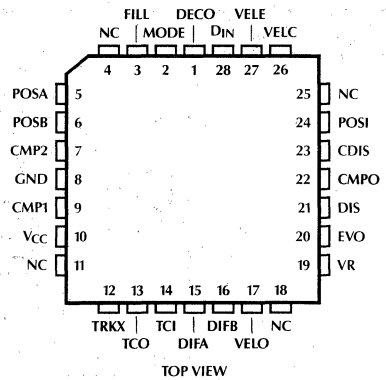
ML4403 20-PIN PLCC



ML4413 24-PIN SKINNY DIP (Prototypes only)



ML4413 28-PIN PLCC



NC = NO CONNECTION

PIN DESCRIPTION

ML4403 DIP and PLCC	ML4413 DIP PLCC		NAME	FUNCTION
1	1	1	DECO	Digital output from the velocity event detector. In application, this output goes to a logic high when the actual actuator velocity reaches the trajectory curve. It remains high through the "braking" or negative acceleration. This pin goes low when velocity is zero and remains low during actuator acceleration. This pin is only allowed to go high during access mode. This output is open collector and requires an external pull-up resistor.
2	2	2	MODE	Digital input used to select Hold mode (low level) or Access mode (high level).
3	3	3	FILL	Analog output that provides a sawtooth waveform that, when summed with stair-step output of the external DAC, provides a smooth trajectory curve. Refer to Figure 3.
4	4	5	POSA	Analog input for quadrature position signals from demodulator (ML4401/4431). Low pass prefiltering is recommended to eliminate peak detector ripple and external noise.
5	5	6	POSB	
6	7	8	GND	Device ground connection.
	8	9	CMP1	Digital outputs that can be used for various control and count schemes. These pins are only available on the ML4413. Timing is shown in Figure 3. These outputs are open collector outputs with an internal pull-up resistor tied to +5V.
	6	7	CMP2	
7	9	10	V _{CC}	+12V power supply connection.
8	10	12	TRKX	Digital output that provides a logic transition at each track crossing which is defined as the point midway between two tracks. Refer to Figure 3. This output is open collector with an internal pull-up resistor tied to +5V.
9	11	13	TCO	Digital output from the on-track detector. Used in Hold mode, this pin goes to logic high when the position signal exceeds an established window. This output is open collector with an internal pull-up resistor tied to +5V.
10	12	14	TCI	Analog input into the on-track detector. The input is normally derived from the position signal.
11	13	15	DIFA	Analog inputs for differentiated quadrature position signals. These inputs are used to generate the velocity signal at output VELO.
12	14	16	DIFB	
13	15	17	VELO	Analog output that provides a continuous velocity (tachometer) signal by time multiplexing/inverting the DIFA, DIFB input signals.
14	16	19	VR	Reference voltage input. This value should typically be +5V, which is obtainable from the V _{REF} output of ML4401/4431.
15	17	20	E _{VO}	Multiplexed analog output of both velocity error and position error signals. This output is used as the input for the servo actuator driving circuitry such as the ML4402.
	18	21	DIS	
16	19	22	CMPO	Digital input that, upon a logic high, electrically shorts pins CDIS and CMPO in order to keep the compensator capacitor discharged after entering hold mode. This pin and function is only available on ML4413. This function is used to reduce settling time when entering the Hold mode. Unlike pins MODE and DIN which float to logic high, this pin floats to logic low when left unconnected.
	20	23	C _{DIS}	Analog connection point for position compensation circuitry that is connected between this pin and POSI.
	21	24	POSI	Used to discharge external position compensation as shown in Figure 5. This pin is only available on ML4413. On the ML4403 this pin is internally connected to pin POSI.
17	21	24	POSI	Analog input for position control amplifier.
18	22	26	VELC	Analog input into velocity comparator. The velocity comparator trigger level is VR and is used for velocity event detection as described below.
19	23	27	VELE	Analog input for velocity error signal generated off-chip, referenced to VR.
20	24	28	D _{IN}	Digital input that controls actuator direction during Seek mode. This input affects the waveforms of outputs FILL, VELO, and E _{VO} . Refer to Figure 3.

ABSOLUTE MAXIMUM RATINGS

Absolute maximum ratings are those values beyond which the device could be permanently damaged. Absolute maximum ratings are stress ratings only and functional device operation is not implied. (All voltages referenced to GND.)

Power Supply Voltage, V_{CC}	14V
Terminal Voltage Range	
VR	-0.3 to 7.0V
POSI	-0.3 to $V_R + 0.3V$
DIN, POSA, POSB, DIFA, DIFB, VELE, VELC, MODE, DIS,	
TCl	-0.3 to $V_{CC} + 0.3V$
Storage Temperature Range	-65°C to +150°C
Junction Temperature	+125°C
Lead Temperature (Soldering, 10sec)	260°C

OPERATING CONDITIONS

Temperature Range	0°C to 70°C
Supply Voltage (V_{CC})	12V \pm 10%

ELECTRICAL CHARACTERISTICS

The following specifications apply over the recommended operating conditions of $V_{CC} = 10.8$ to $13.2V$, and $V_R = 5.0V$, unless otherwise specified. (See Note 1.)

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP NOTE 2	MAX	UNITS
Power Supply						
I_{CC}	V_{CC} Supply Current	Outputs unloaded		38	60	mA
DIGITAL INPUT/OUTPUT CHARACTERISTICS						
Inputs D_{IN} and Mode						
V_{IH}	Logic High Voltage		2.0			V
I_{IH}	Logic High Current	$V_{IH} = 2.4V$	-40	1	40	μA
V_{IL}	Logic Low Voltage				0.8	V
I_{IL}	Logic Low Current	$V_{IL} = 0.4V$	-100	-50	0	μA
Input DIS (ML4413 Only)						
V_{IH}	Logic High Voltage		2.0			V
I_{IH}	Logic High Current	$V_{IH} = 2.4V$	0	180	250	μA
V_{IL}	Logic Low Voltage				0.65	V
Outputs TCO and TRKX						
V_{OL}	Output Low Voltage	$I_{OL} = 1.6mA$	0		0.4	V
V_{OH}	Output High Voltage	$I_{OH} = 50\mu A$	2.4			V
t_{PD}	Propagation Delay	$C_L = 15pF$		200		ns
V_{TH}	Track Comparator Window	+ and - relative to VR	235	257	270	mV
Outputs CMP1 and CMP2 (ML4413 Only)						
V_{OL}	Output Low Voltage	$I_{OL} = 0.4mA$	0		0.4	V
V_{OH}	Output High Voltage	$I_{OH} = -50\mu A$	2.4			V
Output DECO						
V_{OL}	Output Low Voltage	$I_{OL} = 1.6mA$	0		0.5	V

ELECTRICAL CHARACTERISTICS (Continued)

The following specifications apply over the recommended operating conditions of $V_{CC} = 10.8$ to 13.2 V, and $V_R = 5.0$ V, unless otherwise specified. (See Note 1.)

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP NOTE 2	MAX	UNITS
ANALOG INPUT/OUTPUT CHARACTERISTICS						
Outputs Fill, VELO, COMPO, and EVO						
V_{OS1}	Input Offset Voltage EVO, FILL			2		mV
V_{OS2}	Input Offset Voltage COMPO		-5		5	mV
V_{OS3}	VELO Input Offset Voltage Tracking Between 4 Multiplex States	Variation in output level in 4 multiplex states with DIFA = DIFB = 5V	-10		10	mV
SR_1	Slew Rate FILL			4		V/ μ s
SR_2	Slew Rate COMPO, VELO, EVO			1		V/ μ s
V_{OUT}	Output Range All		1.0		9.0	V
I_{SRC1}	Source Current COMP, VELO, FILL		3			mA
I_{SRC2}	Source Current EVO		1.5			mA
I_{SNK1}	Sink Current FILL		0.25			mA
I_{SNK2}	Sink Current EVO, VELE		2			mA
I_{SNK3}	Sink Current COMPO		4			mA
Operational Amplifiers						
V_{OS}	Input Offset Voltage			2		mV
t_C	Average Temperature Coeff of Input Offset Voltage			20		μ V/ $^{\circ}$ C
I_{OS}	Input Offset Current			10		nA
I_B	Input Bias Current			100		nA
A_{VOL}	Open Loop Gain			200		V/mV
GBW	Gain Bandwidth Product			1		MHz
POSA, POSB Comparators						
V_{OS}	Input Offset Voltage			2		mV
V_{HYS}	Hysteresis			± 500		mV
t_C	Average Temp Coeff of Input Offset Voltage			20		μ V/ $^{\circ}$ C
I_{OS}	Input Offset Current			50		nA
I_B	Input Bias Current			500		nA
A_V	Voltage Gain			200		V/mV
P_d	Response Time			500		ns

Note 1: 0° C to $+70^{\circ}$ C operating temperature range devices are 100% tested with temperature limits guaranteed by 100% testing, sampling, or by correlation with worst-case test conditions.

Note 2: Typicals are parametric norm at 25° C.

FUNCTIONAL DESCRIPTION

Power Supply and Reference Requirements

The ML4403/4413 operates from a single $12V \pm 10\%$ power supply, a 5.0V reference is required at pin VR which is available from pin V_{REF} on the ML4401/4431. VR serves as a system reference or "analog ground".

Modes of Operation

The device has two modes of operation, Access and Hold mode, which are controlled by pin MODE. To accomplish this, pin MODE controls the output multiplexer that selects either the velocity or position error signal.

Access Mode

The head actuator servo system uses Access mode to move the recording head(s) from one data track to another. Access mode circuitry within the ML4403/4413 includes analog functions necessary to measure and control head actuator velocity. The head velocity is controlled in a fashion that provides for a fast track-to-track movement and minimum settling time, which results in minimum track access time.

Actuator Trajectory

Similar to racing to the next stop sign, the fastest way to move from one data track to the next is through maximum acceleration and maximum braking (negative acceleration). In a disk drive the acceleration, either positive or negative, is governed by maximum available actuator current. To do this in a controllable manner and land on the target track, an achievable "braking curve" or trajectory function is first defined. At the beginning of Access mode, maximum acceleration is applied until the head velocity reaches this defined braking curve. Following the velocity profile of the trajectory curve, controlled braking stops the head on the target track.

Unlike acceleration, velocity and distance are accurately measurable and therefore controllable parameters. The trajectory function, as shown in Figure 2, is therefore expressed as velocity (track crossing rate) vs. distance (tracks to go). The desirable constant positive and negative acceleration will result in the expression of velocity as a function of the square root of distance. Therefore generation of the trajectory curve, velocity vs. distance, requires a non-linear function.

Actuator Trajectory Generation

At the start of a track access cycle, initial tracks-to-go count is supplied by the microprocessor. As the head moves, the count is decremented by the ML4403/4413 track crossing detector. To generate the analog "desired velocity" signal required for braking control, the tracks-to-go count (distance variable) is converted through a DAC (Digital to Analog Converter) with a non-linear square function included either before or after the conversion. One common approach used to obtain this non-linear function is to pre-process the tracks-to-go count (or multiple thereof) in the microprocessor. This can be performed algorithmically by the use of a look up table.

An alternate method, as shown in the typical application of Figure 5 places the non-linear function after the DAC conversion. The tracks-to-go count is maintained by a simple discrete down-counter that is initialized by the microprocessor. To eliminate the DAC steps and provide a smooth distance signal, the DAC output is summed with the ML4403/4413's FILL output in the external summing amplifier shown. The FILL output generates a sawtooth wave, as shown in Figure 3. This distance signal is then passed through the non-linear trajectory generator which generates the "desired velocity" signal used during braking. Generating a smooth trajectory curve reduces electrical/mechanical system oscillation and target track settling time.

Inductance-caused actuator lag can also create a target track overshoot problem. The trajectory curve generator, as indicated, can be designed to allow the microprocessor to modify the non-linear function in a way to account for this lag. Refer to Figure 2. The amount of lag will depend on duration of braking. Braking duration can be correlated against acceleration duration which is indicated by the timing of pin DECO.

The track crossing detector, which drives the trajectory position counter (see Figure 5), is generated with external logic. The input comparators have a fixed amount of internal hysteresis to provide noise immunity and media dropouts. The CMP1 and CMP2 outputs on the ML4413 can be used to perform more sophisticated sequential track crossing detection schemes. This can further reduce the detector's susceptibility to media dropouts.

Hold Mode

At the end of an Access cycle, the head is stopped, or nearly so, on the target track. Hold mode is then selected to maintain accurate head positioning on that track. In this mode, the compensator output (CMPO) is multiplexed into the error amplifier output (EVO).

Track Selection

Track position is held by maintaining a zero value of the position input signal, with respect to VR. However, to allow selection of one of four track types and maintain error signal polarity, selection of POSA, POSB, or their inverse needs to be possible. Commonly this selection process is accomplished with an analog multiplexer-inverter matrix. The problem inherent with this approach is the track-to-track offset differences, caused by the amplifier input offset differences within the matrix.

The track selection scheme adopted by the ML4401/4431 and ML4403/4413 combination performs the multiplexing within the ML4401/4431. The selection/inversion operation is performed with the external support logic of the ML4401/4431 by changing the peak detector sample timing. This method eliminates the offset problems and allows a higher track density.

Position Amplifier and Compensator Zeroing

During track following mode (mode low), the compensator amplifier acts as an integrator which nulls out the position error. The timing of the transition between access mode and track follow is critical to minimize settling time. The velocity at which this transition occurs can be externally set by resistor RCMP (see Figure 5). During seek mode, the large compensator capacitor (CCMP) is discharged through an internal switch, so that the integrating loop sees no initial charge at the beginning of track follow mode. This can reduce settling time by several milliseconds.

The ML4413 provides a further enhancement of this feature. The switch can remain closed after the beginning of the seek-to-track follow transition by holding pin DIS high. In this way, the time at which the logic switches modes, and the time that integration begins can be controlled independently, and further settling time reduction can be achieved.

On-Track Detector

The on board on-track detector is a window comparator that provides a digital alarm of an off-track condition. This feature is useful as a safety to prevent data transfer during an off-track condition that may occur during track settling or mechanical jarring.

Velocity Control

As a necessary element of velocity control, a velocity signal is generated and is output at pin VELO. To

accomplish this, the quadrature position signals are first differentiated through external RC networks and then input into pins DIFA and DIFB. The ML4403/4413 then time multiplexes these differentiated signals to obtain a continuous velocity signal that is output at pin VELO. It is important to note that the trajectory generator shown in Figure 5 generates a "desired velocity" signal positive with respect to VR, and that VELO creates a negative signal with respect to VR. This allows the use of a simple external resistor bridge to create the velocity error signal.

The summing function can be modified, as illustrated, by the action of pin DECO when the actual velocity reaches the trajectory curve. Modification can also be made just prior to that time with the "optional trajectory overshoot compensation" circuit, shown in Figure 5, that prevents overshoot due to actuator motor inductance.

Inductance-caused actuator lag can also create a track overshoot problem. The trajectory curve generator, as indicated, can be designed to allow the microprocessor to modify the non-linear function in a way to account for this lag as shown in Figure 2. The amount of lag will depend on duration of braking. Braking duration can be correlated against acceleration duration which is indicated by the timing of pin DECO.

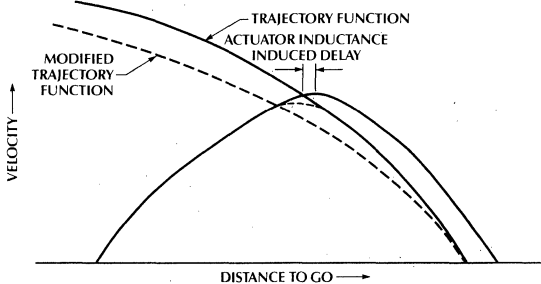


Figure 2.

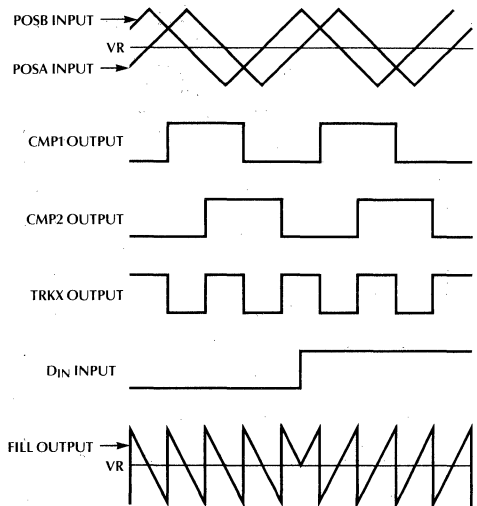
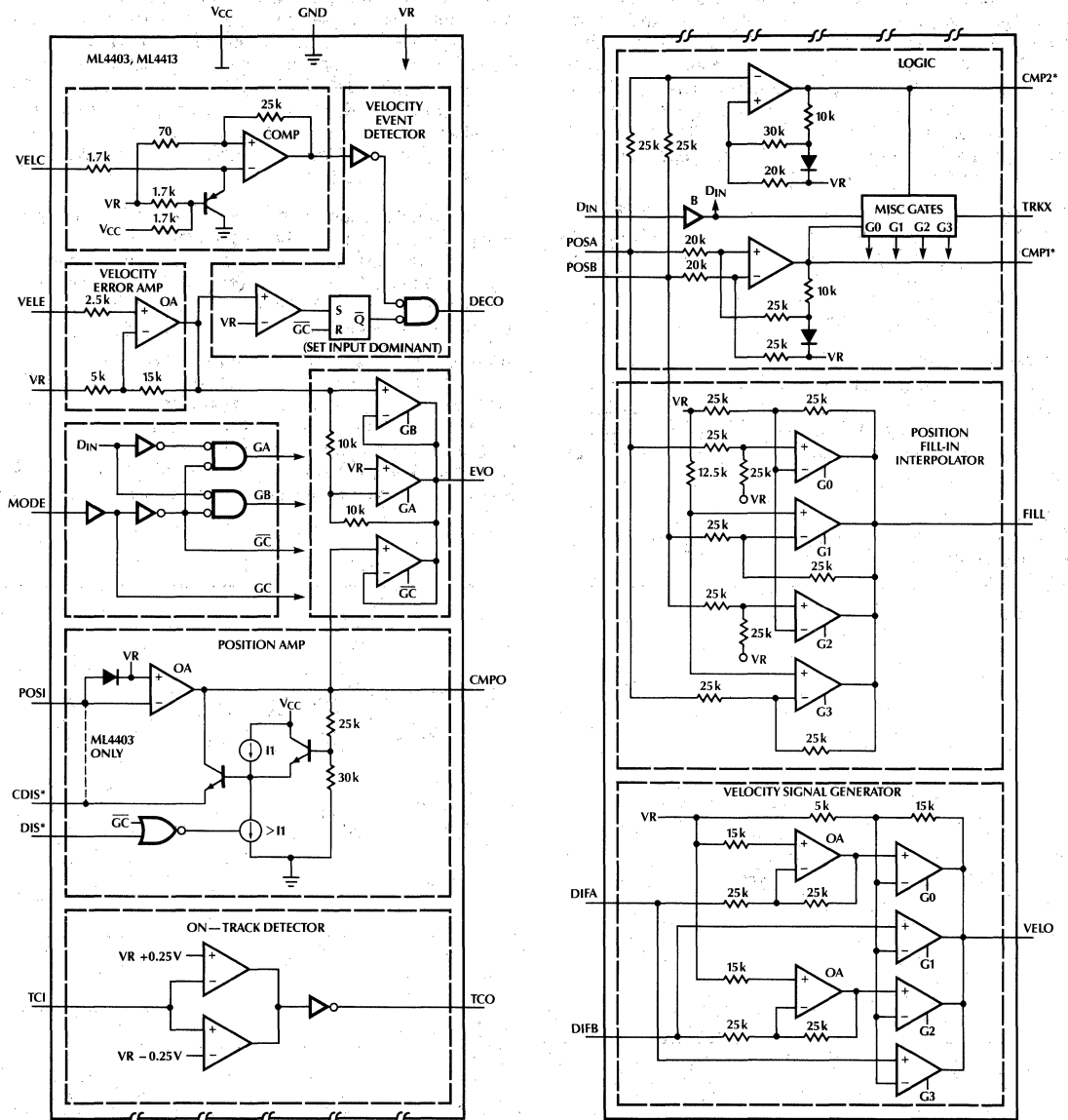


Figure 3.



*NOTE: THESE PINS ONLY AVAILABLE ON ML4413

Detailed Function of Block Diagram of the ML4403/13

Figure 4.

ML4403, ML4413

ORDERING INFORMATION

PART NUMBER	TEMP. RANGE	PACKAGE
ML4403CP	0°C to +70°C	MOLDED DIP (P20)
ML4403CQ	0°C to +70°C	MOLDED PCC (Q20)
ML4413CP	0°C to +70°C	MOLDED DIP (P24)
ML4413CQ	0°C to +70°C	MOLDED PCC (Q28)

Trajectory Generator

GENERAL DESCRIPTION

The ML4404 Trajectory Generator provides the trajectory function for time optimal head positioning systems. The ML4404 receives position and velocity information from a servo controller, such as the ML4403, and generates the desired time optimal velocity trajectory. Desired Velocity is then compared with the actual velocity to create the error signal used by the servo controller.

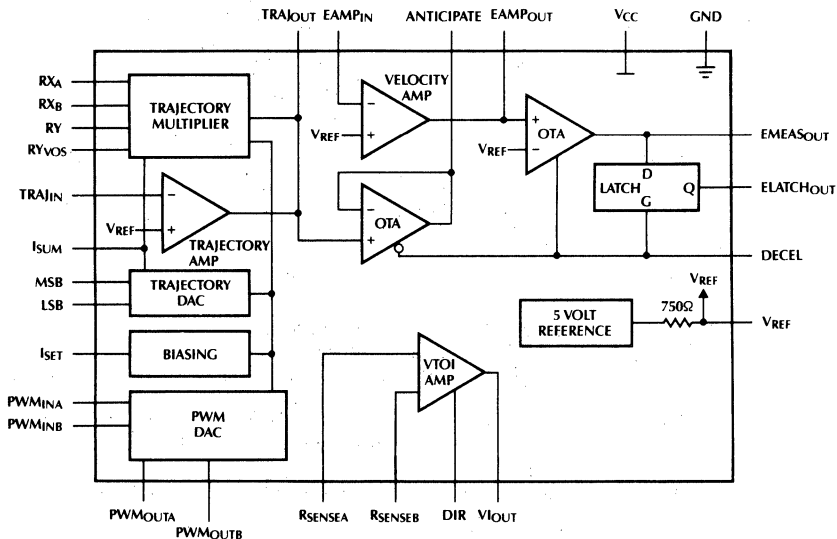
An anticipate function is included to compensate for phase shift error caused by actuator inductance. Another feature on the ML4404 is an error measure output which averages the velocity error during deceleration, so that the control system can monitor and adjust the necessary transducer gain for minimum access time.

The servo system usually requires accurate analog voltages to be set through software control. This is easily accomplished with a duty cycle to current translator function on the ML4404. By controlling the duty cycle of a TTL line, a processor can set an analog voltage on the translator output. These translators are fully independent blocks which can be used anywhere in the control system.

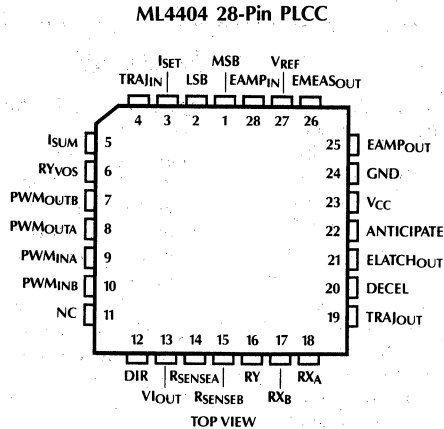
FEATURES

- Flexible architecture allows a user defined trajectory function
- Anticipate function compensates for phase delay caused by actuator inductance
- Feed forward function improves system stability
- Uncommitted PWM to current translators allow an analog voltage to be set with microprocessor control
- Single +12V power supply
- Compatible with Micro Linear's ML4401, ML4431 Demodulator, ML4402, ML4406/07/08 Driver, and ML4403, ML4413 Controller

BLOCK DIAGRAM



PIN CONNECTIONS



PIN DESCRIPTION

PIN #	NAME	FUNCTION
1	MSB	Pulse width modulated (PWM) DAC TTL input (active low). The DAC output current is the position input to the trajectory generator. The MSB/LSB ratio is 8/1. The duty cycle of these two TTL inputs are controlled by a processor to form an 8-bit PWM DAC. The 3 higher order bits are modulated into the MSB.
2	LSB	Pulse width modulated (PWM) DAC TTL input (active low). The 5 lower order bits are modulated into the LSB input.
3	ISET	A resistor (R_{SET}) from this pin to V_{REF} sets the internal bias levels. $I_{BIAS} = 3V/R_{SET}$. The nominal value should be between 0.25 and 0.5 mA.
4	TRAJIN	The trajectory generator input. This node is connected through an external filter to the sum of the PWM DAC output and the multiplier output.
5	ISUM	The trajectory DAC output which is summed with the multiplier output feedback. An external RC filter network from this pin to TRAJIN smooths out PWM DAC ripple.
6	RYVOS	Nulls out the offset of the trajectory curve at the origin. A resistor equal to R_Y is connected from this pin to V_{REF} . This pin is available only on the ML4404.
7	PWMOUTB	PWM to current translator output.
8	PWMOUTA	PWM to current translator output.
9	PWMINA	TTL input for the PWM to current translator. This converter translates a signal's duty cycle to an analog voltage.
10	PWMINB	TTL input for the PWM to current translator. This converter translates a signal's duty cycle to an analog voltage.
11	NC	No Connection.
12	DIR	TTL direction input from the processor. Controls the polarity of the V/I converter output.

PIN DESCRIPTION (Continued)

PIN #	NAME	FUNCTION
ML4404		
13	V _I OUT	The V/I converter output. This circuit block monitors the differential voltage across the sense resistors of an actuator driver (such as the ML4402) and converts it to a bidirectional current whose scale factor is set by two external resistors. This current can be used to compensate for a noise reducing low pass filter in the output of the velocity transducer so that there is no net phase shift in the velocity signal.
14	R _{SENSE} A	A gain setting resistor is connected from this input to the sense resistor on the bridge driver.
15	R _{SENSE} B	A gain setting resistor is connected from this input to the sense resistor on the bridge driver.
16	RY	A resistor (RY) is connected from this pin to V _{REF} . RY and RX set the second order term of the trajectory curve.
17	RX _B	A resistor (RX) is connected between RX _A AND RX _B to set the second order term in the trajectory curve.
18	RX _A	A resistor (RX) is connected between RX _A and RX _B to set the second order term in the trajectory curve. An additional resistor (RK3) can be connected from RX _A to either the trajectory output (TRAJ _{OUT}) or to V _{REF} to set the third order term.
19	TRAJ _{OUT}	The trajectory output. This voltage relative to V _{REF} is proportional to the desired velocity. A resistor and capacitor from this pin to TRAJ _{IN} sets the first order term and the loop compensation.
20	DECEL	Decelerate mode TTL input from the servo controller (such as the ML4403). When low (during accelerate) the anticipate output becomes a voltage follower, the error measure output is a high impedance, and the error sign is latched. When high (during deceleration) anticipate goes to high impedance, error measure integrates the velocity error, and the error sign latch is transparent.
21	ELATCH _{OUT}	The latched sign of the access loop error during deceleration. This TTL output can be used by the processor to adjust the velocity transducer gain to match that required by the mechanical system.
22	ANTICIPATE	Modifies the trajectory curve during acceleration and the accelerate to decelerate transition. This accounts for the time delay error caused by the actuator inductance.
23	V _{CC}	+12V power supply.
24	GND	Ground.
25	EAMP _{OUT}	Error amplifier output. The positive trajectory output (desired velocity) is summed with the negative velocity input (actual velocity) to form a difference output. The velocity input comes from the servo controller (such as the ML4403).
26	EMEAS _{OUT}	Error measure output. This output averages the value of the access loop error during deceleration.
27	V _{REF}	The analog zero reference point. This pin is intended to be driven with the ML4401 V _{REF} output. The ML4404 has an internal 5V reference connected through a current limiting resistor to this pin so that standalone operation/evaluation is available.
28	EAMP _{IN}	Error amplifier input. The INPUT summing node for the trajectory and velocity signals.

ML4404

ABSOLUTE MAXIMUM RATINGS

Absolute maximum ratings are those values beyond which the device could be permanently damaged. Absolute maximum ratings are stress ratings only and functional device operation is not implied. (All voltages referenced to GND.)

Power Supply Voltage, V_{CC}	14V
V_{REF} , $R_{SENSE A}$, $R_{SENSE B}$	-0.3 to +7V
TTL Inputs, I_{BIAS} , $ELATCH_{OUT}$	-0.3 to +7V
$PWM_{OUT A}$, $PWM_{OUT B}$, $PWM_{OUT C}$	-0.3 to $V_{CC} + 0.3V$
Anticipate, V_{IOUT} , $EAMP_{OUT}$	-0.3 to $V_{CC} + 0.3V$
$EAMP_{IN}$, $TRAJ_{IN}$	-0.3 to $V_{CC} + 0.3V$
I_{SUM}	$V_{REF} - 1$ to $V_{REF} + 1V$
$TRAJ_{OUT}$, RX_A , RX_B , RY , RY_{VOS}	$V_{REF} - 1$ to $V_{CC} + 0.3V$
Storage Temperature Range	-65°C to +150°C
Junction Temperature (T_j)	+125°C
Lead Temperature (Soldering, 10sec)	260°C

OPERATING CONDITIONS

(See Figure 7)

Temperature Range	0°C to 70°C
Supply Voltage (V_{CC})	12V \pm 10%
R_{SET}	6.2K
RK1	110K
RY	3K
RY_{VOS}	3K
RX	6.8K
RK3	12K to V_{REF}
CCOMP	56pF
CF1	0.0022 μ F
CF2	Open
RF	10K
R_{LOAD}	20K to V_{REF}

ELECTRICAL CHARACTERISTICS

The following specifications apply over the recommended operating conditions of $V_{CC} = 10.8$ to 13.2V, $V_{REF} = 5V$, $T_A = 0$ to 70°C, $R_{FILT} = 10K$, $RY = RY_{VOS} = 3K$, $RK1 = 100K$, $RX = 6.8K$, $R_{SET} = 6.2K$ and $RK3 = 12K$ to V_{REF} unless otherwise specified. (See Note 1.)

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP NOTE 2	MAX	UNITS
DC CHARACTERISTICS						
I_{CC}	Power Supply Current			23	35	mA
DIGITAL INPUT/OUTPUT CHARACTERISTICS						
(Inputs PWM_{INA}, B, C, MSB, LSB, DIR)						
V_{IH}	Logic High Voltage		2.0			V
I_{IH}	Logic High Current	$V_{IH} = 2.4V$	-40	10	40	μ A
V_{IL}	Logic Low Voltage				0.8	V
I_{IL}	Logic Low Current	$V_{IL} = 0.4V$	-40	1	40	μ A
Inputs (DECEL)						
V_{IH}	Logic High Voltage		2.0			V
I_{IH}	Logic High Current	$V_{IH} = 2.4V$	-250	5	40	μ A
V_{IL}	Logic Low Voltage				0.8	V
I_{IL}	Logic Low Current	$V_{IL} = 0.4V$	-1600	-850	0	μ A
Outputs ($ELATCH_{OUT}$)						
V_{OL}	Output Low Voltage	$I_{OL} = 1.6mA$	0	0.3	0.4	V
V_{OH}	Output High Voltage	$R_L = 5K$ to V_{REF}	2.4	5.0	5.5	V
Trajectory Amplifier (See Note 3)						
I_B	Input Bias Current		0	7	20	nA
A_V	Open Loop Gain			75k		V/V
BW	Unity Gain Bandwidth			1		MHz
PHIM	Phase Margin			75		DEG
Velocity Error Amplifier						
V_{OS}	Input Offset Voltage		-10		10	mV
I_B	Input Bias Current		0	50	300	nA
A_V	Open Loop Gain			120k		V/V
I_{SOURCE}	Source Current		4	8		mA
I_{SINK}	Sink Current		1	2		mA
BW	Unity Gain Bandwidth			1		MHz
PHIM	Phase Margin			75		DEG
V_{OUT}	Output Voltage Range		0.5		$V_{CC} - 3$	V

ELECTRICAL CHARACTERISTICS (Continued)

The following specifications apply over the recommended operating conditions of $V_{CC} = 10.8$ to $13.2V$, $V_{REF} = 5V$, $T_A = 0$ to $70^\circ C$, $R_{FILT} = 10K$, $R_Y = R_{Y_{VOS}} = 3K$, $R_{K1} = 100K$, $R_X = 6.8K$, $R_{SET} = 6.2K$ and $R_{K3} = 12K$ to V_{REF} unless otherwise specified. (See Note 1)

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP NOTE 2	MAX	UNITS
Biasing						
V_{IS}	I_{SET} Voltage		2.00	2.02	2.06	V
PWM to Current Translators						
	I_{CHARGE}/I_{SET} , I_{DIS}/I_{SET}	$PWM_{OUT} = 5.0V$		0.98		mA/mA
	I_{CHARGE}/I_{DIS}	$PWM_{OUT} = 5.0V$	0.910	0.99	1.10	mA/mA
V_{OUT}	Output Voltage Range		1.5		9	V
Transconductance Amps						
V_{OS}	Input Offset Voltage		-10		10	mV
gm	Transconductance			1700		$\mu Mhos$
I_{OUTMAX}	Max Output Current			90		μA
I_B	Input Bias Current			4.5		μA
Latch/Comparator						
I_B	Input Bias Current		0	2	10	μA
V_{OS}	Input Offset Voltage	$V_{TRAJOUT}$ @ 5V	-10		10	mV
A_V	Open Loop Gain			15k		V/V
t_{pd}	Propagation Delay	$C_L = 10pF$, $R_L = 2K$ to V_{REF}		60		ns
V/I Amp						
$I_{OS}/I_{SENSE} * 100$	Sense Current Offset	$I_{SENSE A} = I_{SENSE B}$ $0.1mA < I_{SENSE} < 1mA$ $V_{IOUT} = 5V$	-2	0	2	%
V_{SMAX}	Max R_{SENSE} Voltage		0.5	0.64		V
V_{OUT}	Output Range		1.8		9	V
Trajectory DAC						
I_{MSB}/I_{SET}	MSB Current			0.98		mA/mA
I_{MSB}/I_{LSB}	MSB to LSB Ratio		78	8.00	8.5	mA/mA
Trajectory Multiplier (Note 4)						
V_{OS}	$V_{OUT} - V_{REF}$ at Origin	V_{OUT} at $I_{SUM} = 0$	-5		5	mV
	$V_{TRACK1} : V_{TRACK32}$	$(V_{OUT}$ at $I_{LSB}/32)/(V_{OUT}$ at $I_{LSB})$	0.090		0.140	mV/mV
	$V_{TRACK2} : V_{TRACK32}$	$(V_{OUT}$ at $I_{LSB}/16)/(V_{OUT}$ at $I_{LSB})$	0.165		0.205	mV/mV
	$V_{TRACK4} : V_{TRACK32}$	$(V_{OUT}$ at $I_{LSB}/8)/(V_{OUT}$ at $I_{LSB})$	0.270		0.320	mV/mV
	$V_{TRACK8} : V_{TRACK32}$	$(V_{OUT}$ at $I_{LSB}/4)/(V_{OUT}$ at $I_{LSB})$	0.430		0.470	mV/mV
	$V_{TRACK16} : V_{TRACK32}$	$(V_{OUT}$ at $I_{LSB}/2)/(V_{OUT}$ at $I_{LSB})$	0.660		0.695	mV/mV
V_{LSB}	$V_{TRACK32}$	V_{OUT} at $I_{SUM} = I_{LSB}$	0.935		1.035	V
V_{CROSS}	Crossover Error	$(V_{OUT}$ at $I_{SUM} = I_{LSB}) - (V_{OUT}$ at $I_{SUM} = I_{MSB}/8)$	-25		+10	mV
	$V_{TRACK32} : V_{TRACK256}$	$(V_{OUT}$ at $I_{MSB}/8)/(V_{OUT}$ at $I_{MSB})$	0.305		0.325	mV/mV
	$V_{TRACK64} : V_{TRACK256}$	$(V_{OUT}$ at $I_{MSB}/4)/(V_{OUT}$ at $I_{MSB})$	0.450		0.470	mV/mV
	$V_{TRACK128} : V_{TRACK256}$	$(V_{OUT}$ at $I_{MSB}/2)/(V_{OUT}$ at $I_{MSB})$	0.670		0.685	mV/mV
	$V_{TRACK192} : V_{TRACK256}$	$(V_{OUT}$ at $I_{MSB} * 3/4)/(V_{OUT}$ at $I_{MSB})$	0.840		0.860	mV/mV
V_{MSB}	$V_{TRACK256}$	V_{OUT} at $I_{SUM} = I_{MSB}$ (Full Scale + 1)	3.070		3.225	V
PSRR	Supply Rejection	at Origin at Full Scale		0.2 2		mV/V mV/V

Note 1: $0^\circ C$ to $+70^\circ C$ operating temperature range devices are 100% tested with temperature limits guaranteed by 100% testing, sampling, or by correlation with worst-case test conditions.

Note 2: Typicals are parametric norm at $25^\circ C$.

Note 3: Minimum recommended load resistor is $10k\Omega$ from the trajectory output to V_{REF} .

FUNCTIONAL DESCRIPTION

Power Supply and Reference Requirements

The ML4404 operates from a single 12V power supply. In addition, a 5.0V reference is required at pin V_{REF} which is available from the ML4401, ML4431. V_{REF} serves as a system reference or "analog ground".

Biasing

All of the critical internal biasing on the ML4404 is set as a function of an external resistor, R_{SET} . An internal feed-back loop forces the voltage on I_{SET} (pin 3) to be 2.0V. R_{SET} is connected from this pin to V_{REF} (5.0V) and the resulting current is used in the multiplier and duty cycle-to-current translators.

$$I_{BIAS} = (V_{REF} - 2) / R_{SET} = 3V / R_{SET}$$

The nominal value of I_{BIAS} should be between 0.25 and 0.50mA.

Trajectory Multiplier/Amplifier

The Trajectory Multiplier/Amplifier generates the optimal velocity output from the position-to-go input. The optimal velocity is the TRA_{OUT} voltage relative to V_{REF} . The input position is set by the duty cycle of the MSB and LSB-inputs.

During an access operation, the actuator is first driven to maximum acceleration, and then into braking or deceleration. To minimize access times the trajectory curve (velocity vs position) during deceleration must be accurately controlled so that the head stops exactly on the desired track (without overshooting or undershooting). The head is driven to maximum acceleration until this braking curve is reached. Then the velocity is controlled to follow this optimal trajectory during deceleration.

According to the theory for a second order system, time optimal access is achieved if the position is proportional to the square of velocity ($P = KV^2$, or $V = KP^{1/2}$). However, in the real system environment, this theory needs modification in two important areas. First, as shown in Figure 1, the slope of the trajectory at the origin (final position) is infinite for a pure square root function. This infinite slope would result in an infinite bandwidth servo loop. As a result, a first order linear term needs to be included which will reduce the slope of this curve near the origin. Second, at large velocity, the square root function is not quite optimal, due to non-zero actuator inductance. A higher order term to modify the curve in this region needs to be included.

This trajectory curve, with its first, second, and third order terms is implemented with a multiplier in the feedback loop of an opamp (see Figure 2). The position input is a current which is a fraction of I_{BIAS} , as discussed in the trajectory DAC section below. The first order term is implemented with a feedback resistor (RK1) directly in the feedback path of the op amp. This transfer function of this I/V converter is expressed by $V_{OUT} = RK1 * I_{IN}$. The second order term comes from the multiplier.

$$I_{IN} = (V_{OUT}^2) / (2.25 I_{BIAS} * RX * RY)$$

With both terms together,

$$I_{IN} = \frac{V_{OUT}}{RK1} + \frac{V_{OUT}^2}{2.25 I_{BIAS} * RX * RY}$$

The first order term dominates near the origin, and the second order term dominates, at higher velocities.

The multiplier is modified by the addition of a resistor (RK3) which results in the third order term. This resistor is connected from RXA (pin 18) to either TRA_{OUT} or V_{REF} . As shown in Figure 3, the shape of the trajectory curve for large velocities can be adjusted in either direction from nominal, depending on which pins RK3 is connected between. It should be noted that the above equations are only approximate. The actual multiplier transfer function is not a pure second order function, even with RK3 unconnected. The multiplier is designed to approximate the required trajectory for most typical servo systems. For most applications, very little correction with RK3 will be required. On the ML4404, an additional resistor (RY_{VOS}) equal to RY can be included which nulls the offset of the curve near the origin.

Since the resistors R_{SET} , RK1, RX, RY, RK3, and RY_{VOS} are all external, any desired trajectory can be set. The constraints on these values are as follows:

$$\begin{aligned} V_{OUTMAX} &< 1.5 * RX * I_{BIAS} \\ V_{OUTMAX} &> 1.5 * RY * I_{BIAS} \\ V_{OUTMAX} &< 3.5V \\ 6k &< R_{SET} < 12k \text{ for } V_{REF} = 5V \\ R_{FILT} &< RK1 / 10 \\ R_{FILT} * C_{FILT2} &< RK1 * C_{COMP} \\ R_{LOAD} &> 10k \text{ to } V_{REF} \end{aligned}$$

V_{OUTMAX} is the maximum trajectory output voltage (relative to V_{REF}).

Trajectory DAC

The trajectory DAC creates a position input for the trajectory multiplier. The position input is controlled by the duty cycle of the TTL inputs MSB and LSB. For most applications the position information will be digitally encoded to 8-bit resolution — each code representing one track. Therefore, the full scale input of the trajectory curve is 255 tracks. The input current corresponding to this full scale is I_{BIAS} .

Since the duty cycle of a single line is difficult to control to 0.4% (1/256), the duty cycle to input current translator section is divided into 2 signals (MSB and LSB). As shown in Figure 4, the ratio between these two currents is 8/1. The 5 lower order bits of code are modulated into the LSB input, and the 3 higher order bits into the MSB input. For example, a position input of 32 tracks would correspond to the MSB line always inactive, and the LSB line always active. 256 tracks would be MSB always active, and the LSB always inactive. 1 track would be MSB always inactive, and LSB 1/32 of the period active.

In general for an 8-bit binary code D7 D6 D5 D4 D3 D2 D1 D0 where D7 is the high order bit, the active duty cycle for the MSB input is D7 D6 D5/8 and the active duty cycle for the LSB input is D4 D3 D2 D1 D0 / 32. For example 10100011 (163 tracks) would be active 5/8 of a period on the MSB, and 3/32 of a period on the LSB.

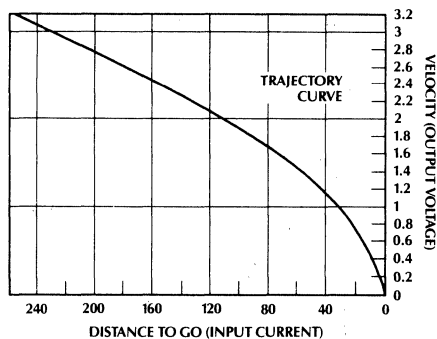


Figure 1.

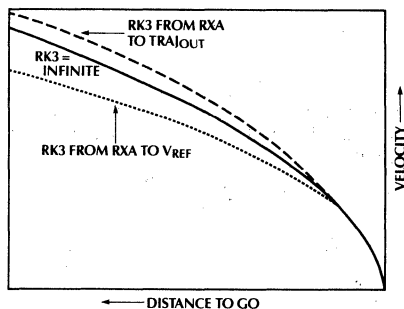


Figure 3.

AC Considerations — Trajectory Amp/Multiplier

The AC response of the trajectory output is primarily controlled by the external components C_{FILT1} , C_{FILT2} , R_{FILT} , $RK1$, and C_{COMP} . Four parameters must be considered to determine the values of these components.

First, the value of $RK1$ is set based on the bandwidth of the servo loop. $RK1$ sets the gain of the trajectory function at the origin. As the system bandwidth increases, more gain is required near the origin, and the value of $RK1$ increases.

Second, the exponential decay rate of the trajectory output during deceleration must be fast enough for the mechanical system to dominate the loop response. As the head approaches the target track, the multiplier (2nd and 3rd order terms) becomes less significant, and the first order term dominates. In this region, the exponential decay is dominated by the $RK1 * C_{COMP}$ product. This product should be set at a frequency which is a few times higher than that of the position loop bandwidth, so that the overall phase margin is not significantly degraded.

Third, the filter components should be set such that the ripple from the trajectory DAC is minimized. Once the values for C_{COMP} and $RK1$ have been set, then the values of the remaining components, C_{FILT1} , C_{FILT2} , and R_{FILT} , can be determined. As the capacitance and resistance of these components increase the PWM ripple from the trajectory DAC is reduced. Due to the nonlinear nature of this circuit block, a mathematical analysis of the ripple is quite cumbersome, so the values of these components are best chosen

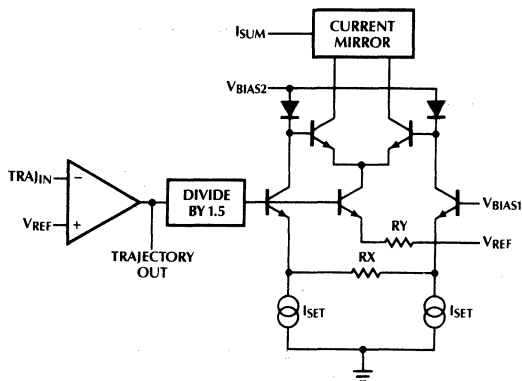


Figure 2.

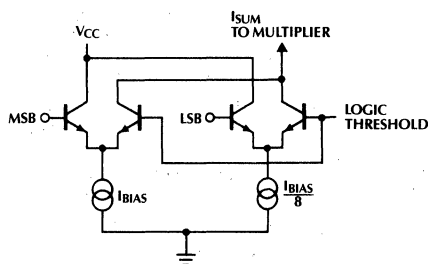


Figure 4.

empirically. The PWM ripple is dependent on R_{FILT} , C_{FILT1} , C_{FILT2} , as well as the duty cycle (50% duty cycle on the MSB will result in the largest ripple), and the frequency of modulation (the ripple is proportional to the square of the period).

Fourth, the R_{FILT} - C_{FILT} combination must be set such that the dynamic response of the trajectory output does not overshoot during deceleration. Ideally, the RC combination should be set such that the system is critically damped to a maximum deceleration input.

Note that a tradeoff exists between the ripple amplitude and the transient response. Too large a value of R_{FILT} - C_{FILT} will cause an overshoot in the transient response, and a low ripple level. Too small a value will provide acceptable transient response, but a large ripple amplitude. A range of values exists for most applications which results in acceptable performance for both ripple and transient response.

Anticipate

The function of the anticipate block is to modify the trajectory curve during acceleration. This compensates for the actuator inductance delay during the accelerate-to-decelerate transition.

At the start of an access operation, the actuator is driven into acceleration. The actuator velocity increases until either the maximum velocity is reached, or the trajectory deceleration curve is reached. As shown in Figure 5, if the trajectory curve is reached first, then the actuator needs to be driven into deceleration so that the trajectory curve can be followed. This accelerate-to-decelerate transition requires that the

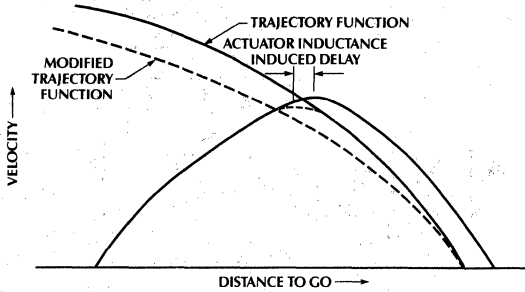


Figure 5.

current in the actuator be reversed. Since this cannot happen instantaneously (due to actuator inductance), a phase shift results. The actuator will then overshoot the desired trajectory, and miss the target track. However, if the curve can be modified (see Figure 5) such that the accelerate-to-decelerate transition begins before the nominal trajectory is reached, this overshoot problem can be eliminated. This function is implemented with a switched transconductance amp. During acceleration (DECEL input low), the anticipate output becomes a voltage follower—the anticipate signal is identical to the $TRAJ_{OUT}$ signal. An external resistor from anticipate to $TRAJ_{IN}$ will modify the trajectory curve as required. During deceleration, the anticipate output becomes a high impedance and the normal trajectory curve is followed.

In addition to the external resistor, an external capacitor to V_{REF} sets the anticipate decay time constant equal to the current reversal time for the actuator.

Velocity Error Amplifier

The function of this block is to subtract the desired velocity (from the trajectory output) with the actual velocity (from the servo controller) to create a velocity error output. This error output is multiplexed through the servo controller into the servo driver during access mode (see ML4403, ML4413 data sheet).

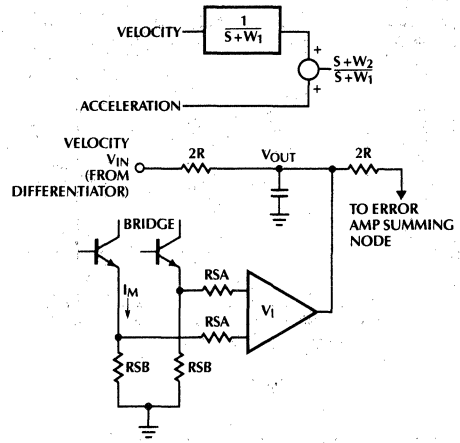
Since the polarity of the velocity input from the ML4403, ML4413 is the opposite of the trajectory output polarity, the difference function is accomplished with a summing amplifier, as shown in the application diagram. The summing resistors and the feedback resistor are external.

V/I Amp — Velocity Filter

The function of this block is to create a low noise velocity input. The velocity error amp requires that a clean, noise-free velocity input be compared with the desired velocity (from the trajectory output) to create a velocity error signal.

One way to create this velocity signal is to differentiate position. However, the differentiated signal will be noisy, and this noise can excite the mechanical resonances in the system. Another way to create velocity is to integrate acceleration. This eliminates the noise problem, however, the integrator DC accuracy will be poor due to the drift.

The ML4404 uses both of these techniques to achieve a low noise tachometer function which will operate at low frequencies noise from the mechanical resonances is attenuated by the RC filter. The filter output is then summed with a current proportional to acceleration.



I_M = ACTUATOR MOTOR CURRENT

Figure 6.

The V/I amp creates this current by monitoring the sense resistors in the bridge driver. The resulting transfer function has both a pole and a zero, and can be expressed by:

$$\frac{V_{OUT}}{V_{IN}} = \frac{R*(M*RSB/(KF*KT*RSA))*S + 1}{R*C*S + 1}$$

- Where KT = The velocity transducer gain (from differentiator)
- KF = Motor force constant
- M = Total moving mass

The above equation is a first order approximation which assumes that the frictional components of the system (such as windage) are negligible. If the pole and zero are set to identical frequencies, then an all pass function is achieved. To do this, first, set the pole ($W = 1/RC$) at a frequency much lower than the mechanical resonances. Then set the scaling resistors, RSA , such that:

$$RSA = M*RSB / (KF*KT*C)$$

Note that setting a lower corner frequency results in increased dependence on the actuator current being an accurate representation of true acceleration. Some frictional terms and bias forces (such as flex cable force), as well as variations in KF , will distort this relationship. The lower limit on this corner frequency will be determined by these non ideal effects.

Error Averaging

The velocity error output should ideally be within a few millivolts of V_{REF} (near zero error) through the deceleration region. However, due to manufacturing tolerances, this error will not be identical for each drive. The $EMAS_{OUT}$ and $ELATCH_{OUT}$ pins, allow this error to be nulled out for each individual system.

During deceleration (DECEL input high) a transconductance amplifier is switched on (see Block Diagram) and the velocity error output is integrated through an external capacitor. This average velocity error is then compared with V_{REF} and sent to $ELATCH_{OUT}$ (TTL output).

During acceleration (DECEL input low) the amplifier is switched off (high impedance output) and the external capacitor is discharged to V_{REF} through an external resistor. In this condition, the TTL output, ELATCH_{OUT}, is held in its previous state. Since the velocity error during acceleration is not of interest, the ELATCH_{OUT} during acceleration is the sign of the average velocity error output at the end of the previous deceleration cycle.

The processor can modify the velocity transducer gain based on the state of the ELATCH_{OUT} signal. During a power-up sequence, this transducer gain can be iteratively adjusted through several seek operations until the velocity error is minimized. As described below, one of the PWM to current translators on the ML4404 could be used to adjust this transducer gain.

PWM To Current Translators

The function of this block is to convert the duty cycle of a TTL input line to an analog output voltage. To optimize a complex servo control system, the manufacturing tolerances of some components must be accounted for. Traditionally, the manufacturing process included an expensive sequence of measurement and adjustment to bring each individual unit within specification. A more cost effective solution is to perform these tasks through software control.

The ML4404 implements this function with TTL to current translators. The external components R_{PWM} and C_{PWM} set the desired characteristics. The operation of these translators can be expressed by:

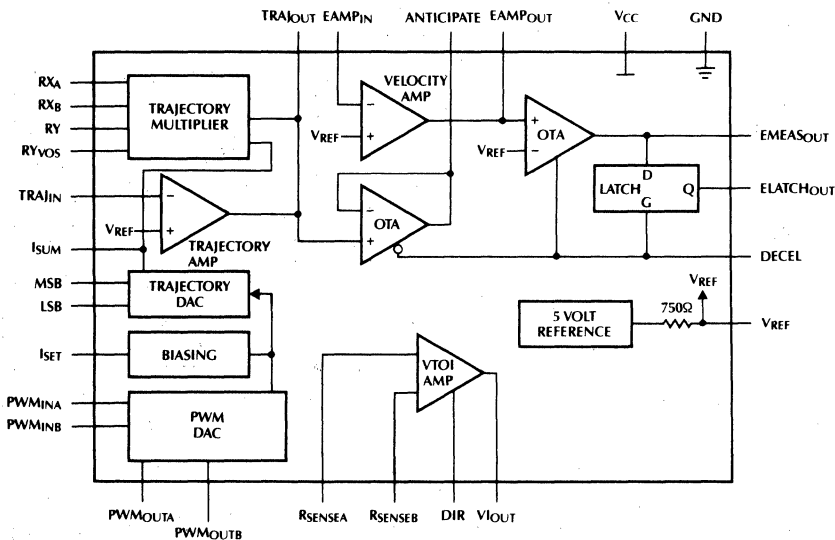
$$VO = V_{REF} + I_{BIAS} * R_{PWM} * (2 * DUTY\ CYCLE / 100 - 1)$$

Thus for a 50% duty cycle, the output voltage equals the reference voltage. The output voltage increases linearly with the input duty cycle.

The external capacitor (C_{PWM}) should be made sufficiently large to smooth out the PWM ripple.

The ML4404 has two translators. These stand-alone converters can be used for any purpose, but their intended functions are:

1. To set the AGC reference voltage on the ML4401, ML4431 servo demodulator.
2. To offset the position loop null location for data recovery (compensator inputs on the ML4403, ML4413 servo controller).
3. To offset the access arrival point for the trajectory (I_{SUM} node on the ML4404).



* Not available on the ML4404
 ** Not available on the ML4414

Figure 7. Block Diagram of Trajectory Generator.

ORDERING INFORMATION

PART NUMBER	TEMP. RANGE	PACKAGE
ML4404YCQ	0°C to +70°C	MOLDED PCC (Q28)

Disk Voice Coil Servo Driver

GENERAL DESCRIPTION

The ML4406 is a voice coil power driver intended for use in Hard Disk servo systems. The ML4406 contains all power and control circuitry necessary to drive the voice coils of most 3.5" drives. In addition, power fail detection and head retraction functions are provided for orderly shut-down of the drive.

The transconductance is programmed by a logic input at 1/4 A/V and 1/24 A/V respectively, using a 1Ω sense resistor. This allows for greater DAC resolution in digitally controlled servos during track follow without compromising dynamic range during seek.

The retraction circuit, main drive circuit, and control circuits are each powered from their own supplies. This allows maximum flexibility and provides for the lowest forward drop.

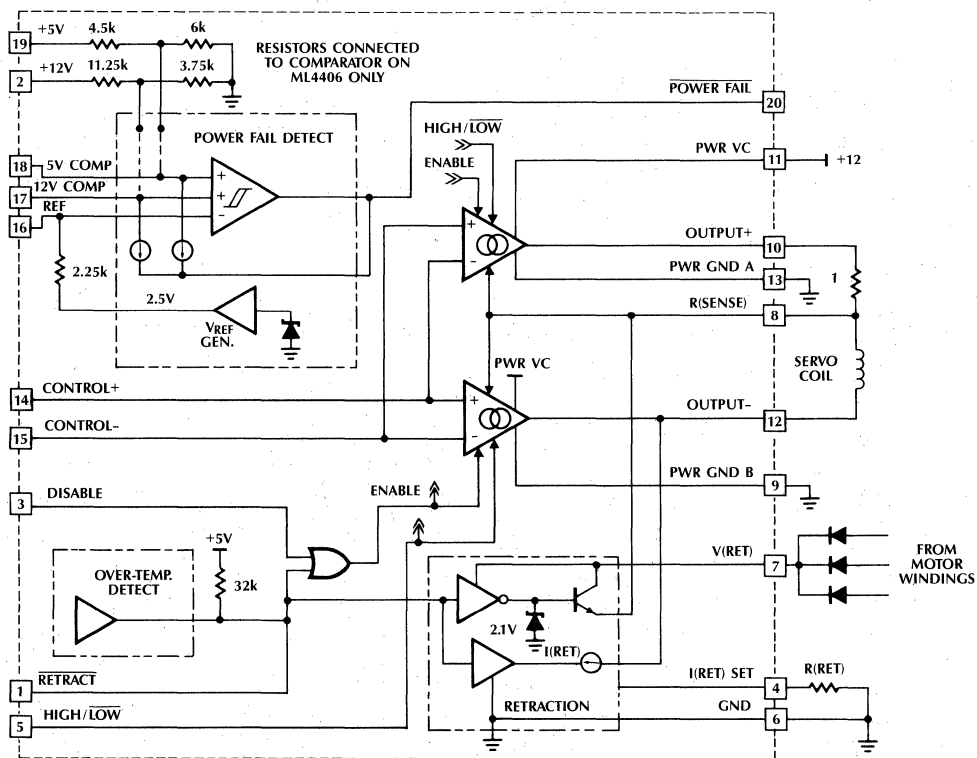
The power fail detection circuit includes a precision 2.5V bandgap reference with the option of either

internally generated power-fail thresholds (ML4406) or open comparator inputs for adjustable thresholds (ML4407).

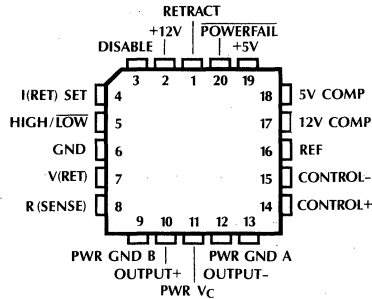
The ML4406 is implemented using Micro Linear's bipolar array technology. This allows for easy customization of the IC for a user's specific application.

FEATURES

- 500mA power output with 1.5V total forward drop
- Low offsets, cross-over distortion and quiescent current
- Pin-programmable transconductance settings
- Retraction circuitry with programmable retract current, voltage limiting, and separate supply pin
- On-chip precision power fail detect circuitry
- Over-temperature protection with flag output
- Logic input available for disabling outputs



PIN CONFIGURATION

ML4406/ML4407
20-Pin PCC

PIN DESCRIPTION

PIN NO.	NAME	FUNCTION	PIN NO.	NAME	FUNCTION
1	RETRACT	A logic "0" input causes the main outputs to tri-state and the retraction circuit to activate. This input also functions as a flag output and will go low in the event of an over-temperature condition.	11	PWR VC	Power supply for bridge amplifier.
2	+12V	12V power to the circuit and input to the power fail detection circuit.	12	OUTPUT-	Output terminal for bridge amplifier.
3	DISABLE	A logic "1" turns off the main outputs.	13	PWR GND A	Ground Terminal for power amplifier.
4	I(RET) SET	A resistor to ground sets the retract current.	14	CONTROL+	Positive input for current command.
5	HIGH/LOW	A logic "1" sets the trans-conductance gain to 1/4 while a logic "0" sets the gain to 1/24. Transconductance gain is voltage across $R_{SENSE} \div$ the input voltage.	15	CONTROL-	Negative input for current command.
6	GND	Analog Signal Ground.	16	REF	Reference input to the Power Fail comparator. Leave open to use internal 2.5V reference.
7	V(RET)	Power supply for the retract circuit.	17	12V COMP	Input to the Power Fail Comparator. Connect to an external resistor divider for the ML4407. Internally connected to a resistor divider from 12V in the ML4406.
8	R(SENSE)	Current sensing resistor terminal.	18	5V COMP	Input to the Power Fail Comparator. Connect to an external resistor divider for the ML4407. Internally connected to a resistor divider from 5V in the ML4406.
9	PWR GND B	Ground Terminal for power amplifier.	19	+5V	5V power supply terminal.
10	OUTPUT+	Output terminal for bridge amplifier.	20	POWER FAIL	Open collector output drives low if pin 17 or pin 18 are below pin 16. Normally tied to pin 1.

ML4406/ML4407

ABSOLUTE MAXIMUM RATINGS

Supply Voltage (Pins 11, 13, 7, 2)	14V
Voltage Pins 19, 18, 17, 16, 1, 3, 5	-3V to +7V
Pins 14, 15	-3 to +V _{CC}
Output Current	±750mA
Retraction Current	80mA
Retract Set Current (Pin 4)	3mA
Junction Temperature	150°C
Storage Temperature Range	-65°C to 150°C
Lead Temperature (soldering 10 sec.)	260°C
Thermal Resistance (θ_{JA})	60°C/W

OPERATING CONDITIONS

Temperature Range	0°C to 70°C
Supply Voltage (PWR VC, +12V)	12V ± 10%
+5V (Pin 19)	5V ± 10%
V(RET) (Pin 7)	2.5V to 16V
Control + Voltage Range (Pin 15 = 5V)	0V to V _{CC}
Control - Voltage Range	-1V to V _{CC} - 1V

Absolute maximum ratings are those values beyond which the device could be permanently damaged. Absolute maximum ratings are stress ratings only and functional device operation is not implied.

ELECTRICAL CHARACTERISTICS

Unless otherwise specified, T_A = Operating Temperature Range, V_{CC} = 12V, R_{SENSE} = 1Ω, R_{LOAD} = 15Ω, CONTROL- (Pin 15) = 5V, R_{SET} (Pin 4) = 1.2kΩ.

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
Amplifier					
Offset				±10	mA
Gain	Pin 5 = 2V	238	250	263	mA/V
	Pin 5 = 0.8V	39.6	41.7	43.8	mA/V
Bandwidth			100		KHz
Sinking Saturation	I _{OUT} = 100mA			.6	V
	I _{OUT} = 300mA			.8	
	I _{OUT} = 500mA			1.0	
Sourcing Saturation	I _{OUT} = 100mA			1.2	V
	I _{OUT} = 300mA			1.3	
	I _{OUT} = 500mA			1.5	
Retraction Circuit					
I(RET)SET			.75		V
Turn On Time			300		ns
Turn Off Time			2		ms
I(RET) Current	Pin 1 = 0.8V	34	50	65	mA
Power Fail Detection Circuit					
Reference Voltage		2.35	2.50	2.65	V
Reference Source Impedance			2.25		kΩ
Comparator Bias Current	ML4407 only, Pin 20 high		50	250	nA
Hysteresis Current	Pin 20 low, ML4407 only		10		μA
Offset Voltage	ML4407 only			10	mV
12V Threshold Hysteresis	ML4406 only	9.5	10	10.5	V
	ML4406 only		120		mV
5V Threshold Hysteresis	ML4406 only	4.40	4.575	4.75	V
	ML4406 only		30		mV
Logic Inputs					
Voltage High (V _{IH})		2	1.4		V
Voltage Low (V _{IL})			1.4	.8	V
Current High (I _{IH})	V _{IN} = 5V			±10	μA
Current Low (I _{IL})	V _{IN} = 0V	Except Pin 1	-40	-10	μA
		Pin 1 Only	-250	-160	μA

ELECTRICAL CHARACTERISTICS (Continued)

Unless otherwise specified, T_A = Operating Temperature Range, $V_{CC} = 12V$, $R_{SENSE} = 1\Omega$, $R_{LOAD} = 15\Omega$, CONTROL- (Pin 15) = 5V, R_{SET} (Pin 4) = 1.2k Ω .

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
Current Consumption					
Pin 19	$I_{LOAD} = 0$		1	2	mA
Pin 7	$I_{RET} = 0$		1	2	mA
Pin 2 + Pin 11	$I_{LOAD} = 0$		10	15	mA

FUNCTIONAL DESCRIPTION

POWER AMPLIFIER

The ML4406 power amplifier circuit is set up as a Howland current source with a fixed gain of 1/4 or 1/24 (set by driving pin 5 high or low respectively). This architecture yields minimal cross-over distortion while maintaining low output cross conduction currents. The gain figure refers to the ratio of input voltage to the output voltage seen across R_{SENSE} . For example, at a 1/4 gain setting with V- input at 2.5 and the V+ input at 4.5V, +500mA would flow through the coil using a 1 Ω sense resistor. Under the same conditions with pin 5 low, the current would be 83mA. The ability to change from low to high gain allows more complete utilization of DAC resolution when in the track follow mode.

The output stage (Figure 2) is designed to provide minimal saturation losses and employs a "composite PNP" for the sourcing drive and a saturable NPN to sink current. Sourcing saturation drop is typically .9V while sinking saturation drop is typically < 0.4V.

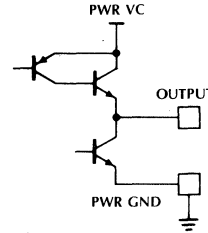


Figure 2. Main Power Output Stage

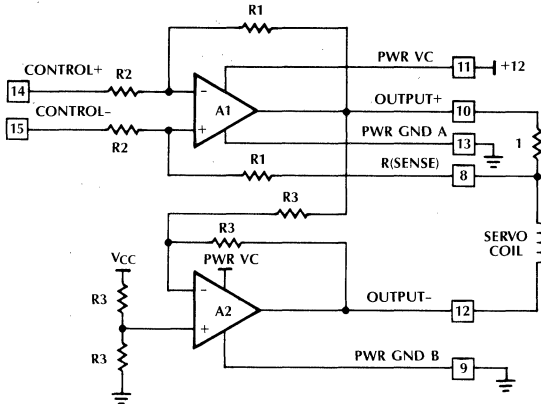


Figure 1. Power Amplifier Topology

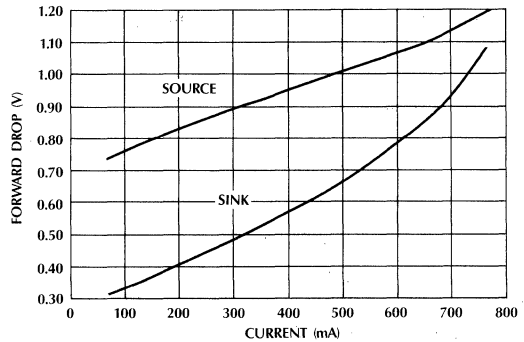


Figure 3. Output Saturation Voltage vs. Output Current (Power $V_C = 12V$)

ML4406, ML4407

Power Fail Detect

The ML4406 power fail detection circuit consists of a precision trimmed reference, resistor dividers, and an "or-function" comparator with hysteresis. The 10 μ A current sink on the comparator input lowers the comparator's positive input by 15mV when the output of the comparator is high. This creates an effective hysteresis of 30mV at the 5V input (on the ML4406). The amount of hysteresis and threshold levels can be programmed by external resistor dividers on the ML4407. The impedance of the external divider sets the amount of hysteresis while the division ratio sets the power fail threshold. The output at pin 20 is open-collector and is normally tied to pin 1 which is internally pulled-up to 5V.

Retract

The retract circuit features a current sink which is programmed via external resistor from pin 4 to ground (R_{RET}). The output of the retract circuit is voltage limited to 1.4V. The current sink provides an acceleration limit during retract while the voltage limited source provides a velocity limit. Pin 1 (Retract Input) also serves as a flag to indicate an over-temperature condition on the die. Pin 1 goes low in the event of over-temperature, which occurs when the die temperature exceeds a safe operating limit (about 160°C).

The retraction current is set by programming R_{RET} (figure 4). The retract circuit works down to 3V on $V_{RETRACT}$ (Pin 7).

Compensation

Figure 6 shows the equivalent AC circuit for the transconductance amplifier.

The amplifier's current bandwidth is limited by C_{OUT} which varies with the value chosen for R_{SENSE}

$$C_{OUT} = \frac{25nF}{R_{SENSE}}$$

With no snubber (R_S and C_S) the bandwidth is limited to

$$F_{-3dB} = \frac{1}{2\pi} \sqrt{\frac{2.414}{L(M) C(OUT)}}$$

Since this is a second order system with $L(M)$ and $C(OUT)$ forming a resonant circuit, some damping is desirable to reduce ringing in the step response. This is accomplished with a resistive snubber. The optimum value of $R(S)$ occurs when the following condition is met:

$$R(S) = \sqrt{\frac{L(VCM)}{C(OUT)}}$$

For a given $C(S)$, setting $R(S)$ to this value will minimize the ringing in the transient response. Larger values of $R(S)$ will result in more ringing and more bandwidth. Smaller values of $R(S)$ will result in more ringing and less bandwidth. $R(S)$ should not exceed 300 Ω .

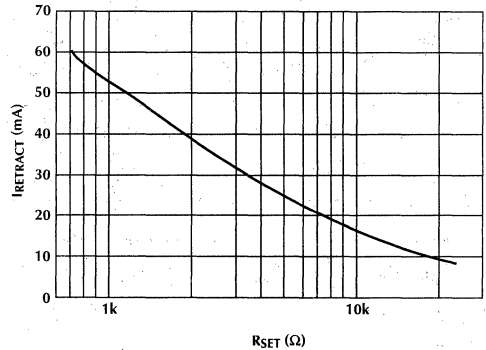


Figure 4. Retract Current vs. R_{SET}

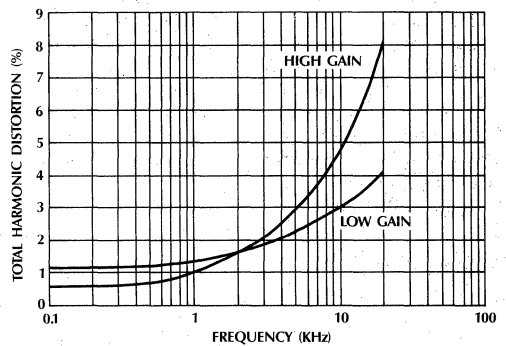


Figure 5. Total Harmonic distortion vs Frequency
Low Gain Setting ($V_{PIN 5} = 0$), $R_{SENSE} = 1\Omega$, $V_{IN} = 2.4V_{P-P}$
High Gain Setting ($V_{PIN 5} = 0$), $R_{SENSE} = 1\Omega$, $V_{IN} = 0.4V_{P-P}$

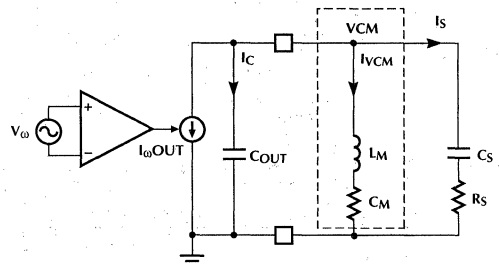


Figure 6. AC Equivalent Circuit for Current Amplifier, Voice Coil Motor (VCM) and Snubber.

$C(S)$ (snubber capacitor) values of between 200nF and 1 μ F are usually necessary to achieve the desired reduction of ringing in the step response. At the optimum value of $R(S)$ larger values of $C(S)$ further reduce the ringing but do not affect the bandwidth.

Tuning the current loop response can be easily done simulating the network in figure 6 with a computer simulator (such as SPICE).

APPLICATIONS

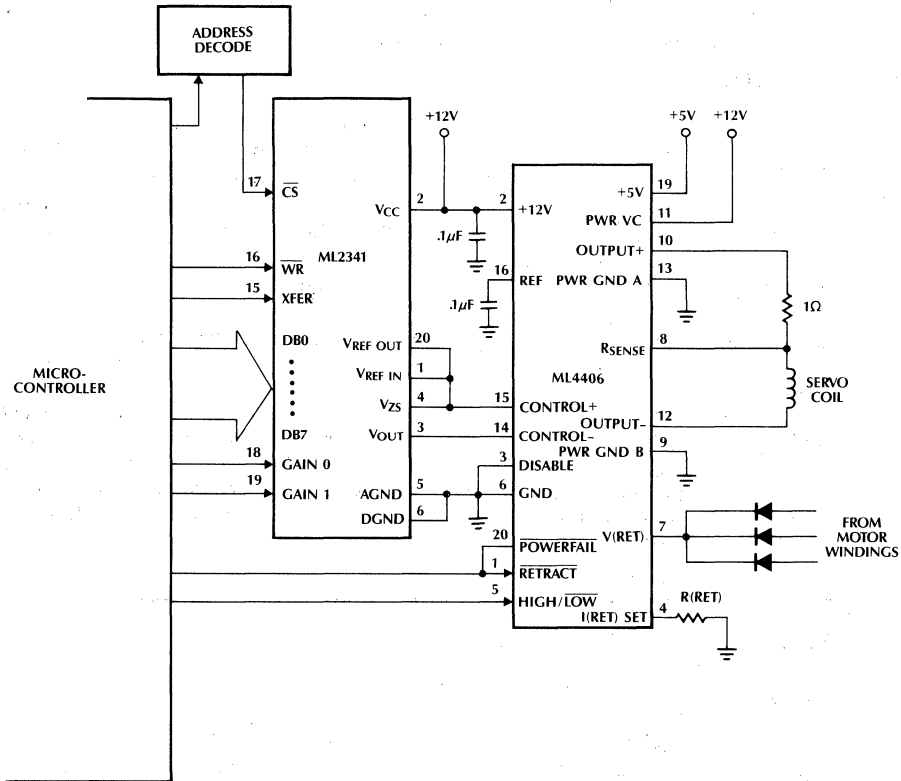


Figure 7. Typical Application: ML4406 used with ML2341 8-bit DAC provides up to 12-bit effective resolution

5

ORDERING INFORMATION

PART NUMBER	TEMP. RANGE	PACKAGE
ML4406CQ	0°C to +70°C	MOLDED PCC (Q20)
ML4407CQ	0°C to +70°C	MOLDED PCC (Q20)

Low Voltage Drop Voice Coil Servo Driver

GENERAL DESCRIPTION

The ML4408 is a voice coil power driver intended for use in High Performance 12V Hard Disk servo systems. The ML4408 contains all control circuitry necessary to drive the voice coils of most small drives. To maximize compliance voltage, the ML4408 includes two 1-Amp NPN drivers and provides drivers for external PNP transistors. In addition, power fail detection and a low voltage head retraction functions are provided for orderly shut-down of the drive.

The transconductance is programmed by a logic input at 1/4 A/V and 1/24 A/V respectively, when using a 1Ω sense resistor. This allows for greater DAC resolution in digitally controlled servos during track follow without compromising dynamic range during seek.

The retraction circuit, main drive circuit, and control circuits are each powered from their own supplies. Retract is self-contained for 12V systems but allows the use of an external PNP retraction with as little as 1V of back EMF from the spindle.

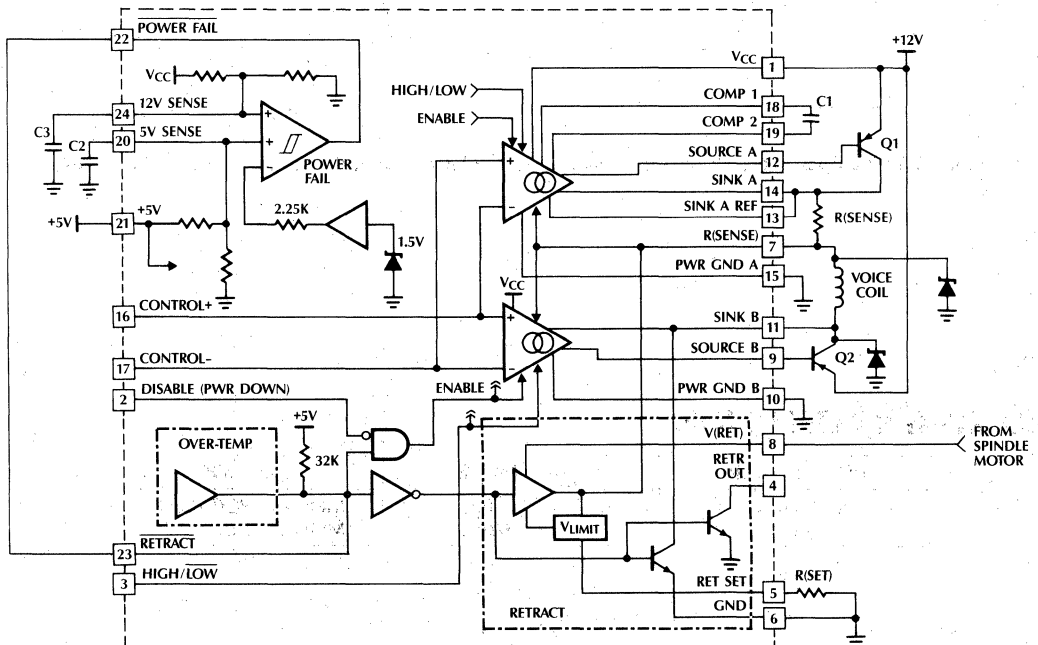
The power fail detection circuit includes a precision 1.5V bandgap reference and a power fail comparator.

The ML4408 is implemented using Micro Linear's bipolar array technology. This allows for customization of the IC for a user's specific application.

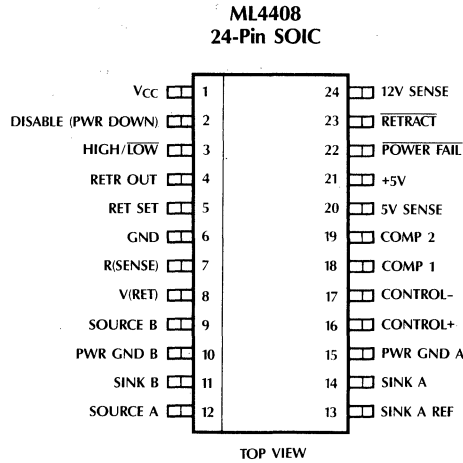
FEATURES

- Low saturation voltage (<1V at 1A)
- No cross-over distortion with low quiescent current
- Pin-programmable transconductance settings
- Retraction circuitry with programmable retract voltage and separate power pin
- On-chip precision power fail detect circuitry
- Over-temperature protection with flag output

BLOCK DIAGRAM



PIN CONFIGURATION



PIN DESCRIPTION

PIN #	NAME	FUNCTION	PIN #	NAME	FUNCTION
1	V _{CC}	Supply input to power amplifiers.	12	SOURCE A	PNP Base drive output for non-inverting power amplifier.
2	DISABLE (PWR DOWN)	A Logic "1" puts the IC into a low power state and disables the power amplifiers.	13	SINK A REF	Kelvin sensing point for power amplifier. Connect to SINK A.
3	HIGH/LOW	A logic "1" sets the transconductance gain to 1/4 while a logic "0" sets the gain to 1/24. Transconductance gain is the $V_{RSENSE} \div V_{CONTROL-}$.	14	SINK A	Current sinking output for non-inverting power amplifier. Connects to voice coil (+) terminal.
4	RETR OUT	Open collector output which pulls low during retract. Used to drive external power transistor to source retract current to the coil and can provide a braking signal to spindle.	15	PWR GND A	Power return pin for non-inverting power amplifier. Normally used for current sensing.
5	RET SET	External set resistor to establish a voltage limit for the internal retract driver.	16	CONTROL+	Positive input for current command.
6	GND	Analog signal ground.	17	CONTROL-	Negative input for current command.
7	R(SENSE)	Current sense resistor terminal.	18	COMP 1	Pin for external compensation capacitor.
8	V(RET)	Supply pin for retract circuits.	19	COMP 2	Pin for external compensation capacitor.
9	SOURCE B	PNP Base drive output for inverting power amplifier.	20	5V SENSE	Center node of a resistor divider from +5V.
10	PWR GND B	Power return pin for inverting power amplifier. Normally used for current sensing.	21	+5V	Input for +5V for power fail detection and logic power supply.
11	SINK B	Current sinking output for inverting power amplifier. Connects to voice coil (-) terminal.	22	POWER FAIL	Open Collector output drives low for low voltage conditions.
			23	RETRACT	A logic "0" initiates retract. Also used as an open-collector over-temperature output flag.
			24	12V SENSE	Input to the power fail comparator from a resistor divider from V _{CC} .

5

ML4408

ABSOLUTE MAXIMUM RATINGS

Absolute maximum ratings are those values beyond which the device could be permanently damaged. Absolute maximum ratings are stress ratings only and functional device operation is not implied.

Supply Voltage (pins 1, 8)	14V
Voltage pins 2, 3, 23	-3V to +7V
pins 4, 7, 9, 11, 12, 13, 14, 16, 17, 22	-3V to V_{CC}
Output Sink Current	$\pm 1A$
Retraction Current	80mA
Retract Set Current (pin 5)	3mA
Junction Temperature	150°C
Storage Temperature Range	-65°C to +150°C
Lead Temperature (Soldering 10 sec)	150°C
Thermal Resistance (θ_{JA})	60°C/W

OPERATING CONDITIONS

Temperature Range	0°C to +70°C
V_{CC} Supply Voltage	
12V operation	10.8V to 13.2V
+5V (pin 21) Supply Voltage	4.5V to 5.5V
V(RET) (pin 8) Supply Voltage	
12V operation	2.5V to 13.2V
Control + Voltage Range (pin 15 = 5V)	0V to V_{CC}
Control - Voltage Range	2V to $V_{CC} - 1.5V$

ELECTRICAL CHARACTERISTICS

Unless otherwise specified, T_A = Operating Temperature Range, V_{CC} = Operating Range, $R_{SENSE} = 1\Omega$, $R_{COIL} = 15\Omega$, CONTROL- (pin 17) = $V_{CC}/2$, $C1 = 30pF$, $Q1, Q2 = MJE210$, $R_{SET} = 3.7K\Omega$.

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
Amplifier					
Offset				± 10	mA
Gain	Pin 5 = 2V	238	250	263	mA/V
	Pin 5 = 0.8V	39.6	41.7	43.8	
Bandwidth			100		KHz
Sinking Saturation	$I_{OUT} = 100mA$.3		V
	$I_{OUT} = 300mA$.4		V
	$I_{OUT} = 500mA$.5		V
Sourcing Saturation	$I_{OUT} = 100mA$.1		V
	$I_{OUT} = 300mA$.2		V
	$I_{OUT} = 500mA$.3		V
Source A/B Base Drive		20			mA
Q1/Q2 Standby Current	$V_{PIN 16} = 5V$		4		mA
Retraction Circuit					
I(RET) SET			.75		V
Turn On Time			800		ns
Turn Off Time			8		μs
Source Voltage	$V_{PIN 23} = 0.8V$, $V_{PIN 8} = 3V$, $I_{PIN 7} = 50mA$	0.95	1.2	1.5	V
Sink Current	$V_{PIN 23} = 0.8V$, $V_{PIN 8} = 1.2V$, $V_{PIN 11} = 0.5V$	36	48	60	mA
RETR OUT V_{OL}	$V_{PIN 23} = 0.8V$, $I_{PIN 4} = 1mA$		0.1	0.4	V
Power Fail Detection Circuit					
12V Threshold		9.5	10	10.5	V
Hysteresis — 12V Sense			120		mV
5V Threshold		4.40	4.575	4.75	V
Hysteresis — 5V Sense			30		mV

ELECTRICAL CHARACTERISTICS (Continued)

Unless otherwise specified, T_A = Operating Temperature Range, V_{CC} = Operating Range, $R_{SENSE} = 1\Omega$, $R_{COIL} = 15\Omega$, $CONTROL-$ (pin 17) = $V_{CC}/2$, $C1 = 30pF$, $Q1, Q2 = MJE210$, $R_{SET} = 3.7K\Omega$.

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
Logic Inputs and Outputs					
Voltage High (V_{IH})		2	1.4		V
Voltage Low (V_{IL})			1.4	.8	V
Current High (I_{IH})	$V_{IN} = 5V$			± 10	μA
Current Low (I_{IL})	$V_{IN} = 0V$, except pin 23	-40	-10		μA
	$V_{IN} = 0V$, pin 23 only	-250	-160		μA
Voltage Low (pins 22, 23)	$I_{OL} = 1mA$.4	V
Over-Temperature Detection					
T_j Threshold			160		$^{\circ}C$
Hysteresis			30		$^{\circ}C$
Current Consumption					
Pin 21	Pin 21 = 5.5V		5	7	mA
Pin 1	$V_{CC} = 13.2V$, $V_{PIN 16} = V_{CC}/2$		5	10	mA
Pin 8	$V_{PIN 8} = 13.2V$, $V_{PIN 23} = 5V$		3.5	5	mA

FUNCTIONAL DESCRIPTION**POWER AMPLIFIER**

The ML4408 power amplifier circuit (figure 1) is set up as a Howland Current source with a fixed gain of 1/4 or 1/24 (set by driving pin 3 high or low respectively). This architecture yields minimal cross-over distortion while maintaining low output cross conduction currents. The gain figure refers to the ratio of input voltage to the output voltage seen across R_{SENSE} . For example, at a 1/4 gain setting, with $V(-)$ input at 2.5V and the $V(+)$ input at 4.5V, +500mA would flow through the coil using a 1Ω sense resistor. Under the same conditions with pin 3 low, the current would be 83mA. If lower input voltage swings and higher currents are desired, the overall transconductance gain may be increased by using a lower value of sense resistor, however offset current will increase proportionally. The ability to change from low to high gain allows more complete utilization of DAC resolution when in the track follow mode.

The output stage is designed to provide minimal saturation losses and employs an external PNP transistor for the sourcing drive and an internal saturable NPN to sink current. Sinking saturation drop is typically under 0.4V. Sourcing saturation drop depends on the external transistors used.

Care should be taken to avoid drawing substrate currents due to negative excursions on any pin of the ML4408. Schottky diodes should be included on both sides of the VCM to prevent negative excursions from forward biasing the substrate diodes on the IC.

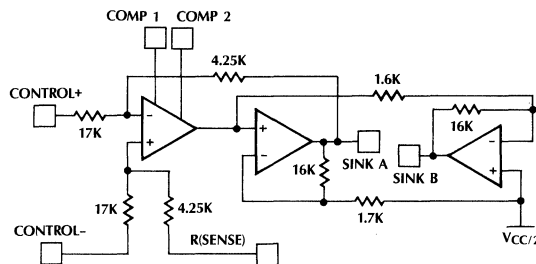


Figure 1. Simplified Power Amplifier Schematic (High Gain Mode)

Two areas should be considered to avoid high frequency oscillation in the output stage:

1. Choose external PNP transistors with a F_T of at least 50MHz.
2. An RC compensation network should be used to cancel the zero presented to the output by the L/R of the voice coil motor as shown in figure 2.

COMPENSATION

Figure 2 shows the equivalent AC circuit for the current amplifier.

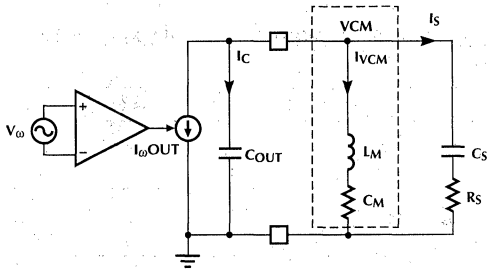


Figure 2. AC Equivalent Circuit for Current Amplifier, Voice Coil Motor (VCM) and Snubber.

The amplifier's current bandwidth is limited by C_{OUT} which varies with the value chosen for R_{SENSE}

$$C_{OUT} \approx \frac{1200 \times (C_{COMP} + 12.8pF)}{R_{SENSE}}$$

Where C_{COMP} is C1 between pins 18 and 19. With no snubber (R_S and C_S) the bandwidth is limited to

$$F_{-3dB} = \frac{1}{2\pi} \sqrt{\frac{2.414}{L(M) C(OUT)}}$$

Since this is a second order system with L(M) and C(OUT) forming a resonant circuit, some damping is desirable to reduce ringing in the step response. This is accomplished with a resistive snubber. The optimum value of R(S) occurs when the following condition is met:

$$R(S) = \sqrt{\frac{L(VCM)}{C(OUT)}}$$

For a given C(S), setting R(S) to this value will minimize the ringing in the transient response. Larger values of R(S) will result in more ringing and more bandwidth. Smaller values of R(S) will result in more ringing and less bandwidth. C(S) (snubber capacitor) values of between 200nF and 1μF are usually necessary to achieve the desired reduction of ringing in the step response. At the optimum value of R(S) larger values of C(S) further reduce the ringing but do not affect the bandwidth.

Tuning the current loop response can be easily done simulating the network in figure 2 with a computer simulator (such as SPICE).

POWER FAIL DETECT CIRCUIT

The ML4408 circuit consists of a precision trimmed reference, resistor dividers and an "or function" comparator with hysteresis. The output (open collector) of this circuit appears on pin 22. When either comparator input (pins 20 and 24) falls below the 1.5V reference, pin 22 pulls low.

RETRACT CIRCUITS

When pin 23 goes low, pin 4 will pull low. The internal NPN transistor will saturate, pulling SINK B (pin 11) low. This portion of the circuit will function with less than 1V on V(RET). An internal voltage limited pull-up transistor is provided which sources current on pin 7 to the VCM. This circuit will operated reliably down to a V(RET) voltage of around 2.5V, making the ML4408 retract circuit adequate for 12V systems where the spindle motor EMF provided is adequate.

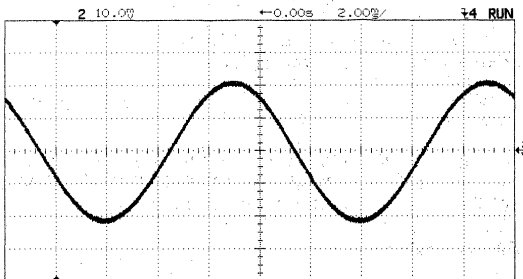


Figure 3. Output Current: V_{IN} = 100 Hz Sine Wave, 1V_{P-P} Low Gain Mode (V_{PIN 3} = 0), R_{SENSE} = 1Ω

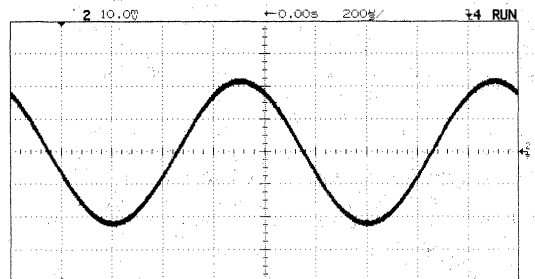


Figure 4. Output Current: V_{IN} = 1 KHz Sine Wave, 1V_{P-P} Low Gain Mode (V_{PIN 3} = 0), R_{SENSE} = 1Ω

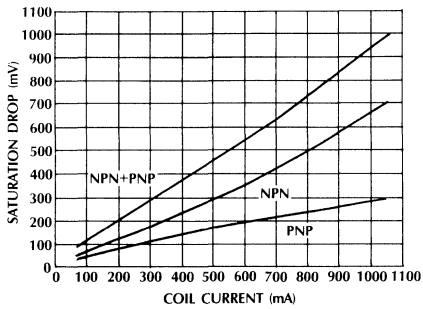


Figure 5. Output Saturation Voltage vs Output Current ($Q_1 = Q_2 = \text{MJE210}$)

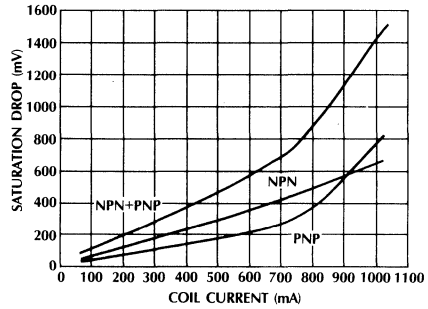


Figure 6. Output Saturation Voltage vs Output Current with BSR31 ($Q_1 = Q_2 = \text{BSR31}$)

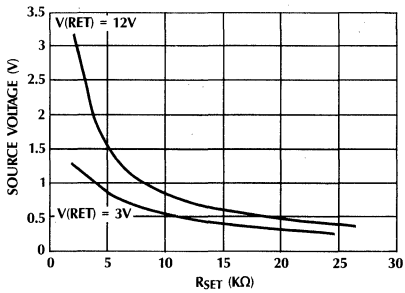


Figure 7. Retract Source Voltage Limit

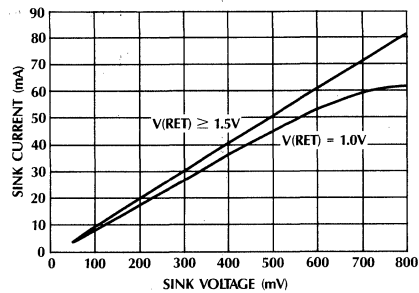


Figure 8. Retract Sink Voltage vs Current

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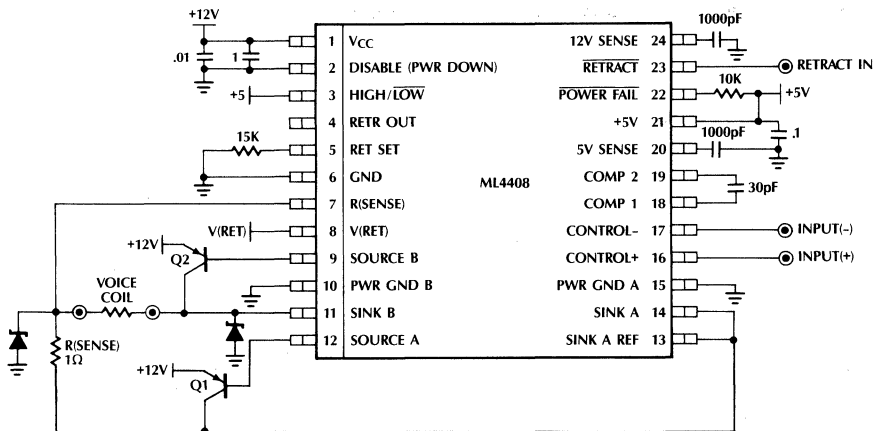


Figure 9. Typical 12V Application

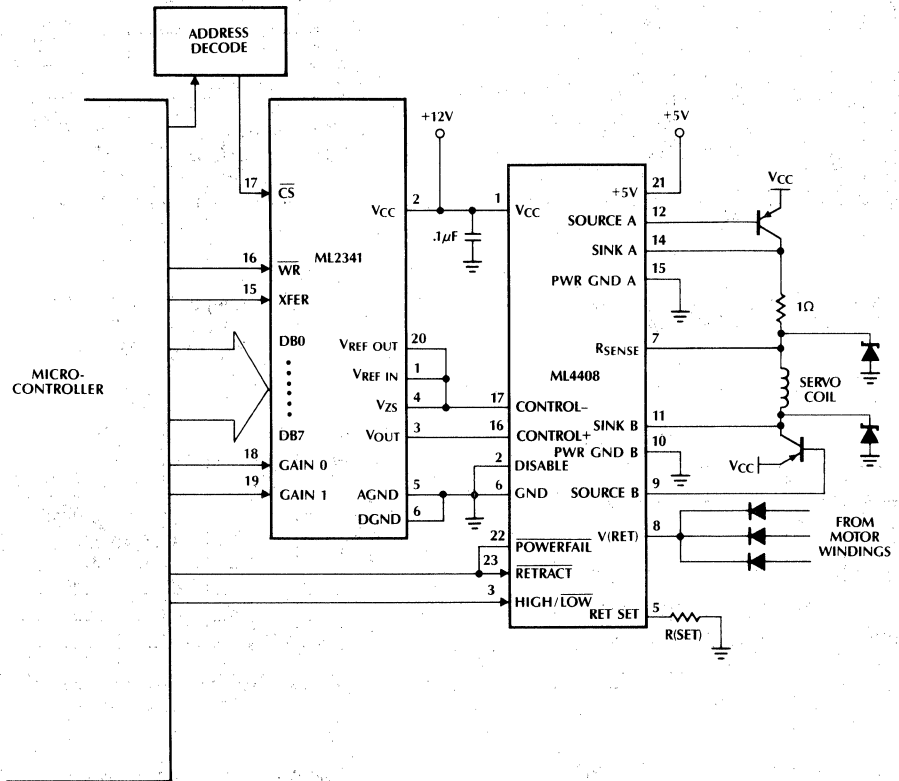


Figure 10. ML4408 Used with ML2341 8-Bit DAC Provides up to 13-Bit Effective Resolution

Sensorless Spindle Motor Controller

GENERAL DESCRIPTION

The ML4410 provides complete commutation for delta or wye wound Brushless DC (BLDC) motors without the need for signals from Hall Effect sensors. This IC senses the back EMF of the 3 motor windings (no neutral required) to determine the proper commutation phase angle using phase lock loop techniques. This technique will commutate virtually any 3-phase BLDC motor and is insensitive to PWM noise and motor snubbing.

Included in the ML4410 is the circuitry necessary for a Hard Disk Drive microcontroller driven control loop. The ML4410 controls motor current with either a constant off-time PWM or linear current control driven by the microcontroller. Speed feedback for the micro is a stable digital frequency equal to the commutation frequency of the motor. All commutation is performed by the ML4410. Braking and Power Fail are also included in the ML4410.

Two different start-up sequencing (minimum start-up time or minimum reverse rotation at start up) algorithms are supported by the ML4410. Since the timing of the start-up sequencing is determined by the

micro, the system can be optimized for a wide range of motors and inertial loads.

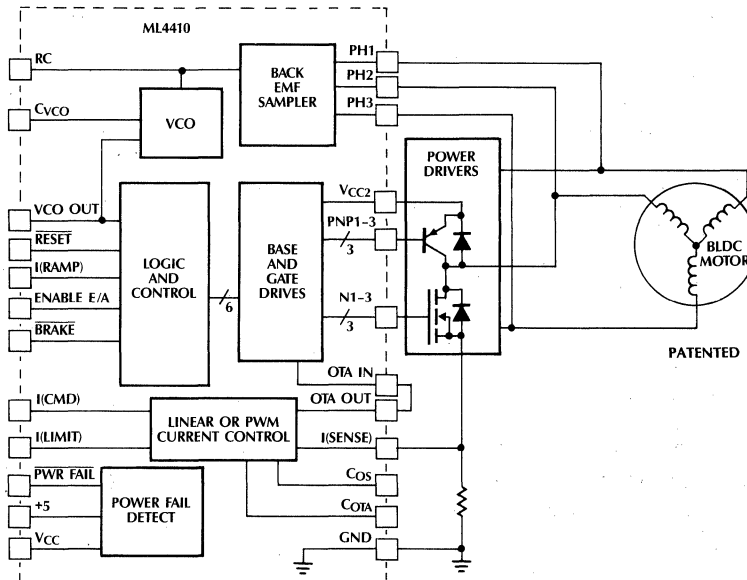
The ML4410 modulates the gates of external N-channel power MOSFETs to regulate the motor current. The IC drives external PNP transistors or P-channel MOSFETs directly. Special circuits are used to save base drive power at low load currents.

FEATURES

- Back-EMF Commutation Provides Maximum Torque for Minimum "Spin-Up" Time for Spindle Motors
- Accurate, Jitter-Free Phase Locked Motor Speed Feedback Output
- Linear or PWM Motor Current Control
- Easy Microcontroller Interface for Optimized Start-Up Sequencing and Speed Control
- Power Fail Detect Circuit with Delayed Braking
- Drives External N-Channel FETs and PNP's or P-Channel FETs

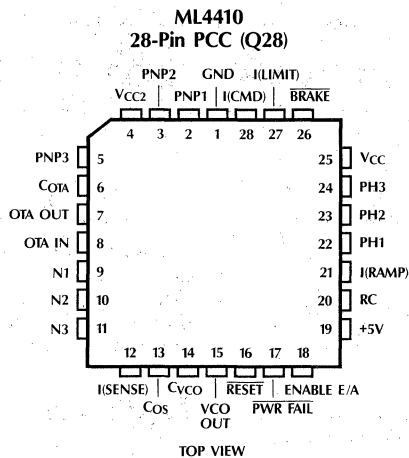
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BLOCK DIAGRAM



ML4410

PIN CONFIGURATION



PIN DESCRIPTION

PIN #	NAME	FUNCTION	PIN #	NAME	FUNCTION
1	GND	Signal and Power Ground.	16	RESET	Input which holds the VCO off and sets the ML4410 to the RESET condition.
2	PNP1	Drives the external PNP power transistor driving motor PH1.	17	PWR FAIL	A "0" output indicates 5V or 12V is under-voltage.
3	PNP2	Drives the external PNP power transistor driving motor PH2.	18	ENABLE E/A	A "1" logic input enables the error amplifier and closes the back-EMF feedback loop.
4	VCC2	12V power and power for the braking function.	19	+5V	5V power supply input.
5	PNP3	Drives the external PNP power transistor driving motor PH3.	20	RC	VCO loop filter components.
6	COTA	Compensation capacitor for linear motor current amplifier loop.	21	I(RAMP)	Current into this pin sets the initial acceleration rate of the VCO during start-up.
7	OTA OUT	Output of motor current error amplifier, normally connected to OTA IN or to external MOSFET gate.	22	PH1	Motor Terminal 1.
8	OTA IN	Driving voltage for N1-N3. Normally tied to OTA OUT.	23	PH2	Motor Terminal 2.
9-11	N1, N2, N3	Drives the external N-channel MOSFETs for PH1, PH2, PH3.	24	PH3	Motor Terminal 3.
12	I(SENSE)	Motor current sense input.	25	VCC	12V power supply. Terminal which is sensed for power fail.
13	Cos	Timing capacitor for fixed off-time PWM current control.	26	BRAKE	A "0" activates the braking circuit.
14	Cvco	Timing capacitor for VCO.	27	I(LIMIT)	Sets the threshold for the PWM comparator.
15	VCO OUT	Open Collector Logic Output from VCO.	28	I(CMD)	Current Command for Linear Current amplifier.

ABSOLUTE MAXIMUM RATINGS

Absolute maximum ratings are those values beyond which the device could be permanently damaged. Absolute maximum ratings are stress ratings only and functional device operation is not implied.

Supply Voltage (pins 4, 25)	14V
Output Current (pins 2, 3, 5, 9, 10, 11)	±150mA
Logic Inputs (pins 16, 17, 18, 26)	-0.3 to 7V
Junction Temperature	150°C
Storage Temperature Range	-65°C to +150°C

Lead Temperature (Soldering 10 sec)	150°C
Thermal Resistance (θ_{JA})	60°C/W

OPERATING CONDITIONS

Temperature Range	0°C to +70°C
V _{CC} Voltage +12V (pin 25)	12V ± 10%
+5V (pin 19)	5V ± 10%
I(RAMP) Current (pin 21)	0 to 100μA
I Control Voltage Range (pins 27, 28)	0V to 7V

ELECTRICAL CHARACTERISTICS

Unless otherwise specified, T_A = Operating Temperature Range, V_{CC} = V_{CC2} = 12V, R_{SENSE} = 1Ω, C_{OTA} = C_{VCO} = .01μF, C_{OS} = .02μF.

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
Oscillator (VCO) Section (Mode 1 or 2 unless otherwise specified)					
Frequency vs. V _{PIN 20}	1V ≤ V _{PIN 20} ≤ 10V		300		Hz/V
Frequency	V _{VCO} = 6V	1450	1800	2150	Hz
	V _{VCO} = .5V	70	140	210	Hz
Reset Voltage at C _{VCO}	Mode = 0		125	250	mV
Sampling Amplifier					
V _{RC}	Mode 0		125	250	mV
I _{RC}	Mode 1, R _{RAMP} = 39KΩ	70	100	130	μA
	Mode 2A, V _{PH2} = 4V	30	50	70	μA
	Mode 2A, V _{PH2} = 6V	-15	2	+15	μA
	Mode 2A, V _{PH2} = 8V	-30	-50	-70	μA
Motor Current Control Section					
I(SENSE) Gain	V _{PIN 27} = 5V, 0V ≤ V _{PIN 28} ≤ 2.5V	4.5	5	5.5	V/V
One Shot Off Time		12	25	33	μs
I(CMD) Transconductance Gain			.19		mmho
Power Fail Detection Circuit					
12V Threshold		9.1	9.8	10.5	V
Hysteresis			150		mV
5V Threshold		3.8	4.25	4.5	V
Hysteresis			70		mV
Logic Inputs					
Voltage High (V _{IH})		2			V
Voltage Low (V _{IL})				.8	V
Current High (I _{IH})	V _{IN} = 2.7V	-10	1	10	μA
Current Low (I _{IL})	V _{IN} = 0.4V	-500	-350	-200	μA

ML4410

ELECTRICAL CHARACTERISTICS (Continued)

Unless otherwise specified, T_A = Operating Temperature Range, $V_{CC} = V_{CC2} = 12V$, $R_{SENSE} = 1\Omega$, $C_{OTA} = C_{VCO} = .01\mu F$, $C_{OS} = .02\mu F$

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
Outputs					
	$I(CMD) = I(LIMIT) = 2.5V$				
I_{PNP} Low		50	75	100	mA
I_{PNP} High	Off State	-100		100	μA
V_N High	$V_{PIN 8} = 10V$	9.7	10	10.3	V
V_N Low			.2	.7	V
A_V Pin 8 to V_N	$V_{PIN 8} = 6V$.95	1	1.05	V/V
LOGIC Low	$I_{OUT} = 0.5mA$.4	V
LOGIC I_{OUT} High			5		μA
Supply Currents					
	(N and PNP Outputs Open)				
5V Current			2	4	mA
V_{CC} Current			38	50	mA
V_{CC2} Current			4	10	mA

Note 1: For explanation of states, see figure 5 and table 1.

FUNCTIONAL DESCRIPTION

The ML4410 provides closed-loop commutation for 3-phase brushless motors. To accomplish this task, a VCO, integrating Back-EMF Sampling error amplifier and sequencer form a phase-locked loop, locking the VCO to the back-EMF of the motor. The IC also contains circuitry to control motor current with either linear or constant off-time PWM modes. Braking and power fail detection functions are also provided on chip. The ML4410 is designed to drive external power transistors (N-channel MOSFET sinking transistors and PNP sourcing transistors) directly, and contains a special circuit to reduce PNP base currents when output current demand is reduced.

Start-up sequencing and motor speed control are accomplished by a microcontroller. Speed sensing is accomplished by monitoring the output of the VCO, which will be a signal which is phased-locked to the commutation frequency of the motor.

BACK-EMF SENSING AND COMMUTATOR

The ML4410 contains a patented back-EMF sensing circuit which samples the phase which is not energized (Shaded area in figure 2) to determine whether to increase or decrease the commutator (VCO) frequency. A late commutation causes the error amplifier to charge the filter (RC) on pin 20, increasing the VCO input while early commutation causes pin 20 discharge. Analog speed control loops can use pin 20 as a speed feedback voltage.

The input impedance of the three PH inputs is about $8K\Omega$ to GND. When operating with a higher voltage motor, the PH inputs should be divided down in voltage so that the maximum voltage at any PH input does not exceed V_{CC} . See ML4411 data sheet for applications.

VCO AND PHASE DETECTOR CALCULATIONS

The VCO should be set so that at the maximum frequency of operation (the running speed of the motor) the VCO

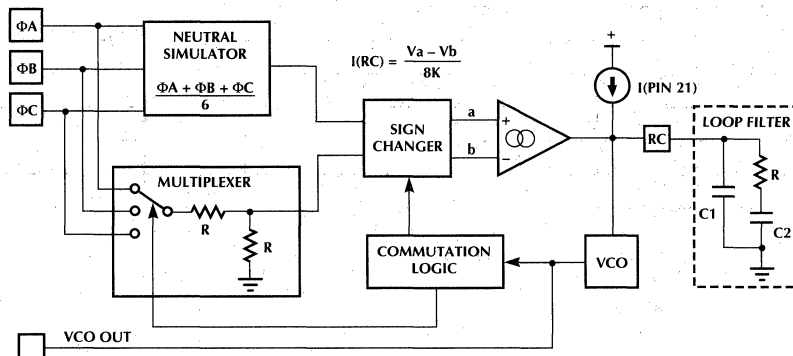


Figure 1. Back EMF Sensing Block Diagram.

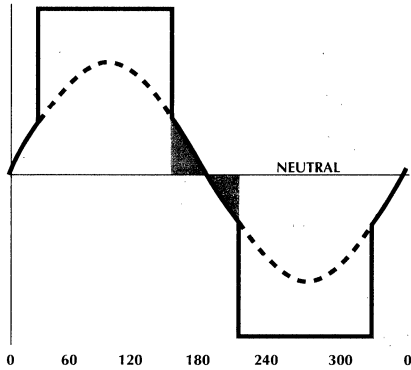


Figure 2. Typical Motor Phase Waveform with Back-EMF Superimposed (Ideal Commutation).

control voltage will be no higher than $V_{CC_{MIN}} - 1V$. The VCO maximum frequency will be:

$$F_{MAX} = 0.05 \times POLES \times RPM$$

where POLES is the number of poles on the motor and RPM is the maximum motor speed in Revolutions Per Minute.

The minimum VCO gain derived from the specification table (using the minimum F_{VCO} at $V_{VCO} = 6V$) is:

$$K_{VCO(MIN)} = \frac{2.42 \times 10^{-6}}{C_{VCO}}$$

Assuming that the $V_{VCO(MAX)} = 9.5V$, then

$$C_{VCO} = \frac{9.5 \times 2.42 \times 10^{-6}}{F_{MAX}}$$

or

$$C_{VCO} = \frac{460}{POLES \times RPM} \mu F$$

Figure 4 shows the transfer function of the Phase Lock Loop with the phase detector formed from the sampled phase through the G_m amplifier with the loop filtered formed by R , C_1 , and C_2 .

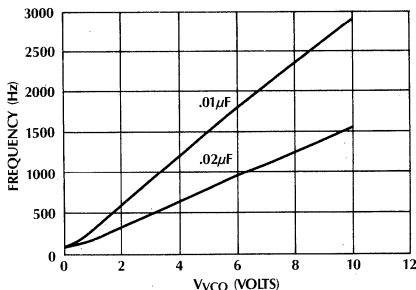


Figure 3. VCO Output Frequency vs. V_{VCO} (Pin 20)

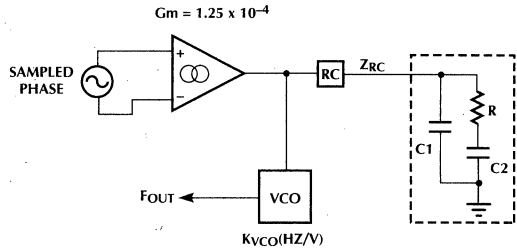


Figure 4. Back EMF Phase Lock Loop Components.

The impedance of the loop filter is

$$Z_{RC}(s) = \frac{1}{C_1 s} \frac{(s + \omega_{LEAD})}{(s + \omega_{LAG})}$$

Where the lead and lag frequencies are set by:

$$\omega_{LEAD} = \frac{1}{RC_2}$$

$$\omega_{LAG} = \frac{C_1 + C_2}{RC_1 C_2}$$

Requiring the loop to settle in 20 PLL cycles with a spread of 10 between $\omega_{LEAD} = 10 \times \omega_{LAG}$ produces the following calculations for R , C_1 and C_2 :

$$C_1 \approx \frac{8.06 \times 10^{-9}}{C_{VCO} \times F_{VCO}^2}$$

$$C_2 = 9 \times C_1$$

$$R = \frac{12.54}{C_2 \times F_{VCO}}$$

START-UP SEQUENCING

When the motor is initially at rest, it is generating no back-EMF. Because a back-EMF signal is required for closed loop commutation, the motor must be started “open-loop” until a velocity sufficient to generate some back-EMF is attained (around 100 RPM). The following steps are a typical procedure for starting a motor which is at rest.

Step 1: The IC is held in reset (state R) with full power applied to the windings (see figure 6). This aligns the rotor to a position which is 30° (electrical) before the center of the first commutation state.

Step 2: Reset is released, and a fixed current is input to pin 21 and appears as a current on pin 20, and will ramp the VCO input voltage, accelerating the motor at a fixed rate.

Step 3: When the motor speed reaches about 100 RPM, the back EMF loop can be closed by pulling pin 18 high.

STATE	OUTPUTS						INPUT SAMPLING
	N1	N2	N3	PNP1	PNP2	PNP3	
R or 0	OFF	ON	OFF	ON	OFF	ON	N/A
A	OFF	OFF	ON	ON	OFF	OFF	PH2
B	OFF	OFF	ON	OFF	ON	OFF	PH1
C	ON	OFF	OFF	OFF	ON	OFF	PH3
D	ON	OFF	OFF	OFF	OFF	ON	PH2
E	OFF	ON	OFF	OFF	OFF	ON	PH1
F	OFF	ON	OFF	ON	OFF	OFF	PH3

Table 1. Commutation States.

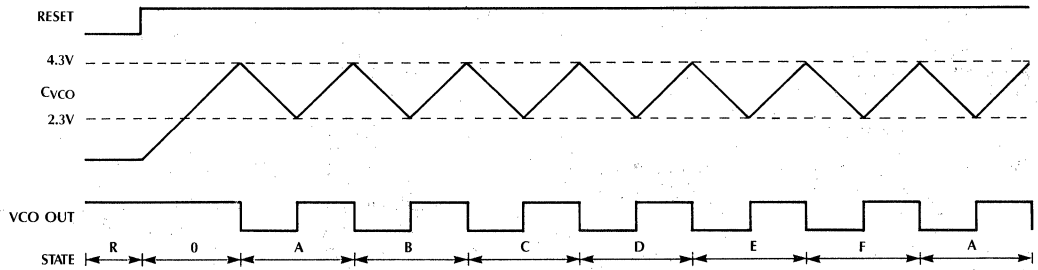


Figure 5. Commutation Timing and Sequencing.

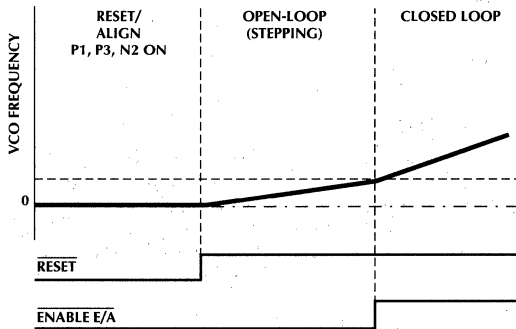


Figure 6. Typical Start-up Sequence.

Step	Pin 16	Pin 18	Pin 21	I(LIMIT) I(CMD)
1	0	0	Fixed	I_{MAX}
2	1	0	Fixed	I_{MAX}
3	1	1	0	I_{MAX}

Table 2. Start-up Sequence.

Using this technique, some reverse rotation is possible. The maximum amount of reverse rotation is $360/N$, where N is the number of poles. For an 8 pole motor, 45° reverse rotation is possible.

ADJUSTING OPEN LOOP STEP RATE

I_{RAMP} should be set so that the VCO's frequency ramp during "open loop stepping" phase of motor starting is less than the motor's acceleration rate. In other words, the motor must be able to keep up with the VCO's ramp rate in open loop stepping mode. The VCO's input voltage ($V_{PIN 20}$) ramp rate is given by:

$$\frac{dV_{VCO}}{dt} \approx \frac{I_{RAMP}}{C_1 + C_2}$$

since

$$F_{VCO} = K_{VCO} \times V_{VCO}$$

$$K_{VCO(MAX)} = \frac{4 \times 10^{-6}}{C_{VCO}}$$

then combining the 3 equations I_{RAMP} can be calculated from the desired maximum open loop stepping rate the motor can follow.

$$I_{RAMP} < \frac{dF_{VCO}}{dt} \frac{C_{VCO} \times (C_1 + C_2)}{4 \times 10^{-6}}$$

The motor will start more consistently and tolerate a wider variation in open loop step rate if there is some damping on the motor (such as head drag) during the open loop modes.

The tolerance of the open loop step VCO acceleration

$\left(\frac{dF_{VCO}}{dt}\right)$ depends on the tolerances of K_{VCO} , I_{RAMP} , C_1 ,

C_2 , and C_{VCO} . For more optimum spin up times, these variables can be digitally "calibrated" out by the microprocessor using the following procedure:

1. Reset the IC by holding pin 16 low for at least 5 μ s.
2. Go into open loop step mode with no current on the motor and measure the difference between the first two complete VCO periods with the PWM signal at 50% duty cycle:

ENABLE E/A = (see below)
 I(CMD) = 0V
 PWM OUT = 50%

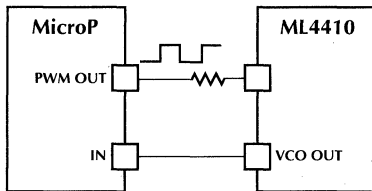


Figure 7. Auto-Calibration of Open-Loop Step Rate.

3. Compute a correction factor to adjust I_{RAMP} current by changing the PWM duty cycle from the Micro (D.C.)

$$D.C.(NEW) = 50\% \times \frac{\Delta F_{VCO} (DESIRED)}{\Delta F_{VCO} (MEASURED)}$$

4. Use new computed duty cycle for open loop stepping mode and proceed with a normal start-up sequence.

If this auto calibration is used ENABLE E/A can be tied permanently high, eliminating a line from the Micro. Since there is offset associated with the Phase Detector Error Amp (E/A), more current than is being injected by I_{RAMP} may be taken out of pin 20 if the offset is positive (into pin 20) if the error amp were enabled during the open loop stepping mode. In that case, V_{VCO} would not rise and the motor would not step properly. The effect of E/A offset can also be canceled out by the auto calibration algorithm described above allowing the E/A to be permanently enabled.

PWM AND LINEAR CURRENT CONTROL

To facilitate speed control, the ML4410 includes two current control loops — linear and PWM (figure 9). The linear control loop senses the motor current on the I(SENSE) terminal through R_{SENSE} . An internal current sense amplifier's (A2) output modulates the gates of the 3 N-channel MOSFET's when OTA OUT is tied to OTA IN, or can modulate a single MOSFET gate tied to OTA OUT. When operated in this mode, OTA IN is tied to 12V, and N1-N3 are saturated switches. This method produces the lowest current ripple at the expense of an extra MOSFET.

The linear current control modulates the gates of the external MOSFET drivers. Amplifier A2 is a transconductance amplifier which amplifies the difference between I(CMD) and I(SENSE). The transconductance gain of A2 is:

$$g_m = 1.875 \times 10^{-4} \text{ } \Omega^{-1}$$

The current loop is compensated by C_{OTA} which forms a pole given by

$$\omega_p = \frac{9.375 \times 10^{-4}}{C_{OTA}}$$

This time constant should be fast enough so that the current loop settles in less than 10% of T_{VCO} at the highest motor speed to avoid torque ripple to V_{TH} mismatch of the N-Channel MOSFETs, or use a separate MOSFET in series with N1-N3 with a lower time constant.

The ML4410 also includes a current mode constant off-time PWM circuit. When motor current builds to the threshold set on I(LIMIT) input (pin 27), a one-shot is fired whose timing is set by C_{OS} . The current in the motor will be controlled by the lower of pin 27 and pin 28.

ML4410

The I(SENSE) input pin should be kept below 1V. If I(SENSE) goes above 1V, a bias current of about $-300\mu\text{A}$ will flow out of pin 12 and the N outputs will be inhibited. Bringing I(SENSE) below .7V removes the bias current to its normal level. For this reason, the noise filter resistor on the I(SENSE) pin ($1\text{K}\Omega$ on Figure 11) should be less than $1.5\text{K}\Omega$.

The noise filter time constant should be less than $1\mu\text{s}$ to avoid excessive phase shift in the I(SENSE) signal.

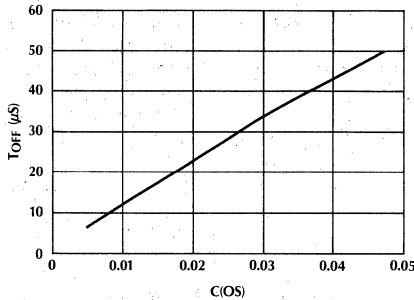


Figure 8. I(LIMIT) Output Off-Time vs. COS.

OUTPUT DRIVERS

The motor's source transistor drivers are open-collector NPN's with internal $50\text{K}\Omega$ pull-up resistors, whose current is controlled according to the current demanded through the motor. To conserve power, the ML4410 sets the current to PNP1, PNP2, and PNP3, proportional to the lower of pin 27 and pin 28.

Drivers N1 through N3 are totem-pole outputs capable of sourcing and sinking 10mA . Switching noise in the external MOSFETs can be reduced by adding resistance in series with the gates.

BRAKING

Applying a 0 on pin 26 activates the braking circuit. The brake circuit turns on PNP1 through PNP3 and turns off NPN1 and NPN3.

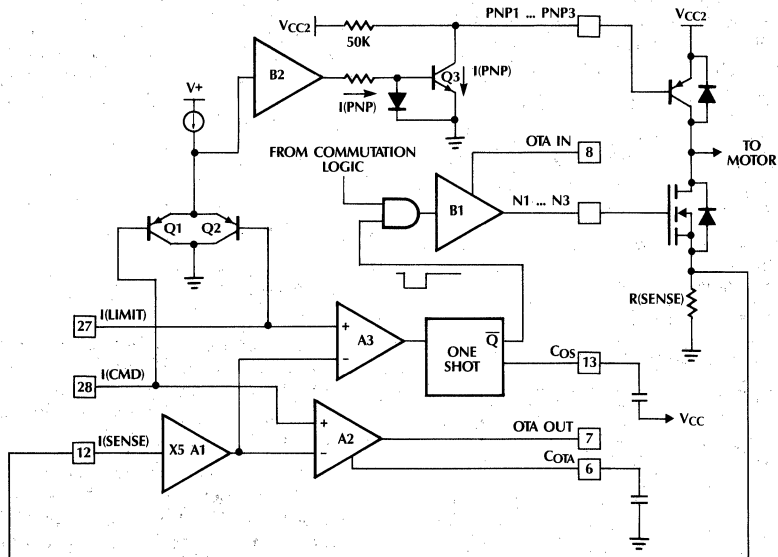


Figure 9. Current Control, Output Drive and Braking Circuits.

ML4410

APPLICATIONS (Continued)

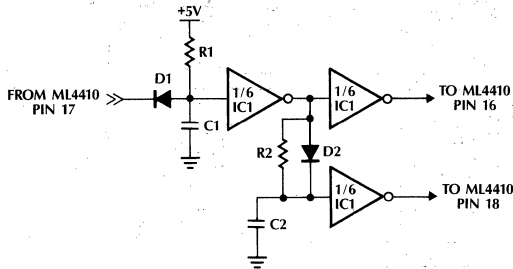
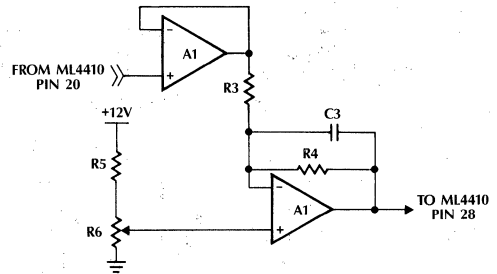


Figure 12. Analog Start-up Circuit.



Symbol	Value
A1	LM358
Q1	74HC14
D1, D2	1N4148
R1	1MΩ
R2	1MΩ
R3	100KΩ

Symbol	Value
R4	100KΩ
R5	50KΩ
R6	50KΩ
C1	3.3μF
C2	3.3μF
C3	.47μF

Figure 13. Analog Speed Control.

Figure 11 shows a typical application of the ML4410 in a hard disk drive spindle control. Although the timing necessary to start the motor in most applications would be generated by a microcontroller, Figure 12 shows a simple "one-shot" start-up timing approach.

Speed control can be accomplished either by:

1. Sensing the VCO OUT frequency with a Microcontroller and adjusting I(CMD) via an analog output from the Micro (PWM DAC).
2. Using analog circuitry for speed control (Figure 13).

OUTPUT STAGE HINTS

Q1, Q2, and Q3 are MJE210 or equivalent. Q4, Q5, and Q6 are IRFU010 or equivalent. Base resistors (100Ω) are included to reduce power dissipation in the IC during start-up. If requested currents are low, these can be eliminated. Switching transients due to commutation can be reduced by increasing the 470Ω gate resistors on Q4-Q6.

Since the output section in a full bridge application consists of three half-H switches, cross-conduction can occur. Cross-conduction is the condition where an N-FET and PNP in the same phase of the bridge conduct simultaneously. This could happen under two conditions (see figure 14):

1. When transitioning from mode 0 to mode A (see table 1) or from braking to mode R, a PNP goes from on to off at the same time N goes from off to on in the same phase. If the PNP turns off slowly and N turns on quickly, cross-conduction may occur. This condition has been prevented inside the IC on later revisions of the ML4410. Consult your Micro Linear representative for date code information. On earlier revision parts, forcing the PNP to turn off more quickly than the NPN turns on will minimize the cross-conduction current.
2. When the MOSFET (or PNP) in the same phase switches on gate current flows due to capacitive coupling of current through the FET's drain to gate capacitance (or PNP's Miller Capacitance). This could cause the device that was off to be turned on.

In Condition 2 above, the PNP is pulled up inside the ML4410 with a 50KΩ resistor. If the current through C(CB) is greater than $0.7V \div 50K$ when the N-FET turns on, the PNP could turn on simultaneously, causing cross-conduction. Adding R1 as shown in figure 14 eliminates this. The size of R1 will depend on the fall time of the phase voltage, and the size of the C(CB).

APPLICATIONS (Continued)

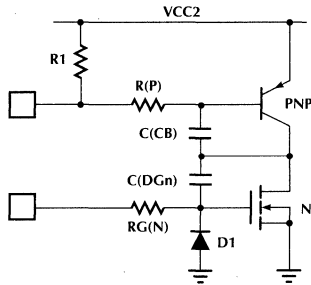


Figure 14. Causes of Cross-conduction.

Adding a series damping resistor to the N-FET gate (R_{Gn}) will slow the fall time. The damping resistor should be low enough to:

Avoid turning on the N-Channel gate when the PNP turns on via the same mechanism outlined in condition 2 above

Not severely increase the switching losses in the N-FET

In higher power applications, when large MOSFETs are used, the N-Output can be pulled below GND, causing the internal substrate diode (Dint) to conduct. The negative substrate current should be limited to less than 2mA, which can be done by adding D1 as shown in figure 14. D1 prevents the gate from going below 0.7V, limiting the substrate current to:

$$\frac{V_{BE(D1)} - V_{BE(Dint)}}{R_{G(N)}}$$

ORDERING INFORMATION

PART NUMBER	TEMPERATURE RANGE	PACKAGE
ML4410CQ	0°C to +70°C	28-Pin Molded PCC (Q28)

Sensorless Spindle Motor Controller

GENERAL DESCRIPTION

The ML4411 provides complete commutation for delta or wye wound Brushless DC (BLDC) motors without the need for signals from Hall Effect sensors. This IC senses the back EMF of the 3 motor windings (no neutral required) to determine the proper commutation phase angle using phase lock loop techniques. This technique will commutate virtually any 3-phase BLDC motor and is insensitive to PWM noise and motor snubbing. The ML4411 is architecturally similar to the ML4410 but with improved braking and brown-out recovery circuitry.

Included in the ML4411 is the circuitry necessary for a Hard Disk Drive microcontroller driven control loop. The ML4411 controls motor current with either a constant off-time PWM or linear current control driven by the microcontroller. Speed feedback for the micro is a stable digital frequency equal to the commutation frequency of the motor. All commutation is performed by the ML4411. Braking and Power Fail are also included in the ML4411.

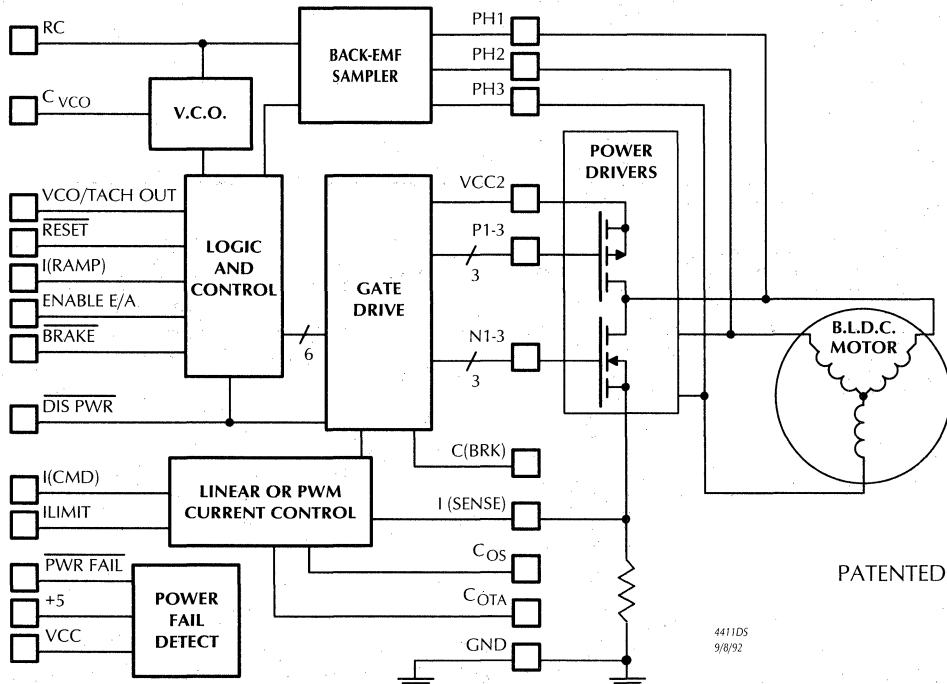
The timing of the start-up sequencing is determined by the micro, allowing the system to be optimized for a wide range of motors and inertial loads.

The ML4411 modulates the gates of external N-Channel power MOSFETS to regulate the motor current. The IC drives P-Channel MOSFETS directly.

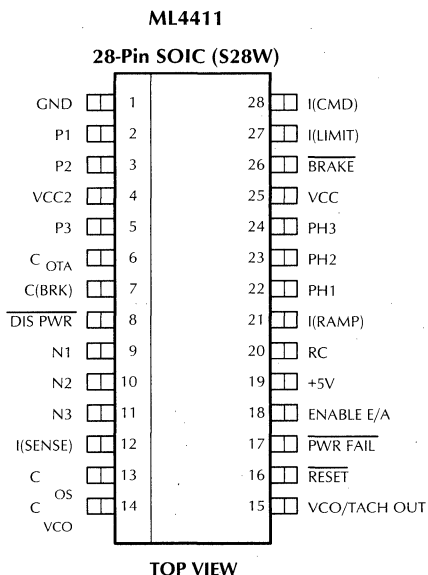
FEATURES

- Back-EMF Commutation Provides Maximum Torque for Minimum "Spin-Up" time for Spindle Motors.
- Accurate, Jitter-Free Phase Locked Motor Speed Feedback Output
- Linear or PWM Motor Current Control
- Easy Microcontroller Interface for Optimized Start-up Sequencing and Speed Control
- Power Fail Detect Circuit with Delayed Braking
- Drives External N-Channel FETs and P-Channel FETs
- Back-EMF comparator detects motor rotation after power fail for fast re-lock after brownout.
- Improved version of ML4410

BLOCK DIAGRAM



PIN CONFIGURATION



PIN DESCRIPTION

PIN# NAME	FUNCTION	PIN# NAME	FUNCTION
1 GND	Signal and Power Ground	15 VCO/TACH OUT	Logic Output from VCO or TACH comparator.
2 P1	Drives the external P-Channel transistor driving motor PH1	16 RESET	Input which holds the VCO off and sets the IC to the RESET condition.
3 P2	Drives the external P-Channel transistor driving motor PH2	17 PWR FAIL	A "0" output indicates 5V or 12V is under-voltage. This is an open collector output with a 4.5KW pull-up to +5V.
4 VCC2	12V power and power for the braking function	18 ENABLE E/A	A "1" logic input enables the error amplifier and closes the back-EMF feedback loop
5 P3	Drives the external P-Channel transistor driving motor PH3	19 +5V	5V power supply input
6 C _{OTA}	Compensation capacitor for linear motor current amplifier loop	20 RC	VCO loop filter components
7 C(BRK)	Capacitor which stores energy to charge N-Channel MOSFETS for braking with power off.	21 I(RAMP)	Current into this pin sets the initial acceleration rate of the VCO during start-up
8 DIS PWR	A logic 0 on this pin turns off the N and P outputs and causes the TACH comparator output to appear on TACH OUT.	22 PH1	Motor Terminal 1
9-11	N1, N2, N3 Drives the external N-channel MOSFETS for PH1, PH2, PH3	23 PH2	Motor Terminal 2
12 I(SENSE)	Motor current sense input	24 PH3	Motor Terminal 3
13 C _{OS}	Timing capacitor for fixed off-time PWM current control	25 VCC	12V power supply. Terminal which is sensed for power fail.
14 C _{VCO}	Timing capacitor for VCO.	26 BRAKE	A "0" activates the braking circuit
		27 I(LIMIT)	Sets the threshold for the PWM comparator
		28 I(CMD)	Current Command for Linear Current amplifier

ML4411

ABSOLUTE MAXIMUM RATINGS

Absolute maximum ratings are those values beyond which the device could be permanently damaged. Absolute maximum ratings are stress ratings only and functional device operation is not implied.

Supply Voltage (pins 4, 25)	14V
Output Current (pins 2,3,5,9,10,11)	±150mA
Logic Inputs (pins 16,17,18,26)	-0.3 to 7V
Junction temperature	150°C
Storage Temperature Range	-65°C to 150°C
Lead Temperature (Soldering 10 sec.)	150°C
Thermal Resistance (θ_{JA})	60°C/W

OPERATING CONDITIONS

Temperature Range	0°C to 70°C
VCC Voltage +12V (pin 25)	12V±10%
+5V (pin 19)	5V±10%
I(RAMP) current (pin 21)	0 to 100µA
I Control Voltage Range (pins 27,28)	0V to 7V

ELECTRICAL CHARACTERISTICS

Unless otherwise specified, T_A =Operating Temperature Range, $V_{CC}=V_{CC2}=12V$, $R_{SENSE}=1\Omega$, $C_{OTA}=C_{VCO}= .01\mu F$, $C_{OS}= .02\mu F$

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
Oscillator (VCO) section					
V _{PIN16} = 5V					
Frequency vs. V _{PIN20}	1V ≤ V _{PIN20} ≤ 10V		300		hz/V
Frequency	V _{VCO} = 6V	1450	1800	2150	hz
	V _{VCO} = .5V	70	140	210	hz
Reset Voltage at C _{VCO}	Mode = 0		125	250	mV
Sampling Amplifier (note 1)					
V _{RC}	State R		125	250	mV
I _{RC}	V _{PIN18} =0V, R _{RAMP} =39KΩ	70	100	130	µA
	V _{PIN18} =5V, State A, V _{PH2} =4V,	30	50	70	µA
	V _{PIN18} =5V, State A, V _{PH2} =6V	-13	2	13	µA
	V _{PIN18} =5V, State A, V _{PH2} =8V	-30	-50	-70	µA
V _{PIN21}	R _{PIN21} = 39KΩ to +5V	1.0	1.1	1.20	V
Motor Current Control Section					
I(SENSE) Gain	V _{PIN27} =5V, 0V ≤ V _{PIN28} ≤ 2.5V	4.5	5	5.5	V/V
One Shot off time		12	25	33	µS
I(CMD) Transconductance Gain			.19		mmho
I(CMD), I(LIM) Bias Current	V _{IN} =0	0	-100	-400	nA
Power Fail Detection Circuit					
12V Threshold		9.1	9.8	10.5	V
Hysteresis			150		mV
5V Threshold		3.8	4.25	4.5	V
Hysteresis			70		mV
Logic Inputs					
Voltage High (V _{IH})		2			V
Voltage Low (V _{IL})				.8	V
Current High (I _{IH})	V _{IN} =2.7V	-10	1	10	µA
Current Low (I _{IL})	V _{IN} =0.4V	-500	-350	-200	µA
Braking Circuit					
V _{PIN17} =0V					
Brake Active Threshold		0.8	1.2	1.6	V
PIN 26 Bias Current	V _{PIN26} = 0V		0.3	1	µA
N-Channel Leakage	V _{CC} , V _{CC2} =0V	0	.06	10	nA
	V _{PIN17} =0V, V _N =4V				
C(BRK) Current	V _{CC} , V _{CC2} = 0V, V _{PIN26} = 3V V _{PIN7} = 6V		20	85	µA

ELECTRICAL CHARACTERISTICS (continued)

Unless otherwise specified, T_A =Operating Temperature Range, $V_{CC}=V_{CC2}=12V$, $R_{SENSE}=1\Omega$, $C_{0TA}=C_{VCO}=.01\mu F$, $C_{OS}=.02\mu F$

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
Outputs					
$I(CMD) = I(LIMIT) = 2.5V$					
I_p low	$V_p = 0.8V$	5	7	19.5	mA
	$V_p = 0.4V$	2	4		mA
V_p high	$I_p = -10\mu A$	$V_{CC2} - 0.4$			V
V_N high	$V_{PIN12} = 0V$	$V_{CC} - 3.2$	10	$V_{CC} - 1.2$	V
V_N low	$I_N = 1mA$		0.2	0.7	V
LOGIC low (V_{OL})	$I_{OUT}=0.4mA$			0.5	V
VCO/TACH V_{OH}	$I_{OUT}=-100\mu A$	2.4			V
POWER FAIL V_{OH}	$I_{OUT}=-10\mu A$	$V_{PIN19}-0.2$	$V_{PIN19}-0.1$	V_{PIN19}	V
Supply Currents (N and P outputs open)					
5V Current			3	4	mA
VCC Current			38	50	mA
VCC2 Current			2	3	mA

Note 1. For explanation of states, see Figure 5 and Table 1.

FUNCTIONAL DESCRIPTION

The ML4411 provides closed-loop commutation for 3-phase brushless motors. To accomplish this task, a VCO, Integrating Back-EMF Sampling error amplifier and sequencer form a phase-locked loop, locking the VCO to the back-emf of the motor. The IC also contains circuitry to control motor current with either linear or constant off-time PWM modes. Braking and power fail detection functions are also provided on chip. The ML4411 is designed to drive external power transistors (N-channel sinking transistors and P-Channel sourcing transistors) directly.

Start-up sequencing and motor speed control are accomplished by a microcontroller. Speed sensing is accomplished by monitoring the output of the VCO, which will be a signal which is phased-locked to the commutation frequency of the motor.

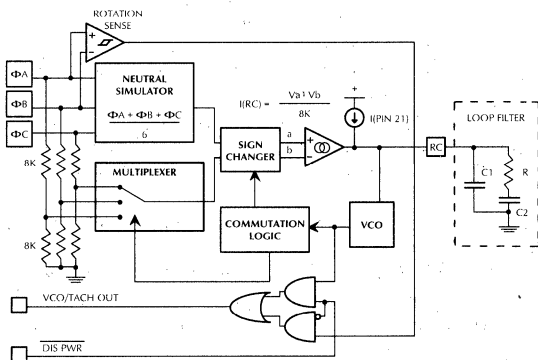


Figure 1. Back EMF sensing block diagram

BACK-EMF SENSING AND COMMUTATOR

The ML4411 contains a patented back-emf sensing circuit which samples the phase which is not energized (Shaded area in fig. 2 below) to determine whether to increase or decrease the commutator (VCO) frequency. A late commutation causes the error amplifier to charge the filter (RC) on pin 20, increasing the VCO input while early commutation causes pin 20 to discharge. Analog speed control loops can use Pin 20 as a speed feedback voltage.

The input impedance of the three PH inputs is about $8K\Omega$ to GND. When operating with a higher voltage motor, the PH inputs should be divided down in voltage so that the maximum voltage at any PH input does not exceed VCC.

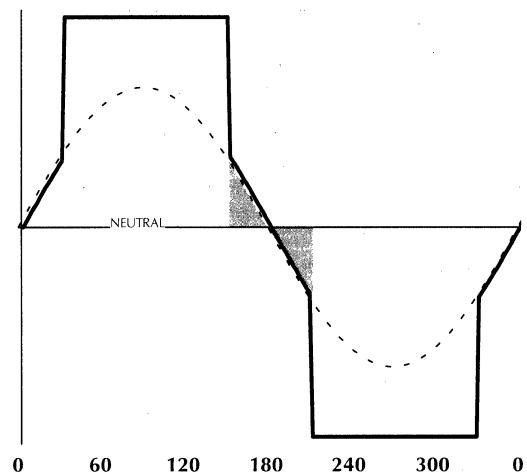


Figure 2. Typical motor phase waveform with Back-EMF superimposed (Ideal Commutation)

ML4411

VCO AND PHASE DETECTOR CALCULATIONS

The VCO should be set so that at the maximum frequency of operation (the running speed of the motor) the VCO control voltage will be no higher than $V_{CCMIN} - 1V$. The VCO maximum frequency will be:

$$F_{MAX} = .05 \times POLES \times RPM$$

where POLES is the number of poles on the motor and RPM is the maximum motor speed in Revolutions Per Minute.

The minimum VCO gain derived from the specification table (using the minimum F_{VCO} at $V_{VCO}=6V$) is:

$$K_{VCO(MIN)} = \frac{2.42 \times 10^{-6}}{C_{VCO}}$$

Assuming that the $V_{VCO(MAX)} = 9.5V$, then

$$C_{VCO} = \frac{9.5 \times 2.42 \times 10^{-6}}{F_{MAX}}$$

or

$$C_{VCO} = \frac{460}{POLES \times RPM} \mu F$$

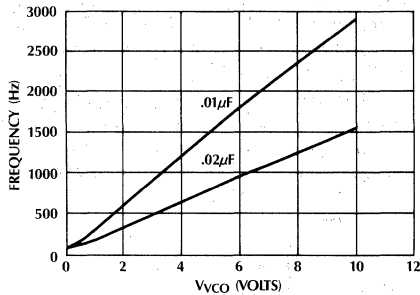


Figure 3. VCO Output Frequency vs. V_{vco} (pin20)

Figure 4 below shows the transfer function of the Phase Lock Loop with the phase detector formed from the sampled phase through the G_m amplifier with the loop filtered formed by R , C_1 , and C_2 .

The impedance of the loop filter is

$$Z_{RC}(s) = \frac{1}{C_1 s} \frac{(s + \omega_{LEAD})}{(s + \omega_{LAG})}$$

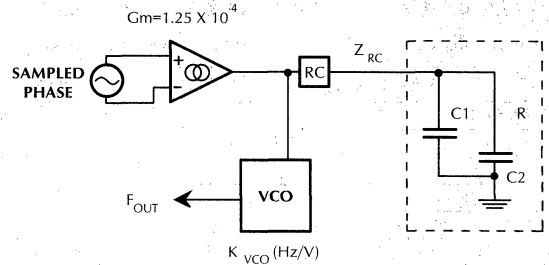


Figure 4. Back EMF Phase Lock Loop Components

Where the lead and lag frequencies are set by:

$$\omega_{LEAD} = \frac{1}{R C_2}$$

$$\omega_{LAG} = \frac{C_1 + C_2}{R C_1 C_2}$$

Requiring the loop to settle in 20 PLL cycles with a spread of 10 between $\omega_{LEAD} = 10 \times \omega_{LAG}$ produces the following calculations for R , C_1 and C_2 :

$$C_1 \approx \frac{4.66 \times 10^{-9}}{C_{VCO} \times F_{VCO}^2}$$

$$C_2 = 9 \times C_1$$

$$R = \frac{12.54}{C_2 \times F_{VCO}}$$

START-UP SEQUENCING

When the motor is initially at rest, it is generating no back-EMF. Because a back-EMF signal is required for closed loop commutation, the motor must be started "open-loop" until a velocity sufficient to generate some back-EMF is attained (around 100 RPM). The following steps are a typical procedure for starting a motor which is at rest. It is possible to determine if the motor is running by polling the VCO/TACH OUT pin with power disabled (Pin 8 = low).

Step 1: The IC is held in reset (state R) with full power applied to the windings (see fig. 6). This aligns the rotor to a position which is 30° (electrical) before the center of the first commutation state.

Step 2: Reset is released, and a fixed current is input to pin 21 and appears as a current on pin 20, and will ramp the VCO input voltage, accelerating the motor at a fixed rate.

Step 3: When the motor speed reaches about 100 RPM, the back EMF loop can be closed by pulling pin 18 high.

STATE	OUTPUTS						INPUT SAMPLING
	N1	N2	N3	P1	P2	P3	
R OR 0	OFF	ON	OFF	ON	OFF	ON	N/A
A	OFF	OFF	ON	ON	OFF	OFF	PH2
B	OFF	OFF	ON	OFF	ON	OFF	PH1
C	ON	OFF	OFF	OFF	ON	OFF	PH3
D	ON	OFF	OFF	OFF	OFF	ON	PH2
E	OFF	ON	OFF	OFF	OFF	ON	PH1
F	OFF	ON	OFF	ON	OFF	OFF	PH3

Table 1. Commutation States

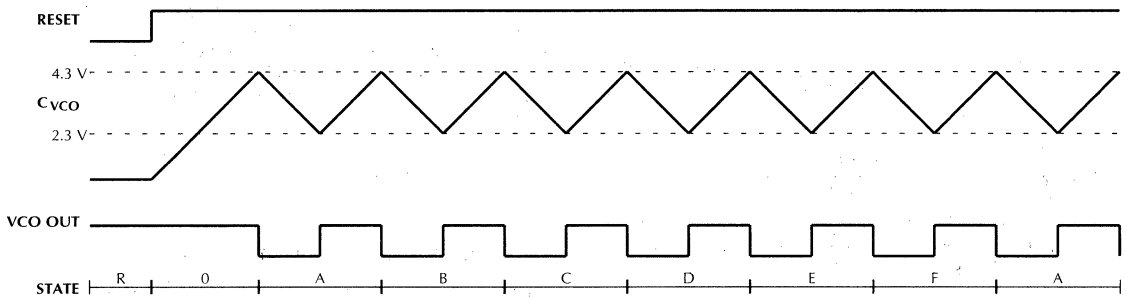


Figure 5. Commutation Timing and Sequencing

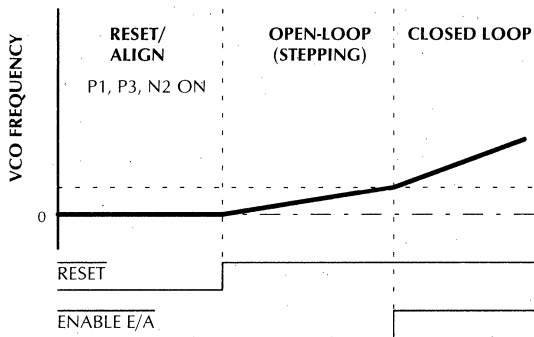


Figure 6. Typical Start-up Sequence

Using this technique, some reverse rotation is possible. The maximum amount of reverse rotation is $360/N$, where N is the number of poles. For an 8 pole motor, 45° reverse rotation is possible.

For quick recovery following a momentary power failure, the following steps can be taken:

STEP	PIN 16	PIN 18	PIN 21	I(LIMIT) I(CMD)
1	0	0	FIXED	I_{MAX}
2	1	0	FIXED	I_{MAX}
3	1	1	0	I_{MAX}

Table 2. Start-up Sequence

Step 1a: The IC is held in reset (state R) with I(CMD) low and DIS PWR low. The Micro Processor monitors the VCO/TACH OUT pin to determine if a signal is present. If a signal is present, the frequency is determined (by measuring the period). If a signal is not present, proceed to the routine described above for starting a motor which is at rest.

Step 2a: Release RESET and DIS PWR. Apply a current to pin 21 and monitor the VCO/TACH OUT pin for VCO frequency.

Step 3a: When the VCO frequency approaches 6 X the motor frequency (or where the motor frequency has decelerated to by coasting during the time the VCO frequency was ramping up) the back EMF loop can be closed by pulling pin 18 high and motor current brought up with I(CMD) or I(LIMIT).

ADJUSTING OPEN LOOP STEP RATE

I_{RAMP} should be set so that the VCO's frequency ramp during "open loop stepping" phase of motor starting is less than the motor's acceleration rate. In other words, the motor must be able to keep up with the VCO's ramp rate in open loop stepping mode. The VCO's input voltage (V_{PIN20}) ramp rate is given by:

$$\frac{dV_{VCO}}{dt} \approx \frac{I_{RAMP}}{C_1 + C_2}$$

since

$$F_{VCO} = K_{VCO} \times V_{VCO}$$

$$K_{VCO(MAX)} \approx \frac{4 \times 10^6}{C_{VCO}}$$

then combining the 3 equations I_{RAMP} can be calculated from the desired maximum open loop stepping rate the motor can follow:

$$I_{RAMP} < \frac{dF_{VCO}}{dt} \frac{C_{VCO} \times (C_1 + C_2)}{4 \times 10^6}$$

The motor will start more consistently and tolerate a wider variation in open loop step rate if there is some damping on the motor (such as head drag) during the open loop modes.

The tolerance of the open loop step VCO acceleration

$\left(\frac{dF_{VCO}}{dt}\right)$ depends on the tolerances of K_{VCO} , I_{RAMP} , C_1 ,

C_2 , and C_{VCO} . For more optimum spin up times, these variables can be digitally "calibrated" out by the microprocessor using the following procedure:

1. Reset the IC by holding pin 16 low for at least $5\mu s$.
2. Go into open loop step mode with no current on the motor and measure the difference between the first two complete VCO periods with the PWM signal at 50% duty cycle:

ENABLE E/A= (see below)

I(CMD) = 0V

PWM OUT=50%

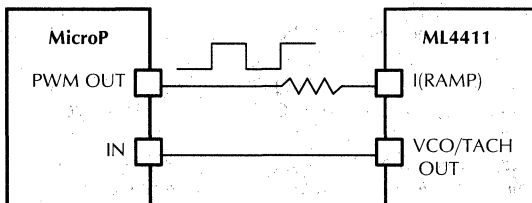


Figure 7. Auto-Calibration of open-loop step rate

3. Compute a correction factor to adjust I_{RAMP} current by changing the PWM duty cycle from the Micro (D.C.)

$$D.C.(NEW) = 50\% \times \frac{\Delta F_{VCO}(DESIRED)}{\Delta F_{VCO}(MEASURED)}$$

4. Use new computed duty cycle for open loop stepping mode and proceed with a normal start-up sequence.

If this auto calibration is used ENABLE E/A can be tied permanently high, eliminating a line from the Micro. Since there is offset associated with the Phase Detector Error Amp (E/A), more current than is being injected by I_{RAMP} may be taken out of pin 20 if the offset is positive (into pin 20) if the error amp were enabled during the open loop stepping mode. In that case, V_{VCO} would not rise and the motor would not step properly. The effect of E/A offset can also be canceled out by the auto calibration algorithm described above allowing the E/A to be permanently enabled.

$$\Lambda_V = \frac{1.875 \times 10^{-4}}{sC_{OTA}}$$

PWM AND LINEAR CURRENT CONTROL

To facilitate speed control, the ML4411 includes two current control loops - linear and PWM (fig. 9). The linear control loop senses the motor current on the I(SENSE) terminal through R_{SENSE} . An internal current sense amplifier's (A2) output modulates the gates of the 3 N-channel MOSFET's.

The ML4411 also includes a current mode constant off-time PWM circuit. When motor current builds to the threshold set on I(LIMIT) input (pin 27), a one-shot is fired whose timing is set by C_{OS} . The current in the motor will be controlled by the lower of pin 27 and pin 28.

The linear current control modulates the gates of the external MOSFET drivers. Amplifier A2 is a transconductance amplifier which amplifies the difference between I(CMD) and I(SENSE). The transconductance gain of A2 is:

$$g_m = 1.875 \times 10^{-4} \text{ } \Omega$$

The current loop is compensated by C_{OTA} which forms a pole given by

$$\omega_p = \frac{9.375 \times 10^{-4}}{C_{OTA}}$$

This time constant should be fast enough so that the current loop settles in less than 10% of T_{VCO} at the highest motor speed to avoid torque ripple to V_{TH} mismatch of the N-Channel MOSFET's.

The I(SENSE) input pin should be kept below 1V. If I(SENSE) goes above 1V, a bias current of about $-300\mu A$ will flow out of pin 12 and the N outputs will be inhibited. Bringing I(SENSE) below .7V returns the bias current to its normal

level. For this reason, the noise filter resistor on the I(SENSE) pin (1KΩ on Figure 10) should be less than 1.5KΩ.

The noise filter time constant should be great enough to filter the leading edge current spike when the N-FETs turn on but small enough to avoid excessive phase shift in the I(SENSE) signal.

OUTPUT DRIVERS

The motor's source drivers (P1 thru P3) are open-collector NPN's with internal 16KΩ pull-up resistors.

Drivers N1 through N3 are totem-pole outputs capable of sourcing and sinking 10mA. Switching noise in the external MOSFETs can be reduced by adding resistance in series with the gates.

BRAKING

As shown in figure 9, the braking circuit pulls the N-Channel MOSFET gates high when BRAKE falls below a 1.4V threshold. After a power failure, C(DLY) is discharged slowly through R(DLY) providing a delay for retract to occur before the braking circuit is activated. The N-Channel buffer (B1) tri-states when the BRAKE pin reaches 2.1V to ensure that no charge from C(BRK) is lost through the pull-down transistor in B1. To brake the motor with external signals, first disable power by pulling pin 8 low, then pull

pin 26 below 1.4V using an open drain (or diode isolated) output.

The bias current for the Braking circuits comes from VCC2. When the N-Channel MOSFETs turn on, no additional power is generated for VCC2 (motor back-EMF rectified through the MOSFET body diodes). After VCC2 drops below 4V, Q2 turns off. Continued braking relies on the C_{GS} of the N-Channel MOSFETs to sustain the MOSFET gate enhancement voltage.

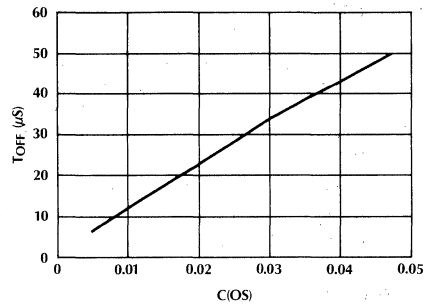


Figure 8. I (LIMIT) output off-time vs. C_{OS}

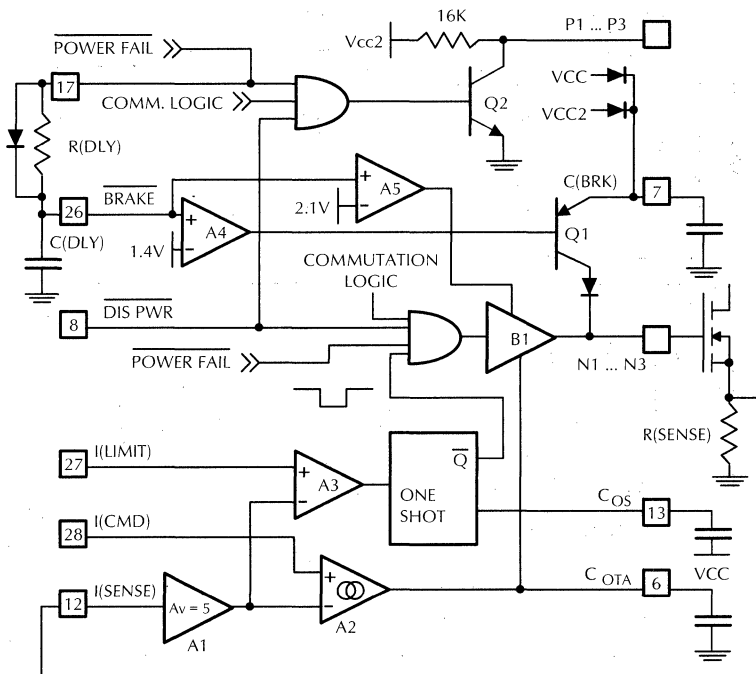


Figure 9. PWM and Linear Current Control, Gate Drive and Braking Circuits

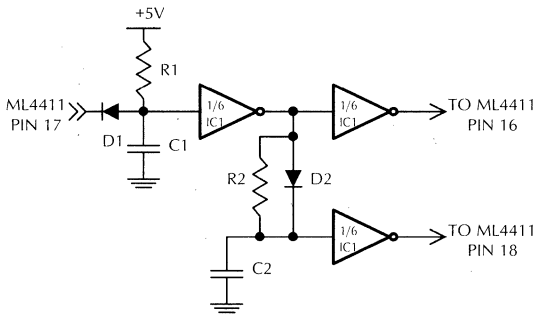
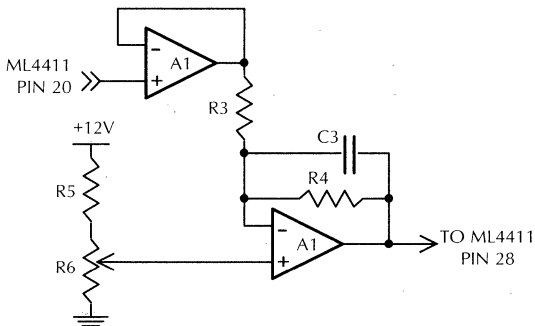


Figure 11. Analog Start-up Circuit



SYMBOL	VALUE	SYMBOL	VALUE
A1	LM358	R4	100KΩ
IC1	74HC14	R5	50KΩ
D1, D2	1N4148	R6	50KΩ
R1	1MΩ	C1	3.3 μF
R2	1MΩ	C2	3.3 μF
R3	100KΩ	C3	.47 μF

Figure 12. Analog Speed Control

P-FET in the same phase of the bridge conduct simultaneously. This could happen under two conditions (see figure 13):

1. When transitioning from mode 0 to mode A (see table 1) P3 goes from on to off at the same time N3 goes from off to on. If P3 turns off slowly and N3 turns on quickly, cross-conduction may occur. This condition has been prevented inside the IC on later revisions of the ML4411. Consult your Micro Linear representative for date code information. On earlier revision parts, forcing P3 to turn off quickly will minimize the cross-conduction current.
2. When the a MOSFET in the same phase switches on gate current flows due to capacitive coupling of current through the MOSFET's drain to gate capacitance. This could cause the MOSFET that was off to be turned on.

In Condition 2 above, the P-Channel MOSFET is pulled up inside the ML4411 with a 16KΩ resistor. If the current through C(DGp) is greater than $V_{TH} \approx 16K$ when the N-FET turns on, the P-FET could turn on simultaneously, causing cross-conduction. Adding R1 as shown in fig. 13 eliminates this. The size of R1 will depend on the fall time of the phase voltage, and the size of the C(DGp). D1 may be needed for high power applications to limit the negative current pulled (through C(DGn)) out of the substrate diode in the ML4411 when P-FET turns off.

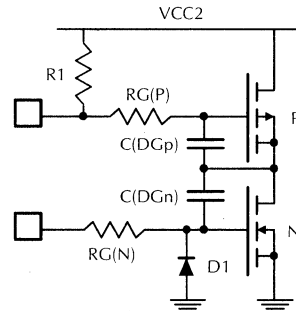


Figure 13. Causes of Cross-conduction

Adding a series damping resistor to the N-FET gate (RGn) will slow the fall time. The damping resistor should be low enough to:

Avoid turning on the N-Channel gate when the P-Channel turns on via the same mechanism outlined in condition 2 above

Not severely increase the switching losses in the N-FET

UNIPOLAR OPERATION

Unipolar mode offers the potential advantage of lower motor drive cost by only requiring the use of 3 transistors to drive the motor. The ML4411 will operate in unipolar mode (fig. 14) provided the following precautions are taken:

1. The IC supplies should not exceed 12V+10%.
2. The phase pins on the IC should not exceed the supply voltage.

In unipolar operation, the motor's windings must be allowed to drive freely to:

$$V_{\Phi(MAX)} = V_{SUPPLY(MAX)} + V_{EMF(MAX)}$$

Therefore, there can be no diodes to clamp the inductive energy to V_{SUPPLY} . This energy must be clamped, however, to avoid an over-voltage condition on the MOSFETs and other components. Typically, a V_{CLAMP} voltage is created to provide the clamping voltage. The inductive energy may either be dissipated (fig 15) or alternately efficiently regenerated back to the system supply (fig 16).

The circuit in Figure 14 is designed to minimize the external components necessary, at some compromise to performance. The 3 resistors from the motor phase windings to the PH inputs work with the ML4411's 8KΩ internal resistance to ground to divide the motor's phase voltage down, providing input signals that do not exceed 12V.

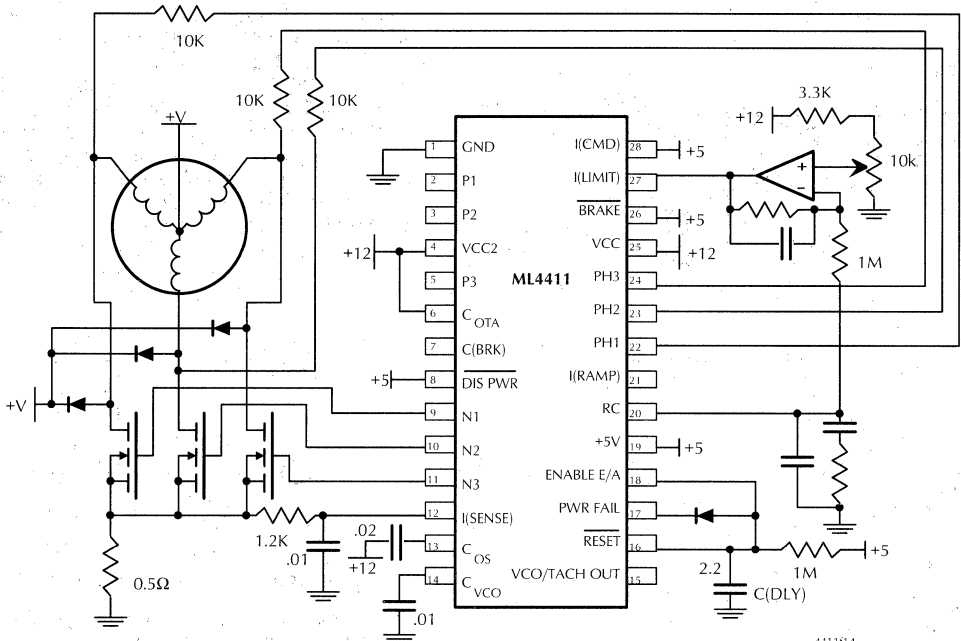


Figure 14. ML4411 Unipolar Drive Application

This circuit uses analog speed regulation. The 1MΩ resistor from pin 20 to the speed regulation op amp provides the function of injecting current into the VCO loop filter for the open loop stepping phase of start-up operation. The "one shot" circuitry to time the reset is replaced by a diode and RC delay from the rising edge or the POWERFAIL signal. The error amplifier is left enabled continuously since at low speeds its current contribution is negligible. The current injected into the loop filter must be greater than the leakage current from the phase detector amplifier for the motor to start reliably.

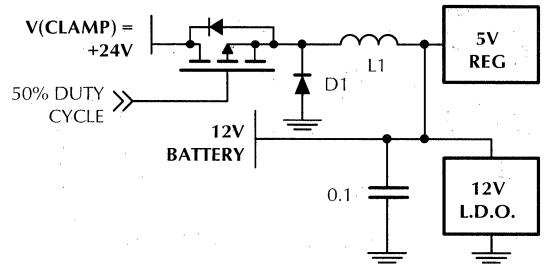


Figure 16. Non-Dissipative Clamping Technique

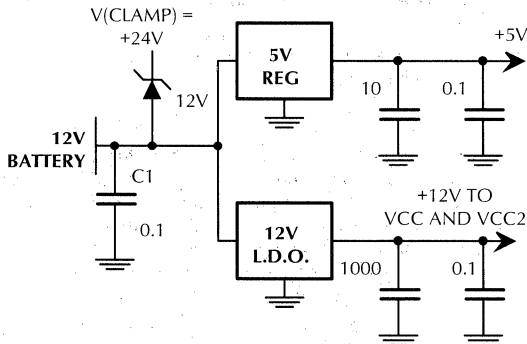


Figure 15. Dissipative Clamping Technique

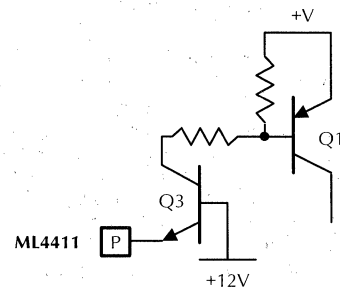


Figure 17. High Voltage Translation using PNP Power Transistor

HIGHER VOLTAGE MOTOR DRIVE

To drive a higher voltage motor, the same precautions regarding ML4411 voltage limitations as were outlined for Unipolar drive above should be followed. Figs. 17-19 provide several methods of translating the ML4411's P outputs to drive a higher voltage.

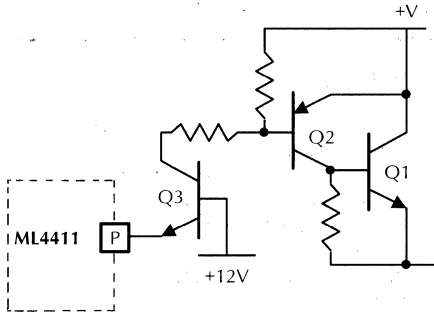


Figure 18. High Voltage Translation using "Composite" PNP Power Transistor

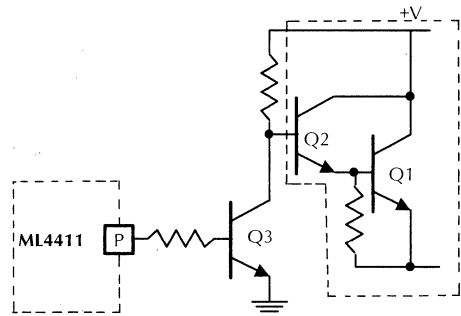


Figure 19. High Voltage Translation with NPN Darlington

ORDERING INFORMATION

PART NUMBER	TEMPERATURE RANGE	PACKAGE
ML4411CS	0°C to +70°C	28-PIN SOIC (S28W)



ML4415, ML4415R ML4416, ML4416R

15 Channel Read/Write Circuit

GENERAL DESCRIPTION

The ML4415, ML4416 devices are bipolar monolithic read/write circuits designed for use with fixed disk ferrite center-tapped recording heads. They provide a low noise read path, write current control, and data protection circuitry for all channels.

These multiplexed read/write data channels exhibit features not found in similar read/write circuits such as improved write current stability and elimination of write current "glitches" during power up.

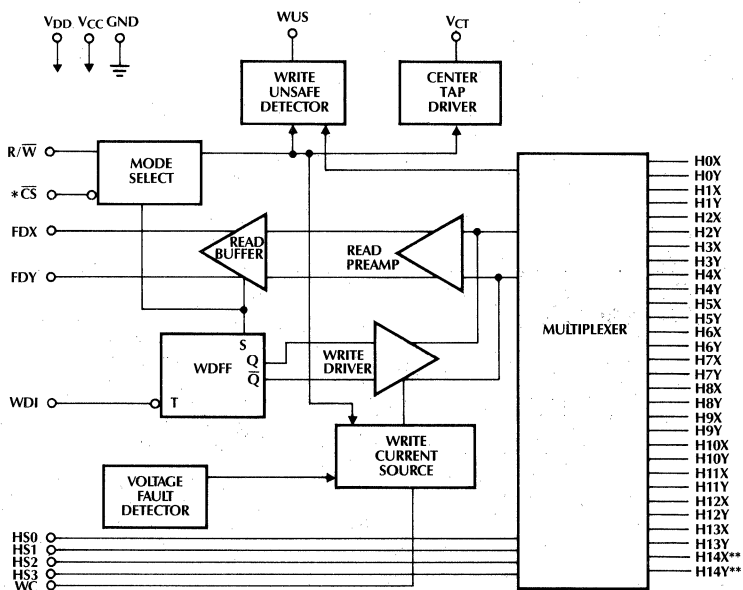
The ML4416 has fourteen read/write data channels and a chip select pin. The chip select pin allows additional read/write circuits in the system by enabling or disabling a particular chip. The ML4415 has fifteen read/write data channels and no chip select pin.

The ML4415R and ML4416R versions include on-chip damping resistors.

FEATURES

- Write current disable during power up
- Enhanced write current stability
- Designed for center-tapped ferrite heads
- ML4415 provides 15 read/write channels
- ML4416 — easily multiplexed for larger systems
- Includes write unsafe detection
- TTL compatible control signals
- Programmable write current source
- +5V, +12V power supplies

BLOCK DIAGRAM

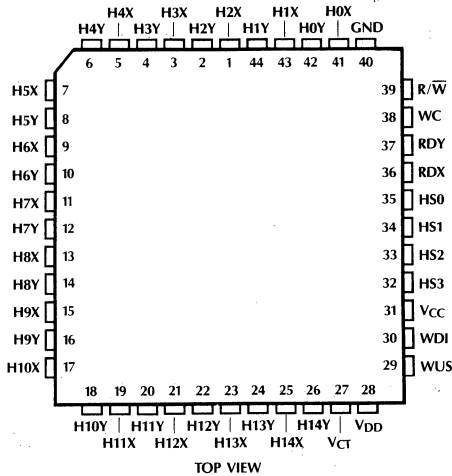


* ML4416 ONLY
** ML4415 ONLY

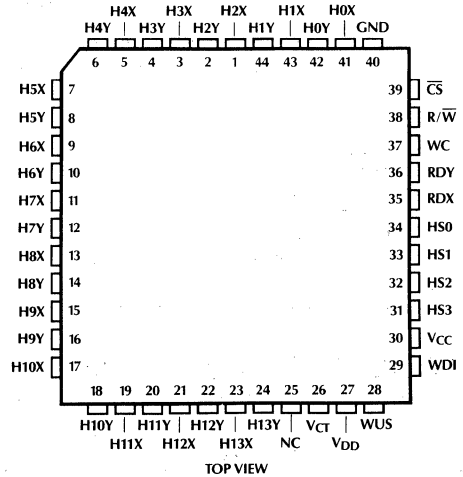
ML4415, ML4415R, ML4416, ML4416R

PIN CONNECTIONS

ML4415CQ, ML4415RCQ
44-Pin PCC



ML4416CQ, ML4416RCQ
44-Pin PCC



PIN DESCRIPTION

NAME	FUNCTION	NAME	FUNCTION
HS0-HS3	Head Select (14 heads for the ML4416, and 15 heads for ML4415).	H0X-H14X	X head connections
$\overline{\text{CS}}$	Chip Select (low level enables, ML4416 only)	H0Y-H14Y	Y head connections
R/W	Read/Write (high level select Read Mode)	RDX, RDY	X, Y Read Data (differential read signal out)
WUS	Write Unsafe, open collector output (high level indicates an unsafe writing condition)	WC	Write Current (used to set the write current magnitude)
WDI	Write Data In (negative transition toggles head current direction)	V _{CT}	Voltage Center Tap (center tap voltage source)
		V _{CC}	+5 volts
		V _{DD}	+12 volts
		GND	Ground

ML4415, ML4415R, ML4416, ML4416R

ABSOLUTE MAXIMUM RATINGS

(Note 1)

Power Supply Voltage Range	
V _{DD1}	-0.3 to 14V _{DC}
V _{DD2}	-0.3 to 14V _{DC}
V _{CC}	-0.3 to 6V _{DC}
Input Voltage Range	
Digital Inputs (CS, R/W, HS, WDI)	-0.3 to V _{CC} +0.3V _{DC}
Head Ports	-0.3 to V _{DD1} +0.3V _{DC}
Write Unsafe (WUS)	-0.3 to 14V _{DC}
Write Current (I _W)	60 mA
Output Current	
Read Data (RDX, RDY)	-10 mA
Center Tap Current (I _{CT})	-60 mA
Write Unsafe (WUS)	12 mA
Storage Temperature	-65°C to 150°C
Junction Temperature (T _J)	135°C
Lead Temperature (Soldering 10 sec.)	300°C

OPERATING CONDITIONS

Supply Voltage	
V _{DD1}	12V ± 10%
V _{CC}	5V ± 10%
Head Inductance	
L _H	5 to 15μH
Damping Resistor (R _D , ML4415R or ML4416R)	
	500 to 2000Ω
RCT Resistor (1/4 Watt)	
	120Ω ± 5%
Write Current (I _W)	
	10 to 40 mA

ELECTRICAL CHARACTERISTICS

Unless otherwise specified V_{DD1} = V_{DD2} = 12V ± 10%, V_{CC} = 5V ± 10%, R_{CT} = 120Ω ± 5%, I_W = 40 mA, 0°C ≤ T_A ≤ 70°C (Notes 2 and 3).

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
DC OPERATING CHARACTERISTICS						
POWER SUPPLY						
I _{CC}	V _{CC} Supply Current	Read or Idle Mode		31	35	mA
		Write Mode		26	30	mA
I _{DD}	V _{DD} Supply Current	Read Mode		29	35	mA
		Write Mode		17 + I _W	20 + I _W	mA
		Idle Mode		17	20	mA
P _D	Power Dissipation	Read Mode		550	655	mW
		Write Mode I _W = 40 mA, R _{CT} = 0Ω		890	960	mW
		Idle Mode		378	455	mW
DIGITAL INPUTS (CS, R/W, HS, WDI)						
V _{IH}	High Voltage		2			V _{DC}
V _{IL}	Low Voltage				0.8	V _{DC}
I _{IH}	High Current	V _{IH} = 2.0V			100	μA
I _{IL}	Low Current	V _{IL} = 0.8V	-0.4			mA
WUS OUTPUT						
V _{OL}	Output Low Voltage	I _{OL} = 8 mA (Safe)			0.5	V _{DC}
I _{OH}	Output High Current	V _{OH} = 5V (Unsafe)			100	μA
CENTER TAP VOLTAGES						
V _{CT}	Read Mode	Read Mode		4		V _{DC}
V _{CT}	Write Mode	Write Mode		6		V _{DC}

ML4415, ML4415R, ML4416, ML4416R

ELECTRICAL CHARACTERISTICS (Continued)

Unless otherwise specified $V_{DD1} = 12V \pm 10\%$, $V_{CC} = 5V \pm 10\%$, $R_{CT} = 120\Omega \pm 5\%$, $I_W = 35mA$, $L_H = 10\mu H$, $R_D = 750\Omega$ (ML4415, ML4416), $f_{DATA} = 5MHz$, C_L (RDX, RDY) $\leq 20pF$, $0^\circ C \leq T_A \leq 70^\circ C$ (Notes 2 and 3) (V_{IN} is referenced to V_{CT} for Read Mode Characteristics).

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
WRITE MODE CHARACTERISTICS						
I_{HCW}	Head Current (per side)	Write Mode $0 \leq V_{CC} \leq 3.7V$ $0 \leq V_{DD1} \leq 8.7V$	-200	0.15	200	μA
I_{WR}	Write Current Range	$I_W = K/R_{WC}$	10		40	mA
K	Write Current Constant		2.375	2.5	2.625	
V_{HD}	Differential Head Voltage Swing		7.0	10.2		V_{PK}
I_{HU}	Unselected Head Transient Current				2	mA_{PK}
C_{OD}	Differential Output Capacitance			8.8	15	pF
R_{OD}	Differential Output Resistance	ML4415, 4416	10k			Ω
		$T_J = 25^\circ C$ ML4415R, 4416R	600		960	Ω
f_{WDI}	WDI Transition Frequency	WUS = Low	250	490		kHz
A_I	I_{WC} to Head Current Gain			0.99		mA/mA
I_L	Unselected Head Leakage	Sum of X & Y Side Leakage Current			85	μA
READ MODE CHARACTERISTICS						
A_V	Differential Voltage Gain	$V_{IN} = 1mV_{P,P}$ @ 300kHz, R_L (RDX, RDY) = 1k Ω	85	106	115	V/V
DR	Dynamic Range	DC Input Voltage (V_I) Where Gain Falls 10%, $V_{IN} = V_I + 0.5mV_{P,P}$ @ 300kHz	-3	± 7	+3	mV
BW	Bandwidth (-3dB)	$ Z_S < 5\Omega$, $V_{IN} = 1mV_{P,P}$	30	40		MHz
e_{IN}	Input Noise Voltage	BW = 15MHz, $L_H = 0$, $R_H = 0$		1.2	1.5	nV/ \sqrt{Hz}
C_{IN}	Differential Input Capacitance	$f = 5MHz$		14	20	pF
R_{IN}	Differential Input Resistance	$f = 5MHz$, $T_J = 25^\circ C$ ML4415, 4416	2k	15K		Ω
		$V_{IN} = 6mV_{P,P}$ ML4415R, 4416R	460		860	Ω
I_{HCR}	Head Current (per side)	Read or Idle Mode $0 \leq V_{CC} \leq 5.5V$ $0 \leq V_{DD1} \leq 13.2V$	-200		200	μA
I_{IN}	Input Bias Current (1 side)			8.5	45	μA
CMRR	Common-Mode Rejection Ratio	$V_{CM} = V_{CT} + 100mV_{P,P}$ @ $f = 5MHz$	50	77		dB
PSRR	Power Supply Rejection Ratio	100mV $_{P,P}$ @ 5MHz on V_{DD1} , V_{DD2} , or V_{CC}	45			dB
CS	Channel Separation	Unselected Channels: $V_{IN} = 100mV_{P,P}$ @ 5MHz and Selected Channel: $V_{IN} = 0mV_{P,P}$	45	57		dB
V_{OS}	Output Offset Voltage	Read Mode	-460	± 29	+460	mV
		Write or Idle Mode	-20	± 1	+20	mV
V_{OCM}	Common-Mode Output Voltage	Read Mode	4.5	5.5	6.5	V
		Write or Idle Mode		5.6		V
R_{OUT}	Single-Ended Output Resistance	$f = 5MHz$			30	Ω
I_L	Leakage Current, RDX, RDY	(RDX, RDY) = 6V Write or Idle Mode	-100	± 15	100	μA
I_O	Output Current	AC Coupled Load, RDX to RDY	± 2.1	± 2.7		mA

ML4415, ML4415R, ML4416, ML4416R

ELECTRICAL CHARACTERISTICS (Continued)

Unless otherwise specified $V_{DD1} = 12V \pm 10\%$, $V_{CC} = 5V \pm 10\%$, $R_{CT} = 120\Omega \pm 5\%$, $I_W = 35mA$, $L_H = 10\mu H$, $R_D = 750\Omega$ (ML4415, ML4416), $f_{DATA} = 5MHz$, $0^\circ C \leq T_A \leq 70^\circ C$ (Notes 2 and 3)

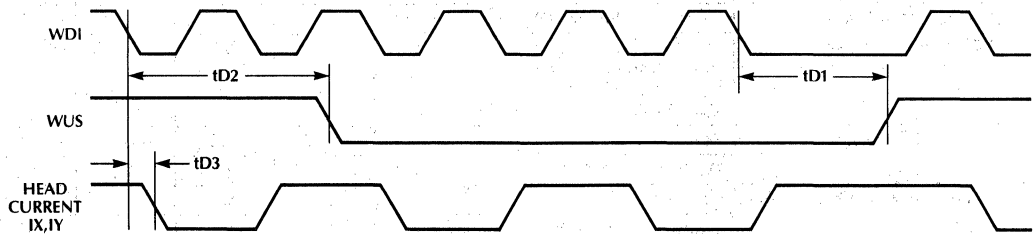
SYMBOL	PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
SWITCHING CHARACTERISTICS						
t_{RW}	R/ \bar{W} to Write Switching Delay	To 90% of Write Current Output		.105	1	μs
t_{WR}	R/ \bar{W} to Read Switching Delay	To 90% of 100mV, 10MHz Read Signal Envelope or to 90% Decay of Write Current		.036	1	μs
t_{IW} or t_{IR}	\bar{CS} to Select Switching Delay	To 90% of Write Current or to 90% of 100mV, 10MHz Read Signal Envelope		.165	1	μs
t_{WI} or t_{RI}	\bar{CS} to Unselect Switching Delay	To 90% Decay of 100mV, 10MHz Read Signal Envelope or to 90% Decay of Write Current		.084	1	μs
t_{HS}	Head Select Switching Delay	To 90% of 100mV, 10MHz Read Signal Envelope		.045	1	μs
t_{D1}	Safe to Unsafe Write Unsafe Delay	$I_W = 35mA$	1.6	3.9	8	us
t_{D2}	Unsafe to Safe Write Unsafe Delay	$I_W = 35mA$.387	1	us
t_{D3}	Prop. Delay Head Current	$L_H = 0$, $R_H = 0$ From 50% points		23	25	ns
	Asymmetry Head Current	WDI has 50% Duty Cycle and 1ns Rise/Fall Time		0.9	2	ns
	Rise/Fall Head Current	10% and 90% Points		5	20	ns

Note 1: Absolute maximum ratings are limits beyond which the life of the integrated circuit may be impaired. All voltages unless otherwise specified are measured with respect to ground.

Note 2: Limits are guaranteed by 100% testing, sampling, or correlation with worst-case test conditions.

Note 3: Maximum junction temperature (T_j) should not exceed 135°C.

TIMING DIAGRAM



Write Mode Timing Diagram

FUNCTIONAL DESCRIPTION

CIRCUIT OPERATION

For any selected head, the ML4415/4416 functions as a read amplifier when in the Read mode, or as a write current switch when in the Write mode. Pins HS0, HS1 and HS2 determine head selection while pin R/W controls the Read/Write mode. A detected "write-unsafe" condition is indicated by pin WUS.

READ MODE

When the ML4415, 4416 is in the Read Mode, it operates as a low-noise differential amplifier on the selected channel. In Read mode the write data flip-flop is set and both the write unsafe detector and the write current source are deactivated. The center tap voltage is also lowered. Pins RDX and RDY provide differential emitter follower outputs which are in phase with the X and Y head input pins.

Note that during the Read or Chip Deselect mode the internal write current is deactivated, thus making external write current gating unnecessary.

WRITE MODE

The ML4415, 4416 operates as a write-current switch when in the Write mode. Write current magnitude is determined by the following relationship:

$$I_W = K/R_{WC}$$

Where: K = Write Current Constant

R_{WC} = Resistance connected between pin WC and GND.

The head current is toggled between the X and Y side of the selected head by a negative transition WDI (Write Data Input). When switching the ML4415, 4416 to write mode, the WDFF (Write Data Flip-Flop) is initialized to pass write current through the X-side of the head.

The ML4415, 4416 exhibit enhanced write current stability, compared to similar read/write circuits, which reduces the problem of oscillation. This is a result of increased internal write current compensation. Also, write current "glitches" during power-up, common in similar read/write circuits, are eliminated with an exclusive write current disabling function.

The WUS (Write Unsafe) pin is an open collector output that gives a logic high level for any of the following unsafe write conditions:

- Open head
- Open head center-tap
- Too low WDI frequency
- Read mode selected
- Device not selected
- No write current

Two negative transitions on WDI are required to clear WUS after the fault condition is removed.

The ML4415, 4416 also offers a voltage fault detection circuit that prevents write current during power-loss or power-up.

Table 1.

Head Select				
HS3	HS2	HS1	HS0	HEAD
0	0	0	0	H0
0	0	0	1	H1
0	0	1	0	H2
0	0	1	1	H3
0	1	0	0	H4
0	1	0	1	H5
0	1	1	0	H6
0	1	1	1	H7
1	0	0	0	H8
1	0	0	1	H9
1	0	1	0	H10
1	0	1	1	H11
1	1	0	0	H12
1	1	0	1	H13
1	1	1	0	H14*

* ML4415 only
 0 = Logic Level Low
 1 = Logic Level High
 X = Don't Care

Table 2.

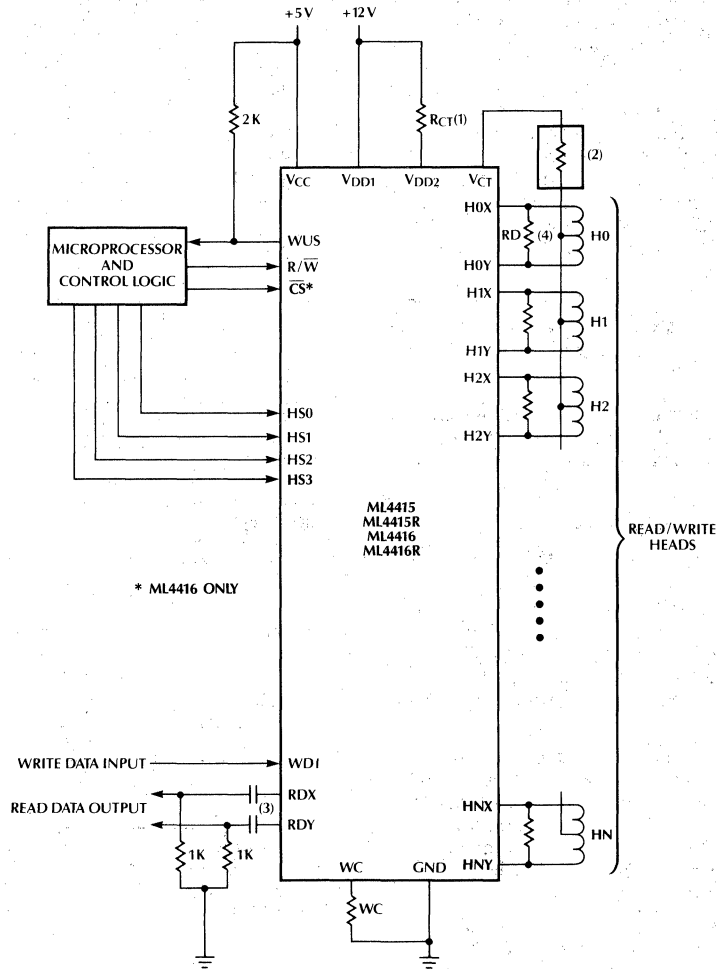
Mode Select		
CS**	R/W	MODE
0	0	Write
0	1	Read
1	X	Idle

** ML4416 only
 0 = Logic Level Low
 1 = Logic Level High
 X = Don't Care



ML4415, ML4415R, ML4416, ML4416R

TYPICAL APPLICATION



NOTES:

1. RCT is optional and is used to limit internal power dissipation (Otherwise connect V_{DD1} to V_{DD2}).
 $RCT (1/2 \text{ Watt}) = 120 (40/I_W)$ ohms
 where I_W = Write Current, in mA
2. Ferrite head optional: used to suppress write current overshoot and ringing. Recommend Ferroxcube 3659065/4A6.
3. RDX and RDY load capacitance 20pF maximum. RDX and RDY output current must be limited to 100 μ A.
4. Damping resistors not required on ML4415R, 4416R.

ML4415, ML4415R, ML4416, ML4416R

ORDERING INFORMATION

PART NUMBER	PACKAGE	NUMBER OF CHANNELS
ML4415CQ ML4415RCQ	MOLDED PCC (Q44) MOLDED PCC (Q44)	15 15
ML4416CQ ML4416RCQ	MOLDED PCC (Q44) MOLDED PCC (Q44)	14 with \overline{CS} 14 with \overline{CS}

Zoned Bit Recording Circuit

GENERAL DESCRIPTION

The ML4417/27 is a bipolar monolithic integrated circuit that simplifies the design of zoned bit recording systems in hard disk drives. It contains a VCO capable of operating at frequencies up to 95 MHz, a charge pump, and the active electronics required for a loop filter to form a variable rate data encoding and decoding system.

The ML4417/27 also includes a code clock output and the dividers required for an interface clock output whose frequency is equal to the code clock output frequency divided by 1.5. This feature simplifies the use of RLL (1, 7) coding for improved storage density.

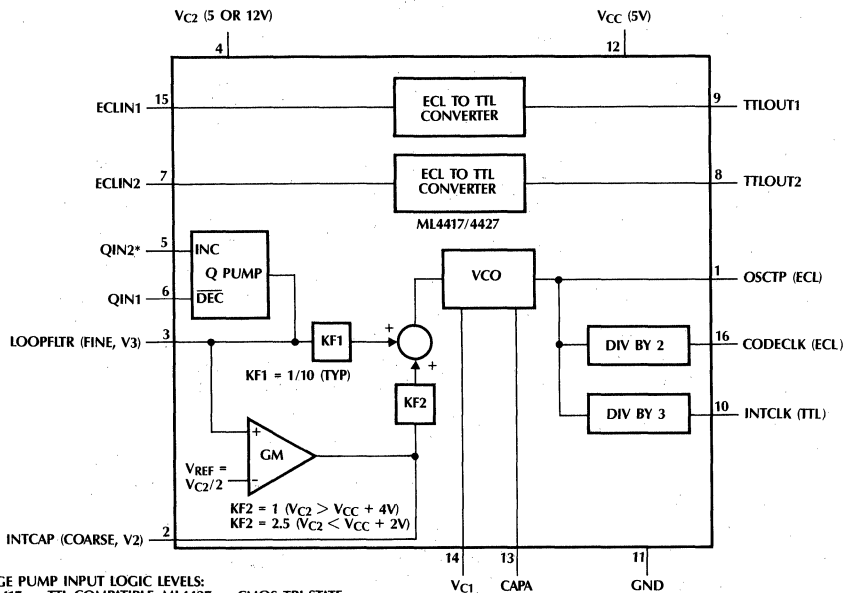
In addition, the ML4417/27 includes two uncommitted ECL to TTL level translators to simplify interfacing with TTL-based systems. The ML4417/27 is designed for operation from 12V and 5V supplies, but may be operated from a single 5V supply if desired.

The ML4417 has TTL-compatible logic input levels on the charge pump, and the ML4427 has a charge pump control input, which, when driven by a CMOS tri-state output, eliminates one logic interface line to the circuit.

FEATURES

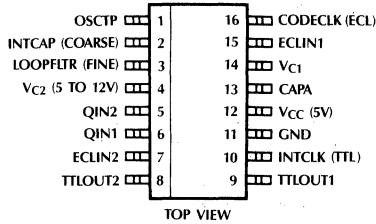
- Wide VCO Range (3:1 Range to 95 MHz)
- Allows RLL (1, 7) or (2, 7) Encoding
- SO-16 (Narrow) Packaging
- Coarse and Fine VCO Control Inputs
- Two Uncommitted ECL to TTL Converters
- 12V, 5V or Single 5V Operation

BLOCK DIAGRAM

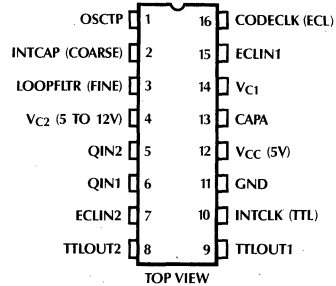


PIN CONNECTIONS

ML4417, ML4427
SOIC-16 (Narrow) Package



ML4417, ML4427
PDIP-16 Package
(Prototypes Only)



PIN DESCRIPTION

PIN NO.	NAME	FUNCTION	PIN NO.	NAME	FUNCTION
1	OSCTP (ECL)	Oscillator Test Point. An ECL output of the VCO that is useful for direct evaluation of the VCO output.	15, 7	ECLIN1, 2	ECL inputs for ECL to TTL level translators.
2	INTCAP (COARSE)	The coarse input for the loop filter time constant setting.	9, 8	TTLOUT1, 2	TTL outputs for ECL to TTL level translators.
3	LOOPFLTR (FINE)	The fine input for loop filter time constant setting.	10	INTCLK (TTL)	Interface clock output. This output is a TTL output at one third of the VCO frequency.
4	VC2	Analog power supply input, nominal 5V or 12V.	11	GND	Ground.
5	QIN2	Increment input on the charge pump. This input is TTL-compatible on the ML4417. On the ML4427, it can be connected, along with pin 6, to a single CMOS tri-state output, eliminating one pin on the controlling gate array. (Active high)	12	VCC (5V)	Logic power supply input, nominally 5V.
6	QIN1	Decrement input on the charge pump. (Active low)	13	CAPA	VCO capacitor connection. This capacitor determines the nominal VCO frequency.
			14	VC1	VC1 should be connected to a well-regulated 5V ± 5% supply.
			16	CODECLK (ECL)	The code clock output. This is an ECL output at half the VCO frequency.

ML4417, ML4427

ABSOLUTE MAXIMUM RATINGS

Power Supply Voltage Range	
V _{C1}	-0.3 to V _{CC} + .3 VDC
V _{C2}	-0.3 to 14 VDC
V _{CC}	-0.3 to 6 VDC
Digital Inputs	
ECLIN1, 2	-0.3 to V _{CC} + 0.3V
QIN1, 2	-0.3 to V _{CC} + 0.3V
Analog Inputs	
LOOPFLTR, INTCAP	-0.3 to V _{C2} + 0.3V
CAPA	-0.3 to V _{C1} + 0.3V
Digital Outputs	
TTOUT1, 2, OSCPT, CODECLK, INTCLK	-0.3 to V _{CC} + 0.3V

TYPICAL OPERATING CONDITIONS

Temperature Range	0°C to +70°C
Analog Supply Voltage (V _{C2})*	5 or 12V
Digital Supply Voltage (V _{CC})	5V
V _{C1}	5V

* This supply voltage is designed for 5V or 12V operation. This data sheet specifies the ML4417/4427 for 12V operation. For 5V specification, please contact Micro Linear.

Absolute maximum ratings are those values beyond which the device could be permanently damaged. Absolute maximum ratings are stress ratings only and functional device operation is not implied. (All voltages are referenced to GND.)

ELECTRICAL CHARACTERISTICS

Unless otherwise specified, V_{C2} = 12V ± 10%, V_{CC} = 5V ± 5%, V_{C1} = 5V ± 5%, T_A = 25°C.

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS	
Power Supply							
I _{CC}	V _{CC} Supply Current ⁽¹⁾	Pin 12		90.0		mA	
I _{C1}	V _{C1} Supply Current	Pin 14		11.0		mA	
I _{C2}	V _{C2} Supply Current	Pin 4		4.5		mA	
Digital Inputs							
V _{IH} (ECL)	High Voltage ECL Input	Pin 15, V _{CC} = 5V	4.0			V	
V _{IL} (ECL)	Low Voltage ECL Input	Pin 15, V _{CC} = 5V			3.6	V	
I _{IH} (ECL)	High Current ECL Input	Pin 15, V _{CC} = 5V			1250	μA	
I _{IL} (ECL)	Low Current ECL Input	Pin 15, V _{CC} = 5V	625		1000	μA	
Digital Outputs (ECL are Open Emitter)							
V _{OH} (TTL)	High Voltage TTL Output	I _{OH} = -0.4mA	TTL OUT _{1, 2, 3} Pins 8, 9, 10, V _{CC} = 5V	3.75		V	
V _{OL} (TTL)	Low Voltage TTL Output	I _{OL} = 1.6mA			0.50	V	
V _{OH} (ECL)	High Voltage ECL Output	I _{OH} = -4mA	ECL Code CLK Pin 16, V _{CC} = 5V	4.05	4.22	4.30	V
V _{OL} (ECL)	Low Voltage ECL Output	I _{OL} = -4mA		2.80	3.22	3.55	V
Voltage Controlled Oscillator (VCO) (Transfer Function Pin 2 to Pin 1 = 7.5MHz/Volt @ 10pF)							
f _{VCO}	VCO Range	C _{OSC} = 10pF Pin 14 to Pin 13 (Pin 2 = 1V to 11V, Pin 3 = 6V) (Pin 14 = V _{CC})		20-95		MHz	
Charge Pump							
I _Q	Charge Pump Current	Pin 3		±125		μA	
V _{QH}	Charge Pump Maximum Voltage	Pin 3		V _{C2} - 1V		V	
V _{QL}	Charge Pump Minimum Voltage	Pin 3		1.0		V	
INC, DEC Inputs							
V _{IH}	High Voltage Input	Pin 6, V _{CC} = 5V	1.9		V _{CC}	V	
V _{IL}	Low Voltage Input	Pin 6	0		0.8	V	
V _{IH}	High Voltage Input	Pin 5 (ML4417), V _{CC} = 5V	1.9		V _{CC}	V	
V _{IL}	Low Voltage Input	Pin 5 (ML4417)	0		0.8	V	

Note 1: This value includes current consumed in 1KΩ terminating resistors from pins 1 and 16 to ground.

ELECTRICAL CHARACTERISTICS (Continued)

Unless otherwise specified, $V_{C2} = 12V \pm 10\%$, $V_{CC} = 5V \pm 5\%$, $V_{C1} = 5V \pm 5\%$, $T_A = 25^\circ C$.

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
INC, DEC Inputs (Continued)						
V_{IH}	High Voltage Input	Pin 5 (ML4427) see figure 2, $V_{CC} = 5V$	4.2		5.0	V
V_{IL}	Low Voltage Input	Pin 5 (ML4427) see figure 2, $V_{CC} = 5V$	0		3.1	V
I_{IH}	High Current Input	Pin 6, $V_{IN} = 1.9V$	-5.0		+1.0	μA
I_{IL}	Low Current Input	Pin 6, $V_{IN} = 0V$	-25		-1.9	μA
I_{IH}	High Current Input	Pin 5 (ML4417), $V_{IN} = 5V$	+30		+200	μA
I_{IL}	Low Current Input	Pin 5 (ML4417), $V_{IN} = 0 \rightarrow 0.9V$	-25		+40	μA
I_{IH}	High Current Input	Pin 5 (ML4427), $V_{IN} = 5V$	+1.0		+20	μA
I_{IL}	Low Current Input	Pin 5 (ML4427), $V_{IN} = 3.1V$	-0.1		+70	μA
ECL Input 2 (Pin 7) at $25^\circ C$, $5MHz < f_{IN} < 35MHz$, $40\% < \text{Duty Cycle} < 60\%$ (If Unused, Pin 7 = V_{CC})						
V_{IH}	High Voltage Input	$V_{CC} = 5V$	3.0	4.2	5.1	V
V_{IL}	Low Voltage Input	$V_{CC} = 5V$	2.5	3.4	4.6	V
V_A	Voltage Swing	$V_{IH} - V_{IL}$, $V_{CC} = 5V$.5		2.0	V
I_{IN}	Input Current	DC Bias Value		35		μA
Transconductance Amplifier						
V_{REF1}	Inverting Input of Amplifier			$V_{C2}/2$		V
G_M	Transconductance	ΔI (Pin 2) \div ΔV (Pin 3)		275		μmho
I_{SAT}	Limiting Value of Output Current	Pin 2		± 120		μA

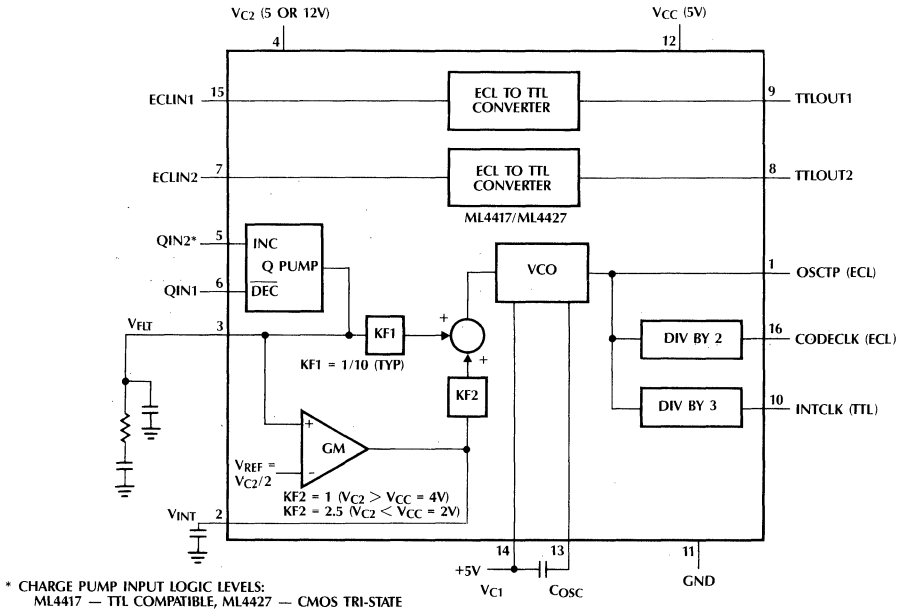


Figure 1. Typical Passive Component Connections

ML4417, ML4427

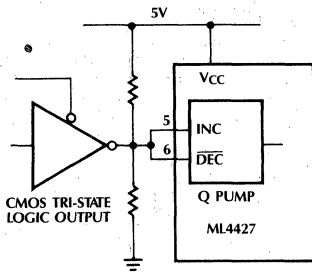


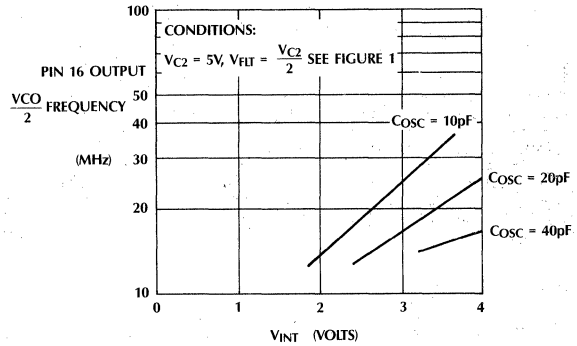
Figure 2.

The ML4427 version has an input on pin 5 that allows a single-line control interface on the charge pump. By connecting pins 5 and 6 together, the charge pump can be controlled from a single CMOS tri-state output as follows: HI = increment, LO = decrement, tri-state = coast. The benefit is a savings of one output pin on a control gate array. A resistive termination to $V_{CC}/2$ is required to establish the logic level during tri-state, as shown.

TYPICAL PERFORMANCE CHARACTERISTICS

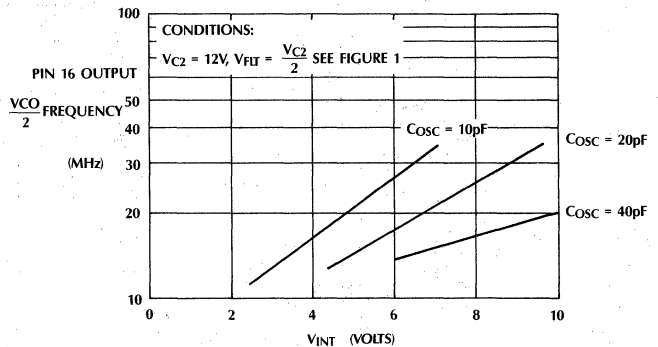
$$\frac{VCO}{2} \text{ (MHz) vs. } V_{INT}$$

$V_{C2} = 5V$
(5V-Only Operation)



$$\frac{VCO}{2} \text{ (MHz) vs. } V_{INT}$$

$V_{C2} = 12V$



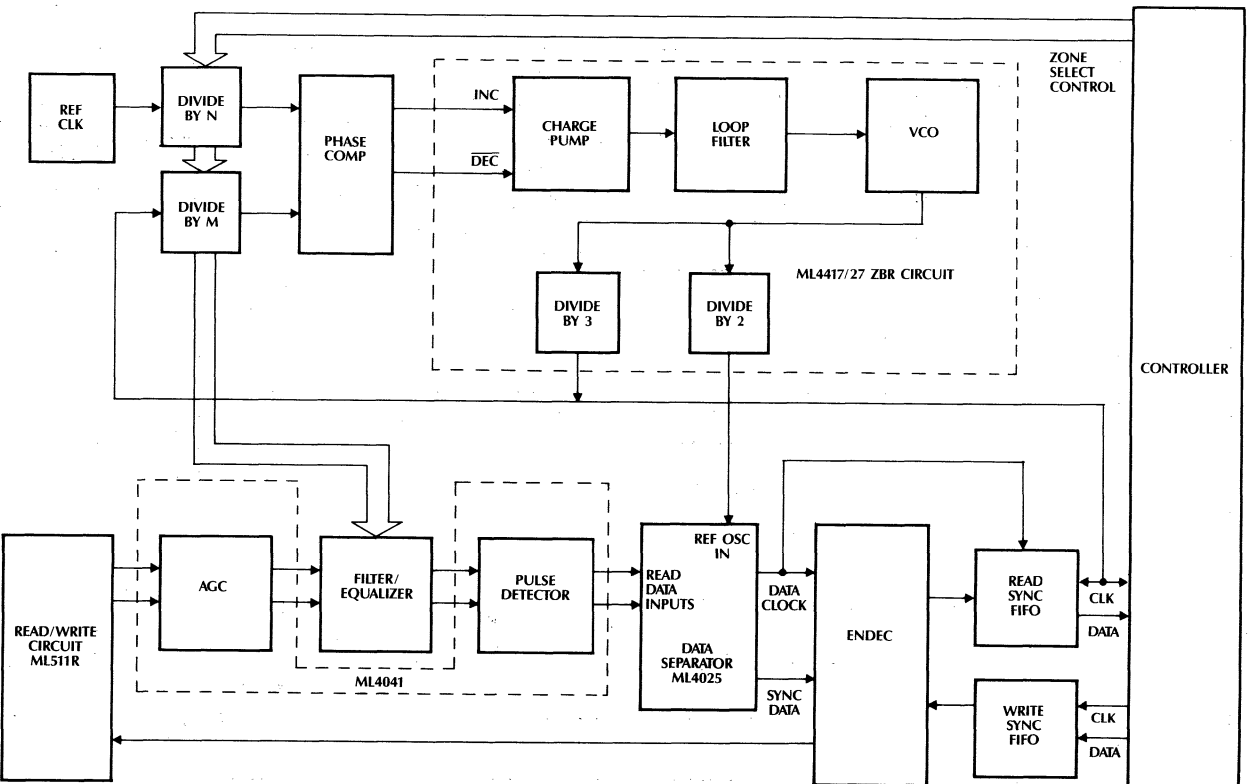
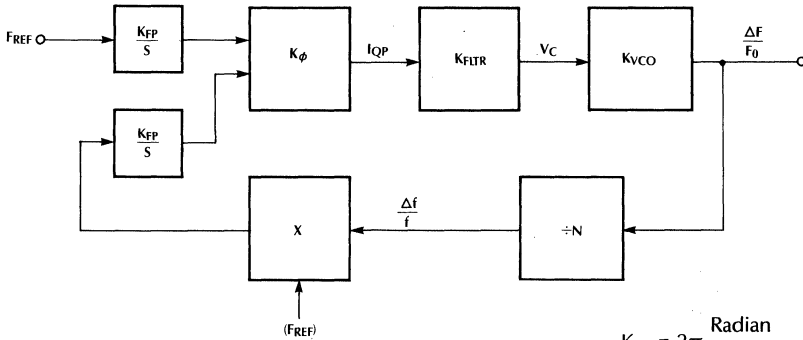


Figure 3. Read Channel Using ML4417/27 ZBR Circuit with RLL (1, 7) Encoding

4417/4427 LOOP RESPONSE

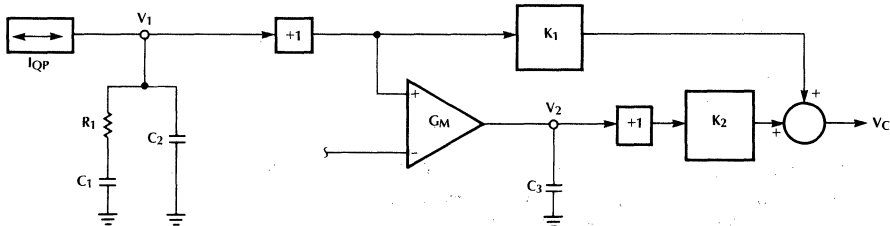


$$K_{FP} = 2\pi \frac{\text{Radian}}{\text{Hertz}}$$

$$K_{\phi} = \frac{.125\text{mA}}{2\pi} \frac{\text{Amp}}{\text{Radian}}$$

$$K_{VCO} = \frac{100\%}{2.5\text{V}} \frac{\%}{\text{Volt}} \quad (C_{OSC} = 10\text{pF})$$

K_{FLTR} — To Be Derived, Units = $\frac{\text{Volts}}{\text{Amp}}$



$$V_1 = I_{QP} * \frac{(R_1 + \frac{1}{sC_1}) \frac{1}{sC_2}}{R_1 + \frac{1}{sC_1} + \frac{1}{sC_2}} = I_{QP} * \frac{SR_1C_1 + 1}{S(C_1 + C_2) (SR_1 \frac{C_1C_2}{C_1 + C_2} + 1)}$$

$$V_2 = V_1 * \frac{G_M}{sC_3}, \quad V_C = V_1K_1 + V_2K_2 = V_1 \left(K_1 + \frac{G_M K_2}{sC_3} \right) = V_1 \frac{S \frac{K_1}{K_2 G_M} C_3 + 1}{S \frac{1}{G_M K_2} C_3}$$

$$\therefore \frac{V_C}{I_{QP}} = \frac{G_M K_2 (SR_1C_1 + 1) \left(S \frac{K_1}{K_2 G_M} C_3 + 1 \right)}{S^2 (C_1 + C_2) C_3 \left(SR_1 \frac{C_1C_2}{C_1 + C_2} + 1 \right)}$$

$$= \frac{(SR_1C_1 + 1) (S * 3.6 * 10^3 C_3 + 1)}{36 * 10^3 S^2 (C_1 + C_2) C_3 \left(SR_1 \frac{C_1C_2}{C_1 + C_2} + 1 \right)} = K_{FLTR}$$

$$K_1 = .1 \frac{\text{V}}{\text{V}}$$

$$K_2 = 1 \frac{\text{V}}{\text{V}}$$

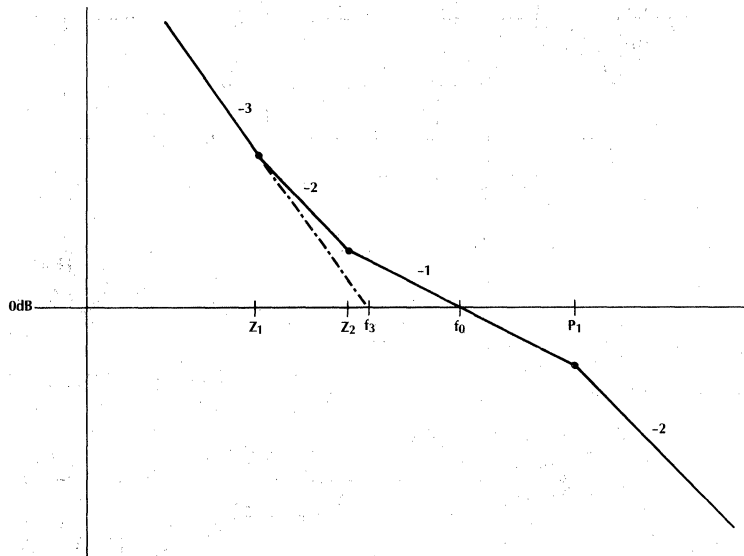
$$G_M = 27.5 * 10^{-6} \frac{\text{Amp}}{\text{Volt}}$$

ML4417, ML4427

Thus complete open loop transfer function T_{OL} :

$$T_{OL} = \frac{K_{FP}}{S} * K_{\phi} * K_{FLTR} * K_{VCO} * F_{REF} = \frac{2\pi}{S} * \frac{.125 * 10^{-3}}{2\pi} * K_{FLTR} * \frac{1}{2.5} * F_{REF}$$

$$= \frac{K_{FLTR}}{S} * \frac{F_{REF}}{20 * 10^3} = \frac{(SR_1 C_1 + 1) (S * 3.6 * 10^3 C_3 + 1) * F_{REF}}{S^3 (C_1 + C_2) C_3 \left(SR_1 \frac{C_1 C_2}{C_1 + C_2} + 1 \right) * 720 * 10^6}$$



Must define desired F_{REF} , f_0 , Z_2 , P_1 , Z_1 ; then can proceed with component value determination.

If $F_{OUT} = 36 * 10^6$, $N = 50$ (typical numbers), then $F_{REF} = 720 * 10^3$
 Assume: $f_0 = 1000\text{Hz}$, $Z_2 = 250\text{Hz}$, $P_1 = 3000\text{Hz}$, $Z_1 = 45\text{Hz}$

1. Set Z_1 with C_3 :

$$3.6 * 10^3 * C_3 = \frac{1}{2\pi Z_1} \rightarrow C_3 = \frac{1}{2\pi * 45 * 3.6 * 10^3} = .982 * 10^{-6} \sim 1\mu\text{F}$$

2. Set -3 intercept frequency f_3 with $(C_1 + C_2)$: $f_3 = (Z_1 Z_2 f_0)^{1/3} = (11.25 * 10^6)^{1/3}$

$$\rightarrow C_1 + C_2 = \frac{720 * 10^3}{(2\pi)^3 * 11.25 * 10^6 * 10^{-6} * 720 * 10^6} = .358 * 10^{-6}$$

3. Ratio $\frac{Z_2}{P_1} = \frac{R_1 \left(\frac{C_1 C_2}{C_1 + C_2} \right)}{R_1 C_1} = \frac{C_2}{C_1 + C_2} \rightarrow C_2 = (C_1 + C_2) \frac{Z_2}{P_1} = .358 * 10^{-6} * \frac{250}{3000} = .0298 * 10^{-6} \sim .030\mu\text{F}$

4. $C_1 = (C_1 + C_2) - C_2 = .358 * 10^{-6} - .030 * 10^{-6} = .328 * 10^{-6} \sim .33\mu\text{F}$

5. Set Z_2 with R_1 : $R_1 C_1 = \frac{1}{2\pi Z_2} \rightarrow R_1 = \frac{1}{2\pi * 250 * .33 * 10^{-6}} = 1.929 * 10^3 \sim 1.91\text{K}$

ORDERING INFORMATION

PART NUMBER	TEMP. RANGE	PACKAGE
ML4417CP ML4417CS	0°C to +70°C 0°C to +70°C	MOLDED DIP (P16) MOLDED SOIC (S16N)
ML4427CP ML4427CS	0°C to +70°C 0°C to +70°C	MOLDED DIP (P16) MOLDED SOIC (S16N)

Low Saturation Voice Coil Servo Driver

GENERAL DESCRIPTION

The ML4418 is a voice coil power driver intended for use in High Performance 12V Hard Disk servo systems. The ML4418 contains all control circuitry necessary to drive the voice coils of most drives. To maximize compliance voltage, the ML4418 includes two 1-Amp NPN drivers and provides base drive for external PNP transistors. In addition, power fail detection and a low voltage head retraction functions are provided for orderly shut-down of the drive. A current sense amplifier is included to enable voice coil current feedback for velocity calculations. Special care has also been taken to maximize system loop bandwidth.

The transconductance programmed by a logic input at 1/2 A/V and 1/7 A/V respectively, when using a 1Ω sense resistor. This allows for greater DAC resolution in digitally controlled servos during track follow without compromising dynamic range during seek.

The retraction circuit, main drive circuit, and control circuits are each powered from their own supplies. Retract is self-contained for 12V systems but allows the use of an external PNP to allow retraction with as little as 1V of back EMF from the spindle.

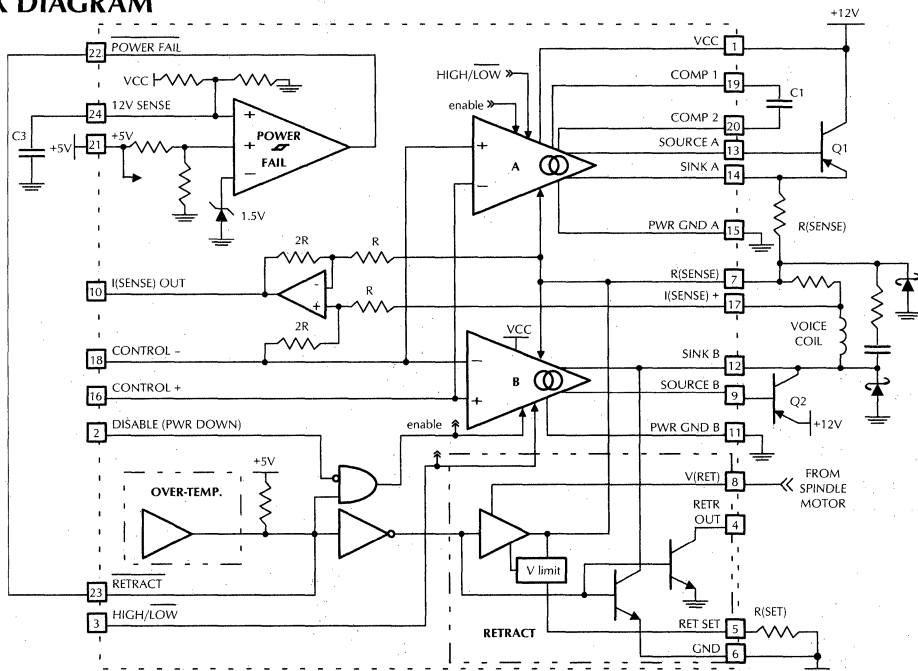
The power fail detection circuit includes a precision 1.5V bandgap reference and a power fail comparator.

The ML4418 is implemented using Micro Linear's bipolar array technology. This allows for customization of the IC for a user's specific application.

FEATURES

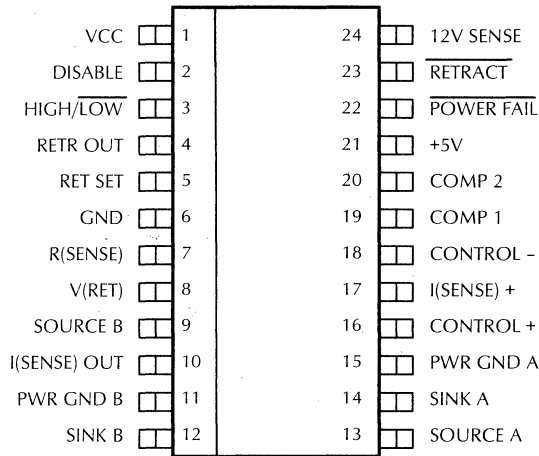
- Low saturation voltage (<1V at 1A.)
- No cross-over distortion with low quiescent current
- VCM coil current output referenced to V_{REF}
- Pin-programmable transconductance settings
- Retraction circuitry with programmable retract voltage and separate power pin operates to 1V
- On-chip precision power fail detect circuitry
- Over-temperature protection with flag output
- Operates from +12 supplies

BLOCK DIAGRAM



PIN CONFIGURATION

ML4418
24-Pin SOIC



TOP VIEW

PIN DESCRIPTION

PIN#	NAME	FUNCTION	PIN#	NAME	FUNCTION
1	VCC	Supply input to power amplifiers	14	SINK A	Current sinking output for non-inverting power amplifier. Connects to voice coil (+) terminal.
2	DISABLE	A Logic "1" puts the IC into a low power state and disables the power amplifiers.	15	PWR GND A	Power return pin for non-inverting power amplifier A.
3	HIGH/LOW	A logic "1" sets the transconductance gain to 1/2 while a logic "0" sets the gain to 1/7. Transconductance gain is the $V_{R(SENSE)} \div V_{CONTROL}$.	16	CONTROL +	Positive input for current command.
4	RETR OUT	Open collector output which pulls low during retract. Used to provide a braking signal to spindle.	17	I(SENSE) +	Positive input for current sense amplifier.
5	RET SET	External set resistor to establish a voltage limit for the internal retract driver.	18	CONTROL -	Negative input for current command.
6	GND	Analog signal ground.	19	COMP 1	Pin for external compensation capacitor.
7	R(SENSE)	Current sense resistor terminal.	20	COMP 2	Pin for external compensation capacitor.
8	V(RET)	Supply pin for retract circuits.	21	+5V	Input for +5V for power fail detection and logic power supply.
9	SOURCE B	PNP Base drive output for inverting power amplifier.	22	POWER FAIL	Open Collector output drives low for low voltage conditions.
10	I(SENSE) OUT	Output of the Current Sense amplifier	23	RETRACT	A logic "0" initiates retract. Also used as an open-collector over-temperature output flag.
11	PWR GND B	Power return pin for inverting power amplifier B.	24	12V SENSE	Input to the power fail comparator from a resistor divider from VCC.
12	SINK B	Current sinking output for inverting power amplifier. Connects to voice coil (-) terminal.			
13	SOURCE A	PNP Base drive output for non-inverting power amplifier.			

ABSOLUTE MAXIMUM RATINGS

Supply Voltage (pins 1,8)	14V
Voltage pins 2, 3, 23, 22	-0.3V to +7V
pins 4,7,8,9,12,13,14,17	-0.3V to VCC
Output Sink Current	± 1A
Retraction Current	80mA
Retract set current (pin 5)	3 mA
Junction temperature	150°C
Storage Temperature Range	-65°C to 150°C
Lead Temperature (Soldering 10 sec.)	260°C
Thermal Resistance (θ_{JA})	60°C/W

OPERATING CONDITIONS

Temperature Range	0°C to 70°C
VCC Supply Voltage	10.8V to 13.2V
+5V (pin 21) Supply Voltage	4.5V to 5.5V
V(RET) (pin 8) Supply Voltage	2.5V to 13.2V
CONTROL + Voltage Range (pins 16)	0V to VCC
CONTROL - Voltage Range (pins 18)	2.4V to 6.0V

Absolute maximum ratings are those values beyond which the device could be permanently damaged. Absolute maximum ratings are stress ratings only and functional device operation is not implied.

ELECTRICAL CHARACTERISTICS

Unless otherwise specified, T_A =Operating Temperature Range, VCC=operating range, $R_{SENSE}=1\Omega$, $R_{COIL}=15\Omega$
 CONTROL - (pin 18) = VCC/2, R_{SET} (pin 5) = 7.5K Ω , $C_1=30pF$, $Q_1=Q_2= MJE210$.

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
Amplifier					
Offset				±12	mA
Common Mode Transconductance	$2V \leq V_{PIN18} \leq 6V$		0.5	1	mA/V
Gain	pin 3 = 2V	475	500	525	mA/V
	pin 3 = 0.8V	136	143	150	mA/V
Maximum Bandwidth			100		Khz
Sinking saturation	$I_{OUT} = 100mA$		0.3	0.6	V
	$I_{OUT} = 300mA$		0.4	0.8	V
	$I_{OUT} = 500mA$		0.5	1	V
Sourcing saturation	$I_{OUT} = 100mA$		0.1		V
	$I_{OUT} = 300mA$		0.2		V
	$I_{OUT} = 500mA$		0.3		V
Source A/B Base Drive		20	30	50	mA
Q1/Q2 Standby Current	$\beta_{PNP} = 200, V_{PIN16} = 5V$	1	4	7	mA
Retraction Circuit					
Turn on time			800		nS
Turn off time			8		μS
Source Voltage	$V_{PIN23}=0.8V, V_{PIN8}= 3V, I_{PIN7}= -50mA$	0.53	0.75	0.97	V
Sink Current	$V_{PIN23}=0.8V, V_{PIN8}= 1.2V, V_{PIN12}= 0.6V$	36	48	150	mA
RETR OUT V_{OL}	$V_{PIN23}=0.8V, I_{PIN4}=1mA$			0.4	V
Power Fail Detection Circuit					
12V Threshold		9.5	10	10.5	V
Hysteresis - 12V Sense		5	120	190	mV
5V Threshold		4.35	4.525	4.70	V
Hysteresis - 5V Sense		5	30	80	mV
Logic Inputs and Outputs					
Voltage High (V_{IH})		2	1.4		V
Voltage Low (V_{IL})			1.4	0.8	V
Current High (I_{IH})	$V_{IN}=5V$			±10	μA
Current Low (I_{IL})	$V_{IN}=0V$, except pin 23	-40	-10		μA
	$V_{IN}=0V$, pin 23 only	-250	-160		μA
Voltage Low (pins 22, 4)	$I_{OL}=1mA$			0.4	V

ELECTRICAL CHARACTERISTICS (continued)

Unless otherwise specified, T_A =Operating Temperature Range, V_{CC} =operating range, $R_{SENSE}=1\Omega$, $R_{COIL}=15\Omega$
 CONTROL - (pin 18) = $V_{CC}/2$, R_{SET} (pin 5) = $7.5K\Omega$, $C1=30pF$, $Q1=Q2= MJE210$.

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
Over-Temperature Detection					
T_j Threshold		150	160		$^{\circ}C$
Hysteresis			30		$^{\circ}C$
Current Sense Amplifier					
Voltage Offset				± 50	mV
Differential Mode Gain		1.95	2	2.05	V/V
Common Mode Gain		-44			dB
Current Consumption					
Pin 21	Pin 21 = 5.5V		5	8	mA
Pin 1	$V_{CC}=13.2V$, $V_{PIN16} = V_{CC}/2$		8	13	mA
Pin 8	$V_{PIN8} = 13.2V$, $V_{PIN23} = 5V$		3.5	5	mA

FUNCTIONAL DESCRIPTION

POWER AMPLIFIER

The ML4418 power amplifier circuit is set up as a Howland Current source with a fixed gain of 1/2 or 1/7 (set by driving pin 3 high or low respectively). This architecture yields minimal cross-over distortion while maintaining low output cross conduction currents.

The gain figure refers to the ratio of input voltage to the output voltage seen across R_{SENSE} . For example, at a 1/2 gain setting, with V(-) input at 2.5V and the V(+) input at 3V, +500mA would flow through the coil using a 0.5Ω sense resistor. Under the same conditions with pin 3 low, the current would be 143mA. The ability to change from low to high gain allows more complete utilization of DAC resolution when in the track follow mode.

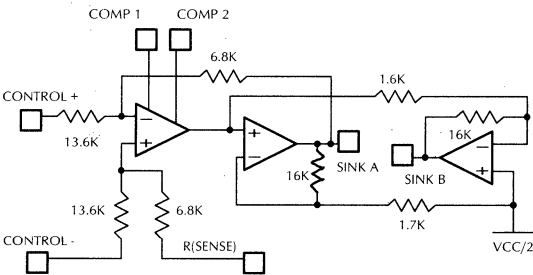


Figure 1. Power Amplifier Topology

The output stage is designed to provide minimal saturation losses and employs an external PNP transistor for the sourcing drive and an internal saturable NPN to sink current. Sinking saturation drop is typically under 0.4V. Sourcing saturation drop depends on the external transistors used. To avoid oscillation in the output stage, PNP transistors with $F_T \geq 50MHz$. should be used.

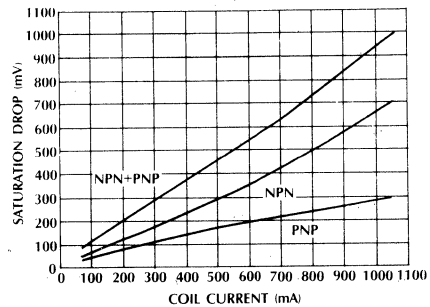


Figure 5. Output Saturation Voltage vs. Output Current (Q1 = Q2 = MJE210)

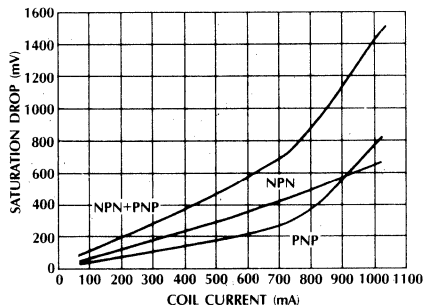


Figure 6. Output Saturation Voltage vs. Output Current (Q1 = Q2 = BSR31)

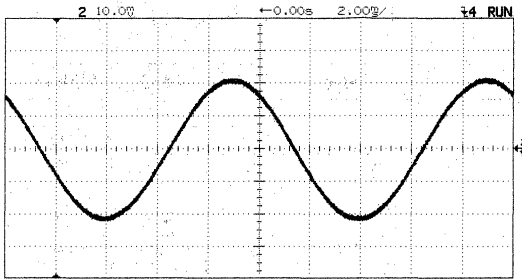


Figure 4. Output Current : $V_{IN} = 100$ Hz Sine Wave, $2.4V_{P-P}$, Low Gain Mode ($V_{PIN3}=0$), $R(SENSE) = 1\Omega$

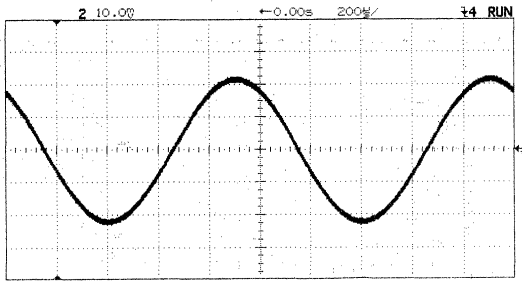


Figure 5. Output Current: $V_{IN} = 1$ KHz Sine Wave, $2.4V_{P-P}$, Low Gain Mode ($V_{PIN3} = 0$), $R(SENSE) = 1\Omega$

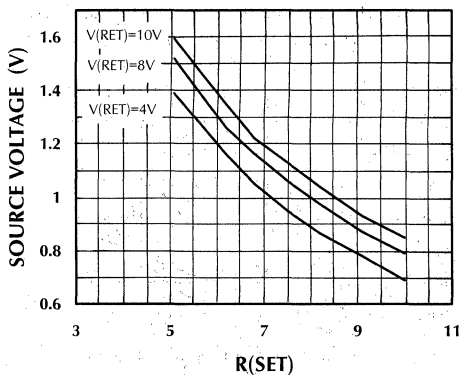


Figure 6. Retract Source Voltage at Pin 7 vs. $R(SET)$ $I_{VCM}=50mA$

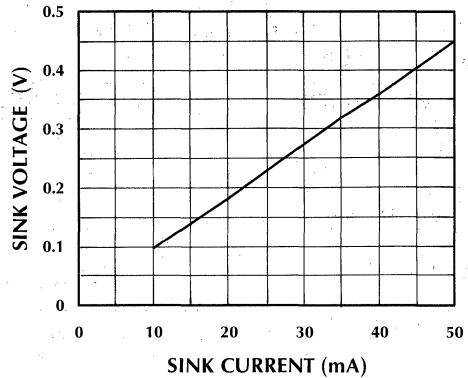


Figure 7. Retract Sink Voltage at Pin 12 vs. $I_{VCM}=50mA$

POWER FAIL DETECT CIRCUIT

The ML4418 circuit consists of a precision trimmed reference, resistor dividers and an "or function" comparator with hysteresis. The output (open collector) of this circuit appears on pin 22. When either comparator input falls below the 1.5V reference, pin 22 pulls low.

RETRACT CIRCUITS

The ML4418 retract circuit provides for spindle EMF energized power fail retraction of the VCM. When pin 23 goes low, pin 4 will pull low, providing a signal which can be delayed for spindle braking. The internal NPN transistor will saturate, pulling SINK B (pin 12) low. This portion of the circuit will function with less than 1V on V(RET). An internal voltage limited pull-up transistor is provided which sources current on pin 7 to the VCM. This circuit will operate reliably down to a V(RET) voltage of around 2.5V, making the ML4418 retract circuit adequate for 12V systems where the spindle motor EMF provided is adequate.

Figure 6 shows the saturation characteristics of the SINK B output ($R_{SAT} \approx 9\Omega$). The R_{SAT} of the pull down transistor does not vary appreciably with V(RET) voltage. Figure 7 shows the voltage sourced at R(SENSE) during Retract vs. R(SET) at various V(RET) input voltages.

CURRENT SENSE AMPLIFIER

The current sense amplifier in the ML4418 creates a signal referenced to CONTROL - (normally the Vref pin of the system's DAC and ADC) appearing across a resistor in series with the VCM. Its output is twice the voltage that appears between pins 17 and 7.

APPLICATIONS

POWER AMPLIFIER COMPENSATION

Figure 8 below shows the equivalent AC circuit for the current amplifier.

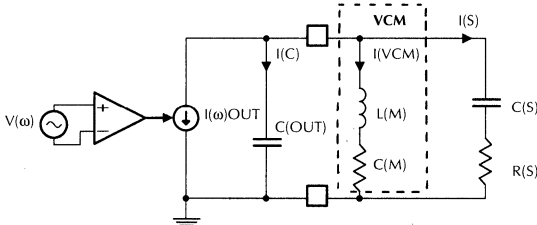


Figure 8. AC Equivalent Circuit for Current Amplifier, Voice Coil Motor (VCM) and Snubber

The amplifier's current bandwidth is limited by C_{OUT} which varies with the value chosen for R_{SENSE}

$$C_{OUT} \approx \frac{1197 \times (C_{COMP} + 3.5pF)}{R_{SENSE}}$$

Where C_{COMP} is C1 between pins 19 and 20. With no snubber (R_S and C_S) the bandwidth is limited to.

$$F_{-3dB} = \frac{1}{2\pi} \sqrt{\frac{2.414}{L(M) C(OUT)}}$$

Since this is a second order system with $L(M)$ and $C(OUT)$ forming a resonant circuit, some damping is desirable to reduce ringing in the step response. This is accomplished with a resistive snubber. The optimum value of $R(S)$ occurs when the following condition is met:

$$R(S) = \sqrt{\frac{L(VCM)}{C(OUT)}}$$

For a given $C(S)$, setting $R(S)$ to this value will minimize the ringing in the transient response. Larger values of $R(S)$ will result in more ringing and more bandwidth. Smaller values of $R(S)$ will result in more ringing and less bandwidth.

$C(S)$ (snubber capacitor) values of between 200nF and 1 μ F are usually necessary to achieve the desired reduction of ringing in the step response. At the optimum value of $R(S)$ larger values of $C(S)$ further reduce the ringing but do not affect the bandwidth.

Tuning the current loop response can be easily done simulating the network in figure 8 with a computer simulator (such as SPICE).

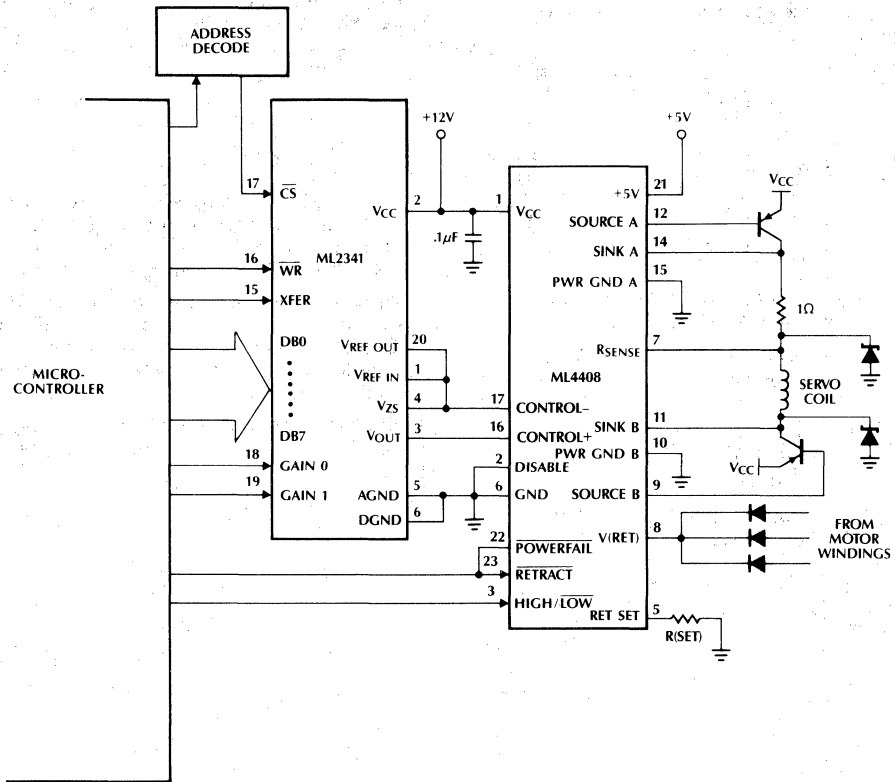


Figure 9. ML4418 Used with ML2341 8-Bit Gain Ranging DAC Provides up to 13-Bit Effective Resolution

ORDERING INFORMATION

PART NUMBER	TEMPERATURE RANGE	PACKAGE
ML4418CS	0°C to +70°C	S20W

Servo Demodulator

GENERAL DESCRIPTION

The ML4431 provides all of the analog circuitry necessary for the demodulation of di-bit servo signal information in Winchester disk drives. It interfaces to the servo head preamp and provides quadrature position signal outputs for the servo controller circuitry.

The ML4431 includes a high-performance 592-type input amplifier and differential AGC circuit. External logic is designed to meet the needs of the particular servo system utilizing the VCO and Charge Pump to create a PLL time base for Peak Detector gating. The SYNC output provides servo channel timing information for the logic.

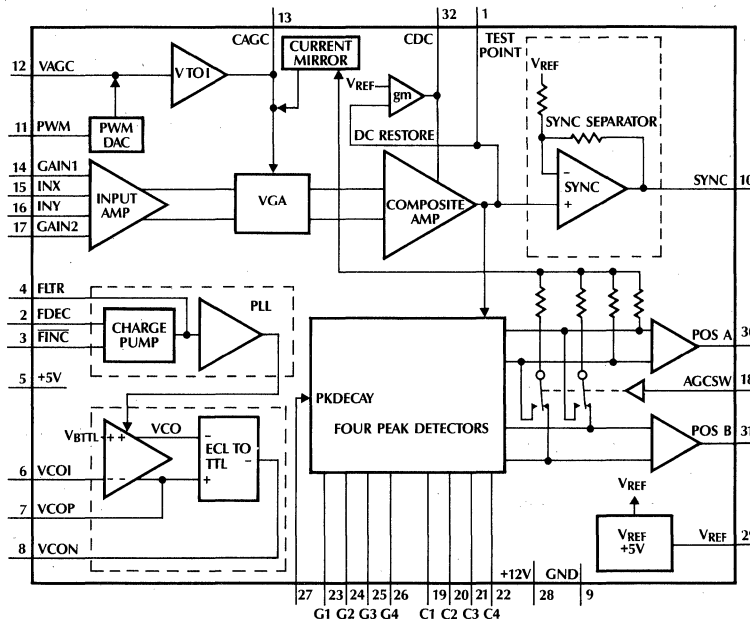
The ML4431 has an ECL-type VCO, with an internal ECL-to-TTL converter for simplified interfacing.

The ML4431, when combined with the ML4402 Servo Driver, the ML4403, ML4413 Servo Controller and the ML4404 Trajectory Generator, provides a flexible closed-loop servo control system.

FEATURES

- Combines all analog di-bit demodulation circuitry
- Logic track-type switching can be used to minimize demodulator offset
- Exponential AGC characteristics makes AGC settling independent of input step size
- External loop compensation of analog blocks
- External digital circuitry allows flexible pattern format
- On-chip band gap voltage reference eliminates external referencing
- Operates from 5V and 12V power supplies
- Programmable Peak Detector Discharge Current
- Digitally-controlled AGC set point
- TTL output VCO
- AGC Sense switchable to "POSA only" or both "POSA and POSB"
- Compatible with Micro Linear's ML4403, ML4413 Servo Controller, ML4402 Servo Driver and ML4404 Trajectory Generator

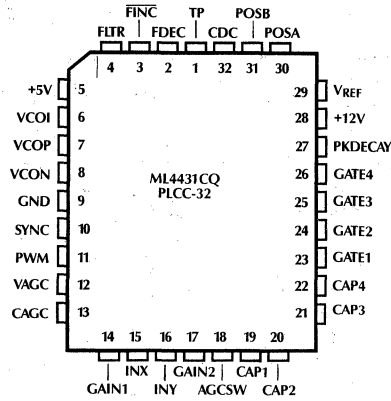
BLOCK DIAGRAM



ML4431

PIN CONNECTIONS

ML4431 32-Pin PCC



PIN DESCRIPTION

PIN NO.	NAME	FUNCTION	PIN NO.	NAME	FUNCTION
1	TP	Composite test point, normally left unconnected.	18	AGCSW	Selects between "POSA only" or "POSA and POSB" AGC sense operation. Logic "0" selects "POSA only" operation. Logic "1" selects "POSA and POSB" operation.
2	FDEC	Charge pump frequency decrement input.	19	CAP1	Peak detector 1 capacitor.
3	FINC	Charge pump frequency increment input.	20	CAP2	Peak detector 2 capacitor.
4	FLTR	PLL loop compensation terminal.	21	CAP3	Peak detector 3 capacitor.
5	+5V	+5V supply.	22	CAP4	Peak detector 4 capacitor.
6	VCOI	VCO input.	23	GATE1	Peak detector 1 gate input (TTL) Logic "1" enabled, "0" disabled.
7	VCOP	VCO positive output, for capacitive feedback to VCOI.	24	GATE2	Peak detector 2 gate input (TTL) Logic "1" enabled, "0" disabled.
8	VCON	VCO negative output, drives resistive feedback to VCOI.	25	GATE3	Peak detector 3 gate input (TTL) Logic "1" enabled, "0" disabled.
9	GND	Ground.	26	GATE4	Peak detector 4 gate input (TTL) Logic "1" enabled, "0" disabled.
10	SYNC	SYNC pulse output.	27	PKDECAV	Sets peak detector discharge current.
11	PWM	PWM DAC input to adjust AGC set point.	28	+12V	+12V supply.
12	VAGC	AGC gain reference voltage input.	29	VREF	Voltage reference output.
13	CAGC	External capacitor to set AGC response.	30	POSA	Position output A. POSA = Peak Detector 1 – Peak Detector 2
14	GAIN1	Input amplifier gain adjusting RC terminal 1	31	POSB	Position output B. POSA = Peak Detector 3 – Peak Detector 4
15	INX	X input into input amplifier.	32	CDC	External capacitor terminal to set DC restore response.
16	INY	Y input into input amplifier.			
17	GAIN2	Input amplifier gain adjusting RC terminal 2.			

ABSOLUTE MAXIMUM RATINGS

Absolute maximum ratings are those values beyond which the device could be permanently damaged. Absolute maximum ratings are stress ratings only and functional device operation is not implied.

Power Supply Voltage Range, V_{CC} 14V

Input Voltages:

GAIN1, GAIN2 -0.3 to 8V

C_{AGC} -0.3 to 70V

V_{AGC} PWM, VCOI -0.3 to 5.3V

CAP1, CAP2, CAP3, CAP4 -0.3 to 10V

GATE1, GATE2, GATE3, GATE4, VCOP -0.3 to 7.5V

INX, INY, VCON, FINC, FDEC, C_{DC} , C_{AGC} , FLTR -0.3 to $V_{CC} + 0.3V$

θ_{JA} for PLCC-32 $\approx 60^{\circ}C/Watt$

Storage Temperature Range $-65^{\circ}C$ to $+150^{\circ}C$

Junction Temperature (T_{JMAX}) $150^{\circ}C$

Lead Temperature (Soldering, 10 sec) $260^{\circ}C$

OPERATING CONDITIONS

Temperature Range $0^{\circ}C$ to $70^{\circ}C$

Supply Voltage $V_{(+12V)}$ $12V_{DC} \pm 10\%$

Supply Voltage $V_{(+5V)}$ $5V_{DC} \pm 10\%$

Input Coupling Capacitance (C_I) $0.01\mu F$

Input Amp Gain Capacitance (C_C) $0.047\mu F$

Input Amp Gain Resistance (R_C) $1k\Omega$

AGC Response Compensation Capacitance (C_A) $0.018\mu F$

Composite DC Restore Capacitance (C_D) $0.018\mu F$

PLL Compensation Components:

C_{CP1} $0.1\mu F$

C_{CP2} $1\mu F$

R_{CP} 910Ω

VCO Components:

C_V $39pF$

R_V 1500Ω

RL 680Ω

Peak Detector Capacitance (CAP1 thru CAP4) $270pF$

On track Base-to-Peak Voltage at pin TP $1.75V$

V_{GA} Gain Control Voltage (at pin C_{AGC}) $\approx 2.4V$

R_{SET} $330K\Omega$

ELECTRICAL CHARACTERISTICS

The following specifications apply over the recommended operating conditions of $T_A = 0^{\circ}C$ to $70^{\circ}C$, $V_{(+12V)} = 10.8$ to $13.2V$, $V_{(+5V)} = 4.5$ to $5.5V$, $V_{VAGC} = 4.0V$, and external components as recommended above, unless otherwise specified (See Note 1)

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
Power Supply						
I_{+12}	Supply Current	$V_{+12} = 12V, V_{+5} = 5V$		73	51	mA
I_{+5}	Supply Current	$V_{+12} = 12V, V_{+5} = 5V$		37	47	mA
TTL Inputs FINC, FDEC, GATE1, GATE2, GATE3, GATE4, PWM, AGCSW						
V_{IH}	High Level Input Voltage		2.0			V
V_{IL}	Low Level Input Voltage				0.8	V
I_{IH}	High Level Input Current	$V_{IH} = 2.4V$	-1		30	μA
I_{IL}	Low Level Input Current	$V_{IL} = 0.4V$	-20		1	μA
SYNC Output						
V_{OL}	Low Level Output Voltage	$I_{OL} = 1.6mA$	0	0.35	0.5	V
V_{THR}	Positive going input threshold			$V_{REF} + 0.9$		V
V_{THF}	Negative going input threshold			V_{REF}		V
$t_{PD\pm}$	Propagation Delay Rising, Falling	$RL = 2k, C_L = 15pF$		50		ns
VCON Output						
V_{OH}	High Level Output Voltage	$I_{OH} = 50\mu A$	2.4			V
V_{OL}	Low Level Output Voltage	$I_{OL} = 1.6mA$	0		0.5	V
VCO and Charge Pump Section						
I_{BIAS}	V_{COI} Input Bias Current		0	20	50	μA
I_{CH}, I_{DIS}	FLTR Charge and Discharge Current		330	450	590	μA
I_{CH}/I_{DIS}	FLTR Charge/Discharge Ratio		0.95	1.00	1.05	$\mu A/\mu A$
I_{OFF}	FLTR OFF State Current	$FINC = 2.0, FDEC = 0.8$	0	25	50	nA
F_{MAX}	MAX VCO Frequency to Maintain + and - 5% Control Range (Note 3)		20			MHz
V_{QH} (FLTR)	Charge Pump Maximum Voltage			$V_{(+12V)} - 12V$		V
V_{QL} (FLTR)	Charge Pump Minimum Voltage			1.0		V

5

ML4431

ELECTRICAL CHARACTERISTICS (Continued)

The following specifications apply over the recommended operating conditions of $T_A = 0^\circ\text{C}$ to 70°C , $V_{(+12V)} = 10.8$ to 13.2V , $V_{(+5V)} = 4.5$ to 5.5V , $V_{AGC} = 4.0\text{V}$, and external components as recommended above, unless otherwise specified (See Note 1)

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
VCO and Charge Pump Section (Continued)						
F_{VCO}	VCO Frequency Range (Note 3)	$T_A = 25^\circ\text{C}$, $V_{+5} = 5\text{V}$, $V_{FLTR} = 6\text{V}$, $C_V = 30\text{pF}$, $R_V = 3.74\text{k}\Omega$, see figure 1	9.7	10.0	10.3	MHz
K_{VCO}	VCO Voltage to Frequency Factor			2		%/V
Input AMP, AGC AMP, and DC Restore						
R_{IN}	INX, INY Differential Input Resistance		7	10	14	k Ω
$I_{GAIN1,2}$	GAIN1, GAIN2 Bias Current		0.66	1.0	1.20	mA
R_{INAGC}	V_{AGC} Input Resistance		7	10	13	k Ω
G_{MAGC}	AGC Transconductance at C_{AGC}			370		μMHOS
R_{AGC}	Control Range of AGC Loop to Regulate Composite Amplitude to within 2% of Nominal			7/1		V/V
BW	Bandwidth from INX, INY to Composite (Note 4)		10	15		MHz
GMDCR	DC Restore Transconductance			500		μMHOS
Peak Detectors						
I_{CH}	Charge Current		5			mA
I_{DIS}	Discharge Current	$T_A = 25^\circ\text{C}$, $R_{SET} = 330\text{K}$	10	15	20	μA
Voltage Reference						
V_{REF}	Reference Voltage	$T_A = 25^\circ\text{C}$	4.75	5.00	5.25	V
TC	Tempco			50		ppm/ $^\circ\text{C}$
R_{OUT}	Load Regulation			2		mV/mA
PSRR	Line Regulation			10		mV/V
I_{SINK}	Maximum SINK Current		0.8			mA
Output Amplifiers (POSA, POSB)						
V_{OS}	Input Offset	$V_{CAP1-4} = 6\text{V}$	-10	0	10	mV
A_V	Gain		1.15	1.20	1.25	V/V
A_{VA}/A_{VB}	Gain Tracking		-3	0	+3	%
V_{OUT}	Output Voltage Range		1.0		9.5	V
I_{SRC}	Output Source Current		3			mA
I_{SNK}	Output Sink Current		2			mA
SR	Slew Rate			2.5		V/ μs
BW	3dB Gain Bandwidth			3		MHz

Note 1: 0°C to 70°C operating temperature range devices are 100% tested with temperature limits guaranteed by 100% testing, sampling, or correlation with worst-case test conditions.

Note 2: Typicals are parametric norm at 25°C .

Note 3: This parameter is guaranteed but not 100% tested and is not used in outgoing quality level calculations.

APPLICATION HINTS

Using a nominal on-track servo signal, amplitude adjustment should be made as follows:

- Set composite signal amplitude, measured at pin TP, by adjusting voltage at pin V_{AGC} (approximately 4.7 volts). The composite signal should be set to 1.75 volts base to peak of an on-track position pulse (an off-track position pulse will be about 3.5 volts maximum).
- Adjust R_g so that the VGA is in mid-range. This is determined by measuring the voltage at pin C_{AGC} ; it should be approximately 0.9 volts. C_{AGC} voltage will vary approximately ± 0.5 volts over the AGC range.

FUNCTIONAL DESCRIPTION

INPUT AMPLIFIER

The input amplifier is equivalent to a wide-band 592 type video amplifier and provides amplification and buffering to the AGC circuitry. The Inputs INX and INY, which must be AC coupled, accept the composite analog signal from the servo head differential preamplifier. Internal input termination resistors eliminate the need for external bias resistors. Prefiltering of the signal is normally desired to eliminate unwanted components. External components R_G and C_G determine the input amplifier's low frequency cutoff and gain as follows:

$$FC = \frac{1}{2\pi(R_G + 60\Omega)C_G} \quad A_V = \frac{1700}{R_G + 60\Omega}$$

Where: C_G = External series capacitance between pins GAIN1 and GAIN2

R_G = External series resistance between pins GAIN1 and GAIN2

AUTOMATIC GAIN CONTROL (AGC)

The purpose of the AGC loop is to maintain a constant peak output voltage level at outputs POSA and POSB. This peak level is established by the reference voltage applied to pin V_{AGC} .

$$V_{P-P} \text{ (Composite Position Pulses)} = K1 \times V_{AGC} + K2$$

Where: $K1 = 0.65$

$$K2 = .13 \times V_{REF}$$

In this closed-loop system, the peak detector output voltages are fed back and combined with the V_{AGC} voltage to provide a gain control current. The current controls the variable gain amplifier (VGA) and is compensated at pin C_{AGC} to provide control of AGC bandwidth. The bandwidth of the entire AGC loop is determined by:

$$BW = \frac{K V_{V_{AGC}}}{2\pi C_A}$$

Where: $K = 2.8 \times 10^{-4}$

$V_{V_{AGC}}$ = External reference voltage at pin V_{AGC}

C_A = External capacitance at pin C_{AGC}

PWM CONTROL OF AGC SET POINT

The PWM input (pin 10) accepts a variable duty-cycle input to control the AGC set point. The relationship between duty-cycle and set point is:

100% duty-cycle AGC set point is equal to V_{REF} .

0% duty-cycle AGC set point equal to $0.6 \times V_{REF}$.

A filter capacitor from pin 11 to ground is required to filter the PWM signal. This capacitor should be sufficiently large relative to the 10K Ω nominal internal termination resistance at pin 11.

The AGC set point may be set manually via direct voltage control of pin 12 if desired. Pin 11 should be grounded in this case.

SWITCHING THE AGC SENSE RESISTORS

The AGCSW input (pin 17) allows selection of the AGC sense. The choices are:

AGCSW low AGC senses POS A peak detector outputs only.

AGCSW high AGC senses POS A and POS B peak detector outputs.

COMPOSITE AMPLIFIER

The input amplifier and AGC circuit of the ML4431 operate in a differential signal mode to provide good common mode and power supply rejection. The composite amplifier converts the differential signal into a buffered single-ended signal for the peak detector circuitry. The DC base line of the composite signal is equal to V_{REF} . The bandwidth of the DC restore function is controlled by capacitor C_D at pin C_{DC} with the following relationship:

$$BW = \frac{gm}{2\pi C_D}$$

Where: $gm = \frac{1}{2K\Omega}$

C_D = External capacitance at pin C_{DC}

The composite signal is available at pin TP and is normally left unconnected. For short circuit protection a 750 Ω resistor is connected in series with pin TP internally.

SYNCHRONIZATION PULSE SEPARATOR

The SYNC pulse separator is a threshold comparator with hysteresis which passes pulses from the composite amplifier above a set threshold. It provides a buffered TTL output. The SYNC output, when gated through an external one-shot, is used to control the external gate timing and PLL logic. Active pull-up differs from ML4401 SYNC.

PEAK DETECTOR

The peak detector circuit captures the peak signal amplitude of the di-bit pulses. The gates are controlled by inputs GATE1 through GATE4. Timing is established by the external logic circuitry. The external peak detector capacitors are connected from pins CAP1 through CAP4 to ground. The peak detector discharge rate (set by CAP1-CAP4 and current out of PKDECAY) determines the maximum track crossing rate during an access operation. The peak detector outputs are fed into internal differential amplifiers that calculate the track error signals and provide buffered outputs POSA and POSB as follows:

$$POSA = 1.20 (CAP1 - CAP2) + V_{REF}$$

$$POSB = 1.20 (CAP3 - CAP4) + V_{REF}$$

PEAK DETECTOR DECAY RATE CONTROL

The decay rate of the peak detector can be programmed by changing the external resistor R_{SET} (pin 26, see connection diagram). The decay rate is determined by the discharge current for the hold capacitors C1 – C4. The relationship between the discharge current and R_{SET} is:

$$I_{DISCHARGE} = \frac{V_{REF}}{R_{SET}}$$

VOLTAGE CONTROLLED OSCILLATOR AND CHARGE PUMP

The VCO and external phase compare logic provide a time base for peak detector gate synchronization. Inputs FINC and FDEC provide increment and decrement signals to the charge pump for changing the oscillator frequency. The FINC and FDEC inputs gate the charge pump for the duration of the pulse width. The RC timing network formed by C_V and R_V at pins VCOI, VCON, and VCOP control the oscillators center frequency. (See Typical Performance Characteristics)

R_V should be greater than 1000Ω , Too low of a value will result in excessive power dissipation. R_L should be about 680Ω .

The VCO output should only be taken from pin VCON. Charge pump capacitor C_{CP1} is connected from pin FLTR to ground. Components R_{CP} and C_{CP2} are also connected in series from pin FLTR to ground to provide VCO loop compensation.

INTERNAL VOLTAGE REFERENCE

V_{REF} is an internal band-gap voltage reference. It is buffered and available at pin V_{REF} and is used by the ML4402, ML4403, ML4404 and other chips requiring a 5 volt reference.

EXTERNAL LOGIC

The external logic provided by the user typically has a complexity of about 150 to 300 equivalent gates. Complexity and architecture depends on the users di-bit pattern and control function.

Note: Stray capacitance should be considered in applying the above relationships when low capacitor values are used. Stray capacitance of the integrated circuit terminal is typically about 2 to 3pF.

TYPICAL PERFORMANCE CHARACTERISTICS

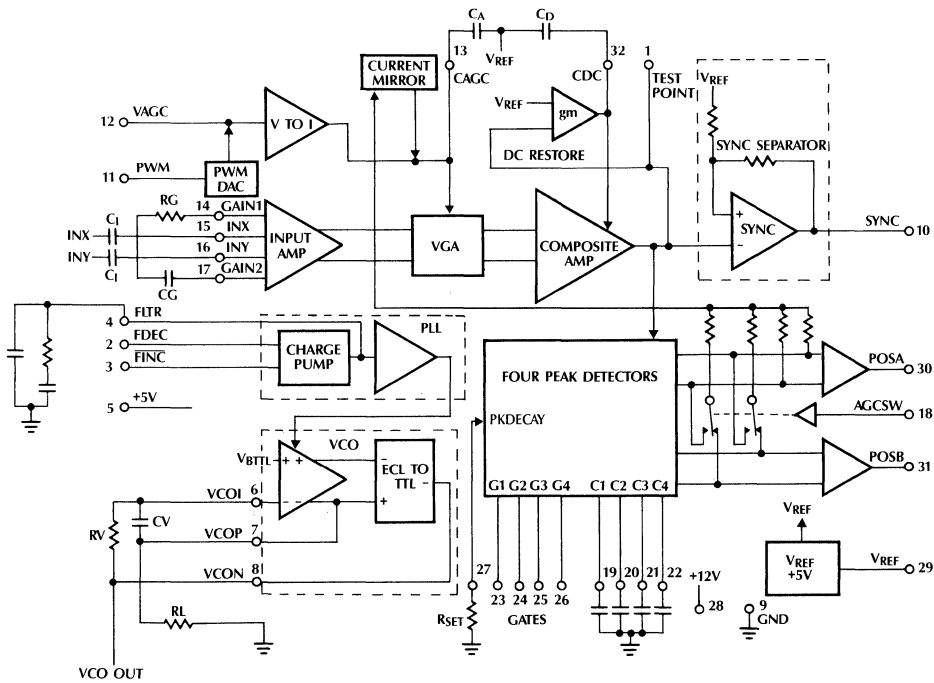
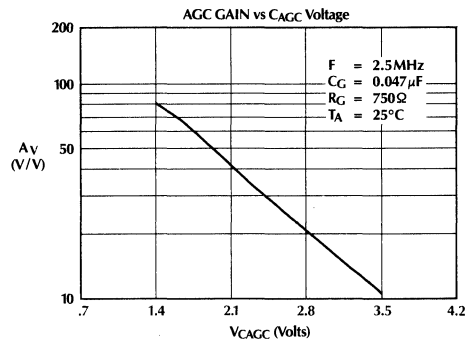


Figure 1. ML4431 Connection Diagram



ML4431

ORDERING INFORMATION

PART NUMBER	TEMP. RANGE	PACKAGE
ML4431CQ	0°C to +70°C	MOLDED PCC (Q32)

5V Disk Voice Coil Servo Driver

GENERAL DESCRIPTION

The ML4506 is a voice coil power driver intended for use in 5V Hard Disk servo systems. The ML4506 contains all power and control circuitry necessary to drive the voice coils of most small form factor drives. In addition, power fail detection and head retraction functions are provided for orderly shut-down of the drive.

The transconductance is programmed by a logic input at 1/4 A/V and 1/24 A/V respectively, using a 1Ω sense resistor. This allows for greater DAC resolution in digitally controlled servos during track follow without compromising dynamic range during seek.

The retraction circuit, main drive circuit, and control circuits are each powered from their own supplies. This allows maximum flexibility and provides for the lowest forward drop by eliminating the need for a blocking diode.

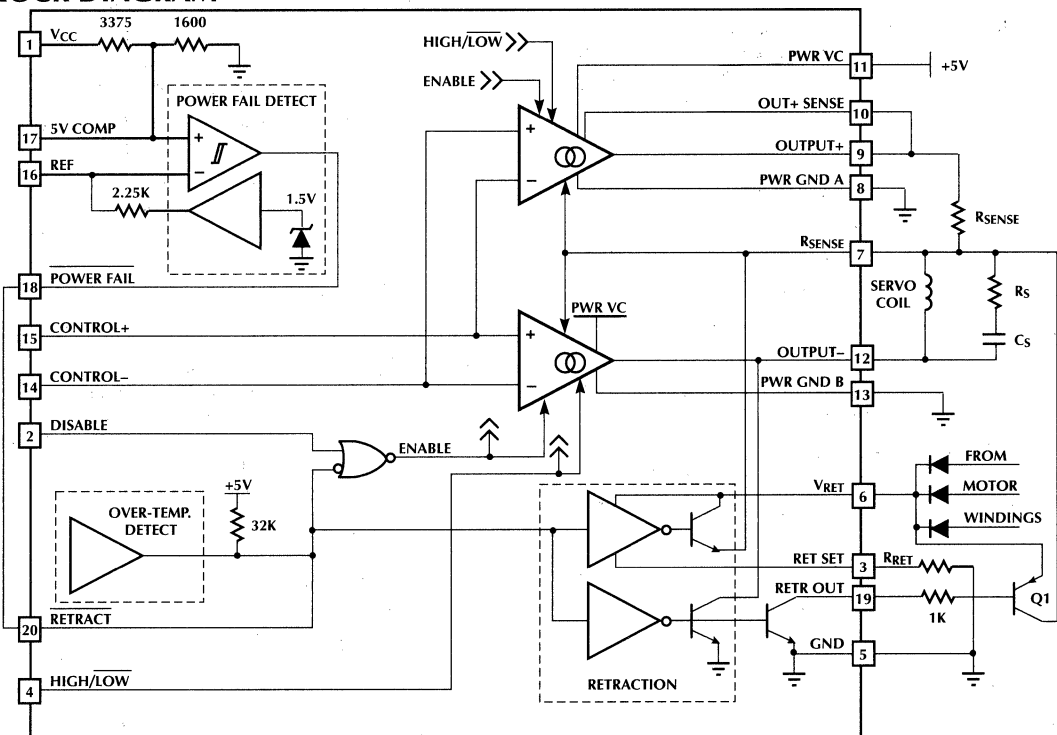
The power fail detection circuit includes a precision 1.5V bandgap reference.

The ML4506 is implemented using Micro Linear's bipolar array technology. This allows for easy customizing of the IC for a user's specific application.

FEATURES

- 500mA power output with 1.3V total forward drop
- Low offsets, cross-over distortion and quiescent current
- Pin-programmable transconductance settings
- Retraction circuitry with programmable retract current, voltage limiting, and separate supply pin.
- On-chip precision power fail detect circuitry
- Over-temperature protection with flag output
- Logic input available for disabling outputs

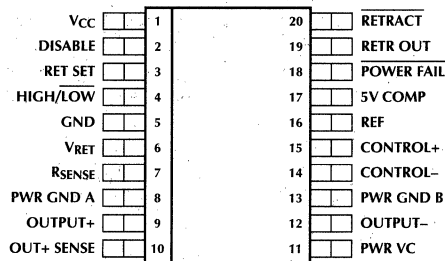
BLOCK DIAGRAM



ML4506

PIN CONNECTION

ML4506
20-PIN SOIC (S20W) OR 20-PIN SSOP (R20W)



TOP VIEW

PIN DESCRIPTION

PIN#	NAME	FUNCTION	PIN#	NAME	FUNCTION
1	V _{CC}	Positive Power supply for the IC. Normally connected to +5V.	12	OUTPUT-	Negative Output terminal for bridge amplifier.
2	DISABLE	A logic "1" turns off the main outputs.	13	PWR GND B	Ground Terminal for power amplifier.
3	RET SET	A Current into this sets up the voltage limit for the internal retract sourcing circuit	14	CONTROL-	Negative input for current command.
4	HIGH/LOW	A logic "1" sets the transconductance gain to 1/4 while a logic "0" sets the gain to 1/24. Transconductance gain is defined as: $\frac{V_{RSENSE}}{(CONTROL+) - (CONTROL-)}$	15	CONTROL+	Positive input for current command.
5	GND	Analog Signal Ground	16	REF	Reference input to the Power Fail comparator. Leave open to use internal 2.5V reference.
6	V _{RET}	Power supply for the retract circuit.	17	5V COMP	Input to the Power Fail Comparator. Can be connected to a bypass capacitor for noise immunity.
7	R _{SENSE}	Current sensing resistor terminal.	18	POWER FAIL	Open collector output drives low if pin 17 or pin 18 are below pin 16. Normally tied to pin 20.
8	PWR GND A	Ground Terminal for power amplifier A.	19	RETR OUT	Open collector output pulls low to drive external PNP for retract if V _{CC} is less than 3.5V and pin 20 is low.
9	OUTPUT+	Positive Output terminal for bridge amplifier.	20	RETRACT	A logic "0" input causes the main outputs to tri-state and the retraction circuit to activate. This input also functions as a flag output and will go low in the event of an over-temperature condition.
10	OUT+ SENSE	Positive Amplifier Kelvin sense terminal. Tie to OUTPUT+.			
11	PWR VC	+5V supply for bridge amplifier			

ABSOLUTE MAXIMUM RATINGS

Absolute maximum ratings are those values beyond which the device could be permanently damaged. Absolute maximum ratings are stress ratings only and functional device operation is not implied.

Supply Voltage (pins 1,6,11)	7V
Voltage Pins 2,4,18,19,20	-0.3V to +7V
Pins 14, 15	-0.3 to +V _{CC}
Output Current	±750mA
Retraction Current	80mA
Retract set current (pin 3)	3mA

Junction Temperature	150°C
Storage Temperature Range	-65°C to 150°C
Lead Temperature (Soldering 10 sec.)	150°C
Thermal Resistance (θ_{JA})	
SOIC Package (S)	55°C/W
SSOP Package (R)	65°C/W

OPERATING CONDITIONS

Temperature Range	0°C to 70°C
Supply Voltage (pins 1,11)	5V ± 10%
V _{RET} (pin 6)	1V to V _{CC}

ELECTRICAL CHARACTERISTICS

Unless otherwise specified, T_A = Operating Temperature Range, V_{CC} = 5V ± 10%, R_{SENSE} = 1Ω, CONTROL- (pin 15) = 2.5V, R_{SET} (pin 3) = 3.7kΩ, Load = 10Ω.

PARAMETER	CONDITION	MIN	TYP	MAX	UNITS
AMPLIFIER					
Control Common Mode Range		0.5		V _{CC} - 1	V
Offset				±10	mV
Transconductance Gain	pin 4 = 2V pin 4 = 0.8V	238 39.6	250 41.7	263 43.8	mA/V mA/V
Bandwidth			100		kHz
Sinking saturation	I _{OUT} = 100mA I _{OUT} = 300mA I _{OUT} = 500mA			0.5 0.6 0.8	V V V
Sourcing saturation	I _{OUT} = 100mA I _{OUT} = 300mA I _{OUT} = 500mA			1.1 1.2 1.3	V V V
RETRACTION CIRCUIT V_{PIN20} = 0.8V, V_{RET} = 2.5V					
I _{RET SET}			0.75		V
Turn on time			300		ns
Turn off time			8		μs
Sink current (I _{PIN12})	V _{PIN12} = 0.4V	34	50	150	mA
Source Voltage (V _{PIN7})	I _{PIN7} = -50mA	0.4	0.6	0.8	V
POWER FAIL DETECTION CIRCUIT					
Reference Voltage		1.35	1.50	1.65	V
Reference Source Impedance			2.25		kΩ
5V Threshold Hysteresis		4.40	4.575 30	4.75	V mV
LOGIC INPUTS					
Voltage High (V _{IH})		2	1.4		V
Voltage Low (V _{IL})			1.4	0.8	V
Current High (I _{IH})	V _{IN} = 5V			±10	mA
Current Low (I _{IL})	V _{IN} = 0V, except pin 20 V _{IN} = 0V, pin 20 only	-40 -250	-10 -160		mA mA
CURRENT CONSUMPTION					
Pin 1 + Pin 11	V _{PIN14} = V _{PIN15} = 2.5V		10	15	mA
Pin 6	V _{PIN14} = 2.5V		2.5	5.0	mA

FUNCTIONAL DESCRIPTION

POWER AMPLIFIER

The ML4506 power amplifier circuit is set up as a Howland Current source with a fixed gain of 1/4 or 1/24 (set by driving pin 4 high or low respectively). This architecture yields minimal cross-over distortion while maintaining low output cross conduction currents.

The gain figure refers to the ratio of input voltage to the output voltage seen across R_{SENSE} . For example, at a 1/4 gain setting, with $V(-)$ input at 2.5V and the $V(+)$ input at 3.5V, +500mA would flow through the coil using a 0.5Ω sense resistor. Under the same conditions with pin 4 low, the current would be 83mA. The ability to change from low to high gain allows more complete utilization of DAC resolution when in the track follow mode.

The output stage (figure 2) is designed to provide minimal saturation losses and employs a "composite PNP" for the sourcing drive and a saturable NPN to sink current. Sourcing saturation drop is typically 0.9V while sinking saturation drop is typically 0.4V.

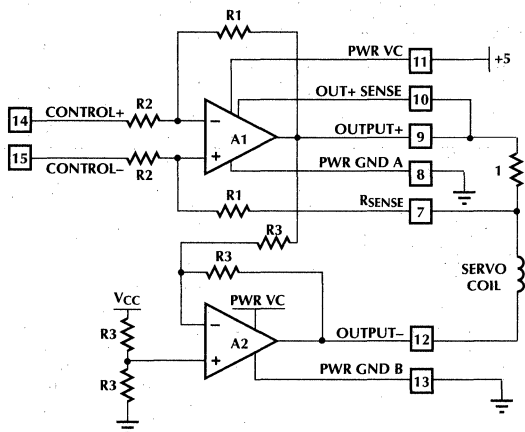


Figure 1. Power Amplifier Topology.

POWER FAIL DETECT

The ML4506 power fail detection circuit consists of a precision trimmed reference, resistor dividers, and a comparator with an effective hysteresis of 30mV. The output at pin 18 is open-collector and is normally tied to pin 1 which is internally pulled-up to 5V.

RETRACT

The retract circuit features provision for very low voltage operation as well as voltage limiting when a "live" retract with 5V on V_{RET} is performed. When pin 20 goes low, the internal NPN transistor will saturate, pulling SINK B (pin 11) low. A RETR OUT signal (open collector) saturates to drive an external PNP source transistor when pin 20 is low and when V_{RET} (pin 6) is below 3.5V. This portion of the circuit will function with less than 1V on V_{RET} .

An internal voltage limited pull-up circuit is provided which sources current on pin 7 to the VCM. This limit is set by an external resistor (see fig. 7) This circuit will operated reliably down to a V_{RET} voltage of around 2.5V. Pin 20 (Retract input) also serves as a flag to indicate an over-temperature condition on the die and goes low when the die temperature exceeds a safe operating limit (about 160°C).

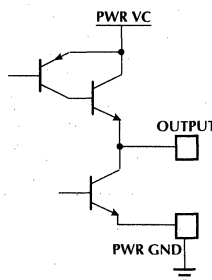


Figure 2. Power Output Stage.

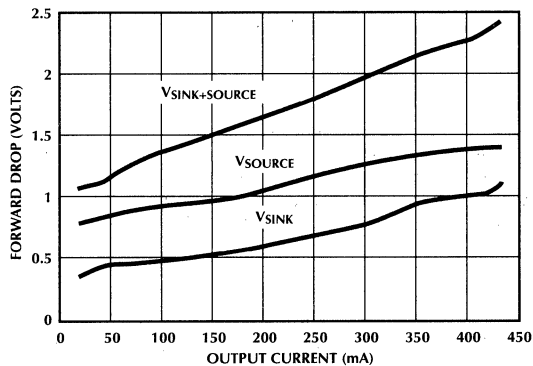


Figure 3. Output Saturation Voltage vs. Output Current.
($V_{CC} = PWR\ VC = 5V$)

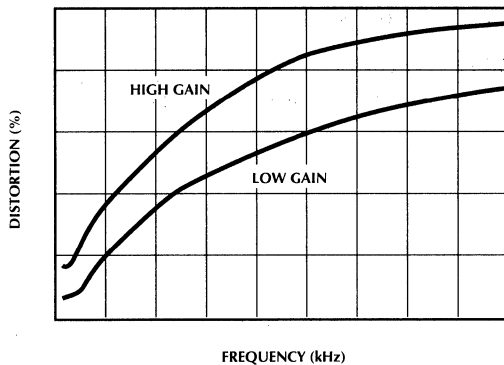


Figure 6. Total Harmonic Distortion vs. Frequency.
Low Gain Setting ($V_{PIN5} = 0$), $R_{SENSE} = 1\Omega$, $V_{IN} = 2.4V_{P-P}$
High Gain Setting ($V_{PIN5} = 0$), $R_{SENSE} = 1\Omega$, $V_{IN} = 0.4V_{P-P}$

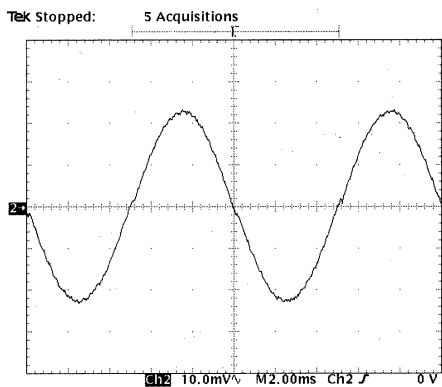


Figure 4. Output Current:
 $V_{IN} = 100Hz$ Sine Wave, $100mA_{P-P}$
Low Gain Mode ($V_{PIN5} = 0$), $R_{SENSE} = 0.5\Omega$, $R_L = 10\Omega$.

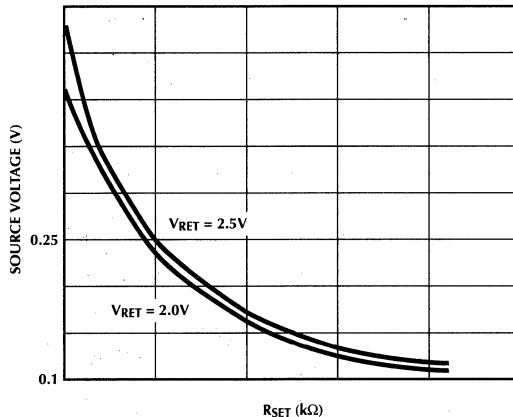


Figure 7. R_{SET} vs. Retract Source Voltage Limit.

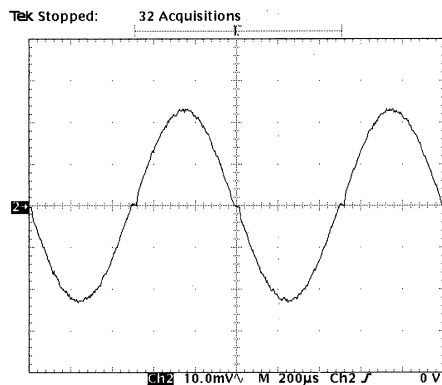


Figure 5. Output Current:
 $V_{IN} = 1kHz$ Sine Wave, $100mA_{P-P}$
Low Gain Mode ($V_{PIN5} = 0$), $R_{SENSE} = 0.5\Omega$, $R_L = 10\Omega$.

ML4506

APPLICATIONS

COMPENSATION

Figure 8 shows the equivalent AC circuit for the transconductance amplifier.

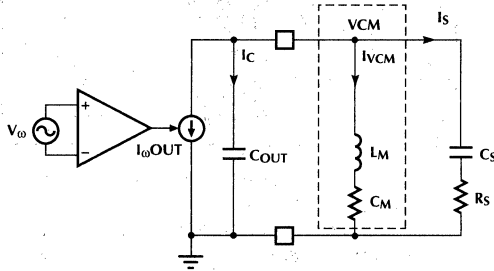


Figure 8. AC Equivalent Circuit for Current Amplifier, Voice Coil Motor (VCM) and Snubber.

The amplifier's current bandwidth is limited by C_{OUT} which varies with the value chosen for R_{SENSE}

$$C_{OUT} = \frac{25nF}{R_{SENSE}}$$

With no snubber (R_S and C_S) the bandwidth is limited to:

$$F_{-3dB} = \frac{1}{2\pi} \sqrt{\frac{2.414}{L(M) C(OUT)}}$$

Since this is a second order system with $L(M)$ and $C(OUT)$ forming a resonant circuit, some damping is desirable to reduce ringing in the step response. This is accomplished with resistive snubber. The optimum value of $R(S)$ occurs when the following condition is met:

$$R(S) = \sqrt{\frac{L(VCM)}{C(OUT)}}$$

For a given $C(S)$, setting $R(S)$ to this value will minimize the ringing in the transient response. Larger values of $R(S)$ will result in more ringing and more bandwidth. Smaller values of $R(S)$ will result in more ringing and less bandwidth. $R(S)$ should not exceed 300Ω .

$C(S)$ (snubber capacitor) values of between $200nF$ and $1\mu F$ are usually necessary to achieve the desired reduction of ringing in the step response. At optimum value of $R(S)$ larger values of $C(S)$ further reduce the ringing but do not affect the bandwidth.

Tuning the current loop response can be best done simulating the network in figure 8 with a computer simulator (such as SPICE).

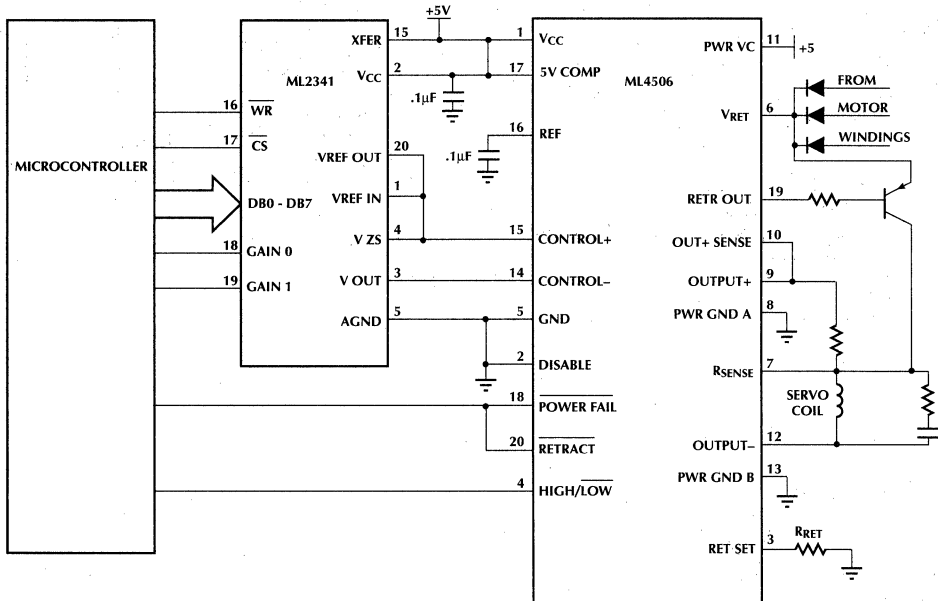


Figure 9. Typical Application: ML4506 used with ML2341 8-bit DAC provides up to 12-bit effective resolution.

ORDERING INFORMATION

PART NUMBER	TEMPERATURE RANGE	PACKAGE
ML4506CS ML4506CR	0°C to 70°C 0°C to 70°C	20-Pin SOIC (S20W) 20-Pin SSOP (R20)

Low Saturation 5V Voice Coil Servo Driver

GENERAL DESCRIPTION

The ML4508 is a voice coil power driver intended for use in both High Performance 5V Hard Disk servo systems. The ML4508 contains all control circuitry necessary to drive the voice coils of most small drives. To maximize compliance voltage, the ML4508 includes two 1 Amp NPN drivers and provides drivers for external PNP transistors. In addition, power fail detection and a low voltage head retraction functions are provided for orderly shut-down of the drive.

The transconductance programmed by a logic input at 1/4 A/V and 1/24 A/V respectively, when using a 1Ω sense resistor. This allows for greater DAC resolution in digitally controlled servos during track follow without compromising dynamic range during seek.

The retraction circuit, main drive circuit, and control circuits are each powered from separate supply pins. Retract is self-contained and allows the use of an external PNP for 5V systems to allow retraction with as little as 1V of back EMF from the spindle.

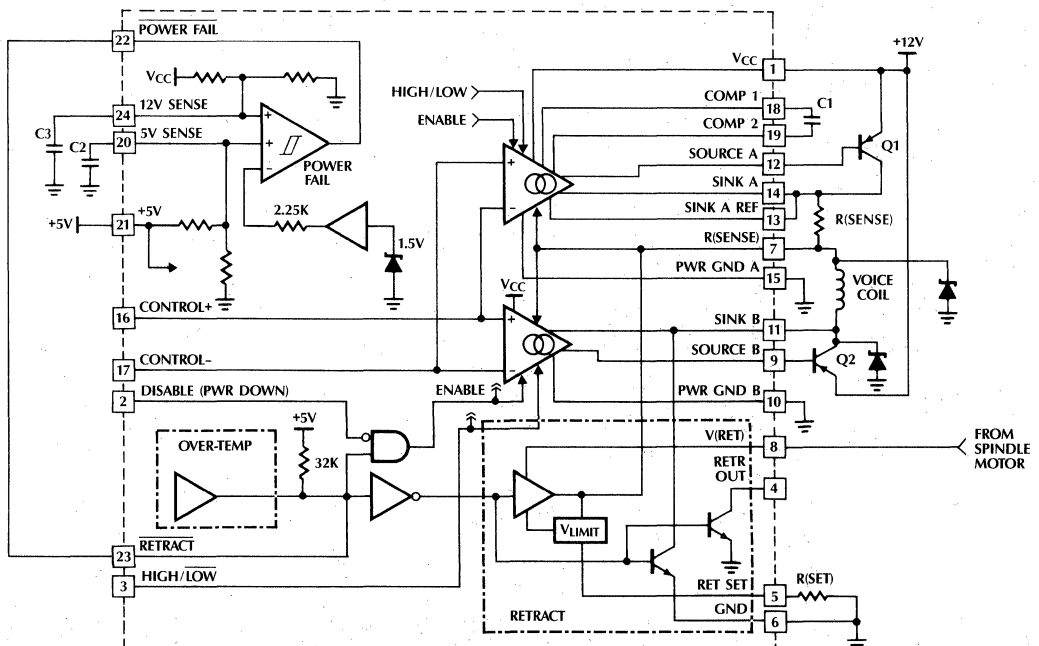
The power fail detection circuit includes a precision 1.5V bandgap reference and a power fail comparator.

The ML4508 is implemented using Micro Linear's bipolar array technology. This allows for customization of the IC for a user's specific application.

FEATURES

- Low saturation voltage (<1V at 1A.)
- No power supply blocking diode necessary
- No cross-over distortion with low quiescent current
- Pin-programmable transconductance settings
- Retraction circuitry with programmable retract voltage and separate power pin operates to 1V
- On-chip precision power fail detect circuitry
- Over-temperature protection with flag output
- Operates from +5V supplies

BLOCK DIAGRAM



5V Sensorless Spindle Motor Controller

GENERAL DESCRIPTION

The ML4510 provides complete commutation for delta or wye wound Brushless DC (BLDC) motors without the need for signals from Hall Effect sensors. This IC senses the back EMF of the 3 motor windings (no neutral required) to determine the proper commutation phase angle using phase lock loop techniques. This technique will commutate virtually any 3-phase BLDC motor and is insensitive to PWM noise and motor snubbing.

Included in the ML4510 is the circuitry necessary for a Hard Disk Drive microcontroller driven control loop. The ML4510 controls motor current with either a constant off-time PWM or linear current control driven by the microcontroller. Speed feedback for the micro is a stable digital frequency equal to the commutation frequency of the motor. All commutation is performed by the ML4510. Braking and Power Fail are also included in the ML4510.

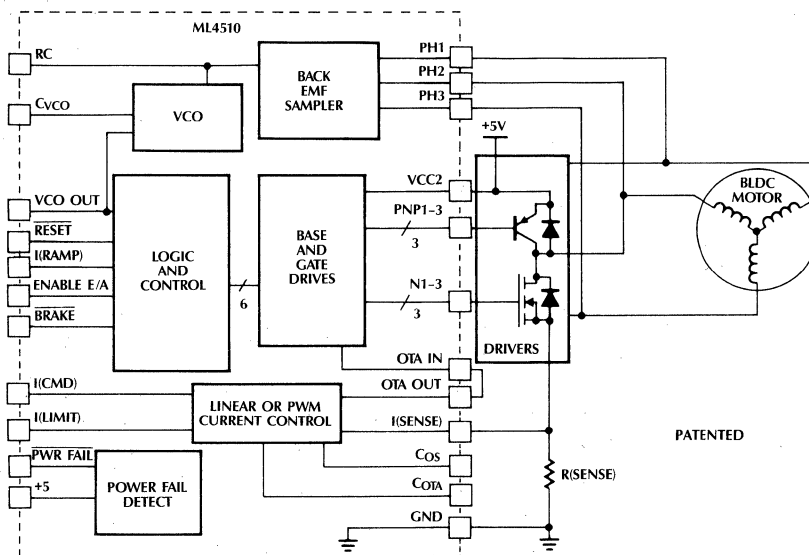
Since the timing of the start-up sequencing is determined by external circuitry, the system can be optimized for a wide range of motors and inertial loads.

The ML4510 modulates the gates of external N-channel power MOSFETs to regulate the motor current. The IC drives external PNP transistors or P-channel MOSFETs directly. Special circuits are used to save base drive power at low load currents.

FEATURES

- Back-EMF Commutation Provides Maximum Torque for Minimum "Spin-Up" Time for Spindle Motors
- Accurate, Jitter-Free Phase Locked Motor Speed Feedback Output
- Operates on Single 5V Power Supply
- Linear or PWM Motor Current Control
- Easy Microcontroller Interface for Optimized Start-Up Sequencing and Speed Control
- Power Fail Detect Circuit
- Drives External N-Channel FETs and PNP's or P-Channel FETs

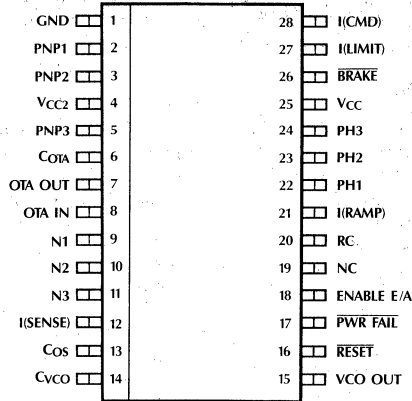
BLOCK DIAGRAM



ML4510

PIN CONFIGURATION

ML4510
28-Pin SOIC (S28W)



PIN DESCRIPTION

PIN #	NAME	FUNCTION	PIN #	NAME	FUNCTION
1	GND	Signal and Power Ground.	16	RESET	Input which holds the VCO off and sets the ML4510 to the RESET condition.
2	PNP1	Drives the external PNP power transistor driving motor PH1.	17	PWR FAIL	A "0" output indicates 5V is under-voltage.
3	PNP2	Drives the external PNP power transistor driving motor PH2.	18	ENABLE E/A	A "1" logic input enables the error amplifier and closes the back-EMF feedback loop.
4	VCC2	5V power.	19	NC	No Electrical Connection.
5	PNP3	Drives the external PNP power transistor driving motor PH3.	20	RC	VCO loop filter components.
6	C _{OTA}	Compensation capacitor for linear motor current amplifier loop.	21	I(RAMP)	Current into this pin sets the initial acceleration rate of the VCO during start-up.
7	OTA OUT	Output of motor current error amplifier, normally connected to OTA IN or to external MOSFET gate.	22	PH1	Motor Terminal 1.
8	OTA IN	Driving voltage for N1-N3. Normally tied to OTA OUT.	23	PH2	Motor Terminal 2.
9-11	N1, N2, N3	Drives the external N-channel MOSFETs for PH1, PH2, PH3.	24	PH3	Motor Terminal 3.
12	I(SENSE)	Motor current sense input.	25	VCC	5V power supply. Terminal which is sensed for power fail.
13	C _{OS}	Timing capacitor for fixed off-time PWM current control.	26	BRAKE	A "0" activates the braking circuit.
14	C _{VCO}	Timing capacitor for VCO.	27	I(LIMIT)	Sets the threshold for the PWM comparator.
15	VCO OUT	Logic output from VCO.	28	I(CMD)	Current Command for Linear Current amplifier.

ABSOLUTE MAXIMUM RATINGS

Absolute maximum ratings are those values beyond which the device could be permanently damaged. Absolute maximum ratings are stress ratings only and functional device operation is not implied.

Supply Voltage (pins 4, 25)	7V
Output Current (pins 2, 3, 5, 9, 10, 11)	±150mA
Logic Inputs (pins 16, 18, 26)	-0.3 to 7V
I(SENSE) Voltage (pin 12)	0.9V
Junction Temperature	150°C
Storage Temperature Range	-65°C to +150°C
Lead Temperature (Soldering 10 sec)	150°C
Thermal Resistance (θ_{JA})	60°C/W

OPERATING CONDITIONS

Temperature Range	0°C to +70°C
V _{CC} Voltage (pins 4, 25)	5V ± 10%
I(RAMP) Current (pin 21)	0 to 100µA
I Control Voltage Range (pins 27, 28)	0V to 3V

ELECTRICAL CHARACTERISTICS

Unless otherwise specified, T_A = Operating Temperature Range, V_{CC} = V_{CC2} = 5V, R_{SENSE} = 1Ω, C_{OTA} = C_{VCO} = .01µF, C_{OS} = .02µF

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
Oscillator (VCO) Section (Mode 1 or 2 unless otherwise specified)					
Frequency vs. V _{PIN 20}	1V ≤ V _{PIN 20} ≤ 3.5V		670		Hz/V
Frequency	V _{VCO} = 2.5V	830	1675	2500	Hz
	V _{VCO} = .5V	120	245	350	Hz
Reset Voltage at C _{VCO}	Mode = 0		125	250	mV
Sampling Amplifier					
V _{RC}	Mode 0		125	250	mV
I _{RC}	Mode 1, R _{RAMP} = 39KΩ	16	33	50	µA
	Mode 2A, V _{PH2} = 0.5V	30	60	90	µA
	Mode 2A, V _{PH2} = 2.5V	-6	2	6	µA
	Mode 2A, V _{PH2} = 4.5V	-30	-60	-90	µA
Motor Current Control Section					
I(SENSE) Gain	V _{PIN 27} = 5V, 0V ≤ V _{PIN 28} ≤ 2.5V	4	5	6	V/V
One Shot Off Time		12	25	33	µs
I(CMD) Transconductance Gain			.19		mmho
Power Fail Detection Circuit					
5V Threshold		3.8	4.25	4.5	V
Hysteresis			70		mV
Logic Inputs					
Voltage High (V _{IH})		2			V
Voltage Low (V _{IL})				.8	V
Current High (I _{IH})	V _{IN} = 2.7V	-10	1	10	µA
Current Low (I _{IL})	V _{IN} = 0.4V	-250	-120	-60	µA

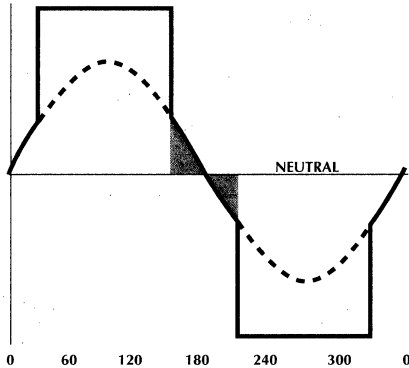


Figure 2. Typical Motor Phase Waveform with Back-EMF Superimposed (Ideal Commutation).

control voltage will be no higher than $V_{CC(MIN)} - 1V$. The VCO maximum frequency will be:

$$F_{MAX} = 0.05 \times POLES \times RPM$$

where POLES is the number of poles on the motor and RPM is the maximum motor speed in Revolutions Per Minute.

The minimum VCO gain derived from the specification table (using the minimum F_{VCO} at $V_{VCO} = 2.5V$) is:

$$K_{VCO(MIN)} = \frac{3.32 \times 10^{-6}}{C_{VCO}}$$

Assuming that the $V_{VCO(MAX)} = 3.2V$, then

$$C_{VCO} = \frac{3.2 \times 3.32 \times 10^{-6}}{F_{MAX}}$$

or

$$C_{VCO} = \frac{212}{POLES \times RPM} \mu F$$

Figure 3 shows the transfer function of the Phase Lock Loop with the phase detector formed from the sampled phase through the Gm amplifier with the loop filtered formed by R, C₁, and C₂.

The impedance of the loop filter is

$$Z_{RC}(s) = \frac{1}{C_1 s} \frac{(s + \omega_{LEAD})}{(s + \omega_{LAG})}$$

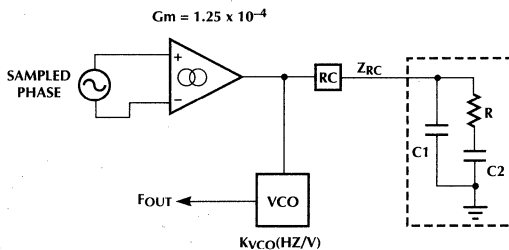


Figure 3. Back EMF Phase Lock Loop Components.

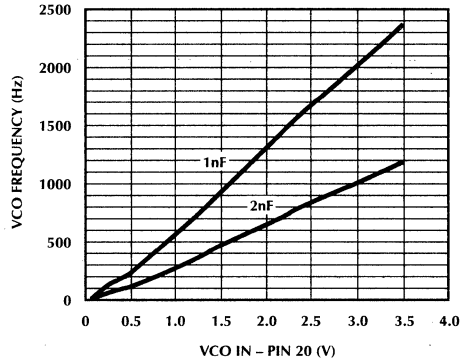


Figure 4. VCO Output Frequency vs. V_{VCO} (Pin 20)

Where the lead and lag frequencies are set by:

$$\omega_{LEAD} = \frac{1}{RC_2}$$

$$\omega_{LAG} = \frac{C_1 + C_2}{RC_1 C_2}$$

Requiring the loop to settle in 20 PLL cycles with a spread of 10 between $\omega_{LEAD} = 10 \times \omega_{LAG}$ produces the following calculations for R, C₁ and C₂:

$$C_1 \approx \frac{1.97 \times 10^{-9}}{C_{VCO} \times F_{VCO}^2}$$

$$C_2 = 9 \times C_1$$

$$R = \frac{12.65}{C_2 \times F_{VCO}}$$

START-UP SEQUENCING

When the motor is initially at rest, it is generating no back-EMF. Because a back-EMF signal is required for closed loop commutation, the motor must be started "open-loop" until a velocity sufficient to generate some back-EMF is attained (around 100 RPM).

Two modes are possible for starting the motor. For the lowest possible starting time, the chip is held in the reset (mode R) state by holding pin 16 low and providing full current to the motor (figure 5).

Step	Pin 16	Pin 18	Pin 21	I(LIMIT) I(CMD)
1	0	0	Fixed	I_{MAX}
2	1	0	Fixed	I_{MAX}
3	1	1	0	I_{MAX}

Figure 5. Minimum Time Start-Up Sequence.

ML4510

Step 1: The IC is held in reset (mode R) with full power applied to the windings (see figure 7). This aligns the rotor to a position which is 30° (electrical) before the center of the first commutation state.

Step 2: A fixed current is input to pin 21 and appears as a current on pin 20, and will accelerate the motor at a fixed rate.

Step 3: When the motor speed reaches about 100 RPM, the back-EMF loop can be closed by pulling pin 18 low.

Using this technique, some reverse rotation is possible. The maximum amount of reverse rotation is $360/N$, where N is the number of poles. For an 8 pole motor, 45° reverse rotation is possible.

STATE	OUTPUTS						INPUT SAMPLING
	N1	N2	N3	PNP1	PNP2	PNP3	
R or 0	OFF	ON	OFF	ON	OFF	ON	N/A
A	OFF	OFF	ON	ON	OFF	OFF	PH2
B	OFF	OFF	ON	OFF	ON	OFF	PH1
C	ON	OFF	OFF	OFF	ON	OFF	PH3
D	ON	OFF	OFF	OFF	OFF	ON	PH2
E	OFF	ON	OFF	OFF	OFF	ON	PH1
F	OFF	ON	OFF	ON	OFF	OFF	PH3

Figure 6. Commutation State Table.

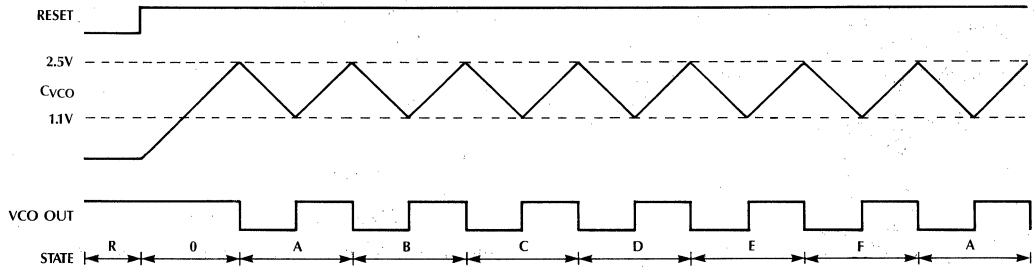


Figure 7. Start-Up Timing and Mode Sequencing.

SPEED CONTROL — CURRENT LOOP

To facilitate speed control, the ML4510 includes two current control loops — linear and PWM (figure 8). The linear control loop senses the motor current on the I(SENSE) terminal through R_{SENSE} . An internal current sense amplifier's output modulates the gates of the 3 N-channel MOSFET's when OTA OUT is tied to OTA IN, or can modulate a single MOSFET gate to control current.

The ML4510 also includes a current mode constant off-time PWM circuit. When motor current builds to the threshold set on I(LIMIT) input (pin 27), a one-shot is fired whose timing is set by C_{OS} . The current in the motor will be controlled by the lower of pin 27 and pin 28.

OUTPUT DRIVERS

The motor's source transistor drivers are open-collector NPN's with internal $8K\Omega$ pull-up resistors, whose current is controlled according to the current demanded through the motor. To conserve power, the ML4510 sets the current to PNP1, PNP2, and PNP3, proportional to the lower of pin 27 and pin 28.

Drivers N1 through N3 are totem-pole outputs capable of sourcing and sinking 10mA. Switching noise in the external MOSFETs can be reduced by adding resistance in series with the gates.

BRAKING

Applying a logic 0 on pin 26 activates the braking circuit. The brake circuit turns on PNP1 through PNP3 and turns off NPN1 through NPN3.

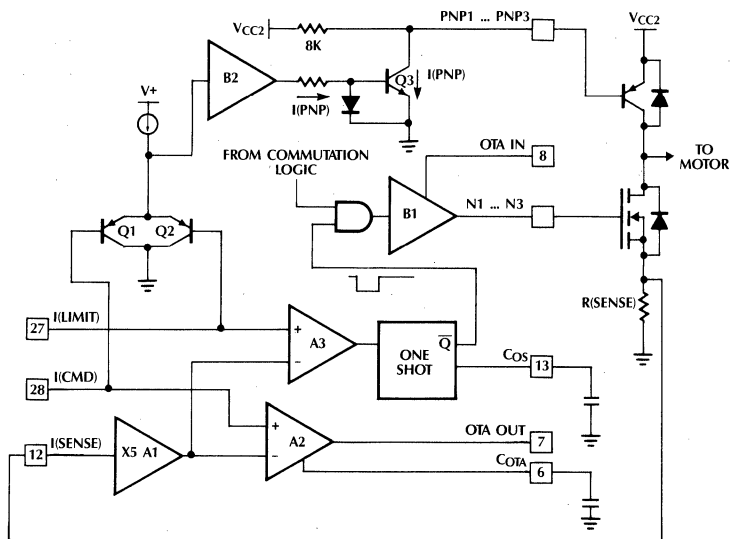


Figure 8. Current Control and Output Drive.

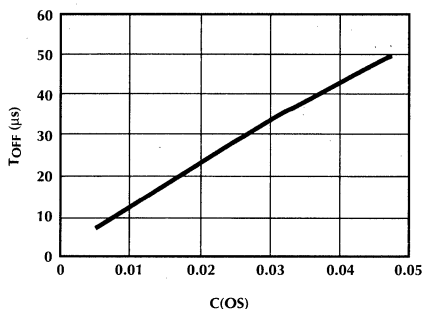


Figure 9. I(LIMIT) Output Off-Time vs. C_{OS} .

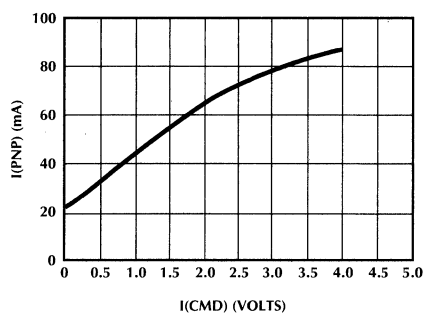


Figure 10. Available PNP Drive Current vs. I(CMD) Input.

ML4510

APPLICATIONS

Figure 11 shows a typical application of the ML4510 in a hard disk drive spindle control. Although the timing necessary to start the motor in most applications would be generated by a microcontroller, Figure 12 shows a simple "one-shot" start-up timing approach.

Speed control can be accomplished either by:

1. Sensing the VCO OUT frequency with a Microcontroller and adjusting I(CMD) via an analog output from the Micro (PWM DAC).

2. Using analog circuitry for speed control (Figure 13).

Q1, Q2, and Q3 are MJE210 or equivalent. Q4, Q5, and Q6 are IRFU010 or equivalent. Base resistors (50Ω) are included to reduce power dissipation in the IC during start-up. If requested currents are low, these can be eliminated. Switching transients due to commutation can be reduced by increasing the 470Ω gate resistors on Q4-Q6.

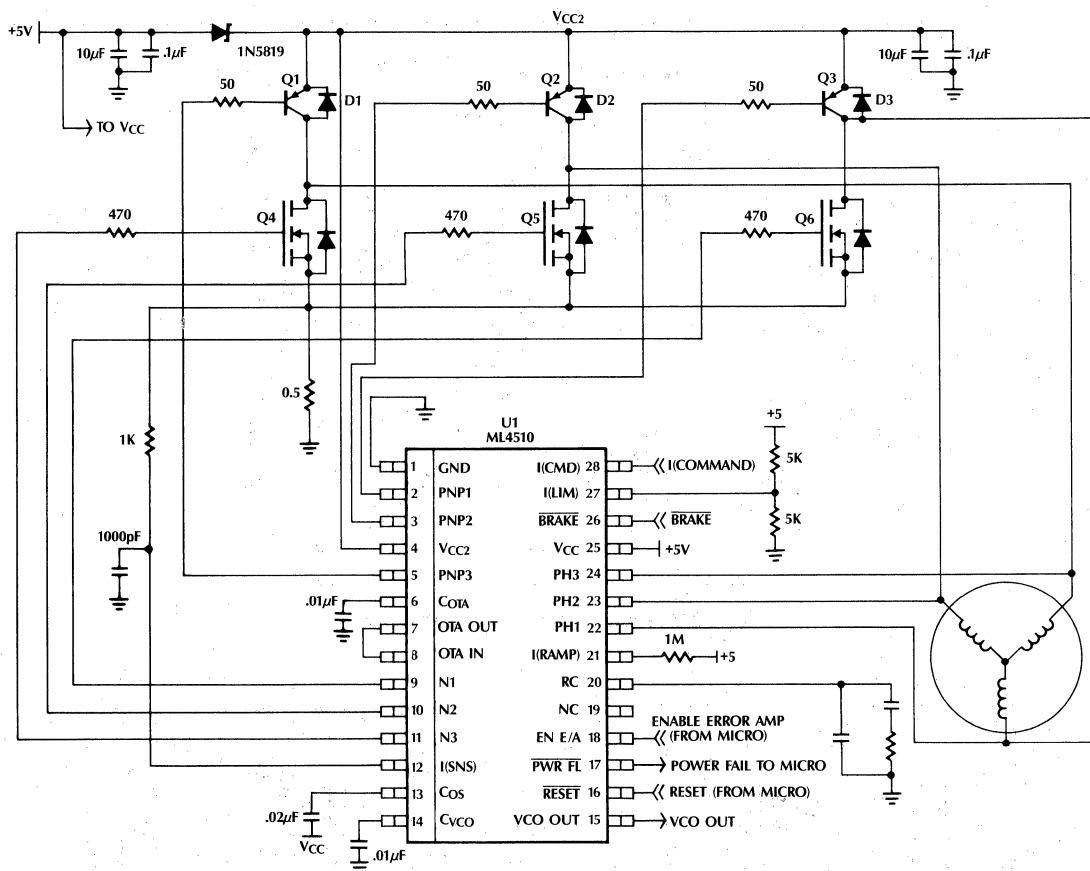


Figure 11. ML4510 Typical Application.

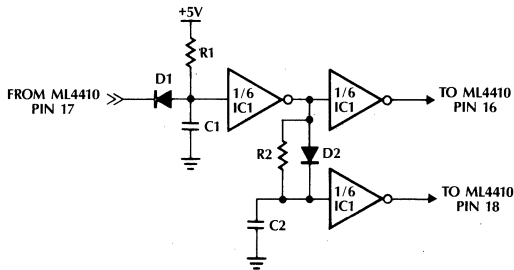
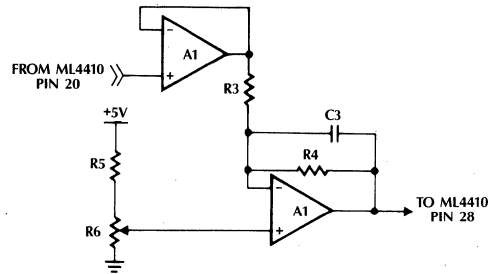


Figure 12. Analog Start-up Circuit.



Symbol	Value
A1	LM358
Q1	74HC14
D1, D2	1N4148
R1	1M Ω
R2	1M Ω
R3	100K Ω

Symbol	Value
R4	100K Ω
R5	50K Ω
R6	50K Ω
C1	3.3 μ F
C2	3.3 μ F
C3	.47 μ F

Figure 13. Analog Speed Control.

ORDERING INFORMATION

PART NUMBER	TEMPERATURE RANGE	PACKAGE
ML4510CS	0°C to +70°C	28-Pin SOIC (S28W)

Servo Burst Area Detector

GENERAL DESCRIPTION

The ML4532 and ML4533 Area Detectors are designed to minimize the pipeline transport delay while accurately quantizing the area of servo bursts in high-speed embedded servo systems. Combined with the ML2261 or ML2264 High Speed A/D Converters, the ML4532 and ML4533 are designed to capture back-to-back servo bursts in a 700ns or larger window. Power dissipation is minimized by the use of a digital power down pin which allows the area detector to be powered down between the servo sectors. The ML4536 is the ML4533 with different reference voltage levels.

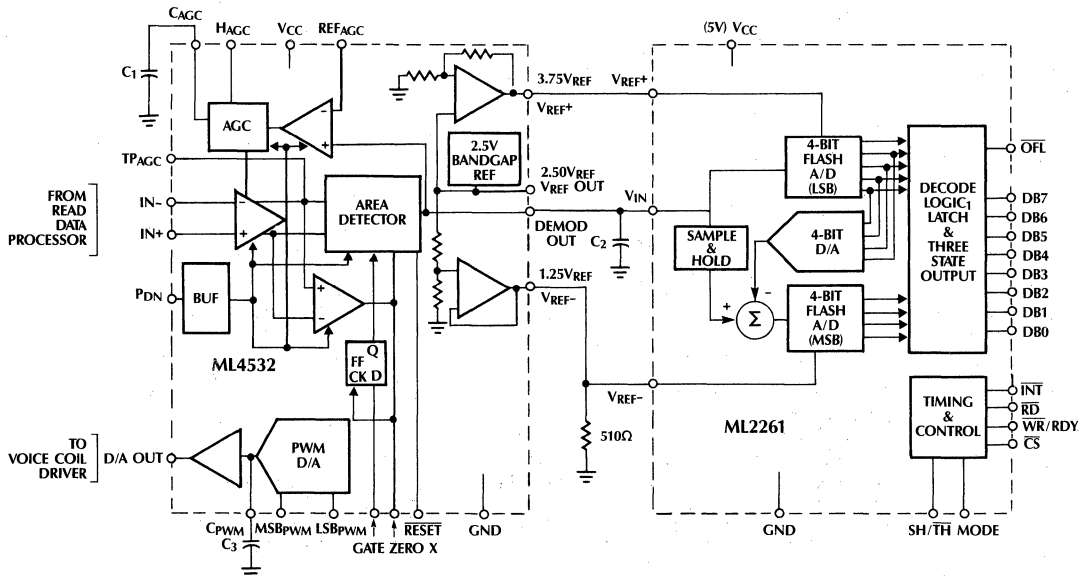
The ML4532 includes a PWM D/A for microprocessor control of the actuator driver, changing the REF_{AGC} pin voltage during head change, or other system control functions.

FEATURES	ML4532	ML4533	ML4536
Package Options	20-Pin PCC or SSOP	16-Pin SOIC	16-Pin SOIC
Zero-Scale Ref. Output Voltage	1.25V	1.25V	1.0V
Full-Scale Ref. Output Voltage	3.75V	3.75V	3.4V
Reference Output Voltage	2.5V	2.5V	2.2V
PWM D/A Onboard	Yes	No	No

FEATURES

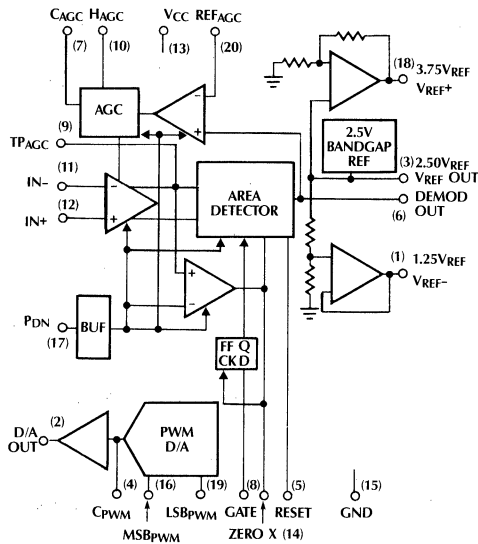
- Allows for Area Detection of 1 μ s back-to-back bursts
- AGC amplifier for maintaining accuracy
- 0.2% nonlinearity between 25% and 75% of input signal range
- 2% nonlinearity over the input signal range
- Provides zero- and full-scale outputs for A/D converter
- 5V supply, at 29mA for ML4533/ML4536, 35mA for ML4532
- Digitally controlled power down for minimizing power between sectors
- Bandgap Reference output
- ML4532 includes PWM D/A for controlling voice coil driver or AGC during head change
- ML4533/ML4536 available in 16-pin SOIC package ML4532 available in 20-pin PCC or SSOP package
- Reference outputs defining the minimum and maximum demodulation output values

TYPICAL APPLICATION

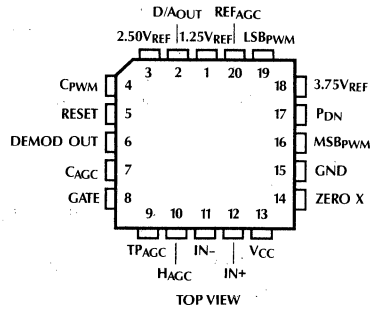


Typical HDD Digital Servo Application

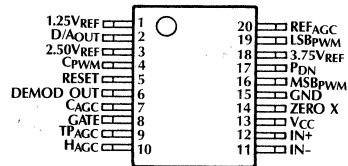
ML4532 BLOCK DIAGRAM AND PIN CONNECTION



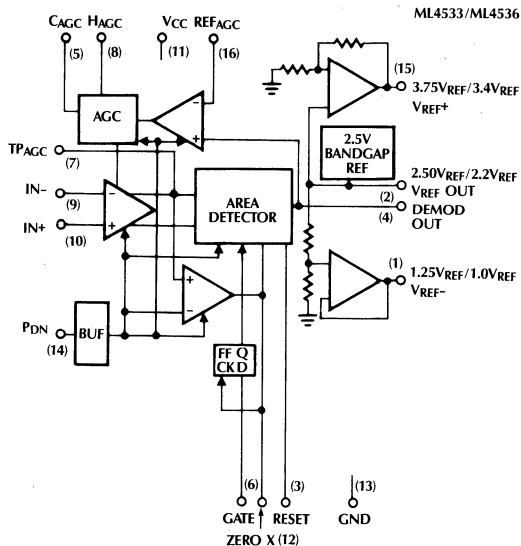
20-Pin PCC



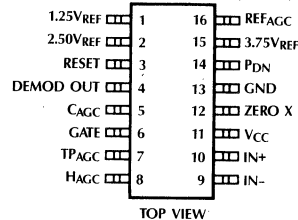
20-Pin SSOP



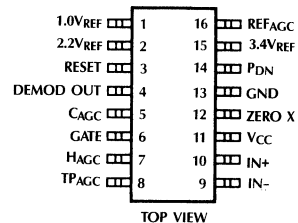
ML4533/ML4536 BLOCK DIAGRAM AND PIN CONNECTION



16-Pin SOIC (ML4533)



16-Pin SOIC (ML4536)



ML4532, ML4533, ML4536

PIN DESCRIPTION

ML4533/ ML4532		ML4533/ ML4536		
PIN #	PIN #	NAME	DESCRIPTION	
1	1	1.25V _{REF} / 1.00V _{REF}	Zero scale reference output.	
2		D/A _{OUT}	Analog output of PWM D/A.	
3	2	2.50V _{REF} / 2.20V _{REF}	2.50 voltage reference output.	
4		C _{PWM}	PWM D/A smoothing capacitor.	
5	3	RESET	Reset, active high.	
6	4	DEMODO OUT	Area detector output.	
7	5	C _{AGC}	AGC capacitor.	
8	6	GATE	Defines area detect window, active high.	
9	7	TP _{AGC}	Output test point for AGC.	
10	8	H _{AGC}	AGC hold input, AGC active when high; AGC constant when low.	
	11	9	IN ₋	Negative input.
	12	10	IN ₊	Positive input.
	13	11	V _{CC}	5V power supply.
	14	12	ZERO X	Zero detector crossing output.
	15	13	GND	Ground.
	16		MSB _{PWM}	PWM D/A most significant bit input.
	17	14	P _{DN}	Power down control, reduces power if logic high.
	18	15	3.75V _{REF} / 3.40V _{REF}	Full scale reference output.
	19		LSB _{PWM}	PWM D/A least significant bit input.
	20	16	REF _{AGC}	AGC voltage reference.

ABSOLUTE MAXIMUM RATINGS

(Note 1)

Power Supply Voltage, V _{CC}	8V
Input Voltage	-0.3V to +8V
Storage Temperature	-65°C to +150°C
Package Dissipation at T _A = 25°C (Board Mount)	
20-Pin PCC	875mW
20-Pin SSOP	750mW
16-Pin SOIC	750mW
Package Lead Temperature	
Soldering (10 sec)	260°C
Vapor Phase (60 sec)	215°C
Infrared (15 sec)	220°C

OPERATING CONDITIONS

Temperature Range	0 to +70°C
Supply Voltage (V _{CC})	5V ± 5%
(I _{IN+}) - (I _{IN-})	1V _{P-P}
C _{AGC}	100pF
REF _{AGC}	2.5V
C _{PWM}01μF
C _H at DEMOD OUT	100pF

ELECTRICAL CHARACTERISTICS

The following specifications apply over the recommended operating conditions of $T_A = 0$ to $+70^\circ\text{C}$, $V_{CC} = 4.75$ to 5.25V , and external component values as recommended above, unless otherwise specified.

PARAMETER	NOTES	CONDITIONS	MIN	TYP (Note 2)	MAX	UNITS
DEMODULATOR						
Differential Input Range	3	For Full Scale Output	.25		2	V_{P-P}
Differential AGC Range	3		0.8		1.5	V/V
Differential Input Resistance				4		$k\Omega$
Differential Input Capacitance				5		pF
Common Mode Input Resistance				2		$k\Omega$
Power Supply RR				40		dB
Differential Nonlinearity	5, 3 5, 4	25% to 75% of Full Scale Zero to Full Scale		.2 2	2 5	% %
DEMOD OUT Offset Current		$V_{IN} = \pm 500\text{mV}$			20	μA
Maximum DEMOD OUT Charge Current	3	GATE = High	500			μA
DEMOD OUT Leakage Current	3	GATE = Low			+5	μA
DEMOD OUT Reset Voltage (ML4532/33) (ML4536)			1.15 0.9	1.25 1.0	1.35 1.1	V V
DEMOD OUT Reset Current		Discharge, RESET = High	2.0			mA
AUTOMATIC GAIN CONTROL						
AGC Dynamic Range	3		2.5			V/V
AGC Output Swing	3		1			V
C_{AGC} Charging Current	3		150		250	μA
C_{AGC} Discharging Current	3		150		250	μA
C_{AGC} Leakage Current	3				5	μA
VOLTAGE REFERENCES						
1.25 V_{REF} Output Voltage (ML4532/33)	3	$T_A = 25^\circ\text{C}$	1.20	1.25	1.30	V
3.75 V_{REF} Output Voltage (ML4532/33)	3	$T_A = 25^\circ\text{C}$	3.60	3.75	3.90	V
2.50 V_{REF} Output Voltage (ML4532/33)	3	$T_A = 25^\circ\text{C}$	2.40	2.50	2.60	V
1.0 V_{REF} Output Voltage (ML4536)	3	$T_A = 25^\circ\text{C}$	0.95	1.0	1.05	V
3.4 V_{REF} Output Voltage (ML4536)	3	$T_A = 25^\circ\text{C}$	3.2	3.4	3.6	V
2.2 V_{REF} Output Voltage (ML4536)	3	$T_A = 25^\circ\text{C}$	2.05	2.2	2.35	V
Load Regulation	3	$0\text{mA} \leq I_{OUT} \leq 5\text{mA}$	-5		+5	mV/mA
Line Regulation			-30		+30	mV/V
DIGITAL AND DC						
Logical "0" Input Voltage	3				.8	V
Logical "1" Input Voltage	3		2.0			V
Logical "0" Input Current	3	$V_{IN} = 0.4\text{V}$			-1.5	mA
Logical "1" Input Current	3	$V_{IN} = 2.5\text{V}$			+100	μA
Logical "0" Output Voltage	3	ZERO X, $I_{OUT} = 1\text{mA}$.5	V
Logical "1" Output Voltage	3	$I_{OUT} = -1\text{mA}$	3.0			V
Supply Current ML4532	3	$P_{DN} = \text{Low}$			45	mA
	3	$P_{DN} = \text{High}$		11	15	mA
ML4533/ML4536	3	$P_{DN} = \text{Low}$			38	mA
	3	$P_{DN} = \text{High}$		5	12	mA
Monotonicity	4		9	10		Bits
LSB to MSB Ratio	3		16.0	16.5	18.0	V/V

ML4532, ML4533, ML4536

ELECTRICAL CHARACTERISTICS (Continued)

The following specifications apply over the recommended operating conditions of $T_A = 0$ to $+70^\circ\text{C}$, $V_{CC} = 4.75$ to 5.25V , and external component values as recommended above, unless otherwise specified.

PARAMETER	NOTES	CONDITIONS	MIN	TYP (Note 2)	MAX	UNITS
D/A CONVERTER (ML4532 Only)						
Output Voltage Swing	3	$R_L = 5\text{K}$	1.25		3.75	V
Logical "0" Input Voltage	3				.8	V
Logical "1" Input Voltage	3		2.0			V
Logical "0" Input Current	3	$V_{IL} = .4\text{V}$			-1	mA
Logical "1" Input Current	3	$V_{IH} = 2.5\text{V}$			300	μA
DYNAMIC AND AC PARAMETERS (Figures 1 and 2)						
f_B , Burst Input Frequency	4		1		10	MHz
t_{CS} , Gate Edge Setup Prior to Burst \downarrow Zero Crossing	4, 6				30	ns
t_{RESET} , Reset Pulse Width	4, 6	$C_H \leq 200\text{pF}$	300			ns
t_{BZX} , Burst Zero Crossing to ZERO X Output	4, 6	$C_L = 50\text{pF}$			25	ns
t_{PG} , Power Down \downarrow to Gate \uparrow	4, 5			200	400	ns

Note 1: Absolute Maximum Ratings are limits beyond which the life of the integrated circuit may be impaired. All voltages unless otherwise specified are measured with respect to ground.

Note 2: Typicals are parametric norm at 25°C .

Note 3: Parameter guaranteed and 100% production tested.

Note 4: Parameter guaranteed. Parameters not 100% tested are not in outgoing quality level calculation.

Note 5: Linearity measured as a percentage of the midpoint between 25% to 75% of full scale.

Note 6: Timing measured at 1.4V.

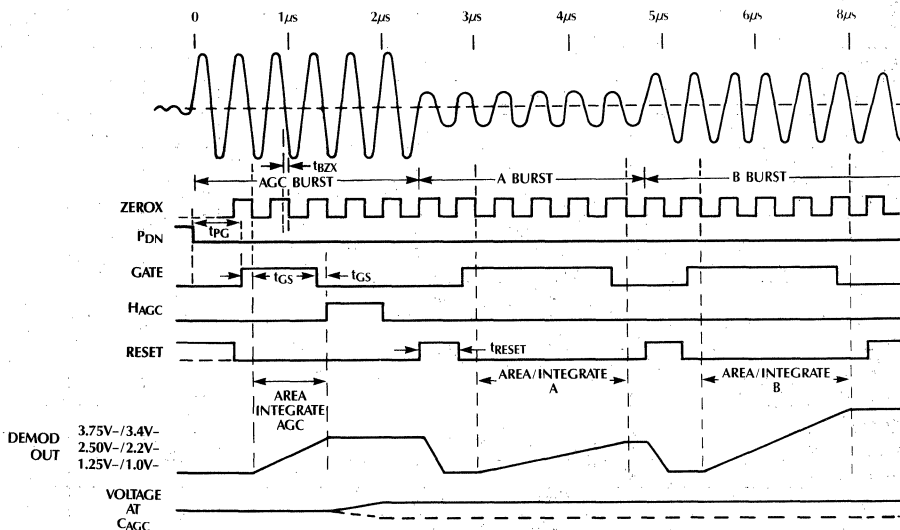


Figure 1. AGC Burst Timing

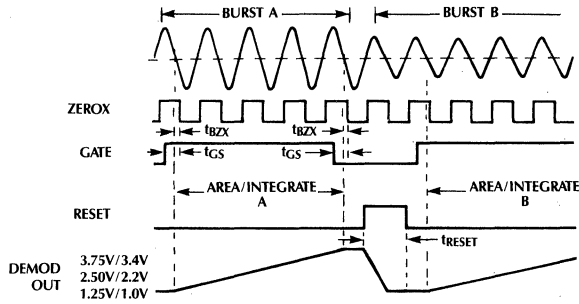


Figure 2. Burst Area Detect Timing

TYPICAL PERFORMANCE CURVES

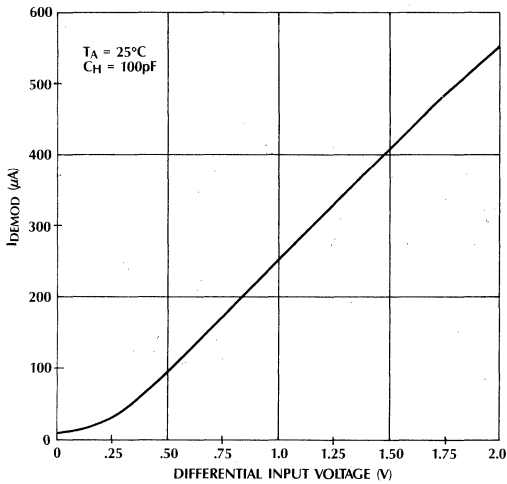


Figure 3. DEMOD Output Current vs. Input

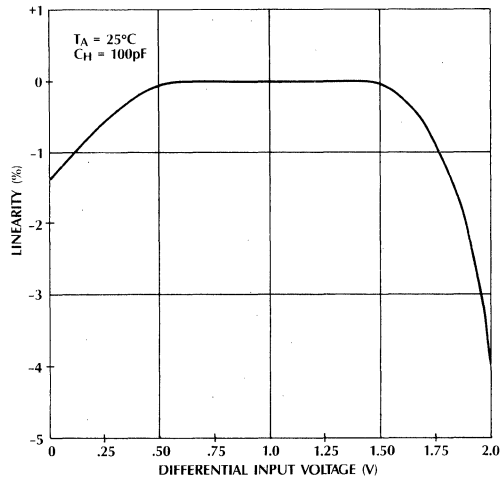


Figure 4. Linearity vs. Input

ML4532, ML4533, ML4536

1.0 FUNCTIONAL DESCRIPTION

The ML4532, ML4533 and ML4536 are composed of an AGC amplifier, an area detector, and a band-gap reference with three buffered outputs. In addition the ML4532 (see Figure 5) includes a pulse width modulation D/A. The ML4536 is essentially the ML4533 with a different set of reference voltages.

1.1 INPUT AMPLIFIER AND AUTOMATIC GAIN CONTROL

The inputs of the ML4532, ML4533 and ML4536 are intended for use at the output of the read channel filter, accepting a $0.25V_{P-P}$ to $2V_{P-P}$ signal range. The input amplifier and AGC circuit of these area detectors operate in a differential signal mode to provide good common mode and power supply rejection. The purpose of the AGC loop is to maintain a constant area detect value that correlates to the zero scale ($1.25V_{REF}/1.0V_{REF}$) and full scale ($3.75V_{REF}/3.4V_{REF}$) output values based upon the minimum and maximum burst value. The sensing for the AGC is at the output of the area detector, allowing signal ranging based on the area of burst rather than the signal level of the burst. The AGC is intended to be updated at every sector of servo position bursts such that the signal variances due to platter radius and differences in read channel data frequencies can be corrected. The initial gain of the AGC circuit is established by the voltage applied to the REF_{AGC} input.

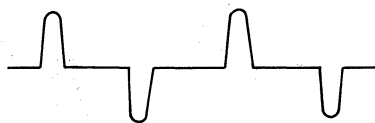
In this closed-loop system, the area detected output voltage is fed back and compared with the REF_{AGC} voltage in the G_M amplifier with a G_M of $1/4000$ ohms, to provide a gain control current, charging and discharging C_{AGC} .

The AGC value is held constant by the hold function and is controlled by H_{AGC} pin. When H_{AGC} is at a logic high the level of gain can change up or down and is held at a constant gain with a logic low input.

A capacitor from ground to the C_{AGC} pin holds the gain setting when H_{AGC} is at a logic low level and the area detector output does not affect the gain setting in this mode. See figure 1 for the AGC burst timing.

1.2 AREA DETECTOR

The area detector provides a measurement of servo burst area during a time window beginning at the first falling zero crossing edge after the GATE input is placed in a logic high state and ends at the first falling zero crossing edge after the GATE input is placed in a logic low state. The Zero crossing output enables the user to time the gate pulse by counting zero crossings. The analog input should be without open baseline by either keeping burst pulse spacing sufficiently close to avoid it or band limiting the signal. In most cases, both are necessary.



- a. May clock anywhere and give multiple transitions, not acceptable.



- b. Proper spacing.



- c. Band limiting.

The value of the area measurement is held on the output hold capacitor (C_H) until the RESET line is asserted. The RESET pin when placed in a logic high state for at least 300ns resets the area detector output to $1.25V$ which is the zero scale reference point and equals the voltage value on the $1.25V_{REF}$ pin. See Figure 2 for position area burst detection timing.

ZERO X Detector Output

The output of the zero crossing detector (comparator) is provided for system synchronization. This signal is internally generated in ECL, but an internal ECL to TTL converter is provided to simplify external interfacing to this signal.

1.3 BANDGAP REFERENCE

A $2.5V$ bandgap reference is included on the ML4532 and ML4533 and a $2.2V$ one in the ML4536, to set up internal biasing and establish the on-track reference level. This is also a buffered output. Full-scale (V_{REF+}) and zero scale (V_{REF-}) outputs are derived and buffered from the bandgap to simplify the interface to A/D converters, such as the ML2261 or ML2264. The $1.25V_{REF}$ pin is tied directly to the V_{REF-} pin of the A/D converter and with a 510Ω resistor to ground. The $3.75V_{REF}$ pin is tied directly to the V_{REF+} pin of the high speed A/D converter. The ML4536 offers a $1.0V_{REF}$ and $3.4V_{REF}$ for interface with the A/D converter on the Zilog type microcontroller devices.

1.4 PWM D/A OF THE ML4532

A D/A is included on the ML4532 for driving the VCM driver to position the head or for any other desired system error compensation, such as processor-controlled AGC set point during head change. This is a PWM D/A and requires a pulse width modulation logic signal from the microcontroller signal to be applied to the MSB_{PWM} and LSB_{PWM} pins. The buffered and filtered output appears at the (D/A OUT) pin. This output voltage swing is centered around the 2.5V_{REF} pin (2.5V). The end-points of the D/A output are defined by:

PWM DAC Output Description:

Range Point	Duty Cycle at MSB _{PWM}	Duty Cycle at LSB _{PWM}	Voltage Output
"Negative" Full-Scale:	0%	0%	1.094V
	0%	100%	1.250V
	100%	0%	3.594V
"Positive" Full Scale:	100%	100%	3.750V

The D/A is designed for 8-bit binary coding with the MSB weighted 16 times the LSB. The MSB_{PWM} and LSB_{PWM} inputs are negative true in that if these inputs are in a low state for 100% of the time the D/A output will be 1.094 volts and if are held in a high state for 100% of the time the output will be at 3.75 volts. The D/A output voltage is 1.25 volts if the MSB_{PWM} input pulse width has a 1/16 or 6.25% positive duty cycle and the LSB_{PWM} is in a constant low state or if the MSB_{PWM} input is held in a low state and the LSB_{PWM} input is held in a high state. The output voltage range that is controlled by the MSB_{PWM} input is 2.5 volts and the LSB_{PWM} input controls 156mV. The time constant for the PWM smoothing filter is approximately $3k\Omega \times C_{PWM}$. The external capacitor (C_{PWM}) should be made sufficiently large to smooth out the PWM ripple.

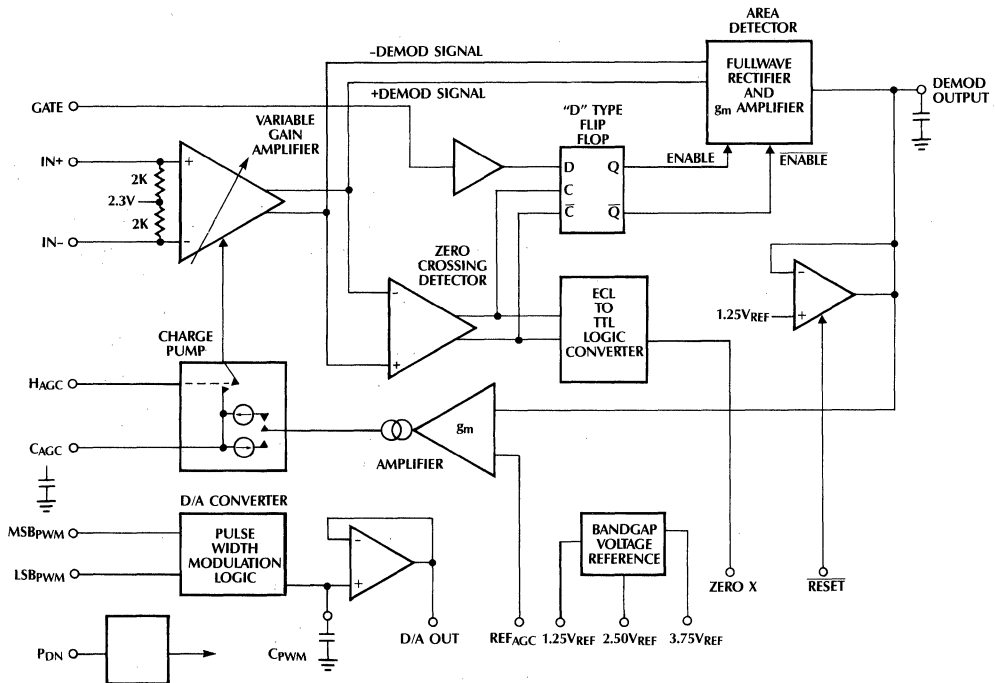


Figure 5. ML4532 Functional Block Diagram

ML4532, ML4533, ML4536

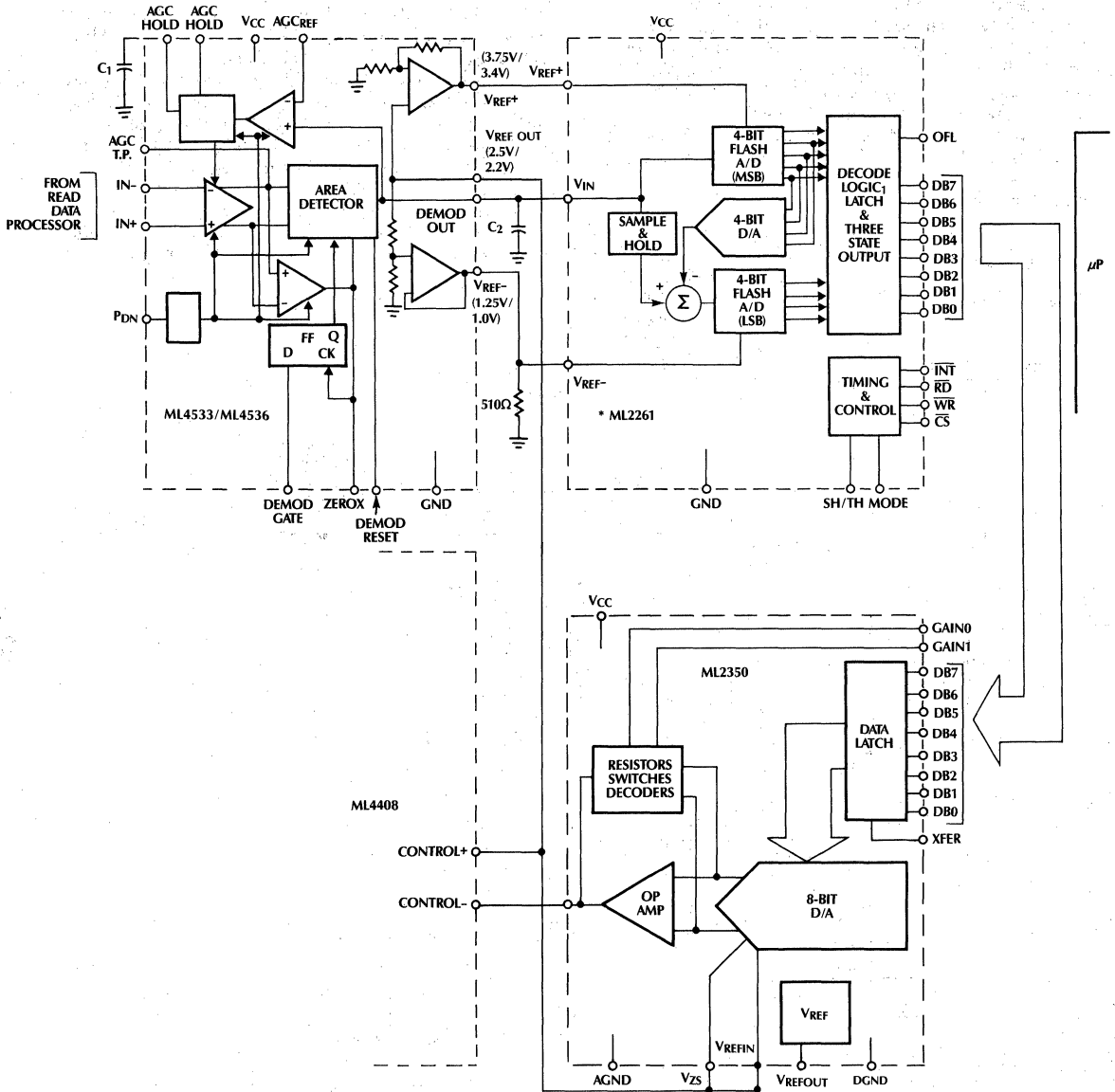


Figure 6. Typical Application ML4533/ML4536

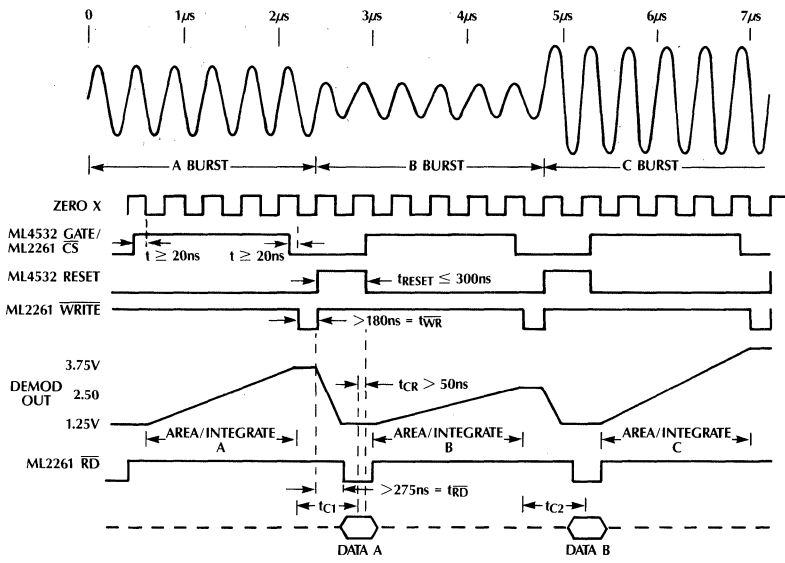


Figure 7. ML4532 Application Timing with ML2261 A/D Converter

ML4532, ML4533, ML4536

ORDERING INFORMATION

PART NUMBER	TEMP. RANGE	PACKAGE
ML4532CQ ML4532CR	0°C to +70°C 0°C to +70°C	MOLDED PCC (Q20) MOLDED SSOP (R20)
ML4533CS ML4536CS	0°C to +70°C 0°C to +70°C	MOLDED SOIC (S16) MOLDED SOIC (S16)

Area Detector Based Embedded Servo Demodulator

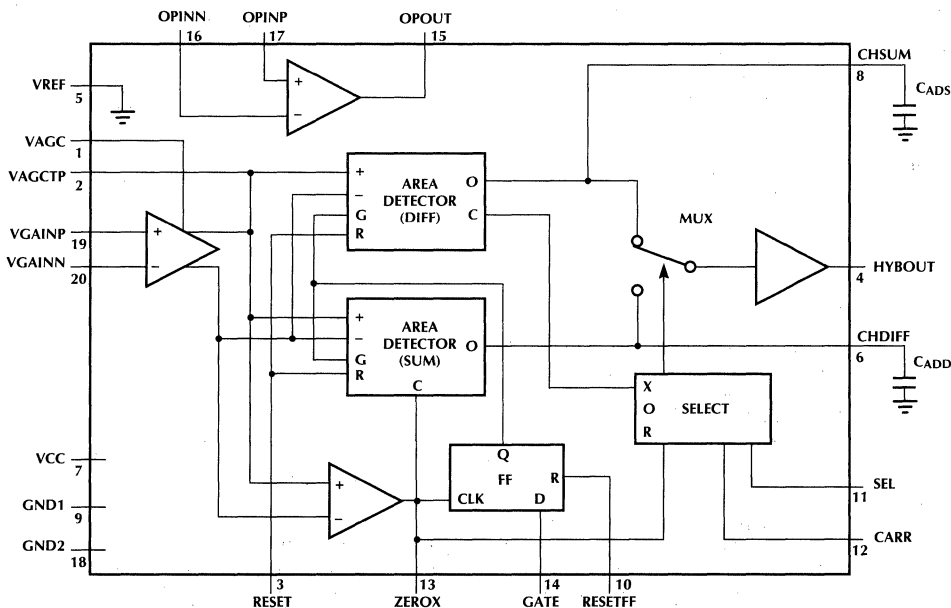
GENERAL DESCRIPTION

The ML4534 Embedded Servo Demodulator IC is designed for use in the hybrid data surface channel of an high-performance disk drive. Hybrid data are interleaved on the data surface with data records and encoded in A/B differential burst format, with a AGC field preceding the burst information. The AGC field is used by the read channel to set AGC gain levels in the burst area, which once established are held fixed for the duration of the servo burst. The demodulator measures burst amplitude using an area detection scheme, for improved noise immunity and provides both (A-B) and (A+B), to permit position error normalization with on-chip synchronization and reset functions. Using the SEL and CARR inputs the on-chip multiplexer allows selection of either (A-B) or (A+B) as the output. The multiplexer and area detection capacitors operate in concert to provide a hold capability for both the (A-B) and (A+B) outputs. Also included is an uncommitted operational amplifier which could be used for voice-coil motor current sensing.

FEATURES

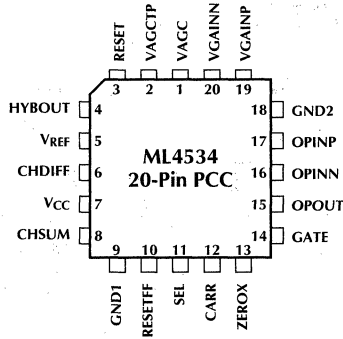
- Allows for Area Detection of back-to-back bursts
- 2% nonlinearity over input signal range
- Reset forces voltage on the Area Detecting capacitors C_{ADD} to V_{REF} & C_{ADS} to $V_{REF}/2$
- Separate Reset provided for Resync Flip-Flop
- Muxed/Selectable (A-B) & (A+B) demodulator output
- General purpose operational amplifier, applicable for use in voice coil motor current sensing
- 5V supply, 20-pin, J-leaded, PLCC package

BLOCK DIAGRAM



ML4534

PIN CONNECTION



PIN DESCRIPTION

PIN NO.	NAME	FUNCTION	PIN NO.	NAME	FUNCTION
1	VAGC	Gain Control input on the VGA	12	(Cont.)	area detector integrates the B burst in a direction opposite to that in which the A burst is integrated, thus realizing the (A-B) differencing operation. Carrier polarity in the (A+B) Area Detector is not affected by the state of the CARR pin. CARR pin in conjunction with the SEL pin, selects the multiplexer output.
2	VAGCTP	Test point connected through an isolation resistor to the output of the VGA	13	ZEROX	This is the output of the Carrier Comparator. +ZEROX is a nominally square wave having transitions coinciding with zero crossings of the output VGA.
3	RESET	Asserting this input pin resets the area detector (DIFF) to $V_{REF} (+2.5V)$ and the area detector (SUM) to $V_{REF}/2$	14	GATE	Asserting this line enables the (A+B) and (A-B) area detectors to measure area of the output signal of the VGA. This signal is re-synchronized to the area detector carrier internally before application to the area detectors.
4	HYBOUT	Output of the multiplexing amplifier, with V_{REF} , (A-B) or (A+B) area detector output, depending on the state of the SEL & CARR pins.	15	OPOUT	Optional operational amplifier (short circuit protected) output.
5	V_{REF}	+2.5V reference voltage input.	16	OPINN	Optional operational amplifier inverting input.
8	CHSUM	The (A+B) area detector integrating capacitor is connected between this pin and the ground.	17	OPINP	Optional operational amplifier non-inverting input.
7	V_{CC}	+5 Volt supply	18	GND	Ground
6	CHDIFF	The (A-B) area detector integrating capacitor is connected between this pin and the ground.	19	VGAINP	VGA non-inverting input. Inputs should be AC coupled
9	GND	Ground	20	VGAINN	VGA inverting input. Inputs should be AC coupled
10	RESETEFF	Active high signal resets the resynch flip-flop			
11	SEL	This pin in conjunction with the CARR pin, governs the multiplexer channel selection as follows : SEL CARR Mux Channel 0 X V_{REF} 1 0 (A-B) 1 1 (A+B)			
12	CARR	Asserting this pin high inverts the carrier input of the (A-B) area detector. CARR should be asserted throughout the B burst of the A/B burst pair. While the CARR pin is asserted, the (A-B)			

NOTE: The value of the CHSUM capacitor should be roughly twice that of the CHDIFF capacitor. It is also advisable to include a small resistor in series with the capacitor on the CHSUM pin and also the CHDIFF pin, to improve settling time.

ABSOLUTE MAXIMUM RATINGS

DC Supply Voltage (V_{CC})	-0.3 to +7 VDC
Storage Temperature (T_{STG})	-65 to +150°C
Package Dissipation	
$T_A = 25^\circ\text{C}$ (Board Mount)	875mW
Package Lead Temperature:	
Soldering (10 sec)	260°C
Vapor Phase (60 sec)	215°C
Infrared (15 sec)	220°C

RECOMMENDED OPERATING CONDITIONS

DC Supply Voltage Range (V_{CC})	$5 \pm 5\%$ VDC
Temperature Range	0 to +70°C
Operating Junction Temperature (T_J)	+25 to +125°C

ELECTRICAL CHARACTERISTICS

The following specifications apply over the recommended operating conditions of $T_A = 0$ to +70°C, $V_{CC} = 4.75$ to 5.25V, and external component values as recommended, unless otherwise stated.

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNIT
I_{CC}	$V_{AGC} = 4.0V$, $V_{REF} = 2.5V$	20	40	60	mA
I_{VREF}	$V_{REF} = 2.5V$	-50	25	200	μA
GATE, CARR, SEL, RESET, RESETFF	For all signals in test program	V_{IH} V_{IL}		2.0	V V
GATE, CARR, SEL, RESET, RESETFF	For $V_{IH} = 2.4V$ and $V_{IL} = 0.8V$	I_{IH} I_{IL}	-250 -400	-10 40	μA μA
VGAINPDC (VINP)	VGAINP, VGAINN open	2.3	2.5	2.7	V
VGAINNDC (VINN)	VGAINP, VGAINN open	2.3	2.5	2.7	V
V_{OH} ZEROX	VGAINP = 3.5, $I_{OH} = -0.4\text{mA}$ VGAINN = 1.5V	2.7		5.0	V
V_{OL} ZEROX	VGAINN = 3.5, $I_{OL} = 2.0\text{mA}$ VGAINP = 1.5V	0		0.5	V

OPERATIONAL AMPLIFIER USED FOR MOTOR CURRENT SENSING

I_{BIAS} Offset	OPINN = OPINP = 1.0V	-200		200	nA
V_{OS} - MCS	$A_V = 2.0$, $V_{IN} = 0$	-15		+15	mV
V_{OH} - MCS	$A_V = 2.0$, $V_{IN} = -1.0$, $I_{SRC} = -1.5\text{mA}$	3.8		5.0	V
V_{OL} - MCS	$A_V = 2.0$, $V_{IN} = 1.0$, $I_{SINK} = 1.5\text{mA}$	0		1.0	V
I_{SINK} - MCS	Openloop, OPINP = 0.0V OPINN = 1.0, OPOUT = V_{CC}	1.5		10	mV
I_{BIAS} - MCS	OPINN = 1.0, OPINP = 1.0 ($I_{OPINN} + I_{OPINP}$)/2	-2.0		0.0	μA
Amplifier Settling Time (t_{SMCS})	$R_{OUT} = 604\Omega$, $C_{OUT} = 36\text{pF}$		0.4	1.0	μs
Amplifier Bandwidth		4	8		MHz
Amplifier Gain (A_V)	Open Loop	58	63		dB

AGC

A_V - VGAMIN	Minimum Gain of AGC with 400mV input	0		1.1	V/V
A_V - VGAMAX	Maximum Gain of AGC with 100mV input	6.6		20	V/V
VAGCBIAS	$V_{AGC} = 1.0$	0		200	μA

ML4534

ELECTRICAL CHARACTERISTICS (Continued)

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNIT
RESET CIRCUITRY					
I RESET SUM, DIFF	RESET = V_{IH}	80		400	μA
I OFF SUM, DIFF	RESET = V_{IL}	-10		10	nA
V CH SUM RESET = V_{IH}	1.245	1.260	1.275	V	
V CH DIFF RESET = V_{IH}		2.490	2.5	2.510	V
I SUM, DIFF UNBAL	GATE = V_{IH} , CLOCK, 1V swing VAGC 1X, Measure Current with VGAINP = $V_{INP} + 0.2$ and VGAINN = $V_{INN} - 0.2$, then do VGAINP = $V_{INP} - 0.2$ and VGAINN = $V_{INN} + 0.2$, Subtract	-40		40	μA
I DIFF UNBAL XOR	CARR = V_{IH}	-40		+40	μA
I PEAK SUM	VGAINP = $V_{INP} + 1.0$ VAGC = 1.0	-540	-400	-265	μA
I PEAK DIFF P	VGAINP = $V_{INP} + 1.0$ VGAINN = $V_{INN} - 1.0$ VAGC = 1.0, CARR = V_{IL}	-540	-400	-265	μA
I PEAK DIFF ON	VGAINP = $V_{INP} + 1.0$ VGAINN = $V_{INN} - 1.0$ VAGC = 1.0, CARR = V_{IH}	265	400	540	μA
VOH SUM	VGAINP = $V_{INP} + 1.0$ VGAINN = $V_{INN} - 1.0$	3.9		5.0	V
VOH DIFF	VGAINP = $V_{INP} + 1.0$ VGAINN = $V_{INN} - 1.0$ CARR = V_{IL}	3.9		5.0	V
VOL DIFF	VGAINP = $V_{INP} + 1.0$ VGAINN = $V_{INN} - 1.0$ CARR = V_{IH}	0.0		1.0	V
IGATE	GATE = V_{IH} , CLOCK, VAGC 1X, $V_{SWING} = 1.0$	-10		10	nA
I RESET FF	RESET FF = V_{IH}	-10		10	nA

ELECTRICAL CHARACTERISTICS (Continued)

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNIT
MUX AMPLIFIER					
VHYBOUT	CARR = V_{IL} , SEL = V_{IL}	2.4		2.6	V
VOS MUX SUM	CARR = SEL = V_{IH} , CHSUM = 2.5	-8		8	mV
VOS MUX DIFF	CARR = V_{IL} , SEL = V_{IH} , CHDIFF = 2.5	-8		8	mV
IBIASSUM	CARR = SEL = V_{IH} , CHSUM = 2.5	0		300	nA
IBIASDIFF	CARR = V_{IL} , SEL = V_{IH} , CHSUM = 2.5	0		300	nA
VOHMUX	CARR = SEL = V_{IH} , CHSUM = 3.95, $I_{SRC} = 1.5\text{mA}$	3.8		5.0	V
VOLMUX	CARR = V_{IL} , SEL = V_{IH} , CHDIFF = 0.95, $I_{SINK} = 1.5\text{mA}$	0		1.0	V
ISINKMUX	CARR = V_{IL} , CHDIFF = 0.95 SEL = V_{IH} , VHYBOUT = V_{CC}	1.5		10	mA
Amplifier settling time (t_{SMUX})	$R_{OUT} = 604\Omega$, $C_{OUT} = 36\text{pF}$		0.4	1	μs
$I_{LEAKAGE}$			10	nA	
Linearity	0 to 1 V_{INPUT} , with VAGC such that $A_V \text{ VAGC} = 1.0$	-5		5	%F.S

FUNCTIONAL DESCRIPTION

The ML4534, +5V Embedded Servo Demodulator IC is designed for use in the hybrid data surface channel of an high-performance disk drive. Hybrid data are interleaved on the data surface with data records and encoded in A/B differential burst format, with an AGC field preceding the burst information. The AGC field is used by the read channel to set AGC gain levels in the burst area, which once established are held fixed for the duration of the servo burst. The demodulator measures burst amplitude using an area detection scheme, for improved noise immunity and provides both (A-B) and (A+B), to permit position error normalization. Using the SEL and CARR inputs, the on-chip multiplexer allows selection of either (A-B) or (A+B) on the output. The multiplexer and area detection capacitors operate in concert to provide a hold capability for both the (A-B) and (A+B) outputs. The area detectors are designed to minimize the pipeline transport delay while accurately quantizing the area of servo bursts in high speed hybrid servo systems. The major functional blocks of the ML4534 are briefly discussed below.

VARIABLE GAIN AMPLIFIER

Hybrid servo burst data from the disk read channel are capacitively coupled into the VGA through the differential input pins (VGAINP, VGAINN). VGA gain is controlled by the voltage on the VAGC pin, and the gain is varied in order to secure constant area of the output signal and counteract the amplitude regulating operation of the read channel AGC loop.

VGABUF COMPARATOR

The VGABUF comparator detects zero crossings of the composite signal delivered by the VGA. The output of this comparator controls the synchronous rectification of the composite VGA output, in the area detectors.

The comparator output is provided at a TTL level on the ZEROX pin. Control logic in the servo channel employs the ZEROX signal to produce an area detector enabling gate, which spans a fixed number of cycles of the composite signal.

AREA DETECTORS (SUM AND DIFF)

The area detectors detect A and B burst levels by area detection. Two area detectors are provided — one to measure the sum of A and B bursts (A+B), and a second one to measure the difference (A-B). Each area detector is

implemented as a gated current — output synchronous rectifier driving an external charge accumulating integrating capacitor. Area detection occurs only while the area detector is enabled under control of the GATE pin. When the detector is disabled, the integrating capacitor is effectively floated. An on-chip binary (FF) re-synchronizes the gating signal to remove any phase shifts due to logic delays in the external gate control logic. Initial conditions on the integrating capacitors are established prior to an area detecting operation by a reset circuit controlled by the RESET pin. A reset operation forces the voltage on the area detecting capacitors to equal the 2.5 volts applied on the V_{REF} pin. Determination of the burst difference (A-B) is accomplished under control of the CARR pin, by inverting the phase of the carrier input to the second area detector, while the burst B is being detected. The inversion is performed by an XOR gate. Accordingly (A-B) is bipolar relative to V_{REF} , while (A+B) is unipolar.

MULTIPLEXER AMPLIFIER

The multiplexer amplifier drives the HYBOUT pin and allows sequential interrogation of the (A-B) and (A+B) measurements, the results of which are stored on the external integrating capacitors. The amplifier is implemented as two independently selectable input stages, driving a common output structure, to form a voltage follower. To minimize the droop of the (A-B) and the (A+B) measurements, both input stages are biased off during periods when neither measurement is required to be routed to the HYBOUT pin. The SEL and CARR pins govern multiplexer channel selection through a decoding network.

Figure 1 shows a typical hybrid servo system application diagram for the ML4534 and also illustrates waveforms characteristic of a hybrid demodulator in a typical application.

OPERATIONAL AMPLIFIER USED FOR MOTOR CURRENT SENSE

This general purpose operational amplifier is intended for use as a differential to single-ended convertor and level shift stage. It performs voice coil motor current sensing by monitoring the voltage developed differentially across current sense resistors, on the ground side of the voice coil power driver bridge.

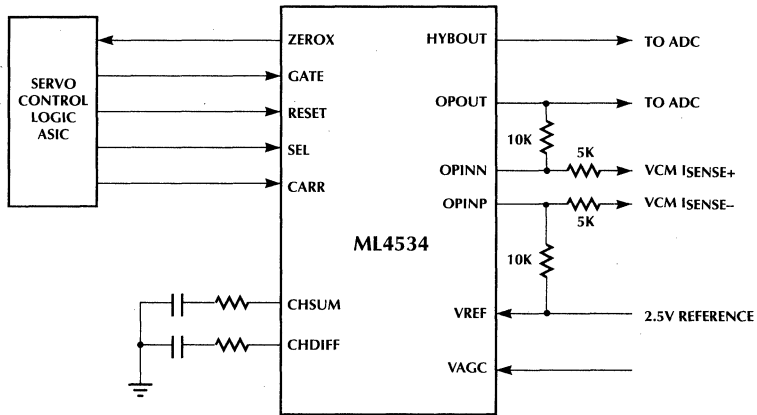


Figure 1. A Typical Servo System Application with the ML4534.

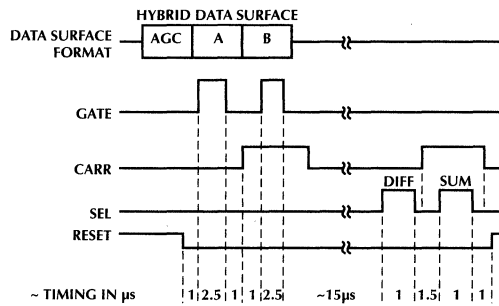


Figure 2. Illustrative Waveforms

ORDERING INFORMATION

PART NUMBER	TEMPERATURE RANGE	PACKAGE
ML4534CQ	0°C to +70°C	20-Pin Molded PLCC (Q20)

Area Detection Based Hybrid Servo Demodulator

GENERAL DESCRIPTION

The ML4535 is a bipolar monolithic hybrid servo circuit that provides area measurement demodulation of both the continuous servo surface (dedicated servo) and the sectored servo data (embedded servo) information in a high performance "hybrid servo" based disk drive. It operates on a single +5V supply and is intended to interface to a moderate speed, successive approximation ADC, with multiplexed inputs and sample and holds, like the ML2377 family.

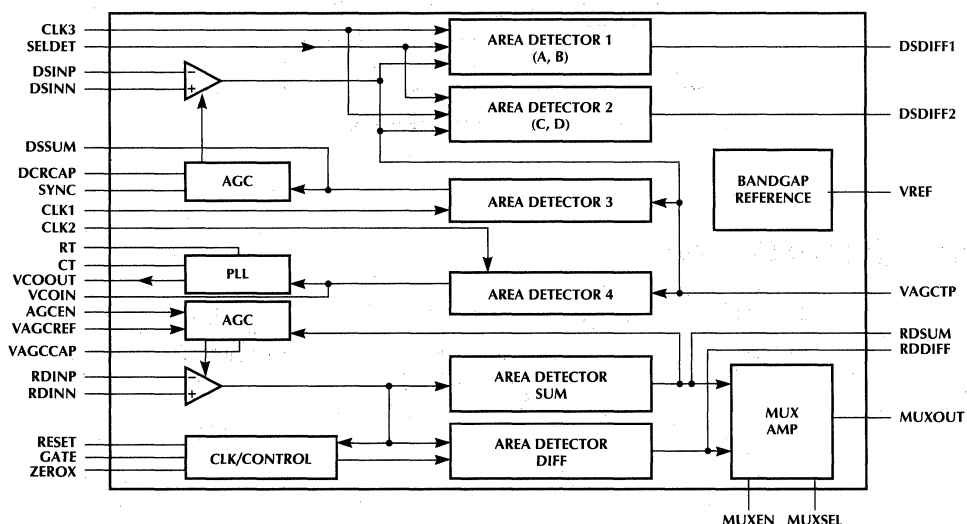
The area detectors are designed to minimize the pipeline transport delay while accurately quantizing the area of servo bursts in high speed servo systems. The data surface (embedded) servo demodulator section of the ML4535 consists of Sum and Difference area detectors along with an AGC control loop, so that the amplitude control function is self contained on the chip. The continuous (dedicated) servo demodulator section of the ML4535 consists of a variable gain amplifier, variable frequency oscillator and four synchronous detectors.

The ML4535 provides a high level of integration for designing the complex Hybrid Servo systems becoming popular in disk drives requiring very high bit and track densities.

FEATURES

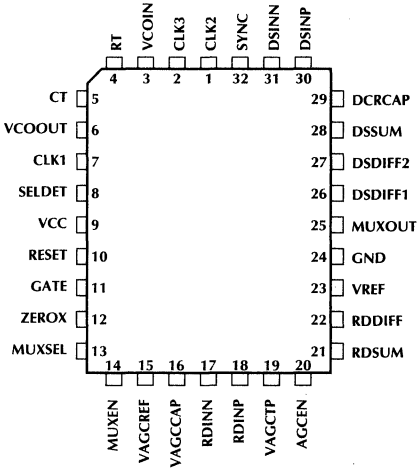
- Allows for area detection of back-to-back bursts
- 2% non-linearity over the input signal range
- Single +5 volt operation
- Internal 2.5V bandgap reference with reference output
- Separate AGC control loop for data surface and servo surface demodulator sections.
- Data surface amplitude control self contained on chip
- Data surface demodulator has muxed/selectable (A-B) and (A+B) outputs.
- Four synchronous area detectors onboard for implementing the continuous servo demodulator.
- Available in 32-pin PLCC package
- Future availability in 32-pin TQFP package.

BLOCK DIAGRAM



PIN CONNECTION

ML4535
32-Pin PLCC



ML4535

PIN DESCRIPTION

PIN#	NAME	FUNCTION	PIN#	NAME	FUNCTION												
29	DCRCAP	DC Restore capacitor input	13	MUXSEL	This pin in conjunction with the MUXEN pin governs the multiplexer channel selection as follows: <table border="1"> <thead> <tr> <th>MUXSEL</th> <th>MUXEN</th> <th>MUX CHANNEL</th> </tr> </thead> <tbody> <tr> <td>0</td> <td>X</td> <td>VREF</td> </tr> <tr> <td>1</td> <td>0</td> <td>Difference (A-B)</td> </tr> <tr> <td>1</td> <td>1</td> <td>Sum (A+B)</td> </tr> </tbody> </table>	MUXSEL	MUXEN	MUX CHANNEL	0	X	VREF	1	0	Difference (A-B)	1	1	Sum (A+B)
MUXSEL	MUXEN	MUX CHANNEL															
0	X	VREF															
1	0	Difference (A-B)															
1	1	Sum (A+B)															
30	DSINP	Differential input to AGC from															
31	DSINN	Continuous (dedicated) servo surface. Inputs must be AC coupled															
8	SELDET	Select signal for synchronous detectors 1 or 2															
2	CLK3	Clock for Area Detectors 1 & 2	14	MUXEN	Asserting this pin inverts the carrier input of the Difference (A-B) area detector. MUXEN should be asserted throughout the B burst of the A/B burst pair. While this pin is asserted, the Difference (A-B) area detector integrates the B burst in a direction opposite to that in which A is integrated, thus realizing the (A-B) operation. Carrier polarity in the (A+B) area detector is not affected by the state of the MUXEN pin. This pin along with the MUXSEL pin, also selects the multiplexer output.												
7	CLK1	Clock for Area Detector 3															
1	CLK2	Clock for Area Detector 4															
4	RT	Pins to connect a resistor/capacitor network for setting the center frequency of the internal VCO; R from RT to CT, C from CT to V _{CC}															
5	CT																
6	VCOOUT	VCO clock output															
3	VCOIN	Pin for connecting the loop filter for the PLL															
32	SYNC	Servo frame sync signal output															
28	DSSUM	Pin for connecting the filter for the AGC loop	12	ZEROX	This is the output of the carrier comparator, nominally a square wave having transitions coinciding with zero crossings of the VGA output.												
27	DSDIFF2	Area Detector #2 (C,D) or the Q output															
23	VREF	2.5V Bandgap reference output															
26	DSDIFF1	Area Detector #1 (A,B) or the N output	11	GATE	Asserting this pin defines the (A+B) and (A-B) area detect windows, to measure the area under the curve of the VGA output. This signal is resynchronized to the area detector carrier internally before application to the area detectors.												
19	VAGCTP	Test point connected through an isolation resistor to the output of the VGA															
16	VAGCCAP	AGC Loop Filter/Hold Capacitor	25	MUXOUT	Output of the multiplexer with (A-B) or (A+B) output												
20	AGCEN	AGC enable pin, defines area detect window	21	RDSUM	The (A+B) area detector integrating capacitor is connected here.												
15	VAGCREF	AGC voltage reference	22	RDDIFF	The (A-B) area detector integrating capacitor is connected here.												
18	RDINP	Differential input to VGA from Data surface (embedded servo). Inputs must be AC coupled.	24	GND	Ground pin												
17	RDINN			9	VCC	+ 5V supply											
10	RESET	Asserting this input pin resets the Area Detector (DIFF) to VREF (+2.5V) and the Area Detector (SUM) to VREF/2															

ABSOLUTE MAXIMUM RATINGS

DC Supply Voltage (VCC)	-0.3 to +7 VDC
Package Dimension, TA = 25°C (board mount)	TBD mW
Package Lead Temperature	
Soldering (10 sec)	260°C
Vapor Phase (60 sec)	215°C
Infrared (15 sec)	220°C
Storage Temperature (tstg)	-65 to +150°C

OPERATING CONDITIONS

DC Supply Voltage (VCC)	5+/-10% VDC
Temperature Range	0 to +70°C
Operating Junction Temperature (Tj)	+25 to +125 °C

ELECTRICAL CHARACTERISTICS

PARAMETER	CONDITIONS	MIN	TYP.	MAX	UNITS
DC CHARACTERISTICS					
ICC Supply Current	VAGCCAP = 4.0V	60	80	110	mA
Bandgap Reference voltage, VREF		2.45	2.5	2.55	V
VIH	For CLK1, CLK2, CLK3 SELDET, GATE, MUXEN, MUXSEL, AGCEN			2.0	V
VIL	For CLK1, CLK2, CLK3 SELDET, GATE, MUXEN, MUXSEL, AGCEN	0.8			V
IIH	For CLK1, CLK2, CLK3 SELDET, GATE, MUXEN, MUXSEL, AGCEN	-40		+40	μA
IIL	For CLK1, CLK2, CLK3 SELDET, GATE, MUXEN, MUXSEL, AGCEN	-400		10	μA
AGC (for dedicated servo)					
DSINPDC, DSINNDC	open	2.4		2.6	V
AvAGC min Voltage gain from input to test point	DSSUM = 4.0V, Measure DSDIFF1 DSINP - DSINN = 0.5V			0.4	V/V
AvAGC max	DSSUM = 1.0V DSINP - DSINN = 10 mV	75			V/V
DCRHIGH (DCR CAP VOLTAGE)	DSSUM = 1.0V DSINP - DSINN = 7mV	3.0	3.3		V
DCRLOW (DCR CAP VOLTAGE)	DSSUM = 1.0V DSINN - DSINP = 7 mV	2.3	2.5	2.7	V
AREA DETECTOR 1 and 2					
DSDIFF HI 1	SELDET = VIH	3.2			V
DSDIFF LO 2	DSINP - DSINN = 0.1 DSSUM = 1V, CLK3=VIH			1.1	V
DSDIFF LO 1	CLK3 = VIL			1.1	V
DSDIFF HI 2		3.2			V

ML4535

ELECTRICAL CHARACTERISTICS

PARAMETER	CONDITIONS	MIN	TYP.	MAX	UNITS
AREA DETECTOR 1 and 2 (continued)					
DSDIFF HI R1	DSINN - DSINP = 0.1	3.2			V
DSDIFF LO R2	CLK3 = VIL			1.1	V
DSDIFF LO R1	DSINN - DSINP = 0.1			1.1	V
DSDIFF HI R2	CLK3 = VIH	3.2			V
DSDIFF UNSEL 1,2	DSINP - DSINN = 1.0V	1.8		2.6	V
DSDIFF OFF	DSINP - DSINN = 0V	1.8		2.6	V
AREA DETECTOR 3					
SYNC LO	CLK1 = VIH, IIL = 1.6 mA DSSUM = 1.0V DSINN - DSINP = 0.1		0.25	0.5	V
SYNC HI	CLK1 = VIH, IIH = -0.4 mA DSSUM = 1.0V DSINN - DSINP = 0.1	2.4			V
I DSSUM LO	DSINN - DSINP = 0V	60	68	200	μA
I DSSUM HI	CLK1 = VIH, DSSUM = 1 V DSINP - DSINN = 0.1	-10	-30	-100	μA
AREA DETECTOR 4, VCO					
I VCOIN HI	CLK2 = VIH, DSSUM = 1 V DSINP - DSINN = 0.1V	-200		-60	μA
I VCOIN LO	CLK2 = VIL	60		200	μA
VCO HI	CT = 4.0V, VCOIN = 4.0V	2.5	0		V
VCO LO	CT = 1.0V, VCOIN = 1.0V		0.25	0.5	V
RDINPDC, RDINNDC	open	2.3	2.5	2.7	V
VOH ZEROX	RDINP - RDINN = 2.0V IOH = -0.4 mA	2.4			V
VOL ZEROX	RDINP - RDINN = 2.0V IOL = 2.0 mA			0.5	V
AGC (for embedded servo)					
VAGCTP	RDINP - RDINN = 2.0 V VAGC = 1.0V	2.4		3.6	V
Av VGA MIN	RDINP - RDINN = 1.0 V VAGC = 4.0V			0.5	V/V
Av VGA MAX	RDINP - RDINN = 0.4 V VAGC = 1.0 V	2.0			V/V
VAGC BIAS	XAGC = 1.0 V	0		200	μA

ELECTRICAL CHARACTERISTICS

PARAMETER	CONDITIONS	MIN	TYP.	MAX	UNITS
RESET LOGIC					
I RESET SUM, DIFF	RESET = VIL	80		400	μ A
I OFF SUM, DIFF	RESET = VIH	-100		100	nA
V SUM	RESET = VIL	1		1.3	V
V DIFF	RESET = VIL	2.1		2.4	V
I SUM, DIFF UNBAL	GATE = VIH, CLOCK VAGC 1X, 1 V _{p-p} swing	-40		40	μ A
I DIFF UNBAL XOR	MUXSEL = VIH	-40		40	μ A
I PEAK SUM	RDINP - RDINN = 1.0 V MUXSEL = VIH	300	400	600	μ A
I PEAK DIFF P	RDINP - RDINN = 1.0 V MUXSEL = VIH, VAGC = 1V	300	400	600	μ A
I PEAK DIFF XOR	RDINP - RDINN = 1.0 V MUXEN = VIH, VAGC = 1.0V	-600	-400	-300	μ A
VOHSUM, VOHDIFF	RDINP - RDINN = 1.0V	3.55			V
VOL DIFF, VOLSUM	RDINP - RDINN = 1.0V MUXSEL = VIH	0		1.0	V
IGATE	GATE = VIH, CLOCK 1X	-100		100	nA
MUXAMPLIFIER					
VOS MUXSUM	MUXSEL = MUXEN = VIL	2.492		2.508	V
VOS MUXDIFF	MUXSEL = VIH, VDIFF = 2.5V, MUXEN = VIL	2.492		2.508	V
IBIASSUM DIFF	SUM = DIFF = 2.5V	0		300	nA
VOHMUX	SUM = 3.9 V	3.55			V
VOLMUX	DIFF = 0.95	0		1.0	V
INSINGMUX	VMUXOUT = VXX	1.5		2.5	mA
Amplifier settling time - t _{smux}	R _{out} = 604 ohms, C _{out} = 36 pF		0.4	1	μ sec
Ileakage				10	nA
Linearity	0 to 1V input with VAGC such that AvVAGC = 1.0V	-5		5	%F.S

FUNCTIONAL DESCRIPTION

The ML4535 provides area measurement demodulation of both the continuous (dedicated) servo surface and the sectored (embedded) servo data on each of the data surfaces of a "hybrid" servo disk drive. It operates on a single +5V supply and is intended to interface to a moderate speed, successive approximation ADC with multiplexed inputs and sample and holds, like the ML2377. In a conventional peak detection based servo scheme, the attack rate of the peak detectors are inherently faster than the decay, high crest factor noise sensitivity is high and rectification must have a very low offset for it to be functionally correct. On the other hand area detection has much better noise rejection and is more tolerant of small rectifier offsets. However it requires that the measurement period be an integer number of signal half cycles. Hence when the timing requirements are satisfied, area detection is certainly more accurate than peak detection schemes.

DATA SURFACE OR EMBEDDED SERVO DEMODULATOR SECTION

The data surface (embedded) servo demodulator section of the ML4535 consists of a standalone AGC control loop so that the amplitude control function is self contained on the chip and two area detectors providing the sum (A+B) and difference (A-B) outputs which are muxed out through a mux amplifier.

Input Amplifier and AGC

The input amplifier and AGC circuit operate with differential inputs in the range of 0.25V_{p-p} to 2V_{p-p}, from the read channel filter's lowpass outputs. The purpose of the AGC loop is to maintain a constant area detect value that correlates to the zero scale and full scale output values based upon the minimum and maximum burst value. The sensing for the AGC is at the output of the area detector, allowing signal ranging based on the area of the burst rather than the peak level of the burst. The AGC is intended to be updated at every sector of servo position bursts such that the signal variances due to the platter radius and differences in the read channel data frequencies can be corrected. In this closed-loop system, the area detected output voltage is fed back and compared with the VAGCCREF voltage to provide a gain control current for charging and discharging the VAGCCAP. The gain is varied to secure constant area of the output signal and provide amplitude control. The AGC gain value is held constant when the AGCEN is at logic low. When is is logic high, the level of gain can change up or down. The capacitor from VAGCCAP to ground holds the gain setting when AGCEN is at logic low and the area detector output does not affect the gain setting in this mode.

Zero X Detector

The output of the zero crossing detector (comparator) is provided for system synchronization. It detects zero crossings of the composite signal delivered by the Variable Gain Amplifier, VGA. The output of this comparator controls the synchronous rectification of the composite VGA output, in the area detectors. This signal is internally generated in ECL, but an internal ECL to TTL converter presents this output as a TTL level on the ZEROX pin. Control logic in the servo channel employs the ZEROX signal to produce an area detector enabling gate, which spans a fixed number of cycles of the composite signal.

Area Detectors (Sum & Difference)

The area detectors detect the A and B burst levels by area detection. Two area detectors — one to measure the sum of the A and B bursts (A+B), and a second one to measure the difference (A-B). Each area detector is implemented as a gated current — output synchronous rectifier, driving an external charge accumulating integrating capacitor. Area detection occurs only while the area detector is enabled under the control of the GATE pin. When the detector is disabled, the integrating capacitor is effectively floated. The on-chip D Flip-Flop resynchronizes the gating signal to remove any timing error due to logic delays in the external gate control logic. Initial conditions on the integrating capacitors are established prior to an area detecting operation by a reset circuit controlled by the RESET pin. A reset operation forces the voltage on the DIFF area detecting capacitor to equal the 2.5 volts applied on the VREF pin and the voltage on the sum area detecting capacitor to equal VREF/2. Determination of the burst difference (A-B), is accomplished under control of the MUXEN pin, by inverting the phase of the carrier input to the second area detector, while the burst B is being detected. The inversion is performed by an XOR gate. Accordingly (A-B) is bipolar relative to VREF, while (A+B) is unipolar.

Multiplexer Amplifier

The multiplexer amplifier drives the MUXOUT pin and allows sequential interrogation of the (A-B) and (A+B) measurements, the results of which are stored on the external integrating capacitors. The amplifier is implemented as two independently selectable input stages, driving a common output structure, to form a voltage follower. To minimize the droop of the (A-B) and the (A+B) measurements, both input stages are biased off during periods when neither measurement is required to be routed to the MUXOUT pin (MUXSEL = 0). The MUXSEL and MUXEN pins govern multiplexer channel selection through a decoding network.

CONTINUOUS OR DEDICATED SERVO DEMODULATOR SECTION

The continuous (dedicated) servo demodulator section of the ML4535 consists of its own variable gain amplifier and AGC loop, a variable frequency oscillator and four synchronous detectors.

The first synchronous detector (AREA DETECTOR #4) is used as a multiplying phase detector to control the variable frequency oscillator and complete the analog portion of the phase locked loop that recovers the clock.

The second synchronous detector (AREA DETECTOR #3) is used for measuring the area of the composite signal, to determine its amplitude for comparison with the on-chip reference of the AGC loop. An amplitude level comparator is also included on this detector's output to provide the logic level output for Frame sync and Index data.

The third and fourth synchronous detectors (AREA DETECTORS #1 & #2) are used to demodulate the normal and quadrature position signals. The normal and quadrature outputs are currents that have been terminated on chip with nominal 19K resistors to a 2.3 volt reference (0.9 x 2.5 volts).

Operational Example

An example continuous servo composite signal encoding and the associated demodulator clock waveforms are shown in figure 1. The VCO operates at twice the frequency of the fundamental of the composite signal and drives two flip flops that generate quadrature and normal phase references. The Clock generation logic circuit and synchronizing circuit for the PLL, to acquire initial lock with type 2 loop, are shown in figure 2. A state counter divides the servo frame into eight intervals which are:

NAME	LENGTH(IN CLOCK CYCLES)
S	2
X ₁ , X ₂ , X ₃	1
A, B, C, D	n where n is an integer like 8

Note that all peaks of the composite signal are on Quad clock phase boundaries, so it contains only one fundamental frequency, which is easily acquired by the phase locked loop. The sync character is 180° out of phase with all others; thus at phase alignment it causes no disturbances to the phase comparator but gives an easily recognizable reverse polarity ripple in the AGC which is easily detected with a level detector to provide a frame sync logic signal to initialize the state counter.

The A, B demodulator is selected from the center of S to the center of X₂ and the C, D demodulator is selected from the center of X₂ to the center of S. The S, X₁, X₂, X₃ information is in quadrature with the position clock and symmetrical so their contribution cancels to zero for the position output but not for AGC and phase compare. The NORMAL clock is used throughout A, B, C, D but inverted 180° as required by the track type to give A-B, C-D, B-A, D-C without additional analog switching in position 1 demodulator for track following.

Filters are needed on each of the four demodulators for removing the carrier ripple and providing frequency compensation for the gain control phase locked loop systems. The current output scaling of all four detectors

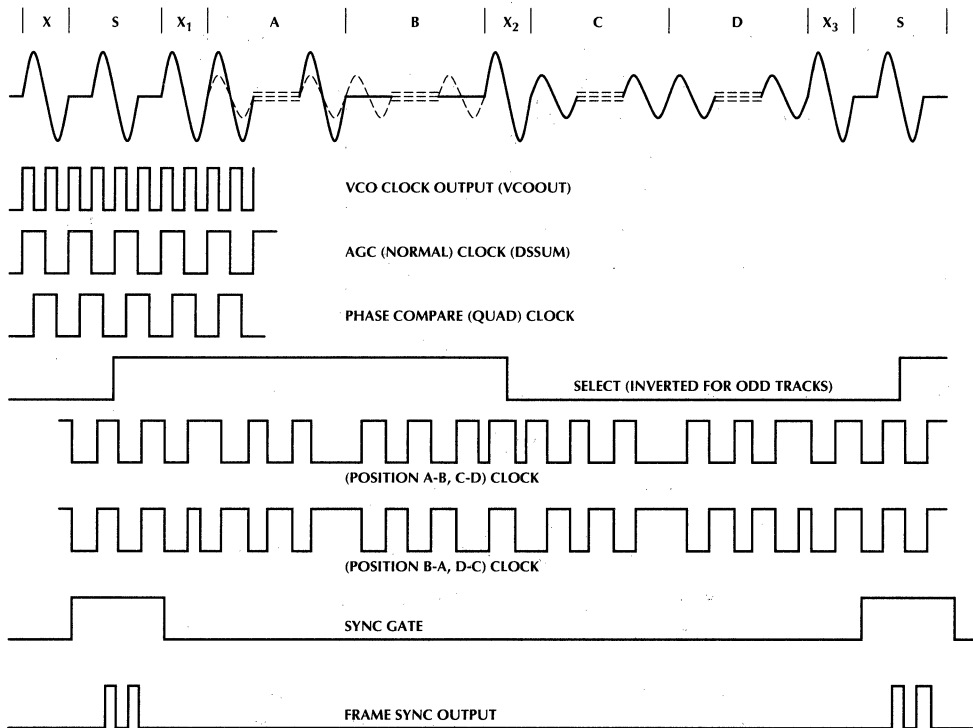


Figure 1. Continuous Servo Encoding and Demodulator Clock Waveforms

ML4535

are the same and set by on-chip resistors. The AGC setpoint is also a current fixed by a tracking on-chip resistors. Thus a capacitor to ground provides an integrating response for the AGC control loop, as well as ripple filtering. For the phase comparator filter, two capacitors and a resistor provide D.C. integration plus a lead-lag for the PLL control loop compensation. For the position outputs, on-chip 19K nominal resistors to $(0.9 \times V_{ref})$ are provided so a capacitor to ground is added to form a low pass ripple filter.

The phase compare detector forms phase only (not phase frequency) characteristic, so the loop will not acquire initial lock with and integrating loop filter, which is needed to assure no steady state phase error. The synchronizing circuit suggested (refer figure 2), senses when the VCO control voltage is near either rail and applies a pulse that ramps it toward the other rail and thus through the operational frequency where it locks. The lock range is much greater than the acquire range, so it retains lock in the presence of the resistor-coupled pulse, with a small phase error, until the pulse goes away and the phase error becomes zero.

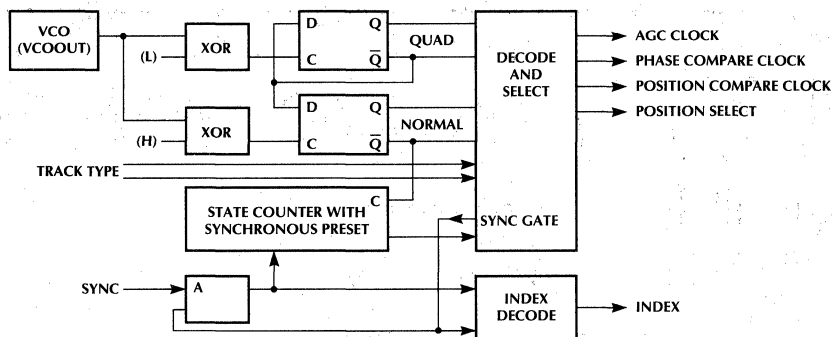
HYBRID SERVO VERSUS ONLY EMBEDDED SERVO

There are a number of merits in using a hybrid servo scheme consisting of a servo surface plus limited data head servo samples over the completely embedded or data head sector servo samples scheme. These are summarized below:

Cost: Lost data surface capacity is less than 1% for a servo surface plus samples compared to 8% for samples only. This suggests that with six or more platters, the servo surface has an advantage. Without a servo surface, it is difficult to generate accurately phase and track center aligned data head (embedded) servo sectors in the drive, requiring them to be done with extra time on an expensive servo writer and moving the cost crossover point nearer to four platters. Drive hardware and costs including assembly and test favour DSP implementations in either case.

Effect on position error sources: These can be very similar for both configurations with optimized control algorithms. The servo surface does have some advantages in being able to obtain higher bandwidths and thus faster settling time and greater reduction of non-repetitive run-out and random disturbances.

CLOCK GENERATION LOGIC



SYNCHRONIZING CIRCUIT FOR THE PLL (To acquire initial lock with type 2 loop)

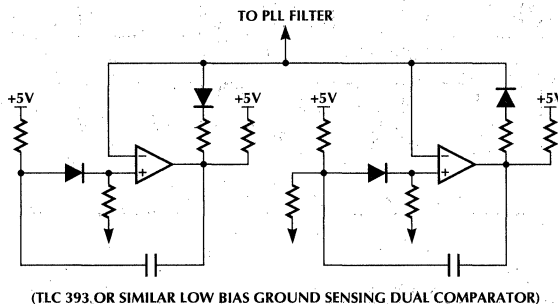


Figure 2. Support Circuitry for the ML4535 Based "Hybrid" Servo Subsystem

Effect on access time: A system with a servo surface has two advantages here. The ability to adjust the control signal at shorter time intervals and a higher small signal bandwidth, both of which reduce settling time. Move times can be equivalent.

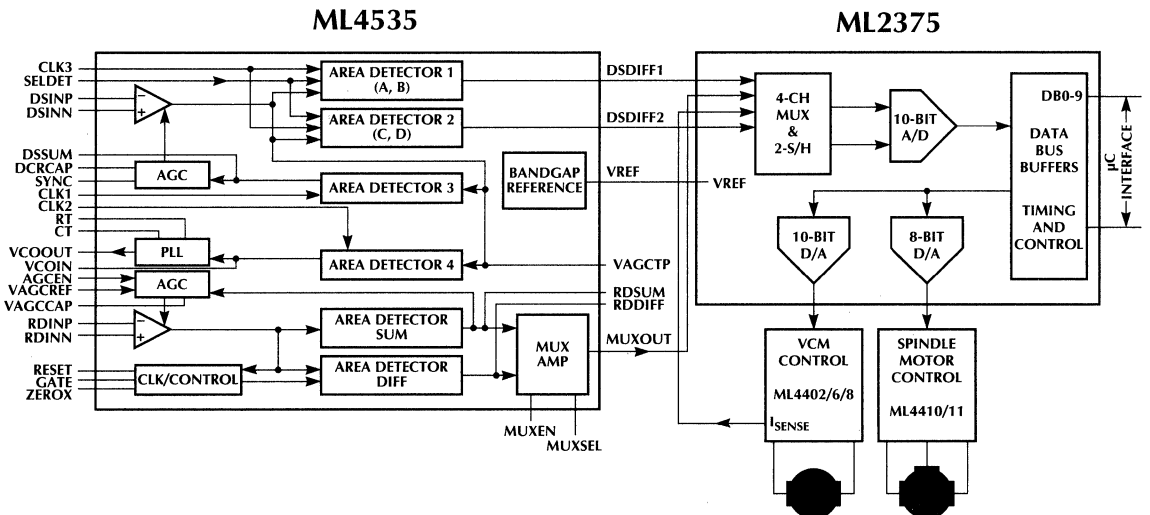
Data integrity: Here there is a clear superiority for a servo surface system in preventing writes which destroy existing data. There are at least two ways in which this can happen. Electronic noise in the sector timing causes servo sectors to be over written, so that the head can no longer be positioned to read the track even if the data is intact. This probability can be made acceptably small by redundancy in the electronics. External mechanical shock while writing a data sector can not only cause improper writes of the new data but also overwrite adjacent tracks. Inherently there is no way to prevent this with servo sectors only, as there is no position data measurement available and estimators do no good for random fast disturbances. Dynamically balanced rotary actuators reduce this exposure compared to linear travel positioners but cannot eliminate it completely.

SERVO DESIGN SUGGESTIONS FOR A HIGH TRACK DENSITY DISK DRIVE

The best design choices for a high track density disk drive with four or more platters are outlined below and the hybrid servo subsystem based on the ML4535 & ML2377, provides the most optimum solution for implementing these design choices and making track densities of 3000 TPI easily achievable.

- 1) Continuous servo surface with quadrature signals plus some position samples and/or calibration tracks for each data head.
- 2) DSP implementation of the position control system for best performance of state estimators and adaptive parameter adjustment.
- 3) Area integration position demodulation on both the servo (dedicated) surface and data head position (embedded) servo samples, for best accuracy and noise rejection.
- 4) Thin film heads with gap edges aligned and perpendicular to the disk surface.
- 5) Dynamically balanced rotary actuator for best rejection of external mechanical shock.

APPLICATION DIAGRAM



DSP Based Hybrid Servo Subsystem.

ML4535

ORDERING INFORMATION

PART NUMBER	TEMPERATURE RANGE	PACKAGE
ML4535CQ	0°C to +70°C	32-Pin PLCC (P32)
ML4535CH	0°C to +70°C	32-Pin TQFP (H32)

Disk Pulse Detector + Embedded Servo Detector

GENERAL DESCRIPTION

The ML4568 is a hard disk pulse detector with two gated peak detectors to demodulate embedded servo information. The pulse detector section includes a wide bandwidth differential amplifier with automatic gain control (AGC), a precision full wave rectifier, time channel and gate channel. The embedded servo peak detector section includes a full-wave rectifier, two gated peak detectors, buffered peak detector outputs, and a difference output. A 2.25V bandgap reference is also included on-chip.

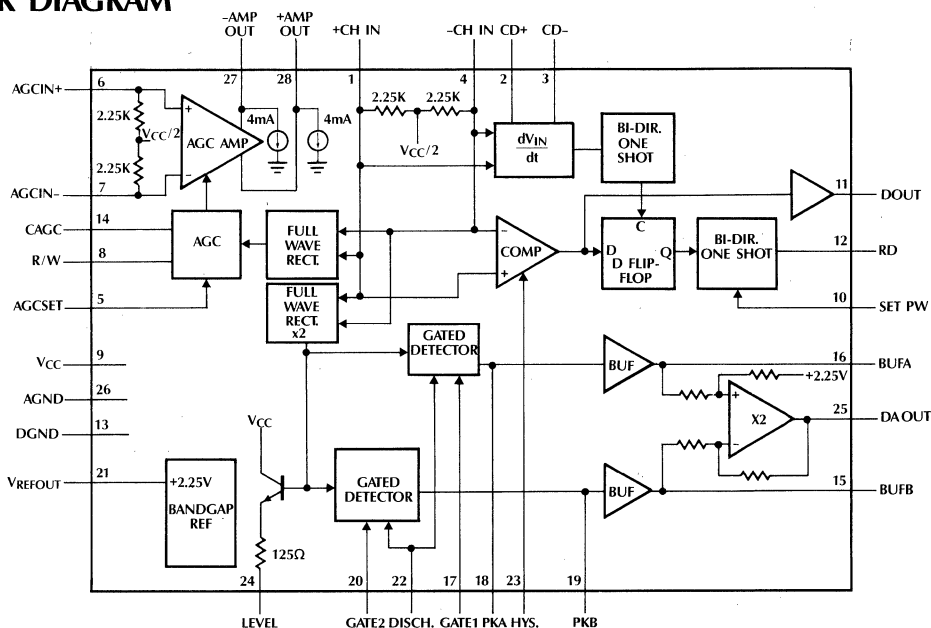
The ML4568 is a 5V-only upgrade for 8468-type devices. Upgraded features include increased data rate operation (to 24 MB/s with RLL(1, 7) coding), improve pulse pairing (1ns), and reduced power consumption (400mW typical) resulting from 5V-only operation.

The ML4568 pulse detector section detects amplitude peaks, producing a TTL-compatible output which accurately indicates the time position of signal peaks. In hard disk applications, these signal peaks represent flux reversals in the magnetic medium.

FEATURES

- 5V-only operation
- Low power consumption (400mW typical)
- Supports 24 MB/s RLL(1, 7) coding
- Less than ± 1 ns Pulse Pairing
- Wide input signal amplitude range (10mV_{PP} to 100mV_{PP})
- On-chip differential gain controlled amplifier, differentiator, comparator gating circuitry, and output pulse generator
- Adjustable comparator hysteresis
- Dynamic hysteresis tracks signal amplitude
- AGC and differentiator time constants set by external components
- TTL compatible digital inputs and outputs
- Built in embedded servo detector
- On chip buffers provide low impedance servo output voltages
- User adjustable servo time constants

BLOCK DIAGRAM

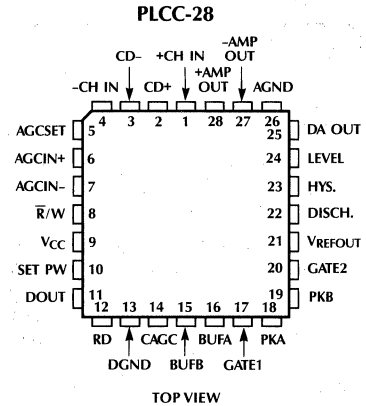


ML4568

GENERAL DESCRIPTION (Continued)

The ML4568 also incorporates two gated detectors which detect embedded servo information, used for head positioning. The ML4568 provides two buffered low impedance voltage outputs which represent the peak detected level of each servo burst. The ML4568 also provides a buffered output that represents the voltage difference between the two servo channels, centered about V_{REF} .

PIN CONFIGURATION



PIN DESCRIPTION

PIN #	NAME	FUNCTION
Power Supply		
9	V_{CC}	+5V \pm 10% supply.
21	$V_{REF OUT}$	Internal 2.25 V reference voltage output.
26	ANALOG GROUND	Analog signals should be referenced to this pin.
13	DIGITAL GROUND	Digital signals should be referenced to this pin.

Analog Signals

6	AMP IN+	These are the differential inputs to the Amplifier. The output of the read/write head amplifier should be capacitively coupled to these pins.
7	AMP IN-	
28	AMP OUT+	These are the differential outputs of the Amplifier. These outputs should be capacitively coupled to the channel filter.
27	AMP OUT-	
4	-CH IN	These are the differential inputs to the time, gating and servo channels. These inputs must be capacitively coupled to the channel filter at the amp. outputs. The maximum differential peak-to-peak swing at this input is 1.5 V_{p-p} .
1	+CH IN	
2	CD+	The external differentiator network is connected between these two pins.
3	CD-	
23	HYS.	The DC voltage on this pin sets the amount of hysteresis on the differential comparator.

PIN #	NAME	FUNCTION
Analog Signals (Continued)		
24	LEVEL	This is a Peak Detector Output signal that is used in conjunction with the set hysteresis pin 23 to provide a dynamic hysteresis function.
5	AGCSET	The AGC circuit adjusts the gain of the gain controlled amplifier to make the differential peak to peak voltage at the Channel inputs equal to four times the DC voltage on this pin. $V_{AGCSET} = \frac{1}{2}V_{CC} + \frac{1}{4}V_{PP}$ where V_{PP} is the peak-peak differential voltage on the channel input.
14	CAGC	The external capacitor for the AGC is connected between this pin and Analog Ground.
18	PKA	The peak detected servo signal voltage appears across the RC networks connected from these pins to analog ground.
19	PKB	
16	BUFA	These low impedance pins, output the DC level at pins 18 and 19 respectively, level shifted down by two diode drops.
15	BUFB	
25	DA OUT	This low impedance pin outputs the difference in voltage between pins 16 and 15 about a zero level set by the voltage on pin 21.

PIN DESCRIPTION (Continued)

PIN #	NAME	FUNCTION
Digital Signals		
10	SET PW	An external capacitor to control the pulse width of the Encoded Data Out (RD) is connected between this pin and Digital Ground. See Figure 1.
8	$\overline{R/W}$	If this pin is low, the Pulse Detector is in the read mode and the chip is active. When this pin goes high, the pulse detector is forced into a stand-by mode. This is a standard TTL input.
11	DOUT	This is the buffered, open collector, output of the differential comparator with hysteresis.
12	RD	This is the standard TTL output whose leading edge indicates the time position of the peaks.

PIN #	NAME	FUNCTION
Digital Signals (Continued)		
17	GATE 1	These inputs accept TTL levels. When a low level is present the embedded servo signal is allowed to charge the RC network at pins 18 and 19 respectively. A high level will force a hold condition of the DC voltage across the RC network and will also disable the servo channel.
20	GATE 2	
22	DISCH.	This input accepts a TTL level. A high level connects a 1.5K internal resistor to ground on pins 18 and 19.

ABSOLUTE MAXIMUM RATINGS

Supply Voltage	
Pin 9	14V
TTL Input Voltage	
Pins 8, 17, 20, 22	5.5V
TTL Output Voltage	
Pins 12, 11	5.5V
Input Voltage	
Pins 23, 5	5.5V
Minimum Input Voltage	
Pins 23, 5	-0.5V
Differential Input Voltage	
Pins 6-7, 4-1	3V or -3V
ESD susceptibility rating is to be determined	
Storage Temperature	-65°C to +150°C
Lead Temperature (Soldering 10 sec.)	300°C
Maximum Power Dissipation at 25°C:	
PLCC Package (derate TBD mW/°C above 25°C)	500mW

OPERATING CONDITIONS

V _{CC}	4.5V to 5.5V
Ambient Temperature, T _A	0°C to +70°C

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ELECTRICAL CHARACTERISTICS

Over recommended operating conditions.

Set Hysteresis = 0V, $V_{PIN\ 17} = 2V$, READ/WRITE = 0.4V, $V_{PIN\ 22} = 0.4V$, unless otherwise noted.

Symbol	Pins	Parameter	Conditions	Min	Typ	Max	Units
Amplifier							
Z_{INAI}	6, 7	Amp In Impedance (Note 1)	$T_A = 25^\circ C$	1.8	2.4	3.0	K Ω
A_{VMIN}	28, 27	Minimum Voltage Gain Differential	AC Output 3 V_{PP}		6	15	V/V
A_{VMAX}	28, 27	Maximum Voltage Gain Differential	AC Output 3 V_{PP}	250	300		V/V
Channel							
Z_{INCI}	4, 1	Channel Input Impedance	$T_A = 25^\circ C$ (Note 1)		2.5		K Ω
I_{CAGC^-}	14	Pin 14 Current which Charges C_{AGC}	$V_{PIN\ 14} = 2.2V$	5.0	5.8		mA
I_{CAGC^+}	14	Pin 14 Current which Discharges C_{AGC}	$V_{PIN\ 14} = 2.2V$		0.5	2	μA
I_{AGCSET}	5	AGCSET Input Bias Current			8	100	μA
I_{IL}	23	Set Hysteresis Input Bias Current	$V_{PIN\ 23} = 0$			-20	μA
I_{CD}	2, 3	Current into Pin 2 and 3 that Discharges C_D		0.8	1.0		mA
HYS	23	Peak Hys. vs V_{HYS}	$V_{PIN\ 23} = 1V$	0.25	0.4	0.55	V_{PK}/V_{DC}
Write Mode							
Z_{INAI}	6, 7	Amp In Impedance in Write Mode	$V_{PIN\ 8} = 2.0V$		350	450	Ω
I_{AGC^-}	14	Pin 14 Current in Write Mode	$V_{PIN\ 8} = 2.0V, V_{PIN\ 14} = 2.2V$		0.2	1.0	μA
Digital Pins							
V_{IH}	8, 17, 20, 22	High Level Input Voltage		2			V
V_{IL}	8, 17, 20, 22	Low Level Input Voltage				0.8	V
I_{IH}	8, 17, 20, 22	High Level Input Current	$V_{5V} = \text{Max}, V_I = 2.7V$			20	μA
I_{IL}	8, 17, 20, 22	Low Level Input Current	$V_{5V} = \text{Max}, V_I = 0.5V$		140	200	μA
V_{OH}	12	High Level Output Voltage	$V_{5V} = \text{Min}, I_{OH} = -400\mu A$ (Note 2)	2.4			V
V_{OL12}	12	Low Level Output Voltage	$V_{5V} = \text{Min}, I_{OL} = 800\mu A$ (Note 2)			0.5	V
I_{LH}	11	High Level Output Leakage Current	$V_{PIN\ 11} = V_{CC}$ Measure Current into Pin 11			50	μA
V_{OL11}	11	Low Level Output Voltage	$I_{PIN\ 11} = 800\mu A$			0.5	V
Servo Channel							
Z_{DIS}	18, 19	Discharge Impedance	$V_{PIN\ 22} = 2V$ (discharge) Force 2.5V on Pins 18 or 19	0.5	1.8	2.5	K Ω
V_{BOQ}	15, 16	Buffer Quiescent Output Level	$V_{PIN\ 17, 20, 22} = 0.4V, V_{CI} = 0V$ Pull 0mA from Pins 15 and 16	1.0	1.6	2.0	V
$V_{LEVEL\ Q}$	24	Level Quiescent Output Level	$V_{CI} = 0V$ Pull 200 μA from Pin 24		0.2	0.5	V
I_L	18, 19	Gated Off Leakage Current	$V_{PIN\ 22} = 0.4V, V_{PIN\ 20} = V_{PIN\ 17} = 2V$ Force 3V on Pin 18 or Pin 19	-1		1	μA

ELECTRICAL CHARACTERISTICS (Continued)

Over recommended operating conditions.

Set Hysteresis = 0V, $V_{PIN\ 17} = 2V$, READ/WRITE = 0.4V, $V_{PIN\ 22} = 0.4V$, unless otherwise noted.

Symbol	Pins	Parameter	Conditions	Min	Typ	Max	Units
Servo Channel (Continued)							
V_{OSBO}	16, 15	Buffer Output Offset Voltage for $V_{CI} = 1V_{PK-PK}$	$V_{PIN\ 17, 20, 22} = 0.4V$, $V_{PIN\ 1} = 2.75V$ Pull 0mA from Pins 15 and 16 $V_{PIN\ 4} = 2.25V$, $V_{OSBO} = V_{PIN\ 16} - V_{PIN\ 15}$		2	± 15	mV
V_{OSYS}	25, 21	System Output Offset Voltage for $V_{CI} = 0.75V_{PK-PK}$ Pull 0mA from Pin 25 $V_{OSYS} = V_{PIN\ 25} - V_{PIN\ 21}$	$V_{PIN\ 17, 20, 22} = 0.4$ $V_{PIN\ 1} = 2.688V$, $V_{PIN\ 4} = 2.313V$		± 5	± 20	mV
$AV_{DA} (1V)$	25, 21	Difference Amplifier Gain, 1V Differential Input	$V_{PIN\ 17, 20} = 2V$ $V_{PIN\ 19} = 1.5V$, $V_{PIN\ 18} = 2.5V$, $V_{PIN\ 22} = 0.4V$	1.6	2	2.4	V/V
$AV_{DA} (.5V)$	25, 21	Difference Amplifier Gain, 0.5V Differential Input	$V_{PIN\ 17, 20} = 2V$ $V_{PIN\ 19} = 1.75V$, $V_{PIN\ 18} = 2.25V$, $V_{PIN\ 22} = 0.4V$	1.6	2	2.4	V/V
GL_{DA}	25	Difference Amplifier Gain Linearity			0.2	2.5	%
Z_{LEVEL} SOURCE	24	Level Out Output Impedance	$V_{PIN\ 17, 20, 22} = 0.4V$, $V_{CI} = 0.75V$ Measure $V_{PIN\ 24}$ with 200 μ A and 3mA pulled out of the pin. $Z_{LEVEL} = \text{change in } V_{PIN\ 24} \text{ SOURCE}$ 3mA - 0.2mA	100	180	250	Ω
AV_{GD} (1.5V)	15, 16	Gated Detector Gain for $V_{CI} = 1.5V_{PK-PK}$	$V_{PIN\ 22, 20, 17} = 0.4V$ $V_{PIN\ 1} = 2.875V$, $V_{PIN\ 4} = 2.125V$	1.45	1.8	2.25	V/V
AV_{GD} (0.75V)	15, 16	Gated Detector Gain for $V_{CI} = 0.75V_{PK-PK}$	$V_{PIN\ 22, 20, 17} = 0.4V$ $V_{PIN\ 1} = 2.688V$, $V_{PIN\ 4} = 2.313V$	1.45	1.7	2.25	V/V
AV_{LEVEL} (1.5V)	24	Level Voltage Gain For $V_{CI} = 1.5V_{PK-PK}$	$V_{PIN\ 1} = 2.875V$, $V_{PIN\ 4} = 2.125V$	1.45	1.8	2.25	V/V
AV_{LEVEL} (0.75V)	24	Level Voltage Gain For $V_{CI} = 0.75V_{PK-PK}$	$V_{PIN\ 1} = 2.687V$, $V_{PIN\ 4} = 2.312V$	1.6	1.9	2.4	V/V
GL_{GD}	15, 16	Gated Detector Gain Linearity			± 0.1	± 2.5	%
I_{CC}	9	V_{CC} Supply Current	$V_{CC} = \text{Max}$	40	90	110	mA
V_{REF}	21	V_{REF} Voltage		2.0	2.25	2.5	V

ML4568

AC ELECTRICAL CHARACTERISTICS

Over Recommended Operating Temperature and Supply Range refer to AC Test Setup.
 $f = 2.5\text{MHz}$ unless otherwise indicated. PKA, PKB = $1\text{K}\Omega + 10\text{nF}$ to GND.

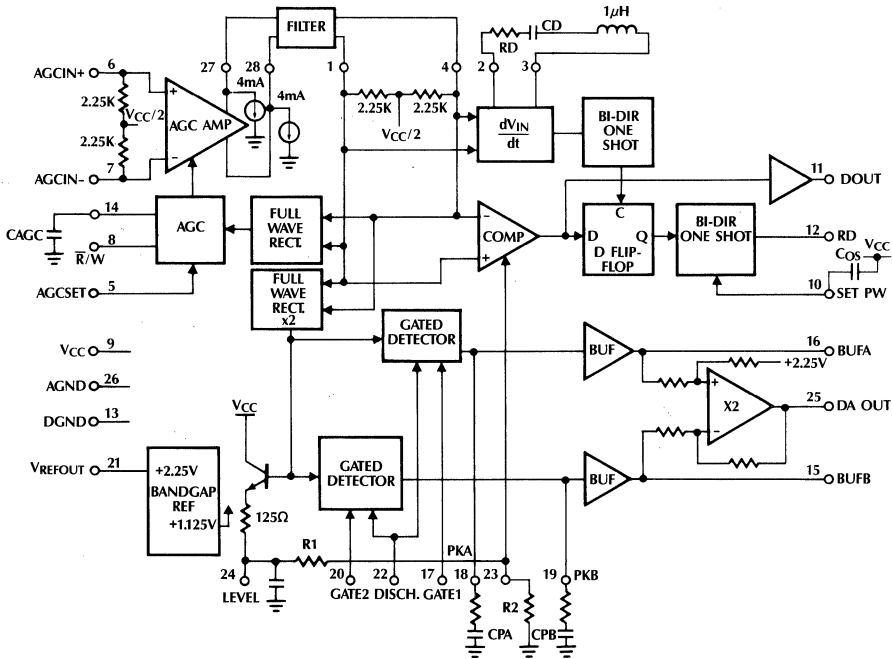
Symbol	Pins	Parameter	Conditions	Min	Typ	Max	Units
t_{CHARGE}	15, 16	Gated Detector Charge Time	$V_{\text{CI}} = 1.5V_{\text{PP}}$, $V_{\text{PIN 22}} = 0.3V$, With PKA and PKB discharged, measure the time from Pin 17 or 20 going from 2V to 0.3V, to V_{BO1} or V_{BO2} respectively, reaching 90% of their final value		1.0		μs
$t_{\text{DISCHARGE}}$	15, 16	Gated Detector Discharge Time	$V_{\text{CI}} = 1.5V_{\text{PP}}$. With LP1 charged, measure the time from Pin 22 going from 0.3V to 2V, to the voltage at V_{BO1} or V_{BO2} reaching 90% of their final value		70		μs
t_{ON}	18, 19	Gated Detector Turn ON Time	$V_{\text{CI}} = 0.35V_{\text{DC}}$, $V_{\text{PIN 22}} = 0.3V$. With LP1 discharged, measure the time from Pin 17 going from 2V to 0.3V, to the voltage on Pin 18 increasing 0.1V. Do a similar measurement with LP2, Pin 20 and Pin 19		0.2		μs
t_{OFF}	18, 19	Gated Detector Turn OFF Time	$V_{\text{CI}} = 0.35V_{\text{DC}}$, $V_{\text{PIN 22}} = 2V$. Measure the time from Pin 17 going from 0.3V to 2V, to the voltage on Pin 18 decreasing by 0.1V. Do a similar measurement with Pins 20 and 19		0.4		μs
t_{PP}	12	Pulse Pairing ML4568-1	$f = 2.5\text{MHz}$ and $V_{\text{CI}} = 1V_{\text{PP}}$ differential			± 1	ns
t_{PP}	12	Pulse Pairing ML4568-2	$f = 2.5\text{MHz}$ and $V_{\text{CI}} = 1V_{\text{PP}}$ differential			± 3	ns

Notes:

- The temperature coefficient of the input impedance is typically 0.05% per °C.
- To prevent inductive coupling from the digital outputs to Amp In, the TTL outputs should not drive more than one ALS TTL load each. Pin 11 is an open collector output which is tested with an external 1K pullup resistor to the 5V supply.

ML4568 CONNECTION DIAGRAM

PLCC-28 Version



Note 1: $K = R1/R2$

Note 2: Hysteresis Level = $0.6 \times K/V_{IN P-P}$

Note 3: RC on pins PKA and PKB basically tuned to minimize ripple.

ML4568

APPLICATION INFORMATION

SETTING THE OUTPUT PULSEWIDTH

The RD output pulsewidth is dependent on the value of C_{OS} , which is connected from pin 10 to V_{CC} . This relationship is shown in figure 1.

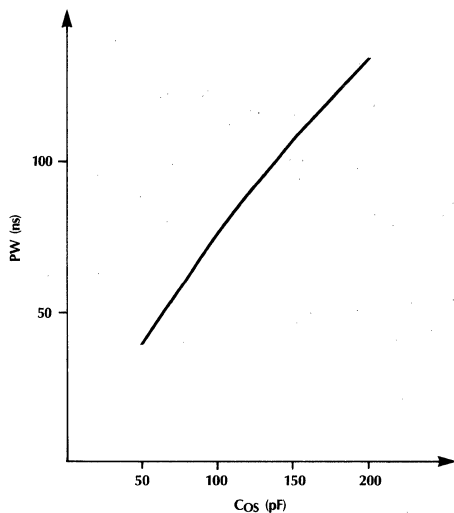


Figure 1. RD Output Pulsewidth as a Function of C_{OS}
 $PW \approx 0.5C_{OS}$

SELECTING C_D

The following table summarizes the maximum C_D value allowed for different data rates. These values are derived using

$$C_D (\text{max}) = \frac{176}{f_{\text{MAX}}}, R_D = 0$$

Data Rate	f_{MAX}	C_D (max)
75 MB/s	2.81 MHz	62.6 pF
24 MB/s	9 MHz	19.6 pF

Table 1. Maximum C_D Value Allowed for a 1.5 V_{p-p} Differential Signal Using RLL (1, 7) Code

	1, 7 RLL	2, 7 RLL
f_{MAX}	3/8 x Data Rate	1/3 x Data Rate
f_{MAX}	3/32 x Data Rate	1/8 x Data Rate

Table 2.

ORDERING INFORMATION

PART NUMBER	PACKAGE	PULSE PAIRING
ML4568-1CQ	MOLDED PCC (Q28)	±1 ns
ML4568-2CQ	MOLDED PCC (Q28)	±3 ns

ML4610R, ML4611R

5V, 2-, 4-Channel Thin-Film Read/Write Circuit

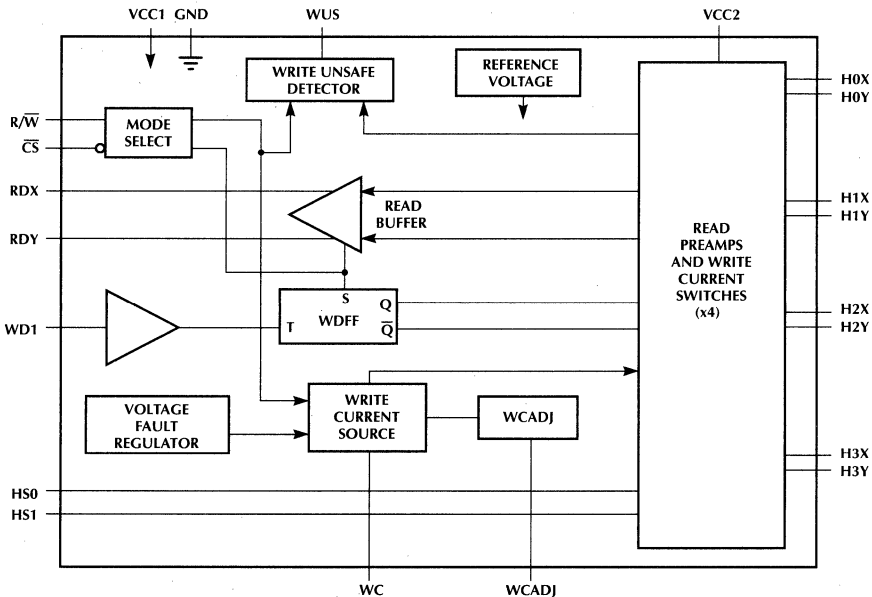
GENERAL DESCRIPTION

The ML4610R/4611R is a bipolar monolithic read/write circuit designed for use with two-terminal thin-film recording heads. They provide a low noise read amplifier, write current control, and data protection circuitry for up to four channels. The ML4610R/4611R incorporates internal 700 ohm damping resistors which dampen the write signals to the disk. When the device is switched to read mode, the damping resistor is switched out to allow the full signal to be amplified. Power supply fault protection is provided by disabling the write current generator during power sequencing. System write to read recovery time is significantly improved by controlling the read channel common mode output voltage shift in the write mode. The ML4611R option also provides an user controllable write current adjustment capability, available in the 24-Pin package only.

FEATURES

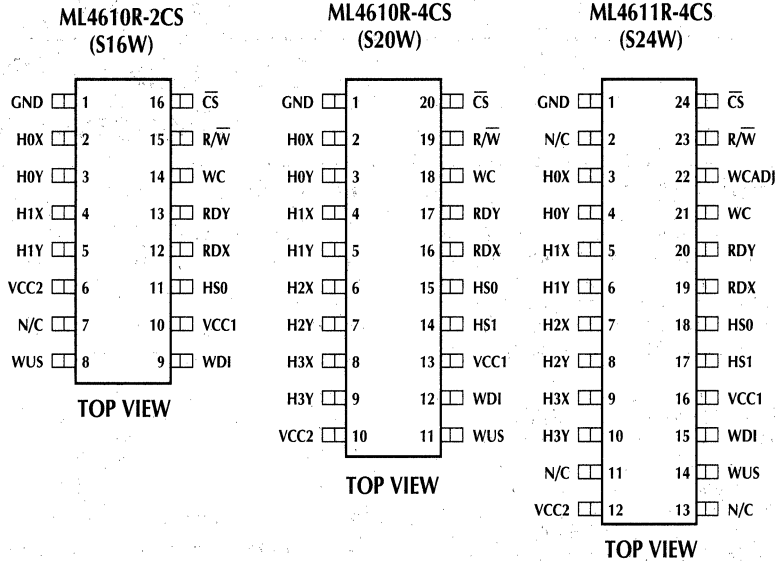
- Compatible to SSIs 32R4610R/4611R
- Can drop into SSIs 32R2020R series sockets
- Single +5 volt operation
- Low Power, $P_{IDLE} < 50mW$, $P_{MAX} < 200mW$
- Read Mode gain = 200V/V
- Damping resistors switched out in Read mode
- Input noise = 0.85nV/ \sqrt{Hz} max
- Input capacitance = 35pF max
- Write Current range = 10–35mA
- Programmable write current source
- Enhanced system write to read recovery time
- Power supply fault protection
- Head short to ground protection
- 24-pin SOIC (4 channel with WCADJ)
- 20-pin SOIC (4 channel without WCADJ)
- 16-pin SOIC (2 channel without WCADJ)

BLOCK DIAGRAM



ML4610R, ML4611R

PIN CONNECTION



PIN DESCRIPTION

NAME	TYPE	FUNCTION	NAME	TYPE	FUNCTION
HS0, HS1	I	Head Select: Selects one of four heads	RDX, RDY	O	X, Y Read Data: Differential read data output
\overline{CS}	I	Chip Select: A high inhibits the chip	WC		Write Current: Used to set the magnitude of the write current
R/\overline{W}	I	Read/Write: A high selects read mode	WCADJ		Write Current Adjust: Used to decrease the write current
WUS	O	Write Unsafe: A high indicates an unsafe writing condition	VCC1	I	+5 volt supply
WDI	I	Write Data In: Changes the direction of the current in the head	VCC2	I	+5 volt supply for write current drivers
H0X - H3X H0Y - H3Y	I/O	X,Y Head Connectors	GND	I	Ground

ABSOLUTE MAXIMUM RATINGS

DC Supply Voltage (V_{CC1})-0.3 to +7VDC
DC Supply Voltage (V_{CC2})-0.3 to +7VDC
Write Current (I_W)80mA
Digital Input Voltage (V_{IN})-0.3 to $V_{CC1} + 0.3$ VDC
Head Port Voltage (V_H)-0.3 to $V_{CC1} + 0.3$ VDC
Output Current: (RDX, RDY IO)-10mA
Output Current: (WUS)+12mA
Storage Temperature T_{STG}-65 to +150°C

RECOMMENDED OPERATING CONDITIONS

DC Supply Voltage (V_{CC1})5 ±5% VDC
DC Supply Voltage (V_{CC2})5 ±5% VDC
Operating Junction Temperature (T_J)+25° to +110°C

ELECTRICAL CHARACTERISTICS

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP	MAX	UNIT
V_{CC1}	Supply Current	Read Mode			33	mA
		Write Mode			27	mA
		Idle Mode			12	mA
V_{CC2}	Supply Current	Read Mode			11	mA
		Write Mode			10 + I_W	mA
		Idle Mode			0.4	mA
	Power Dissipation	Read Mode			200	mW
		Write Mode			150 + 4 I_W	mW
		Idle Mode			50	mW

DIGITAL INPUTS

V_{IL}	Input Low Voltage				0.8	VDC
V_{IH}	Input High Voltage		2.0			VDC
	Input Low Current	$V_{IL} = 0.8$ V	-0.4			mA
	Input High Current	$V_{IH} = 2.0$ V			100	μA
V_{OL}	WUS Output Low Voltage	$I_{OL} = 2$ mA max			0.5	VDC
	V_{CC1} Fault Voltage	$I_W < 0.2$ mA	3.8	4.0	4.2	VDC

WRITE CHARACTERISTICS

	Write Current Constant "K"			0.99		
V_{WC}	Write Current Voltage		1.15	1.25	1.35	V
	WCADJ Voltage	$I_{WCADJ} = 0$ to 0.5 mA	2.0	$V_{CC}/2$	3.0	VDC
	I_{HEAD} (DECREASE) / I_{WCADJ}		26	29	32	mA/mA
	I_{WCADJ} Range		0.0		0.5	mA
	Differential Head Voltage Swing		3.4	6		V_{P-P}
	Unselected Head Current				1	mA (pk)
	Head Differential Load Capacitance				25	pF
	Head Differential Load Resistance	R_D (ML4610R/4611R)	560	700	950	Ω
	WDI Transition Frequency	WUS = low	1.0			MHz
	Write Current Range (I_W)		10		35	mA

ML4610R, ML4611R

ELECTRICAL CHARACTERISTICS

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
READ CHARACTERISTICS CL (RDX,RDY) < 20 pF, RL (RDX, RDY) = 1kΩ						
	Differential Voltage Gain	$V_{IN} = 1mV_{P-P} @ 1MHz$	160	200	240	V/V
	Voltage BW -1dB -3dB	$ Z_{sl} < 5\Omega, V_{IN} = 1mV_{P-P}$	20 35			MHz MHz
	Input Noise Voltage	$BW = 15MHz, L_H = 0, R_H = 0$		0.6	0.85	nV/√Hz
	Differential Input Capacitance	$V_{IN} = 1mV_{P-P}, f = 5MHz$		27	35	pF
	Differential Input resistance	$V_{IN} = 1mV_{P-P}, f = 5MHz$	1000			Ω
	Dynamic Range	AC input voltage where gain falls to 90% of its small signal gain value, $f = 5MHz$	3			mV _{P-P}
	Common Mode Rejection Ratio	$V_{IN} = 0 \text{ volts DC} + 100mV_{P-P} @ 5MHz$	45			dB
	Power Supply Rejection Ratio	$100mV_{P-P} @ 5MHz \text{ on } V_{CC}$	40			dB
	Channel Separation	Unselected channels driven with $V_{IN} = 0 \text{ volts DC} + 100mV_{P-P}$	45			dB
	Output Offset Voltage		-200		+200	mV
	Single-Ended Output Resistance	$f = 5MHz$			40	Ω
	Output Current	AC coupled load, RDX to RDY	1.5			mA
RDX, RDY	Common Mode Output		2.0	2.8	3.5	VDC
SWITCHING CHARACTERISTICS $I_W = 20mA, R_H = 30\Omega, L_H = 1\mu H, f_{DATA} = 5MHz$						
R/\overline{W}	Read to Write	R/\overline{W} to 90% of write current		0.1	1.0	μs
R/\overline{W}	Write to Read	R/\overline{W} to 90% of 100mV Read signal envelope		0.5	1.0	μs
\overline{CS}	Unselect to Select	\overline{CS} to 90% of write current or 90% of 100mV, 10MHz		0.4	1.0	μs
\overline{CS}	Select to Unselect	\overline{CS} to 10% of write current		0.4	1.0	μs
	H50-1 to any head	To 90% of 100mV 10MHz Read signal envelope		0.2	1.0	μs
	WUS Safe to Unsafe (TD1) Unsafe to Safe (TD2)		0.6	2.0 0.2	3.6 1.0	μs μs
	Head Current: WDI to Ix - Iy (TD3) Asymmetry Rise/Fall Time	$L_H = 0, R_H = 0$ From 50% points WDI has 1ns rise/fall time 10% to 90% points			32 1.0 12	ns ns ns

TIMING DIAGRAM

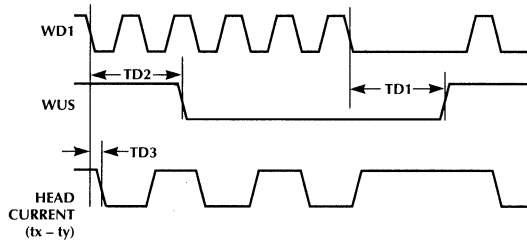


Figure 1. Write Mode.

MODE SELECT

\overline{CS}	R/\overline{W}	MODE
0	0	Write
0	1	Read
1	0	Idle
1	1	Idle

HEAD SELECT

HS1	HS0	HEAD
0	0	0
0	1	1
1	0	2
1	1	3

FUNCTIONAL DESCRIPTION

The ML4610R/4611R has the ability to address up to 4 two-terminal thin-film heads and provide write drive or read amplification. Head selection and mode control are described in the tables below. The TTL inputs R/\overline{W} and \overline{CS} have internal pull-up resistors to prevent an accidental write condition. HS0 and HS1 have internal pull-downs. Internal clamp circuitry will protect the ML4610R/4611R from a head short to ground condition in any mode. The damping resistors are switched out during read mode, as identified by the R/\overline{W} pin.

WRITE MODE OPERATION

Taking both \overline{CS} and R/\overline{W} low selects write mode which configures the ML4610R/4611R as a current switch and activates the Write Unsafe (WUS) detector circuitry. Head current is toggled between the X and Y side of the selected head on each high to low transition of the Write Data Input (WDI). A preceding read or idle mode select initializes the Write Data Flip-Flop to pass write current through the "X" side of the head. The current calculations are shown below:

Write current (peak) is given by:

$$I_w = \frac{K \times V_{WC}}{R_{WC}}$$

where

R_{WC} is connected from pin WC to GND

Actual head current is given by:

$$I_{X, Y} = \frac{I_w}{1 + \frac{R_H}{R_D}}$$

where

R_H = head + external wire resistance

R_D = damping resistance

The ML4610R/4611R adds a feature which allows the user to adjust the I_w current by a finite amount using the WCADJ pin, while writing to the disk. It is used by switching a separate write current adjust resistor in and out on the WCADJ pin or by connecting a DAC to that pin to sink a controllable amount of current. It is nominally biased to $V_{CC}/2$. Sinking current from this pin to ground, will divert a proportional amount of current from the actual head current while maintaining a constant current through the WC resistor and V_{CC} . Allowing WCADJ to float or pulling it high will cut off the circuit and it will have no effect. For example, if the nominal head current is set to 30mA through WC with WCADJ open, then for a

ML4610R, ML4611R

7.25mA head current decrease, a 10kΩ resistor would be connected from the WCADJ pin to ground. A TTL gate could be used as a switch with a small degradation in accuracy. A DAC could be programmed to sink 0.25mA from the WCADJ pin, for achieving the same function.

$$I_W \text{ head (decrease)} = \frac{29 \times V_{WCADJ}}{R_{WCADJ}}$$

where

V_{WCADJ} = Voltage on the WCADJ pin

R_{WCADJ} = Write current adjust setting resistor

VOLTAGE FAULT DETECTION

A voltage fault detection circuit improves data security by disabling the write current generator during a voltage fault or power start-up, regardless of mode. The Write Unsafe (WUS) open collector output goes high under the conditions given below. After the fault condition is removed, a negative transition on WDI is required to clear WUS.

- Write Data Input frequency too low
- Device in Read Mode
- Chip is disabled or head is open
- No write current

READ MODE OPERATION

The Read mode configures the ML4610R/4611R as a low noise differential amplifier and deactivates the write current generator. The RDX and RDY output are driven by emitter followers. They should be AC coupled to the load. The (X,Y) inputs are non-inverting to the (X,Y) outputs. In the Idle or Write mode, the read amplifier is deactivated and RDX, RDY outputs become high impedance. This facilitates multiple R/W applications (wired-OR RDX, RDY) and minimizes voltage drifts when switching from Write to Read mode. The write current source is also deactivated for both the Read and Idle mode. In addition the ML4610R/4611R supports the feature by which the internal damping resistors are switched out in the read mode, which allows the full signal to be amplified.

IDLE MODE OPERATION

Taking \overline{CS} high selects the idle mode which switches the RDX and RDY outputs into a high impedance state and deactivates the device. Power consumption in this mode is held to a minimum, less than 50mW.

ORDERING INFORMATION

PART NUMBER	TEMPERATURE RANGE	PACKAGE
ML4610R-2CS	0°C to +70°C	16-Pin SOIC (S16W)
ML4610R-4CS	0°C to +70°C	20-Pin SOIC (S20W)
ML4611R-4CS	0°C to +70°C	24-Pin SOIC (S24W)

24 Mbps Read Channel Filter/Equalizer

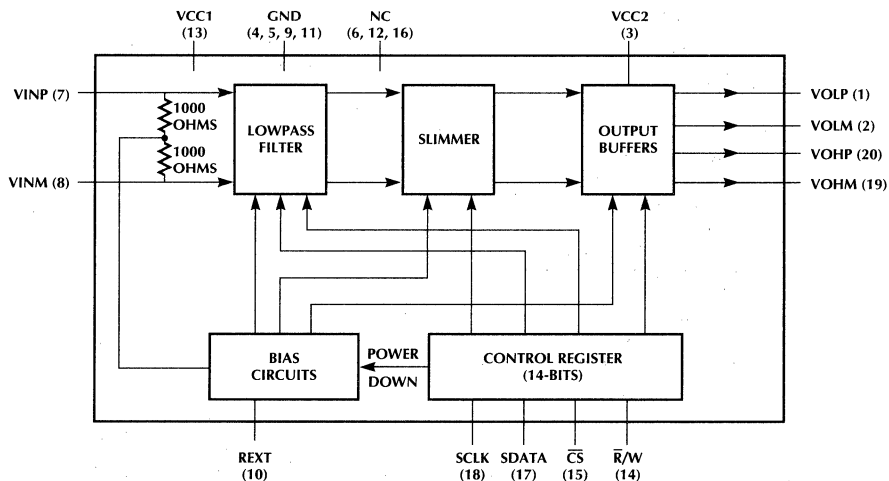
GENERAL DESCRIPTION

The ML6005 is a monolithic analog filter/equalizer intended for hard disk drive read channel applications, capable of handling disk data rates upto 24Mbits/s, with an operating power dissipation of less than 300mW. Its architecture consists of a continuous type filter based on a transistor and a high speed parasitic free active integrator, allowing complete independence of the filter response from interconnect parasitics, thus realizing a family of frequency response curves optimized for disk drive read channel equalization. It consists of a programmable 6-pole 2-zero lowpass filter stage, two pairs of high-speed drivers, and a serial microprocessor interface. The poles of the transfer function approximate a maximally flat group delay (Bessel) response, whereas the symmetric zeros provide the high-frequency boost necessary for pulse slimming. The user can independently adjust both the corner frequency, as well as the slimming level. The desired frequency response is programmed by a 14-bit serial input data stream which includes one bit for power-down, one bit for read/write control, and one bit for auto-zero control. The auto-zero circuitry, if enabled, reduces the output offsets to less than 20mV. The read/write control is also provided by a hardware pin. The ML6005 is well suited for constant density recording systems (Zoned-bit recording) as well as for constant data rate systems. A 36Mbits/s version, ML6006 is also available.

FEATURES

- 6-pole, 2-zero continuous time filter with < -45dB harmonic distortion
- Disk Data rates up to 24Mbit/s
- Programmable filter cutoff frequency (4.3:1 range in 64 steps) ($f_c = 3.13$ to 13.5MHz)
- 32 step programmable pulse slimming equalization, 0 to 10dB boost at f_c .
- Power-down, Auto-zero, R/W modes programmable through bits in the Control Register
- Lowpass output and Differentiated Lowpass (Bandpass) output provided.
- Fully I/O balanced architecture with TTL/CMOS compatible interface
- High speed (upto 25MHz clock) three wire serial microprocessor interface
- Double buffered data latch for synchronous or asynchronous data loading.
- Single $5V \pm 10\%$ power supply
- $0^\circ\text{C} - 70^\circ\text{C}$ operating temperature
- Available in 20-pin SSOP package.
- 4 GHz/1.5 μ BiCMOS process
- Power Dissipation — $P_{opr} = 300\text{mW}$, $P_{dn} = 7.5\text{mW}$

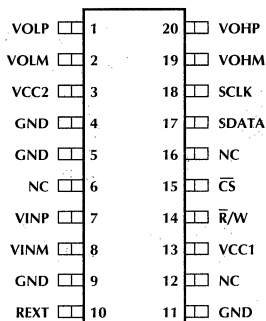
BLOCK DIAGRAM



ML6005

PIN CONNECTION

20-Pin SSOP



TOP VIEW

PIN DESCRIPTION

PIN#	NAME	FUNCTION	PIN#	NAME	FUNCTION
1	VOLP	Normal Lowpass outputs	15	\overline{CS}	Control Register Enable. A logical low level allows the SCLK input to clock data into the control register via the SDATA input line. A logical high level latches the control register contents and issues the information to the appropriate circuitry. A TTL input.
2	VOLM		17	SDATA	Control Register Data. A TTL input
3	VCC2	Positive supply for the output drivers, $5V \pm 10\%$	18	SCLK	Control Register Clock. Negative edge triggered control register clock input. A TTL input.
4, 5, 9, 11	GND	Ground	19	VOHM	Differentiated lowpass outputs
7	VINP	Signal Inputs	20	VOHP	
8	VINM		6, 12, 16	NC	No Connects, reserved for future use.
10	REXT	A 10K resistor between this pin and ground sets the filters corner frequency			
13	VCC1	Positive supply, $5V \pm 10\%$			
14	$\overline{R/W}$	Read/Write Control pin. A low input level allows normal operation of the filter in the read mode. A high level input puts the filter in the write mode, where the input impedance is lowered to prevent the transients generated during write to read transitions from affecting the filter response. A TTL input. Additionally a metal mask option is available to configure this pin as either power down enable or frequency boost disable			

ABSOLUTE MAXIMUM RATINGS

VCC1, VCC2	+6.5 volts
VINP, VINM, REXT, \overline{CS} , SCLK, SDATA, \overline{RW}	GND - 0.3V to VCC1 + 0.3V
VOLP, VOLM, VOHP, VOHM	GND - 0.3V to VCC2 + 0.3V
Input Current per pin	± 25 mA
Package Dissipation at Ta = 25°C (Surface Mount)	1.5 Watts
Junction Temperature	+150°C
Storage Temperature	-65°C to +150°C

OPERATING CONDITIONS

VCC1 = VCC2	+ 5 volts $\pm 10\%$
VIN = (VINP-VINM)	1 Vp-p
Rext	10 Kohms
Serial Clock Frequency (SCLK)	< 25 MHz
AC Coupling Capacitors	> 0.0001 μ F

ELECTRICAL CHARACTERISTICS

The following specifications apply over the recommended operating conditions, unless otherwise stated. Please refer to the application/test setup diagram:

VCC1 = VCC2 = 5 volt $\pm 10\%$, Ta = 0°C to 70°C, Rext = 10 Kohms
 VIN = (VINP - VINM) = 1 Vp-p sinewave input
 VOL = (VOLP - VOLM) and VOH = (VOHP - VOHM)
 Input and Output coupling capacitors = 0.0047 μ F
 RB1 = 750 ohms (pins 1 & 2), RB2 = 750 ohms (pins 19 & 20)
 RL = 1000 (1000) ohms and CL = 50 (50) pF on pins 1 (19) and 2 (20)
 Serial Clock Frequency = 20 MHz, Power Down, Auto Zero, Read/Write bits = 0
 Digital timing measured at 1.4V midpoint
 Input control signals from 10% - 90% of VCC1 with ($t_r = t_f$) < 5 ns.

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP.	MAX	UNITS
DC CHARACTERISTICS						
I _{CC}	VCC Supply Current	RB1 = RB2 = INF		60	74	mA
I _{pd}	Standby Current	VIN = 0		1.5	2	mA
DIGITAL INPUT CHARACTERISTICS (SCLK, SDATA, \overline{CS})						
V _{IL}	Low Voltage				0.8	V
V _{IH}	High Voltage		2.0			V
I _{IH}	High Current				1.0	μ A
I _{IL}	Low Current				-1.0	μ A
I _{CIN}	Input Current			5		pF
DIGITAL TIMING CHARACTERISTICS (SCLK, SDATA, \overline{CS})						
t _{PW-\overline{CS}}	Width of \overline{CS} , High/Low		25			ns
t _{SU-SDATA}	SDATA Setup time to SCLK		15			ns
t _{H-SDATA}	SDATA Hold Time		5			ns
t _{SU-\overline{CS}}	\overline{CS} Setup Time to SCLK		15			ns
t _{H-\overline{CS}}	\overline{CS} Hold Time to SCLK		0			ns
t _{PH-SCLK}	SCLK Pulse Width		20			ns
t _{H-SCLK}	\overline{CS} Inactive to SCLK Active		125			ns

ML6005

ELECTRICAL CHARACTERISTICS

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP.	MAX	UNITS	
EQUALIZER (NORMAL AND LOWPASS OUTPUT)							
AG	Absolute Gain	S0-S4 = 0, F0-F5 = 0 at 1MHz	-1.5	-0.5	0.5	dB	
CF	Cutoff Frequency, -3dB (f _{ref} = 1MHz)	S0-S4 = 0, (-3dB slimming)					
		F5 F4 F3 F2 F1 F0 (f _c)	0 0 0 0 0 0	12.15	13.50	14.85	MHz
		0 0 0 0 0 1	11.54	12.82	14.10	MHz	
		0 0 0 0 1 0	11.04	12.27	13.50	MHz	
		0 0 0 1 0 0	10.13	11.25	12.38	MHz	
		0 0 1 0 0 0	8.68	9.64	10.60	MHz	
		0 1 0 0 0 0	6.75	7.50	8.25	MHz	
		1 0 0 0 0 0	4.67	5.19	5.71	MHz	
1 1 1 1 1 1	2.82	3.13	3.44	MHz			
SL	Slimming Level (Gain at CF Referred to AG, V _{out} = 1Vp-p)	F0-F5 = 0; at CF					
		S4 S3 S2 S1 S0	0 0 0 0 1	-1.4	-2.4	-3.4	dB
		0 0 0 1 0	-0.9	-1.9	-2.9	dB	
		0 0 1 0 0	-0.0	-0.9	-1.9	dB	
		0 1 0 0 0	-0.2	0.8	1.8	dB	
		1 0 0 0 0	2.4	3.4	4.4	dB	
1 1 1 1 1	5.9	6.9	7.9	dB			
GD	Diff Group Delay	Fref = 5.0MHz, F0-F5 = 0	-1		+1	ns	
HD	Harmonic Distortion Second and Third related to Fundamental	F0-F5 = 0, V _{out} = 1.5Vp-p, Fin = 9.0MHz					
		S0-S4 = 0 (no slimming)			-45	dB	
		S0-S4 = 1 (full slimming)			-40	dB	
ICN	Idle Channel Noise (VIN = 0, DC - 78MHz)	F0-F5 = 0, VOLP S0-S4 = 0 (no slimming) S0-S4 = 1 (full slimming)			2 6	mVrms mVrms	
DR	Dynamic Range (Signal/(Noise + Distor)) Signal = 1Vp-p	F0-F5 = 0, Fin = 13.5MHz					
		S0-S4 = 0 (no slimming)			-41	dB	
		S0-S4 = 1 (full slimming)			-35	dB	
PSRR	Power Supply Rejection	100mVp-p sinewave on Vcc F0-F5 = 0, S0-S4 = 0, Vin = 0					
		Fin = 1.0MHz Fin = 40MHz		40 30		dB dB	
DELP HI	Phase Shift between LP and HP Output	All F's and S's = 0 Vin = 1Vp-p, Fin = 9.0MHz	88	90	92	Degree	
ANALOG							
VIP	Input Signal Monotonicity	All F's and S's = 0, (VINP - VINM) Fin = 9.0MHz		1	2	Vp-p	
RID	Differential Input Resistance	VIN = 100mVp-p at 6.7MHz	1.3	2	3	Kohms	
CID	Differential Input Capacitance	VIN = 100mVp-p at 6.7MHz		5		pF	
ZIC	Common-mode Input Impedence			1		Kohms	
RPD	Recovery from Pwr Dn	Auto Zero function OFF Auto Zero function ON		10 TBD		μs	
VOS	Output Offset Voltage	Differential VOLP or VOHP					
		Auto Zero ON (S0-S4 = 0 or 1)			20	mV	
		Auto Zero OFF (S0-S4 = 0)			300	mV	
		Auto Zero OFF (S0-S4 = 1)			TBD	mV	

ELECTRICAL CHARACTERISTICS

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP.	MAX	UNITS
ANALOG (Continued)						
ROD	Output Resistance	Differential VIN = 0; at 6.7MHz		5		Ohms
COD	Output Capacitance	Differential VIN = 0; at 6.7 MHz		8		pF
ROC	Output Resistance Common Mode	Common mode VIN = 0; at 6.7MHz		5		Ohms
COC	Output Capacitance Common Mode	Common mode VIN = 0; at 6.7MHz		15		pF
CLSE	Load Capacitance	VOLP; RB1 = 750 ohms VOHP; RB2 = 750 ohms			50 50	pF pF
RLSE	Load Resistance	VOLP VOHP	400 400			Ohms Ohms
RLOZ	Input Resistance	Diff; PD and/or RW bit = 1			350	Ohms
I _{OB}	Output Buffer Bias Current	VOLP or VOHP, VOLM or VOHM	1	1.4		mA
I _{OSC}	Short Circuit Output Current	VOLP or VOHP, VOLM or VOHM		50	100	mA
SBA	Stopband Attenuation	S0 = 0 at 2CF	TBD			dB

FUNCTIONAL DESCRIPTION

INTRODUCTION

Many of the high-frequency continuous-time filters have principally utilized a basic integrator consisting of a transconductance stage driving a passive integrating capacitor. These approaches are susceptible to frequency response variations due to the parasitic capacitances associated with the parasitic-sensitive output nodes of the integrator. This type of transconductance stage also often has low open-circuit voltage gain, resulting in limited practical Q range in the filter. The use of an active parasitic-insensitive integrator, has generally been avoided in these filters because of the additional excess phase that the amplifier contributes.

The ML6005 is a continuous-time filter based on a transconductor and a high-speed parasitic free active integrator, allowing complete independence of the filter response from interconnect parasitics and a very wide range of realizable filter Q. A unique approach to cancelling the excess phase contributed by the hi-speed BiCMOS amplifier allows this filter to achieve reproducible responses at 13.5 MHz filter bandwidth in a 1.5 μ /4GHz BiCMOS process. This active integrator incorporates a novel technique for setting the transconductance G_m value as a function of an external precision resistor, independent of temperature, supply, in conjunction with a wafer-sort trim technique to adjust capacitor process tolerances, thus eliminating the need for an on-chip PLL for tuning.

The ML6005 filter consists of a 6th order Bessel low-pass and a 2nd order cosine equalizer stage. It is made up of three biquads with lowpass and bandpass outputs. Both outputs of the last stage are available with matched group-delay characteristics. The corner frequency is digitally

programmable to 64 values over a 4 to 1 range, through the serial microprocessor interface. This is accomplished internally by changing the integrating capacitor value. Slimming equalization is done by digitally programming two real-symmetric zeroes, through the serial microprocessor interface. This boosts the high frequency response in 32 steps from 0 to 10 dB.

In a typical application, the ML6005 is used together with a pulse detector such as the ML541, ML4041 or the ML8464, making up a section of the AGC loop. Thus, the output of the AGC amplifier is AC coupled to the ML6005 input and the output of the ML6005 is AC coupled not only to the rectifier input thus closing the AGC loop, but also to the pulse detector input. The ML6005 provides two sets of fully balanced outputs. The lowpass outputs and the differentiated lowpass outputs. The ML6005 input and output common mode voltage biases are generated on-chip. The ML6005 consists of an input common bias circuit, a programmable continuous type equalizer filter with normal and differentiated lowpass outputs, followed by output buffers, and a high speed serial microprocessor interface. The ML6005 processes only differential input signals, common mode inputs are rejected. The output should also be taken differentially in order to obtain the best performance.

INPUT COMMON MODE

The input common mode bias consists of two resistors as shown in the block diagram, and a buffer which biases the center point with a well defined voltage required by the internal circuitry. These resistors are 1000 Ohms each and together with the external coupling capacitor define the lower corner frequency of the transfer function.

ML6005

EQUALIZER FILTER

The filter transfer function is composed of a second order numerator and a sixth order denominator. The low frequency attenuation is set internally to 0dB. The numerator realizes two zeros symmetrical to the imaginary axis, one in the left and the other in the right half plane. The location of the zeros is programmable. This realizes a digitally programmable pulse slimming function in order to overcome intersymbol interference and thus contribute to increasing bit density. The slimming level is controlled by 5 bits in the control register, thus providing 32 different choices between 0 to 10 dB. The denominator approximately realizes a maximally flat group delay (Bessel) function with a digitally programmable corner frequency controlled by 6 bits in the control register, thus providing 64 different cutoff frequencies.

SLIMMING LEVEL

The slimming levels generated by the slimming bits are shown below. There are 5 bits of control, S0 - S4. The typical gain of the equalizer at the cutoff frequency is shown in the table below. The gain at f_c in dB is also given by the formula :

$$\text{Gain (dB)} = 20 \times \text{Log} (0.707 \times (1 + 0.06868 \times K))$$

where $K = 0, 1, \dots, 31$

CUTOFF FREQUENCY

There are 6 bits in the control register that controls the position of the cutoff frequency, F0 - F5. The typical values of the cutoff (-3dB) frequency are shown in the table below for the case when S0 - S4=0 (no slimming). There are a total of 64 frequencies available from 13.5MHz down to 3.13MHz. Bits F1 - F5 will select one of 32 frequency settings in a monotonic fashion. Bit F0 is used to shift the whole frequency setting range by 5% lower than each of the 32 settings given by F1 - F5. This offers a scheme to increase the effective resolution of the cutoff frequency programmability. This feature is specially useful in the higher frequency range, where the granularity is coarse.

For example :

By setting $F_0 = 0$,

Cutoff frequency = 13.50 MHz with $F_5 - F_1 = 00000$ and

Cutoff frequency = 12.27 MHz with $F_5 - F_1 = 00001$, the next consecutive setting.

Frequency delta between consecutive settings = 1.23 MHz or about 9% of 13.50 MHz.

By setting $F_0 = 1$, we can shift the consecutive cutoff frequency settings as follows :

Cutoff frequency = 13.50 MHz with $(F_5 - F_1, F_0) = (00000, 0)$

Cutoff frequency = 12.82 MHz with $(F_5 - F_1, F_0) = (00000, 1)$ Delta = 0.68 MHz

TABLE 1: TABLE OF SLIMMING LEVEL PROGRAMMING VALUES

S4	S3	S2	S1	S0	K	GAIN AT F_c (dB)	STEPS (dB)
0	0	0	0	0	0	- 3.0	
0	0	0	0	1	1	- 2.4	0.6
0	0	0	1	0	2	- 1.9	0.5
0	0	0	1	1	3	- 1.4	0.5
0	0	1	0	0	4	- 0.9	0.5
0	0	1	0	1	5	- 0.4	0.5
0	0	1	1	0	6	- 0.0	0.4
0	0	1	1	1	7	0.4	0.4
0	1	0	0	0	8	0.8	0.4
0	1	0	0	1	9	1.2	0.4
0	1	0	1	0	10	1.5	0.3
0	1	0	1	1	11	1.9	0.4
0	1	1	0	0	12	2.2	0.3
0	1	1	0	1	13	2.5	0.3
0	1	1	1	0	14	2.8	0.3
0	1	1	1	1	15	3.1	0.3
1	0	0	0	0	16	3.4	0.3
1	0	0	0	1	17	3.7	0.3
1	0	0	1	0	18	4.0	0.3
1	0	0	1	1	19	4.2	0.2
1	0	1	0	0	20	4.5	0.3
1	0	1	0	1	21	4.7	0.2
1	0	1	1	0	22	5.0	0.3
1	0	1	1	1	23	5.2	0.2
1	1	0	0	0	24	5.4	0.2
1	1	0	0	1	25	5.7	0.3
1	1	0	1	0	26	5.9	0.2
1	1	0	1	1	27	6.1	0.2
1	1	1	0	0	28	6.3	0.2
1	1	1	0	1	29	6.5	0.2
1	1	1	1	0	30	6.7	0.2
1	1	1	1	1	31	6.9	0.2

Cutoff frequency = 12.27 MHz with $(F_5 - F_1, F_0) = (00001, 0)$ Delta = 0.55 MHz

Hence frequency delta between consecutive settings is lower, thus higher resolution.

In the table 2 below, the cutoff frequencies are shown as two columns depending on the F0 bit being zero or one. The monotonicity is guaranteed within the individual frequency columns, however because of the limitations of the 5% frequency circuitry, the monotonicity between the two columns cannot be guaranteed. This is especially significant at the lower end of the frequency range, where the difference in frequencies between the two column settings becomes very close (< 1%). Further tuning of the cutoff frequency down to the 1 to 10% range can be achieved by modifying the value of the external resistor

from its ideal 10 Kohms value by 1 to 10%, which shifts the whole response. Larger changes are not recommended for proper operation of the filter. The corner frequency is given by the formula outlined below :

$$f_c = \left(\frac{13.5 \times (1 - F0 \times 0.05)}{[1 + 0.1 \times \text{INT}(N/2)]} \times \frac{10 \text{Kohms}}{\text{Rext}} \right) \text{MHz}$$

OUTPUT BUFFER

The output buffer is the final stage of the ML6005 for both the normal and differentiated outputs. This is a fully differential buffer with unity gain. Only 1.4 mA of sinking current is provided on chip. More drive can be obtained by connecting external resistors to ground. The common mode output voltage is typically 2V.

SERIAL MICROPROCESSOR INTERFACE

The serial microprocessor interface consists of a simple three-wire serial port. It consists of a fourteen bit serial shift register with a double buffered latch for synchronous and asynchronous loading. A timing diagram and the control word definition are shown below. The 14-bit data word present on the SDATA line is serially shifted into the register on falling edges of the serial shift clock, SCLK, provided the CS pin is active (logical zero). F0 should be shifted in first, and F13 (the power-down bit) shifted in last as shown below. When the CS pin is inactive (logical one), SDATA and SCLK are ignored, and the previously shifted information is latched at the rising edge of CS becoming inactive (logical one). It is recommended that the SCLK input be kept inactive low till such time when it is in use. The SCLK input can run upto speeds of 25 MHz. The Autozero function, if enabled, minimizes the offsets at the filter outputs to 20mV.

TABLE 2: TABLE OF CUTOFF FREQUENCY PROGRAMMING VALUES IN MHZ

F5	F4	F3	F2	F1	N	f _c with F0 = 0	N	f _c with F0 = 1
0	0	0	0	0	0	13.5		
0	0	0	0	0			1	12.82
0	0	0	0	1	2	12.27		
0	0	0	0	1			3	11.66
0	0	0	1	0	4	11.25		
0	0	0	1	0			5	10.69
0	0	0	1	1	6	10.38		
0	0	0	1	1			7	9.87
0	0	1	0	0	8	9.64		
0	0	1	0	0			9	9.16
0	0	1	0	1	10	9.0		
0	0	1	0	1			11	8.55
0	0	1	1	0	12	8.44		
0	0	1	1	0			13	8.02
0	0	1	1	1	14	7.94		
0	0	1	1	1			15	7.54
0	1	0	0	0	16	7.50		
0	1	0	0	0			17	7.13
0	1	0	0	1	18	7.11		
0	1	0	0	1			19	6.75
0	1	0	1	0	20	6.75		
0	1	0	1	1	22	6.43		
0	1	0	1	0			21	6.41
0	1	1	0	0	24	6.14		
0	1	0	1	1			23	6.11
0	1	1	0	1	26	5.87		
0	1	1	0	0			25	5.83
0	1	1	1	0	28	5.63		
0	1	1	0	1			27	5.58
0	1	1	1	1	30	5.40		
0	1	1	1	0			29	5.34
1	0	0	0	0	32	5.19		

F5	F4	F3	F2	F1	N	f _c with F0 = 0	N	f _c with F0 = 1
0	1	1	1	1			31	5.13
1	0	0	0	1	34	5.00		
1	0	0	0	0			33	4.93
1	0	0	1	0	36	4.82		
1	0	0	0	1			35	4.75
1	0	0	1	1	38	4.66		
1	0	0	1	0			37	4.58
1	0	1	0	0	40	4.50		
1	0	0	1	1			39	4.42
1	0	1	0	1	42	4.35		
1	0	1	0	0			41	4.28
1	0	1	1	0	44	4.22		
1	0	1	0	1			43	4.14
1	0	1	1	1	46	4.09		
1	0	1	1	0			45	4.01
1	1	0	0	0	48	3.97		
1	0	1	1	1			47	3.89
1	1	0	0	1	50	3.86		
1	1	0	0	0			49	3.77
1	1	0	1	0	52	3.75		
1	1	0	0	1			51	3.66
1	1	0	1	1	54	3.65		
1	1	0	1	0			53	3.56
1	1	1	0	0	56	3.55		
1	1	1	0	0			55	3.47
1	1	1	0	1	58	3.46		
1	1	1	0	0			57	3.38
1	1	1	1	0	60	3.38		
1	1	1	0	1			59	3.29
1	1	1	1	1	62	3.29		
1	1	1	1	0			61	3.21
1	1	1	1	1			63	3.13

Note: N is the decimal value of the cutoff frequency bits (F5 - F0), in the control register

ML6005

TIMING DIAGRAM

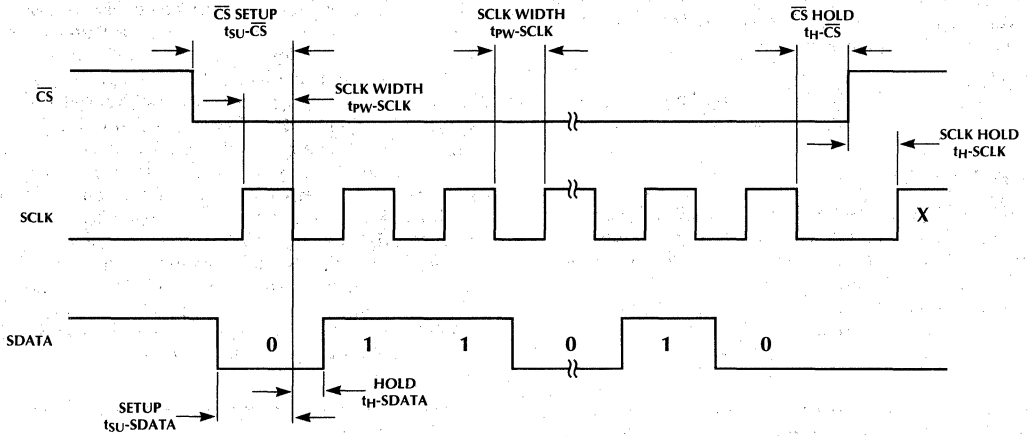
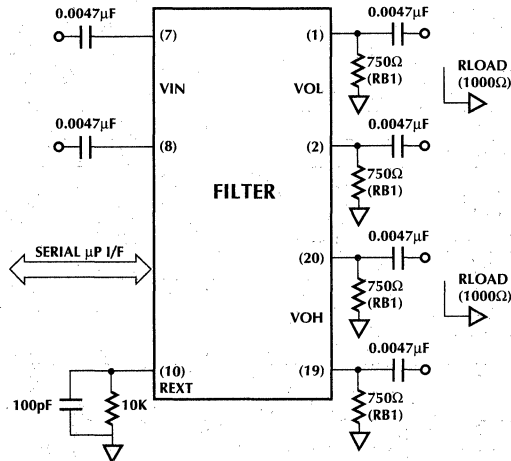


Figure 1.

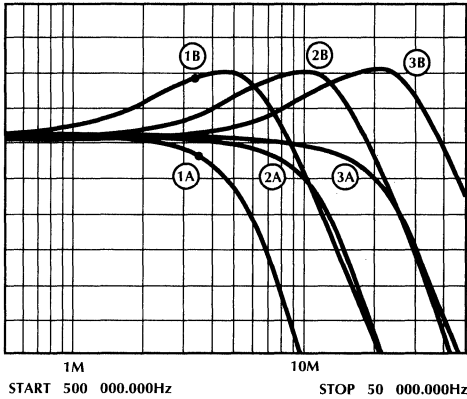
CONTROL REGISTER DEFINITION

F13	F12	F11	F10	F9	F8	F7	F6	F5	F4	F3	F2	F1	F0
AZ	PD	$\overline{R/W}$	SLIMMING CONTROL				FREQUENCY CONTROL						
AZ	AutoZero		1 = Autozero circuitry activated 0 = Autozero circuitry inactive										
PD	Power Down		1 = Chip is in power down mode 0 = Chip is fully powered up										
$\overline{R/W}$	Read/Write		1 = Write data mode 0 = Read data mode										

APPLICATIONS CIRCUIT/TEST SETUP



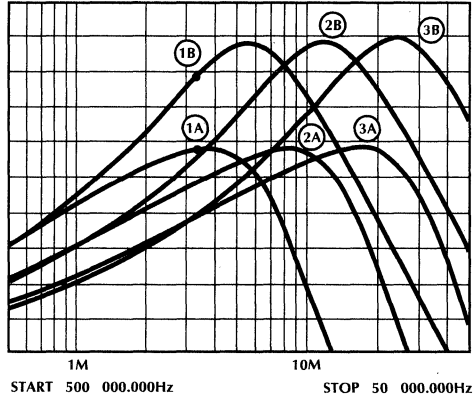
REF LEVEL /DIV
 -15.000dB 5.000dB
 -15.000dB 5.000dB
 MARKER 3 385 752.200Hz
 MAG (UDF) -25.348dB
 MARKER 3 385 752.200Hz
 MAG (D4) -36.371dB



Filter Response (Lowpass Output)

Shown are the ML6005 filter response at three different cutoff frequency (f_c) settings. Setting 1 = 3.13 MHz, 2 = 6.75 MHz and 3 = 13.5 MHz. At each of the f_c settings, the filter response is shown with no slimming (A) and with full slimming (B) activated.

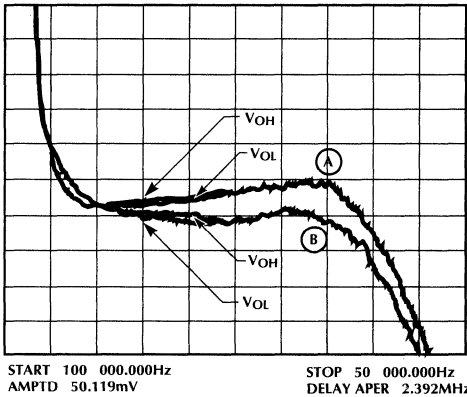
REF LEVEL /DIV
 -15.000dB 5.000dB
 -15.000dB 5.000dB
 MARKER 3 385 752.200Hz
 MAG (UDF) -25.270dB
 MARKER 3 385 752.200Hz
 MAG (D4) -36.157dB



Filter Response (Bandpass Output)

Shown are the ML6005 filter response at three different cutoff frequency (f_c) settings. Setting 1 = 3.13 MHz, 2 = 6.75 MHz and 3 = 13.5 MHz. At each of the f_c settings, the filter response is shown with no slimming (A) and with full slimming (B) activated.

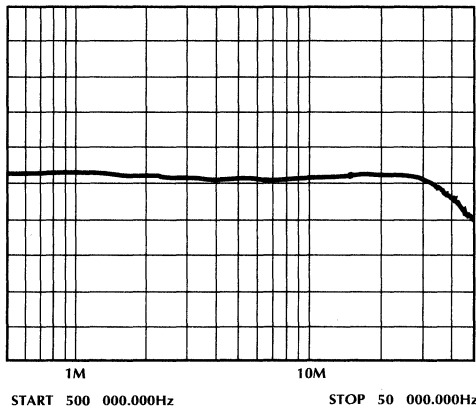
REF LEVEL /DIV
 40.000nSEC 1.000nSEC
 40.000nSEC 1.000nSEC
 MARKER 10 415 500.00Hz
 DELAY (UDF) 39.387nSEC
 MARKER 10 415 500.00Hz
 DELAY (UDF) 39.413nSEC



ML6005 Filter/Equalizer group Delay Tracking

Shown are the curves to demonstrate group delay tracking between the lowpass (V_{OL}) and bandpass (V_{OH}) outputs, at an f_c of 13.5 MHz, with on slimming activated (A) and full slimming activated (B). It can be seen that the group delay tracking between the lowpass and bandpass outputs is well within 1 ns.

REF LEVEL /DIV
 -90.000deg 1.000deg
 MARKER 13 632 170.100Hz
 PHASE (UDF) -89.709deg



Phase Difference between Lowpass and Bandpass Outputs

Shown is the delta in the phase between the lowpass and bandpass outputs. Ideally the bandpass output should be -90° . The curve shows that this is within 1° for a frequency range of 50 MHz to 10 MHz.

ML6005 FILTER/EQUALIZER CHARACTERISTICS

ML6005

ORDERING INFORMATION

PART NUMBER	TEMPERATURE RANGE	PACKAGE
ML6005CR	0°C to +70°C	20-Pin SSOP (R20)

36 Mbps Read Channel Filter/Equalizer

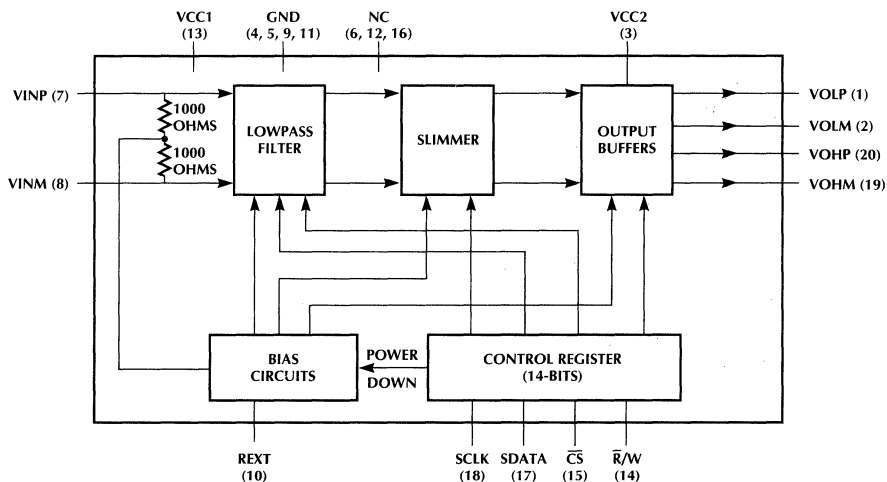
GENERAL DESCRIPTION

The ML6006 is a monolithic analog filter/equalizer intended for hard disk drive read channel applications, capable of handling disk data rates upto 36Mbits/s, with an operating power dissipation of less than 350mW. Its architecture consists of a continuous type filter based on a transistor and a high speed parasitic free active integrator, allowing complete independence of the filter response from interconnect parasitics, thus realizing a family of frequency response curves optimized for disk drive read channel equalization. It consists of a programmable 6-pole 2-zero lowpass filter stage, two pairs of high-speed drivers, and a serial microprocessor interface. The poles of the transfer function approximate a maximally flat group delay (Bessel) response, whereas the symmetric zeros provide the high-frequency boost necessary for pulse slimming. The user can independently adjust both the corner frequency, as well as the slimming level. The desired frequency response is programmed by a 14-bit serial input data stream which includes one bit for power-down, one bit for read/write control, and one bit for auto-zero control. The read/write control is also provided by a hardware pin. The ML6006 is well suited for constant density recording systems (Zoned-bit recording) as well as for constant data rate systems. A 24 Mbits/s version, ML6005 is also available.

FEATURES

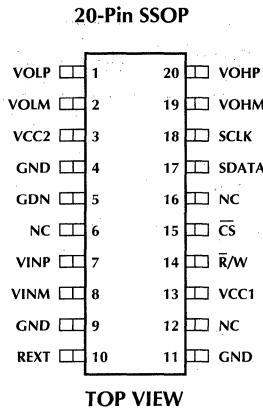
- 6-pole, 2-zero continuous time filter with $< -45\text{dB}$ harmonic distortion
- Disk Data rates up to 36 Mbit/s
- Programmable filter cutoff frequency (4.3:1 range in 64 steps) ($f_c = 4.69$ to 20.25 MHz)
- 32 step programmable pulse slimming equalization, 0 to 10dB boost at f_c .
- Power-down, Auto-zero, R/W modes programmable through bits in the Control Register
- Lowpass output and Differentiated Lowpass (Bandpass) output provided.
- Fully I/O balanced architecture with TTL/CMOS compatible interface
- High speed (upto 25MHz clock) three wire serial microprocessor interface
- Double buffered data latch for synchronous or asynchronous data loading.
- Single $5\text{V} \pm 10\%$ power supply
- $0^\circ\text{C} - 70^\circ\text{C}$ operating temperature
- Available in 20-pin SSOP package.
- 4 GHz/ $1.5\mu\text{m}$ BiCMOS process
- Power Dissipation: $P_{opr} = 350\text{mW}$, $P_{dn} = 7.5\text{mW}$

BLOCK DIAGRAM



ML6006

PIN CONNECTION



PIN DESCRIPTION

PIN#	NAME	FUNCTION	PIN#	NAME	FUNCTION
1	VOLP	Normal Lowpass outputs	15	\overline{CS}	Control Register Enable. A logical low level allows the SCLK input to clock data into the control register via the SDATA input line. A logical high level latches the control register contents and issues the information to the appropriate circuitry. A TTL input.
2	VOLM		17	SDATA	Control Register Data. A TTL input
3	VCC2	Positive supply for the output drivers, $5V \pm 10\%$	18	SCLK	Control Register Clock. Negative edge triggered control register clock input. A TTL input.
4, 5, 9, 11	GND	Ground	19	VOHM	Differentiated lowpass outputs
7	VINP	Signal Inputs	20	VOHP	
8	VINM		6, 12, 16	NC	No Connects, reserved for future use.
10	REXT	A 10K resistor between this pin and ground sets the filters corner frequency			
13	VCC1	Positive supply, $5V \pm 10\%$			
14	$\overline{R/W}$	Read/Write Control pin. A low input level allows normal operation of the filter in the read mode. A high level input puts the filter in the write mode, where the input impedance is lowered to prevent the transients generated during write to read transitions from affecting the filter response. A TTL input. Additionally a metal mask option is available to configure this pin as either power down enable or frequency boost disable			

ABSOLUTE MAXIMUM RATINGS

VCC1, VCC2 +6.5 volts
 VINP, VINM, REXT, ~CS, SCLK,
 SDATA, ~R/W GND - 0.3V to VCC1 + 0.3V
 VOLP, VOLM,
 VOHP, VOHM GND - 0.3V to VCC2 + 0.3V
 Input Current per pin ± 25 mA
 Package Dissipation
 at Ta = 25°C (Surface Mount) 1.5 Watts
 Junction Temperature +150°C
 Storage Temperature -65°C to +150°C
 Lead Temperature (Soldering 10 sec) 260°C

OPERATING CONDITIONS

VCC1 = VCC2 + 5 volts ± 10%
 VIN = (VINP - VINM) 1 Vp-p
 Rext 10 Kohms
 Serial Clock Frequency (SCLK) < 25 MHz
 AC Coupling Capacitors > 0.0001 µF

ELECTRICAL CHARACTERISTICS

The following specifications apply over the recommended operating conditions, unless otherwise stated. Please refer to the application/test setup digram:

VCC1 = VCC2 = 5 volt ± 10%, Ta = 0°C to 70°C, Rext = 10 Kohms
 VIN = (VINP - VINM) = 1 Vp-p sinewave input
 VOL = (VOLP - VOLM) and VOH = (VOHP - VOHM)
 Input and Output coupling capacitors = 0.0047 µF
 RB1 = 750 ohms (pins 1 & 2), RB2 = 750 ohms (pins 19 & 20)
 RL = 1000 (1000) ohms and CL = 50 (50) pF on pins 1 (19) and 2 (20)
 Serial Clock Frequency = 20 MHz, Power Down, Auto Zero, Read/Write bits = 0
 Digital timing measured at 1.4V midpoint
 Input control signals from 10% - 90% of VCC1 with (tr = tf) < 5 ns.

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP.	MAX	UNITS
DC CHARACTERISTICS						
I _{CC}	VCC Supply Current	RB1 = RB2 = INF		70	84	mA
I _{pd}	Standby Current	VIN = 0		1.5	2	mA
DIGITAL INPUT CHARACTERISTICS (SCLK, SDATA, \overline{CS})						
VIL	Low Voltage				0.8	V
VIH	High Voltage		2.0			V
I _{IH}	High Current				1.0	µA
I _{IL}	Low Current				-1.0	µA
CIN	Input Current			5		pF
DIGITAL TIMING CHARACTERISTICS (SCLK, SDATA, \overline{CS})						
t _{PW-\overline{CS}}	Width of \overline{CS} , High/Low		25			ns
t _{SU-SDATA}	SDATA Setup time to SCLK		15			ns
t _{H-SDATA}	SDATA Hold Time		5			ns
t _{SU-\overline{CS}}	\overline{CS} Setup Time to SCLK		15			ns
t _{H-\overline{CS}}	\overline{CS} Hold Time to SCLK		0			ns
t _{PH-SCLK}	SCLK Pulse Width		20			ns
t _{H-SCLK}	\overline{CS} Inactive to SCLK Active		125			ns

ELECTRICAL CHARACTERISTICS

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP.	MAX	UNITS	
EQUALIZER (NORMAL AND LOWPASS OUTPUT)							
AG	Absolute Gain	S0-S4 = 0, F0-F5 = 0 at 1MHz	-1.5	-0.5	0.5	dB	
CF	Cutoff Frequency, -3dB ($f_{ref} = 1\text{MHz}$)	S0-S4 = 0, (-3dB slimming)					
		F5 F4 F3 F2 F1 F0 (f_c)		18.23	20.25	22.28	MHz
		0 0 0 0 0 0		17.32	19.24	21.16	MHz
		0 0 0 0 0 1		16.57	18.41	20.25	MHz
		0 0 0 1 0 0		15.19	16.88	18.57	MHz
		0 0 1 0 0 0		13.01	14.46	15.91	MHz
		0 1 0 0 0 0		10.13	11.25	12.38	MHz
		1 0 0 0 0 0		7.01	7.79	8.57	MHz
1 1 1 1 1 1		4.22	4.69	5.16	MHz		
SL	Slimming Level (Gain at CF Referred to AG, $V_{out} = 1V_{p-p}$)	F0-F5 = 0; at CF					
		S4 S3 S2 S1 S0		-1.4	-2.4	-3.4	dB
		0 0 0 0 1		-0.9	-1.9	-2.9	dB
		0 0 0 1 0		-0.0	-0.9	-1.9	dB
		0 1 0 0 0		-0.2	0.8	1.8	dB
		1 0 0 0 0		2.4	3.4	4.4	dB
1 1 1 1 1		5.9	6.9	7.9	dB		
GD	Diff Group Delay	$f_{ref} = 7.5\text{MHz}$, F0-F5 = 0	-1		+1	ns	
HD	Harmonic Distortion Second and Third related to Fundamental	F0-F5 = 0, $V_{out} = 1.5V_{p-p}$, $f_{in} = 13.5\text{MHz}$			-45	dB	
		S0-S4 = 0 (no slimming) S0-S4 = 1 (full slimming)			-40	dB	
ICN	Idle Channel Noise ($V_{IN} = 0$, DC - 78MHz)	F0-F5 = 0, VOLP S0-S4 = 0 (no slimming) S0-S4 = 1 (full slimming)			2 -6	mVrms mVrms	
DR	Dynamic Range (Signal/(Noise + Distor)) (Signal = $1V_{p-p}$)	F0-F5 = 0, $f_{in} = 13.5\text{MHz}$ S0-S4 = 0 (no slimming) S0-S4 = 1 (full slimming)			-41 -35	dB dB	
PSRR	Power Supply Rejection	100mVp-p sinewave on Vcc					
		F0-F5 = 0, S0-S4 = 0, $V_{in} = 0$ $f_{in} = 1.0\text{MHz}$ $f_{in} = 40\text{MHz}$		40 30		dB dB	
DELPHI	Phase Shift between LP and HP Output	All F's and S's = 0 $V_{in} = 1V_{p-p}$, $f_{in} = 13.5\text{MHz}$	88	90	92	Degree	
ANALOG							
VIP	Input Signal Monotonicity	All F's and S's = 0, (VINP - VINM) $f_{in} = 13.5\text{MHz}$		1	2	Vp-p	
RID	Differential Input Resistance	$V_{IN} = 100mV_{p-p}$ at 10MHz	1.3	2	3	Kohms	
CID	Differential Input Capacitance	$V_{IN} = 100mV_{p-p}$ at 10MHz		5		pF	
ZIC	Common-mode Input Impedance			1		Kohms	
RPD	Recovery from Pwr Dn	Auto Zero function OFF Auto Zero function ON		10 TBD		μs	
VOS	Output Offset Voltage	Differential VOLP or VOHP					
		Auto Zero ON (S0-S4 = 0 or 1)			20	mV	
		Auto Zero OFF (S0-S4 = 0) Auto Zero OFF (S0-S4 = 1)			300 TBD	mV mV	

ELECTRICAL CHARACTERISTICS

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP.	MAX	UNITS
ANALOG (Continued)						
ROD	Output Resistance	Differential VIN = 0; at 10MHz		5		Ohms
COD	Output Capacitance	Differential VIN = 0; at 10 MHz		8		pF
ROC	Output Resistance Common Mode	Common mode VIN = 0; at 10MHz		5		Ohms
COC	Output Capacitance Common Mode	Common mode VIN = 0; at 10MHz		15		pF
CLSE	Load Capacitance	VOLP; RB1 = 750 ohms VOHP; RB2 = 750 ohms			50 50	pF pF
RLSE	Load Resistance	VOLP VOHP	400 400			Ohms Ohms
RLOZ	Input Resistance	Diff; PD and/or RW bit = 1			350	Ohms
I _{OB}	Output Buffer Bias Current	VOLP or VOHP, VOLM or VOHM	1	1.4		mA
I _{OSC}	Short Circuit Output Current	VOLP or VOHP, VOLM or VOHM		50	100	mA
SBA	Stopband Attenuation	S0 = 0 at 2CF	TBD			dB

FUNCTIONAL DESCRIPTION

INTRODUCTION

Many of the high-frequency continuous-time filters have principally utilized a basic integrator consisting of a transconductance stage driving a passive integrating capacitor. These approaches are susceptible to frequency response variations due to the parasitic capacitances associated with the parasitic-sensitive output nodes of the integrator. This type of transconductance stage also often has low open-circuit voltage gain, resulting in limited practical Q range in the filter. The use of an active parasitic-insensitive integrator, has generally been avoided in these filters because of the additional excess phase that the amplifier contributes.

The ML6006 is a continuous-time filter based on a transconductor and a high-speed parasitic free active integrator, allowing complete independence of the filter response from interconnect parasitics and a very wide range of realizable filter Q. A unique approach to cancelling the excess phase contributed by the hi-speed BiCMOS amplifier allows this filter to achieve reproducible responses at 20 MHz filter bandwidth in a 1.5 μ /4GHz BiCMOS process. This active integrator incorporates a novel technique for setting the transconductance G_m value as a function of an external precision resistor, independent of temperature, supply, in conjunction with a wafer-sort trim technique to adjust capacitor process tolerances, thus eliminating the need for an on-chip PLL for tuning.

The ML6006 filter consists of a 6th order Bessel low-pass and a 2nd order cosine equalizer stage. It is made up of three biquads with lowpass and bandpass outputs. Both outputs of the last stage are available with matched group-delay characteristics. The corner frequency is digitally

programmable to 64 values over a 4 to 1 range, through the serial microprocessor interface. This is accomplished internally by changing the integrating capacitor value. Slimming equalization is done by digitally programming two real-symmetric zeroes, through the serial microprocessor interface. This boosts the high frequency response in 32 steps from 0 to 10 dB.

In a typical application, the ML6006 is used together with a pulse detector such as the ML541, ML4041 or the ML8464, making up a section of the AGC loop. Thus, the output of the AGC amplifier is AC coupled to the ML6006 input and the output of the ML6006 is AC coupled not only to the rectifier input thus closing the AGC loop, but also to the pulse detector input. The ML6006 provides two sets of fully balanced outputs. The lowpass outputs and the differentiated lowpass outputs. The ML6006 input and output common mode voltage biases are generated on-chip. The ML6005 consists of an input common bias circuit, a programmable continuous type equalizer filter with normal and differentiated lowpass outputs, followed by output buffers, and a high speed serial microprocessor interface. The ML6005 processes only differential input signals, common mode inputs are rejected. The output should also be taken differentially in order to obtain the best performance.

INPUT COMMON MODE

The input common mode bias consists of two resistors as shown in the block diagram, and a buffer which biases the center point with a well defined voltage required by the internal circuitry. These resistors are 1000 Ohms each and together with the external coupling capacitor define the lower corner frequency of the transfer function.

ML6006

EQUALIZER FILTER

The filter transfer function is composed of a second order numerator and a sixth order denominator. The low frequency attenuation is set internally to 0dB. The numerator realizes two zeros symmetrical to the imaginary axis, one in the left and the other in the right half plane. The location of the zeros is programmable. This realizes a digitally programmable pulse slimming function in order to overcome intersymbol interference and thus contribute to increasing bit density. The slimming level is controlled by 5 bits in the control register, thus providing 32 different choices between 0 to 10 dB. The denominator approximately realizes a maximally flat group delay (Bessel) function with a digitally programmable corner frequency controlled by 6 bits in the control register, thus providing 64 different cutoff frequencies.

SLIMMING LEVEL

The slimming levels generated by the slimming bits are shown below. There are 5 bits of control, S0 - S4. The typical gain of the equalizer at the cutoff frequency is shown in the table below. The gain at f_c in dB is also given by the formula :

$$\text{Gain (dB)} = 20 \times \text{Log} (0.707 \times (1 + 0.06868 \times K))$$

where $K = 0, 1, \dots, 31$

CUTOFF FREQUENCY

There are 6 bits in the control register that controls the position of the cutoff frequency, F0 - F5. The typical values of the cutoff (-3dB) frequency are shown in the table below for the case when S0 - S4=0 (no slimming). There are a total of 64 frequencies available from 20.25MHz down to 4.69MHz. Bits F1 - F5 will select one of 32 frequency settings in a monotonic fashion. Bit F0 is used to shift the whole frequency setting range by 5% lower than each of the 32 settings given by F1 - F5. This offers a scheme to increase the effective resolution of the cutoff frequency programmability. This feature is specially useful in the higher frequency range, where the granularity is coarse.

For example :

By setting $F0 = 0$,

Cutoff frequency = 20.25 MHz with $F5 - F1 = 00000$ and

Cutoff frequency = 18.41 MHz with $F5 - F1 = 00001$, the next consecutive setting.

Frequency delta between consecutive settings = 1.84 MHz or about 9% of 20.25 MHz.

By setting $F0 = 1$, we can shift the consecutive cutoff frequency settings as follows :

Cutoff frequency = 20.25 MHz with $(F5 - F1, F0) = (00000, 0)$

Cutoff frequency = 19.24 MHz with $(F5 - F1, F0) = (00000, 1)$ Delta = 1.01 MHz

TABLE 1: TABLE OF SLIMMING LEVEL PROGRAMMING VALUES

S4	S3	S2	S1	S0	K	GAIN AT F_c (DB)	STEPS (DB)
0	0	0	0	0	0	-3.0	
0	0	0	0	1	1	-2.4	0.6
0	0	0	1	0	2	-1.9	0.5
0	0	0	1	1	3	-1.4	0.5
0	0	1	0	0	4	-0.9	0.5
0	0	1	0	1	5	-0.4	0.5
0	0	1	1	0	6	-0.0	0.4
0	0	1	1	1	7	0.4	0.4
0	1	0	0	0	8	0.8	0.4
0	1	0	0	1	9	1.2	0.4
0	1	0	1	0	10	1.5	0.3
0	1	0	1	1	11	1.9	0.4
0	1	1	0	0	12	2.2	0.3
0	1	1	0	1	13	2.5	0.3
0	1	1	1	0	14	2.8	0.3
0	1	1	1	1	15	3.1	0.3
1	0	0	0	0	16	3.4	0.3
1	0	0	0	1	17	3.7	0.3
1	0	0	1	0	18	4.0	0.3
1	0	0	1	1	19	4.2	0.2
1	0	1	0	0	20	4.5	0.3
1	0	1	0	1	21	4.7	0.2
1	0	1	1	0	22	5.0	0.3
1	0	1	1	1	23	5.2	0.2
1	1	0	0	0	24	5.4	0.2
1	1	0	0	1	25	5.7	0.3
1	1	0	1	0	26	5.9	0.2
1	1	0	1	1	27	6.1	0.2
1	1	1	0	0	28	6.3	0.2
1	1	1	0	1	29	6.5	0.2
1	1	1	1	0	30	6.7	0.2
1	1	1	1	1	31	6.9	0.2

Cutoff frequency = 18.41 MHz with $(F5 - F1, F0) = (00001, 0)$ Delta = 0.83 MHz

Hence frequency delta between consecutive settings is lower, thus higher resolution.

In the table 2 below, the cutoff frequencies are shown as two columns depending on the F0 bit being zero or one. The monotonicity is guaranteed within the individual frequency columns, however because of the limitations of the 5% frequency circuitry, the monotonicity between the two columns cannot be guaranteed. This is especially significant at the lower end of the frequency range, where the difference in frequencies between the two column settings becomes very close (< 1%). Further tuning of the cutoff frequency down to the 1 to 10% range can be achieved by modifying the value of the external resistor

from its ideal 10 Kohms value by 1 to 10%, which shifts the whole response. Larger changes are not recommended for proper operation of the filter. The corner frequency is given by the formula outlined below :

$$f_c = \left(\frac{20.25 \times (1 - F0 \times 0.05)}{[1 + 0.1 \times \text{INT}(N/2)]} \times \frac{10\text{Kohms}}{\text{Rext}} \right) \text{MHz}$$

OUTPUT BUFFER

The output buffer is the final stage of the ML6006 for both the normal and differentiated outputs. This is a fully differential buffer with unity gain. Only 1.4 mA of sinking current is provided on chip. More drive can be obtained by connecting external resistors to ground. The common mode output voltage is typically 2V.

SERIAL MICROPROCESSOR INTERFACE

The serial microprocessor interface consists of a simple three-wire serial port. It consists of a fourteen bit serial shift register with a double buffered latch for synchronous and asynchronous loading. A timing diagram and the control word definition are shown below. The 14-bit data word present on the SDATA line is serially shifted into the register on falling edges of the serial shift clock, SCLK, provided the CS pin is active (logical zero). F0 should be shifted in first, and F13 (the power-down bit) shifted in last as shown below. When the CS pin is inactive (logical one), SDATA and SCLK are ignored, and the previously shifted information is latched at the rising edge of CS becoming inactive (logical one). It is recommended that the SCLK input be kept inactive low till such time when it is in use. The SCLK input can run upto speeds of 25 MHz. The Autozero function, if enabled, minimizes the offsets at the filter outputs to 20mV.

TABLE 2: TABLE OF CUTOFF FREQUENCY PROGRAMMING VALUES IN MHZ

F5	F4	F3	F2	F1	N	f _c with F0 = 0	N	f _c with F0 = 1
0	0	0	0	0	0	20.25		
0	0	0	0	0			1	19.24
0	0	0	0	1	2	18.41		
0	0	0	0	1			3	17.49
0	0	0	1	0	4	16.88		
0	0	0	1	0			5	16.03
0	0	0	1	1	6	15.58		
0	0	0	1	1			7	14.80
0	0	1	0	0	8	14.46		
0	0	1	0	0			9	13.74
0	0	1	0	1	10	13.50		
0	0	1	0	1			11	12.83
0	0	1	1	0	12	12.66		
0	0	1	1	0			13	12.02
0	0	1	1	1	14	11.91		
0	0	1	1	1			15	11.32
0	1	0	0	0	16	11.25		
0	1	0	0	0			17	10.69
0	1	0	0	1	18	10.66		
0	1	0	0	1			19	10.13
0	1	0	1	0	20	10.13		
0	1	0	1	1	22	9.64		
0	1	0	1	0			21	9.62
0	1	1	0	0	24	9.20		
0	1	0	1	1			23	9.16
0	1	1	0	1	26	8.80		
0	1	1	0	0			25	8.74
0	1	1	1	0	28	8.44		
0	1	1	0	1			27	8.36
0	1	1	1	1	30	8.10		
0	1	1	1	0			29	8.02
1	0	0	0	0	32	7.79		

F5	F4	F3	F2	F1	N	f _c with F0 = 0	N	f _c with F0 = 1
0	1	1	1	1			31	7.70
1	0	0	0	1	34	7.50		
1	0	0	0	0			33	7.40
1	0	0	1	0	36	7.23		
1	0	0	0	1			35	7.12
1	0	0	1	1	38	6.98		
1	0	0	1	0			37	6.87
1	0	1	0	0	40	6.75		
1	0	0	1	1			39	6.63
1	0	1	0	1	42	6.53		
1	0	1	0	0			41	6.41
1	0	1	1	0	44	6.33		
1	0	1	0	1			43	6.21
1	0	1	1	1	46	6.14		
1	0	1	1	0			45	6.01
1	1	0	0	0	48	5.96		
1	0	1	1	1			47	5.83
1	1	0	0	1	50	5.79		
1	1	0	0	0			49	5.66
1	1	0	1	0	52	5.63		
1	1	0	0	1			51	5.50
1	1	0	1	1	54	5.47		
1	1	0	1	0			53	5.34
1	1	1	0	0	56	5.33		
1	1	1	0	0			55	5.20
1	1	1	0	1	58	5.19		
1	1	1	0	0			57	5.06
1	1	1	1	0	60	5.06		
1	1	1	0	1			59	4.93
1	1	1	1	1	62	4.94		
1	1	1	1	0			61	4.81
1	1	1	1	1			63	4.69

Note: N is the decimal value of the cutoff frequency bits (F5 - F0), in the control register

ML6006

TIMING DIAGRAM

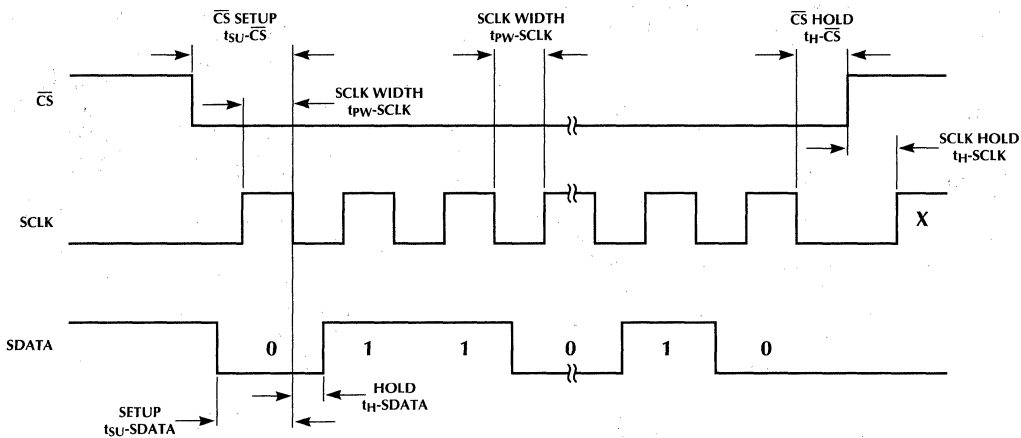
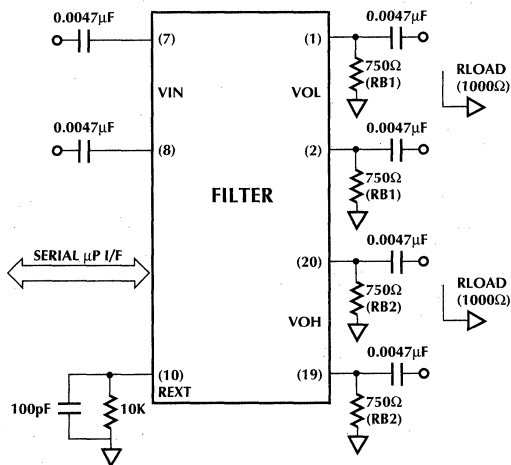


Figure 1.

CONTROL REGISTER DEFINITION

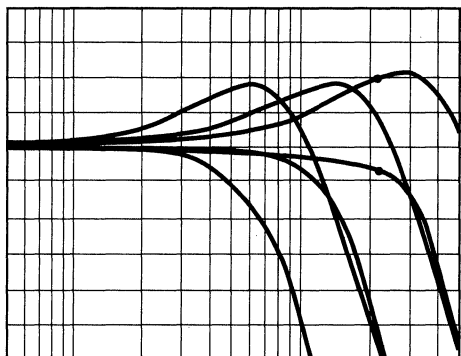
F13	F12	F11	F10	F9	F8	F7	F6	F5	F4	F3	F2	F1	F0
AZ	PD	RW	SLIMMING CONTROL				FREQUENCY CONTROL						
AZ	AutoZero		1 = Autozero circuitry activated 0 = Autozero circuitry inactive										
PD	Power Down		1 = Chip is in power down mode 0 = Chip is fully powered up										
RW	Read/Write		1 = Write data mode 0 = Read data mode										

APPLICATIONS



REF LEVEL /DIV
25.000dB 5.000dB
25.000dB 5.000dB

MARKER 21 848 415.800Hz
MAG (UDF) 9.617dB
MARKER 21 848 415.800Hz
MAG (D4) -3.166dB



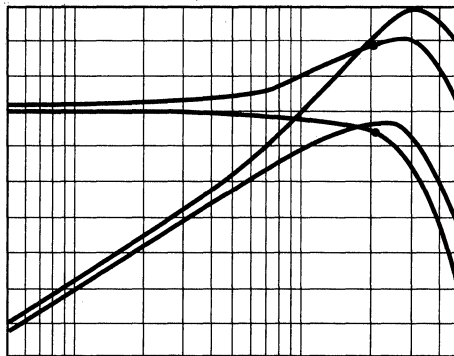
START 500 000.000Hz STOP 50 000.000Hz

Filter Response (Lowpass Output)

Shown are the ML6006 filter response at three different cutoff frequency (f_c) settings. At each of the f_c settings, the filter response is shown with no slimming and with full slimming activated.

REF LEVEL /DIV
15.000dB 5.000dB
15.000dB 5.000dB

MARKER 20 391 123.200Hz
MAG (UDF) 9.380B
MARKER 20 391 123.200Hz
MAG (D4) -3.190dB



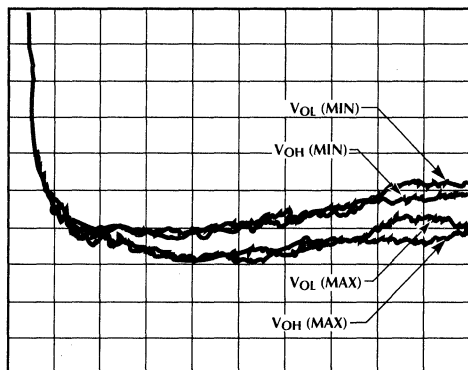
START 500 000.000Hz STOP 50 000.000Hz

Filter Response (Lowpass and Bandpass Outputs)

Shown are the ML6006 filter characteristic curves for the lowpass and bandpass outputs, with no slimming and full slimming activated.

REF LEVEL /DIV
31.000nSEC 1.000nSEC
31.000nSEC 1.000nSEC

MARKER 20 338 750.000Hz
DELAY (UDF) 29.593nSEC
MARKER 20 338 750.000Hz
DELAY (UDF) 30.446nSEC



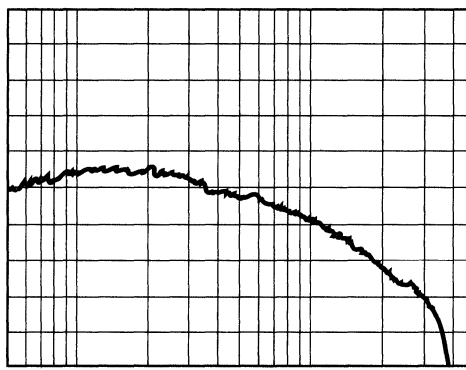
START 500 000.000Hz STOP 30 000.000Hz

ML6006 Filter/Equalizer Group Delay Tracking

Shown are the curves to demonstrate group delay tracking between the lowpass (V_{OL}) and bandpass (V_{OH}) outputs, with no slimming activated (min) and full slimming activated (max). It can be seen that the group delay tracking between the lowpass and bandpass outputs is well within 1 ns.

REF LEVEL /DIV
-90.000deg 0.500deg

MARKER 13 478 921.800Hz
PHASE (UDF) -90.659deg



START 500 000.000Hz STOP 50 000.000Hz

Phase Difference between Lowpass and Bandpass Outputs

Shown is the delta in the phase between the lowpass and bandpass outputs. Ideally the bandpass output should be -90° . The curve shows that this is within 1° for a frequency range of 50 MHz to 10 MHz.

ML6006

ORDERING INFORMATION

PART NUMBER	TEMPERATURE RANGE	PACKAGE
ML6005CR	0°C to +70°C	20-Pin SSOP (R20)

48 Mbps Read Channel Filter/Equalizer

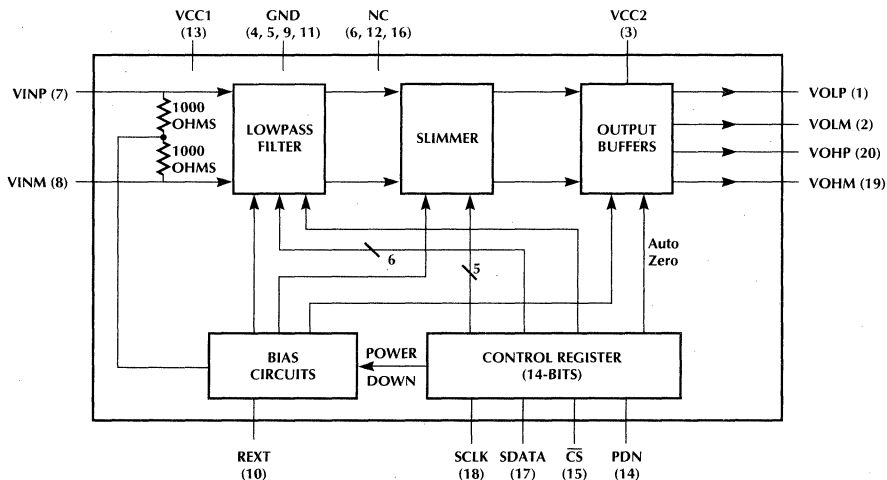
GENERAL DESCRIPTION

The ML6007 is a monolithic analog filter/equalizer intended for hard disk drive read channel applications, capable of handling disk data rates upto 48Mbits/s, with an operating power dissipation of less than 400mW. Its architecture consists of a continuous type filter based on a transistor and a high speed parasitic free active integrator, allowing complete independence of the filter response from interconnect parasitics, thus realizing a family of frequency response curves optimized for disk drive read channel equalization. It consists of a programmable 6-pole 2-zero lowpass filter stage, two pairs of high-speed drivers, and a serial microprocessor interface. The poles of the transfer function approximate a maximally flat group delay (modified Bessel type) response, whereas the symmetric zeros provide the high-frequency boost necessary for pulse slimming. The user can independently adjust both the corner frequency, as well as the slimming level. The desired frequency response is programmed by a 14-bit serial input data stream which includes one bit for power-down, one bit for read/write control, and one bit for auto-zero control. Real-time power down control is provided by a hardware pin. Metal mask options are available to configure this pin to realize Read/Write or Slimmer disable functions. The ML6007 is well suited for constant density recording systems (Zoned-bit recording) as well as for constant data rate systems.

FEATURES

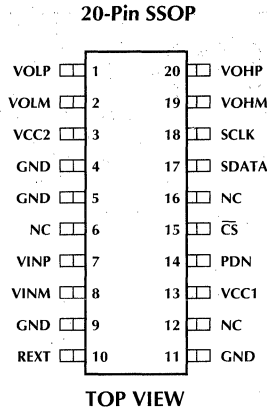
- 6-pole, 2-zero continuous time filter with < -45dB harmonic distortion
- Disk Data rates up to 48 Mbit/s
- Programmable filter cutoff frequency (2.8:1 range in 64 steps) ($f_c = 9.37$ to 27 MHz)
- 32 step programmable pulse slimming equalization, 0 to 10dB boost at f_c .
- Power-down, Auto-Zero, R/W modes programmable through bits in the Control Register
- Lowpass output and Differentiated Lowpass (Bandpass) output provided.
- Fully I/O balanced architecture with TTL/CMOS compatible interface
- High speed (upto 25MHz clock) three wire serial microprocessor interface
- Double buffered data latch for synchronous or asynchronous data loading.
- Single 5V \pm 10% power supply
- Available in 20-pin SSOP package.
- 4 GHz/1.5 μ BiCMOS process
- Power Dissipation: $P_{opr} = 400mW$, $P_{dn} = 10mW$

BLOCK DIAGRAM



ML6007

PIN CONNECTION



PIN DESCRIPTION

PIN#	NAME	FUNCTION	PIN#	NAME	FUNCTION
1	VOLP	Normal Lowpass outputs	15	\overline{CS}	Control Register Enable. A logical low level allows the SCLK input to clock data into the control register via the SDATA input line. A logical high level latches the control register contents and issues the information to the appropriate circuitry. A TTL input.
2	VOLM		17	SDATA	
3	VCC2	Positive supply for the output drivers, $5V \pm 10\%$	18	SCLK	Control Register Clock. Negative edge triggered control register clock input. A TTL input.
4, 5, 9, 11	GND	Ground	19	VOHM	Differentiated lowpass outputs
7	VINP	Signal Inputs	20	VOHP	
8	VINM		A 10K resistor between this pin and ground sets the filters corner frequency	6, 12, 16	NC
10	REXT				
13	VCC1	Positive supply, $5V \pm 10\%$			
14	PDN	Power Down Control pin. A low level input allows normal operation of the filter. A high level input puts the filter in the power down mode, a TTL input. Additionally a metal mask option is available to configure this pin as either READ/WRITE or FREQUENCY BOOST DISABLE.			

Dave McKay
Ram Gopalan

ML600X – Filter/Equalizer Eval Board

1.0 INTRODUCTION

The ML6005/6 eval board provides a vehicle to conveniently evaluate the ML6005/6 Filter/Equalizer chips, intended for hard disk read channel applications. Detailed specifications of the part can be found in the ML6005/6 data sheet. The eval board consists of three BNC connectors for the input signal to the filter, the normal lowpass filter output and the differentiated bandpass output. The programming of the filter is accomplished with an interface to an IBM PC through the standard parallel printer port. An executable program is provided, which provides the user interface to program the various parameters of the filter, through this parallel port interface.

2.0 EVALUATION BOARD DESCRIPTION & SETUP

The filter response from the demo board can be evaluated with the help of a network analyzer (NA). Refer to a detailed schematic of the board attached. The demo board has one analog input and two analog outputs. The BNC from the SOURCE output of the NA is connected to the input BNC, V_{IN} , of the demo board, which drives the one-to-one input transformer. The output side of the input path transformer is terminated with 50Ω to match the SOURCE, and is AC coupled to the input of the ML6005/6. The input of the ML6005/6 has an impedance of 1000Ω so the input corner frequency, with $C = 0.0047\mu F$, is 33.8kHz .

Both the outputs from the ML6005/6, the normal lowpass output and the differentiated bandpass output, are handled in the same way. Consider the low-pass, V_{OL} path. The plus and minus outputs are both biased by 750Ω and are then AC coupled to 1000Ω resistors in series with the transformer inputs.

The output side of the transformer goes directly to the NA where it is terminated by 50Ω . The 1000Ω resistors in series with the transformer inputs results in an attenuation, since the transformer reflects the NA 50Ω load back to the input side; $1/2$ of the 50Ω is in the plus-side circuit and $1/2$ is in the minus-side circuit so the attenuation is $25/(1025) = 1/41$ (-32.26dB). The NA gain/attenuation setting has to take this into account. This is for normalization and does not effect the shape of the frequency response curve. The eval board output coupling capacitors are $0.0047\mu F$ and therefore each has a break point with its 1000Ω series resistor at 33.8kHz , the same as the corner frequency of the input coupling capacitors.

The corner frequency and boost of the ML6005/6 are programmable through the serial interface. The upper left hand corner of the schematic shows the serial interface connection. As pointed out earlier, the programming of the ML6005/6 serial interface is done through the parallel port of an IBM PC. The interface program supplied with the demo board implements the three wire serial interface protocol to transmit the 14 control bits to the ML6005/6 internal shift register. The LSB is transmitted first. For details on the timing relationship of the serial interface, please refer to the data sheet. CS is a active low signal that is asserted before the first data bit is transmitted. This signals the start of data being clocked into the control register and is de-asserted after the last data bit is transmitted, which latches the data internally. SCLK is the clock signal of the three wire serial interface and has the negative edge active. **NOTE:** The last active low going edge must remain low at least 150ns after the CS signal is de-asserted. SDATA is the active high serial data line.

3.0 GETTING STARTED

To start evaluation of the ML6005/6 using the Eval board, the following equipment will be required :

- IBM PC-AT (or compatible) with 640K of RAM, a parallel printer port card and a mouse, (with a math co-processor, not mandatory)
- A parallel printer cable (shielded preferred) with a male connector on both ends.
- A +5 volt power supply and banana plugs.
- A HP 3577A Network Analyzer (or some other analyzer to see the frequency response).
- 3 BNC cables, short cables recommended
- ML6005/6 Eval board and interface program, (ver 2.0).

The socket on the demo board has 24 pins whereas the ML6005/6 is a 20-pin SSOP package. Hence, ensure that pin one of the part is justified to pin one of the socket. Holding the board with the socket lid opening to the right, pin one is on the top left corner of the socket. Set up the PC and connect the parallel cable from LPT1 or LPT2 to the demo board. Connect the +5V power and ground to the demo board with banana jacks. Connect the NA output or any other input source to V_{IN} of the demo board. Connect the NA inputs to V_{OL} and V_{OH} of the demo board.

5

ML600X User Guide

The demo program is provided on a diskette and comes with a number of files. It is important that all the files reside in the same directory that the ".exe" file is executed from. Hence follow these steps to install the program on drive C :

— type **A: <enter>**

— type **INSTALL.BAT <enter>**

The program will be installed on your C: drive and will automatically start execution. If the user quits from the program and wishes to restart the demo program again, then while in the **ml600x directory**

— type **uLDEMO.BAT <enter>**

The first screen (refer screen#1 in attached figure), comes up with the Micro Linear logo and asks the user to let the program know the parallel port selected (**LPT1** or **LPT2**) and the device under test (**ML6005** or **ML6006**). In addition the Eval board shows the frequency response of a 6th order Bessel including the effect due to the (0.0047 μ F) input and output coupling capacitors. It asks the user to program the start and stop frequency values for this frequency response display. **NOTE:** This ideal Bessel response is shown to give a feeling of the frequency response. The actual filter response is embellished by the finer realizations in actual silicon implementation. Having done that, click on the run box to get the operating screen which allows the user to change the parameters of the filter, or click on the quit box to end the program altogether.

The main operating screen (refer screen #2 in attached figure), allows the entry of three pieces of information; the mode bits, the boost values and the corner frequency values. The mode bits viz. power down, read/write, and autozero are shown as toggle switches and can be turned **on** or **off**, as desired, using the mouse. The cutoff frequency is seen as a sub-window in the main dialog window.

This sub-window shows the cutoff frequency in MHz, the 6 bit binary value of the cutoff frequency setting in the control register, and also the decimal value of this setting. The slimming boost options are seen as another sub-window in the main dialog window. This sub-window shows the boost values in dB, the five bit binary value of the boost setting in the control register, and also the decimal value of this setting. The values in the sub-windows can be changed by clicking on the up and down arrows seen at the edge of the box or if the user clicks once with the cursor inside the sub-window, a menu pops up showing the various values available for that sub-window. Keeping the mouse button pressed at the last line scrolls the sub-window. Moving the mouse one line up, with the button pressed, stops the scrolling. Move the mouse to the desired setting, while keeping the button pressed, and release for selecting that setting. In the bottom of the dialog box is a sub-window showing the current programmed contents of the control register. In addition, there is a **SERIAL SEND** box, clicking on which, sends the programmed information, over the serial

interface to the chip. However, if the **AUTO SEND** box is selected, then whenever any of the information is updated in the dialog window, the control register is automatically updated in a real-time manner. When the user clicks on the **PLOT** box, the frequency response is plotted on the screen, based on the current control register parameters. If the **AUTO PLOT** box is selected, then every time the control register parameters are updated, the plot on the screen is also automatically updated. The **ERASE** box is used to clear the plot on the screen. If the **OVERLAY** box is selected then the multiple frequency response curves are overlaid, one on top of the previous one, as long as the original scale can handle the responses. However if the most current response plot requires a change of scale on the Y-axis, then all the previous plots will be erased. Hence, if the previous plots need to be retained, even if the most current plot needs a different scale, then the **LOCK GRAF** box needs to be selected.

Clicking on the **RETURN** box, takes the user back to the first screen. It should be noted that the frequency response shown on the screen is an ideal Bessel type response, and hence is not an exact representation of the filter transfer function implemented inside the chip.

4.0 SAMPLE EVALUATION EXPERIMENTS

Caution : Before running any tests on the eval board using the Network Analyzer, be sure to normalize it for its own internal phase errors.

Normalization on the HP network analyzer is done by setting the receiver ATTN to 20dB and connecting the SOURCE via a BNC cable directly back to INPUT A. Set up the frequency range etc. and press TRIGGER to get a sweep and then press MEASR CAL. After calibration set the receiver ATTN to 0dB. Outlined below are some of the evaluation experiments.

4.1 POWER DOWN MODE TEST

Put a current meter in series with the demo board's +5V power supply and monitor the current in both the power down state and the non power down state. This measured current includes the drop through the external bias resistors. Therefore, to get an accurate value subtract the current due to the 750 Ω external bias resistors, approximately 10.7mA, and the 50 μ A current due to the pull-up resistor R9.

4.2 READ/WRITE CONTROL FUNCTION TEST

Set the RW bit on and off in the control register and measure the input impedance in each state. In the read mode, with the bit low, the differential input impedance is 2k, while in the write mode the impedance is 350 Ω . The Read/Write mode can be controlled through the serial interface or by the switch provided on the eval board (optional). The filter provides a clamping of the input to protect against transients generated during a write to read transition.

4.3 FREQUENCY RESPONSE MEASUREMENTS

At maximum f_C , no slimming and full slimming

At minimum f_C , no slimming and full slimming

— Normal lowpass output (V_{OL}), refer graph 1

— Differentiated Bandpass output (V_{OH}), refer graph 2

4.4 GROUP DELAY TRACKING

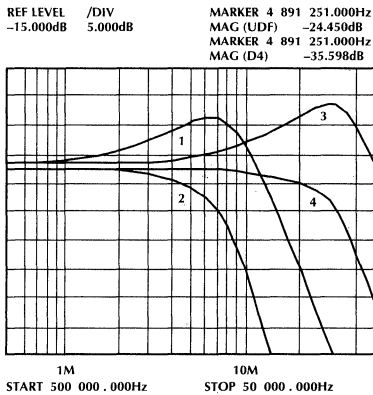
Tracking between the normal lowpass output (V_{OL}) and the differentiated bandpass output

(V_{OH}) for both cases of no slimming and full slimming.

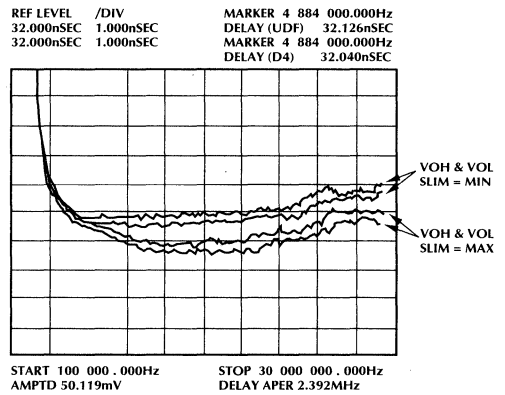
Refer to graph 3, showing that the group delay tracking is within 1ns within the frequency range of interest.

4.5 LOG PHASE

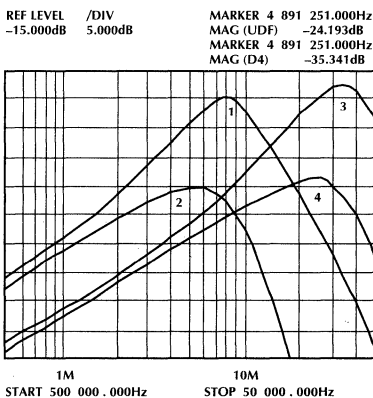
Phase Difference between the normal lowpass output and differentiated bandpass output. Refer to graph 4, which shows that the phase difference is actually within 1° for the frequency range of 5 to 20MHz.



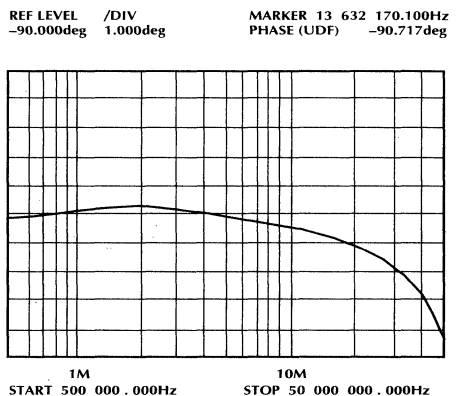
Graph 1
(Frequency Response - Lowpass)



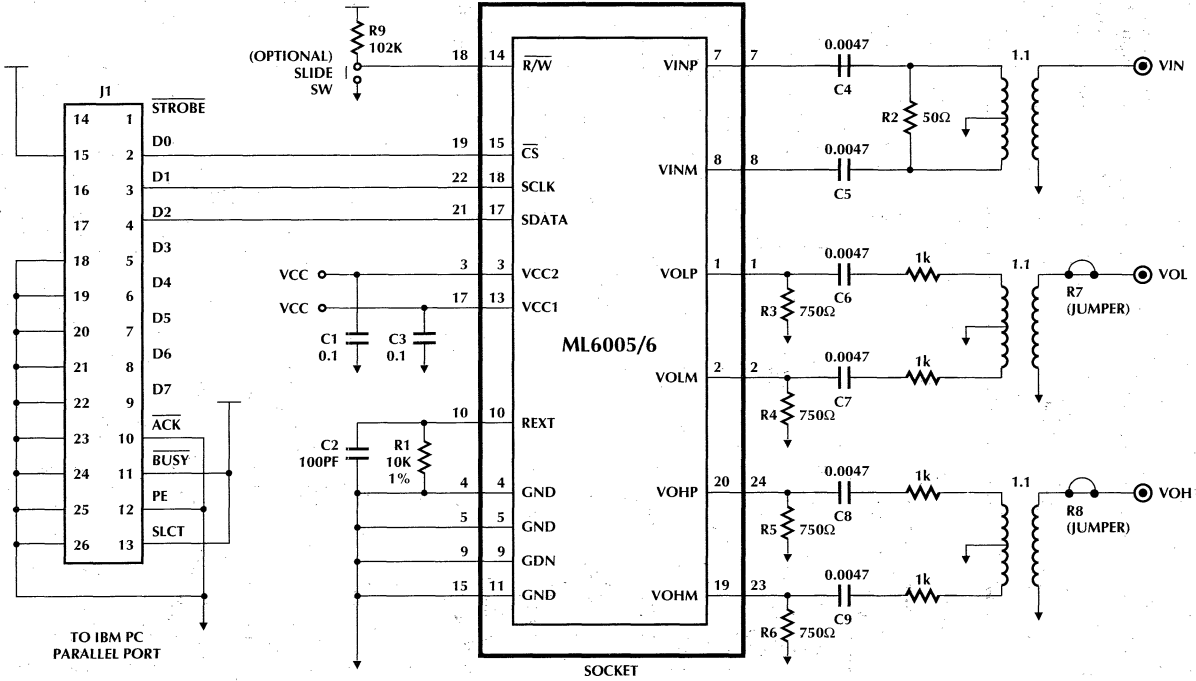
Graph 3
(Group Delay)



Graph 2
(Frequency Response - Bandpass)



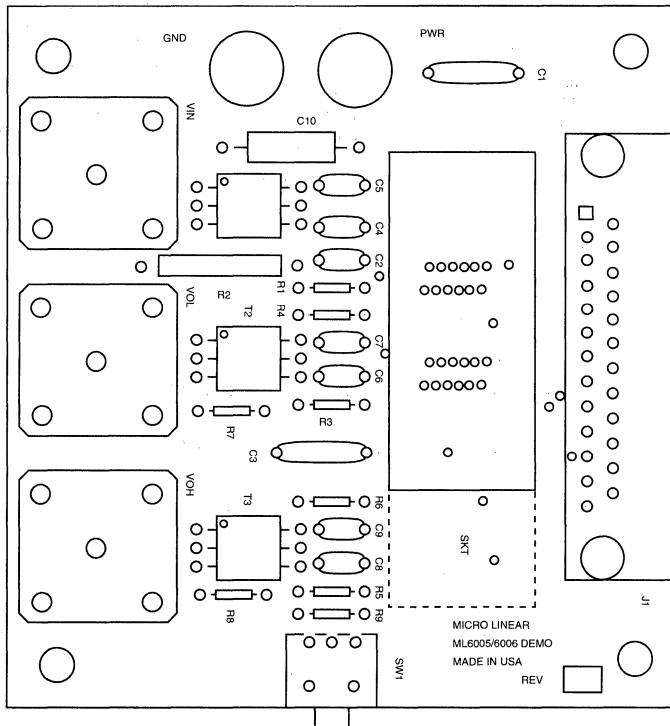
Graph 4
(Phase Error - V_{OH} vs. V_{OL})



ML6005/ML6006 Eval Board Schematic.


ML6005/6 EVAL BOARD BOM

DESCRIPTION	VENDOR	PART#	QUAN
Transformer T1, T2, T3	Mini-circuits	T1-6T	3
25 pin D-shell connector, J1	Pan Pacific	DRB-25SR	1
BNC, right angle, female, PC mount	Pomona	model 4788	3
Power inputs, banana jacks	Pomona		2
Slide Switch, SW1	Alco	tss11	1
Resistors, metal film 1/4 watt			
R2	50Ω		1
R3,R4,R5,R6	750Ω		4
Added transformer input resistors (in series with C6, C7, C8, C9)	1000Ω		4
R7, R8 (now wire jumperd)			
Capacitors			
C10, Tantalum	1.0μF		1
C1, C2 ceramic	0.1μF		2
C6,C7,C8,C9	0.0047μF		4
C2	100pF		1
IC - ML6006	Micro Linear		1



Eval Board Component Layout

ML600X User Guide



Micro Linear
2092 Concourse Dr. San Jose, CA 95131

*** WELCOME to MICRO LINEAR's FILTER DEMO ***

Select printer port for Demo Board:

LPT1

LPT2

Set 'Ideal Bessel Plot' Param's:


Start Frequency:

Stop Frequency:

Select Device:

ML6005

ML6006



Micro Linear
2092 Concourse Dr. San Jose, CA 95131

* ML6005/6 Programmable Electronic Filter *

Gain @ Fc (dB)

10 9 8 7 6

00 -3.0 0 0 0 0

Cutoff Frequency (MHz)

5 4 3 2 1 0

00 20.25 0 0 0 0

Bit 13 Auto Zero On Off

Bit 12 PDB On Off

Bit 11 RW Write Read

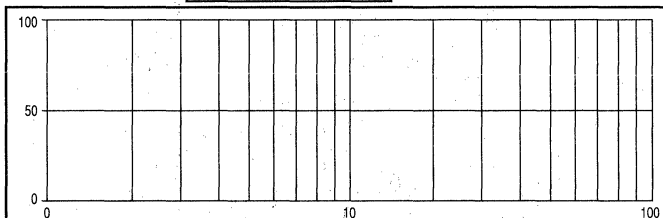
Buffer

N K RW PDB AZ

00 00 0 0 1

Auto Send Auto Plot

*** Ideal Bessel Filter Response ***



Over Lay

Lock Grf

ML6006/ML6005 Eval Board User Interface Templates.

Integrated Disk Read Channel Processor

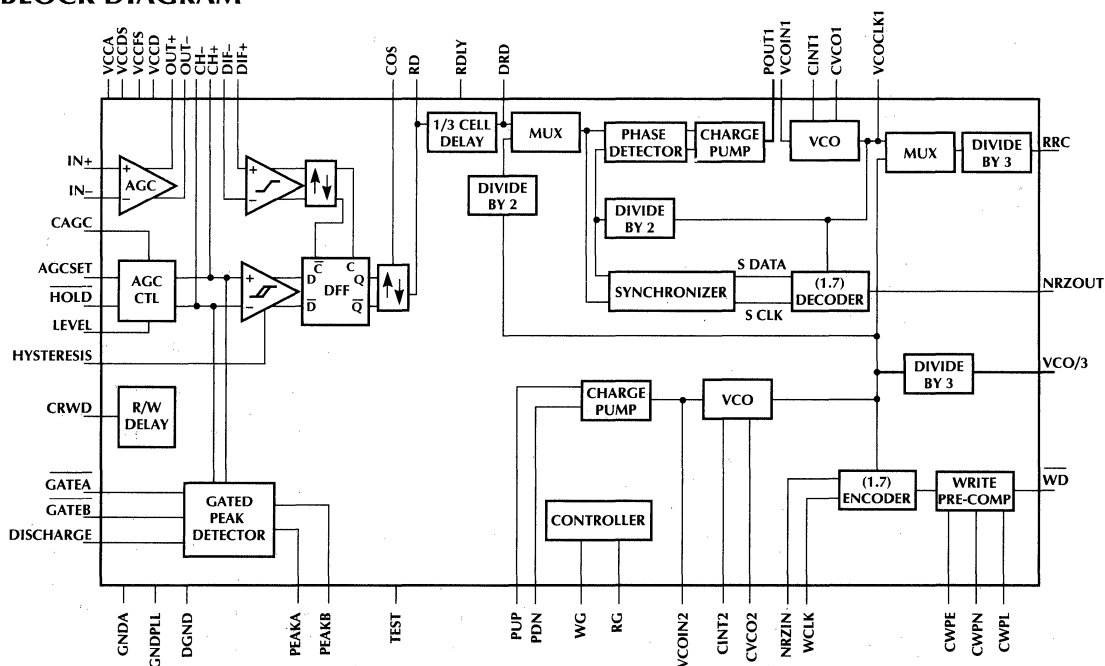
GENERAL DESCRIPTION

The ML6010 Integrated Disk Read Channel Processor, incorporates a pulse detector, two channel gated servo peak detectors, a data synchronizer, a partial frequency synthesizer, a (1,7) RLL encoder/decoder and write precompensation circuitry onto a single chip, providing a complete read channel solution for hard disk drive subsystems. The chip receives serial NRZ data from the disk controller, encodes the data into (1,7) RLL code with write precompensation and writes to the disk. In the read mode it peak detects preamplified read pulses for both data and embedded servo information, resynchronizes the data, and decodes the (1,7) RLL back to NRZ. The chip operates at data rates up to 36 Mb/s with three levels of write precompensation, set with an external capacitor and a pulse detector with less than 1 ns of pulse pairing. The ML6010 supports constant density recording (CDR) applications with an onboard charge pump and VCO for the frequency synthesizer and requires an external phase detector and M & N dividers to realize a complete frequency synthesizer. It is set to interface directly to ML6005/6 family of BICMOS filter/equalizer chips with programmable cutoff frequency and pulse slimming (equalization) capability.

FEATURES

- NRZ data rates up to 36 Mb/s
- Single +5 volt power supply +/- 5%
- Operating power dissipation 700mW
- Industry standard pulse detector circuitry with less than 1 ns pulse pairing
- Pattern—insensitive wide bandwidth AGC amplifier
- Two channel gated servo peak detectors for embedded servo recovery
- Industry standard fast acquisition PLL with zero phase start capability
- VCO/Charge pump has greater than 3:1 tuning range
- On-board frequency synthesizer charge pump and VCO for addressing ZBR applications
- Interface to industry standard channel filtering/equalizer chips like ML6005/6
- Three level Write Data Precompensation support
- (1,7) RLL encoding/decoding support
- Available in a 52 pin QFP package

BLOCK DIAGRAM



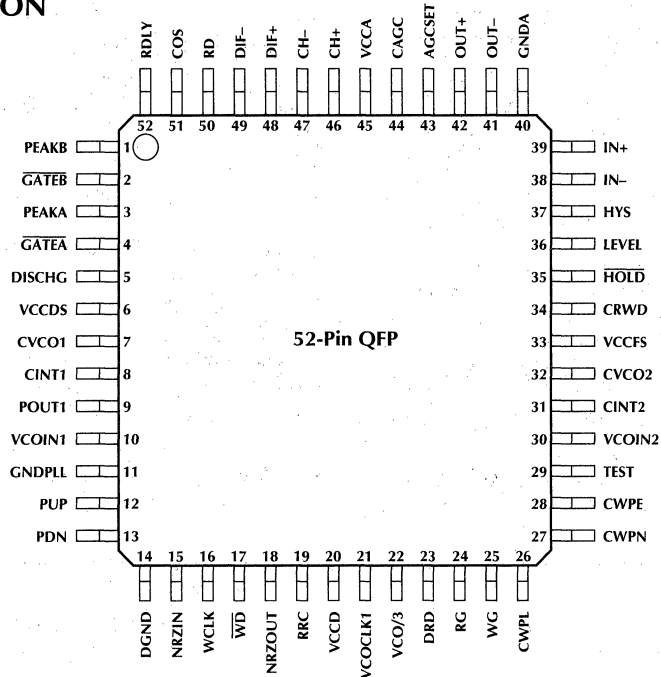
ML6010

GENERAL DESCRIPTION (Continued)

The ML6010 is fabricated in a BiCMOS process (4GHz npn f_t ; 1.5 μ CMOS) and operates off of a single 5 volt supply. The ML6010 is based on a semi-standard tile array (FC3560) with built in uncommitted gain stages for the flexibility of user defined channel pulse detector and/or passive differentiator or matched delay applications and a

digital gate array for adding user specific functions, thus allowing for customizable options of this feature set, based on the user requirements. Some examples are the optional servo demod outputs—unbuffered A,B,C,D, or buffered A, B and A-B outputs; a 2,7 RLL Endec instead of the 1,7 RLL Endec or maybe implement the M & N dividers instead of the Endec, digital glue, etc.

PIN CONNECTION



PIN DESCRIPTION

PIN#	NAME	FUNCTION	PIN#	NAME	FUNCTION
1	PEAKB	A capacitor or a capacitor in series with a resistor, connected between this pin and GNDA functions as a servo sample and hold for channel B.	5	DISCHG	TTL input (active high). When this pin is forced to a TTL high, both the servo peak detectors are discharged.
2	GATEB	TTL input (active low). When this pin is forced TTL low, servo peak detector B is enabled.	6	VCCDS	Analog + 5 volt supply for data synchronizer related blocks.
3	PEAKA	A capacitor or a capacitor in series with a resistor, connected between this pin and GNDA functions as a servo sample and hold for channel A.	7	CVCO1	Capacitor between this pin and VCCDS sets up the VCO center frequency for the data synchronizer.
4	GATEA	TTL input (active low). When this pin is forced TTL low, servo peak detector A is enabled.	8	CINT1	Coarse input for data synchronizer loop filter time constant setting.
			9	POUT1	Data synchronizer's charge pump output, drives the loop filter input.
			10	VCOIN1	Data synchronizer's VCO control input, driven by the loop filter output.

PIN DESCRIPTION (Continued)

PIN#	NAME	FUNCTION	PIN#	NAME	FUNCTION
11	GNDPLL	Analog ground for data synchronizer and frequency synthesizer.	27	CWPN	RC network to setup the normal delay time constant for write precompensation.
12	PUP	Charge pump input for the frequency synthesizer (pump up).	28	CWPE	RC network to setup the early delay time constant for write precompensation.
13	PDN	Charge pump input for the frequency synthesizer (pump down).	29	TEST	Reserved for test purposes. Must be tied to DGND for normal operation.
14	DGND	Digital ground.	30	VCOIN2	Fine input for the frequency synthesizer loop filter time constant setting.
15	NRZIN	NRZ write data input from disk controller. This pin can be connected to the NRZOUT pin to form a single bi-directional NRZ port if desired.	31	CINT2	Coarse input for frequency synthesizer loop filter time constant setting.
16	WCLK	Write clock input synchronous with the NRZ Write data input.	32	CVCO2	Capacitor between this pin and VCCA sets up the VCO center frequency for the frequency synthesizer PLL.
17	\overline{WD}	Write precompensated, active low (1,7) RLL encoded write data output to the read/write amplifier.	33	VCCFS	Analog +5 volt supply for frequency synthesizer.
18	NRZOUT	NRZ read data output to the disk controller. This pin can be connected to the NRZIN pin to form a single bi-directional NRZ port if desired.	34	CRWD	Capacitor between this pin and GNDA determines the write to read (input clamp) delay time, e.g. a 1800 pF capacitor gives a delay time of 3 us typ.
19	RRC	Read/Reference clock : a multiplexed clock source used by the disk controller. During mode change there will be no glitches on this line and no more than two lost clock pulses will occur. When Read Gate goes high, RRC synchronized to the NRZ read data will be available after 19 read data pulses.	35	\overline{HOLD}	TTL input pin (active low). When this pin is forced low, all the charging and discharging paths on the CAGC pin are disabled. The AGC amplifier now acts as a fixed gain amp. with the gain determined by the voltage on the CAGC pin.
20	VCCD	Digital VCC +5 volts.	36	LEVEL	Emitter follower output, provides rectified signal level, which can be used through a resistor divider network as a dynamic hysteresis control signal to the hysteresis pin.
21	VCOCLK1	An open emitter ECL output for testing purposes.	37	HYS	The voltage applied to this pin determines the threshold level of the qualification channel. In a typical application, the signal on this pin is derived from the LEVEL pin.
22	VCO/3	Divide by three output of the frequency synthesizer VCO clock output.	38	IN-	AGC amplifier differential input, AC coupled from the R/W amplifier. Signal range recommended 15 mV to 150 mVp-p differential.
23	DRD	1/3 cell delayed read data for testing purposes.	39	IN+	
24	RG	Active high read gate input from the disk controller. This signal is used to select the PLL reference input.	40	GNDA	Analog ground.
25	WG	Active high write gate input from the disk controller. This signal is used to enable the write mode.			
26	CWPL	RC network to setup the late delay time constant for write precompensation.			

ML6010

PIN DESCRIPTION (Continued)

PIN#	NAME	FUNCTION	PIN#	NAME	FUNCTION
41	OUT-	AGC amplifier differential outputs, AC coupled to ML6005/ML6006 filter chip. These pins have a current sink capability of 2 mA typical.			of the ML6005/ML6006 filter/equalizer chip. Peak to Peak differential signal at these pins are determined by the DC voltage applied to the AGCSET pin.
42	OUT+				
43	AGCSET	Voltage on this pin sets up the peak to peak differential voltage at the CH+/CH- pins when the AGC amplifiers are settled. $V_{p-p} \text{ diff @CH+/CH-} = 4 * (V_{AGCSET} - V_{CCA}/2)$	48	DIF+	Time Channel zero crossing comparator differential inputs. These are AC coupled from the differentiated (bandpass) outputs of the ML600X filter/equalizer chip.
			49	DIF-	
44	CAGC	AGC loop capacitor to GNDA. Lead-Lag network may be used for different loop filter characteristics, if needed.	50	RD	Pulse detector raw read data output. This signal internally goes to the 1/3 cell delay block enroute to the data synchronizer. This is an ECL output.
45	VCCA	Analog +5 volt supply for pulse detector and servo peak detector.	51	COS	The capacitor between this pin and the VCCA sets up the raw read data pulse width from the pulse detector section to the data synchronizer.
46	CH+	Gating channel differential inputs to hysteresis comparator, AGC fullwave rectifier and two-channel servo peak detector. These are AC coupled from the lowpass output	52	RDLY	Resistor between this pin and VCCDS sets the charging current in 1/3 cell delay based on data rate range.
47	CH-				

ABSOLUTE MAXIMUM RATINGS

DC Supply Voltage (VCCD)	-0.3 to +7 V
TTL Output Voltage (Vout)	5.5 V
TTL Input Voltage (Vin)	5.5 V
Differential Input Voltage (Vdif)	3.0 or -3.0 V
Analog Inputs	-0.3 to VCCA + 0.3 V
Storage Temperature (Tstg)	-65 to +150 °C
Maximum Junction Temperature (Tjmax)	125°C

OPERATING CONDITIONS

DC Supply Voltage (VCCD)	5 +/- 5% VDC
DC Supply Voltage (VCCA)	5 +/- 5% VDC
Operating Temperature Range	0 to 70°C

ELECTRICAL CHARACTERISTICS

PARAMETER	CONDITIONS	MIN	TYP.	MAX	UNITS
AGC Amplifier and I/P Z-Clamp					
AGC 3dB bandwidth (Note 1)	VCAGC = 2V, VIN(Diff) = 15 mVp-p		60		MHz
Maximum AGC gain	VCAGC = 2V, VIN(Diff) = 15 mVp-p	100	150		V/V
Minimum AGC gain	VCAGC = 3.5V, VIN(Diff) = 150 mVp-p		5	10	V/V
AGC Control Range			27		dB
AGC Control Sensitivity			37		dB/V
Input Signal Range (Differential)		15		150	mV
Output Signal Swing (Differential)	VCAGC = 2.0V, VIN(Diff) = +/- 100 mVp-p		1.5	2.5	Vp-p

ELECTRICAL CHARACTERISTICS (Continued)

PARAMETER	CONDITIONS	MIN	TYP.	MAX	UNITS
AGC Amplifier and I/P Z-Clamp (continued)					
Input Voltage Noise (Note 1)	VIN = 0 V		20		nV/√Hz
Differential Input Resistance	Read Mode Vin+ = 2.575V Vin- = 2.425	4.5	6.0	7.5	kohm
Differential Input Resistance	Write Mode Vin+ = 2.575V Vin- = 2.425	200	300	400	ohm
Common Mode Rejection Ratio (Note 1)	Vin+ = Vin- = 100 mV VCAGC = 2V		45		dB
AGC offset (I/P referred)	VIN+ = VIN- = VCCA/2 VCAGC = 2 V	-4		4	mV
Power supply Rejection Ratio	VCCA = 5V + 100 mVp-p VCAGC = 2V,		45		dB
AGC output common mode bias	VCCA = 5V, Vin+ = Vin-		2.7	3.0	V
AGC capacitor bias current	VCAGC = 3.5 V		1	20	μA
AGC input offset voltage	VCAGC=2 V, Vin+ = Vin-			400	mV
AGC output sink current	VCCA = 5V	1	2	3	mA
AGC output resistance	Measured on OUT+/OUT-		40		ohms
CAGC voltage range		2.2		3.4	V
Clamp Off time delay (Note 1)	Crwd=1800 pF, VCAGC=2.0V, Vin (diff)=5 mV	2	2.5	3.0	μs
Clamp Propagation Delay R W (Note 1)	Crwd=1800 pF, VCAGC=2.0V, Vin (diff)=5 mV		15	100	ns
AGCVOS vs Gain	VCAGC=VCC/2 to VCC/2 + 0.25V	-400		400	mV
AGC Rectifier and Comparator					
Input signal range (Note 1)				2.0	Vp-p
Input Bias voltage		40	50	60	%VCCA
Input resistance (differential)		3.75	5	6.25	Kohms
CAGC Output voltage	@CAGC, Vin > VAGCSET RCAGC = 100K VAGCSET = VCCA/2 + 100mV		VCCA- .8		V
CAGC charge resistance	Vin > VAGCSET VAGCSET = VCCA/2 + 100mV		300		ohms
CAGC clamped voltage (CAGCLMPV)	Vin < VAGCSET VAGCSET = VCCA/2 + 100mV		2.2		V
AGCSET bias current	VAGCSET = 2.5 V		1.5	10	μA
CAGC Decay current	Read , VCAGC=VCCA		4.7		μA
CAGC leak current	Hold Mode VCAGC=CAGCLMPV + 0.4V		0.03	0.5	μA
Hold On & Hold Off time				1	μs

ML6010

ELECTRICAL CHARACTERISTICS (Continued)

PARAMETER	CONDITIONS	MIN	TYP.	MAX	UNITS
Zero Crossing Comparator					
Differential input signal range (Note 1)				1.5	Vp-p
Differential input bias current	@ VCCA/2			20	μA
Diff comparator offset voltage	HYS = 0V	-3.0		+ 3.0	mV
Differentail input resistance		4.5	6.0	7.5	kohms
Differential bias voltage		40	50	60	%VCCA
Zero crossing comparator gain (Note 1)			65		dB
Hysteresis Comparator					
HYS input signal range (Note 1)				1.5	Vp-p
Peak Hysterisis vs HYS voltage		0.44	0.5	0.56	V
HYS bias current				20	μA
LEVEL pin max output current		2			mA
LEVEL pin output resistance			130		ohms
Internal and Pulse Qualification One-shot					
Internal one-shot (t _{pw}) (Note 1)			15		ns
Pulse Qualification one-shot (t _{pw})	Cos = 47 pF		35		ns
Pulse Detector raw data VOH			VCCA-2Vbe		V
Pulse Detector raw data VOL			VOH - 0.73		V
Pulse Pairing V _{diff} =1Vp-p diff @ 5 MHz			0.5	1.0	ns
Data Sperator PLL					
Phase Detector range			+/- pi		radians
Charge pump gain	V/I - VCOIN1 @DSVCC/2		125		μA
Data Synch fVCO max	CVCO1 = 10pF VCCDS RG = 1, WG = 0, VCINT1 = VCCDS - 1 VVCOIN1 = VCCDS/2	108			MHz
Data Synch VCO range	CVCO1 = 10pF VCCDS RG = 1, WG = 0, VCINT1 = 1V to 4.2V	3:1	4:1		
Data Synch VCO Course Gain	CVCO1 = 10pF VCCDS RG = 1, WG = 0, VCINT1 = 2V to 2.5V VVCOIN1 = VCCDS/2		300		Mrad/s /V
Data Sync VCO fine Gain	CVCO1 = 10pF VCCDS RG = 1, WG = 0, VCINT1 = VCCDS/2 VVCOIN1 = 2V to 3V		TBD		Mrad/s /V
PLL jitter specifications			0.7		ns

ELECTRICAL CHARACTERISTICS (Continued)

PARAMETER	CONDITIONS	MIN	TYP.	MAX	UNITS
Data Separator PLL (continued)					
Data sync window center offset			1%Tw +/-2		ns
Maximum data rate (1,7 RLL)	CVCO1 = 10pF VCCDS RG = 1, WG = 0,		28	36	Mbits/s
Frequency Synthesizer PLL					
Charge pump gain	V/I - VCOIN2 @VCCFS/2		125		μA
Freq Synthesizer fVCO max	CVCO2 = 10pF VCCFS RG = 0, WG = 1, VCINT2 = VCCFS -1 VVCOIN2 = VCCFS/2	108			MHz
Freq Synthesizer VCO range	CVCO2 = 10pF VCCFS VCINT2 = 1V to 4.2V	3:1	4:1		
Freq Synthesizer fine VCO gain	CVCO2 = 10pF VCCFS VVCOIN2 = 2V to 3V		TBD		Mrad/s /V
Freq Synthesizer coarse VCO gain	CVCO2 = 10pF VCCFS VCINT2 = 2V to 2.5V VVCOIN2 = VCCFS/2		300		Mrad/s /V
Servo Peak Detector					
Input signal range (differential) (Note 1)				1.5	Vp-p
Servo demod voltage gain			2		V/V
Offset mismatch between two chnls	Vin = 1.5 Vp-p differential	-20		+20	mV
Output leakage current (ch disabled)	GATE A/B =1, DISCHG=0			10	μA
Servo demod discharge current	R = 750 ohms		650		μA
Miscellaneous					
Operating power dissipation			750	1000	mW
Write Precompensation time	Cwpm=20pF, Cwpl=39pF fWCLK = 36 MHz		2		ns
Write Precompensation time	Cwpm=100pF, Cwpl=200pF fWCLK = 7.2 MHz		20		ns
TTL compatible inputs and outputs					
Input low voltage (VIL)		-0.3		0.8	V
Input high voltage (VIH)		2.0		VCC + 0.3	V
Input low current (IIL)	VIL = 0.4V			0.4	mA
Input high current (IIH)	VIH = 2.7V			100	μA
Output low voltage (VOL)	IOL = 1 mA			0.4	V
Output high voltage (VOH)	IOH = -400 μA	2.4			V

Note 1: These parameters are guaranteed by design. They are not 100% tested and are not in outgoing quality level calculation.

ML6010

FUNCTIONAL DESCRIPTION

The ML6010 provides the integration of most of the functions associated with the implementation of disk read channel design up to 36 Mbits/s data rates. It incorporates a pulse detector, two gated servo peak detectors, a data separator with fast acquisition capability, the charge pump and VCO functions for implementing a frequency synthesizer, write precompensation circuitry and a (1,7) RLL Endec. It is targeted at one/two platter 3 1/2" and high capacity 2 1/2" drives, where performance and capacity requirement take priority to power requirements, although the overall power requirements are much lower than earlier discrete block implementations.

The **Pulse Detector** section includes a wide bandwidth differential amplifier with automatic gain control, a precision full wave rectifier, time channel and gate channel. User programmable equalization or pulse slimming, and CDR band selection is supported through an external filter chip (ML6006). The Pulse detector will support pulse pairing specifications less than +/- 1ns.

Two Gated Servo Peak Detectors are incorporated for recovery of embedded servo information. Optionally it can provide buffered (A,B and a position error signal—PES) low impedance outputs which represent the peak detected level of each servo burst. These voltages are suitable for digitizing by an A/D converter and processed by the controlling processor, for head positioning.

The **Data Synchronizer** incorporates a fast acquisition phase lock loop with zero phase start capability and a 3:1 tuning range. Precise decode window control is

achievable using external components and the 1/3 cell delay automatically tunes to the synthesized frequency depending on the zone. The settling time is typically less than 2 μ s which is well within the requirements, e.g. for 36 Mbps data rate with 8 bytes of 3T preamble, and 1,7RLL code the minimum settling time required equals :

$$1/36 * 2 * 8 * 8 = 3.55 \mu s$$

The partial **Frequency Synthesizer** generates all necessary clocks for data encoding and synchronizer reference. The synthesizer requires external logic for the input divider (N) and feedback divider (M) programming and also uses an external loop filter, giving the user full control over the PLL's dynamics.

The **Endec** employs the 2/3 (1,7) RLL code type and supports a hard sectored drive implementation. After the index/sector pulse has been detected, an internal counter counts negative transitions of the incoming read data looking for three consecutive "3T" pattern. Once detected the VCO lock process is established. The **Write Precompensation** circuitry, provides control of the normal, early and late settings.

The ML6010 is fabricated in a BiCMOS process (4GHz npn f_t ; 1.5 μ CMOS) and operates off of a single 5 volt supply and is based on the **FC3560 semi-standard array**. The tile array allows a number of configurable features to realize a ML6010 like read channel combo chip, with customized features and pinout. For more information on the configurable options contact the factory:

ORDERING INFORMATION

PART NUMBER	TEMPERATURE RANGE	PACKAGE
ML6010CG	0°C to +70°C	52-Pin QFP (G52)

Pulse Detector

GENERAL DESCRIPTION

The ML8464 is a Pulse Detector designed for use in magnetic disk applications to detect the amplitude peaks on the output of the read/write amplifier. These signal peaks are caused by flux reversal on the disk media, which when connected to the read/write amplifier result in an output consisting of a series of pulses of alternating polarity. The relative time position of these signal peaks is indicated by the leading edge of the TTL output pulses. The Pulse Detector accurately represents the time position of these peaks.

The ML8464 contains three major blocks. The amplifier block contains a wide bandwidth differential amplifier with Automatic Gain Control (AGC) and a precision full wave rectifier. The time channel block includes a programmable differentiator followed by a bidirectional one shot multivibrator. The gate channel block includes a differential comparator with programmable hysteresis, a D flip-flop and an output bi-directional one shot multivibrator. The ML8464C internally connects the time channel output to the D flip-flop.

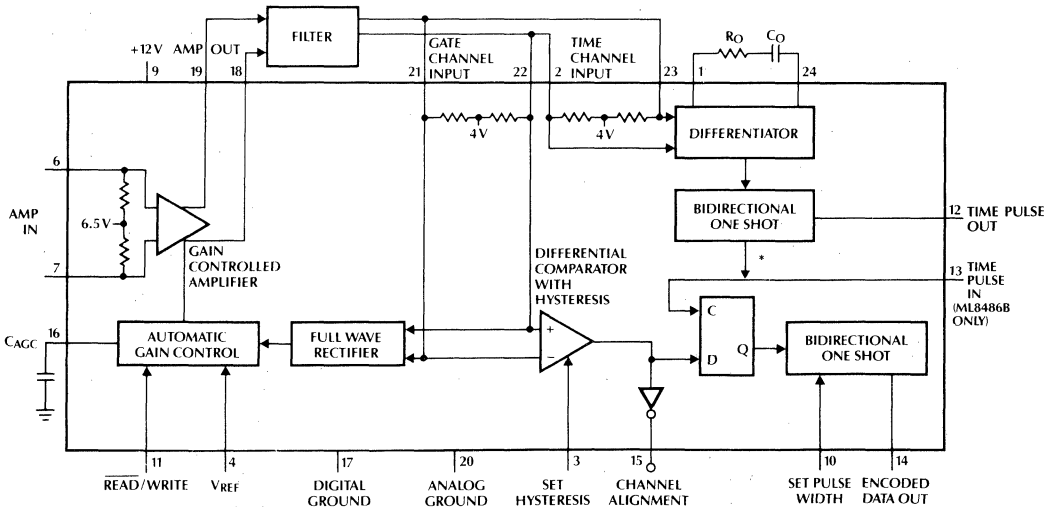
FEATURES

- Wide differential input signal range 20–660 mV_{p-p}
- TTL compatible digital Inputs and Output
- Externally gain controlled input differential amplifier
- Variable hysteresis comparator with gating circuitry
- Differentiator with externally programmable time constants
- Standard 12V power requirement
- Available in 24-pin DIP package, or a 28-pin surface mount PCC
- Improved pulse pairing (± 1 ns max.)
- Handles RLL (1, 7) or (2, 7) data to 24 MB/s

ML8464B FEATURES

- Direct replacement for DP8464B

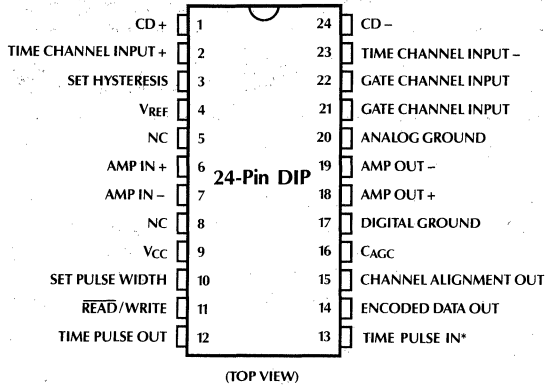
BLOCK DIAGRAM



* ML8464C ONLY

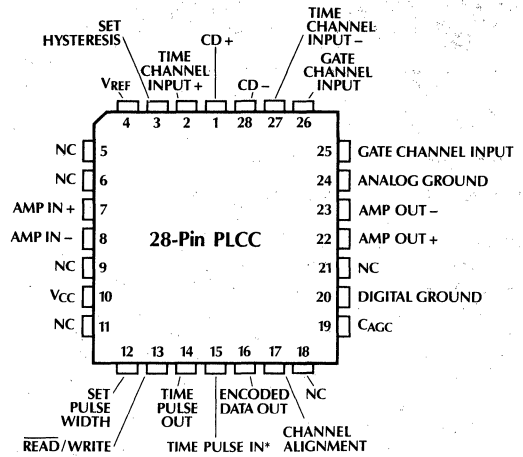
ML8464B, ML8464C

PIN CONNECTIONS



NC = No Connect

* THIS PIN IS A NO CONNECT ON THE ML8464C.



PIN DESCRIPTION

NAME	FUNCTION	NAME	FUNCTION
Amp In+, Amp In-	Differential inputs to the Amplifier. The output of the read/write head amplifier should be capacitively coupled to these pins.	Set Pulse Width	External capacitor between this pin and Digital ground is connected to control the pulse width of the Encoded Data Out.
Amp Out+, Amp Out-	Differential outputs of the Amplifier. These outputs should be capacitively coupled to the gating channel filter and to the time channel filter.	Read/Write	TTL input. When low, the chip is in read mode and active. When High, the chip is forced into stand by mode.
Gate Channel Inputs	Differential inputs to the AGC block and the gating channel. Must be capacitively coupled from the Amp Out.	Channel Alignment	Buffered output of the differential comparator with hysteresis. This output is TTL on the ML8464B, and is open emitter on ML8464C. The ML8464C is specified with a 2K Ω pull-down resistor to ground.
Time Channel Input+, Time Channel Input-	Differential inputs to the time channel differentiator. A filter is required between these pins and Amp Out pins to band limit the noise and to correct for any phase distortion due to read circuitry. Also inputs must be capacitively coupled to prevent disturbing the DC input level.	Time Pulse In (ML8464B only)	This is the TTL input to the clock of the D flip-flop. Usually it is connected to the Time Pulse Out pin.
CD+, CD-	External differentiator network is connected between these two pins.	Time Pulse Out	ML8464B: This is the TTL output from the bidirectional one shot following the differentiator. Usually it is connected to the Time Pulse In pin. ML8464C: Open emitter-follower test point.
Set Hysteresis	DC voltage on this pin sets the amount of hysteresis on the differential comparator.	Encoded Data Out	TTL output. Leading edge of this pin indicates the time position of the peaks.
V _{REF}	AGC circuit adjusts the gain of the amplifier to make the differential peak to peak voltage on the Gate Channel. Input is four times the DC voltage on this pin.	V _{CC}	12V power supply.
C _{AGC}	External capacitor between this pin and Analog ground is connected for the AGC.	GND	Digital ground. Digital signals should be referenced to this pin.
		AGND	Analog ground. Analog signals should be referenced to this pin.

FUNCTIONAL DESCRIPTION

The output from the read/write amplifier is AC coupled to the amp input of the ML8464. The amplifier's output voltage is fed back via an external filter to an internal fullwave rectifier and compared against the external voltage on the V_{REF} pin. The AGC circuit adjusts the gain of the amplifier to make the peak to peak differential voltage on the Gate Channel Input four times the DC voltage on the V_{REF} . Typically the signal on the amp out will be set for $4V_{P-P}$ differential. Since the filter usually has a 6dB loss, the signal on the Gate Channel Input will be $2V_{P-P}$ differential. The user should therefore set 0.5V on V_{REF} which can be done with a simple voltage divider from the +12V supply or other suitable reference.

The peak detection is performed by feeding the output of the amplifier through an external filter to the differentiator. The differentiator output changes state when the input pulse changes direction, generally this will be at the peaks. However, if the signal exhibits shouldering, the differentiator will also respond to noise near the baseline. To avoid this problem, the signal is also fed to a gating channel which is used to define a level either side of the baseline. This gating channel is

comprised of a differential comparator with hysteresis and a D flip-flop. The hysteresis for this comparator is externally set via the Set Hysteresis pin. In order to have data out, the input amplitude must first cross the hysteresis level which will change the logic level on the D input of the flip-flop. The peak of the input signal will generate a pulse out of the differentiator and bidirectional one shot. This pulse will clock the new data at the D input through to the output. In this way, when the differentiator is responding to noise at the baseline, the output of the D flip-flop is not changing since the logic level into the D input has not been changed. The comparator circuitry is therefore a gating channel which prevents any noise near the baseline from contaminating the data. The amount of hysteresis is twice the DC voltage on the Set Hysteresis pin. For instance, if the voltage on the Set Hysteresis pin is 0.3V, the differential AC signal across the gate channel input must be larger than 0.6V before the comparator will change states. In this case, the hysteresis is 30% of a 2V peak to peak differential signal at the gate channel input.

ABSOLUTE MAXIMUM RATINGS

Supply Voltage	14V
TTL Input Voltage	5.5V
TTL Output Voltage	5.5V
Input Voltage	5.5V
Differential Input Voltage	+3V
θ_{JA} for 24-Pin Plastic DIP (Copper Lead Frame)	60°C/Watt
θ_{JA} for 28-Pin PLCC (Copper Lead Frame)	60°C/Watt
Storage Temperature Range	-65°C to +150°C

Absolute maximum ratings are those values beyond which the device could be permanently damaged. Absolute maximum ratings are stress ratings only and functional device operation is not implied.

ML8464B, ML8464C

DC ELECTRICAL CHARACTERISTICS

Over recommended operating conditions of $T_A = 0$ to 70°C , $V_{CC} = 12.0\text{V} \pm 10\%$, $V_{REF} = 0.5\text{V}$, Set Hysteresis = 0.3V , Read/Write = 0.8V unless otherwise noted. (All pin numbers refer to DIP package.)

SYMBOL	PARAMETER	MIN	TYP	MAX	UNITS	CONDITIONS
AMPLIFIER						
Z_{INAI}	Amp In Impedance	0.8	1.0	1.5	$\text{k}\Omega$	
$A_{V\text{MIN}}$	Min Voltage Gain			6.0	V/V	AC Output $4V_{P-P}$ Differential
$A_{V\text{MAX}}$	Max Voltage Gain	180			V/V	AC Output $4V_{P-P}$ Differential
V_{CAGC}	Voltage on C_{AGC}	2.8	4.5 3.4	5.5	V	$A_V = 6.0$ $A_V = 180$
GATE CHANNEL						
Z_{INGCI}	Gate Channel Input Impedance	1.75	2.5	3.25	$\text{k}\Omega$	
I_{CAGC^-}	Current that charges C_{AGC}	-1.5	-2.5	-3.5	mA	Pin 16 = 3.9V Pin 21 - Pin 22 = 1.3V
I_{CAGC^+}	Current that discharges C_{AGC}		1	5	μA	Pin 16 = 5.0V Pin 21 - Pin 22 = 0.7V
I_{VREF}	V_{REF} Input Bias Current		-0.01	-100	μA	
V_{THAGC}	AGC Threshold	0.88	1.0	1.12	V	Pin 16 = 4.2V See Note 1
I_{SH}	Set Hysteresis Bias Current		-60	-100	μA	
V_{THSH}	Set Hysteresis Threshold	0.48	0.6	0.72	V	See Note 2
TIME CHANNEL						
Z_{INTC}	Time Channel Input Impedance	3.5	5	6.5	$\text{k}\Omega$	
I_{CD}	Current into pins 1 & 24 that discharges C_D	2.1	2.7	3.4	mA	
WRITE MODE						
Z_{INAI}	Amplifier Input Impedance in Write Mode	100		500	Ω	Pin 11 = 2V
I_{CAGC}	Pin 16 Current in Write Mode		1.0	5.0	μA	Pin 11 = 2V Pin 16 = 3.9V Pin 21 - Pin 22 = 1.3V
DIGITAL PINS						
V_{IH}	High Level Input Voltage	2.0			V	ML8464B: Pins 11, 13 ML8464C: Pin 11
V_{IL}	Low Level Input Voltage			0.8	V	
V_I	Input Clamp Voltage			-1.5	V	$V_{CC} = 10.8\text{V}$, $I_I = -18\text{mA}$
I_{IH}	High Level Input Current			20	μA	$V_{CC} = 13.2\text{V}$, $V_I = 2.7\text{V}$
I_I	Input Current at Maximum Input Voltage			1	mA	$V_{CC} = 13.2\text{V}$, $V_I = 5.5\text{V}$
I_{IL}	Low Level Input Current			-200	μA	$V_{CC} = 13.2\text{V}$, $V_I = 0.5\text{V}$
V_{OH}	High Level Output Voltage	2.4			V	$V_{CC} = 10.8\text{V}$, $V_{IOH} = -40\mu\text{A}$ See notes 3, 7
V_{OL}	Low Level Output Voltage			0.5	V	$V_{CC} = 10.8\text{V}$, $I_{OL} = 800\mu\text{A}$, see note 7
I_{OSC}	Output Short Circuit Current			-100	mA	$V_{CC} = 13.2\text{V}$, $V_O = 0\text{V}$
I_{CC}	Supply Current		54	75	mA	$V_{CC} = 13.2\text{V}$

DC ELECTRICAL CHARACTERISTICS (Continued)

Over recommended operating conditions of $T_A = 0$ to 70°C , $V_{CC} = 12.0\text{V} \pm 10\%$, $V_{REF} = 0.5\text{V}$, Set Hysteresis = 0.3V , Read/Write = 0.8V unless otherwise noted. (All pin numbers refer to DIP package.)

SYMBOL	PARAMETER	MIN	TYP	MAX	UNITS	CONDITIONS
DIGITAL PINS (Continued)						
V_{OHCA}	Channel Alignment Pin V_{OH}	2.4	7.6		V	(Note 3) $I_{OH} = -40\mu\text{A}$ 10k Ω Load to GND
	ML8464B ML8464C				V	
V_{OLCA}	Channel Alignment Pin V_{OL}		6.9	0.4	V	(Note 3) $I_{OL} = 800\mu\text{A}$ 10k Ω Load to GND
	ML8464B ML8464C				V	
V_{OHTP}	Time Pulse Out Pin V_{OH}	2.4	9.6		V	10k Ω Load to GND 10k Ω Load to GND
	ML8464B ML8464C				V	
V_{OLTP}	Time Pulse Out Pin V_{OL}		8.6	0.4	V	10k Ω Load to GND 10k Ω Load to GND
	ML8464B ML8464C				V	

AC ELECTRICAL CHARACTERISTICS

Over recommended operating temperature and supply range of $V_{CC} = 10.8$ to 13.2V , $T_A = 0$ to 70°C .

SYMBOL	PARAMETER	MIN	TYP	MAX	UNITS	CONDITIONS
ML8464-1 t_{p-p}	Pulse Pairing		± 0.5	± 1.0	ns	f = 2.5MHz $V_{IN} = 40\text{mV}_{p-p}$ differential See note 4
ML8464-1.5 t_{p-p}	Pulse Pairing ⁶		± 0.8	± 1.5	ns	
ML8464-2 t_{p-p}	Pulse Pairing		± 1.5	± 3.0	ns	

Note 1: The AGC threshold is defined as the voltage across the gate channel input when the voltage on C_{AGC} is 4.2V.

Note 2: The Set Hysteresis threshold is defined as the voltage across the gate channel input when the channel alignment output voltage changes state.

Note 3: To prevent inductive coupling from the digital outputs to amplifier inputs, the TTL outputs should not drive more than one ALS TTL load.

Note 4: The filter and differentiator network are described in the pulse pairing set-up.

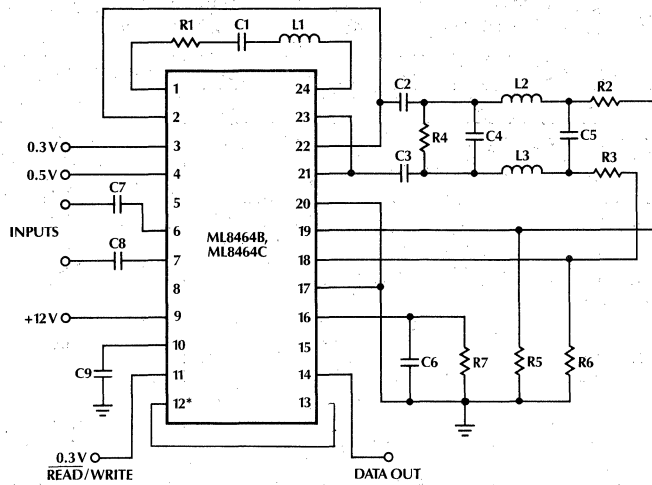
Note 5: All limits are guaranteed by 100% testing or alternate methods.

Note 6: The 1.5 ns pulse pairing specification is available only on the ML8464C, not the ML8464B.

Note 7: ML8464B: Pins 12, 14, 15.

ML8464C: Pins 14 and 15 only.

ML8464B, ML8464C



PULSE PAIRING SET UP

PARTS LIST

R1	220 Ω	C1	82pF
R4	680 Ω	C2, C3, C6	0.01 μ F
R2, R3	240 Ω	C4	100pF
R5, R6	3.3k Ω	C5	15pF
R7	100k Ω	C7, C8	0.0022 μ F
L1	1.5 μ H	C9	47pF
L2, L3	4.7 μ H		

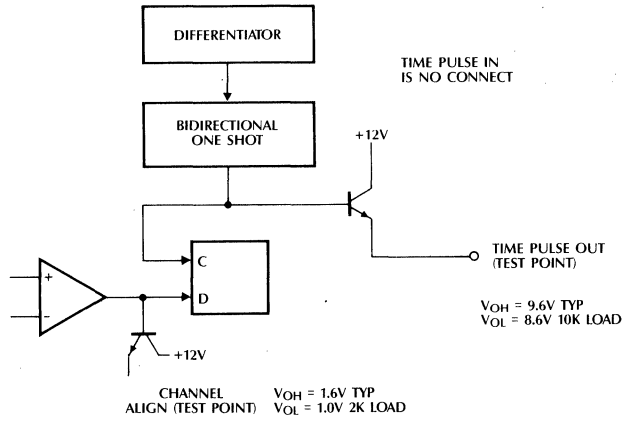
* The connection between pins 12 and 13 is required only for the ML8464B.

PULSE PAIRING MEASUREMENT

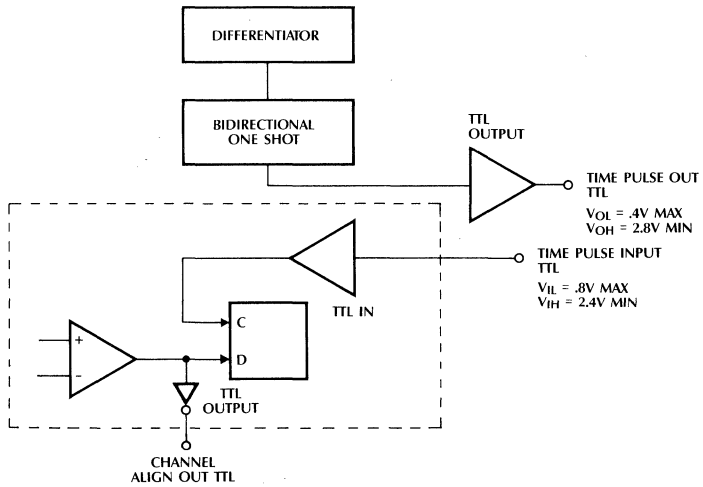
The scope probe is connected to pin 14 (Encoded Data Out) and triggered off of its positive edge. The trigger holdoff is adjusted so that the scope first triggers off the pulse associated with the positive peak and then off

the pulse associated with the negative peak. Pulse pairing is displayed on the second pair of pulses on the display. If the second pair of pulses are separated by 6ns, then the pulse pairing for the part is ± 3 ns.

ML8464C



ML8464B



ML8464B, ML8464C

DIFFERENCES BETWEEN ML8464C AND ML8464B

THE EXTERNAL DELAY

The ML8464B open circuits the digital signal at pins 12 and 13. This allows for the insertion of an external delay filter. The ML8464C has no TTL buffers at these pins and closes the signal path internally bringing out a test point at pin 12. Hence, the ML8464 does not allow for the external delay.

TEST POINTS

The ML8464B has two TTL test points at pins 12 and 15. The ML8464C uses open emitter followers in an ECL configuration. Hence, the voltage levels are not similar at pins 12 and 15 on both devices. The typical voltage level at pins 12 are $V_{OH} = 9.6V$, $V_{OL} = 8.6V$ and at pin 15 are $V_{OH} = 1.6V$, $V_{OL} = 1.0V$.

AGC GAIN CONTROL FACTOR

The AGC reference level is a DC voltage externally set at V_{REF} (pin 4). Increasing this DC voltage will increase the gain of the gain controlled amplifier.

AGC gain control factor =

$$V_{OUT\ PEAK} = \text{peak of the AGC amp} \frac{V_{REF}}{0.5V_{DC}}$$

$$\begin{aligned} \text{AGC gain control factor} &= \frac{2.5V_{PP}}{0.5V_{DC}} = 5 \text{ for ML8464B} \\ &= \frac{2.0V_{PP}}{0.5V_{DC}} = 4 \text{ for ML8464C} \end{aligned}$$

Thus, at $V_{REF} = 0.5V_{DC}$, $V_{OUT\ AGC} = 2.5V$ for ML8464B and $2.0V$ for ML8464C. This smaller signal amplitude should be taken into consideration at the hysteresis comparator. To set the desired amount of hysteresis, and external DC control voltage is used. The particular settings for V_{REF} and control voltage at pin 3 that optimizes the ML8464B performance may not necessarily optimize the ML8464C performance.

ORDERING INFORMATION

PART NUMBER	TEMP. RANGE	PACKAGE	PULSE PAIRING
ML8464C-1CP	0°C to +70°C	MOLDED DIP (P24)	±1 ns
ML8464C-1CQ	0°C to +70°C	MOLDED PCC (Q28)	±1 ns
ML8464C-1.5CP	0°C to +70°C	MOLDED DIP (P24)	±1.5 ns
ML8464C-1.5CQ	0°C to +70°C	MOLDED PCC (Q28)	±1.5 ns
ML8464C-2CP	0°C to +70°C	MOLDED DIP (P24)	±3 ns
ML8464C-2CQ	0°C to +70°C	MOLDED PCC (Q28)	±3 ns
ML8464B-1CP	0°C to +70°C	MOLDED DIP (P24)	±1 ns
ML8464B-1CQ	0°C to +70°C	MOLDED PCC (Q28)	±1 ns
ML8464B-2CP	0°C to +70°C	MOLDED DIP (P24)	±3 ns
ML8464B-2CQ	0°C to +70°C	MOLDED PCC (Q28)	±3 ns

Power Supply ICs

Section 6

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SWITCH MODE POWER SUPPLY CONTROL

Micro Linear offers high performance Switch Mode Power Supply IC controllers for PWM, Phase Modulation, Resonant and Power Factor Correction as well as Power Controllers for Battery powered systems. These IC's are the highest frequency IC's available and include unique features for enhanced stability, easy synchronization and improved fault management. These controllers can be tailored to meet your unique design requirements using Micro Linear's array-based Semi-Standard capability.

For Power Factor Correction, four IC's are available to meet the needs of a variety of different applications. Boost and Buck Boost dedicated PFC control IC's are available. Also, a new IC, the ML4819, combines a boost PFC stage with a Current Mode PWM control section. This new "Combo" controller is the first IC available which controls an entire PFC corrected power supply on a single chip.

Also, a new Average Current Sensing Boost PFC controller, the ML4821, achieves the lowest harmonic distortion and highest power factor.

Two new resonant controllers are now available, one for zero voltage switching and one multi-mode controller. Both IC's offer unique overload protection features, high current output drivers, and low cross conduction. The multi-mode controller supports both series resonant converters operating above resonance and ZCS topologies.

The ML4818 Phase Modulation Controller combines the low-loss zero-voltage switching transitions of a resonant topology with the efficient energy transfer characteristics of square wave PWM.

Our new ML4861 and ML4862 offer integrated system solutions for palmtop and notebook or laptop applications. High efficiency is achieved through synchronous rectification.

These IC's are available in Commercial (0°C to +70°C), Industrial (-40°C to +85°C) and Military (-55°C to +125°C) temperature ranges in both DIP and Surface Mount packages.

PULSE WIDTH MODULATION IC's

- 1MHz Operation
- Voltage Mode or Current Mode Operation
- High Current (2A peak) High Speed Totem Pole Outputs
- Precision (+1%) 5.1V Reference
- Soft Start Latch Ensures Full Soft Start Cycle
- Semi-Standard Options Available
- **Unique Features to Enhance Synchronization Stability Fault Protection**

POWER FACTOR CONTROLLERS

- ML4812 General Purpose Boost Mode (Peak I_{SENSE})
- ML4821 Average Current Sense Boost Mode
- ML4813 Flyback Converter for Low Power Systems
- ML4819 Boost PFC and PWM "Combo"

BATTERY POWERED IC's

- ML4862 Laptop Power Controller
- ML4861 Palmtop Boost Regulator

SOFT-SWITCHING AND RESONANT CONTROLLERS

ML4815 Single Ended Zero Voltage Switching Controller

- Ideal for Low Input Voltage DC to DC Converter Modules
- Operation to 1.5MHz

ML4816 Multi-Mode Push-Pull Resonant Controller

- Supports All Major Topologies: ZVS and ZCS
- Constant Off-Time or Constant On-Time Control

ML4818 Phase Modulation/Soft Switching Controller

- Full Bridge Zero Voltage Switching Operation
- Power DIP Package

ML4830 ELECTRONIC BALLAST CONTROLLER

- A complete solution for Electronic Ballasts
- Power Factor Controller
- PWM or Frequency Modulation

Power Supplies

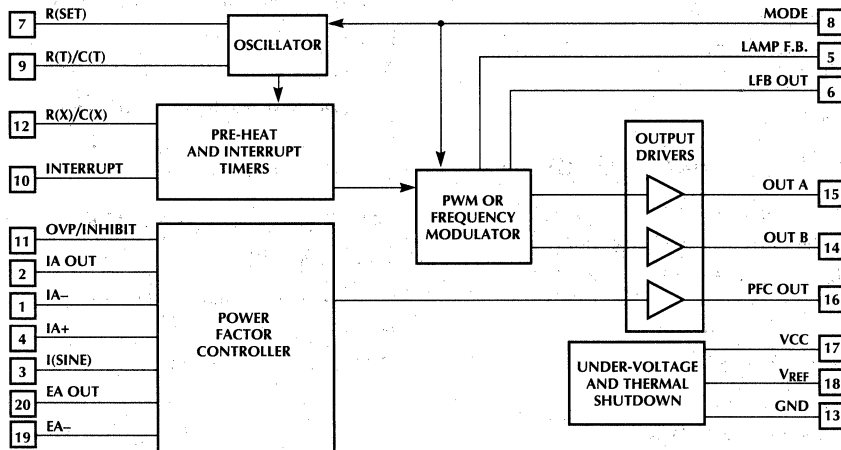
BATTERY POWERED CONTROL IC's

New in 1992 are two IC's for battery powered systems. The ML4862 is a complete control solution for multi-cell Laptop and Notebook computer systems. The ML4861 Palmtop Boost is the most compact boost converter on the market, providing 5V (or 3.3V) output for input voltages from 1V to 4.5V. Both use synchronous rectification for high efficiency.

	ML4862	ML4861
Application	Notebook/Laptop	Palmtop
Input Voltage Range	5.5 to 22V	1 to 4.5V
Output Voltages	3.3, 5V, 12V, 5V (Linear)	5V or 3.3V or Adjustable
Output Current	Determined by External Components	80mA (1 Cell) 350mA (3 Cell)
Other Features	Logic to MOSFET Drivers All N-Channel Design	Only 2 External Components

ELECTRONIC BALLAST CONTROL

The ML4830 is the first commercially available IC designed specifically for Electronic Ballasts. This IC contains an average current sensing boost mode power factor controller, similar to the ML4821, as well as a ballast control section. Flexibility is built in to the ML4830, allowing it to operate with programmable starting scenarios and either PWM or Frequency modulated ballast outputs.



POWER FACTOR CONTROLLERS

Micro Linear offers more choices for Power Factor Correction than any other company. All of Micro Linear's PFC systems will easily meet IEC-555 requirements for AC input current harmonic distortion.

For the lowest harmonic distortion available over a wide range of input line and output load conditions, Micro Linear has introduced the ML4821 Average Current Sensing PFC controller. The ML4813 flyback controller

offers the user a choice of output voltages and is easily isolated. The ML4819 PFC Combo IC provides a complete control solution for power factor corrected supplies, providing both the PFC pre-regulation and PWM control outputs. The ML4821 uses the lossless peak current sensing method for the highest efficiency possible.

All of Micro Linear's PFC control IC's provide Over-Voltage Protection and high current outputs for fast MOSFET drive.

Power Factor Controller Comparison

FEATURE	ML4812	ML4813	ML4819	ML4821
Topology	Continuous Boost	Flyback	Boost + PWM Combo	Continuous Boost
Efficiency	Highest	Moderate	High	High
System Power Factor	0.99	0.99	0.99	0.99
Usable Power Range	Over 50W	Under 250W	Under 400W	Over 50W
Output Voltage	$V_{OUT} > V_{IN}$	V_{OUT} Independent of V_{IN}	$V_{OUT} > V_{IN}$	$V_{OUT} > V_{IN}$
Over-Voltage Protection	YES	YES	YES	YES
Peak Output Drive	1A	1A	1A	1A
Control Method	Peak Current	Voltage Mode	Peak Current	Average Current
Sync Input				YES
Output can be Isolated	NO	YES	YES	NO

HIGH FREQUENCY PWM CONTROLLERS

Micro Linear offers a complete family of 1 MHz PWM control IC's with superior performance over the Industry Standard PWM controllers. Our Family of PWM controllers feature:

- 1MHz Operation
- 2A Peak Fast Output Drive
- High Speed Cycle-by-Cycle Current Limit
- Soft Start with Complete Reset
- Under-Voltage Lockout with Low Current Start-up
- Current Mode or Voltage Mode Operation

PWM Control IC Comparison

FEATURE	ML4809	ML4810	ML4811	ML4817	ML4823	ML4825
Push-Pull	YES	YES	YES			YES
Single-Ended				YES	YES	
Integrating Fault Detection		YES	YES	YES		
Oscillator Sync Input	YES		YES			
Slope Compensation	YES					
Precision Duty-Cycle Limit				YES		
Soft-Start Delay	YES	YES	YES			
Separate Error Amp Output Pin	YES					

Power Supplies

RESONANT AND PHASE MODULATION CONTROLLERS

Adding to the family of Switch Mode Power Supply Controllers are three IC's for high frequency Zero Voltage or Zero Current Switching. Micro Linear offers both Resonant and Phase Modulation Controllers.

We have introduced two new Resonant IC's. The ML4815 is optimized for single-ended DC to DC Zero Voltage Switching converters. The ML4816 multi-mode controller is designed for push-pull or half-bridge resonant

topologies and can run in either constant on-time mode (for Zero Current Switching) or constant off-time mode (for Zero Voltage Switching). The ML4816 includes programmable frequency limits for both upper and lower frequencies, ensuring that the IC can be used in supplies that operate either above or below resonance.

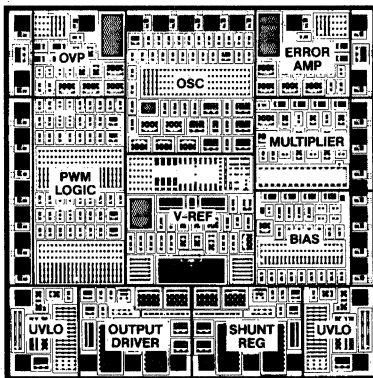
The ML4818 Phase Modulation Controller. Combines the Zero Voltage Switching characteristics of a resonant supply with the efficient energy transfer and simplicity of PWM.

Power Factor Controller Comparison

FEATURE	ML4815	ML4816	ML4818
Zero Voltage Switching	YES	YES	YES
Zero Current Switching		YES	
Control Type	Constant Off-Time	Constant On or Constant Off	Phase Modulation
Integrating Fault Detection	YES	YES	YES
Intended Application	DC-DC	Offline or DC-DC	Offline or DC-DC

SEMI-STANDARD, CUSTOMIZED CONTROLLERS

Micro Linear's unique Tile Array based designs make it possible to modify any of our standard products to suit your unique application needs. Semi-Standard is a low risk path to modify the IC to change its functionality, package, temperature range, or parametric performance. Call your Micro Linear representative for more information.



Power Supply Array Comparison

ARRAY	STANDARD PRODUCTS	PERFORMANCE
FB3480 High Frequency PWM	ML4823/25 ML4810/11 ML4809, ML4817	1MHz Operation 2 x 2A Output Drivers 50ns Prop. Delay
FB3490 General Purpose SMPS	ML4812, ML4813 ML4819 ML4821	500KHz Operation 2 x 1A Output Drivers 100ns Prop. Delay
FB3491 Resonant	ML4816	2MHz Operation 2 x 1.5A Output Drivers 50ns Prop. Delay
FB3492 Resonant	ML4818	2MHz Operation 4 x 1.5A Output Drivers 50ns Prop. Delay
FB3680	ML4830	12V Bipolar 3 x 200mA Drivers 50ns Prop. Delay
FC3580	ML4861	5V Operation 1A MOS Power MicroPower BiCMOS
FB3430	ML4862	36V Bipolar 4 Outputs

High Frequency Power Supply Controller

GENERAL DESCRIPTION

The ML1825 High Frequency PWM Controller is an IC controller optimized for use in Switch Mode Power Supply designs running at frequencies to 1MHz. Propagation delays are minimal through the comparators and logic for reliable high frequency operation while slew rate and bandwidth are maximized on the error amplifier. This controller is designed to work in either voltage or current mode and provides for input voltage feed forward.

A 1V threshold current limit comparator provides cycle-by-cycle current limit while exceeding a 1.4V threshold initiates a soft-start cycle. The soft start pin doubles as a maximum duty cycle clamp. All logic is fully latched to provide jitter-free operation and prohibit multiple pulsing. An under-voltage lockout circuit with 800mV of hysteresis assures low startup current and drives the outputs low.

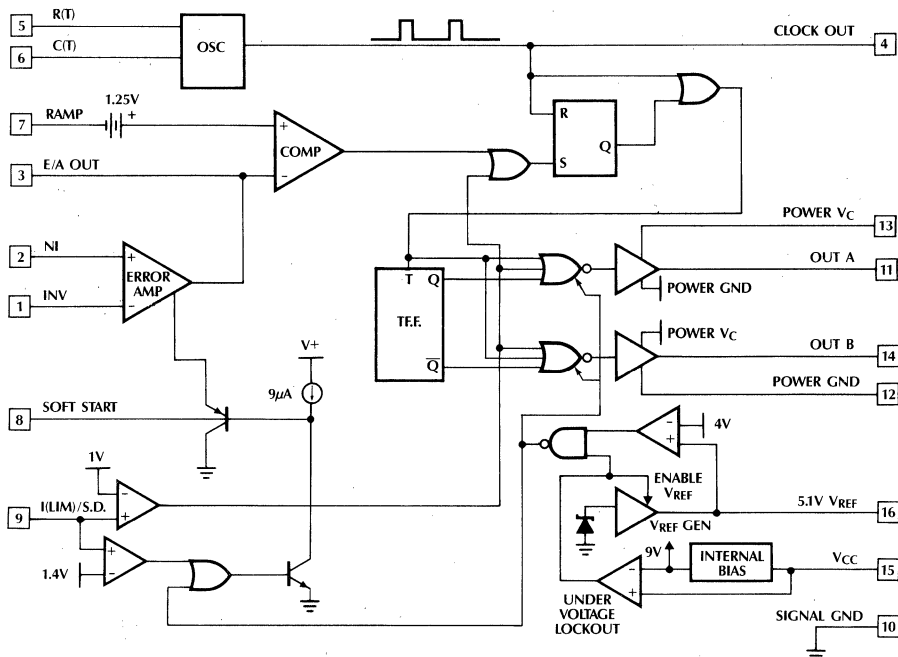
The ML1825 is fabricated on a 40V bipolar process from the FB3480 Power Supply Controller Array. Customized versions of this controller are therefore easily implemented. Please refer to the FB3480 datasheet for more information.

This controller is a pin for pin replacement for the UC1825 controller.

FEATURES

- Practical Operation at Switching Frequencies to 1.0MHz
- High Current (2A peak) Dual Totem Pole Outputs
- Wide Bandwidth Error Amplifier
- Fully Latched Logic with Double Pulse Suppression
- Pulse-by-Pulse Current Limiting
- Soft Start and Max. Duty Cycle Control
- Under Voltage Lockout with Hysteresis
- 5.1V, $\pm 1\%$ Trimmed Bandgap Reference
- Pin Compatible Replacement for UC1825

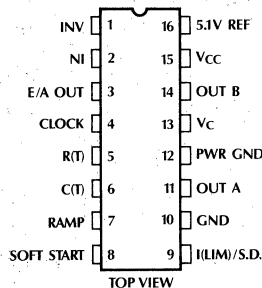
BLOCK DIAGRAM



ML1825

PIN CONNECTION

ML1825
16-Pin DIP



PIN DESCRIPTION

PIN #	NAME	FUNCTION	PIN #	NAME	FUNCTION
1	INV	Inverting input to error amp.	9	I(LIM)/S.D.	Current limit sense pin. Normally connected to current sense resistor.
2	NI	Non-inverting input to error amp.	10	GND	Analog Signal Ground.
3	E/A OUT	Output of error amplifier and input to main comparator.	11	OUT A	High Current Totem pole output. This output is the first one energized after Power On Reset.
4	CLOCK	Oscillator output.	12	PWR GND	Return for the High Current Totem pole outputs.
5	R(T)	Timing Resistor for Oscillator — sets charging current for oscillator timing capacitor (pin 6).	13	V _c	Positive Supply for the High Current Totem pole outputs.
6	C(T)	Timing Capacitor for Oscillator.	14	OUT B	High Current Totem pole output.
7	RAMP	Non-Inverting input to main comparator. Connected to C(T) for Voltage Mode operation or to current sense resistor for current mode.	15	V _{CC}	Positive Supply for the IC.
8	SOFT START	Normally connected to Soft Start Capacitor.	16	5.1V REF	Buffered output for the 5.1V voltage reference.

ABSOLUTE MAXIMUM RATINGS

Supply Voltage (Pins 15, 13)	30V
Output Current, Source or Sink (Pins 11, 14)	
DC	0.5A
Pulse (0.5 μ s)	2.0A
Analog Inputs	
(Pins 1, 2, 7)	-0.3V to 7V
(Pins 9, 8)	-0.3V to 6V
Clock Output Current (Pin 4)	-5mA
Error Amplifier Output Current (Pin 3)	5mA
Soft Start Sink Current (Pin 8)	20mA
Oscillator Charging Current (Pin 5)	-5mA
Junction Temperature	
ML4825M	150°C
ML4825I, ML4825C	125°C
Storage Temperature	-65°C to +150°C
Lead Temperature (Soldering 10 sec.)	+260°C
Thermal Resistance (θ_{JA})	
Ceramic DIP	65°C/W

OPERATING CONDITIONS

Temperature Range	
ML1825M	-55°C to +125°C

Absolute maximum ratings are those values beyond which the device could be permanently damaged. Absolute maximum ratings are stress ratings only and functional device operation is not implied.

ELECTRICAL CHARACTERISTICS

Unless otherwise specified, $R_T = 3.65K\Omega$, $C_T = 1000pF$, $T_A =$ Operating Temperature Range, $V_{CC} = 15V$.

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
Oscillator					
Initial Accuracy	$T_J = 25^\circ C$, (note 1)	360	400	440	KHz
Voltage Stability	$10V < V_{CC} < 30V$, (note 1)		0.2	2	%
Temperature Stability	(note 1)		5		%
Total Variation	line, temp, (note 1)	340		460	KHz
Clock Out High		3.9	4.5		V
Clock Out Low			2.3	2.9	V
Ramp Peak	(note 1)	2.6	2.8	3.0	V
Ramp Valley	(note 1)	0.7	1.0	1.25	V
Ramp Valley to Peak	(note 1)	1.6	1.8	2.0	V
Reference Section					
Output Voltage	$T_J = 25^\circ C$, $I_O = 1mA$	5.05	5.10	5.15	V
Line Regulation	$10V < V_{CC} < 30V$		2	20	mV
Load Regulation	$1mA < I_O < 10mA$		5	20	mV
Temperature Stability	$-55^\circ C < T_J < 150^\circ C$, (note 1)		.2	.4	%
Total Variation	line, load, temp (note 1)	5.0		5.20	V
Output Noise Voltage	10Hz to 10KHz		50		μV
Long Term Stability	$T_J = 125^\circ C$, 1000 hrs, (note 1)		5	25	mV
Short Circuit Current	$V_{REF} = 0V$	-15	-50	-100	mA
Error Amplifier Section					
Input Offset Voltage				10	mV
Input Bias Current			.6	3	μA
Input Offset Current			.1	1	μA
Open Loop Gain	$1 < V_O < 4V$	60	95		dB

ML1825

ELECTRICAL CHARACTERISTICS (Continued)

Unless otherwise specified, $R_T = 3.65K\Omega$, $C_T = 1000pF$, $T_A =$ Operating Temperature Range, $V_{CC} = 15V$.

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
Error Amplifier Section (Continued)					
CMRR	$1.5 < V_{CM} < 5.5V$	75	95		dB
PSRR	$10 < V_{CC} < 30V$	85	110		dB
Output Sink Current	$V_{PIN\ 3} = 1V$	1	2.5		mA
Output Source Current	$V_{PIN\ 3} = 4V$	-5	-1.3		mA
Output High Voltage	$I_{PIN\ 3} = -0.5mA$	4.0	4.7	5.0	V
Output Low Voltage	$I_{PIN\ 3} = 1mA$	0	0.5	1.0	V
Unity Gain Bandwidth	(note 1)	3	5.5		MHz
Slew Rate	(note 1)	6	12		V/ μs
PWM Comparator Section					
Pin 7 Bias Current	$V_{PIN\ 7} = 0V$		-1	-5	μA
Duty Cycle Range		0		80	%
Pin 3 Zero DC Threshold	$V_{PIN\ 7} = 0V$	1.1	1.25		V
Delay to Output	(note 1)		50	80	ns
Soft-Start Section					
Charge Current	$V_{PIN\ 8} = 0.5V$	3	9	20	μA
Discharge Current	$V_{PIN\ 8} = 1V$	1			mA
Current Limit/Shutdown Section					
Pin 9 Bias Current	$0V < V_{PIN\ 9} < 4V$			+15	μA
Current Limit Threshold		.9	1	1.1	V
Shutdown Threshold		1.25	1.4	1.55	V
Delay to Output	(note 1)		50	80	ns
Output Section					
Output Low Level	$I_{OUT} = 20mA$.25	.4	V
	$I_{OUT} = 200mA$		1.2	2.2	V
Output High Level	$I_{OUT} = -20mA$	13.0	13.5		V
	$I_{OUT} = -200mA$	12.0	13.0		V
Collector Leakage	$V_C = 30V$		100	500	μA
Rise/Fall Time	$C_L = 1000pF$, (note 1)		30	60	ns
Under-Voltage Lockout Section					
Start Threshold		8.8	9.2	9.6	V
UVLO Hysteresis		.4	.8	1.2	V
Supply Current					
Start Up Current	$V_{CC} = 8V$		1.1	2.5	mA
I_{CC}	$V_{PIN\ 1, 7, 9} = 0V$, $V_{PIN\ 2} = 1V$,		22	33	mA

Note 1: This parameter not 100% tested in production but guaranteed by design.

FUNCTIONAL DESCRIPTION

OSCILLATOR

The ML1825 oscillator charges the external capacitor (C_T) with a current (I_{SET}) equal to $3/R_{SET}$. When the capacitor voltage reaches the upper threshold (Ramp Peak), the comparator changes state and the capacitor discharges to the lower threshold (Ramp Valley) through Q1. While the capacitor is discharging, Q2 provides a high pulse.

The Oscillator period can be described by the following relationship:

$$T_{OSC} = T_{RAMP} + T_{DEADTIME}$$

where: $T_{RAMP} = C (Ramp Valley to Peak) / I_{SET}$

and: $T_{DEADTIME} = C (Ramp Valley to Peak) / I_{Q1}$

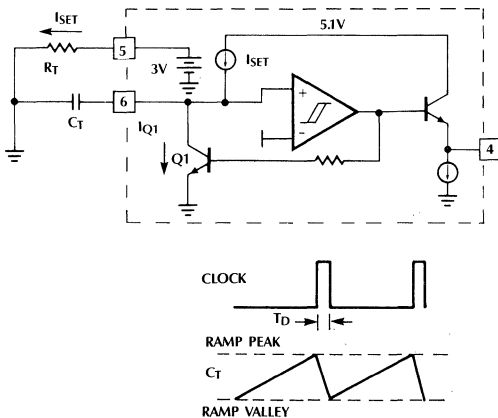


Figure 1. Oscillator Block Diagram

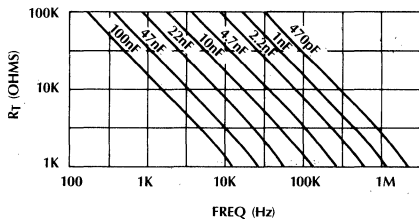


Figure 2. Oscillator Timing Resistance vs Frequency

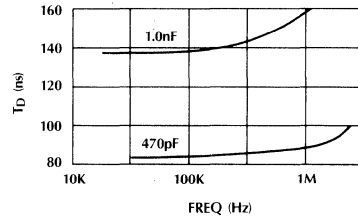


Figure 3. Oscillator Deadtime vs Frequency

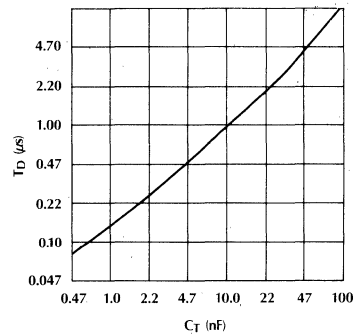


Figure 4. Oscillator Deadtime vs $C(T)$ ($3K\Omega \leq R(T) \leq 100K\Omega$)

ERROR AMPLIFIER

The ML1825 error amplifier is a 5.5MHz bandwidth $12V/\mu s$ slew rate op-amp with provision for limiting the positive output voltage swing (Output Inhibit line) for ease in implementing the soft start function.

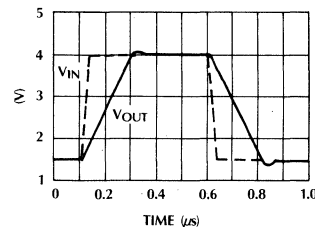


Figure 5. Unity Gain Slew Rate

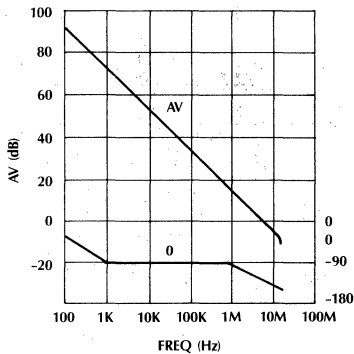


Figure 6. Open Loop Frequency Response

OUTPUT DRIVER STAGE

The ML1825 Output Driver is a 2A peak output high speed totem pole circuit designed to quickly switch the gates of capacitive loads, such as power MOSFET transistors.

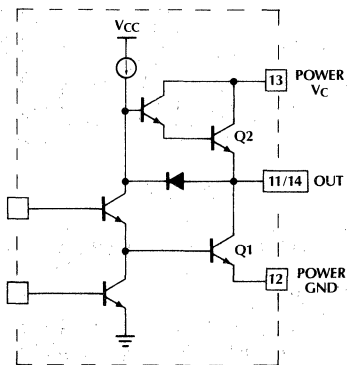


Figure 7. Simplified Schematic

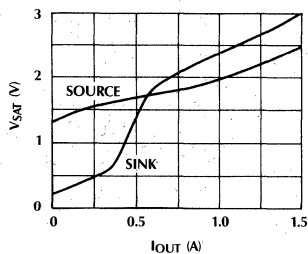


Figure 8. Saturation Curves

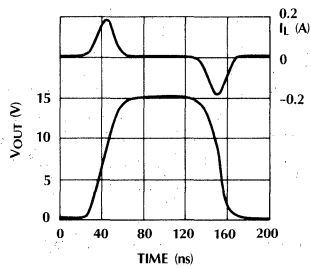


Figure 9. Rise/Fall Time ($C_L = 1000pF$)

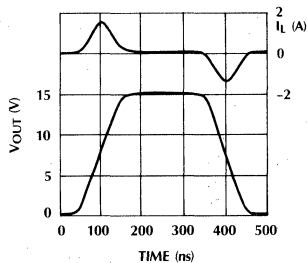


Figure 10. Rise/Fall Time ($C_L = 10,000pF$)

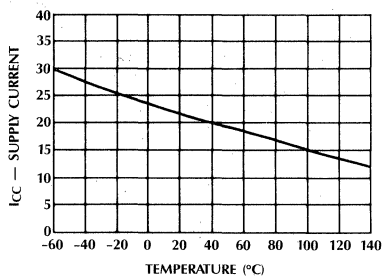


Figure 11. Supply Current vs Temperature

SOFT START AND CURRENT LIMIT

The ML1825 employs two current limits. When the voltage at pin 9 exceeds 1V, the outputs are immediately shut off and the cycle is terminated for the remainder of the oscillator period by resetting the RS flip flop.

If the output current is rising quickly such that the voltage on pin 9 reaches 1.4V before the outputs have turned off, a soft start cycle is initiated and the soft start capacitor (pin 8) is discharged. The duty cycle on start up is limited by limiting the output voltage of the error amplifier voltage to the voltage at pin 8.

ORDERING INFORMATION

PART NUMBER	TEMPERATURE RANGE	PACKAGE
ML1825MJ	-55°C to +125°C	Hermetic DIP (J16)

6

High Frequency PWM Controller

GENERAL DESCRIPTION

The ML4809 High Frequency PWM Controller is a full-featured IC controller optimized for use in Switch Mode Power Supply designs running at frequencies to 1MHz. Propagation delays are minimized while slew rate and bandwidth are maximized for reliable high frequency operation. This controller is designed to work in either voltage or current mode and provides for input voltage feed forward.

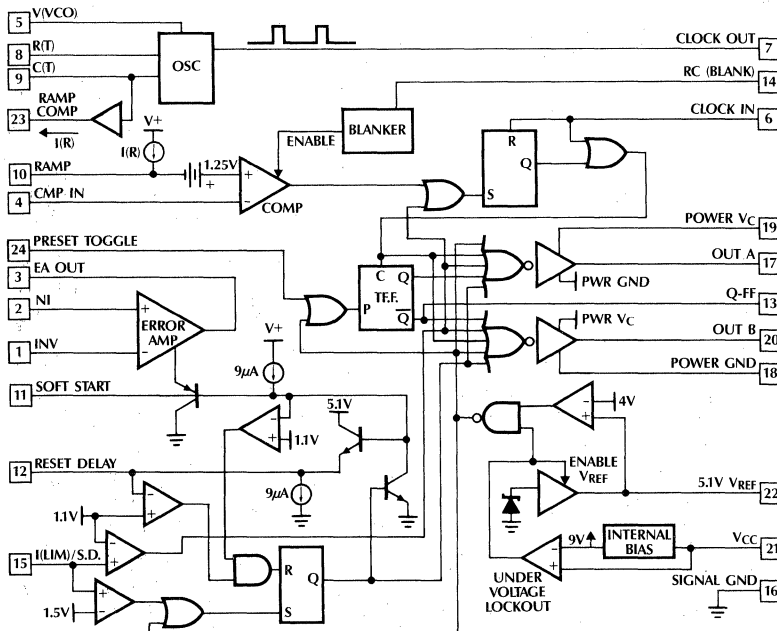
A 1.1V threshold current limit comparator provides cycle-by-cycle current limit while exceeding a 1.5V threshold initiates a soft-start cycle. The soft start pin doubles as a maximum duty cycle clamp. All logic is fully latched to provide jitter-free operation and prohibit multiple pulsing. An under-voltage lockout circuit with 7V of hysteresis assures low startup current and drives the outputs low.

The ML4809 is fabricated on a 40V bipolar process from the FB3480 Power Supply Controller Array. Customized versions of this controller are easily implemented. This controller is similar to the UC1825 controller, however the ML4809 includes many features not found on the 1825. These features are set in *Italics*.

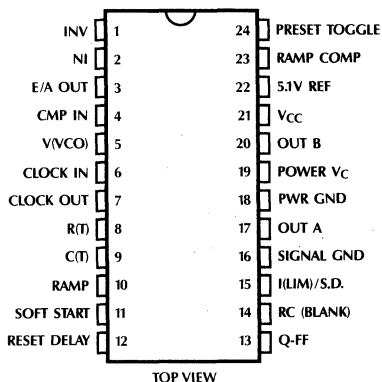
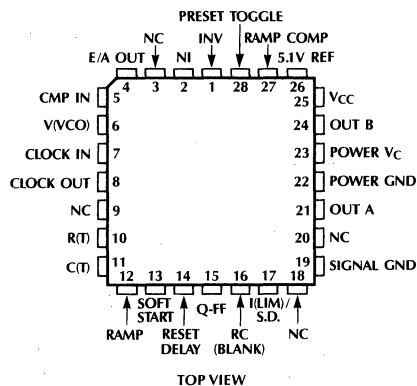
FEATURES

- Practical Operation at Switching Frequencies to 1.0MHz
- High Current (2A peak) Dual Totem Pole Outputs
- Wide Bandwidth Error Amplifier
- Fully Latched Logic with Double Pulse Suppression
- Pulse-by-Pulse Current Limiting
- Soft Start and Max. Duty Cycle Control
- 5.1V, $\pm 1\%$ Trimmed Bandgap Reference
- *Under Voltage Lockout: 16V Start with 7V Hysteresis*
- *Programmable Ramp Compensation Circuit*
- *VCO Input for Synchronization or Frequency Control*
- *External Clock Input for Synchronization*
- *Toggle Preset for Synchronization*
- *Comparator Blanker for Better Noise Immunity/Stability*
- *Separate Error Amplifier Output Pin for Loop Filtering Versatility*
- *Fast Shut Down Path from Current Limit to Outputs*
- *Outputs Preset to Known Condition After Under Voltage Lockout*
- *Soft Start Latch Ensures Full Soft Start Cycle*
- *Programmable Soft Start Delay*

BLOCK DIAGRAM



PIN CONNECTIONS

ML4809
24-Pin DIPML4809
28-Pin PCC

PIN DESCRIPTION

PIN #	NAME	FUNCTION	PIN #	NAME	FUNCTION
1	INV	Inverting input to error amp.	13	Q-FF	An Emitter Follower output which is High for B active.
2	NI	Non-inverting input to error amp.	14	RC (BLANK)	Connect resistor and capacitor to ground for blanker function.
3	E/A OUT	Output of error amplifier.	15	I(LIM)/S.D.	Current limit sense pin. Normally connected to sense resistor.
4	CMP IN	Main Comparator Input.	16	GND	Analog Signal Ground.
5	V(VCO)	A control voltage input which sets the VCO frequency. May be tied to 5.1V REF (22) for fixed frequency operation.	17	OUT A	High Current Totem pole output. This output is the first one energized after Power On Reset.
6	CLOCK IN	A "1" level blanks the outputs and prepares the chip for the next cycle by toggling the T flip flop.	18	PWR GND	Return for the High Current Totem pole outputs.
7	CLOCK OUT	Oscillator output. This is an emitter follower output.	19	POWER V _c	Positive Supply for the High Current Totem pole outputs.
8	R(T)	Timing Resistor for Oscillator — sets charging current for oscillator timing capacitor (pin 9).	20	OUT B	High Current Totem pole output.
9	C(T)	Timing Capacitor for Oscillator.	21	V _{CC}	Positive Supply for the IC.
10	RAMP	Non-Inverting input to main comparator. Connected to C(T) for Voltage Mode operation or to current sense resistor for current mode.	22	5.1V REF	Buffered output for the 5.1V voltage reference.
11	SOFT START	Normally connected to Soft Start Capacitor.	23	RAMP COMP	Connect resistor to GND for ramp compensation.
12	RESET DELAY	Connect to capacitor for time delay before new soft-start cycle begins after 1.4V current limit is reached.	24	PRESET TOGGLE	Presets the toggle flip-flop. Tie to GND to disable.

ML4809

ABSOLUTE MAXIMUM RATINGS

Supply Voltage (Pins 21, 19)	36V
Output Current, Source or Sink (Pins 17, 20)	
DC	0.5A
Pulse (0.5 μ s)	2.0A
Input Voltage	
(Pins 1, 2, 4, 5, 10)	-0.3V to 7V
(Pins 8, 9, 11, 12, 15, 24)	-0.3V to 6V
Logic Output Current (Pins 7, 13)	-5mA
Blanker Charge Current (Pin 14)	-5mA
Error Amplifier Output Current (Pin 3)	5mA
Soft Start Sink Current (Pin 11)	20mA
Oscillator Charging Current (Pin 8)	-5mA
Junction Temperature	
ML4809M	150°C
ML4809I, ML4809C	125°C
Storage Temperature Range	-65°C to +150°C
Lead Temperature (Soldering 10 sec.)	+260°C

Thermal Resistance (θ_{JA})	
Plastic DIP	50°C/W
Ceramic DIP	55°C/W
Plastic Chip Carrier (PCC)	55°C/W

OPERATING CONDITIONS

Temperature Range	
ML4809M	-55°C to +125°C
ML4809I	-40°C to +85°C
ML4809C	0°C to +70°C

Absolute maximum ratings are those values beyond which the device could be permanently damaged. Absolute maximum ratings are stress ratings only and functional device operation is not implied.

ELECTRICAL CHARACTERISTICS

Unless otherwise specified, $R_T = 6.2K\Omega$, $C_T = 1000pF$, $V(VCO) = V_{REF}$, R_L (Pins 7, 13) = $5K\Omega$, T_A = Operating Temperature Range, $V_{CC} = 15V$. (note 3)

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS	
Oscillator						
Initial Accuracy	$T_J = 25^\circ C$, (note 1)	390	430	470	KHz	
Voltage Stability	$10V < V_{CC} < 30V$, (note 1)		0.2	4	%	
Temperature Stability	(note 1)			5	%	
Total Variation	line, temp, (note 1)	370	430	490	KHz	
Clock Out High		3.9	4.5		V	
Clock Out Low			2.3	2.9	V	
Ramp Peak	(note 1)	2.6	2.8	3.0	V	
Ramp Valley	(note 1)	0.7	1.0	1.25	V	
Ramp Valley to Peak	(note 1)	1.6	1.8	2.0	V	
V(VCO) Control Range		1		5.5	V	
Reference Section						
Output Voltage	ML4809C	$T_J = 25^\circ C$, $I_O = 1mA$	5.00	5.10	5.20	V
	ML4809M, ML4809I		5.05	5.10	5.15	V
Line Regulation	$10V < V_{CC} < 30V$		2	20	mV	
Load Regulation	$1mA < I_O < 10mA$		5	20	mV	
Temperature Stability	$-55^\circ C < T_J < 150^\circ C$, (note 1)		.2	.4	%	
Total Variation	ML4809C	line, load, temp, (note 1)	4.95		5.25	V
	ML4809M, ML4809I		5.0		5.20	V
Output Noise Voltage	10Hz to 10KHz		50		μV	
Long Term Stability	$T_J = 125^\circ C$, 1000 hrs, (note 1)		5	25	mV	
Short Circuit Current	$V_{REF} = 0V$, $T_A \geq 0^\circ C$	-15	-50	-100	mA	
	$T_A < 0^\circ C$	-15	-50	-120	mA	
Under-Voltage Lockout Section						
Start Threshold		15	16	17	V	
UVLO Hysteresis		6.5	7	7.5	V	

ELECTRICAL CHARACTERISTICS (Continued)

Unless otherwise specified, $R_T = 6.2\text{k}\Omega$, $C_T = 1000\text{pF}$, $V(\text{VCO}) = V_{\text{REF}}$, R_L (Pins 7, 13) = $5\text{k}\Omega$,
 T_A = Operating Temperature Range, $V_{\text{CC}} = 15\text{V}$. (note 3)

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
Error Amplifier Section					
Input Offset Voltage	$T_A = 25^\circ\text{C}$			± 15	mV
				± 20	mV
Input Bias Current			.6	3	μA
Input Offset Current			.1	1	μA
Open Loop Gain	$1 < V_O < 4\text{V}$	60	96		dB
CMRR	$1.5 < V_{\text{CM}} < 5.5\text{V}$, $T_A \geq 0^\circ\text{C}$	75	95		dB
	$1.5 < V_{\text{CM}} < 5.5\text{V}$, $T_A < 0^\circ\text{C}$	65	95		dB
PSRR	$10 < V_{\text{CC}} < 30\text{V}$, $T_A \geq 0^\circ\text{C}$	80	110		dB
	$10 < V_{\text{CC}} < 30\text{V}$, $T_A < 0^\circ\text{C}$	70	110		dB
Output Sink Current	$V_{\text{PIN } 3} = 1\text{V}$	1	2.5		mA
Output Source Current	$V_{\text{PIN } 3} = 4\text{V}$	-5	-1.3		mA
Output High Voltage	$I_{\text{PIN } 3} = 1\text{mA}$	3.5	4.3	5.0	V
Output Low Voltage	$I_{\text{PIN } 3} = 1\text{mA}$	0	0.5	1.0	V
Unity Gain Bandwidth	(note 1)	3	5.5		MHz
Slew Rate	(note 1)	6	12		V/ μs
PWM Comparator Section					
Pin 10 Bias Current	$V_{\text{PIN } 10} = 0\text{V}$, $V_{\text{PIN } 23} = \text{open}$, $V_{\text{PIN } 9} = 2\text{V}$		-1	-10	μA
Duty Cycle Range	$T_A \geq 0^\circ\text{C}$	0		75	%
	$T_A < 0^\circ\text{C}$	0		70	%
Pin 4 Zero DC Threshold	$V_{\text{PIN } 7} = 0\text{V}$	1.1	1.25		V
Delay to Output	(note 1)		50	80	ns
Ramp Compensation					
Pin 10 Current	$V_{\text{PIN } 9} = 2\text{V}$, $R_{\text{PIN } 23} = 6.8\text{k}\Omega$	270	300	330	μA
Soft-Start/Reset Delay Section					
Charge Current (Pin 11)	$V_{\text{PIN } 11} = 0.5\text{V}$	-3	-9	-20	μA
Discharge Current (Pin 11)	$V_{\text{PIN } 11} = 1\text{V}$	1			mA
Discharge Current (Pin 12)	$V_{\text{PIN } 12} = 1\text{V}$	3	9	20	μA
Charge Current (Pin 12)	$V_{\text{PIN } 12} = 0.5\text{V}$	-4			mA
Current Limit/Shutdown Section					
Pin 15 Bias Current	$0\text{V} < V_{\text{PIN } 15} < 4\text{V}$			± 20	μA
Current Limit Threshold		1.0	1.1	1.2	V
Shutdown Threshold		1.35	1.50	1.65	V
Delay to Output	(note 1)		40	70	ns
Blanker Section					
T_{BLANK}	(note 1), $R_C = 5.1\text{k}\Omega$, 68pF	80	100	120	ns

ML4809

ELECTRICAL CHARACTERISTICS (Continued)

Unless otherwise specified, $R_T = 6.2K\Omega$, $C_T = 1000\mu F$, $V(VCO) = V_{REF}$, R_L (Pins 7, 13) = $5K\Omega$, T_A = Operating Temperature Range, $V_{CC} = 15V$. (note 3)

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
Output Section					
Output Low Level	$I_{OUT} = 20mA$.25	.4	V
	$I_{OUT} = 200mA$		1.2	2.2	V
Output High Level	$I_{OUT} = -20mA$	13.0	13.5		V
	$I_{OUT} = -200mA$	12.0	13.0		V
Collector Leakage	$V_C = 30V$		100	500	μA
Rise/Fall Time	$C_L = 1000\mu F$, (note 1)		30	60	ns
Logic Inputs/Outputs					
Pin 24 Threshold	(note 2)		$V_{REF} - .98$		V
Pin 13 V_{OH}	(note 2)		$V_{REF} - .65$		V
Pin 13 V_{OL}	(note 2)		$V_{REF} - 1.3$		V
Supply Current					
Start Up Current (note 2)	$V_{CC} = 8V$, $T_A = 25^\circ C$		1.1	3.0	mA
	$V_{CC} = 8V$, $T_A < 25^\circ C$			3.5	mA
I_{CC}	$V_{PIN 1, 10, 15} = 0V$, $V_{PIN 2} = 1V$, $T_A = 25^\circ C$		29	38	mA

Note 1: This parameter not 100% tested in production but guaranteed by design.

Note 2: The thresholds on the logic input pins are set by a reference generator that is: $V_{TH} = V_{REF} - (1.5 * V_{BE})$. The logic outputs swing from: $V_{OH} = V_{REF} - V_{BE}$ to $V_{OL} = V_{REF} - 2 * V_{BE}$. V_{BE} is nominally .65V and varies with temperature. Logic inputs and outputs will track each other with temperature variation.

Note 3: Since the Under Voltage Lockout start-up threshold is 16V, the supply is first raised to 20V to activate the IC and then lowered to 15V to conduct the electrical testing.

Note 4: Reference short circuit current, Supply current and Start-up I_{CC} decrease with increasing temperature.

FUNCTIONAL DESCRIPTION

OSCILLATOR

The ML4809 Voltage Controlled Oscillator charges the external capacitor (C_T) with a current (I_{CHARGE}) equal to $V(VCO)/R_T$. When the capacitor voltage reaches the upper threshold (Ramp Peak), the comparator changes state and the capacitor discharges to the lower threshold (Ramp Valley) through Q1. While the capacitor is discharging, Q2 provides a high pulse. For Fixed Frequency Operation, $V(VCO)$ can be tied to V_{REF} .

The Oscillator period can be described by the following relationship:

$$T_{OSC} = T_{RAMP} + T_{DEADTIME}$$

where: $T_{RAMP} = C$ (Ramp Valley to Peak)/ I_{CHARGE}

and: $T_{DEADTIME} = C$ (Ramp Valley to Peak)/ I_{DIS}

An approximate expression for the oscillator frequency in fixed frequency operation (where $V(VCO) = V_{REF}$) is:

$$F_{OSC} \approx \frac{2.48}{R_T C_T}$$

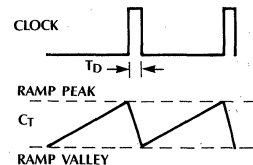
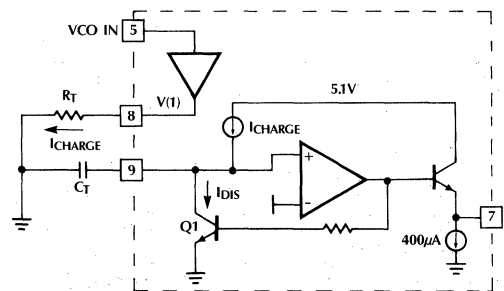


Figure 1. Oscillator Block Diagram

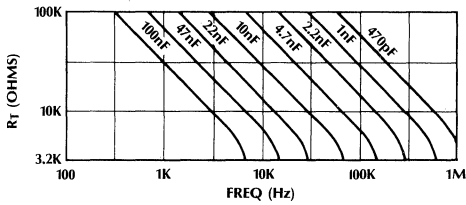


Figure 2. Timing Resistance vs Frequency (V(VCO) = 5.1V)

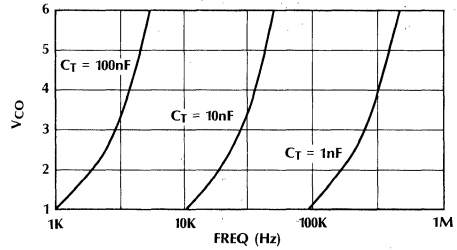


Figure 3. Oscillator Frequency vs V(VCO)
(RC = 6.2KΩ, 1000pF)

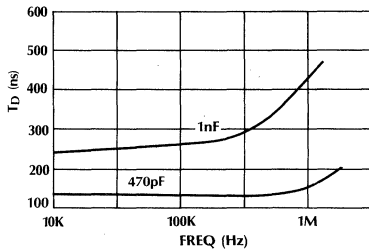


Figure 4. Oscillator Deadtime vs Frequency (V(VCO) = 5.1V)

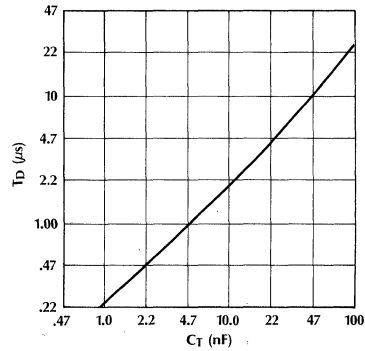


Figure 5. Oscillator Deadtime vs C(T) (3KΩ ≤ R(T) ≤ 100KΩ)

ERROR AMPLIFIER

The ML4809 error amplifier is a 3.5MHz bandwidth 6V/μs slew rate op-amp with provision for limiting the

positive output voltage swing (Output Inhibit line) for ease in implementing the soft start function.

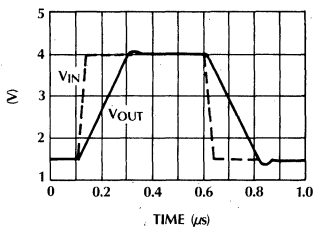


Figure 6. Unity Gain Slew Rate

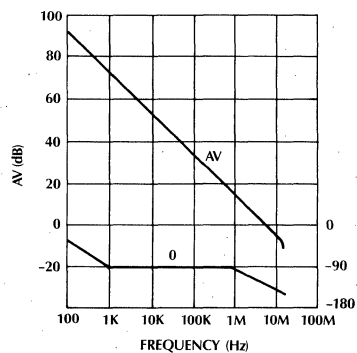


Figure 7. Open Loop Frequency Response

OUTPUT DRIVER STAGE

The ML4809 Output Driver is a 2A peak output high speed totem pole circuit designed to drive capacitive loads, such as power MOSFET transistors.

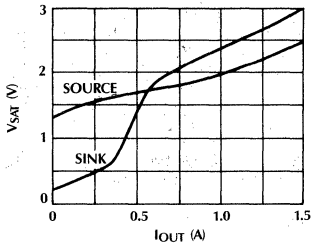


Figure 8. Saturation Curves

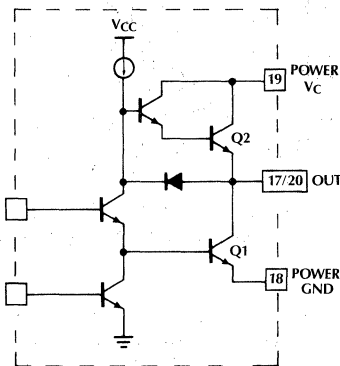


Figure 9. Simplified Schematic

SOFT START, CURRENT LIMIT, AND RESET DELAY

The ML4809 employs two current limits. When the voltage at pin 15 (I(LIM)/S.D.) exceeds 1.1V, the outputs immediately pull low and the cycle is terminated for the remainder of the oscillator period by resetting the RS flip flop.

If the output current is rising quickly (usually due to transformer saturation) such that the voltage on pin 15 reaches 1.5V before the outputs have turned off, a soft start cycle is initiated. The soft start capacitor (pin 11) is discharged and outputs are held "off" until the voltage at pin 11 reaches 1.1V, ensuring a complete soft start cycle. The duty cycle on start up is limited by limiting the output voltage of the error amplifier voltage to the voltage at pin 11.

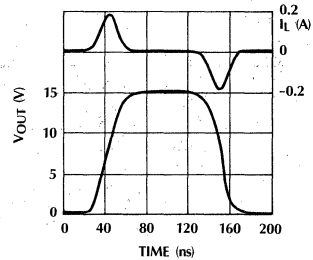


Figure 10a. Rise/Fall Time ($C_L = 1000pF$)

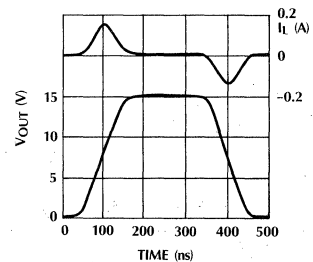


Figure 10b. Rise/Fall Time ($C_L = 10000pF$)

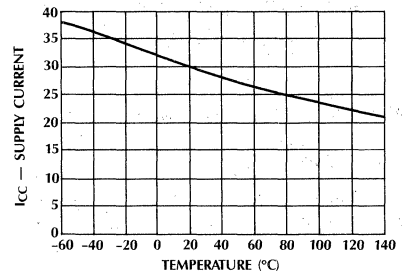


Figure 11. Supply Current vs. Temperature

The ML4809 also includes a delay circuit which inhibits the outputs from coming on until a time determined by the RESET DELAY capacitor on pin 12. This capacitor is normally charged to a voltage equal to $V_{PIN 11} - .7V$ and is limited to V_{REF} . After the 1.5V limit is reached, the capacitor is allowed to slowly discharge through the $9\mu A$ current sink. When this capacitor and the Soft Start Capacitor both have discharged to 1.1V, the outputs are enabled and the new soft start cycle begins. During Under Voltage Lockout, both capacitors will be discharged to prepare for a new cycle.

Since the emitter follower which drives pin 12 presents a load on Pin 11, the Soft Start Capacitor's effective value will be increased by:

$$C_{EFFECT} = C_{PIN 11} + (C_{PIN 12}/\beta)$$

where β varies from 50 to 250. Should this cause unacceptable variation on the soft start capacitor value, this effect can be mitigated by connecting a resistor from V_{REF} to pin 11 to charge the Soft Start Cap (select a resistor which keeps the charge current below 2mA).

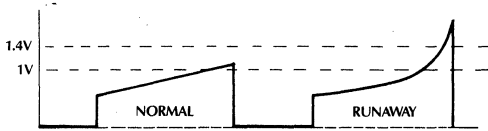


Figure 12. Normal (Cycle by Cycle) and "Runaway" Current Timing

RAMP COMPENSATION

In order to allow stable operation of a current mode regulator above 50% duty cycle, some of the oscillator ramp needs to be added to the current signal.

Notice that the waveform of (1) and the waveform of ramp (2) have different average current values. (1) is an example of a waveform for high line and (2) an example of low line. Since the controllers all regulate based on the peak value of the current in the circuit, and the control variable really wants to be the average current, adding some of the oscillator ramp to comparator input (shown here for clarity as a subtraction of the comparator reference input, which is the output of the error amplifier) allows the peak current control to more closely approximate the average current.

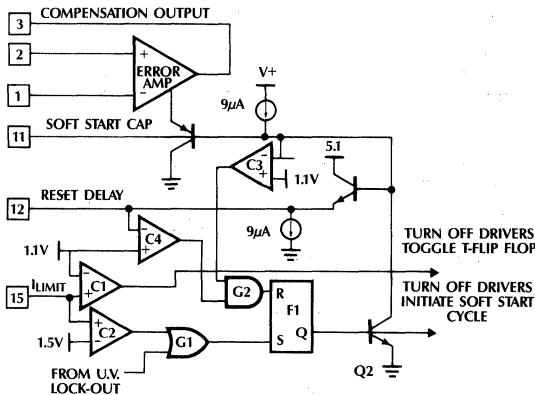


Figure 13. Current Limit, Soft Start and Reset Delay

UNDERVOLTAGE LOCKOUT

In the circuit in Figure 14, the ML4809 remains in a low quiescent drain (1.1mA) during T1 while C1 charges through R(S) to 16V. After V_{CC} rises to 16V the ML4809 begins running. C1 provides the energy needed to run the gate drive and ML4809 until the auxiliary winding can provide sustaining energy for the control circuit, preventing C1 from draining below the 9V lockout threshold. The 7V of hysteresis in the Undervoltage Lockout circuit allows the ML4809 to start from a bleed resistor/capacitor easily. While the ML4809 is in the standby (Lockout) condition, OUTA and OUTB will be pulled low.

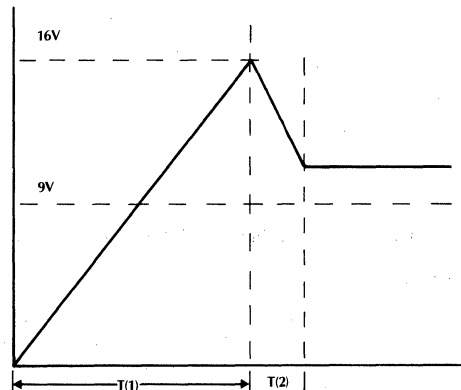
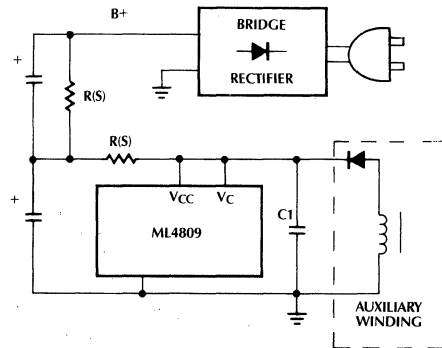


Figure 14. Typical Off-Line Start-Up Circuit and Timing

In the actual implementation, an external resistor (pin 23) sets a current which will be equal to $V_{RAMP}/R1$ and will appear on the comparator input pin. Since the sense resistor is a low impedance point, putting another resistor ($R2$) in series with the V_{SENSE} pin (10) causes a voltage to add to the ramp voltage which is equal to $V_{RAMP} (R2/R1)$.

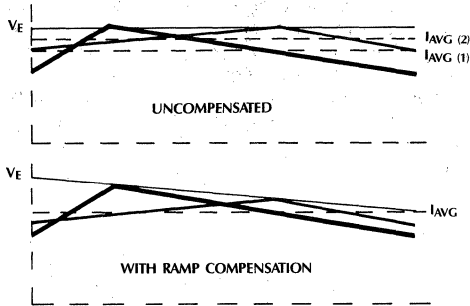


Figure 15. The Effect of Ramp Compensation

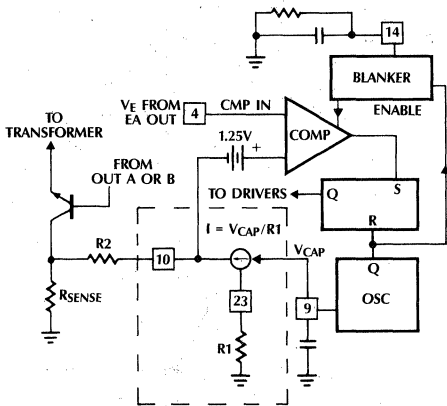


Figure 16. Ramp Compensation and Blanker Block Diagram

MAIN COMPARATOR BLANKER

When CMP IN (EA OUT) is at a low level, spikes which occur on RAMP (which is connected to a current sense resistor or transformer) when the power MOSFETs first turn on can cause the cycle to terminate early. The result of early termination can cause instabilities. Three problems occur which all contribute to this spike.

1. Inductance in the sense resistor.
2. Inter-winding capacitance in the transformer.
3. Reverse recovery current in the rectifier in the opposite FET intrinsic diode (or from the secondary diodes).

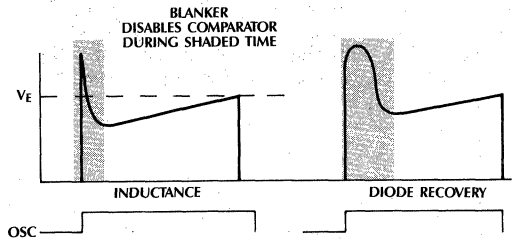


Figure 17. Unintended Early Cycle Termination

The first two problems usually cause a fairly short spike which is easy to filter out with just a simple RC before the comparator input without causing unacceptable phase delay at the input, since there is not much area underneath the spike. The third problem can have significant energy, and a filter with a low enough pole to reduce the "spike" to a level low enough not to cause early cycle termination would cause excessive phase shift.

The solution is to provide a blanking pulse to the comparator at the beginning of the cycle. The width of this pulse is programmed by an external RC. When CLOCK IN is high, a buffer in the ML4809 charges the capacitor on pin 14 to 4V. When CLOCK goes low, the capacitor discharges through the external resistor. The outputs are held low until the voltage at pin 14 falls below 3.2V. The buffer driving pin 14 is limited to 5mA output current. The Blanking period can be calculated by the expression:

$$T_{BLANK} \approx \frac{R_{BLANK} C_{BLANK}}{2.83}$$

SYNCHRONIZATION INPUTS AND OUTPUTS

When using the Clock (pin 7) or Q (pin 13) outputs, a 5KΩ pull down resistor is recommended. These outputs are open emitters. Clock has an internal (375μA) current sink load while Q is unloaded. Both will exhibit significant timing skew due to PC board capacitance if not loaded.

Clock Output and External Clock Input

Used to synchronize multiple supplies. For synchronized operation of multiple ML4809's, tie the CLOCK OUT from the "master" to the CLOCK IN of the slaves.

Toggle Preset and Q Output

In multiple supply systems, this is important for synchronization. To synchronize multiple chips, connect the Q output from the "master" ML4809 to the Preset Input of the "slave" in a "daisy chain". For non-synchronized operation this input would be connected to GND.

OTHER FEATURES

Fast Shut Down Path from Current Limit to Outputs

Provides a 30ns path to the outputs which begins to turn off the outputs while the longer latching path is propagating. In a normal UC1825, it can be as much as 80ns until the over-current condition shuts down the outputs.

Separate Error Amplifier Output Pin for Loop Filtering Versatility

This is especially useful for:

1. Diagnostic purposes, to see what the chip is really doing, it is useful to break the feedback loop.
2. High power supplies — current sharing: in system design with more than one supply running, in order to ensure that the supplies share current equally it is often necessary to have a “master” circuit control the PWM operation of each of the “slaves”. This is most easily accomplished by an “or” (where the lowest output dominates) of the Error Amp outputs, which is impossible if the output of the amp is internally connected to the input of the comparator.

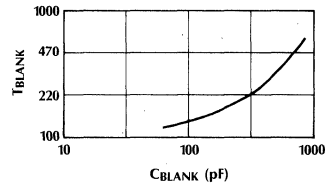


Figure 18. T_{BLANK} vs. C_{BLANK} (R_{BLANK} = 5.1KΩ)

APPLICATIONS

Figure 19 shows the ML4809 in a push-pull non-isolated application. Note the Schottky Diodes on pins 17 and 20. These diodes are necessary in order to prevent transients from driving these pins negative with respect to GND which would cause the IC to malfunction.

Care should be exercised in layout:

1. Avoid Ground Loops. Use “star” grounding.
2. Bypass the V_{CC} line with a high frequency capacitor which is physically close to the IC.
3. Avoid running signal lines near power lines.
4. Employ “ground planing”.

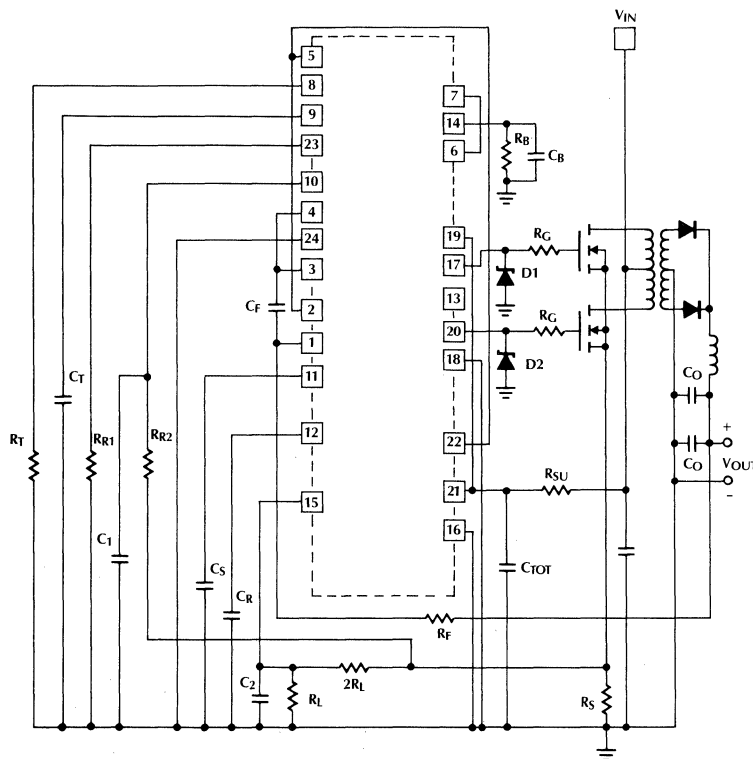


Figure 19. ML4809 Typical Application

ML4809

ORDERING INFORMATION

PART NUMBER	TEMPERATURE RANGE	PACKAGE
ML4809CP	0°C to +70°C	Molded DIP (P24N)
ML4809CQ	0°C to +70°C	Molded PCC (Q28)
ML4809IP	-40°C to +85°C	Molded DIP (P24N)
ML4809IQ	-40°C to +85°C	Molded PCC (Q28)
ML4809MJ	-55°C to +125°C	Hermetic DIP (J24N)

ML4810, ML4811

High Frequency Power Supply Controller

GENERAL DESCRIPTION

The ML4810 and ML4811 High Frequency PWM Controllers are optimized for use in Switch Mode Power Supply designs running at frequencies to 1MHz. The ML4810/11 contain a unique overload protection circuit which helps to limit stress on the output devices and reliably performs a soft-start reset. Propagation delays are minimal through the comparators and logic for reliable high frequency operation and slew rate and bandwidth are maximized on the error amplifier. These controllers are designed to work in either voltage or current mode and provide for input voltage feed forward.

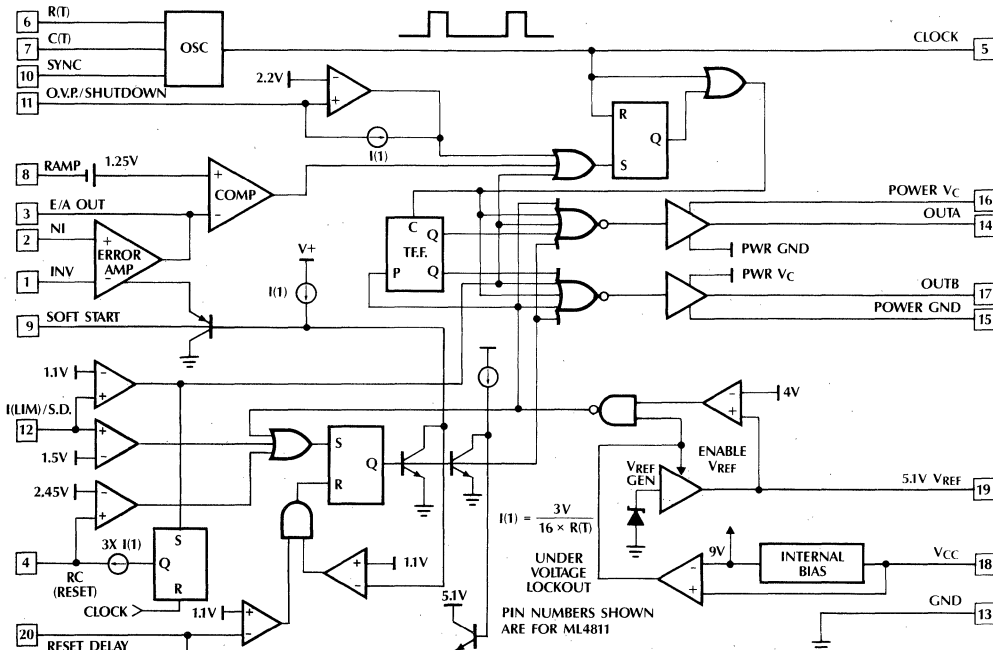
A 1.1V threshold current limit comparator provides a cycle-by-cycle current limit. An integrating circuit "counts" the number of times the 1.1V limit was reached. A soft-start cycle is initiated if the cycle-by-cycle current limit is repeatedly activated. A reset delay function is provided on the ML4811. All logic is fully latched to provide jitter-free operation and prevent multiple pulsing. An under-voltage lockout circuit with 7V of hysteresis assures low startup current and drives the outputs low during fault condition.

The ML4810/11 are fabricated on a 40V bipolar process from the FB3480 Power Supply Controller Array. Customized versions of this controller can therefore be easily implemented. Please refer to the FB3480 datasheet for more information. These controllers are similar to the UC1825 controller, however these controllers include many features not found on the 1825. These features are set in *Italics*.

FEATURES

- *Integrating Soft Start Reset*
- High Current (2A peak) Dual Totem Pole Outputs
- Practical Operation to 1MHz (f_{OSC})
- 5.1V, $\pm 1\%$ Trimmed Bandgap Reference
- *Under Voltage Lockout with 7V Hysteresis*
- *Soft Start Reset Delay (ML4811)*
- *Oscillator Synchronization Function (ML4811)*
- *Soft Start latch ensures full soft start cycle*
- *Outputs pull low for undervoltage lockout*
- *Accurately controlled Oscillator ramp discharge current*
- *All timing currents "slaved" to R(T) for precise control*

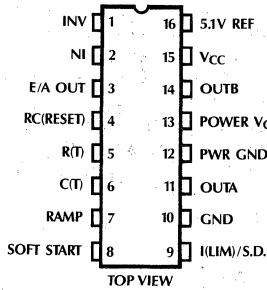
ML4811 BLOCK DIAGRAM


6

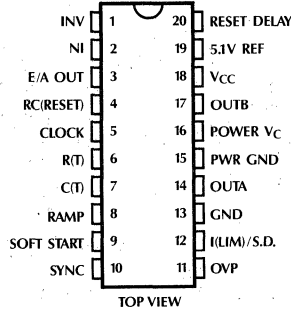
ML4810, ML4811

PIN CONNECTIONS

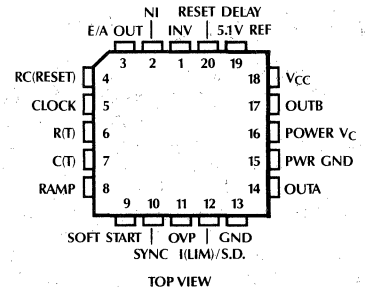
ML4810
16-Pin DIP



ML4811
20-Pin DIP



ML4811
20-Pin PCC



PIN DESCRIPTION (Pin numbers shown for ML4811)

PIN NO.	NAME	FUNCTION	PIN NO.	NAME	FUNCTION
1	INV	Inverting input to error amp	11	OVP	Exceeding 2.5V terminates the PWM cycle and inhibits the outputs
2	NI	Non-inverting input to error amp	12	I(LIM)/S.D.	Current limit sense pin. Normally connected to current sense resistor
3	E/A Out	Output of error amplifier and input to main comparator	13	GND	Analog Signal Ground
4	RC(RESET)	Timing elements for Integrating Soft Start reset	14	OUTA	High Current Totem pole output. This output is the first one energized after Power On Reset
5	CLOCK	Oscillator output.	15	PWR GND	Return for the High Current Totem pole outputs
6	R(T)	Timing Resistor for Oscillator — sets charging current for oscillator timing capacitor (Pin 6)	16	V _C	Positive Supply for the High Current Totem pole outputs
7	C(T)	Timing Capacitor for Oscillator	17	OUTB	High Current Totem pole output
8	RAMP	Non-Inverting input to main comparator. Connected to C(T) for Voltage Mode operation or to current sense resistor for current mode	18	V _{CC}	Positive Supply for the IC
9	SOFT START	Normally connected to Soft Start Capacitor	19	5.1V REF	Buffered output for the 5.1V voltage reference
10	SYNC	A high going pulse terminates the PWM cycle and discharges C(T)	20	RESET DELAY	Timing Capacitor to determine the amount of delay between fault

ABSOLUTE MAXIMUM RATINGS

Supply Voltage (Pins 18, 16)	25V
Output Current, Source or Sink (Pins 14, 17)	
DC	0.5A
Pulse (0.5 μ S)	2.0A
Analog Inputs	
(Pins 1, 2, 8)	-0.3V to 7V
(Pins 9, 10, 11, 12, 20)	-0.3V to 6V
Clock Output Current (Pin 5)	-5mA
Error Amplifier Output Current (Pin 3)	5mA
Junction Temperature	
ML4811M	150°C
ML4811I, ML4810C, ML4811C, ML4810I	125°C
Storage Temperature Range	-65°C to +150°C

Lead Temperature (Soldering 10 sec.)	+260°C
Thermal Resistance (θ_{JA})	
Plastic DIP	65°C/W
Ceramic DIP	65°C/W
Plastic SOIC	65°C/W

OPERATING CONDITIONS

Temperature Range	
ML4811M	-55°C to +125°C
ML4811I	-40°C to +85°C
ML4810C, ML4811C	0°C to +70°C

Absolute maximum ratings are those values beyond which the device could be permanently damaged. Absolute maximum ratings are stress ratings only and functional device operation is not implied. Pin numbers given for ML4811.

ELECTRICAL CHARACTERISTICS

Unless otherwise specified, $R_T = 3.65K\Omega$, $C_T = 1000pF$, $T_A =$ Operating Temperature Range, $V_{CC} = 15V$.

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS	
Oscillator						
Initial Accuracy	$T_J = 25^\circ C$ (note 1)	360	400	440	KHz	
Voltage Stability	$10V < V_{CC} < 25V$, (note 1)		0.2	4	%	
Temperature Stability	(note 1)			5	%	
Total Variation	line, temp. (note 1)	340		460	KHz	
Clock Out High		3.9	4.5		V	
Clock Out Low			2.3	2.9	V	
Ramp Peak			2.8		V	
Ramp Valley			1.0		V	
Ramp Valley to Peak		1.6		2.3	V	
Sync Input Threshold		0.8	1.0	1.4	V	
Sync Input Current	$V_{PIN 10} = 4V$				μA	
Reference Section						
Output Voltage	ML4810/11C	$T_J = 25^\circ C, I_O = 1mA$	5.00	5.10	5.2	V
	ML4811M, ML4810/11I		5.05	5.10	5.15	V
Line Regulation	$10V < V_{CC} < 25V$		2	20	mV	
Load Regulation	$1mA < I_O < 10mA$.5	20	mV	
Temperature Stability	$-55^\circ C < T_J < 150^\circ C$, (note 1)		.2	.4	%	
Total Variation	ML4810/11C	line, load, temp. (note 1)	4.95		5.25	V
	ML4811M, ML4810/11I		5.0		5.20	V
Output Noise Voltage	10Hz to 10KHz		50		μV	
Long Term Stability	$T_J = 125^\circ C$, 1000 Hrs (note 1)		5	25	mV	
Short Circuit Current	$V_{REF} = 0V$	-15	-50	-100	mA	
Under-Voltage Lockout Section						
Start Threshold		15	16	17	V	
UVLO Hysteresis		6.5	7	7.5	V	

ML4810, ML4811

ELECTRICAL CHARACTERISTICS (Continued)

Unless otherwise specified, $R_T = 3.65k\Omega$, $C_T = 1000pF$, $T_A =$ Operating Temperature Range, $V_{CC} = 15V$.

PARAMETER		CONDITIONS	MIN	TYP	MAX	UNITS
Supply Current						
Start Up Current	ML4810	$V_{CC} = 8V, T_A \geq 0^\circ C$ $V_{CC} = 8V, T_A < 0^\circ C$		2.0 2.5	3.5 4.0	mA mA
	ML4811	$V_{CC} = 8V, T_A \geq 0^\circ C$ $V_{CC} = 8V, T_A < 0^\circ C$		2.5 3	4.0 4.5	mA mA
I_{CC}	ML4810	$V_{PIN\ 1, 7, 9} = 0V, V_{PIN\ 2} = 1V, T_A = 25^\circ C$		32	46	mA
	ML4811	$V_{PIN\ 1, 7, 9} = 0V, V_{PIN\ 2} = 1V, T_A = 25^\circ C$		38	55	mA
Error Amplifier Section						
Input Offset Voltage					± 20	mV
Input Bias Current				.6	3	μA
Input Offset Current				.1	1	μA
Open Loop Gain		$1 < V_O < 4V$	60	96		dB
CMRR		$1.5 < V_{CM} < 5.5V$	65	95		dB
PSRR		$10 < V_{CC} < 30V$	75	90		dB
Output Sink Current		$V_{PIN\ 3} = 1V$	1	2.5		mA
Output Source Current		$V_{PIN\ 3} = 4V$	-5	-1.3		mA
Output High Voltage		$I_{PIN\ 3} = -0.5mA$	4.0	4.7	5.0	V
Output Low Voltage		$I_{PIN\ 3} = 1mA$	0	0.5	1.0	V
Unity Gain Bandwidth		(note 1)	3	5.5		MHz
Slew Rate		(note 1)	6	12		V/ μs
PWM Comparator Section						
Pin 8 Bias Current		$V_{PIN\ 8} = 0V$		-1	-5	μA
Duty Cycle Range			0		75	%
Pin 3 Zero DC Threshold			1.1	1.25		V
Delay to Output		(note 1)		50	80	ns
Soft-Start Section						
Charge Current (Pin 9)	ML4811	$V_{PIN\ 9} = 1V, V_{PIN\ 4, 12} = 0$	-35	-55	-75	μA
Discharge Current (Pin 9)		$V_{PIN\ 9} = 3V, V_{PIN\ 4} > 2.5$	1	5		mA
		$V_{PIN\ 9} = 3V, V_{PIN\ 12} > 1.65, V_{PIN\ 4} < 2$	1	5		mA
Charge Current (Pin 20)		$V_{PIN\ 20} = 1V$.1	5		mA
Discharge Current (Pin 20)		Requires external discharge resistor		0		μA
Current Limit/Shutdown Section						
Pin 12 Bias Current	ML4810/11C	$0V < V_{PIN\ 12} < 4V$			+15	μA
	ML4811M, ML4811I	$0V < V_{PIN\ 12} < 4V$			+10	μA
Current Limit Threshold	ML4810		1.2	1.3	1.4	V
	ML4811		.95	1.1	1.3	V
Reset Threshold (Pin 12)	ML4810	$V_{PIN\ 4} < 2V$	1.60	1.75	1.90	V
	ML4811	$V_{PIN\ 4} < 2V$	1.4	1.50	1.8	V
Delay to Output		(note 1)		40	70	ns
Pin 4 Charging Current		$V_{PIN\ 12} = 2V$	120	150	180	μA
Restart Threshold (Pin 4)			2	2.45	3	V
OVP Shutdown Threshold (Pin 11)			2.4	2.7	2.8	V
OVP Input Current		$V_{PIN\ 11} = 3V$	40	50	60	μA
Charge Current (Pin 8)	ML4810	$V_{PIN\ 8} = 1V, V_{PIN\ 4, 9} = 0$	-40	-50	-60	μA

out" the surge with the switch protected by the cycle by cycle limit. R1 and C1 can be selected to track diode heating, or to ride out various system surge requirements as required.

If the high current demanded is caused by a short circuit, the duty cycle will be short and the output diodes will carry the current for the majority of the PWM cycle. C2 charges fastest for low duty cycles (since F3 will be on for a longer time) providing for quicker shutdown during short-circuit when the output diodes are being maximally stressed.

OSCILLATOR

The ML4811 oscillator charges the external capacitor (C_T) with a current (I_{SET}) equal to 3/R_T. When the capacitor voltage reaches the upper threshold (Ramp

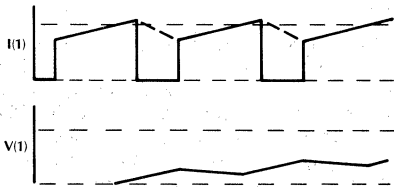


Figure 4. Switching Current and Pin 4 Voltage — Normal

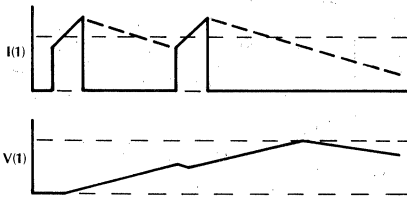


Figure 5. Switching Current and Pin 4 Voltage — Load Fault

Peak), the comparator changes state and the capacitor discharges to the lower threshold (Ramp Valley) through Q1. While the capacitor is discharging, Q2 provides a high pulse. A discharge of the oscillator can be initiated by applying a high level to the Sync pin. A short pulse of a frequency higher than the oscillator's free running frequency can be used to synchronize the ML4811 to an external clock. The pulse can be equal to the desired deadtime (T_D) or the deadtime can be determined by I_{DIS} and C_T, whichever is greater.

The Oscillator period can be described by the following relationship:

$$T_{OSC} = T_{RAMP} + T_{DEADTIME}$$

where: $T_{RAMP} = C (Ramp\ Valley\ to\ Peak) / I_{SET}$
and: $T_{DEADTIME} = C (Ramp\ Valley\ to\ Peak) / I_{Q1}$

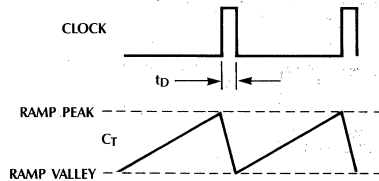
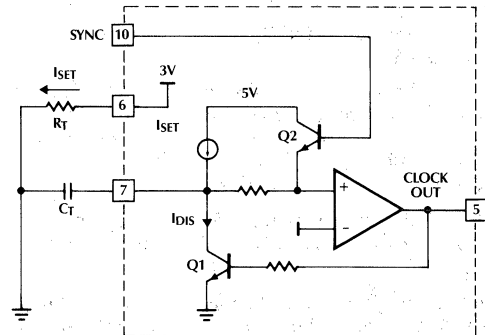


Figure 6. Simplified Oscillator Block Diagram and Timing

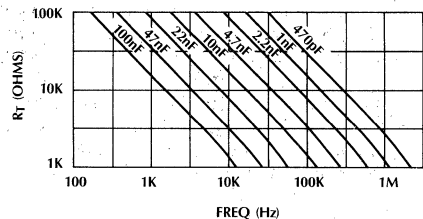


Figure 7. Oscillator Timing Resistance vs. Frequency

ERROR AMPLIFIER

The ML4811 error amplifier is a 5.5MHz bandwidth 12V/ μ sec slew rate op-amp with provision for limiting the positive output voltage swing (Output Inhibit line) for ease in implementing the soft start function.

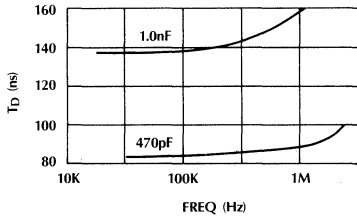


Figure 8. Oscillator Deadtime vs Frequency

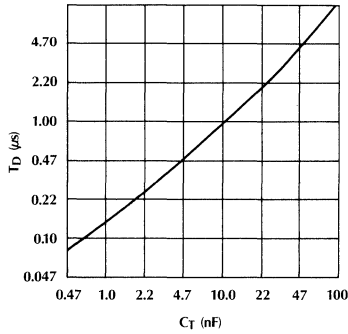


Figure 9. Oscillator Deadtime vs $C(T)$ ($3 \leq R(T) \leq 100K\Omega$)

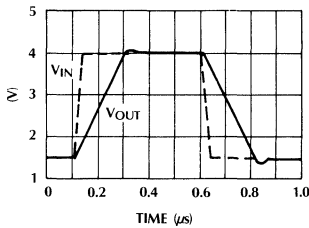


Figure 10. Unity Gain Slew Rate

OUTPUT DRIVER STAGE

The ML4811 Output Driver is a 2A peak output high speed totem pole circuit designed to quickly switch the gates of capacitive loads, such as power MOSFET transistors.

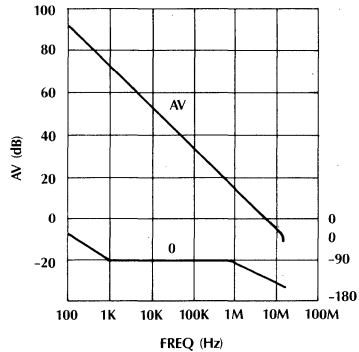


Figure 11. Open Loop Frequency Response

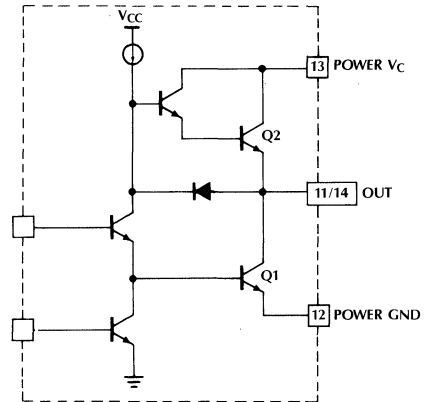


Figure 12. Simplified Schematic

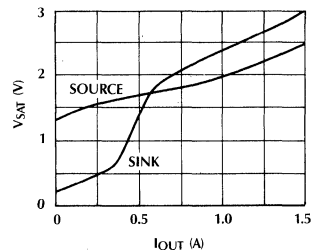


Figure 13. Saturation Curves

ML4810, ML4811

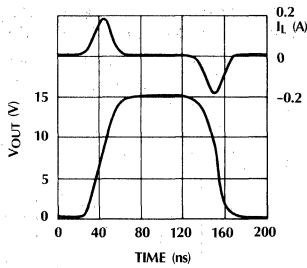


Figure 14. Rise/Fall Time ($C_L = 1000\text{pF}$)

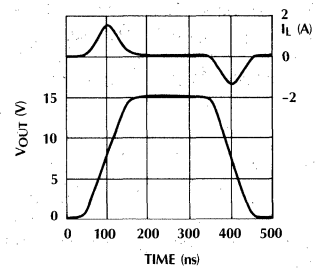


Figure 15. Rise/Fall Time ($C_L = 10,000\text{pF}$)

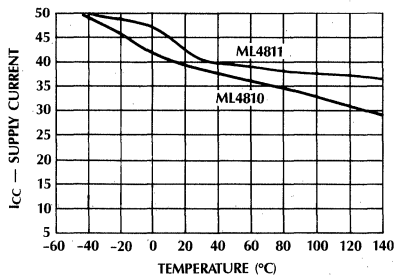


Figure 16. Supply Current vs. Temperature

ORDERING INFORMATION

PART NUMBER	TEMPERATURE RANGE	PACKAGE
ML4810CP	0°C to +70°C	16-Pin MOLDED DIP (P16)
ML4810CS	0°C to +70°C	16-Pin MOLDED SOIC (S16W)
ML4811CP	0°C to +70°C	20-Pin MOLDED DIP (P20)
ML4811CS	0°C to +70°C	20-Pin MOLDED SOIC (S20W)
ML4811IP	-40°C to +85°C	20-Pin MOLDED DIP (P20)
ML4811MJ	-55°C to +125°C	20-Pin HERMETIC DIP (J20)

Power Factor Controller

GENERAL DESCRIPTION

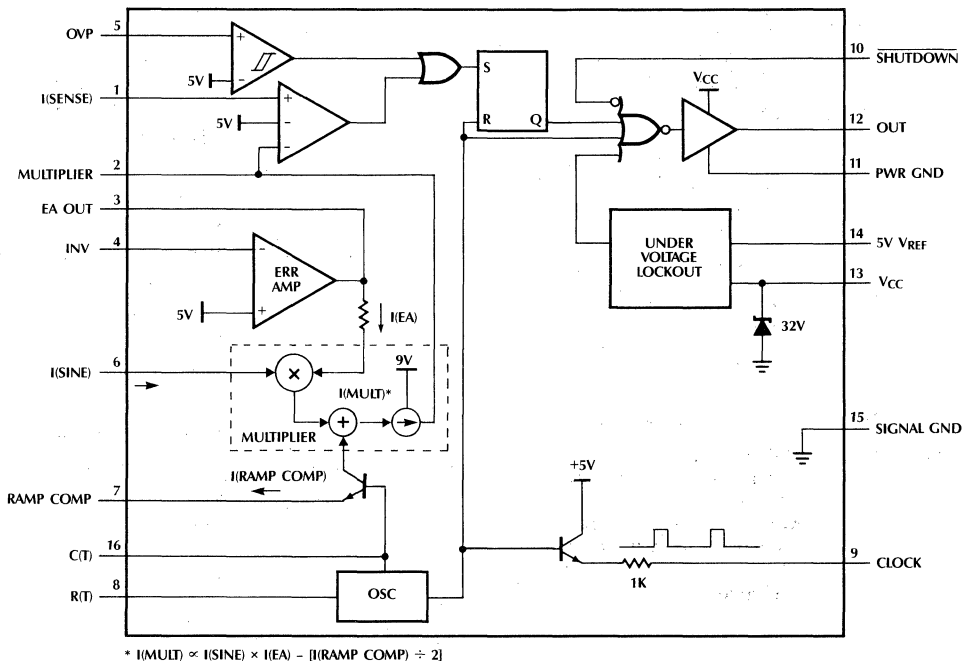
The ML4812 is designed to optimally facilitate a "boost" type power factor correction system. Special care has been taken in the design of the ML4812 to increase system noise immunity. The circuit includes a precision reference, multiplier, error amplifier, over-voltage protection, ramp compensation, as well as a high current output. In addition, start-up is simplified by an under-voltage lockout circuit with 6V hysteresis.

In a typical application, the ML4812 functions as a current mode regulator. The current which is necessary to terminate the cycle is a product of the sinusoidal line voltage times the output of the error amplifier which is regulating the output DC voltage. Ramp compensation is programmable with an external resistor, to provide stable operation when the duty cycle exceeds 50%.

FEATURES

- Precision buffered 5V reference ($\pm 0.5\%$)
- Current Input Multiplier reduces external components and improves noise immunity
- Programmable Ramp Compensation circuit
- 1A Peak Current Totem-Pole Output Drive
- Over-Voltage comparator eliminates output "runaway" due to load removal
- Wide common mode range in current sense comparators for better noise immunity
- Large oscillator amplitude for better noise immunity

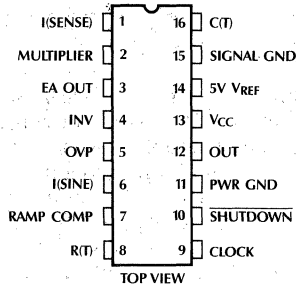
BLOCK DIAGRAM (Pin Out shown is for DIP)



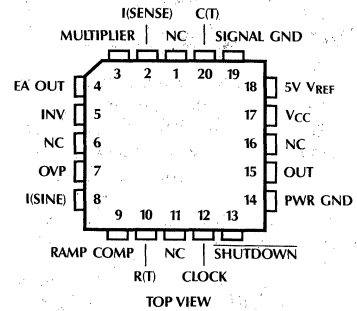
ML4812

PIN CONNECTIONS

ML4812
16-Pin DIP



ML4812
20-Pin PCC



PIN DESCRIPTION (DIP)

PIN NO.	NAME	FUNCTION	PIN NO.	NAME	FUNCTION
1	I(SENSE)	Input from the Current Sense Transformer (T1) to the PWM comparator (+).	8	R(T)	Oscillator timing resistor pin. A 5V source sets a current in the external resistor which is mirrored to charge C(T).
2	MULTIPLIER	Output of Current Multiplier. A resistor to ground on this pin converts the current to a voltage. This pin is clamped to 5V and tied to the PWM comparator (-).	9	CLOCK	Digital clock output.
3	EA OUT	Output of error amplifier.	10	SHUTDOWN	A TTL compatible low level on this pin turns off the output.
4	INV	Inverting input to error amplifier.	11	PWR GND	Return for the High Current Totem pole output.
5	OVP	Input to over voltage comparator.	12	OUT	High Current Totem pole output.
6	I(SINE)	Current Multiplier Input.	13	VCC	Positive Supply for the IC.
7	RAMP COMP	Buffered output from the Oscillator Ramp [C(T)]. A resistor to ground sets the current which is internally subtracted from the product of I(SINE) and I(EA) in the multiplier.	14	5V VREF	Buffered output for the 5V voltage reference.
			15	SIGNAL GND	Analog signal ground.
			16	C(T)	Timing Capacitor for the Oscillator.

ABSOLUTE MAXIMUM RATINGS

Supply Current (I_{CC})	30mA
Output Current, Source or Sink (Pin 12)	
DC	1.0A
Output Energy (capacitive load per cycle)	5 μ J
Multiplier I(SINE) Input (pin 6)	1.2mA
Error Amp Sink Current (pin 3)	10mA
Oscillator Charge Current	2mA
Analog Inputs (pins 1, 4, 5)	-0.3V to 5.5V
Junction Temperature	150°C
Storage Temperature Range	-65°C to +150°C
Lead Temperature (soldering 10 sec.)	+260°C

Thermal Resistance (θ_{JA})

Plastic Chip Carrier (PCC) — Q	60°C/W
Plastic DIP — P	65°C/W
Ceramic DIP — J	65°C/W

OPERATING CONDITIONS

Temperature Range

ML4812C	0°C to +70°C
ML4812I	-40°C to +85°C
ML4812M	-55°C to +125°C

Absolute maximum ratings are those values beyond which the device could be permanently damaged. Absolute maximum ratings are stress ratings only and functional device operation is not implied.

ELECTRICAL CHARACTERISTICS

Unless otherwise specified, $R_T = 14K\Omega$, $C_T = 1000pF$,
 T_A = Operating Temperature Range, $V_{CC} = 15V$ (note 2), Pin numbers refer to 16-pin DIP package.

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
Oscillator					
Initial Accuracy	$T_J = 25^\circ C$	91	98	105	KHz
Voltage Stability	$12V < V_{CC} < 18V$		0.3		%
Temperature Stability			2		%
Total Variation	line, temp.	90		108	KHz
Ramp Valley to Peak			3.3		V
R(T) Voltage		4.8	5.0	5.2	V
Discharge Current (pin 8 open)	$T_J = 25^\circ C, V_{PIN 16} = 2V$	7.8	8.4	9.0	mA
	$V_{PIN 16} = 2V$	7.3	8.4	9.3	V
Clock Out Voltage Low	$R_L = 16K\Omega$		0.2	0.5	V
Clock Out Voltage High	$R_L = 16K\Omega$	3.0	3.5		V
Reference Section					
Output Voltage	$T_J = 25^\circ C, I_O = 1mA$	4.95	5.00	5.05	V
Line Regulation	$12V < V_{CC} < 25V$		2	20	mV
Load Regulation	$1mA < I_O < 20mA$		2	20	mV
Temperature Stability			0.4		%
Total Variation	line, load, temp	4.9		5.1	V
Output Noise Voltage	10Hz to 10KHz		50		μ V
Long Term Stability	$T_J = 125^\circ C, 1000$ Hrs (note 1)		5	25	mV
Short Circuit Current	$V_{REF} = 0V$	-30	-85	-180	mA
Error Amplifier Section					
Input Offset Voltage				± 15	mV
Input Bias Current			-0.1	-1.0	μ A
Open Loop Gain	$1 < V_{PIN 3} < 5V$	60	75		dB
PSRR	$12V < V_{CC} < 25V$	60	75		dB
Output Sink Current	$V_{PIN 3} = 1.1V, V_{PIN 4} = 6.2V$	2	12		mA
Output Source Current	$V_{PIN 3} = 5.0V, V_{PIN 4} = 4.8V$	-0.5	-1.0		mA
Output High Voltage	$I_{PIN 3} = -0.5mA, V_{PIN 4} = 4.8V$	5.3	5.5		V
Output Low Voltage	$I_{PIN 3} = 1mA, V_{PIN 4} = 6.2V$		0.5	1.0	V
Unity Gain Bandwidth			1.0		MHz

ML4812

ELECTRICAL CHARACTERISTICS (Continued)

Unless otherwise specified, $R_T = 14K\Omega$, $C_T = 1000pF$, $T_A =$ Operating Temperature Range, $V_{CC} = 15V$ (note 2), Pin numbers refer to 16-pin DIP package.

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
Multiplier					
I(SINE) Input Voltage	I(SINE) = 500 μ A	.4	.7	.9	V
Output Current (pin 2)	I(SINE) = 500 μ A, PIN 4 = $V_{REF} - 20mV$	460	480	510	μ A
	I(SINE) = 500 μ A, PIN 4 = $V_{REF} + 20mV$		3	10	μ A
	I(SINE) = 1mA, PIN 4 = $V_{REF} - 20mV$	900	950	1020	μ A
	I(SINE) = 500 μ A, PIN 4 = $V_{REF} - 20mV$, $I_{PIN 7} = 50\mu A$		455		μ A
Bandwidth			200		KHz
PSRR	$12V < V_{CC} < 25V$		70		dB
OVP Comparator					
Input Offset Voltage	Output Off	-25		+5	mV
Hysteresis	Output On	95	105	115	mV
Input Bias Current			-0.3	-3	μ A
Propagation Delay			150		nS
PWM Comparator: I(SENSE)					
Input Offset Voltage				± 15	mV
Input Offset Current				± 1	mA
Input Common Mode Range		-0.2		-5.5	V
Input Bias Current			-2	-10	μ A
Propagation Delay			150		nS
I_{LIMIT} Trip Point	$V_{PIN 2} = 5.5V$	4.8	5	5.2	V
Output Section					
Output Voltage Low	$I_{OUT} = -20mA$		0.1	0.4	V
	$I_{OUT} = -200mA$		1.6	2.2	V
Output Voltage High	$I_{OUT} = 20mA$	13	13.5		V
	$I_{OUT} = 200mA$	12	13.4		V
Output Voltage Low in UVLO	$I_{OUT} = -5mA$, $V_{CC} = 8V$		0.1	0.8	V
Output Rise/Fall Time	$C_L = 1000pF$		50		nS
Shut Down Input	V_{IH}	2.0			V
	V_{IL}			0.8	V
	I_{IL} , $V_{PIN 10} = 0V$			-1.5	mA
	I_{IH} , $V_{PIN 10} = 5V$			10	μ A
Under-Voltage Lockout					
Start-Up Threshold		15	16	17	V
Shut-Down Threshold		9	10	11	V
V_{REF} Good Threshold			4.4		V
Total Device					
Supply Current	Start-Up, $V_{CC} = 14V$, $T_J = 25^\circ C$.8	1.2	mA
	Operating, $T_J = 25^\circ C$		20	25	mA
Internal Shunt Zener Voltage	$I_{CC} = 30mA$	25	30	34	V

Note 1: This parameter not 100% tested in production but guaranteed by design.

Note 2: V_{CC} is raised above the Start-Up Threshold first to activate the IC, then returned to 15V.

FUNCTIONAL DESCRIPTION

OSCILLATOR

The ML4812 oscillator charges the external capacitor (C_T) with a current (I_{SET}) equal to $5/R_{SET}$. When the capacitor voltage reaches the upper threshold, the comparator changes state and the capacitor discharges to the lower threshold through Q1. While the capacitor is discharging, Q2 provides a high pulse.

The Oscillator period can be described by the following relationship:

$$T_{OSC} = T_{RAMP} + T_{DEADTIME}$$

where:

$$T_{RAMP} = \frac{C_T \times V_{RAMP \text{ VALLEY TO PEAK}}}{I_{SET}}$$

and:

$$T_{DEADTIME} = \frac{C_T \times V_{RAMP \text{ VALLEY TO PEAK}}}{8.4mA - I_{SET}}$$

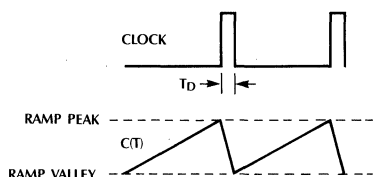
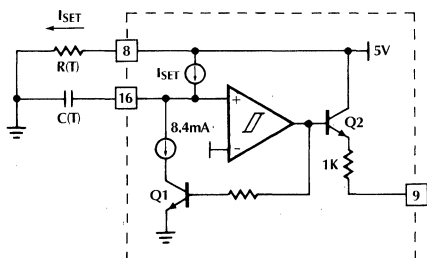


Figure 1. Oscillator Block Diagram

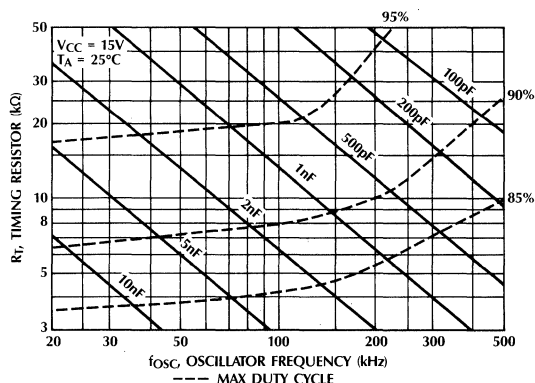


Figure 2. Oscillator Timing Resistance vs. Frequency

OUTPUT DRIVER STAGE

The ML4812 Output Driver is a 1A peak output high speed totem pole circuit designed to quickly drive capacitive loads, such as power MOSFET gates.

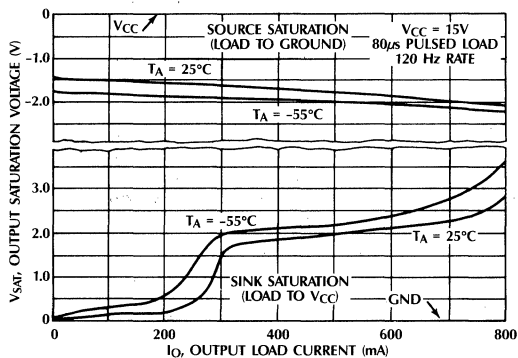


Figure 3. Output Saturation Voltage vs. Output Current

ERROR AMPLIFIER

The ML4812 error amplifier is a high open loop gain, wide bandwidth, amplifier.

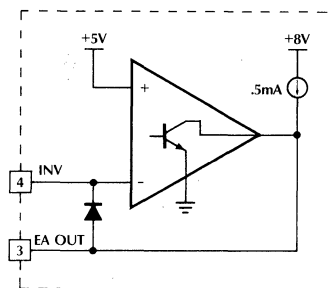


Figure 4. Error Amplifier Configuration

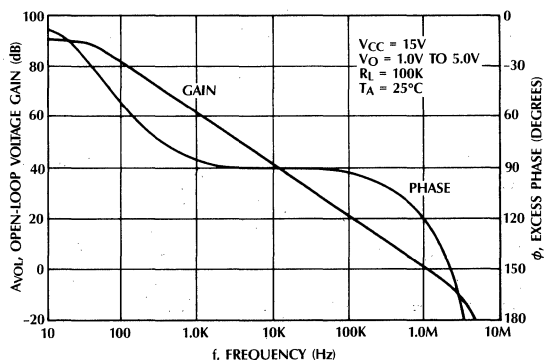


Figure 5. Error Amplifier Open-Loop Gain and Phase vs. Frequency

MULTIPLIER

The ML4812 multiplier is a linear current input multiplier to provide high immunity to the disturbances caused by high power switching. The rectified line input sine wave is converted to a current via a dropping resistor. In this way, small amounts of ground noise produce an insignificant effect on the reference to the PWM comparator.

The output of the multiplier is a current proportional to:

$$I_{OUT} \propto I(SINE) \times I(EA)$$

where $I(SINE)$ is the current in the dropping resistor, and $I(EA)$ is a current proportional to the output of the error amplifier. When the error amplifier is saturated high, the output of the multiplier is approximately equal to the $I(SINE)$ input current.

The multiplier output current is converted into the reference voltage for the PWM comparator through a resistor to ground on the multiplier output. The multiplier output is clamped to 5V to provide current limiting.

Ramp compensation is accomplished by subtracting 1/2 of the current flowing out of pin 7 through a buffer transistor driven by C(T) which is set by an external resistor.

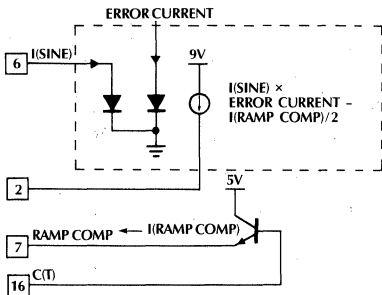


Figure 6. Multiplier Block Diagram

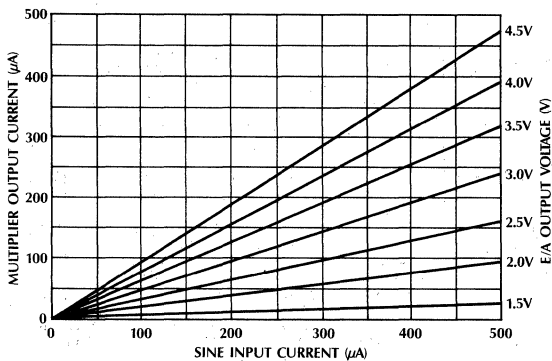


Figure 7. Multiplier Linearity

UNDER VOLTAGE LOCKOUT

On power-up the ML4812 remains in the UVLO condition; output low and quiescent current low. The IC becomes operational when V_{CC} reaches 16V. When V_{CC} drops below 10V, the UVLO condition is imposed. During the UVLO condition, the 5V V_{REF} pin is "off", making it usable as a "flag" for starting up a downstream PWM converter.

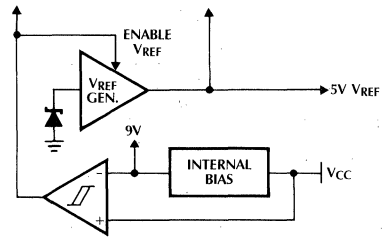


Figure 8. Under-Voltage Lockout Block Diagram

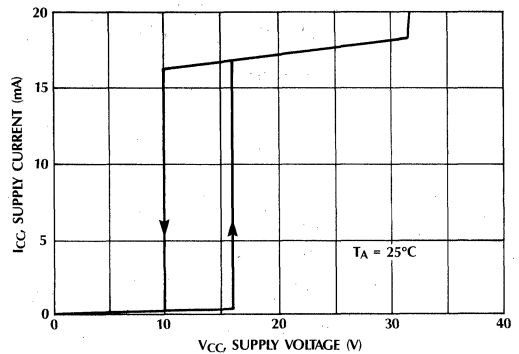


Figure 9a. Total Supply Current vs. Supply Voltage

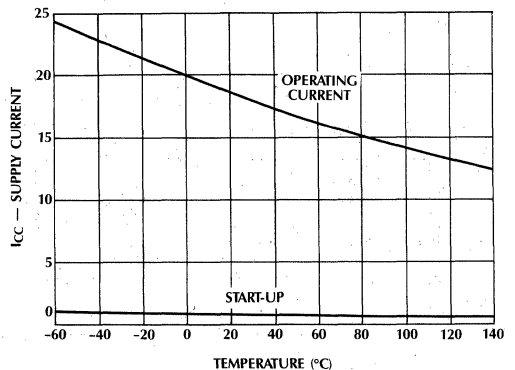


Figure 9b. Supply Current (I_{CC}) vs. Temperature

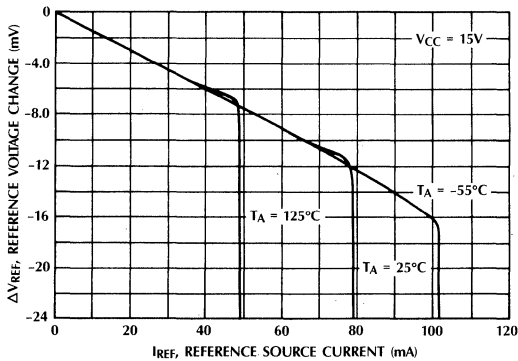


Figure 10. Reference Load Regulation

APPLICATIONS

INPUT INDUCTOR (L1) SELECTION

The central component in the regulator is the input boost inductor. The value of this inductor controls various critical operational aspects of the regulator. If the value is too low, the input current distortion will be high and will result in low power factor and increased noise at the input. This will require more input filtering. In addition, when the value of the inductor is low the inductor dries out (runs out of current) at low currents. Thus the power factor will decrease at lower power levels and/or higher line voltages. If the inductor value is too high, then for a given operating current the required size of the inductor core will be large and/or the required number of turns will be high. So a balance must be reached between distortion and core size.

One more condition where the inductor can dry out is analyzed below where it is shown to be maximum duty cycle dependent.

For the boost converter at steady state:

$$V_{OUT} = \frac{V_{IN}}{1 - D_{ON}} \quad (1)$$

Where D_{ON} is the duty cycle $[T_{ON}/(T_{ON} + T_{OFF})]$. The input boost inductor will dry out when the following condition is satisfied:

$$V_{IN}(t) < V_{OUT} \times (1 - D_{ON}) \quad (2)$$

or

$$V_{INDRY} = [1 - D_{ON}(\max)] \times V_{OUT} \quad (3)$$

V_{INDRY} : Voltage where the inductor dries out.

V_{OUT} : Output dc voltage.

Effectively, the above relationship shows that the resetting volt-seconds are more than setting volt-seconds. In energy transfer terms this means that less energy is stored in the inductor during the ON time than it is asked to deliver during the OFF time. The net result is that the inductor dries out.

The recommended maximum duty cycle is 95% at 100KHz to allow time for the input inductor to dump its energy to the output capacitors.

For example:

$$\begin{aligned} \text{if: } V_{OUT} &= 380V \text{ and} \\ D_{ON}(\max) &= 0.95 \end{aligned}$$

then substituting in (3) yields $V_{INDRY} = 20V$. The effect of drying out is an increase in distortion at low voltages.

For a given output power, the instantaneous value of the input current is a function of the input sinusoidal voltage waveform, i.e. as the input voltage sweeps from zero volts to a maximum value equal to its peak so does the current.

The load of the power factor regulator is usually a switching power supply which is essentially a constant power load. As a result, an increase in the input voltage will be offset by a decrease in the input current.

By combining the ideas set forth above, some ground rules can be obtained for the selection and design of the input inductor:

Step 1: Find minimum operating current.

$$I_{IN}(\min)_{PEAK} = \frac{1.414 \times P_{IN}(\min)}{V_{IN}(\max)} \quad (4)$$

$$V_{IN}(\max) = 260V$$

$$P_{IN}(\min) = 50W$$

then: $I_{IN}(\min)_{PEAK} = 0.272A$

Step 2: Choose a minimum current at which point the inductor current will be on the verge of drying out. For this example 40% of the peak current found in step 1 was chosen.

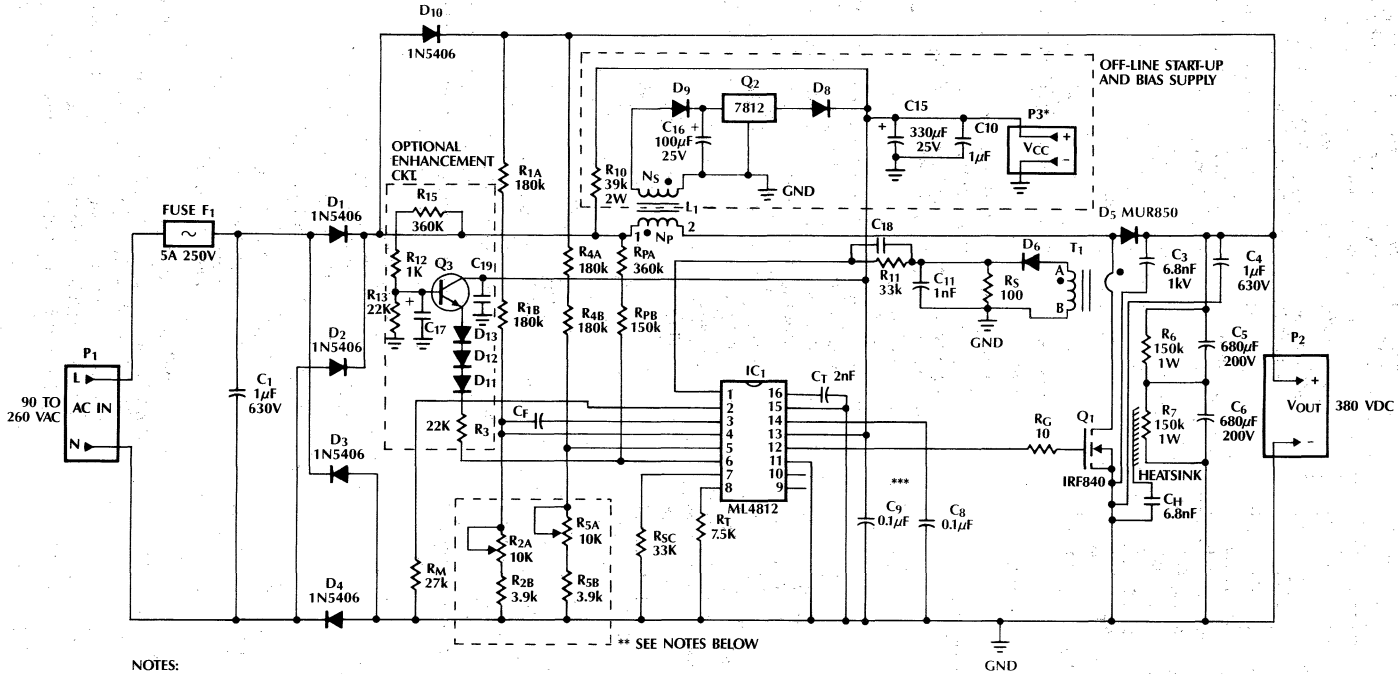
then: $I_{LDRY} = 100mA$

Step 3: The value of the inductance can now be found using previously calculated data.

$$\begin{aligned} L1 &= \frac{V_{INDRY} \times D_{ON}(\max)}{I_{LDRY} \times f_{OSC}} \quad (5) \\ &= \frac{20V \times 0.95}{100mA \times 100KHz} = 2mH \end{aligned}$$

The inductor can be allowed to decrease in value when the current sweeps from minimum to maximum value. This allows the use of smaller core sizes. The only requirement is that the ramp compensation must be adequate for the lower inductance value of the core so that there is adequate compensation at high current.

Figure 11. Typical Application, 200W Power Factor Correction Circuit



NOTES:

1. ALL UNSPECIFIED DIODES ARE 1N4148.
2. ALL UNSPECIFIED RESISTORS ARE 1/4 WATT.
3. ALL UNSPECIFIED CAPACITOR VOLTAGE RATINGS ARE 50V.
4. ADJUST R_{2A} AND R_{5A} WITH CAUTION TO AVOID OVER VOLTAGE CONDITIONS.

Q₃ = 2N2222 OR EQUIVALENT

** SEE NOTES BELOW

GND

* P3 IS USED AT INITIAL TURN-ON TO CHECK THE IC FOR PROPER OPERATION. APPLY = 16VDC.

** FIXED RESISTORS CAN BE USED FOR THE SENSING COMPONENTS. BELOW ARE 1% STANDARD RESISTORS THAT WILL FORCE THE CORRECT OUTPUT VOLTAGES R_{1A}, R_{1B}, R_{4A}, R_{4B} = 178K 1% R_{2B} = 4.75K 1%, R_{5B} = 4.53K 1%. USE JUMPERS INSTEAD OF R_{2A} AND R_{5A} (POTS).

*** FOR HIGHER POWER USE MORE V_{CC} DECOUPLING. 2µF OR MORE MAY BE REQUIRED AT 1KW LEVELS.

Step 4: The presence of the ramp compensation will change the dry out point, but the value found above can be considered a good starting point. Based on the amount of power factor correction the above value of L1 can be optimized after a few iterations.

Gapped Ferrites, Molypermalloy, and Powdered Iron cores are typical choices for core material. The core material selected should have a high saturation point and acceptable losses at the operating frequency.

One ferrite core that is suitable at around 200W is the #4229PL00-3C8 made by Ferroxcube. This ungapped core will require a total gap of 0.180" for this application.

OSCILLATOR COMPONENT SELECTION

The oscillator timing components can be calculated by using the following expression:

$$f_{OSC} = \frac{1.36}{R_T \times C_T} \quad (6)$$

For example:

Step 1: At 100KHz with 95% duty cycle $T_{OFF} = 500\text{ns}$ calculate C_T using the following formula:

$$C_T = \frac{T_{OFF} \times I_{DIS}}{V_{OSC}} = 1000\text{pF} \quad (7)$$

Step 2: Calculate the required value of the timing resistor.

$$R_T = \frac{1.36}{f_{OSC} \times C_T} = \frac{1.36}{100\text{KHz} \times 1000\text{pF}} \quad (8)$$

$$= 13.6\text{K}\Omega \text{ choose } R_T = 14\text{K}\Omega.$$

CURRENT SENSE AND SLOPE (RAMP) COMPENSATION COMPONENT SELECTION

Slope compensation in the ML4812 is provided internally. Rather than adding slope to the noninverting input of the PWM comparator it is actually subtracted from the voltage present at the inverting input of the PWM comparator. The amount of slope compensation should be at least 50% of the downslope of the inductor current during off time as reflected to the inverting input of the PWM comparator. Note that slope compensation is required only when the inductor current is continuous and the duty cycle is more than 50%. The downslope of the inductor current at the verge of discontinuity can be found using the expression given below:

$$\frac{di_L}{dt} = \frac{V_{OUT} - V_{IN DRY}}{L} = \frac{380\text{V} - 20\text{V}}{2\text{mH}} \quad (9)$$

$$= 0.18 \text{ A}/\mu\text{s}$$

The downslope as reflected to the input of the PWM comparator is given by:

$$S_{PWM} = \frac{V_{OUT} - V_{IN DRY}}{L} \times \frac{R_S}{N_C} \quad (10)$$

Where R_S is the current sense resistor and N_C is the turns ratio of the current transformer (T1) used. In general, current transformers simplify the sensing of switch currents especially at high power levels where the use of sense resistors is complicated by the amount of power they have to dissipate. Normally the primary side of the transformer consists of a single turn and the secondary consists of several turns of either enameled magnet wire or insulated wire. The diameter of the ferrite core used in this example is 0.5" (SPANG/Magnetics F41206-TC). The rectifying diode at the output of the current transformer can be a 1N4148 for secondary currents up to 75mA average.

Sense FETs or resistive sensing can also be used to sense the switch current, the sensed signal has to be amplified to the proper level before it is applied to the ML4812.

The value of the ramp compensation (SC_{PWM}) as seen at the inverting terminal of the PWM comparator is:

$$SC_{PWM} = \frac{2.5 \times R_M}{R_T \times C_T \times R_{SC}} \quad (11)$$

The required value for R_{SC} can therefore be found by equating:

$$SC_{PWM} = A_{SC} \times S_{PWM}$$

where A_{SC} is the amount of slope compensation and solving for R_{SC} .

The value of R_M (pin 2) depends on the selection of R_p (pin 6)

$$R_p = \frac{V_{IN(max)PEAK}}{I_{SINE(peak)}} = \frac{260 \times 1.414}{0.5\text{mA}} = 750\text{K} \quad (12)$$

$$R_M = \frac{V_{CLAMP} \times R_p}{V_{IN(min)PEAK}} = \frac{4.9 \times 750\text{K}}{90 \times 1.414} = 28.8\text{K} \quad (13)$$

The peak of the inductor current can be found approximately by:

$$I_{LPEAK} = \frac{1.414 \times P_{OUT}}{V_{IN(min)RMS}} = \frac{1.414 \times 200}{90} = 3.14\text{A} \quad (14)$$

Selection of N_C which depends on the maximum switch current, assume 4A for this example is 80 turns.

$$R_S = \frac{V_{CLAMP} \times N_C}{I_{LPEAK}} = \frac{4.9 \times 80}{4} = 100\Omega \quad (15)$$

Where R_S is the sense resistor, and V_{CLAMP} is the current clamp at the inverting input of the PWM comparator. This clamp is internally set to 5V. In actual application it is a good idea to assume a value less than 5V to avoid unwanted current limiting action due to component tolerances. In this application V_{CLAMP} was chosen as 4.9V.

Having calculated R_S the value S_{PWM} and of R_{SC} can now be calculated:

$$S_{PWM} = \frac{380\text{V} - 20}{2\text{mH}} \times \frac{100}{80} = 0.225\text{V}/\mu\text{s}$$

$$R_{SC} = \frac{2.5 \times R_M}{A_{SC} \times S_{PWM} \times R_T \times C_T} \quad (16)$$

$$R_{SC} = \frac{2.5 \times 28.8K}{0.7 \times (.225 \times 10^6) \times 14K \times 1nF} = 33K$$

The following values were used in the calculation:

$$R_M = 28.8K \quad A_{SC} = 0.7$$

$$R_T = 14K \quad C_T = 1nF$$

VOLTAGE REGULATION COMPONENTS

The values of the voltage regulation loop components are calculated based on the operating output voltage. Note that voltage safety regulations require the use of sense resistors that have adequate voltage rating. As a rule of thumb if 1/4W resistors are available, two of them should be used in series. The input bias current of the error amplifier is approximately $0.5\mu A$, therefore the current available from the voltage sense resistors should be significantly higher than this value. Since two 1/4W resistors have to be used the total power rating is 1/2W. The operating power is set to be 0.4W then with 380V output voltage the value can be calculated as follows:

$$R_1 = (380V)^2 / 0.4W = 360K \quad (17)$$

Choose two 178K, 1% connected in series.

Then R_2 can be calculated using the formula below:

$$R_2 = \frac{V_{REF} \times R_1}{V_{OUT} - V_{REF}} = \frac{5V \times 356K}{380V - 5V} = 4.747K \quad (18)$$

Choose 4.75K, 1%. One more critical component in the voltage regulation loop is the feedback capacitor for the error amplifier. The voltage loop bandwidth should be set such that it rejects the 120Hz ripple which is present at the output. If this ripple is not adequately attenuated it will cause distortion on the input current waveform. Typical bandwidths range anywhere from a few Hertz to 15Hz. The main compromise is between transient response and distortion. The feedback capacitor can be calculated using the following formula:

$$C_F = \frac{1}{3.142 \times R_1 \times BW} \quad (19)$$

$$C_F = \frac{1}{3.142 \times 356K \times 2Hz} = 0.44\mu F$$

OVERVOLTAGE PROTECTION (OVP) COMPONENTS

The OVP loop should be set so that there is no interaction with the voltage control loop. Typically it should be set to a level where the power components are safe to operate. Ten to fifteen volts above V_{OUT} seems to be adequate. This sets the maximum transient output voltage to about 395V.

By choosing the high voltage side resistor of the OVP circuit the same way as above i.e. $R_4 = 356K$ then R_5 can be calculated as:

$$R_5 = \frac{V_{REF} \times R_4}{V_{OVP} - V_{REF}} = \frac{5V \times 356K}{395V - 5V} = 4.564K \quad (20)$$

Choose 4.53K, 1%.

Note that R_1 , R_2 , R_4 and R_5 should be tight tolerance resistors such as 1% or better.

CONTROLLER SHUTDOWN

The ML4812 provides a shutdown pin which could be used to shutdown the IC. Care should be taken when this pin is used because power supply sequencing problems could arise if another regulator with its own bootstrapping follows the ML4812. In such a case a special circuit should be used to allow for orderly start up. One way to accomplish this is by using the reference voltage of the ML4812 to inhibit the other controller IC or to shut down its bias supply current.

OFF-LINE START-UP AND BIAS SUPPLY GENERATION

The ML4812 can be started using a "bleed resistor" from the high voltage bus. After the voltage on pin 13 (V_{CC}) exceeds 16V, the IC starts up. The energy stored on the $330\mu F$, $C15$, capacitor supplies the IC with running power until the supplemental winding on L1 can provide the power to sustain operation.

The values of the start-up resistor $R10$ and capacitor $C15$ may need to be optimized depending on the application. The charging waveform for the secondary winding of L1 is an inverted chopped sinusoid which reaches its peak when the line voltage is at its minimum. In this example, $C9 = .1\mu F$, $C15 = 330\mu F$, $D8 = 1N4148$ $R10 = 39K$, 2W.

ENHANCEMENT CIRCUIT

The theory of operation of the power factor enhancement circuit (inside the dotted lines) in Figure 11 is described in APPLICATION NOTE 11 in detail. It improves the power factor and lowers the input current harmonics. Note that the circuit meets the proposed IEC 555 specifications (with the enhancement) on the harmonics with a large margin while correcting the input power factor to better than 0.99 under most steady state operating conditions.

CONSTRUCTION AND LAYOUT TIPS

High frequency power circuits require special care during breadboard construction and layout. Double sided printed circuit boards with ground plane on one side are highly recommended. All critical switching leads (power FET, output diode, IC output and ground leads, bypass capacitors) should be kept as small as possible. This is to minimize both the transmission and pick-up of switching noise.

There are two kinds of noise coupling; inductive and capacitive. As the name implies inductive coupling is due to fast changing (high di/dt) circulating switching currents. The main source is the loop formed by $Q1$, $D5$, and $C3-C4$. Therefore this loop should be as small as possible, and the above capacitors should be good high frequency types.

The second form of noise coupling is due to fast changing voltages (high dv/dt). The main source in this case is the drain of the power FET. The radiated noise in this case can be minimized by insulating the drain of the FET from the heatsink and then tying the heatsink to the source of the FET with a high frequency capacitor (C_H in Figure 11).

The IC has two ground pins named PWR GND and Signal GND. These two pins should be connected together with a very short lead at the printed circuit board exit point. In general grounding is very important and ground loops should be avoided. Star grounding schemes are preferred.

Component Values/Bill of Materials for Figure 11

Reference	Description
C1, C4	1 μ F, 630V Film (250 VAC)
C3, C_H	6.8nF 1KV Ceramic disk
C5, C6	680 μ F 200V Electrolytic
C8, C9	.1 μ F 50V Ceramic
C10, C19	1 μ F 50V Ceramic
C11	.001 μ F 50V Ceramic
C15	330 μ F 25V Electrolytic
C16	100 μ F 25V Electrolytic
C17	10 μ F 25V Electrolytic
C_F	.47 μ F 50V Ceramic
C_T	.002 μ F 50V Ceramic
D1, D2, D3, D4, D10	1N5406 (Motorola)
D5	MUR850 (Motorola)
D6, D8, D9, D11, D12, D13	1N4148
F1	5A 250V 3AG with clips
IC1	ML4812CP (Micro Linear)
L1	2mH, 4A I_{PEAK} (see below)
Q1	IRF840 or MTPN8N50

Reference	Description
Q2	LM7815CT
Q3	2N2222 or equivalent
R1A, R1B, R4A, R4B	180K Ω
R2A, R5A	10K Ω TRIMPOT BOURNS 3299 or equivalent
R2B, R5B	3.9K Ω
R3, R13	22K Ω
R6, R7, RPB	150K Ω
R10	39K Ω , 2W
R11	33K Ω
R12	1K Ω
RG	10 Ω
RM	27K Ω
RPA, R15	360K Ω
RS	100 Ω
RSC	33K Ω
RT	75K Ω
T1	SPANG F41206-TC $N_S = 80$, $N_P = 1$ (see attached)

Notes: All resistors 1/4W unless otherwise specified. Some reference designators are skipped (e.g. C2, C12, etc.) and do not appear on the schematic. These designators were used in previous revisions of the board and are not used on this revision. Additional information on key components is included in the attached appendix.

Magnetics Tips (Refer to Figure 11)

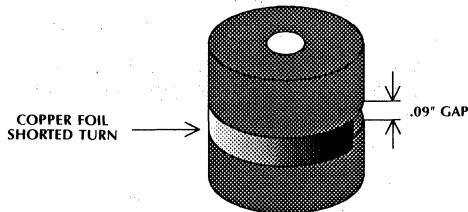
L1 — Main inductor:

One of several toroidal cores can be used for L1:

Material	Manufacturer	Part #	Turns (#24AWG)
Powdered Iron	Micrometals	T225-8/90	200
Powdered Iron	Micrometals	T184-40	120
Molypermalloy	SPANG (Mag. Inc.)	58076-A2 (high flux)	180

The T184-40 core above is the most economical, but has lower inductance at high current. This would yield higher ripple current and require more line EMI filtering. The value for RSC (slope compensation resistor on Pin 7) was calculated for the T225-8/90 and should be recalculated for other inductor characteristics. Selected pages of the Micrometals iron powder core data sheets are attached for your convenience. The core manufacturer also has additional applications literature available.

A gapped ferrite core can also be used in place of the powdered iron core. One such core is a Ferroxcube core #4229PL00-3C8. This is an un-gapped core. Using 145 turns of #24 AWG wire, a total air gap of .180" is required to give a total inductance of about 2mH. Since 1/2 of the gap will be on the outside of the core and 1/2 the gap on the inside, putting a .09" spacer in the center will yield a .180" total gap. To prevent leakage fields from generating RFI, a shorted turn of copper tape should be wrapped around the gap as shown below:



For production, a gapped center leg can be ordered from most core vendors, eliminating the need for the external shorted copper turn when using a pot core.

T1 — Sense Transformer

In addition to the core type mentioned in the parts list, the following Siemens cores should be suitable for substitution and may be more readily available in Europe.

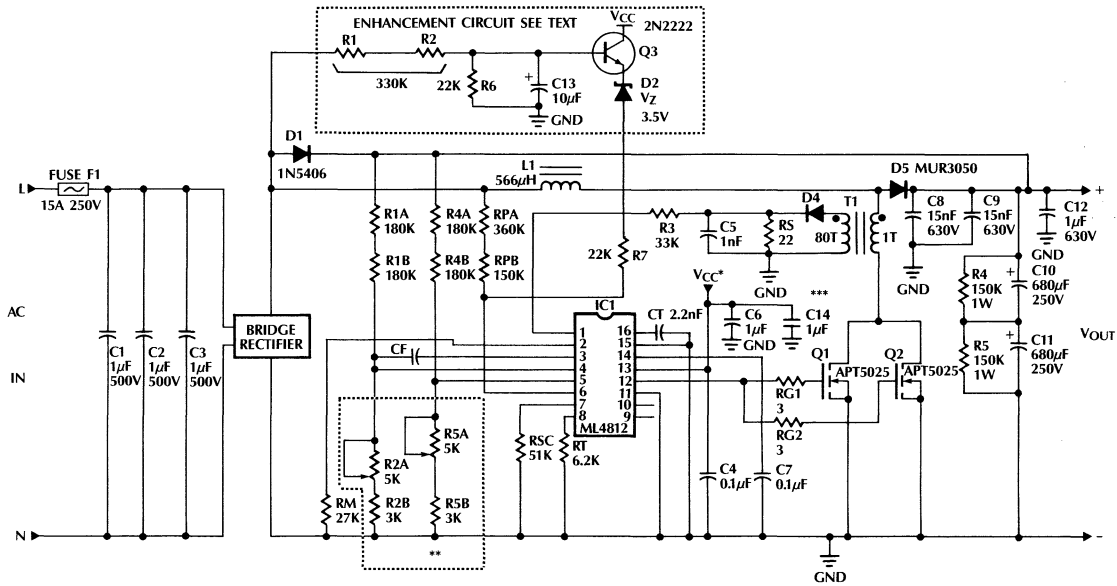
Material	Size Code	Part #
N27	R16/6.3	B64290-K45-X27
N30	R16/6.3	B64290-K45-X830

The N27 material is for high frequency and will work better above 100kHz but both are adequate. In addition, Ferroxcube/Phillips Magnetics core 768T188-3C8 can be used.

U.S. Core Vendors:

Manufacturer	Phone Number
SPANG/Magnetics Inc.	(412) 282-8282
Micrometals	(800) 356-5977
Ferroxcube/Phillips Magnetics	(818) 998-7311

Figure 12. 1KW Input Power, Power Factor Correction Circuit



NOTES:

1. ALL UNSPECIFIED DIODES ARE 1N4148.
2. ALL UNSPECIFIED RESISTORS ARE 1/4 WATT
3. ALL UNSPECIFIED CAPACITOR VOLTAGE RATINGS ARE 50V.
4. ADJUST R_{2A} AND R_{5A} WITH CAUTION TO AVOID OVER VOLTAGE CONDITIONS.

Q₃ = 2N2222 OR EQUIVALENT

* AT INITIAL TURN-ON TO CHECK THE IC FOR PROPER OPERATION, APPLY ≈ 16VDC.

** FIXED RESISTORS CAN BE USED FOR THE SENSING COMPONENTS. BELOW ARE 1% STANDARD RESISTORS THAT WILL FORCE THE CORRECT OUTPUT VOLTAGES R_{1A}, R_{1B}, R_{4A}, R_{4B} = 178K 1% R_{2B} = 4.75K 1%, R_{5B} = 4.53K 1%. USE JUMPERS INSTEAD OF R_{2A} AND R_{5A} (POTS).

*** FOR HIGHER POWER USE MORE V_{CC} DECOUPLING.

ML4812

ORDERING INFORMATION

PART NUMBER	TEMPERATURE RANGE	PACKAGE
ML4812CP	0°C to +70°C	MOLDED DIP (P16)
ML4812CQ	0°C to +70°C	MOLDED PCC (Q20)
ML4812IP	-40°C to +85°C	MOLDED DIP (P16)
ML4812IQ	-40°C to +85°C	MOLDED PCC (Q20)
ML4812MJ	-55°C to +125°C	HERMETIC DIP (J16)

ML4812EVAL

Power Factor Controller Evaluation Kit

GENERAL DESCRIPTION

The ML4812EVAL kit provides a convenient vehicle to evaluate the ML4812 Power Factor Correction circuit. The board implements a 200W "boost" type power factor correction system. Special care has been taken in the layout of this PC board to provide adequate space for probes and a large area for ground plane to increase system noise immunity.

This kit includes a blank PC board, schematic of a complete power factor correction system and specifications for the key external components necessary to build a prototype Power Factor Correction front end. The unit is designed to operate over 90VAC to 256VAC line range and can run from no load to a full 200W. Higher power levels can be achieved using this board by using larger external components.

This boost mode converter is set to run with a 380V output and achieves power factors of better than .99 over a wide range of input line and output load.

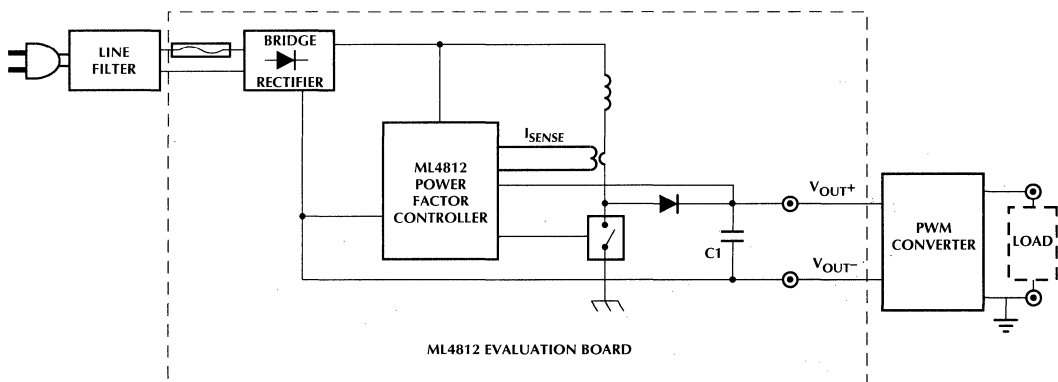
FEATURES

- Power Factor > .99
- Harmonic currents well below proposed IEC555-2 limits.
- 90 to 256VAC input, 380V output to 200W
- 380VDC output to 200W
- Over-Voltage Protection
- Peak Current sense circuit protects Power MOSFET
- PC board and ML4812CP controller included
- Line and Load regulation better than 2%
- Complete documentation and applications information

KIT COMPONENTS

- User's Guide
- ML4812 Datasheet
- ML4812 Sample
- Evaluation Board
- Powder Iron Toroidal Core

BLOCK DIAGRAM



Flyback Power Factor Controller

GENERAL DESCRIPTION

The ML4813 is designed to optimally facilitate a discontinuous "flyback" or "buck-boost" type power factor correction system for low power, low cost applications. Special care has been taken in the design of the ML4813 to reject system noise. The circuit includes a precision reference, oscillator, error amplifier, over-voltage comparator, over-current comparator, and an extra op-amp as well as a high current output. In addition, start-up is simplified by an under-voltage lockout circuit with 6V hysteresis.

In a typical application, the ML4813 functions as a voltage mode regulator. By maintaining a constant duty cycle, the current follows the input voltage, making the impedance of the entire circuit appear purely resistive. With the flyback circuit, power factors of .99 are easily achievable with a small output inductor and a minimum of external components.

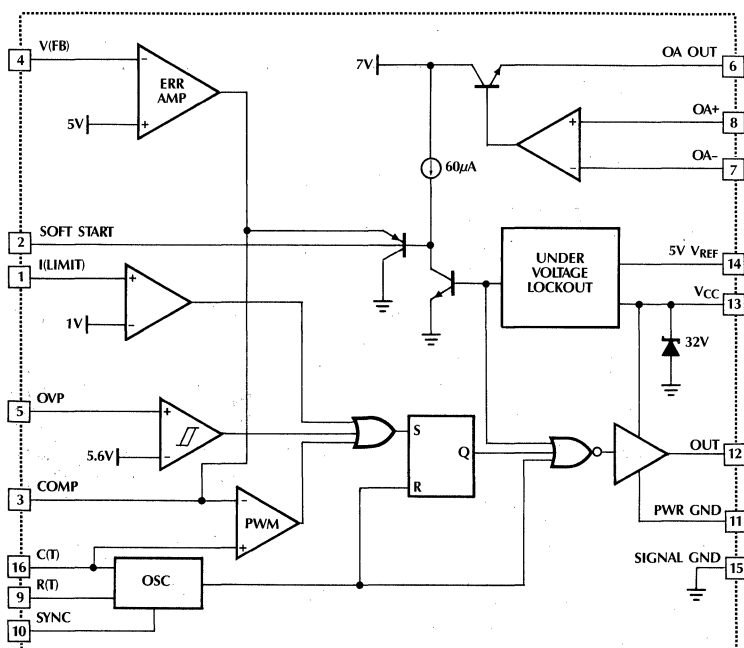
FEATURES

- Precision buffered 5V reference ($\pm 1\%$)
- Extra op-amp for output voltage instrumentation amplifier
- Over Current comparator for switch protection
- Soft Start and 6V hysteresis under-voltage lockout for easy low surge off-line starting
- 1 A Peak Current Totem-Pole Output Drive
- Over-Voltage comparator eliminates output "runaway" due to load removal
- Large oscillator amplitude for better noise immunity

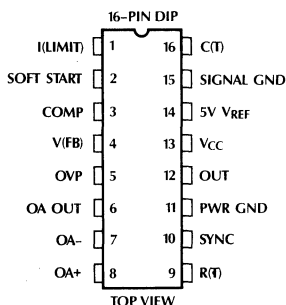
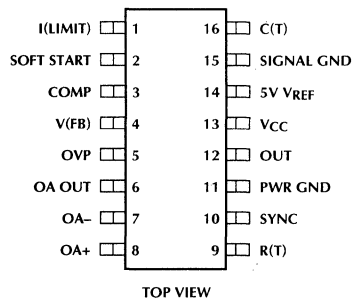
APPLICATIONS

- PC power supplies
- Lamp Ballasts

BLOCK DIAGRAM (Pin out shown is for DIP)



PIN CONFIGURATION

ML4813
16-Pin DIPML4813
16-Pin SOIC

PIN DESCRIPTION (DIP)

PIN #	NAME	FUNCTION	PIN #	NAME	FUNCTION
1	I(LIMIT)	Current limit sense pin. Normally connected to sense resistor. When this pin exceeds 1V, the PWM cycle is terminated.	9	R(T)	Oscillator timing resistor pin. A 5V source sets a current in the external resistor which is mirrored to charge C(T).
2	SOFT START	Normally connected to a Soft Start capacitor.	10	SYNC	Input used to synchronize the oscillator to an external source.
3	COMP	Output of error amplifier and input to PWM comparator.	11	PWR GND	Return for the High Current Totem pole output.
4	V(FB)	Control loop feedback voltage.	12	OUT	High Current Totem pole output.
5	OVP	Input to over voltage comparator.	13	V _{CC}	Positive Supply for the IC.
6	OA OUT	Output of uncommitted op-amp.	14	5V V _{REF}	Buffered output for the 5V voltage reference.
7	OA-	Negative input of uncommitted op-amp.	15	SIGNAL GND	Analog signal ground.
8	OA+	Positive input of uncommitted op-amp.	16	C(T)	Timing Capacitor for the Oscillator.

ML4813

ABSOLUTE MAXIMUM RATINGS

Absolute maximum ratings are those values beyond which the device could be permanently damaged. Absolute maximum ratings are stress ratings only and functional device operation is not implied.

Supply Current (I_{CC})	40mA
Output Current, Source or Sink (Pin 12)	
DC	1.0A
Output Energy (capacitive load per cycle)	5 μ J
Error Amp Sink Current (pin 3)	10mA
Oscillator Charge Current	5mA
Analog Inputs (pins 1, 3-8)	-0.3V to 5.5V

Junction Temperature	150°C
Storage Temperature Range	-65°C to +150°C
Lead Temperature (Soldering 10 sec.)	+260°C
Thermal Resistance (θ_{JA})	
Plastic DIP or SOIC	65°C/W

OPERATING CONDITIONS

Temperature Range	
ML4813C	0°C to 70°C

ELECTRICAL CHARACTERISTICS

Unless otherwise specified, $R_T = 14K\Omega$, $C_T = 1000pF$, $T_A =$ Operating Temperature Range, $V_{CC} = 15V$ (Note 2)

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
Oscillator					
Initial Accuracy	$T_J = 25^\circ C$	90	97	104	KHz
Voltage Stability	$12V < V_{CC} < 18V$		0.3		%
Temperature Stability			2		%
Total Variation	line, temp.	88		108	KHz
Ramp Valley			1.0		V
Ramp Peak			4.3		V
R(T) Voltage		4.8	5.0	5.2	V
Discharge Current (pin 8 open)	$T_J = 25^\circ C, V_{PIN 16} = 2V$	7.5	8.4	9.3	mA
	$V_{PIN 16} = 2V$	7.2	8.4	9.5	mA
Sync Pulse Threshold		.8	1.4	2.0	V
Sync Input Bias Current			350	800	μA
Reference Section					
Output Voltage	$T_J = 25^\circ C, I_O = 1mA$	4.95	5.00	5.05	V
Line Regulation	$12V < V_{CC} < 25V$		6	20	mV
Load Regulation	$1mA < I_O < 20mA$		3	20	mV
Temperature Stability			.4		%
Total Variation	line, load, temp.	4.9		5.1	V
Output Noise Voltage	10Hz to 10KHz		50		μV
Long Term Stability	$T_J = 125^\circ C, 1000$ Hrs, (note 1)		5	25	mV
Short Circuit Current	$V_{REF} = 0V$	-30	-85	-180	mA
Error Amplifier Section					
Input Offset Voltage		-15		15	mV
Input Bias Current			-0.1	-1.0	μA
Open Loop Gain	$1 < V_{PIN 3} < 5V$	60	75		dB
PSRR	$12V < V_{CC} < 25V$	60	70		dB
Output Sink Current	$V_{PIN 3} = 1.1V, V_{PIN 4} = 6.2V$	2	12		mA
Output Source Current	$V_{PIN 3} = 5.0V, V_{PIN 4} = 4.8V$	-0.5	-1.0		mA
Output High Voltage	$I_{PIN 3} = -0.5mA, V_{PIN 4} = 4.8V$	5.3	6.4		V
Output Low Voltage	$I_{PIN 3} = 2mA, V_{PIN 4} = 6.2V$		0.5	1.0	V
Unity Gain Bandwidth			1.0		MHz

ELECTRICAL CHARACTERISTICS (Continued)Unless otherwise specified, $R_T = 14k\Omega$, $C_T = 1000pF$, $T_A =$ Operating Temperature Range, $V_{CC} = 15V$ (note 2)

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
Un-Committed Op Amp					
Input Offset Voltage		-10		10	mV
Input Bias Current			-0.1	-2.0	μA
Input Offset Current		-0.35		0.35	μA
Open Loop Gain			90		dB
PSRR		80	125		dB
Output High Voltage	$I_{PIN\ 3} = -10mA$	6.5	8		V
Output Low Voltage	$R_{L(PIN\ 6)} = 10K\Omega$.2	.5	V
I(LIMIT) Comparator					
Input Trip Point	Output Off	.8	1.0	1.2	V
Input Bias Current			-2	-15	μA
Propagation Delay			150		nS
OVP Comparator					
Input Trip Point	Output Off	5.5	5.6	5.7	V
Hysteresis	Output On		100		mV
Input Bias Current			-0.3	-3	μA
PWM Comparator					
Input Common Mode Range		-0.2		5.5	V
Input Bias Current			-2	-10	μA
Propagation Delay			150		nS
Soft Start Section					
Soft Start Current (pin 2)	$V_{PIN\ 2} = 1V$	40	60	80	μA
Output Section					
Output Voltage Low	$I_{OUT} = 10mA$		0.1	0.4	V
	$I_{OUT} = 200mA$		1.2	2.2	V
Output Voltage High	$I_{OUT} = -20mA$	13	13.6		V
	$I_{OUT} = -200mA$	12	13.4		V
Output Voltage Low in UVLO	$I_{OUT} = 5mA, V_{CC} = 8V$		0.1	0.8	V
Output Rise/Fall Time	$C_L = 1000pF$		50		nS
Under-Voltage Lockout					
Start-up Threshold		15	16	17	V
Shut-Down Threshold		9	10	11	V
V_{REF} Good Threshold			4.4		V
Total Device					
Supply Current	Start-up, $V_{CC} = 14V$.9	1.5	mA
	Operating		20	30	mA
Internal Shunt Zener Voltage	$I_{CC} = 30mA$	25	30	34	V

Note 1: This parameter not 100% tested in production but guaranteed by design.

Note 2: V_{CC} is raised above the Start-up Threshold first to activate the IC, then returned to 15V.

FUNCTIONAL DESCRIPTION

OSCILLATOR

The ML4813 oscillator charges the external capacitor (C_T) with a current (I_{SET}) equal to $5/R_{SET}$. When the capacitor voltage reaches the upper threshold, the comparator changes state and the capacitor discharges to the lower threshold through Q1. While the capacitor is discharging, Q2 provides a high pulse.

The Oscillator period can be described by the following relationship:

$$T_{OSC} = T_{RAMP} + T_{DEADTIME}$$

where:

$$T_{RAMP} = C \text{ (Ramp Valley to Peak)} \div I_{SET}$$

and:

$$T_{DEADTIME} = C \text{ (Ramp Valley to Peak)} \div (8.4\text{mA} - I_{SET})$$

A pulse of a duration shorter than $T_{DEADTIME}$ from an external frequency source set to a higher frequency than f_{OSC} can be applied to pin 10 to synchronize the oscillator. R(SYNC) and C(SYNC) shorten longer pulses.

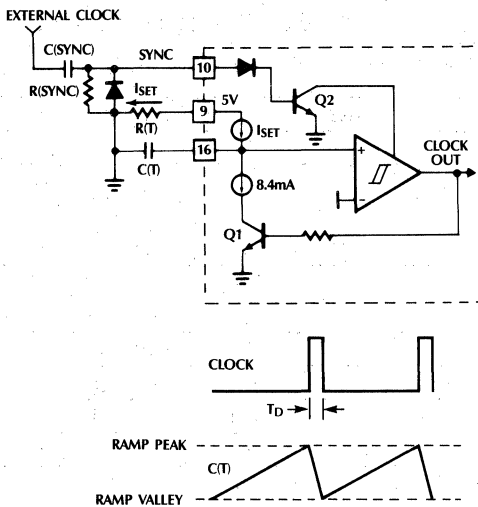


Figure 1. Oscillator Block Diagram

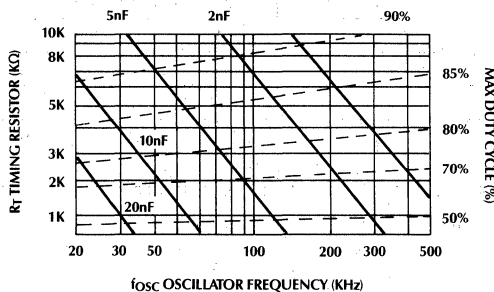


Figure 2. Oscillator Timing Resistance vs. Frequency

OUTPUT DRIVER STAGE

The ML4813 Output Driver is a 1A peak output high speed totem pole circuit designed to quickly drive capacitive loads, such as power MOSFET gates.

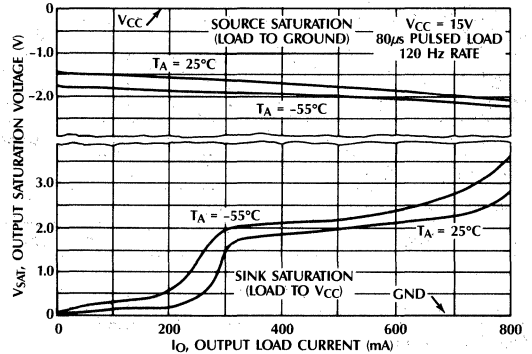


Figure 3. Output Saturation Voltage vs. Output Current

ERROR AMPLIFIER

The ML4813 error amplifier is a high open loop gain, wide bandwidth, amplifier.

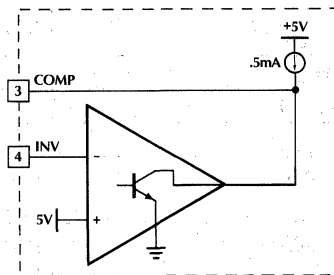


Figure 4. Error Amplifier Configuration

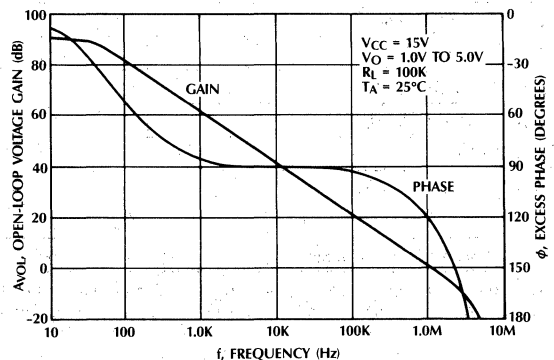


Figure 5. Error Amplifier Open-Loop Gain and Phase vs. Frequency

UN-COMMITTED OP-AMP

The ML4813 contains an un-committed op-amp which is normally configured as a differencing amplifier to sense the output voltage. The output voltage in the flyback configuration is not ground referenced. The op-amp in the ML4813 is a PNP input amplifier similar to the LM324 but with an open emitter output stage (class A).

UNDER VOLTAGE LOCKOUT

On power-up the ML4813 remains in the UVLO condition; output low and quiescent current low. The IC becomes operational when V_{CC} reaches 16V. When V_{CC} drops below 10V, the UVLO condition is imposed. During the UVLO condition, the 5V V_{REF} pin is "off", making it usable as a "flag" for starting up a downstream PWM converter.

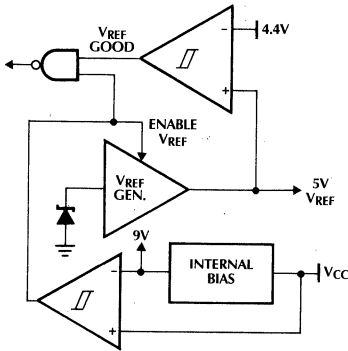


Figure 6. Under-Voltage Lockout Block Diagram

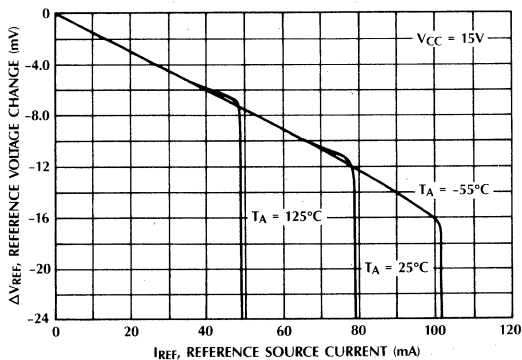


Figure 7. Reference Load Regulation

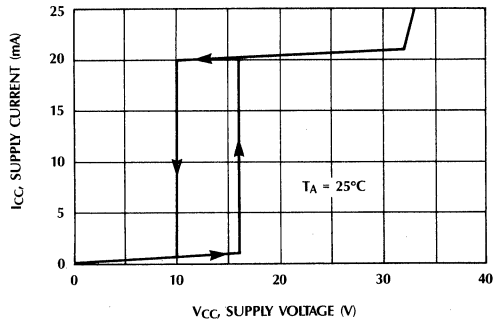


Figure 8. Total Supply Current vs. Supply Voltage

APPLICATIONS

The ML4813 is used to implement a discontinuous mode flyback (buck-boost) power factor regulator. This topology is particularly well suited for low power applications such as: fluorescent ballasts; and low power switching supplies. Also it is a useful topology when there is a requirement for the output voltage to be lower than the peak input voltage, or where an isolated output is required. This is not possible with the boost topology, where the output voltage must always be higher than the maximum peak of the input voltage range. The typical input range for the flyback power factor regulator is from 90 VAC to 260 VAC.

The regulator operates in the discontinuous inductor current conduction mode. The inductor energy stored during the "ON" time of the power switch Q is completely delivered to the output capacitance during the "OFF" time. At steady state conditions, the inductor current at the beginning of the "ON" time starts to ramp-up from 0 Amps to a value that is determined by the instantaneous value of the input full wave rectified voltage; the "ON" time as it is set by the error amplifier and the PWM comparator; and finally by the inductor itself (L).

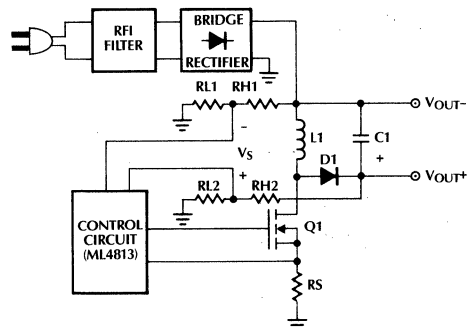


Figure 9. Block Diagram of the Regulator

The expression for the inductor peak current is given by:

$$I_L(\theta) = \frac{V_{IN}(\theta)t_{ON}}{L} \quad (1)$$

Where:

- $I_L(\theta)$ = The instantaneous peak inductor current.
- t_{ON} = Power switch "on" time.
- $V_{IN}(\theta) = V_P \sin \theta$ = Instantaneous Input Voltage.
- V_P = Input Peak Voltage.

Figure 10, is a diagram of the relationship between the low frequency envelope and the high frequency inductor current. Note that for clarity the scale between the two waveforms has not been preserved. Normally for 60Hz input line and 100KHz switching frequency, each half of the sine wave contains approximately 833 high frequency triangular waveforms.

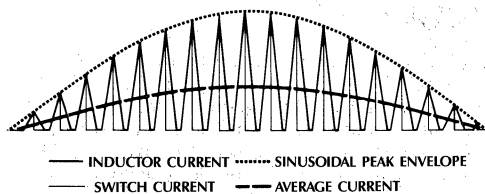


Figure 10. Switch and Line Currents in the Flyback PFC Circuit

The envelope of the peaks of the switch current, which in this case represent the current drawn from the input source, have a sinewave shape. This relationship is shown as:

$$I_L(\theta) = I_P \sin \theta \quad (2)$$

Combining (1) and (2) the following useful relationship is obtained:

$$t_{ON} = \frac{L I_P}{\sqrt{2} V_{RMS}} \quad (3)$$

Note that $V_{IN}(\theta) = V_P \sin \theta$, and also $V_P = \sqrt{2} V_{RMS}$. The average value of the input triangular current is:

$$I_{AVG}(\theta) = \frac{t_{ON}}{2T} I_P \sin \theta \quad (4)$$

Where:

- I_{AVG} = Average value of the switch current. This is the value of the current at the input of the regulator after filtering.
- t_{ON} = Switch "on" time.
- T = Period of the switch cycle.

Substitution of (3) into (4) yields.

$$I_{AVG}(\theta) = \frac{L I_P^2}{2.828 T V_{RMS}} \sin \theta \quad (5)$$

Equation (5) clearly shows that the average value of the switch current is sinusoidal and in phase with the input voltage. The peak value of the average current is:

$$I_{AVG(PEAK)} = \frac{L I_P^2}{2.828 T V_{RMS}} \sin \theta \quad (6)$$

Also:

$$I_{AVG(PEAK)} = \frac{\sqrt{2} P_{IN}}{V_{RMS}} \quad (7)$$

Solving equations (7) and (6) for P_{IN} :

$$P_{IN} = \frac{1}{4} L I_P^2 f \quad (8)$$

For optimum performance and the lowest inductor peak currents, the inductor current should be at the verge of continuity at the lowest operating voltage point and at full load. The above can be satisfied if:

$$I_P \leq \frac{V_{IN} V_{OUT}}{f L (V_{IN} + V_{OUT})} \quad (9)$$

Where: $V_{IN} = \sqrt{2} \times V_{IN \text{ MIN}} \text{ (RMS)}$

Finally (8) and (9) can be combined to derive an upper bound for the inductor value that will guarantee that the regulator always stays in the discontinuous mode of operation. If the regulator were to operate in the continuous mode the average input current would not be sinusoidal.

$$L \leq \left[\frac{V_{IN} V_{OUT}}{2\sqrt{f} P_{IN} (V_{IN} + V_{OUT})} \right]^2 \quad (10)$$

FLYBACK INDUCTOR CALCULATION

Equation (10) gives the upper bound for the inductor value for any set of specified operating conditions. Normally a few iterations may be required, for finalizing the value. The reason for this is that equation (10) does not contain parameters to correct for second or third order effects. All this means that a good initial value for the inductor is probably 10 to 20% lower than the value calculated by the right hand side expression in (10).

Several core materials are candidates for the inductor, such as: powder iron cores, gapped ferrites, moly permalloy cores, etc. In the application that will be described later, a gapped ferrite core is used.

There are no particular restrictions on the inductor except that the inductance is of correct value and the losses are acceptable.

INPUT BYPASS CAPACITANCE

The triangular high frequency current is bypassed by the input capacitor (C_i) labeled C_7 in Figure 12. This is a high quality film capacitor with low ESR value for minimum losses and heating. A polyester, polypropylene or x-type (for line side) is a good candidate. Typical values, depending on the power level, can range anywhere from 0.33 μ F to 1.5 μ F. The

next filtering stage of the RFI filter which has an inductor as input isolates C_7 from the other capacitors which may be present at the input circuit. Note that C_1 (C_7) can be on either side of the bridge rectifier. The preferred location for low crossover distortion is at the input side. The voltage ripple across this capacitor is:

$$V_{C(P-P)} = \frac{D}{C_1 f} \sqrt{\frac{P_{IN}}{L_f}} - \frac{\sqrt{2} P_{IN}}{C_1 f V_{IN}} \quad (11)$$

Where:

$V_{C(P-P)}$ = Peak to peak worst case high frequency capacitor voltage.
 D = Switch Duty Cycle.

Therefore the RFI filter that follows has to be able to attenuate $V_{C(P-P)}$ to the levels set by the relevant regulatory specifications.

INPUT TRANSIENT OVERVOLTAGE PROTECTION

Careful examination of the power circuits reveals that there is no large capacitance at the input of the regulator. The only capacitances present are the RFI filter capacitors. These capacitors have a combined value in the range of a couple of microfarads. Thus their ability to absorb and minimize any line induced transients is almost non-existent. Transients can occur also under sudden load removal. If the line impedance is inductive, hazardous drain source voltages may be generated leading to the destruction of the power switch. To keep this from happening a transient over-voltage protection device should be chosen such that enough safety margin is allowed for the power switch. A good rule of thumb is:

$$B_{VDSS} > V_{ZA} + V_{OUT(OVP)} \quad (12)$$

Where:

B_{VDSS} = Drain-Source breakdown voltage for the FET.
 V_{ZA} = Activation or clamping voltage of the over-voltage transient protector.
 $V_{OUT(OVP)}$ = Maximum output voltage. This is set by the OVP function of the controller, and will be covered later.

THE OUTPUT CIRCUIT

The output circuit for this topology, although it is non-isolated, does not share the same ground with the power circuit. Therefore connecting the two grounds with the measuring leads of instruments should be avoided. This is a common mistake especially with the oscilloscope leads.

The output voltage "rides" on the input voltage when the (+) output is measured with respect to PGND (figure 11).

The extra OP-AMP provided in the ML4813 is used to sense the output voltage for regulation and over voltage conditions. This op-amp is connected as a difference amplifier with its output referenced to PGND. Resistors R_{H1} , R_{H2} , R_{L1} , R_{L2} are used to scale down the voltage.

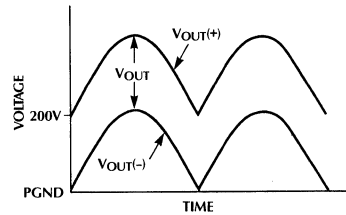


Figure 11. Output Voltage with Respect to PGND

Normally $R_{H1} = R_{H2} = R_H$ and $R_{L1} = R_{L2} = R_L$. Then the voltage designated as V_S in Figure 9 is given by:

$$V_S = V_{OUT} \frac{R_L}{R_H + R_L} \quad (13)$$

OUTPUT CAPACITANCE

The output capacitance should be calculated such that it has the required output ripple at the worst case operating point. In addition the ESR should be sufficiently low to prevent dissipation due to RMS currents. The first criterion can be met by choosing the value of the output capacitor based on the following:

$$C \geq \frac{P_{IN}}{2\pi f_L \Delta V_R V_{OUT}} \quad (14)$$

Where:

C = Total output capacitance.
 P_{IN} = Total input power.
 ΔV_R = Peak output capacitor ripple voltage.
 V_{OUT} = Output Voltage.
 f_L = Line Frequency times 2 (120 for 60Hz line).

The second criterion for the selection of the output capacitor can be satisfied by choosing a component with adequately low ESR value, that can safely bypass the RMS currents.

OUTPUT DIODE

The output diode can be a "fast" or ultrafast" type depending on the operating frequency. Reverse recovery losses are low since at steady state and under normal operating conditions the regulator operates in discontinuous current mode. The diode should be rated to handle the output current. The resulting power dissipation will be the forward drop of the diode times the output current.

POWER SWITCH

If a power FET is used, it should be sized for the required efficiency. Lower $R_{DS(ON)}$ devices will yield lower losses, but if they are operated at high frequencies (100KHz) higher charge dumping losses ($1/2 C_{DS} V_{DS}^2 f$) will be experienced. The RMS current value through the power FET and the sensing resistor is:

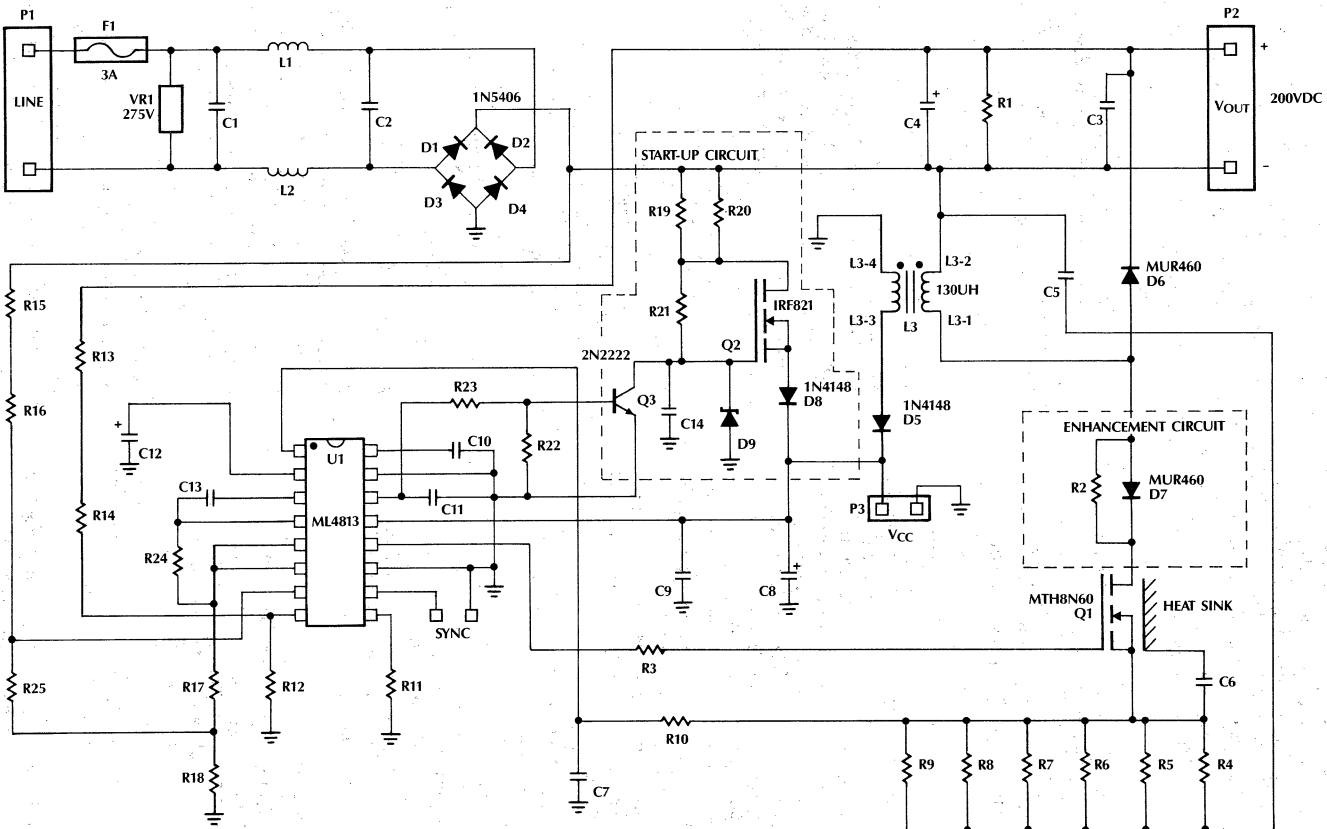


Figure 12. ML4813 Typical Application: 80W Flyback Power Factor Regulator

$$I_{RMS} = \sqrt{\frac{L I_p^3 f_L}{3\sqrt{2} V_{RMS}}} \sqrt{\sum_{k=1}^r \sin^2 \frac{k\pi}{r}} \quad (15)$$

Where:

- I_{RMS} = Total RMS current through the power FET and sense resistor.
- f_L = Line Frequency times 2 (120 for 60Hz line).
- $r = f_{SWITCH}/f_L$.

Table 1 is provided to assist in calculating (15) above. When the power switch is a bipolar transistor (constant V_{CE} drop) then the power dissipation produced can be calculated by using (16):

$$P_D = \frac{0.9 P_{IN}}{V_{RMS}} V_{CE} \quad (16)$$

Where:

- P_D = Power dissipation by the transistor (conduction losses).
- V_{RMS} = RMS value of the minimum input voltage.
- V_{CE} = Collector Emitter forward drop of the power transistor.

OFF-LINE START-UP AND BIAS SUPPLY GENERATION

A fast starting circuit is shown in figure 12. MOSFET Q2 quickly charges the IC's V_{CC} capacitor (C8) when the supply is initially turned on. This allows the supply to come on less than 1 second after AC power is applied. A simpler start-up circuit may be used which replaces the active circuit with a 39KΩ 2W resistor but starts more slowly (up to 15 seconds under low line conditions). Systems which do not require quick starting can reduce cost with the latter start-up method.

f_{SWITCH} (KHz)	r	$\sqrt{\sum_{k=1}^r \sin^2 \frac{k\pi}{r}}$
20	167	9.1
30	250	11.2
40	333	12.9
50	417	14.4
60	500	15.8
70	583	17.1
80	667	18.3
90	750	19.4
100	833	20.4
110	917	21.4
120	1000	22.4
130	1083	23.3
140	1167	24.2
150	1250	25.0
160	1333	25.7
170	1417	26.5
180	1500	27.3
190	1583	28.0
200	1667	28.9

Table 1. Figures for Calculating I_{RMS} (eq. 15)

POWER FACTOR ENHANCEMENT

Some combinations of line and load may exhibit distortion of the input current waveform. This distortion is usually caused by the inductor "ringing" with the C_{DS} of the power MOSFET, resulting in a non-zero inductor current at the beginning of the next cycle. This ringing can be dampened by using R2 and D7 in figure 12. Applications which can get by with slightly worse power factor can eliminate these components.

ADJUSTING THE OUTPUT VOLTAGE

The error amplifier creates an error voltage from the difference between the output voltage presented on pin 6 and the 5V internal reference. Since the output voltage is not ground referenced, the ML4813's internal op-amp is connected as an instrumentation amplifier (figure 13).

The output voltage is set by a combination of resistors which determine the relationship between ($V_{OUT+} - V_{OUT-}$) and the output of the op-amp (pin 6). For the following discussion, $R15' = R15 + R16$ and $R14' = R14 + R13$. The differencing amplifier operation depends on the following relationships:

$$R15' = R14' \text{ and } R12 = R25 + (R18 \parallel R17)$$

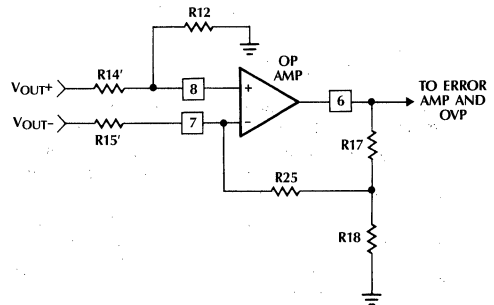


Figure 13. Ground Referencing the Output Voltage

Then:

$$V_{OUT} = \left(\frac{5V \times R18}{R17 + R18} \right) \left(\frac{R15'}{R15' + R25} \right) \left(\frac{R14'}{R12} + 1 \right)$$

Since R25 is a low value compared to R15', the second term reduces to approximately 1. The third term is set at approximately 200. Therefore the above equation reduces to:

$$V_{OUT} \cong 1000 \times \left(\frac{R18}{R17 + R18} \right)$$

The over voltage comparator has a threshold that is set for $1.12 \times V_{OUT}$ when pin 5 and pin 6 are connected directly. Figure 14 shows the connection for setting an OVP trip point higher than $1.12 \times V_{OUT}$, where:

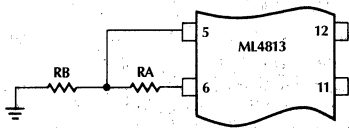


Figure 14. Setting OVP for a $V_{OVP} > 1.12 \times V_{OUT}$

$$V_{OVP} \cong 1.12 \times V_{OUT} \times \left(\frac{RA + RB}{RB} \right)$$

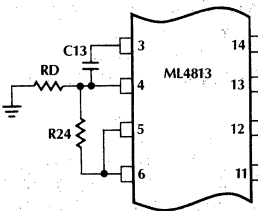


Figure 15. Setting OVP for a $V_{OVP} < 1.12 \times V_{OUT}$

Figure 15 shows OVP set for a voltage lower than $1.12 \times V_{OUT}$ where:

$$V_{OVP} \cong 1.12 \times V_{OUT} \times \left(\frac{RD + R24}{RD} \right)$$

INDUCTOR INFORMATION

L3 is the flyback inductor and also provides the operating power for the control circuitry. A gapped ferrite pot core was chosen for this application for its modest high frequency losses with high ripple current operation. Some possible choices are:

Manufacturer	Part #	Total Gap	N _p
Magnetics Inc.	F43019	.05"	32
Ferroxcube (Phillips)	3019 PL00-3F3	.05"	32
Ferroxcube (Phillips)	3019 PA125-3C8	.07"	38

The first 2 cores are sold ungapped and require the use of a .025" spacer to gap the center leg to yield a total gap length of .05". If an ungapped core is used, a "shorted turn" should be employed as shown below (figure 16) to prevent radiated EMI. The third core listed is sold with its center leg pre-gapped (.07" total), hence the outside of the core closes completely providing shielding without the shorted turn being required. N_s should be 3 turns. All windings #24AWG wire.

L1 and L2 inductors are constructed using a powdered iron. This is a suitable material for these inductors since the high frequency ripple currents (and resulting flux excursions) are much less severe than for L3. The core selected is:

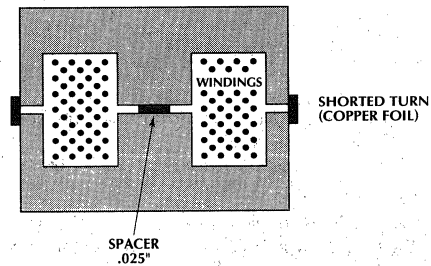


Figure 16. Construction of EMI Shield for Ungapped Cores

Manufacturer	Part #	Turns
MicroMetals	T68-26D	80T #24AWG

COMPONENT	DESCRIPTION
C1, C2	0.68μF, 630V
C3, C5, C6	.01μF, 1KV
C4	330μF, 250V
C7	1000pF, 50V
C8	1000μF, 16V
C9	1μF, 50V
C10	6800pF, 50V
C11, C14	0.1μF, 50V
C12	4.7μF, 50V
C13	0.22μF, 50V
D1 thru D4	1N5406
D5, D8	1N4148
D6, D7	MUR460
D9	22V Zener, 1/4 W
F1	3AC, 3A, 250V
Heat Sink	Thermalloy 6398-U-P3
L1, L2	500μH, 1.5A RMS
L3	160μH, 5A peak
Q1	MTH8N60
Q2	IRF821
Q3	2N2222
R1	220KΩ
R2, R19, R20	4.3KΩ
R3	10Ω
R4 thru R9	1Ω
R10	100Ω
R11	1.8KΩ
R12	4.02KΩ, 1%
R13 thru R16	402KΩ, 1%
R17	806Ω, 1%
R18	200Ω
R21	510KΩ
R22, R23	2KΩ
R24	100KΩ
R25	3.83KΩ, 1%
U1	ML4813CP
VR1	TNR12G431KM

Note: All resistor values 1/4 W ± 5% unless otherwise specified. All capacitor values ±10% unless otherwise specified.

Table 2. Component Values for Figure 8

ORDERING INFORMATION

PART NUMBER	TEMPERATURE RANGE	PACKAGE
ML4813CP ML4813CS	0°C to +70°C 0°C to +70°C	Molded DIP (P16) Molded SOIC (S16W)

Zero Voltage Switching Resonant Controller

GENERAL DESCRIPTION

The ML4815 is designed to facilitate zero-voltage switched (ZVS) resonant converters requiring constant off-time and variable on-time control. Since the power MOSFET is turned on at zero voltage in ZVS resonant converters, power dissipation due to charge-dumping of the MOSFET drain-source capacitance is eliminated, allowing high frequency operation and power density to be maximized. MOSFET parasitic drain-source capacitance can also be used as part of the resonant circuit, minimizing component count.

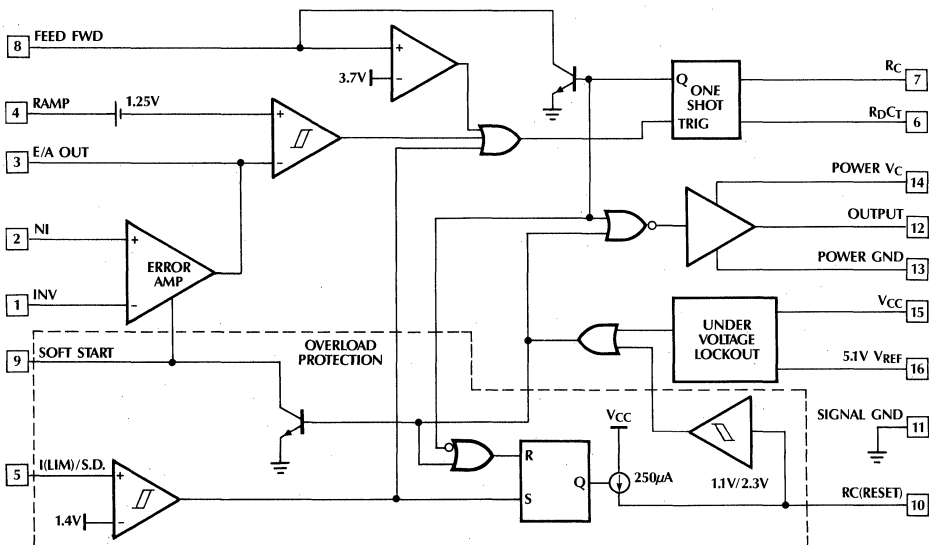
The ML4815 features a monostable multivibrator for precise off-time setting. The on-time is modulated through a ramp comparator in a manner similar to PWM converters. Either current-mode control with maximum on-time clamp or voltage-mode control with input feedforward can be selected.

ML4815 supports pulse-by-pulse (peak) current limiting as well as "hiccup" mode for fault protection. The controller is designed for operation up to 2MHz. ML4815 also includes a wide band error amplifier and a high peak current output driver which minimizes cross-conduction current.

FEATURES

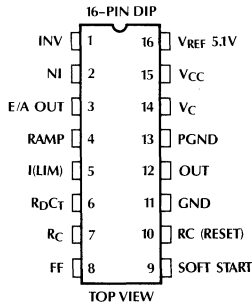
- Supports Single-Switch ZVS Resonant Topology with Minimal External Components
- Ideal for Simple, High Density DC to DC Converters
- Small Converter Frequency Variation from No-Load to Full-Load
- High Current (2A Peak) Totem-Pole Output Drive with Low Cross Conduction
- Precision Buffered 5.1V Reference ($\pm 2\%$)
- Wideband (5.5MHz), High Slew Rate ($12V/\mu S$) Error Amp.
- Under-Voltage Lockout with Low Current Start-Up
- Integrating Fault Detection/Soft-Start Reset

BLOCK DIAGRAM (Pin out shown is for 16-pin DIP)

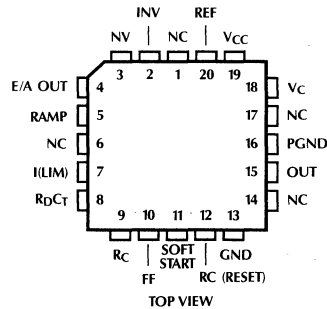


PIN CONNECTIONS

ML4815
16-Pin DIP



ML4815
20-Pin PCC



PIN DESCRIPTION

PIN #	NAME	FUNCTION	PIN #	NAME	FUNCTION
1	INV	Inverting input to error amp.	9	SOFT START	Normally connected to Soft Start Capacitor and charging resistor.
2	NI	Non-inverting input to error amp.	10	RC (RESET)	Timing Capacitor for over-current integration and restart-delay.
3	E/A OUT	Output of error amplifier and input to main comparator.	11	GND	Analog Signal Ground.
4	RAMP	Non-inverting input to main comparator. Connected to pin 8 for feedforward voltage-mode control or to pin 5 for current-mode control.	12	OUT	High Current Totem pole output.
5	I(LIM)	Current limit sense pin. Normally connected to current sense resistor.	13	PGND	Return for the High Current Totem pole output.
6	RDCT	Off-time setting capacitor and resistor.	14	VC	Positive Supply for the High Current Totem pole output.
7	RC	Resistor to pin 6 to limit CT charging rate.	15	VCC	Positive Supply for the IC.
8	FF	Capacitor to generate feedforward ramp.	16	5.1V REF	Buffered output for the 5.1V voltage reference.

6

ML4815

ABSOLUTE MAXIMUM RATINGS

Absolute maximum ratings are those values beyond which the device could be permanently damaged. Absolute maximum ratings are stress ratings only and functional device operation is not implied.

Supply Voltage (Pins 14, 15)	30V
Output Current, Source or Sink (Pin 12)	
DC	0.5A
Pulsed (0.5 μ s)	2A
Analog Inputs (Pins 1, 2, 4, 5, 8, 9, 10)	-0.3V to 6V
Error Amplifier Output Current (pin 3)	-5mA
Soft Start Sink Current (Pin 9)	100mA
Feedforward Sink Current (Pin 8)	80mA
C _T Charging Current (Pin 7)	-50mA

Junction Temperature

ML4815C	125°C
Storage Temperature Range	-65°C to +150°C
Lead Temperature (Soldering 10 sec)	+260°C
Thermal Resistance (θ_{JA})	
Plastic DIP	65°C/W
Plastic Chip Carrier (PCC)	60°C/W

OPERATING CONDITIONS

Temperature Range

ML4815C	0°C to 70°C
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ELECTRICAL CHARACTERISTICS

Unless otherwise noted, these specifications apply for C_T = 330pF, R_C = 100 Ω , R_D = 2K Ω , V_{CC} = 15V, T_A = Operating Temperature Range. Pin numbers refer to 16-pin DIP.

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
Reference Section					
Output Voltage	T _J = 25°C, I _O = 1mA	5.00	5.10	5.20	V
Line Regulation	10V < V _{CC} < 30V		2	20	mV
Load Regulation	1mA < I _O < 10mA		5	20	mV
Temperature Stability	-55°C < T _J < 125°C, (note 1)		.2	.4	%
Total Variation	line, load, temp (note 1)	4.95		5.25	V
Output Noise Voltage	10Hz to 10KHz		50		μ V
Long Term Stability	T _J = 125°C, 1000 hrs, (note 1)		5	25	mV
Short Circuit Current	V _{REF} = 0V	-15	-50	-100	mA
Error Amplifier Section					
Input Offset Voltage				20	mV
Input Bias Current			.6	3	μ A
Input Offset Current			.1	1	μ A
Open Loop Gain	1 < V _O < 4V	60	96		dB
CMRR	1.5 < V _{CM} < 5.5V	75	95		dB
PSRR	12 < V _{CC} < 25V	75	110		dB
Output Sink Current	V _{PIN 3} = 1V	1	2.5		mA
Output Source Current	V _{PIN 3} = 4V	-.5	-1.3		mA
Output High Voltage	I _{PIN 3} = -0.5mA	4.0	4.7	5.0	V
Output Low Voltage	I _{PIN 3} = 1mA	0	0.5	1.0	V
Unity Gain Bandwidth	(note 1)	3	5.5		MHz
Slew Rate	(note 1)	6	12		V/ μ s
RAMP Comparator Section					
Pin 4 Bias Current	V _{PIN 7} = 0		-0.7		μ A
Pin 3 Zero DC Threshold	V _{PIN 2} = 2V, V _{PIN 1} = V _{PIN 3} V _{PIN 5} = 0, V _{PIN 6} = 1.5V V _{PIN 8} = 2V	1.05	1.20	1.55	V
Delay to Output	C _L = 0, (note 1)		55		ns

ELECTRICAL CHARACTERISTICS (Continued)

Unless otherwise noted, these specifications apply for $C_T = 330\text{pF}$, $R_C = 100\Omega$, $R_D = 2\text{K}\Omega$, $V_{CC} = 15\text{V}$, $T_A = \text{Operating Temperature Range}$. Pin numbers refer to 16-pin DIP.

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
Current Limit Comparator					
Pin 5 Input Bias Current	$0 < V_{PIN\ 5} < 4\text{V}$		2		μA
Current Limit Threshold			1.41		V
Hysteresis			30		mV
Delay to Output	$V_{PIN\ 10} = 0$, $C_L = 0$ (note 1)		50		ns
One-Shot Section					
Off-Time Initial Accuracy	$C_L = 0$, $T_A = 25^\circ\text{C}$		0.45		μs
Off-Time Voltage Stability	$C_L = 0$, $12\text{V} < V_{CC} < 25\text{V}$		5		%
Off-Time Temperature Stability	$C_L = 0$ (note 1)		5		%
Off-Time Total Variation	$C_L = 0$, line, temp (note 1)		6		%
Feedforward/Maximum On-Time Clamp Section					
Discharge Current	$V_{PIN\ 8} = 2.5\text{V}$		30		mA
On-Time Initial Accuracy	$C_{FF} = 330\text{pF}$, $R_{FF} = 2.7\text{K}\Omega$ to V_{REF} , $C_L = 0$		1.0		μs
Shutdown/Restart Section					
Pin 10 Charging Current			-250		μA
Overload Shutdown Threshold			2.3		V
Restart Threshold			1.1		V
Soft-Start Section					
Input Bias Current	$V_{PIN\ 9} = 4\text{V}$		1		μA
Discharge Current	$V_{PIN\ 9} = 1\text{V}$		25		mA
Under-Voltage Lockout Section					
Start Threshold			13.4		V
UVLO Hysteresis			3.6		V
Output Section					
Output Low Level	$I_{OUT} = 20\text{mA}$		0.25	0.40	V
	$I_{OUT} = 200\text{mA}$		1.2	2.2	V
Output High Level	$I_{OUT} = -20\text{mA}$		13.0		V
	$I_{OUT} = -200\text{mA}$		12.7		V
Rise/Fall Time	$C_L = 1\text{nF}$ (note 1)		30		ns
Supply Current					
Start Up Current	$V_{CC} = 8\text{V}$, $T_J = 25^\circ\text{C}$		2.0	3.0	mA
	$V_{CC} = 8\text{V}$, $T_J = -40^\circ\text{C}$		2.5	3.5	mA
Operating I_{CC}			28	38	mA

ML4815

FUNCTIONAL DESCRIPTION

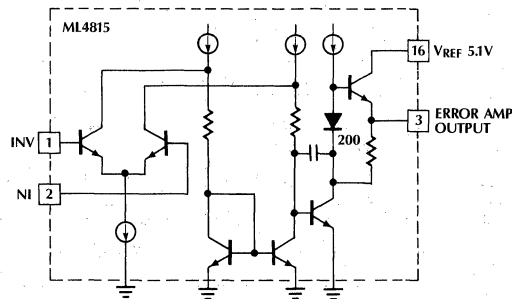
ML4815 PRINTED CIRCUIT BOARD LAYOUT CONSIDERATIONS

High speed circuits demand careful attention to layout and component placement. To assure proper performance of the ML4815, follow these rules: 1) Use a ground plane. 2) Damp or clamp parasitic inductive kick energy from the gate of driven MOSFET. Don't allow the output pins to ring below ground. A series gate resistor or a shunt 1 Amp Schottky diode at the

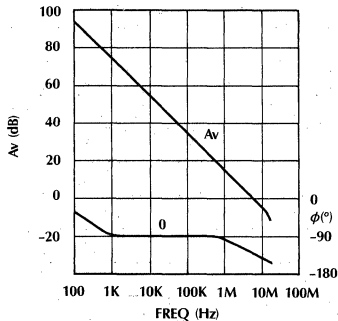
output pin will serve this purpose. 3) Bypass V_{CC} , V_C and V_{REF} . Use $1\mu F$ monolithic ceramic capacitors for V_{CC} and V_C with low equivalent series inductance. Allow less than 1 cm of total lead length for each capacitor between the bypassed pin and the ground plane. 4) Treat the off-time setting capacitor, C_T , like a bypass capacitor.

ERROR AMPLIFIER CIRCUIT

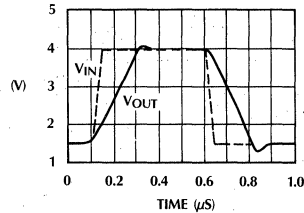
Simplified Schematic



Open-Loop Frequency Response



Unity Gain Slew Rate

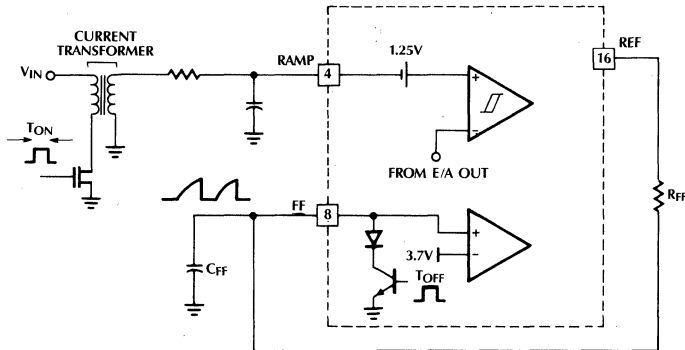


CONTROL METHODS

In current-mode control, the current transformer output is fed into the RAMP comparator input. The current-sense waveform is used as the on-time

modulating ramp. The on-time can be clamped to a maximum by using R_{FF} and C_{FF} as shown.

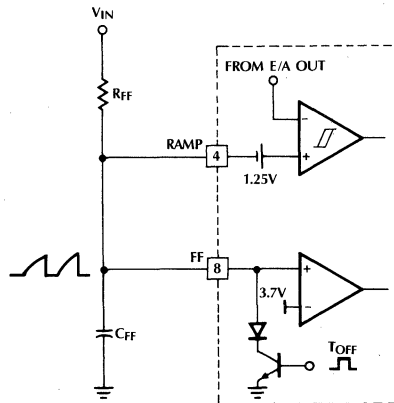
Current-Mode Control with Maximum On-Time Clamp



In feedforward voltage-mode control, the on-time modulating ramp is generated with an external capacitor C_{FF} from pin 8 to the ground. C_{FF} is charged through an external resistor R_{FF} . The maximum on-time

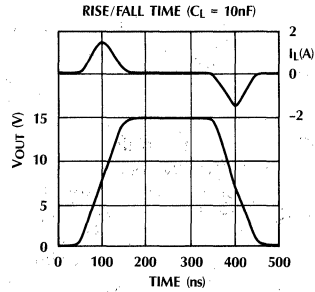
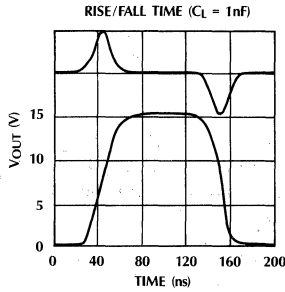
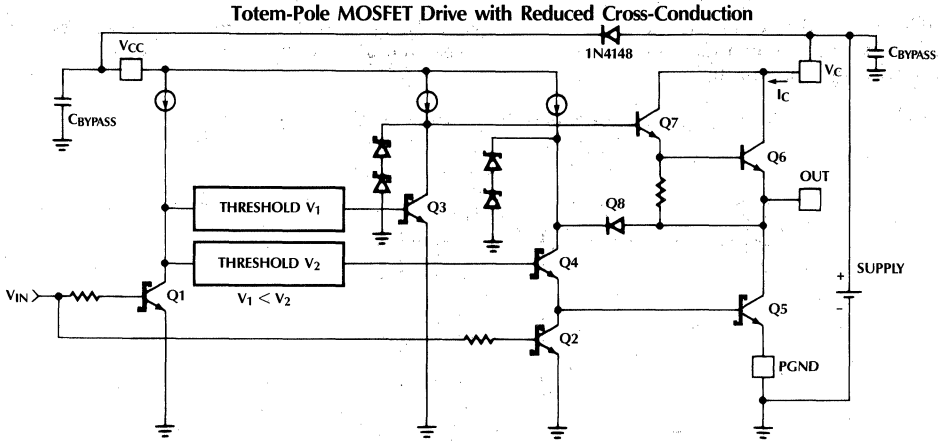
is the time taken to charge C_{FF} to 3.7V. Since the charging current depends on V_{IN} , the resulting maximum on-time varies with V_{IN} .

Feedforward Control

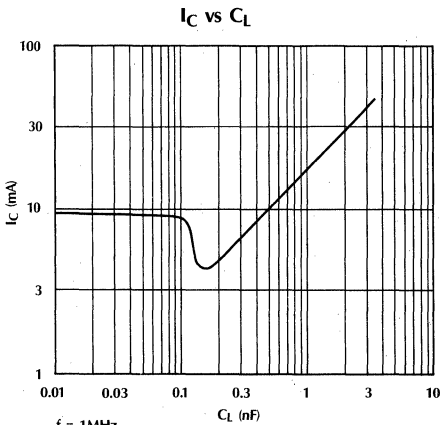


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OUTPUT SECTION

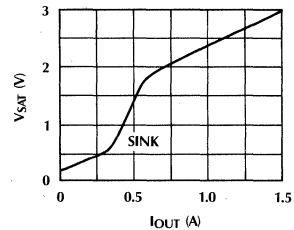


When driving power MOSFET's with high equivalent gate capacitance ($C_G > 3\text{ nF}$), it is advisable to use an external 1N4148 diode between V_{CC} and V_C pins (figure above) to reduce extra power dissipation caused by slow turn-off of Q_7 . In this case both V_{CC} and V_C pins should have bypass capacitors ($C = 1\mu\text{F}$) as close as possible to the IC pins.

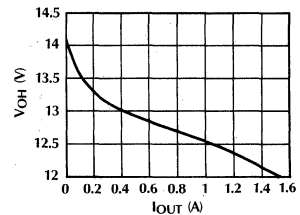


$f = 1\text{ MHz}$
 $T_J = 150^\circ\text{C}$
 $I_C = \text{TOTEM-POLE DRIVER CURRENT}$

V_{OL} Curve



V_{OH} Curve

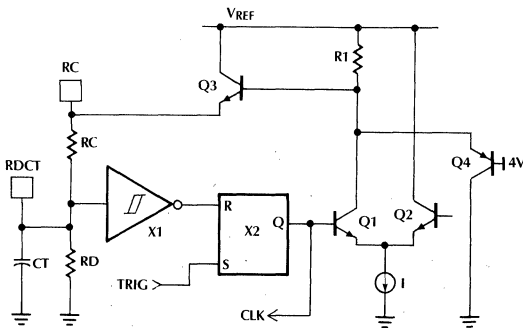


ONE-SHOT

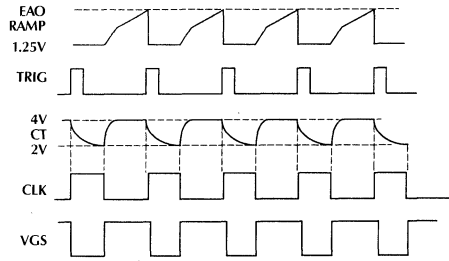
The figure below shows the detailed block diagram of the one-shot. The one-shot is programmed with external resistors R_C , R_D and capacitor C_T . Assuming that CLK is low and Q_2 conducts initially, the timing capacitor C_T is charged to 4V through R_C and Q_3 . This corresponds to the switch conduction cycle (on-time). When either the feedforward ramp or the sensed current signal exceeds the error amplifier output voltage, a trigger pulse is sent to the one-shot, setting the R-S latch X_2 and disabling Q_3 . C_T is immediately discharged through R_D until C_T voltage reaches the lower threshold (2V) of the Schmitt-trigger X_1 . At this point, X_1 output goes high, resetting X_2 . Q_1 turns off, allowing Q_3 to recharge C_T to 4V. This time interval corresponds to the switch off-time. Since the off-time is simply the discharge time of C_T , one can express

$$T_{OFF} = 0.69 R_D C_T$$

Block Diagram of the One-Shot



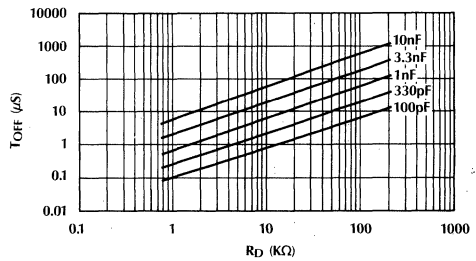
Timing Diagram of the One-Shot



The purpose of R_C is to slow the charging transient of C_T in order to widen the internal reset pulse. R_C is usually chosen such that the following inequality is satisfied.

$$\frac{R_C}{R_C + R_D} < 0.05$$

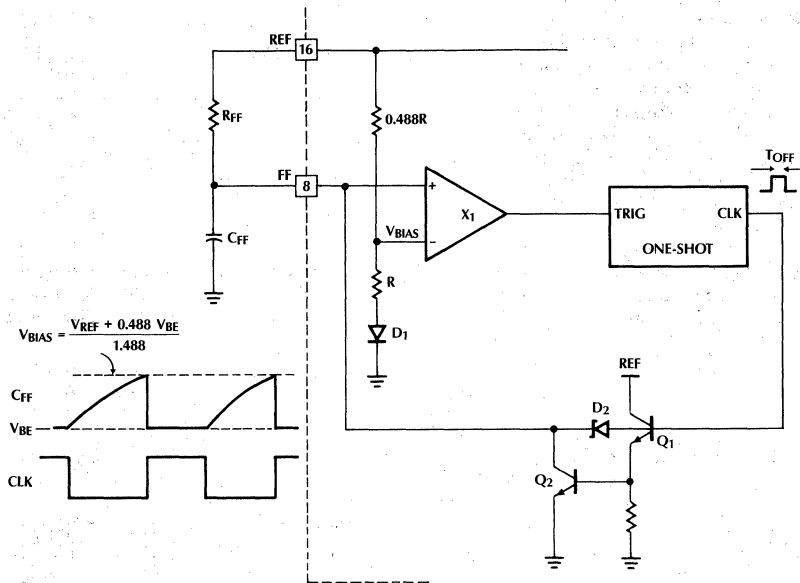
T_{OFF} vs R_D



ML4815

CONSTANT ON-TIME CLAMP (In Current-Mode Only)

Constant On-Time Clamp

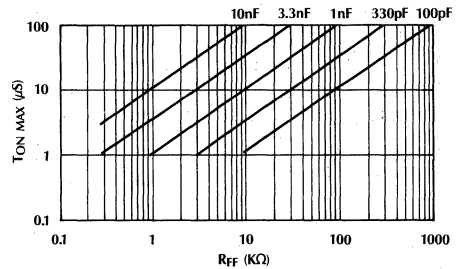


In current-mode control, the maximum on-time can be clamped by using the comparator X_1 (figure above). The internal transistors Q_1 , Q_2 and diode D_2 discharges C_{FF} to approximately V_{BE} . The time taken to charge C_{FF} from V_{BE} to V_{BIAS} ($= \frac{V_{REF} + 0.488 V_{BE}}{1.488}$) sets the maximum on-time. The diode D_1 compensates the V_{BE} dependent C_{FF} valley voltage. It can be shown that

$$T_{ON(MAX)} \approx 1.115 R_{FF} C_{FF}$$

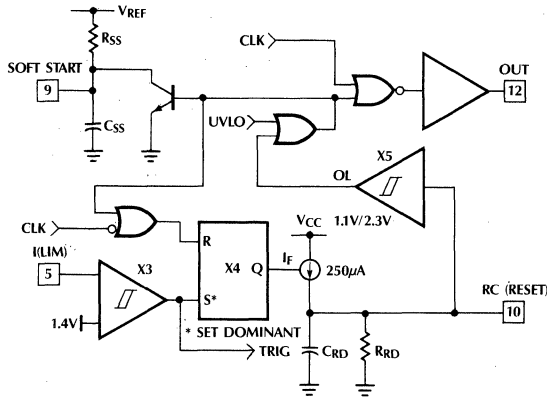
and $T_{ON(MAX)}$ is relatively independent of temperature.

T_{ON} vs R_{FF}



CURRENT-LIMITING, OVERLOAD SHUTDOWN AND FAULT MANAGEMENT

Overload Protection and Fault Management

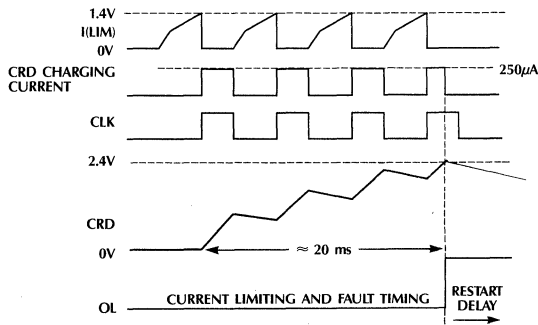


ML4815 features a unique overload protection scheme. The power transistor current is compared with the current-limit threshold (1.4V) of X₃. When the sensed current exceeds this threshold, the one-shot is triggered and the R-S latch X₄ is set. The one-shot blanks the gate drive and X₄ turns on the current source I_F. The external capacitor C_{RD}, which is normally fully discharged, is charged towards an overload threshold of 2.3V. The packet of charge delivered to C_{RD} in each over-current cycle is I_F × T_{OFF} (figure below). X₄ is reset after the off-time elapses. If output short is removed before C_{RD} reaches the overload threshold, C_{RD} will be discharged through R_{RD} and normal operation will resume. Under persistent output short circuit, C_{RD} is

charged until it reaches 2.3V. The gate drive is immediately terminated and the soft-start capacitor is discharged. C_{RD} then discharges through R_{RD} towards the restart threshold (1.1V). Gate drive remains off until C_{RD} is discharged below 1.1V. The time taken for C_{RD} to discharge to the restart threshold is the restart-delay time. This delay reduces the average power delivered to the load during overload, thus protecting both the load and the controller. If overload persists, the controller will continue to hiccup until the cause of overload is removed. The controller undergoes soft-start at each restart. The overload shutdown and restart sequence for a converter with non-bootstrapped power supply V_{CC} is illustrated in figure.

6

Current Limiting Overload Shutdown and Restart Sequence (Non-Bootstrapped Operation)

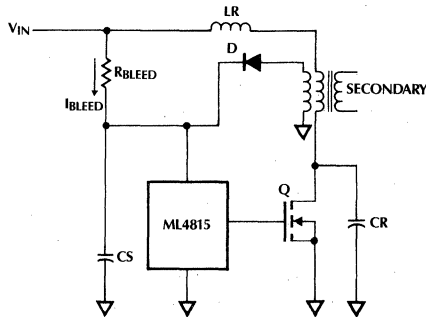


ML4815

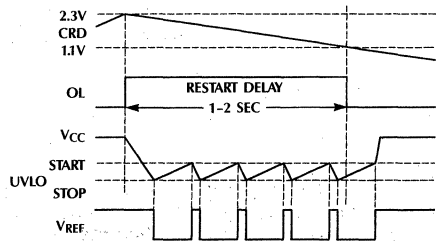
For a bootstrapped converter (where controller V_{CC} is obtained from an auxiliary winding of the main transformer), overload shutdown causes both the converter output and the controller V_{CC} to collapse. Undervoltage lockout (UVLO) is activated and the on-chip bandgap reference is disabled. ML4815 dissipates only 2mA of supply current during shutdown. Since

I_{BLEED} is higher than the start-up current, C_S will be charged towards the UVLO start threshold. When this happens, the entire controller becomes operational except that the gate drive remains off. I_{CC} jumps to its full operational value. Since V_{CC} bootstrapping is not yet available, I_{CC} will discharge C_S below the UVLO stop threshold. The on-chip reference will again be disabled with the controller supply current reduced to 2mA. I_{BLEED} will again charge C_S towards the UVLO start threshold. The process repeats until CRD is discharged below the restart threshold. The shutdown and restart sequence is illustrated with the timing diagram below.

Simplified V_{CC} Bootstrapping Scheme



Overload Shutdown UVLO and Restart Sequence (Bootstrapped Operation)



ML4815

ORDERING INFORMATION

PART NUMBER	TEMPERATURE RANGE	PACKAGE
ML4815CP	0°C to +70°C	Molded DIP (P16)
ML4815CQ	0°C to +70°C	Molded PCC (Q20)

High Frequency Multi-Mode Resonant Controller

GENERAL DESCRIPTION

The ML4816 controller IC is suitable for a wide range of resonant converter topologies. This controller can be used with Zero Current Switched (ZCS) Quasi Resonant Converters (QRC) requiring constant on-time and modulated off-time, as well as frequency modulated converters such as Series Resonant Converters operating above resonance.

The ML4816's oscillator features independent control of charging and discharging currents (on-time and off-time). Output frequency can be obtained either proportional or inversely proportional to the controlling voltage. In addition, both upper and lower frequency limits (f_{MIN} and f_{MAX}) can be independently set.

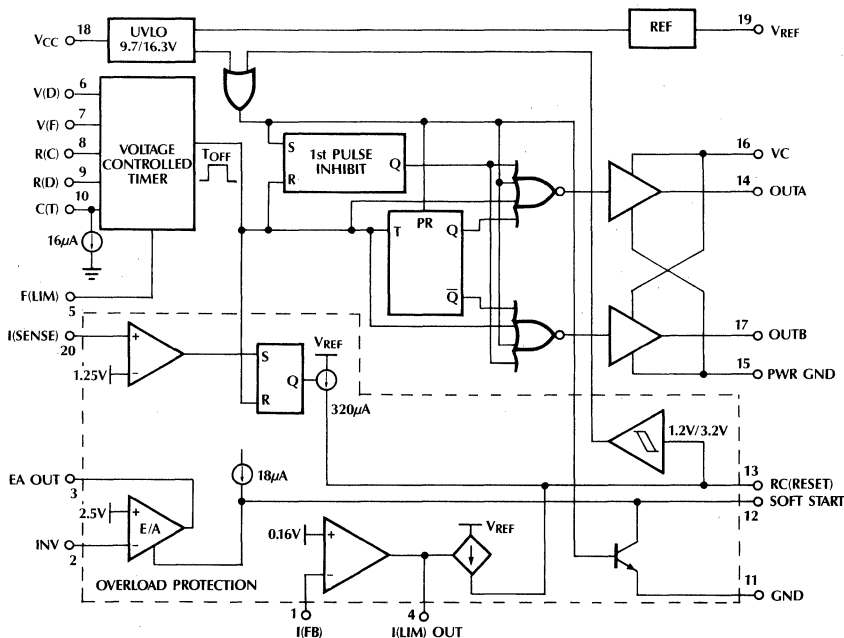
Both pulse-by-pulse and DC current limiting are provided for. Overload protection (shutdown) is triggered after a programmable delay time. Restart after overload shutdown can be delayed by a programmable time. Internal logic disables the gate drive until the oscillator is stable.

The ML4816 includes under-voltage lockout with 6V hysteresis and high current high speed totem pole output drivers for high speed drive of external MOSFETs.

FEATURES

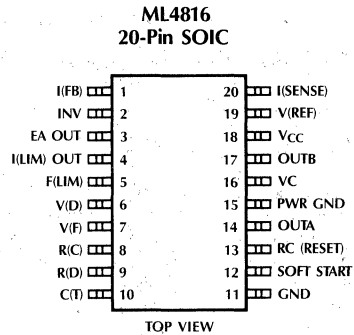
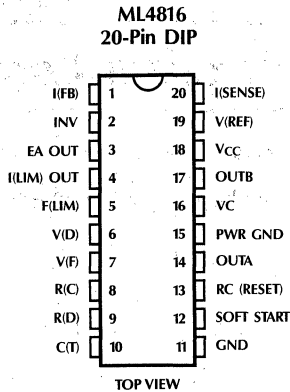
- Supports Zero Current Switched (ZCS) Quasi-Resonant Converters
- Supports Series Resonant (ZVS) converters operating above resonance
- Wide oscillator frequency range
- Programmable f_{MIN} and f_{MAX} limits
- Practical Operation to 2.5MHz (f_{OSC})
- Low Start-up Current and Under-Voltage Lockout Circuits support Off-Line Operation
- Pulse by Pulse or DC Current Limiting
- Integrating Soft Start Reset (Fault Integration) with Programmable Restart Delay
- High current (1.5A peak) totem-pole output drive
- Precision buffered 5V Reference ($\pm 1\%$)

BLOCK DIAGRAM



ML4816

PIN CONFIGURATION



PIN DESCRIPTION

PIN #	NAME	DESCRIPTION	PIN #	NAME	DESCRIPTION
1	I(FB)	Input for load current limit.	9	R(D)	External resistor from this pin to GND sets the oscillator discharge current (off time).
2	INV	Inverting input to error amp.	10	C(T)	Timing capacitor for Oscillator.
3	EA OUT	Output of error amplifier.	11	GND	Signal ground.
4	I(LIM) OUT	Output for load current limit amplifier.	12	SOFT START	Normally connected to Soft Start capacitor.
5	F(LIM)	A voltage input sets the maximum on time for the timer.	13	RC(RESET)	Timing elements for Integrating fault detection and reset delay circuits.
6	V(D)	Controls the C(T) discharge current and oscillator off time. Connected to error amplifier output for off-time modulation and to V(REF) for constant off time.	14	OUTA	High Current Totem pole output A.
7	V(F)	Controls the charging current and oscillator on time. Connected to error amplifier for on time modulation and connected to GND for constant on time.	15	PWR GND	Return for the High Current Totem Pole outputs.
8	R(C)	External timing resistor to either GND or V(REF) sets the charging current (oscillator on time). This pin can either source or sink current.	16	VC	Supply for the High Current Totem Pole outputs.
			17	OUTB	High Current Totem pole output B.
			18	V _{CC}	Positive supply for the IC.
			19	V(REF)	Buffered output for the 5.0V voltage reference.
			20	I(SENSE)	Primary current sense input for current limit.

ABSOLUTE MAXIMUM RATINGS

Supply Voltage (Pins 16, 18)	30V
Output Current, Source or Sink (Pin 12)	
DC	0.5A
Pulse (0.5 μ s)	1.5A
Analog Inputs	
(Pins 1, 2, 5, 6, 7, 13)	-0.3V to 6.3V
Amplifier Output Currents (Pins 3, 4)	5mA
Soft Start Sink Current (Pin 8)	100mA
R(C) Current (Pin 8)	-0.5 to +0.5mA
R(D) Current (Pin 9)	-4mA
Junction Temperature	150°C
Storage Temperature Range	-65°C to +150°C

Lead Temperature (Soldering 10 sec.)	+260°C
Thermal Resistance (θ_{JA})	
Plastic DIP or SOIC	65°C/W

OPERATING CONDITIONS

Temperature Range	0°C to +70°C
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Absolute maximum ratings are those values beyond which the device could be permanently damaged. Absolute maximum ratings are stress ratings only and functional device operation is not implied.

ELECTRICAL CHARACTERISTICS

Unless otherwise specified, $C_T = 470\text{pF}$, $V_{CC} = 15\text{V}$. V_{CC} is adjusted above the start threshold before settling at 15V.

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
Reference Section					
Output Voltage	$T_A = 25^\circ\text{C}$, $I_O = -1\text{mA}$	4.90	5.00	5.10	V
Line Regulation	$12\text{V} \leq V_{CC} \leq 25\text{V}$		2	20	mV
Load Regulation	$1\text{mA} \leq I_O \leq 10\text{mA}$		5	20	mV
Temperature Stability	$T_{\text{MIN}} \leq T_A \leq T_{\text{MAX}}$ (note 1)		0.2	0.4	mV/°C
Total Variation	line, load, temp.	4.85		5.15	V
Output Noise Voltage	$10\text{Hz} < f < 10\text{KHz}$		50		μV
Long Term Stability	$T_J = 125^\circ\text{C}$, 1000 Hrs (note 1)		5	25	mV
Short Circuit Current	$V_{\text{REF}} = 0$	-40	-70	-100	mA
Error Amplifier Section					
Non-Inverting Input Voltage		2.37	2.47	2.57	V
Input Bias Current				3	μA
Open-Loop Gain	$1 \leq V_O \leq 4\text{V}$	60			dB
Unity Gain Bandwidth	(note 1)	2.5	2.8		MHz
PSRR	$12\text{V} \leq V_{CC} \leq 25\text{V}$	70			dB
Output Sink Current	$V_{\text{PIN } 2} = 2.7\text{V}$, $V_{\text{PIN } 3} = 1\text{V}$	1	2.8		mA
Output Source Current	$V_{\text{PIN } 2} = 2.3\text{V}$, $V_{\text{PIN } 3} = 4\text{V}$	-0.5	-2.2		mA
Output High Voltage	$I_{\text{PIN } 3} = -0.5\text{mA}$	5.0	5.5	6.0	V
Output Low Voltage	$I_{\text{PIN } 3} = 1\text{mA}$		0.5	1.0	V
Slew Rate			8.5		V/ μs
Current-Limit Amplifier Section					
Non-Inverting Input Voltage		0.145	0.16	0.175	V
Input Bias Current				-1	μA
Open-Loop Gain	$1 \leq V_O \leq 4\text{V}$	65			dB
Unity Gain Bandwidth	(note 1)	1.0	1.5		MHz
PSRR	$12\text{V} \leq V_{CC} \leq 25\text{V}$	55			dB
Output Sink Current	$V_{\text{PIN } 1} = 1\text{V}$, $V_{\text{PIN } 4} = 1\text{V}$	1	1.6		mA

ML4816

ELECTRICAL CHARACTERISTICS (Continued)

Unless otherwise specified, $C_T = 470\text{pF}$, $V_{CC} = 15\text{V}$. V_{CC} is adjusted above the start threshold before settling at 15V.

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS	
Current-Limit Amplifier Section (Continued)						
Output Source Current	$V_{PIN\ 1} = 0\text{V}$, $V_{PIN\ 4} = 4\text{V}$	-0.5	-1.1		mA	
Output High Voltage	$I_{PIN\ 4} = -0.5\text{mA}$	6.0	7.2	8.0	V	
Output Low Voltage	$I_{PIN\ 4} = 1\text{mA}$		0.7	1.0	V	
Slew Rate	(note 1)		0.9		V/ μs	
Current-Sense Section						
Input Bias Current	$V_{PIN\ 20} = 0$			-2	μA	
Current-Sense Threshold		1.20	1.25	1.30	V	
Delay to Pin 13	(note 1)		80	150	ns	
Soft-Start Section						
Discharging Current	$V_{PIN\ 13} = 4\text{V}$, $V_{PIN\ 12} = 1\text{V}$	20	35		mA	
Charging Current	$V_{PIN\ 13} = 0$, $V_{PIN\ 12} = 1\text{V}$	-16	-21	-26	μA	
Overload Protection Section						
Overload Threshold		3.0	3.2	3.5	V	
Restart Threshold		1.0	1.2	1.4	V	
Pulse-by-pulse Charging Current	$V_{PIN\ 20} = 1.35\text{V}$, $V_{PIN\ 13} = 2\text{V}$		-320		μA	
Current-Limit Amp. Controlled Current	$V_{PIN\ 1} = 0$, $V_{PIN\ 13} = 2\text{V}$ $V_{PIN\ 4} = 1\text{V}$ $V_{PIN\ 4} = 2.5\text{V}$		-2.2 -0.9		mA mA	
Voltage-Controlled Timer						
C_T Minimum Discharging Current	$V_{PIN\ 6} = 0$, $V_{PIN\ 10} = 3\text{V}$	15	18.5	22	μA	
C_T Peak Voltage			3.75		V	
C_T Valley Voltage			2.1		V	
R(C) Minimum Voltage	$V_{PIN\ 5} = V_{PIN\ 7} = 0$, 25K Ω from Pin 8 to GND	$0.446V_{REF}$	$0.455V_{REF}$	$0.464V_{REF}$	V	
R(C) Voltage	$V_{PIN\ 5} = \frac{8}{11} V_{REF}$, $V_{PIN\ 7} = 5\text{V}$, 25K Ω from Pin 8 to GND	$0.713V_{REF}$	$0.727V_{REF}$	$0.742V_{REF}$	V	
R(D) Minimum Voltage	$V_{PIN\ 6} = 0$, 3K Ω from Pin 9 to GND			0	V	
R(D) Maximum Voltage	$V_{PIN\ 6} = 5\text{V}$, 3K Ω from Pin 9 to GND	$0.425V_{REF}$	$0.45V_{REF}$	$0.475V_{REF}$	V	
T_{ON}	$T_A = 25^\circ\text{C}$	$V_{PIN\ 5} = V_{PIN\ 7} = 0$, $V_{PIN\ 6} = 3\text{V}$, 25K Ω from Pin 8 to GND, 3K Ω from Pin 9 to GND	0.62	0.68	0.70	μs
	Total Variation	$12\text{V} \leq V_{CC} \leq 25\text{V}$ (note 1) $T_{MIN} \leq T_A \leq T_{MAX}$	0.60	0.71	0.79	μs
Output Dead Time	$T_A = 25^\circ\text{C}$ (note 1)	$V_{PIN\ 5} = V_{PIN\ 7} = 0$, $V_{PIN\ 6} = 5\text{V}$, 25K Ω from Pin 8 to GND, 3K Ω from Pin 9 to GND	100	120	145	ns
	Total Variation	$12\text{V} \leq V_{CC} \leq 25\text{V}$ (note 1) $T_{MIN} \leq T_A \leq T_{MAX}$	100	120	155	ns

ELECTRICAL CHARACTERISTICS (Continued)Unless otherwise specified, $C_T = 470\text{pF}$, $V_{CC} = 15\text{V}$. V_{CC} is adjusted above the start threshold before settling at 15V.

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS	
Voltage-Controlled Timer (Continued)						
Frequency	$f_{\text{MAX}(1)}$	$V_{\text{PIN } 5} = V_{\text{PIN } 7} = 0$, $V_{\text{PIN } 6} = 5\text{V}$ 25K Ω from Pin 8 to GND 3K Ω from Pin 9 to GND	1.1	1.2	1.3	MHz
	$f_{\text{MIN}(1)}$	$V_{\text{PIN } 5} = V_{\text{PIN } 7} = 0$, $V_{\text{PIN } 6} = 1.4\text{V}$ 25K Ω from Pin 8 to GND 3K Ω from Pin 9 to GND	17	22	28	KHz
	$f_{\text{MAX}(2)}$	$V_{\text{PIN } 5} = \frac{8}{11} V_{\text{REF}}$, $V_{\text{PIN } 7} = 2\text{V}$, $V_{\text{PIN } 6} = 5\text{V}$ 22K Ω from Pin 8 to Pin 19 3K Ω from Pin 9 to GND	1.2	1.45	1.55	MHz
	$f_{\text{MIN}(2)}$	$V_{\text{PIN } 5} = \frac{8}{11} V_{\text{REF}}$, $V_{\text{PIN } 7} = 5\text{V}$, $V_{\text{PIN } 6} = 5\text{V}$ 22K Ω from Pin 8 to Pin 19 3K Ω from Pin 9 to GND	650	800	825	KHz

Under Voltage Lockout Section

Start Threshold		15.8	16.3	16.8	V
Stop Threshold		9.2	9.7	10.2	V

Supply Current

Start-Up Current	$V_{CC} = 15.5\text{V}$	1.2	1.5	2.2	mA
Operating Supply Current	$V_{\text{PIN } 5} = V_{\text{PIN } 7} = 0$, $V_{\text{PIN } 6} = 5\text{V}$ 25K Ω from Pin 8 to GND 3K Ω from Pin 9 to GND $C_{\text{LA}} = C_{\text{LB}} = 0$, $T_A = 25^\circ\text{C}$,	26	32	38	mA
	$T_{\text{MIN}} \leq T_A \leq T_{\text{MAX}}$			53	mA

Output Section

Output Low Level	$I_{\text{SINK}} = 20\text{mA}$		0.1	0.4	V
	$I_{\text{SINK}} = 200\text{mA}$		0.7	2.2	V
Output High Level	$I_{\text{SOURCE}} = 20\text{mA}$	12.0	13.5		V
	$I_{\text{SOURCE}} = 200\text{mA}$	11.5	13.0		V
Rise Time	$C_{\text{LA}} = C_{\text{LB}} = 1\text{nF}$ (note 1)			60	ns
Fall Time	$C_{\text{LA}} = C_{\text{LB}} = 1\text{nF}$ (note 1)			60	ns

Note 1: This parameter is not 100% tested in production but guaranteed by design.

FUNCTIONAL DESCRIPTION

OSCILLATOR

The oscillator is the core of the ML4816 and is designed to allow the maximum flexibility. This oscillator can be used in two basic modes of operation:

1. On time proportional to V_{IN} , fixed off time with a maximum on time limit (where V_{IN} is the output of the error amplifier).
2. Off time inversely proportional to V_{IN} , fixed on time.

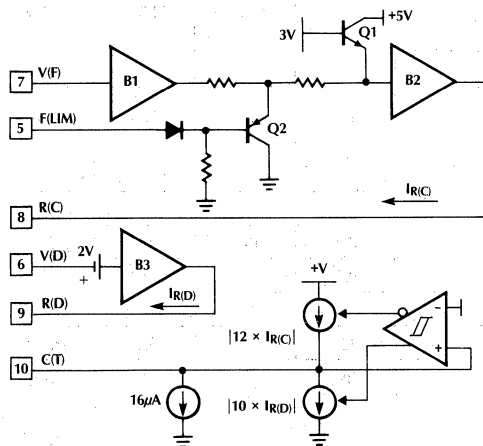
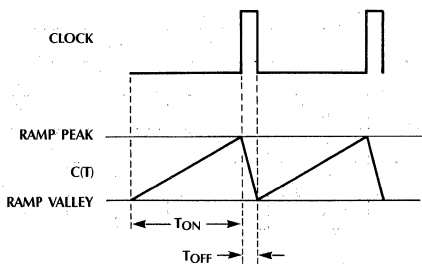


Figure 1. Oscillator Block Diagram



The internal CLOCK signal, shown above, turns the outputs off at its rising edge. Clock remains high and the outputs stay off as long as C(T) is discharging. The discharge time (T_{OFF}) of C(T) is:

$$T_{OFF} = \frac{1.65 C(T) R(D)}{10 (V(D) - 2V) + 16\mu A R(D)} \quad (1)$$

Variable Off-Time, Constant On-Time (Figure 2)

When using a variable off time control, V(D) is tied to the output of the error amplifier. Off time is given by equation (1) while the $16\mu A$ current sink prevents the off time from becoming infinite, thereby providing an upper limit to T_{OFF} of:

$$\text{Max } (T_{OFF}) = C(T) \times 1.03 \times 10^5 \quad (2)$$

The on time is given by:

$$T_{ON} = 0.0605 R(C) C(T) \quad (3)$$

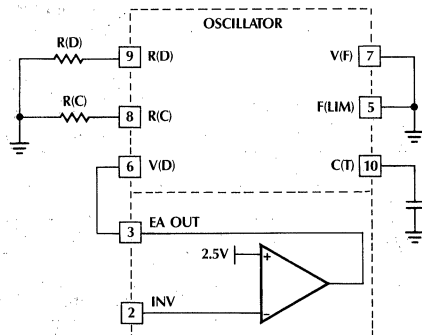


Figure 2. Variable Off Time, Constant On Time Oscillator Connections

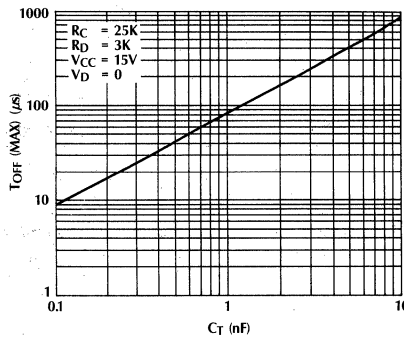


Figure 2a. Max (T_{OFF}) vs. C_T

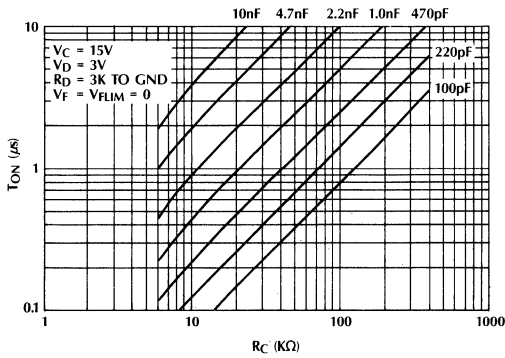


Figure 2b. T_{ON} vs. $R(C)$

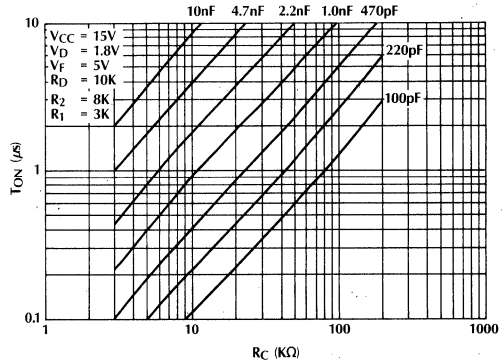


Figure 3a. Minimum T_{ON} for Constant Off-Time

Variable On-Time, Constant Off-Time

The on time (T_{ON}) is controlled by the current flowing from $V(REF)$ through $R(C)$ into B2. The output of B2 is internally limited to be no less than 2.27V and no greater than $F(LIM)$.

Configuration with $V_{FLIM} = \frac{8}{11} \times V_{REF}$

The on time for figure 3 is given by:

$$T_{ON} = \frac{0.138 R(C) C(T)}{V(REF) - V(F)} \tag{4}$$

The maximum on time is given by:

$$T_{ON(MAX)} = \frac{0.138 R(C) C(T)}{V(REF) - F(LIM)} \tag{5}$$

where $F(LIM) > 2.27V$. The minimum on time is:

$$T_{ON(MIN)} = 0.0506 R(C) C(T) \tag{6}$$

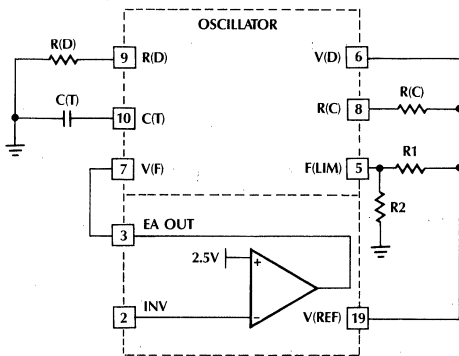
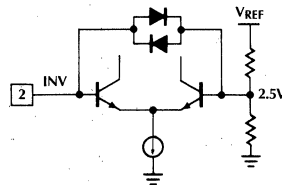


Figure 3. Variable On Time, Constant Off Time Oscillator Connections (T_{ON} Proportional to EA OUT)

ERROR AMPLIFIER

The ML4816 error amplifier is a 2.5MHz bandwidth, 8.5V/ μ sec slew rate op-amp with provision for limiting the positive output voltage swing to implement the soft start function.



The Error Amplifier input contains protection diodes as shown above. INV should not be driven lower than $2.5V - V_{BE}$ or higher than $2.5V + V_{BE}$.

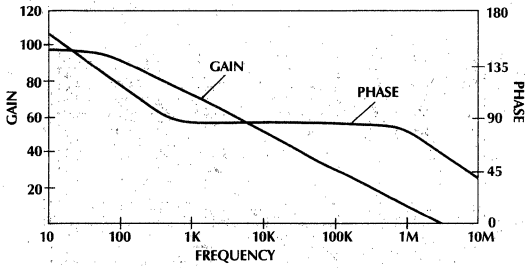


Figure 5. Error Amplifier Open-Loop Gain and Phase vs. Frequency

OUTPUT DRIVER STAGE

The ML4816 has two high current high speed totem pole output drivers each capable of 1.5A peak output, designed to quickly switch the gates of capacitive loads, such as power MOSFET transistors.

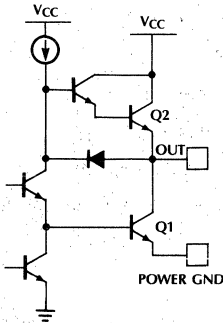


Figure 6. Power Driver Simplified Schematic

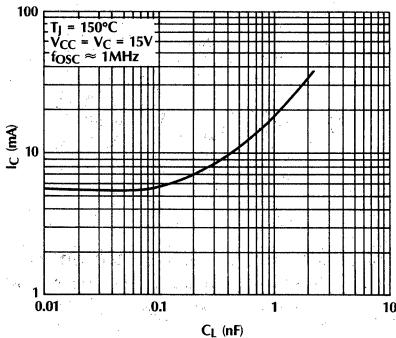


Figure 6a. Output Driver Current Consumption I(C) vs. Output Load Capacitance

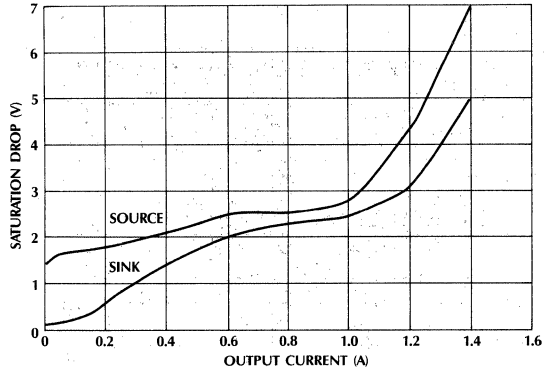


Figure 7. Output Saturation Voltage vs. Output Current

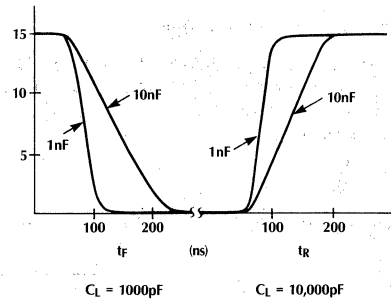


Figure 8. Rise/Fall Time

CURRENT LIMIT, FAULT DETECTION AND SOFT START

The ML4816 has two modes of current limiting: Primary pulse-by-pulse over-current protection and secondary DC average current limiting.

Primary Pulse-by-Pulse Current Limit Circuit

In this mode, the primary current is compared with a 1.25V threshold in comparator X1. When the sensed current exceeds the 1.25V threshold of comparator X1, the R-S latch X2 is set, turning on the 320μA current source to charge CRST. IF1 remains on until CLOCK goes high (TOFF). When CRST has charged to 3.2V, a soft start reset occurs. The number of times the outputs reach current limit are "remembered" on CRST. Over time, CRST is discharged by RST providing a measure of "forgetting" when the over-current condition no longer occurs. If the output fault is removed before CRST reaches 3.2V, CRST discharges slowly through RST and normal operation resumes.

Over-Current Sensing, Overload Shutdown and Fault Management

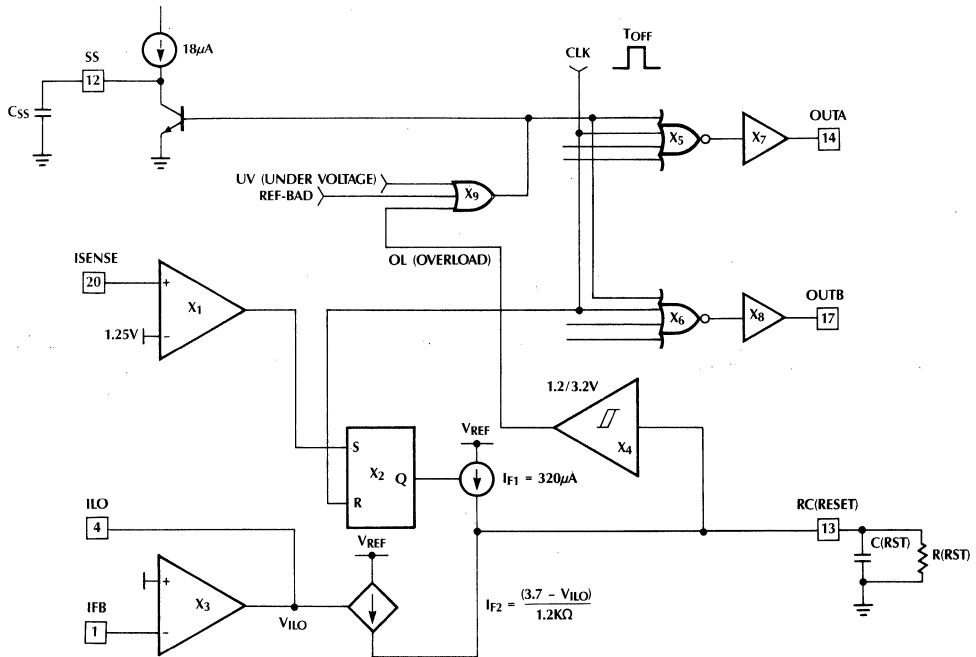


Figure 9. Overload Protection and Fault Management

Secondary dc Current Limit Circuit

In secondary dc current-limiting, the currents in the output rectifiers are sensed, full-wave rectified and smoothed. The smoothed signal is fed into the current-limiting amplifier X_3 . If the sensed current is below the 0.16V threshold, the output of X_3 will go above V_{REF} and I_{F2} will be off. As the sensed current exceeds the current-limit threshold, V_{ILO} starts to fall and

$$I_{F2} \left(\approx \frac{V_{REF} - V_{ILO} - 2V_{BE}}{1200\Omega} \right) \text{ turns on. } I_{F2} \text{ charges } C_{RST}$$

towards the overload threshold (3.2V) of X_4 . C_{RST} charging and temporary recovery through R_{RST} here are similar to the pulse-by-pulse over-current sensing case except that I_{F2} is continuous.

Under persistent output short circuit with either form of over-current protection, C_{RST} is charged until it reaches 3.2V. The gate drives are immediately terminated and the soft-start capacitor C_{SS} is discharged. C_{RST} then discharges through R_{RST} toward the restart threshold (1.2V). Gate drives remain off until C_{RST} is discharged below 1.2V. The time taken for C_{RST} to discharge to the restart threshold is the restart-delay time. This delay reduces the average power delivered to the load during overload, thus protecting both the load and the controller. If overload persists, the controller will continue to hiccup until the cause of overload is removed. The controller undergoes soft-start at each restart.

The overload shutdown and restart sequences for both over-current protection schemes with non-bootstrapped V_{CC} are illustrated in Figures 10 and 11.

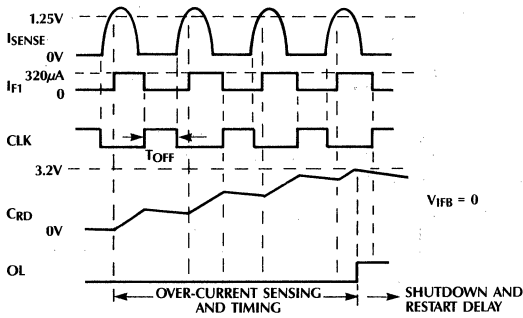


Figure 10. Over-Current Sensing, Overload Shutdown and Restart Sequence (Non-Bootstrapped V_{CC}) (ZCS-QRC Transistor Current Shown)

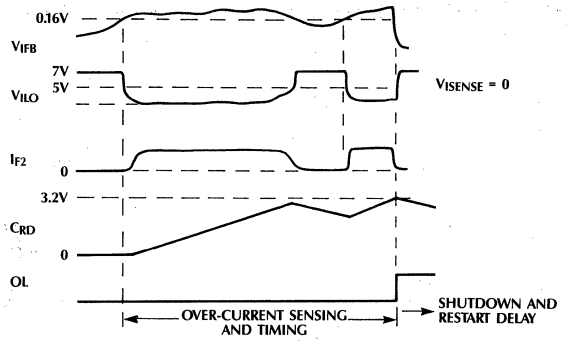


Figure 11. Secondary dc Current Sensing, Overload Shutdown and Restart Sequence (Non-Bootstrapped V_{CC})

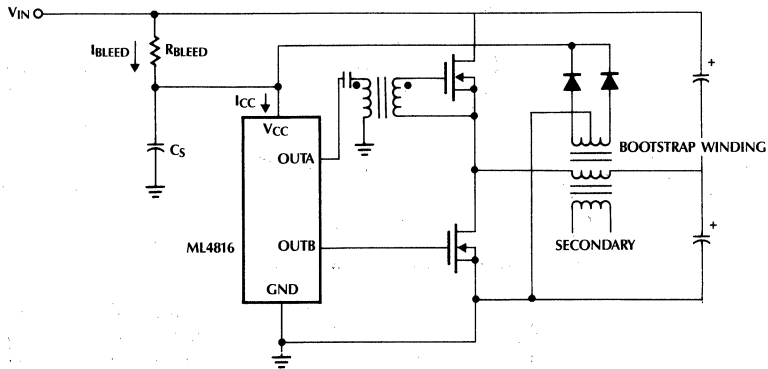


Figure 12. Simplified V_{CC} Bootstrapping Scheme in Half-Bridge Configuration

For a bootstrapped converter, where controller V_{CC} is obtained from an auxiliary winding of the main transformer, (see Figure 12) overload shutdown causes both the converter output and the controller V_{CC} to collapse. Undervoltage lockout (UVLO) is activated and the on-chip bandgap reference is disabled. ML4816 dissipates only 1.5mA during shutdown. Since I_{BLEED} is higher than the start-up current, C_S will be charged towards the UVLO start threshold. When this happens, the entire controller becomes operational except that the gate drives remain off. I_{CC} jumps to its full operational value. Since V_{CC} bootstrapping is not yet available, I_{CC} will discharge C_S below the UVLO stop threshold. The on-chip reference will again be disabled with the controller supply current reduced to 1.5mA. I_{BLEED} will again charge C_S towards the UVLO start threshold. The process repeats until C_{RST} is discharged below the restart threshold. The shutdown and restart sequence is illustrated in Figure 13.

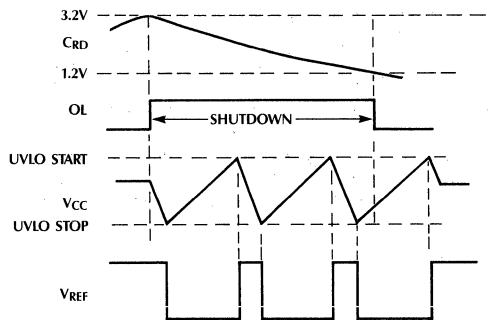


Figure 13. Overload Shutdown, UVLO and Restart Sequence (Bootstrapped Operation)

The over-current timing and shutdown sequence can be disabled by grounding pin 13.

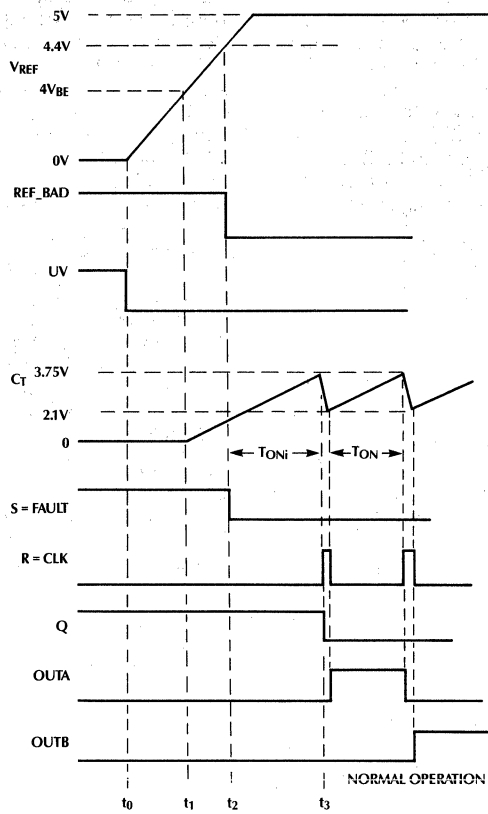
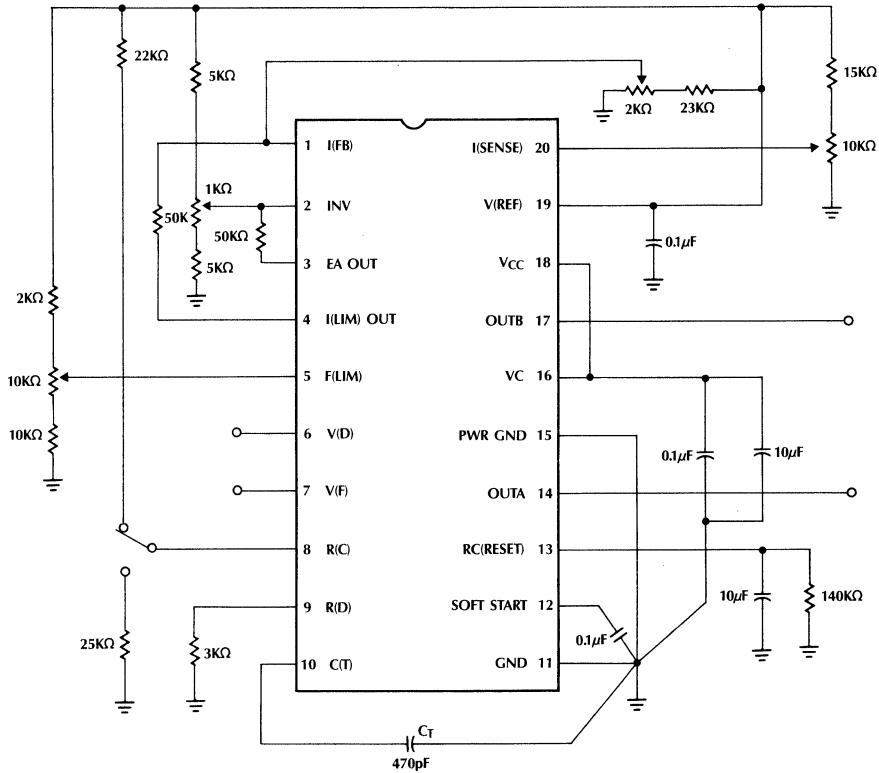


Figure 16. Timing Diagram Illustrating Initial Start-Up and the First-Pulse Inhibit

Open Loop Laboratory Test Fixture



This test fixture is useful for exercising many of the ML4816's functions and measuring their specifications. As with any wideband circuit, careful grounding and bypass procedures should be followed. The use of a ground plane is highly recommended.

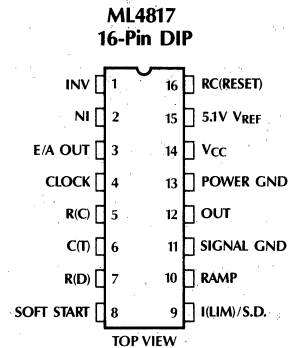
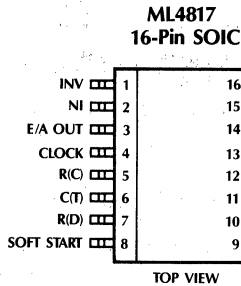
ML4816

ORDERING INFORMATION

PART NUMBER	TEMPERATURE RANGE	PACKAGE
ML4816CP	0°C to +70°C	Plastic DIP (P20)
ML4816CS	0°C to +70°C	Plastic SOIC (S20W)

ML4817

PIN CONFIGURATION



PIN DESCRIPTION

PIN #	NAME	FUNCTION	PIN #	NAME	FUNCTION
1	INV	Inverting input to error amp.	10	RAMP	Non-inverting input to main comparator. Connected to C(T) for Voltage Mode operation or to current sense resistor for current mode.
2	NI	Non-inverting input to error amp.	11	SIGNAL GND	Analog Signal Ground.
3	E/A OUT	Output of error amplifier and input to main comparator.	12	OUT	High Current Totem pole output.
4	CLOCK	Oscillator output.	13	POWER GND	Return for the High Current Totem pole outputs.
5	R(C)	Timing Resistor for Oscillator — sets charging current for oscillator timing capacitor (Pin 6).	14	V _{CC}	Positive Supply for the IC.
6	C(T)	Timing Capacitor for Oscillator.	15	5.1 V _{REF}	Buffered output for the 5.1V voltage reference.
7	R(D)	Resistor which sets discharge current for oscillator timing capacitor.	16	RC(RESET)	Timing elements for integrating fault detection and reset delay circuits.
8	SOFT START	Normally connected to Soft Start Capacitor and charging resistor.			
9	I(LIM)/S.D.	Current limit sense pin. Normally connected to current sense resistor.			

ABSOLUTE MAXIMUM RATINGS

Absolute maximum ratings are those values beyond which the device could be permanently damaged. Absolute maximum ratings are stress ratings only and functional device operation is not implied.

Supply Voltage (Pin 14)	30V
Output Current, Source or Sink (Pin 12)	
DC	0.5A
Pulsed (0.5 μ s)	2.0A
Analog Inputs	
(Pins 1, 2, 8, 9, 10, 16)	-0.3V to 6V
Clock Output Current (Pin 4)	-5mA
Error Amplifier Output Current (Pin 3)	5mA

Soft Start Sink Current (Pin 8)	100mA
Oscillator Charging Current (Pin 5)	-5mA
Junction Temperature	125°C
Storage Temperature Range	-65°C to +150°C
Lead Temperature (Soldering 10 sec)	+260°C
Thermal Resistance (θ_{JA})	
Plastic DIP	65°C/W
Plastic SOIC	65°C/W

OPERATING CONDITIONS

Temperature Range	0°C to +70°C
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ELECTRICAL CHARACTERISTICS

Unless otherwise specified, R(C) = 2540 Ω , R(D) = 2470 Ω , C_T = 470pF, T_A = Operating Temperature Range, V_{CC} = 15V.

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
Oscillator					
Initial Accuracy	T _A = 25°C	500	525	550	KHz
Voltage Stability	12V < V _{CC} < 25V		0.2		%
Temperature Stability	(note 1)			TBD	%
Total Variation	line, temp (note 1)	TBD		TBD	KHz
Maximum Duty Cycle	V _{PIN 1} = 2.3V, V _{PIN 2} = 2.5V V _{PIN 9} = V _{PIN 10} = 0V, T _A = 25°C	44	45	46	%
Maximum Duty Cycle	line, temp (note 1)	42		48	%
C(T) Discharge Current	V _{PIN 6} = 4V, V _{PIN 7} = 3V		4.5		mA
Clock Out High		4.0	4.5		V
Clock Out Low				2.2	V
Ramp Peak			3.75		V
Ramp Valley			2.15		V
Ramp Valley to Peak			1.60		V
Reference Section					
Output Voltage	T _A = 25°C, I _O = 1mA	5.00	5.10	5.20	V
Line Regulation	12V < V _{CC} < 25V		2	20	mV
Load Regulation	1mA < I _O < 10mA		5	20	mV
Temperature Stability	T _{MIN} < T _A < T _{MAX} (note 1)		.2	.4	mV/°C
Total Variation		4.95		5.25	V
Output Noise Voltage	10Hz to 10KHz		50		μ V
Long Term Stability	T _J = 125°C, 1000 hrs, (note 1)		5	25	mV
Short Circuit Current	V _{REF} = 0V	-15	-50	-100	mA

ELECTRICAL CHARACTERISTICS

Unless otherwise specified, R(C) = 2540Ω, R(D) = 2470Ω, C_T = 470pF, T_A = Operating Temperature Range, V_{CC} = 15V.

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
Error Amplifier Section					
Input Offset Voltage				15	mV
Input Bias Current			.6	3	μA
Input Offset Current			.1	1	μA
Open Loop Gain	1 < V _O < 4V	60	95		dB
CMRR	1.5 < V _{CM} < 5.5V	60	95		dB
PSRR	12 < V _{CC} < 25V	80	110		dB
Output Sink Current	V _{PIN 3} = 1V	1	2.5		mA
Output Source Current	V _{PIN 3} = 4.0V	-.5	-1.3		mA
Output High Voltage	I _{PIN 3} = -0.5mA	5.3			V
Output Low Voltage	I _{PIN 3} = 1mA	0	0.5	1.0	V
Unity Gain Bandwidth	(note 1)	3	5.5		MHz
Slew Rate	(note 1)	6	12		V/μs
PWM Comparator Section					
Pin 10 Bias Current	V _{PIN 10} = 0V		-1	-5	μA
Pin 3 Zero D.C. Threshold	V _{PIN 10} = 0V	1.08	1.25	1.45	V
Delay to Output	(note 1)		50	80	ns
Soft Start Section					
Pin 8 Bias Current	V _{PIN 8} = 4V			10	μA
Discharge Current	V _{PIN 8} = 1V	10			mA
Current Limit/Shutdown Section					
Pin 9 Bias Current	0V < V _{PIN 9} < 4V			+10	μA
Current Limit Threshold	V _{PIN 16} = 0V	1.35		1.65	V
Delay to Output	(note 1)		40	70	ns
Pin 16 Shutdown Threshold		2.05		2.55	V
Pin 16 Restart Threshold		0.9		1.3	V
Pin 16 Charging Current	V _{PIN 9} = 2V, V _{PIN 16} = 1.5V	-150	-210	-275	μA
Output Section					
Output Low Level	I _{OUT} = 20mA		.25	.4	V
	I _{OUT} = 200mA		1.2	2.2	V
Output High Level	I _{OUT} = -20mA	12.0	13.5		V
	I _{OUT} = -200mA	11.5	13.0		V
Rise/Fall Time	C _L = 1000pF, (note 1)		30	60	ns
Under-Voltage Lockout Section					
Start Threshold		12.0	13.8	15.0	V
UVLO Hysteresis		3.0	3.6	4.2	V
Supply Current					
Start Up Current			1.8	2.5	mA
I _{CC}	V _{PIN 1} = 2.3V, V _{PIN 2} = 2.5V V _{PIN 9, 10} = 0V, C _L = 0, T _A = 25°C		34	42	mA

Note 1. This parameter is not 100% tested in production but guaranteed by design.

FUNCTIONAL DESCRIPTION

OSCILLATOR

The ML4817 oscillator charges the external capacitor, C_T , with a current (I_{SET}) equal to $2/R_C$. When the C_T voltage reaches the upper threshold (Ramp Peak), the comparator changes state, turning off the current source and turning on the 4.5mA current sink which is voltage clamped to 1.05V by Q1. The capacitor then discharges to the lower threshold (Ramp Valley) with a time constant determined by R_D and C_T .

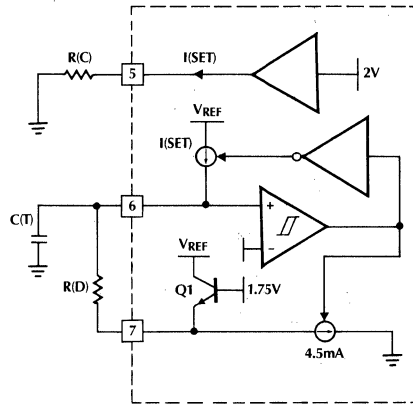
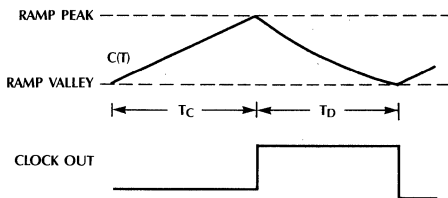


Figure 1. Oscillator Block Diagram

Oscillator period can be determined by the following formula:

$$T_{OSC} = T_C + T_D \tag{1}$$

$$T_C = \frac{(RAMP\ PEAK - VALLEY) C_T R_C}{2} \tag{2}$$

$$T_C = 0.8 (C_T R_C) \tag{3}$$

$$T_D = R_D C_T \ln \left(\frac{RAMP\ PEAK - 1.05}{RAMP\ VALLEY - 1.05} \right) \tag{4}$$

$$T_D = 0.90 (R_D C_T) \tag{5}$$

since: $f_{OSC} = \frac{1}{T_C + T_D}$

then: $f_{OSC} = \frac{1}{C_T (.8R_C + .90R_D)}$

since: Duty Cycle = $\frac{T_C}{T_C + T_D}$

then: Duty Cycle = $\frac{1}{1 + 1.12 \left(\frac{R_D}{R_C} \right)}$

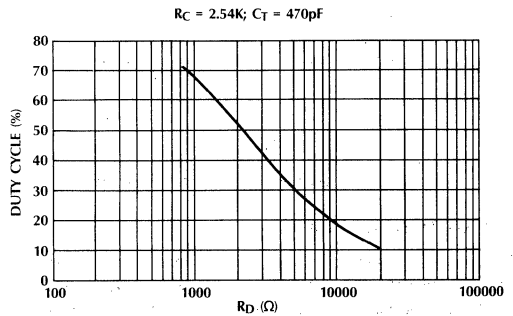


Figure 2. Duty Cycle vs R(D)

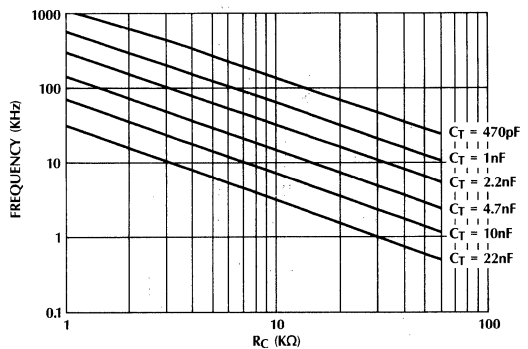


Figure 3. Oscillator Frequency vs $R_D = \frac{R_C}{1.03 R(C)}$ for 50% Duty Cycle

ERROR AMPLIFIER

The ML4817 error amplifier is a 5.5MHz bandwidth, $12V/\mu s$ slew rate op-amp with provision for limiting the positive output voltage swing for ease in implementing the soft start function.

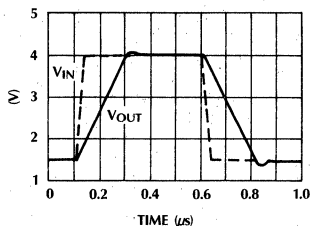


Figure 4. Unity Gain Slew Rate

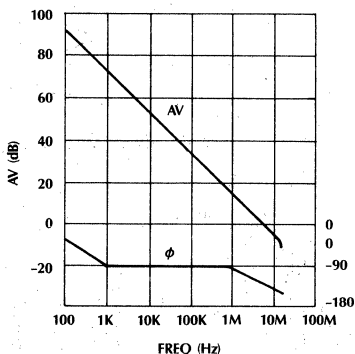


Figure 5. Open Loop Frequency Response

OUTPUT DRIVER STAGE

The ML4817 Output Driver is a 2A peak output high speed totem pole circuit designed to quickly switch capacitive loads, such as power MOSFET transistors.

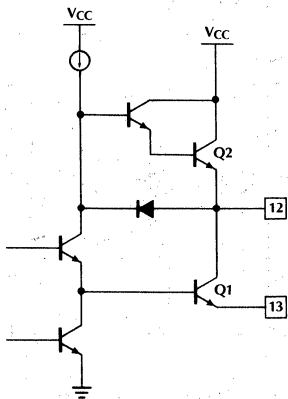
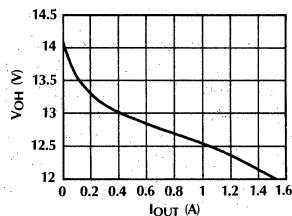


Figure 6. Power Driver Simplified Schematic

VOH Curve



VOL Curve

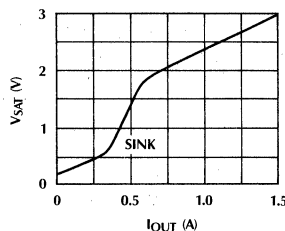


Figure 7. Saturation Curves

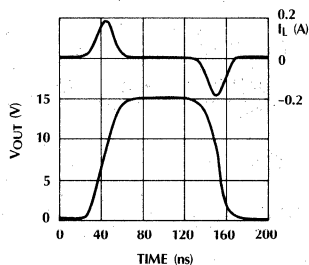


Figure 8. Rise/Fall Time ($C_L = 1000pF$)

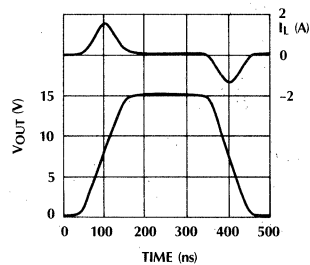


Figure 9. Rise/Fall Time ($C_L = 10,000pF$)

UNDER-VOLTAGE LOCKOUT

When V_{CC} is below 13.8V, the IC draws very little current (1.8mA typ.) and V_{REF} is disabled. When V_{CC} rises above 13.8V, the IC becomes active and V_{REF} is enabled and will stay in that condition until V_{CC} falls below 10.2V.

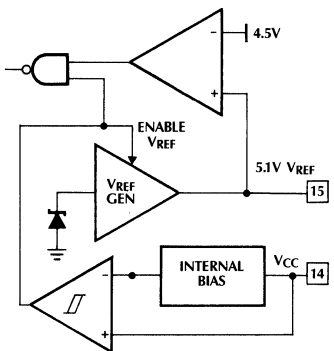


Figure 10. Under-Voltage Lockout Circuit

CURRENT LIMIT, FAULT DETECTION AND SOFT START

When the current (sensed on pin 9) reaches the 1.5V limit, the PWM cycle is terminated. The flip flop (figure 10) turns on current source I(1) to charge C_{RST} and remains on until CLOCK goes high. The magnitude of current source I(1) is $.25 \times I_{SET}$ where I_{SET} is the oscillator charging current. When C_{RST} is charged to 2.3V, a soft start reset occurs. The number of times the PWM cycle is terminated due to over-current is "remembered" on C(RST). Over time, C(RST) is discharged by R(RST) providing a measure of "forgetting" when the over-current condition no longer occurs.

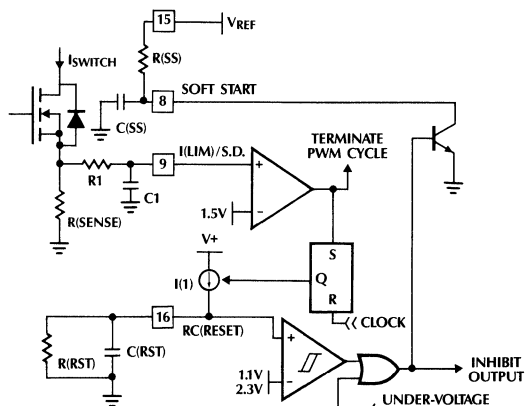


Figure 11. Over-Current, Soft Start, and Integrating Fault Detect Circuits

Since the per cycle charge on RC(RESET) is proportional to how early in the PWM cycle the reset occurs, a reset will occur more quickly under output short circuit conditions (figures 12c and 12d) than during a load surge (figures 12a and 12b).

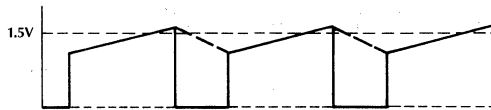


Figure 12a. Pin 9 (I_{LIMIT}) Waveform During Load Surge

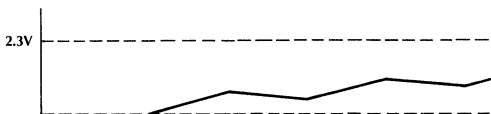


Figure 12b. Corresponding Waveform on Pin 16 (RC_{RESET})



Figure 12c. Current Waveform During Short Circuit (Pin 9)

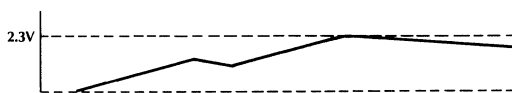


Figure 12d. RC(RESET) (Pin 16) Increases More Quickly During Short Circuit Condition

When the soft start reset occurs, the output is inhibited and the soft start capacitor is discharged. The output will remain off until C(RST) discharges to 1.1V through R(RST), providing a reset delay. When the IC restarts, the error amplifier output voltage is limited to the voltage at pin 8, thus limiting the duty cycle.

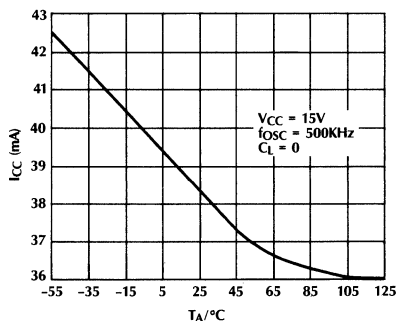


Figure 13. Supply Current vs. Temperature

ORDERING INFORMATION

PART NUMBER	TEMPERATURE RANGE	PACKAGE
ML4817CP	0°C to +70°C	DIP (P16)
ML4817CS	0°C to +70°C	SOIC (S16W)

Phase Modulation/Soft Switching Controller

GENERAL DESCRIPTION

The ML4818 is a complete phase modulation control IC suitable for full bridge soft switching converters. Unlike conventional PWM circuits, the phase modulation technique allows for zero voltage switching transitions and square wave drive across the transformer. The IC modulates the phases of the two sides of the bridge to control output power.

The ML4818 can be operated in current mode. The delay times for the outputs are externally programmable to allow the zero voltage switching transitions to take place.

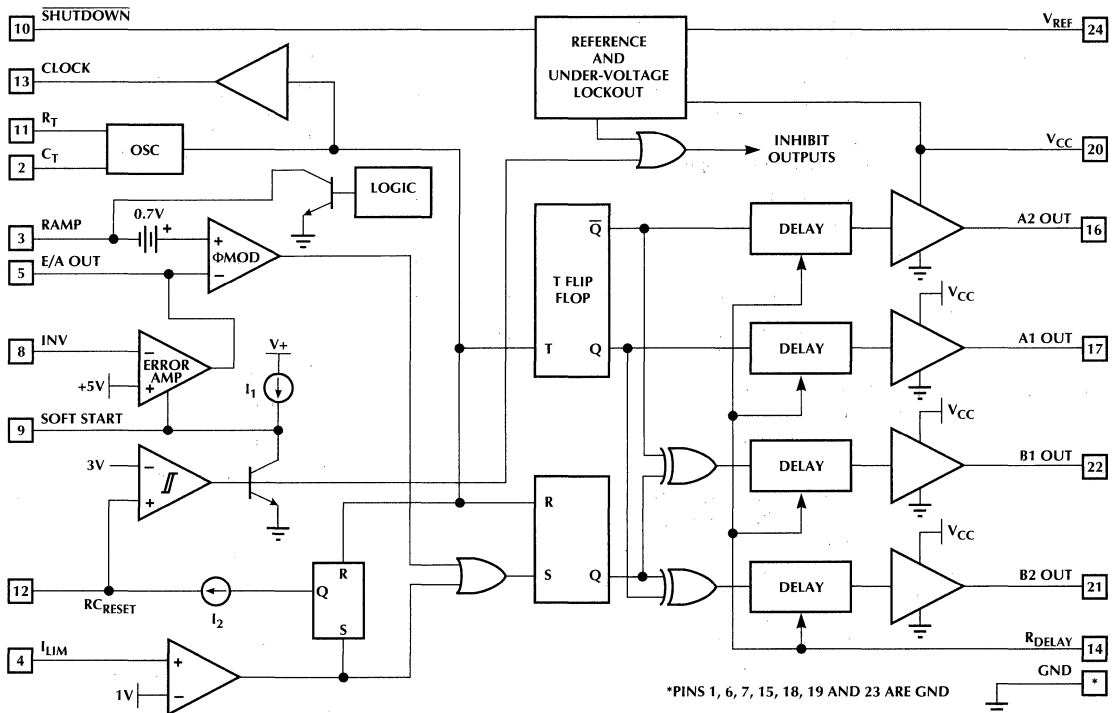
Both pulse-by-pulse current limit and integrating fault detection and soft start reset are provided. The under-voltage lockout circuit features a 6V hysteresis with a low starting current to allow off-line start up with a bleed resistor. A shutdown function powers down the IC, putting it into a low quiescent state.

The circuit can be operated at frequencies above 1MHz. The ML4818 contains four high current totem pole outputs which feature high slew rate with low cross conduction.

FEATURES

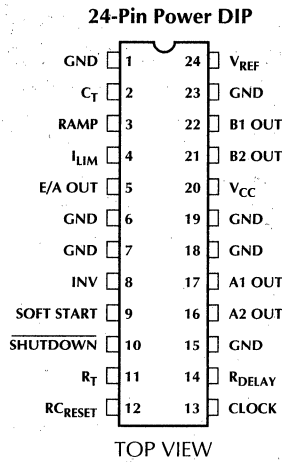
- Full Bridge Phase Modulation **Zero Voltage Switching** Circuit with Programmable ZV transition times
- Constant Frequency Operation to 1.5MHz
- Current Mode Operation
- Cycle-by-Cycle Current Limiting with Integrating Fault Detection and Restart Delay
- Precision buffered 5V Reference (+ 1%)
- Four 1.5 A Peak Current Totem-Pole Output Drivers
- Under-Voltage Lockout circuit with 6V Hysteresis.
- Power DIP package allows higher dissipation

BLOCK DIAGRAM



ML4818

PIN CONNECTION



PIN DESCRIPTION

PIN#	NAME	FUNCTION	PIN#	NAME	FUNCTION
1	GND	Ground	12	R _{CRESET}	Timing elements for Integrating fault detection and reset delay circuits
2	C _T	Timing Capacitor for Oscillator	13	CLOCK	Oscillator output
3	RAMP	Non-Inverting input to main comparator. Connected to current sense resistor for current mode	14	R _{DELAY}	Resistor to ground on this pin programs the amount of delay from the time an output turns off until its complementary output turns on
4	I _{LIM}	Current limit sense pin. Normally connected to current sense resistor	15	GND	Ground
5	E/A OUT	Output of error amplifier and input to PWM comparator	16	A2 OUT	High Current Totem pole output A1
6,7	GND	Ground and Substrate	17	A1 OUT	High Current Totem pole output A2
8	INV	Inverting input to error amp	18,19	GND	Ground and Substrate
9	SOFT START	Normally connected to Soft Start Capacitor	20	V _{CC}	Positive Supply for the IC
10	SHUTDOWN	Pulling this pin low puts the IC into a power down mode and turns off all outputs. This pin is internally pulled up to V _{ref} .	21	B2 OUT	High Current Totem pole output B1
11	R _T	Resistor which sets discharge current for oscillator timing capacitor	22	B1 OUT	High Current Totem pole output B2
			23	GND	Ground
			24	V _{REF}	Buffered output for the 5.0 V voltage reference

ABSOLUTE MAXIMUM RATINGS

Supply Voltage (Pin 20)	30V
Output Current, Source or Sink (Pins 16,17,21,22)	
DC	0.5A
Pulse (0.5 μ s)	1.5 A
Analog Inputs	
(Pins 2 thru 5, 8 thru 10, 12)	-0.3V to 6V
Clock Output Current (Pin 11)	-5mA
Error Amplifier Output Current (Pin 5)	5mA
Soft Start Sink Current (Pin 9)	50 mA
Oscillator Charging Current (Pin 2).....	-5mA
Junction Temperature	150°C

Storage Temperature Range	-65°C to +150°C
Lead Temperature (Soldering 10 Sec.)	+260°C
Thermal Resistance (θ_{JA}) (see fig 13,14)	
Plastic Power DIP	40°C/W

OPERATING CONDITIONS

Temperature Range 0°C to +70°C

Absolute maximum ratings are those values beyond which the device could be permanently damaged. Absolute maximum ratings are stress ratings only and functional device operation is not implied.

ELECTRICAL CHARACTERISTICS

Unless otherwise specified, $R_T = 12.7K\Omega$, $C_T = 250pF$, $R_{CLK} = 3K\Omega$, $R_{DELAY} = 5K\Omega$, $T_A =$ Operating Temperature Range, $V_{CC} = 15V$.

PARAMETER	CONDITIONS	MIN	TYP.	MAX	UNITS
OSCILLATOR					
Initial accuracy	$T_A = 25^\circ C$	460	500	540	KHz
Voltage stability	$12V < V_{CC} < 25V$		-0.3		%/V
Temperature stability	(note 1)		0.2		%
Total Variation	line, temp.	428		551	KHz
C_T Discharge Current	$V_{PIN2} = 2V$	4.7	5.5	6.3	mA
Clock out High		2.4	3.1		V
Clock out Low			0	0.4	V
Ramp Peak			4.1		V
Ramp Valley			1.5		V
Ramp Valley to Peak			2.6		V
REFERENCE SECTION					
Output Voltage	$T_A = 25^\circ C, I_O = 1mA$	4.95	5.0	5.05	V
Line regulation	$12V < V_{CC} < 25V$		2	20	mV
Load regulation	$1mA < I_O < 10 mA$		3	20	mV
Temperature stability	(note 1)		.2		mV/°C
Total Variation		4.85		5.15	V
Output Noise Voltage	10Hz to 10 KHz		50		mV
Long Term Stability	$T_j = 125^\circ C, 1000 hrs, (note 1)$		5	25	mV
Short Circuit Current	$V_{REF} = 0V$	-30	-50	-200	mA
ERROR AMPLIFIER SECTION					
Input Offset Voltage		-35		+25	mV
Input Bias Current			0.6	3	μA
Input Offset Current			0.1	1	μA
Open Loop Gain	$1 < V_O < 4V$	85	120		dB
PSRR	$12 < V_{CC} < 25V$	80	100		dB
Output Sink Current	$V_{PIN5} = 1V$	1	3.2		mA

ML4818

ELECTRICAL CHARACTERISTICS (Continued)

PARAMETER	CONDITIONS	MIN	TYP.	MAX	UNITS
ERROR AMPLIFIER SECTION (Continued)					
Output Source Current	$V_{PIN\ 5} = 5.1V$	-0.5	-2.2		mA
Output High Voltage	$I_{PIN\ 5} = -0.5mA$	5.0	5.5	6.0	V
Output Low Voltage	$I_{PIN\ 5} = 1mA$			0.5	V
Unity Gain Bandwidth	(note 1)	2.0	2.8		MHz
Slew Rate (note 1)			8.5		V/ μs
PHASE MODULATOR SECTION					
Pin 3 Bias Current	$V_{PIN\ 3} = 0V$		-1	-10	μA
Pin 5 Zero D.C. Threshold	$V_{PIN\ 3} = 0V$	0.4	0.6	0.8	V
T_{PDT} , pin 3 to Output	(note 1)		50	80	ns
T_{DELAY}		150	200	270	ns
V_{PIN14}			4.3		V
SOFT-START SECTION					
Pin 9 Charge Current	$V_{PIN\ 9} = 4V$	-20	-25	-30	μA
Discharge Current	$V_{PIN\ 9} = 1V$	10	20	30	mA
CURRENT LIMIT/SHUTDOWN SECTION					
Pin 4 Bias Current	$0V < V_{PIN\ 4} < 4V$		-1	-10	μA
Current Limit Threshold	$V_{PIN\ 10} = 0V$	0.92	1.02	1.12	V
T_{PDT} , pin 4	(note 1)		50		ns
Pin 12 Shutdown Threshold		3.15	3.4	3.65	V
Pin 12 Restart Threshold		1.0	1.3	1.6	V
Pin 12 Charging Current	$V_{PIN\ 4} = 2V, V_{PIN\ 12} = 1.5V$	-445	-523	-600	μA
Pin 10 Shutdown Threshold		2.0	2.4	2.8	V
Pin 10 Input Bias Current	$V_{PIN\ 10} = 0$		-25	-100	μA
OUTPUT SECTION					
Output Low Level	$I_{OUT} = 20\ mA$ $I_{OUT} = 200\ mA$		0.1 0.7	0.4 2.2	V V
Output High Level	$I_{OUT} = -20\ mA$ $I_{OUT} = -200\ mA$	12.0 11.5	13.5 13.0		V V
Rise/Fall Time	$C_L = 1000pF$, (note 1)		30	60	ns
UNDER-VOLTAGE LOCKOUT SECTION					
Start Threshold		15.8	16.5	17.2	V
Stop Threshold		9.7	10.2	10.7	V
SUPPLY CURRENT					
Start Up Current	$V_{CC} < 15.8V$		3	4	mA
I_{CC}	$V_{PIN\ 8} = 4V, V_{PIN\ 3,4} = 0V, C_L = 0$ $T_A = 25^\circ C$		48	60	mA

Note 1: This parameter not 100% tested in production but guaranteed by design.

FUNCTIONAL DESCRIPTION

PHASE MODULATOR

In the ML4818, power is controlled by modulating the phase of the A and B sides full bridge power section (figure 1). As shown in timing diagram, figure 2, we begin the power cycle with A2 and B1 high (Q2 and Q1 are on). Power is being delivered to the output through the transformer and the following sequence of events takes place:

1. After either the Φ MOD or I_{LIM} comparator trips, B1 goes low, turning off Q2, allowing the parasitic capacitances on Q2 and Q3 to charge to $+V_{IN}$ driven by the leakage inductance's current ($I_{LEAKAGE}$).
2. B2 goes high after T_{DELAY} which is set by the resistor R_{DELAY} from pin 14 to GND. At this point the source of Q3 has been charged to $+V_{IN}$, turning on

Q3 at zero voltage. The transformer is now effectively shorted through Q1 and Q3, allowing $I_{LEAKAGE}$ to recirculate.

3. Clock goes high and A2 goes low while A1 remains low for time T_{DELAY} . During this time both Q1 and Q4 are off. $I_{LEAKAGE}$ discharges the parasitic capacitances on Q1 and Q4 until Q4's drain is at 0V.
4. A1 goes high after T_{DELAY} which is set by the resistor R_{DELAY} from pin 14 to GND. At this point the drain of Q4 has been discharged to 0V, making the turning on Q4 at zero voltage. At this point, Q4 and Q3 are both on and power is being delivered to the main transformer.

The above sequence is then repeated but with the opposite polarity on all outputs.

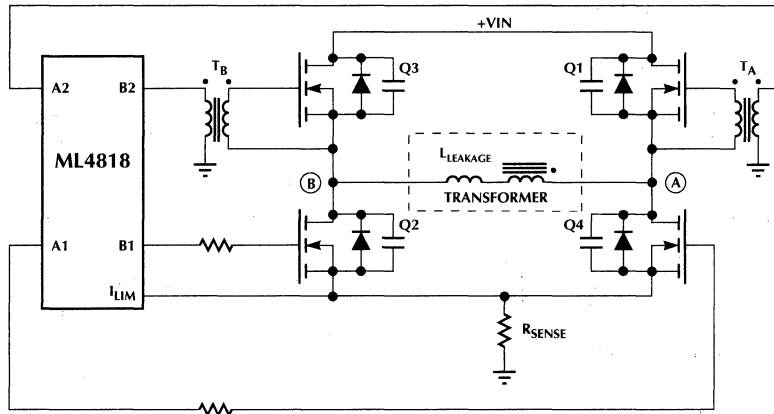


Figure 1. Simplified diagram of Phase Modulated power Outputs.

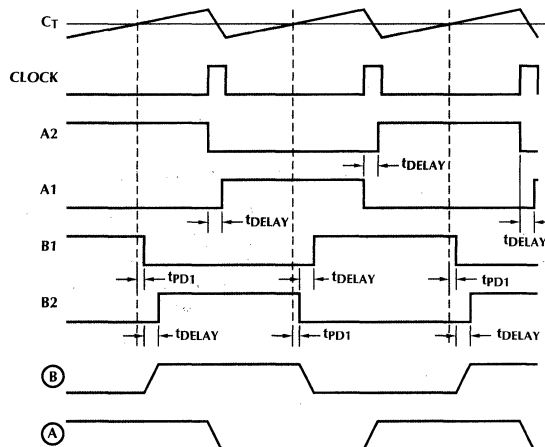


Figure 2. Phase Modulation control waveforms (Shaded areas indicate a power cycle).

ML4818

The ML4818 can also be used in current mode by sensing load current on the RAMP input (pin 3).

The four output delay timers are programmed via an external R_{DELAY} resistor as shown below. This resistor value should be no less than $1K\Omega$. Expressing R_{DELAY} in $K\Omega$ the delay, in ns is:

$$T_{DELAY} = 33 \times R_{DELAY} + 45 \quad (1)$$

The ML4818 contains special logic circuits to provide for voltage mode feed-forward and lock out long pulses into the internal logic. This prevents instability from occurring when the Φ Comparator trips in voltage mode.

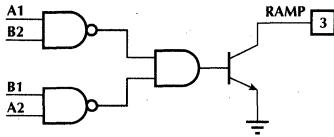


Figure 4. Voltage Feed-Forward Circuit.

The transistor in figure 4 is open only during a power cycle. When the power cycle terminates, pin 3 is pulled low. In voltage mode operation, a capacitor is connected from pin 3 to GND with a resistor from pin 3 to V_{IN} to provide input voltage feed forward.

OSCILLATOR

The ML4818 oscillator charges the external capacitor, C_T , with a current (I_{SET}) equal to $5/R_T$. When the C_T voltage reaches the upper threshold (Ramp Peak), the comparator changes state, turning on the current sink which discharges C_T to the lower threshold (Ramp Valley). The C_T pin is clamped to Ramp Valley by Q1 (figure 1) to prevent inaccuracy due to undershoot on C_T .

To use the Clock Output for driving external synchronization circuitry, a pull-down resistor is required from CLOCK to GND.

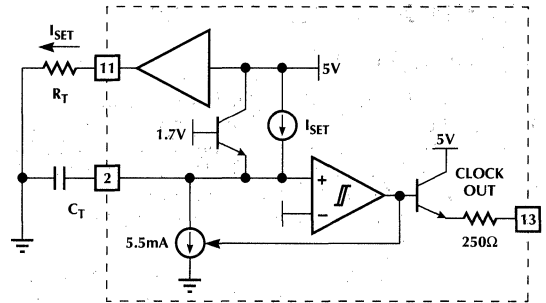
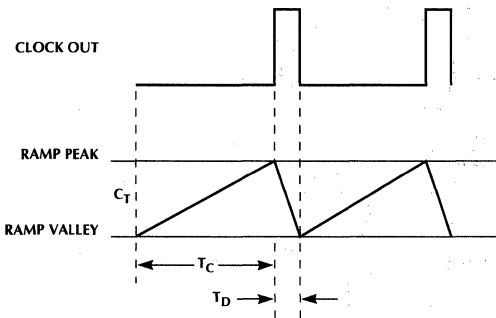


Figure 5. Oscillator Block Diagram.

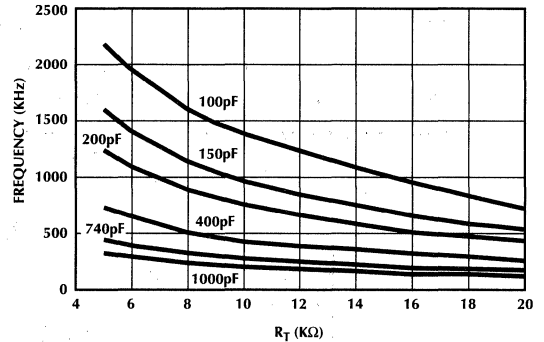


Figure 6. Oscillator Timing Resistance vs. Frequency.

Figure 6 should be used when oscillator frequencies of greater than 300KHz are required.

For frequencies of less than 300KHz, oscillator period can be determined by the following formulae:

$$T_{OSC} = T_C + T_D \quad (2)$$

$$T_C \approx 0.52 R_T C_T \quad (3)$$

$$T_D \approx 500 C_T \quad (4)$$

since:
$$f_{OSC} = \frac{1}{T_C + T_D}$$

then:
$$f_{OSC} = \frac{1}{0.52 C_T R_T + 500 C_T} \quad (5)$$

ERROR AMPLIFIER

The ML4818 error amplifier is a 2.5 MHz. bandwidth, $8.5V/\mu s$ slew rate op-amp with provision for limiting the positive output voltage swing (Output Inhibit line) to implement the soft start function. The error amplifier output source current is limited to 4.5mA.

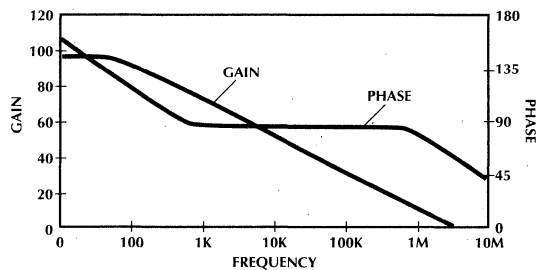


Figure 7. Error Amplifier Open-Loop Gain and Phase vs. Frequency.

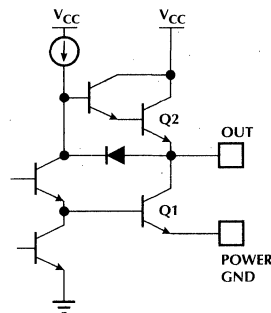


Figure 8. Power Driver Simplified Schematic.

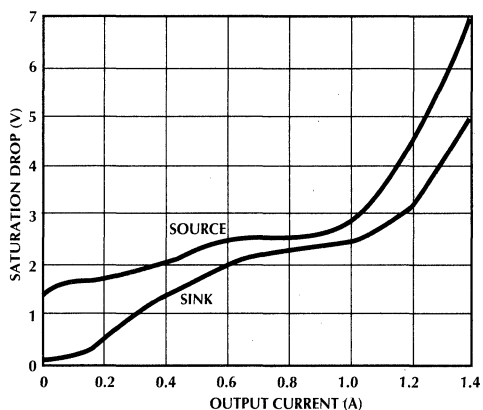


Figure 9. Output Saturation Voltage vs. Output Current.

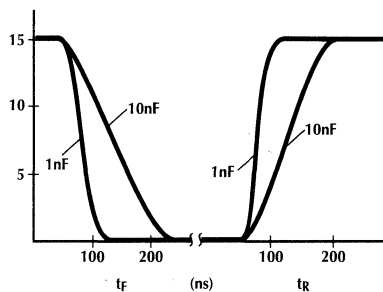


Figure 10. Rise/Fall Time.

OUTPUT DRIVER STAGE

The ML4818 has four high current high speed totem pole output drivers each capable of 1.5A peak output, designed to quickly switch the gates of capacitive loads, such as power MOSFET transistors.

ML4818

CURRENT LIMIT, FAULT DETECTION AND SOFT START

When the current (sensed on pin 4) reaches the 1V limit, the PWM cycle is terminated. The flip flop (figure 11) turns on the current source to charge C_{RST} and remains on for the duration of the clock period. When C_{RST} has charged to 3.4V, a soft start reset occurs. The number of times the PWM cycle is terminated due to over-current is "remembered" on C_{RST} . Over time, C_{RST} is discharged by R_{RST} providing a measure of "forgetting" when the over-current condition no longer occurs.

Since the per cycle charge on C_{RST} is proportional to how early in the power cycle the over-current occurs, a reset will occur more quickly under output short circuit conditions (figures 12a and 12b) than during a load surge (figures 12c and 12d).

When the soft start reset occurs, the output is inhibited and the soft start capacitor is discharged. The output will remain off until C_{RST} discharges to 1.3V through R_{RST} , providing a reset delay. When the IC restarts, the error amplifier output voltage is limited to the voltage at pin 9, thus limiting the duty cycle.

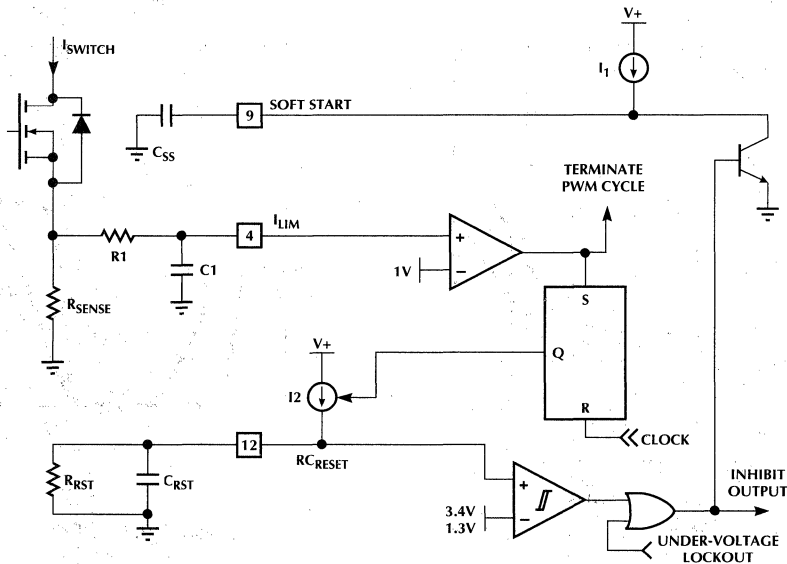


Figure 11. Over-Current, Soft-Start, and Integrating Fault Detect Circuits.

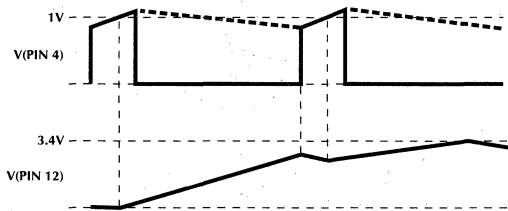


Figure 12a, 12b. I(LIMIT) and Resulting RC(RESET) Waveforms During Short Circuit.

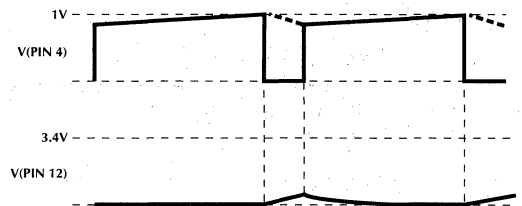


Figure 12c, 12d. I(LIMIT) and Resulting RC(RESET) Waveforms During Load Surge.

UNDER-VOLTAGE LOCKOUT

On power up, when V_{CC} is below 16V, the IC draws very little current (1.1mA typ.) and V_{REF} is disabled. When V_{CC} rises above 16V, the IC becomes active and V_{REF} is enabled and will stay in that condition until V_{CC} falls below 10.2V.

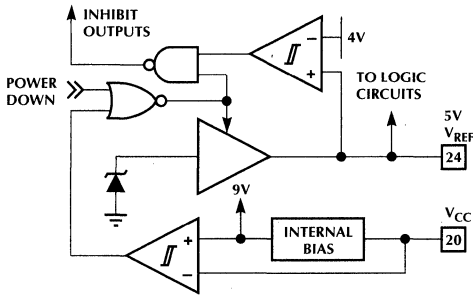


Figure 13. Under-Voltage Lockout and Reference Circuits.

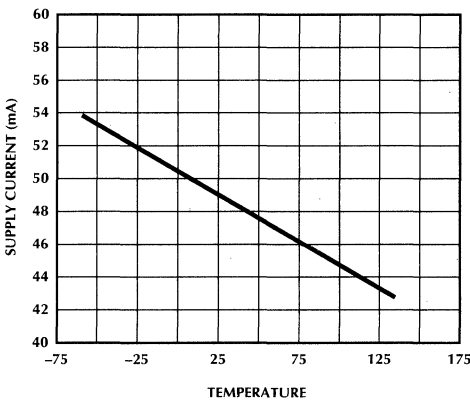


Figure 14. Supply Current vs. Temperature (°C).

THERMAL INFORMATION

The ML4818 is offered in a Power DIP package. This package features improved thermal conduction through the leadframe. Much of the heat is conducted through the center 4 grounded leads. Thermal dissipation can be improved with this package by using copper area on the board to function as a heat sink. Increasing this area can reduce the θ_{JA} (see figures 15 and 16), increasing the power handling capability of the package. Additional improvement may be obtained by using an external heat sink (available from Staver).

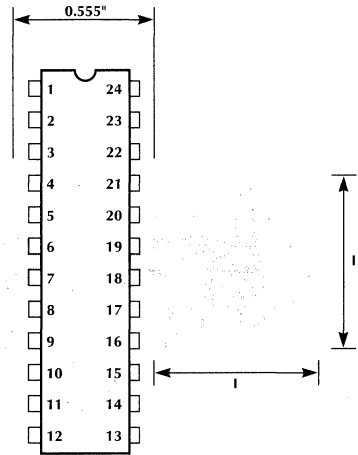


Figure 15. PC Board Copper Area Used as a Heat Sink.

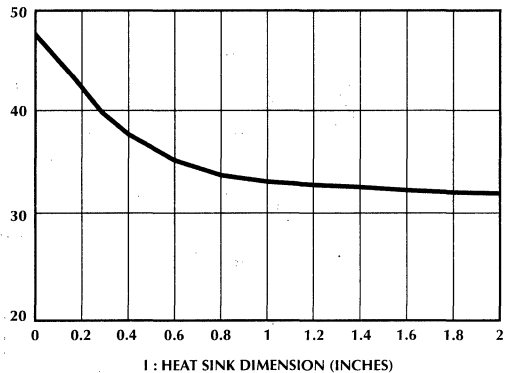


Figure 16. θ_{JA} as a Function of l (see figure 15).

APPLICATIONS

The application, in figure 17, features the ML4818 in a primary-side controlled voltage mode application with feed-forward. Input voltage is rectified 120VAC (nominal). Feed-forward is provided by the ramp on pin 3 via the resistor connected to the high voltage input. Current is sensed through sense transformer T4.

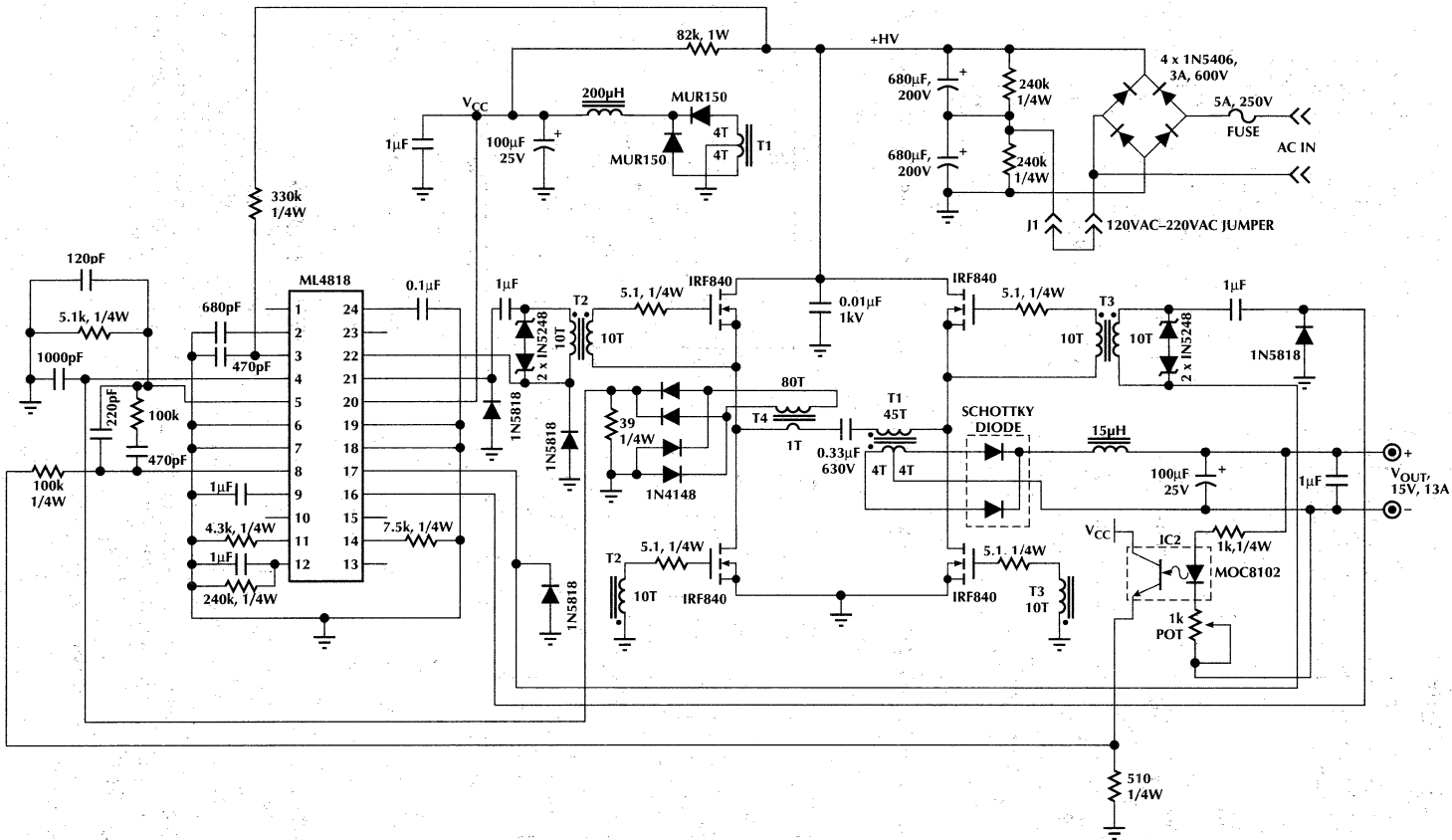


Figure 17. Offline Full Bridge Converter.

ORDERING INFORMATION

PART NUMBER	TEMPERATURE RANGE	PACKAGE
ML4818CP	0°C to +70°C	Power DIP (P24P)*

*Same dimensional outline as (P24N)

ML4818EVAL

Phase Modulation Controller Evaluation Kit

GENERAL DESCRIPTION

The ML4818EVAL kit provides a convenient vehicle to evaluate the ML4818 Phase Modulation Control IC. The board implements a 200W Phase Modulated Power Supply.

The application circuit is designed to show the performance of a 200W isolated AC to DC converter circuit. The circuit topology is a full bridge type, suitable for high power and very high frequency operation with zero voltage switching for high efficiency. The oscillator frequency of the circuit is 500KHz.

The PC board is designed around a 200W single output application circuit, however the circuit can be modified for other power levels and operating conditions.

KIT COMPONENTS

- User's Guide - Includes operating specification, procedure, kit component list, a complete parts list, performance data, and a detailed schematic.
- ML4818 Datasheet.
- Application Note #19: Phase Modulated PWM Topology with the ML4818
- Blank PC Board
- Kit Components - A sample of the ML4818 and additional components which may be difficult to procure in small quantities.

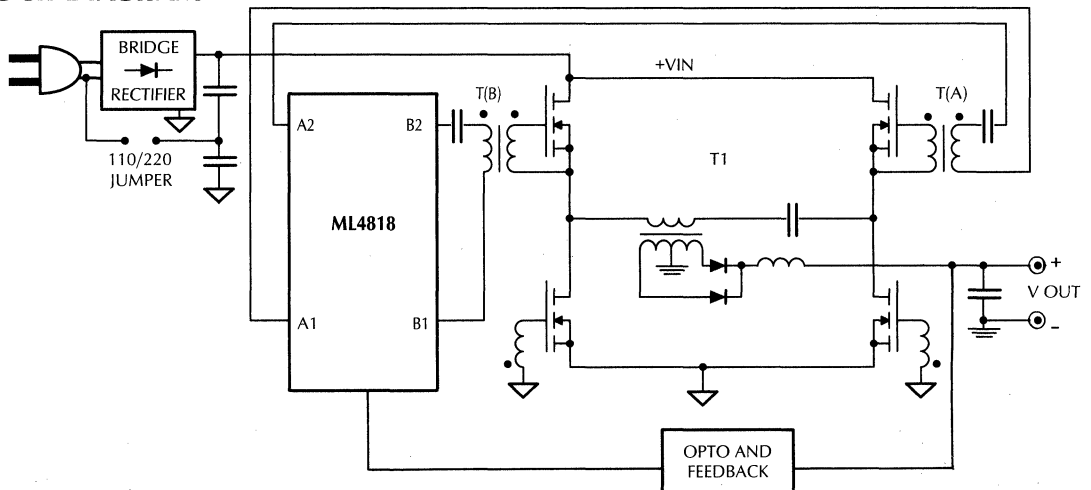
FEATURES

- Zero Voltage Switching (ZVS), high efficiency AC to DC power converter circuit
- High frequency operation resulting in smaller reactive components
- Lower conducted and radiated noise due to slower rise and fall times
- Switching and CV^2 losses greatly minimized
- Feed-forward compensated voltage mode control enables automatic compensation against instantaneous line changes
- Complete documentation and applications information

OPERATING SPECIFICATIONS

Input Voltage Range (switchable)	90 to 260V
Output Voltage	15V
Output Power	200W
Output Current	13A
Switching Frequency	250KHz.
Efficiency ($V_{IN}=120V$, $P_{OUT}=200W$)	85%

BLOCK DIAGRAM



Power Factor and PWM Controller "Combo"

GENERAL DESCRIPTION

The ML4819 is a complete boost mode Power Factor Control (PFC) which also contains a PWM controller. The PFC circuit is similar to the ML4812 while the PWM controller can be used for current or voltage mode control for a second stage converter. Since the PWM and PFC circuits share the same oscillator, synchronization of the two stages is inherent. The outputs of the controller IC provide high current (>1A peak) and high slew rate to quickly charge and discharge MOSFET gates. Special care has been taken in the design of the ML4819 to increase system noise immunity.

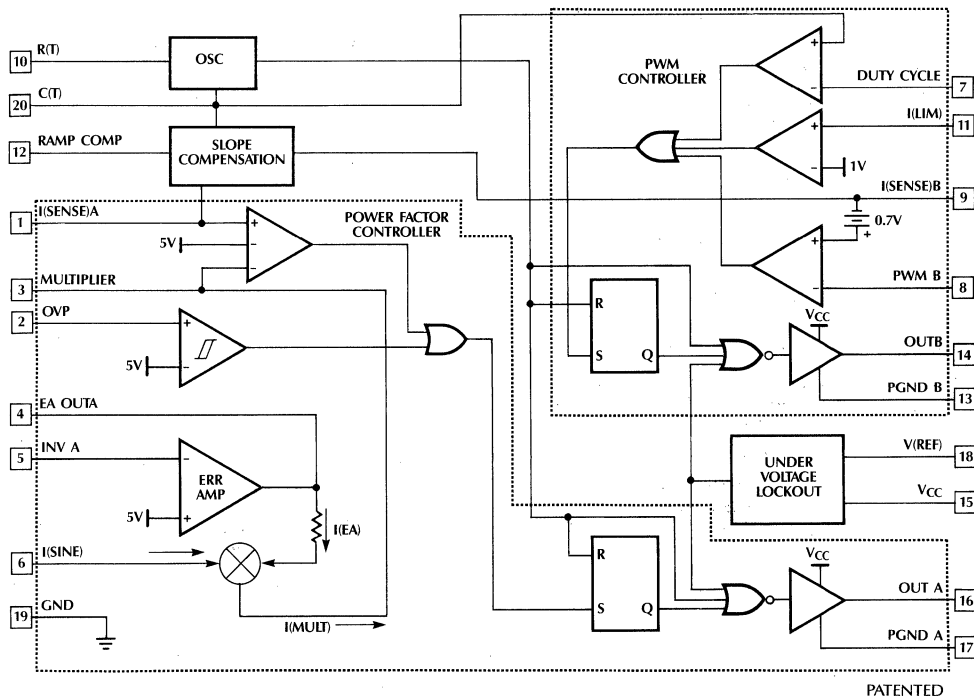
The PFC section is a peak current sensing control which uses a current sense transformer or SENSE FET to non-dissipatively sense switch current, giving the system improved overall efficiency over the average current sensing control method.

The PWM section includes cycle by cycle current limiting, precise duty cycle limiting for single ended converters, and slope compensation.

FEATURES

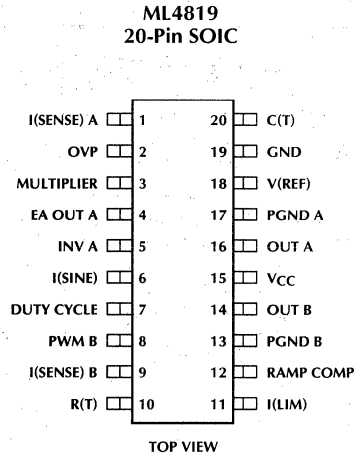
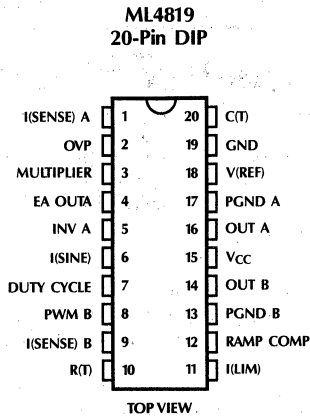
- Two 1A Peak Current Totem-Pole Output Drivers
- Precision buffered 5V Reference ($\pm 1\%$)
- Large oscillator amplitude for better noise immunity
- Precision duty cycle limit for PWM section
- Current input multiplier reduces external components and improves noise immunity
- Programmable Ramp Compensation circuit
- Over-Voltage comparator eliminates output "runaway" due to load removal
- Wide common mode range in current sense comparators for better noise immunity
- Under-Voltage Lockout circuit with 6V hysteresis

BLOCK DIAGRAM (Pin out shown is for DIP)



ML4819

PIN CONFIGURATIONS



PIN DESCRIPTION

PIN #	NAME	FUNCTION	PIN #	NAME	FUNCTION
1	I(SENSE) A	Input from the PFC Current Sense Transformer to the PWM comparator (+). Current Limit occurs when this point reaches 5V.	11	I(LIM)	Cycle by cycle PWM current limit. Exceeding 1V threshold on this pin terminates the PWM cycle.
2	OVP	Input to over voltage comparator.	12	RAMP COMP	Buffered output from the Oscillator Ramp (C(T)). A resistor to ground sets a current 1/2 of which is sourced on pins 9 and 11.
3	MULTIPLIER	Output of Current Multiplier. A resistor to ground on this pin converts the current to a voltage.	13	GND B	Return for the high current totem pole output of the PWM controller.
4	EA OUT A	Output of error amplifier.	14	OUT B	PWM controller totem pole output.
5	INV A	Inverting input to error amplifier.	15	V _{CC}	Positive Supply for the IC.
6	I(SINE)	Current Multiplier input.	16	OUT A	PFC controller totem pole output.
7	DUTY CYCLE	PWM controller duty cycle is limited by setting this pin to a fixed voltage.	17	GND A	Return for the high current totem pole output of the PFC controller.
8	PWM B	Error voltage feedback input.	18	V(REF)	Buffered output for the 5V voltage reference.
9	I(SENSE) B	Input for Current Sense resistor for current mode operation or for Oscillator ramp for voltage mode operation.	19	GND	Analog signal ground.
10	R(T)	Oscillator timing resistor pin. A 5V source across this resistor sets the charging current for C(T).	20	C(T)	Timing Capacitor for the Oscillator.

ABSOLUTE MAXIMUM RATINGS

Absolute maximum ratings are those values beyond which the device could be permanently damaged. Absolute maximum ratings are stress ratings only and functional device operation is not implied.

Supply Voltage (V_{CC})	35V
Output Current, Source or Sink (Pin 12)	
DC	1.0A
Output Energy (capacitive load per cycle)	5 μ J
Multiplier I(SINE) Input (Pin 6)	1.2mA
Error Amp Sink Current (Pin 3)	10mA
Oscillator Charge Current	2mA
Analog Inputs (Pins 1, 4, 5)	-0.3V to 5.5V

Junction Temperature	150°C
Storage Temperature Range	-65°C to +150°C
Lead Temperature (Soldering 10 sec)	+260°C
Thermal Resistance (θ_{JA})	
Plastic DIP or SOIC	65°C/W

OPERATING CONDITIONS

Temperature Range	
ML4819C	0°C to +70°C

ELECTRICAL CHARACTERISTICS

Unless otherwise specified, $R_T = 14K\Omega$, $C_T = 1000pF$, $T_A =$ Operating Temperature Range, $V_{CC} = 15V$ (note 2)

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
Oscillator					
Initial Accuracy	$T_J = 25^\circ C$	90	97	104	KHz
Voltage Stability	$12V < V_{CC} < 25V$		0.2		%
Temperature Stability			2		%
Total Variation	line, temp	88		106	KHz
Ramp Valley			0.9		V
Ramp Peak			4.3		V
R(T) Voltage		4.8	5.0	5.2	V
Discharge Current (pin 8 open)	$T_J = 25^\circ C, V_{PIN 16} = 2V$	7.5	8.4	9.3	mA
	$V_{PIN 16} = 2V$	7.2	8.4	9.5	mA
Duty Cycle Limit Comparator					
Input Offset Voltage		-15		15	mV
Input Bias Current			-2	-10	μA
Duty Cycle	$V_{PIN 7} = V_{REF/2}$	43	45	49	%
Reference Section					
Output Voltage	$T_J = 25^\circ C, I_O = 1mA$	4.95	5.00	5.05	V
Line Regulation	$12V < V_{CC} < 25V$		2	20	mV
Load Regulation	$1mA < I_O < 20mA$		8	25	mV
Temperature Stability			.4		%
Total Variation	line, load, temp	4.9		5.1	V
Output Noise Voltage	10Hz to 10KHz		50		μV
Long Term Stability	$T_J = 125^\circ C, 1000$ hrs, (note 1)		5	25	mV
Short Circuit Current	$V_{REF} = 0V$	-30	-85	-180	mA
Error Amplifier Section					
Input Offset Voltage		-15		15	mV
Input Bias Current			-0.1	-1.0	μA
Open Loop Gain	$1 < V_{PIN 4} < 5V$	60	75		dB
PSRR	$12V < V_{CC} < 25V$	60	90		dB
Output Sink Current	$V_{PIN 4} = 1.1V, V_{PIN 5} = 5.2V$	2	12		mA

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ELECTRICAL CHARACTERISTICS (Continued)

Unless otherwise specified, $R_T = 14K\Omega$, $C_T = 1000pF$, $T_A =$ Operating Temperature Range, $V_{CC} = 15V$ (note 2)

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
Error Amplifier Section (Continued)					
Output Source Current	$V_{PIN\ 4} = 5.0V$, $V_{PIN\ 5} = 4.8V$	-0.5	-1.0		mA
Output High Voltage	$I_{PIN\ 4} = -0.5mA$, $V_{PIN\ 5} = 4.8V$	6.5	7.0		V
Output Low Voltage	$I_{PIN\ 4} = 2mA$, $V_{PIN\ 5} = 5.2V$		0.7	1.0	V
Unity Gain Bandwidth			1.0		MHz
Multiplier					
I(SINE) Input Voltage	$I(SINE) = 500\mu A$.4	.7	.9	V
Output Current (pin 2)	$I(SINE) = 500\mu A$, Pin 5 = $V_{REF} - 20mV$	460	495	505	μA
	$I(SINE) = 500\mu A$, Pin 5 = $V_{REF} + 20mV$		0	10	μA
	$I(SINE) = 1mA$, Pin 5 = $V_{REF} - 20mV$	900	990	1005	μA
Bandwidth			200		KHz
PSRR	$12V < V_{CC} < 25V$		70		dB
Slope Compensation Circuit					
RAMP COMP Voltage (pin 12)			$V_{PIN\ 20} - 1$		V
I_{OUT} (pin 1 or pin 9)	$I_{PIN\ 12} = 100\mu A$ (note 3)	45	48	51	μA
OVP Comparator					
Input Offset Voltage	Output Off	-15		15	mV
Hysteresis	Output On	100	120	140	mV
Input Bias Current			-0.3	-3	μA
Propagation Delay			150		ns
I(SENSE) Comparators A and B					
Input Common Mode Range		-0.2		5.5	V
Input Offset Voltage	I(SENSE) A	-15		15	mV
	I(SENSE) B	+0.4	0.7	+0.9	V
Input Bias Current			-3	-10	μA
Input Offset Current		-3	0	+3	μA
Propagation Delay			150		ns
I_{LIMIT} (A) Trip Point	$V_{PIN\ 3} = 5.5V$	4.8	5	5.2	V
I(LIM) Comparator					
I_{LIMIT} Trip Point		.95	1.0	1.05	V
Input Bias Current			-2	-10	μA
Propagation Delay			150		ns
Output Section (A and B)					
Output Voltage Low	$I_{OUT} = -20mA$		0.1	0.4	V
	$I_{OUT} = -200mA$		1.6	2.2	V
Output Voltage High	$I_{OUT} = 20mA$	13	13.5		V
	$I_{OUT} = 200mA$	12	13.4		V
Output Voltage Low in UVLO	$I_{OUT} = -1mA$, $V_{CC} = 8V$		0.1	0.8	V
Output Rise/Fall Time	$C_L = 1000pF$		50		ns

ELECTRICAL CHARACTERISTICS (Continued)

Unless otherwise specified, $R_T = 14K\Omega$, $C_T = 1000pF$, $T_A =$ Operating Temperature Range, $V_{CC} = 15V$ (note 2)

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
Under Voltage Lockout					
Start-Up Threshold		15	16	17	V
Shut-Down Threshold		9	10	11	V
V_{REF} Good Threshold			4.4		V
Total Device					
Supply Current	Start-Up, $V_{CC} = 14V$.6	1.2	mA
	Operating, $T_j = 25^\circ C$		25	35	mA

Note 1: This parameter not 100% tested in production but guaranteed by design.

Note 2: V_{CC} is raised above the Start-up Threshold first to activate the IC, then returned to 15V.

Note 3: PWM comparator bias currents are subtracted from this reading.

FUNCTIONAL DESCRIPTION

OSCILLATOR

The ML4819 oscillator charges the external capacitor (C_T) with a current (I_{SET}) equal to $5/R_{SET}$. When the capacitor voltage reaches the upper threshold, the comparator changes state and the capacitor discharges to the lower threshold through Q1. While the capacitor is discharging, Q2 provides a high pulse.

The Oscillator period can be described by the following relationship:

$$T_{OSC} = T_{RAMP} + T_{DEADTIME}$$

where:

$$T_{RAMP} = \frac{C \text{ (Ramp Valley to Peak)}}{I_{SET}}$$

and:

$$T_{DEADTIME} = \frac{C \text{ (Ramp Valley to Peak)}}{(8.4mA - I_{SET})}$$

The maximum duty cycle of the PWM section can be limited by setting a threshold on pin 7. When the $C(T)$ ramp is above the threshold at pin 7, the PWM output is held off and the PWM flip-flop is set:

$$D_{LIMIT} \cong \frac{D_{OSC} \times (V_{PIN 7} - 0.9)}{3.4}$$

Where:

D_{LIMIT} = Desired duty cycle limit

D_{OSC} = Oscillator duty cycle

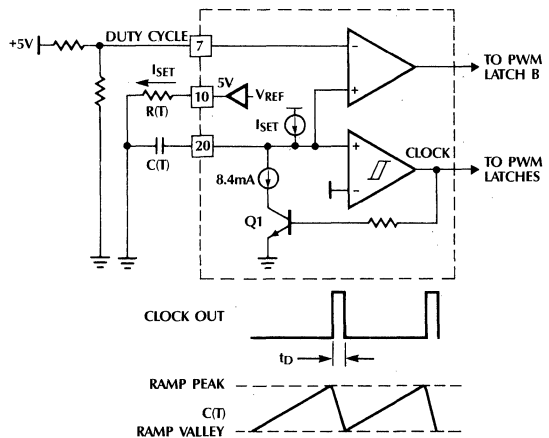


Figure 1. Oscillator Block Diagram

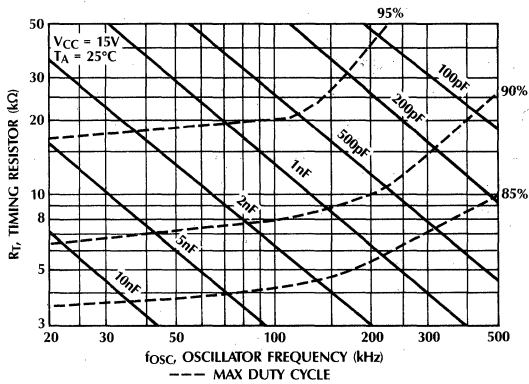


Figure 2. Oscillator Timing Resistance vs. Frequency

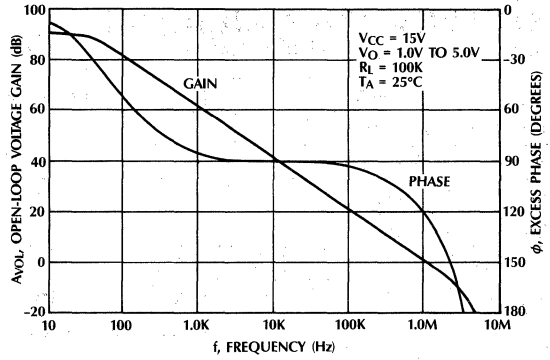


Figure 5. Error Amplifier Open-Loop Gain and Phase vs. Frequency

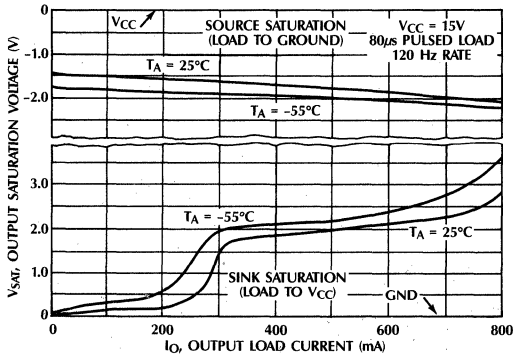


Figure 3. Output Saturation Voltage vs. Output Current

ERROR AMPLIFIER

The ML4819 error amplifier is a high open loop gain, wide bandwidth, amplifier.

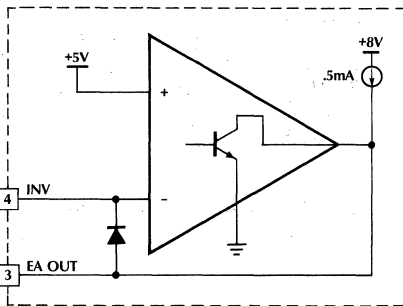


Figure 4. Error Amplifier Configuration

MULTIPLIER

The ML4819 multiplier is a linear current input multiplier to provide high immunity to the disturbances caused by high power switching. The rectified line input sine wave is converted to a current via a dropping resistor. In this way, small amounts of ground noise produce an insignificant effect on the reference to the PWM comparator.

The output of the multiplier is a current proportional to:

$$I_{OUT} \propto I(SINE) \times I(EA)$$

where I(SINE) is the current in the dropping resistor, and I(EA) is a factor which varies from 0 to 1 proportional to the output of the error amplifier. When the error amplifier is saturated high, the output of the multiplier is approximately equal to the I(SINE) input current.

The multiplier output current is converted into the reference voltage for the PWM comparator through a resistor to ground on the multiplier output (pin 3).

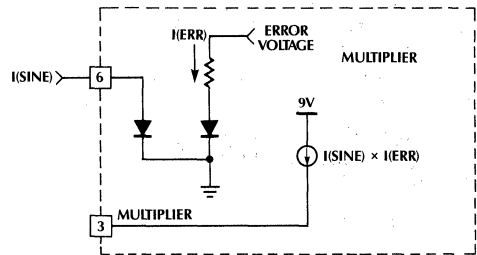


Figure 6. Multiplier Block Diagram

SLOPE COMPENSATION

Slope compensation is accomplished by adding 1/2 of the current flowing out of pin 12 to pin 1 (for the PFC section) and pin 9 (for the PWM section). The amount of slope compensation is equal to $(I_{PIN\ 12}/2) \times R_L$ where R_L is the impedance to GND on pin 1 or pin 9. Since most of the PWM applications will be limited to 50% duty cycle, slope compensation should not be needed for the PWM section. This can be defeated by using a low impedance load to the current sense on pin 9.

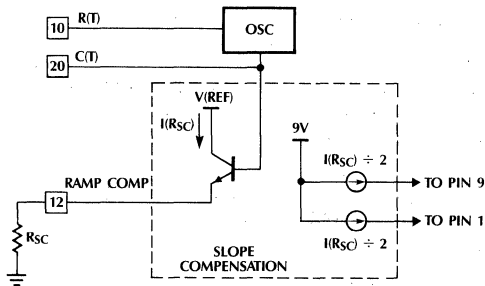


Figure 7. Slope Compensation Circuit

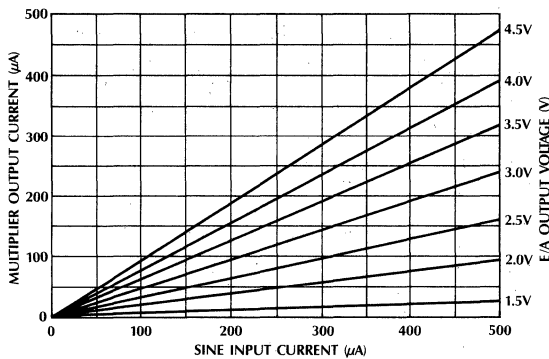


Figure 8. Multiplier Linearity

UNDER VOLTAGE LOCKOUT

On power-up the ML4819 remains in the UVLO condition; output low and quiescent current low. The IC becomes operational when V_{CC} reaches 16V. When V_{CC} drops below 10V, the UVLO condition is imposed. During the UVLO condition, the 5V V_{REF} pin is "off", making it usable as a "flag".

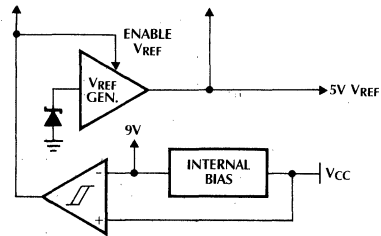


Figure 9. Under-Voltage Lockout Block Diagram

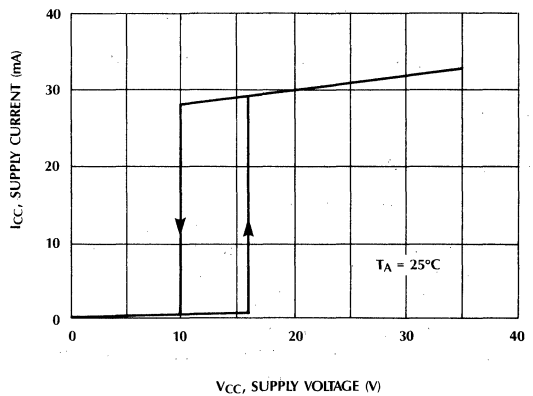


Figure 10a. Total Supply Current vs. Supply Voltage

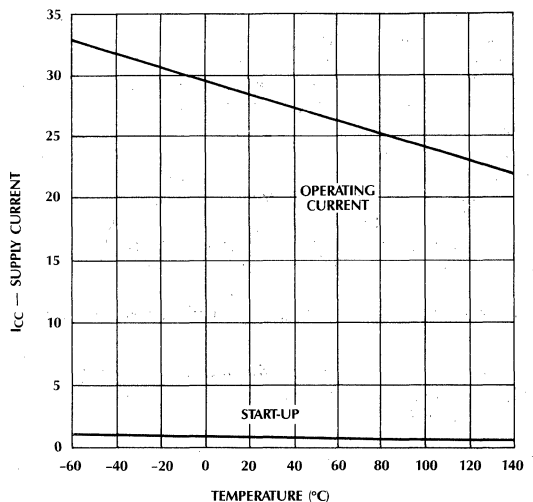


Figure 10b. Total Supply Current (I_{CC}) vs. Temperature

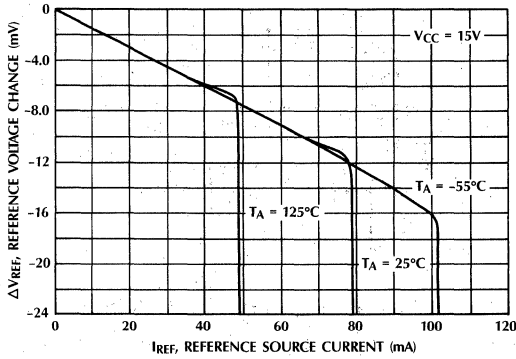


Figure 11. Reference Load Regulation

APPLICATIONS

POWER FACTOR SECTION

The power factor section in the ML4819 is similar to the power factor section in the ML4812 with the exception of the operation of the slope compensation circuit. Please refer to the ML4812 data sheet for more information.

The following calculations refer to figure 12. The component designers in the equations below refer to the following components in figure 12:

$$R_T = R16, C_T = C6.$$

INPUT INDUCTOR (L1) SELECTION

The central component in the regulator is the input boost inductor. The value of this inductor controls various critical operational aspects of the regulator. If the value is too low, the input current distortion will be high and will result in low power factor and increased noise at the input. This will require more input filtering. In addition, when the value of the inductor is low the inductor dries out (runs out of current) at low currents. Thus the power factor will decrease at lower power levels and/or higher line voltages. If the inductor value is too high, then for a given operating current the required size of the inductor core will be large and/or the required number of turns will be high. So a balance must be reached between distortion and core size.

One more condition where the inductor can dry out is analyzed below where it is shown to be maximum duty cycle dependent.

For the boost converter at steady state:

$$V_{OUT} = \frac{V_{IN}}{1 - D_{ON}} \quad (1)$$

Where D_{ON} is the duty cycle $[T_{ON}/(T_{ON} + T_{OFF})]$. The input boost inductor will dry out when the following condition is satisfied:

$$V_{IN}(t) < V_{OUT} \times (1 - D_{ON}) \quad (2)$$

or

$$V_{INDRY} = [1 - D_{ON}(\max)] \times V_{OUT} \quad (3)$$

V_{INDRY} : Voltage where the inductor dries out.
 V_{OUT} : Output dc voltage.

Effectively, the above relationship shows that the resetting volt-seconds are more than setting volt-seconds. In energy transfer terms this means that less energy is stored in the inductor during the ON time than it is asked to deliver during the OFF time. The net result is that the inductor dries out.

The recommended maximum duty cycle is 95% at 100KHz to allow time for the input inductor to dump its energy to the output capacitors.

For example:

$$\text{if: } \begin{aligned} V_{OUT} &= 380V \text{ and} \\ D_{ON}(\max) &= 0.95 \end{aligned}$$

then substituting in (3) yields $V_{INDRY} = 20V$. The effect of drying out is an increase in distortion at low voltages.

For a given output power, the instantaneous value of the input current is a function of the input sinusoidal voltage waveform, i.e. as the input voltage sweeps from zero volts to a maximum value equal to its peak so does the current.

The load of the power factor regulator is usually a switching power supply which is essentially a constant power load. As a result, an increase in the input voltage will be offset by a decrease in the input current.

By combining the ideas set forth above, some ground rules can be obtained for the selection and design of the input inductor:

Step 1: Find minimum operating current.

$$I_{IN}(\min)_{PEAK} = \frac{1.414 \times P_{IN}(\min)}{V_{IN}(\max)} \quad (4)$$

$$V_{IN}(\max) = 260V$$

$$P_{IN}(\min) = 50W$$

$$\text{then: } I_{IN}(\min)_{PEAK} = 0.272A$$

Step 2: Choose a minimum current at which point the inductor current will be on the verge of drying out. For this example 40% of the peak current found in step 1 was chosen.

$$\text{then: } I_{LDRY} = 100mA$$

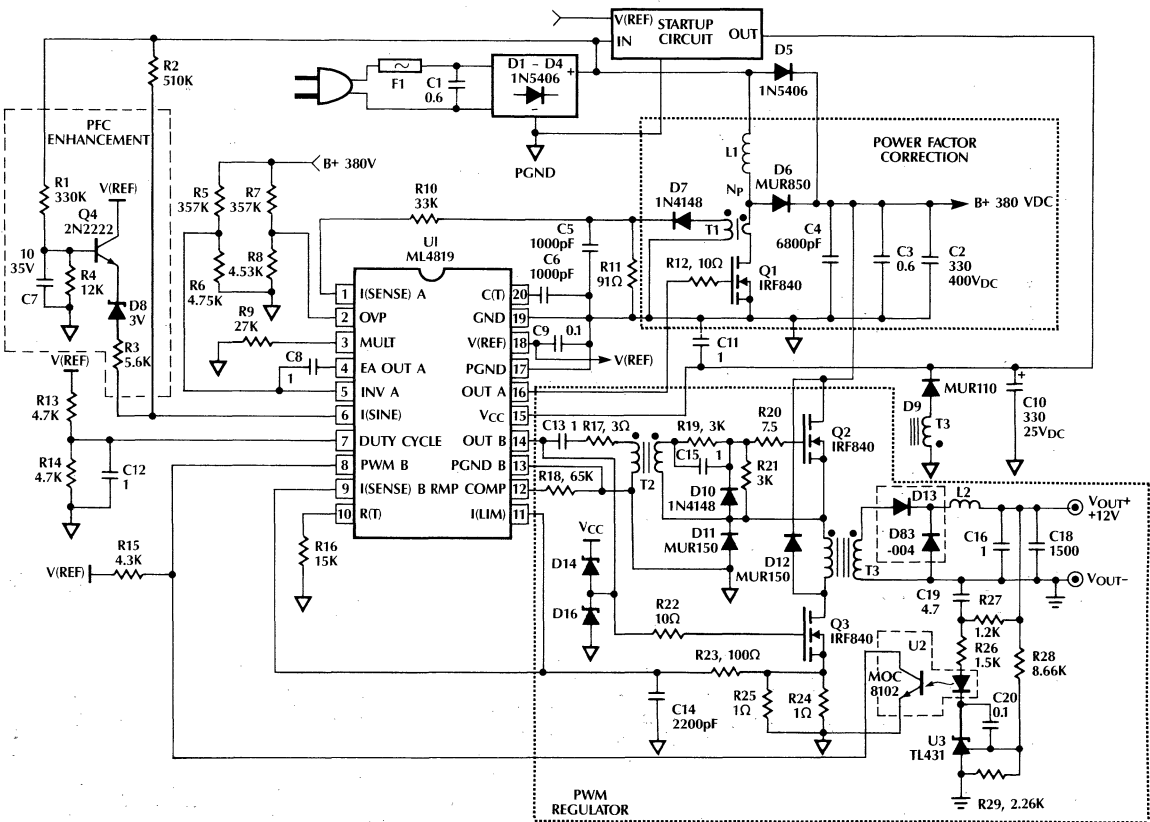


Figure 12. Typical Application, 180W Power Factor Corrected 12V Output Power Supply

Step 3: The value of the inductance can now be found using previously calculated data.

$$L1 = \frac{V_{INDRY} \times D_{ON(max)}}{I_{LDRY} \times f_{OSC}} \quad (5)$$

$$= \frac{20V \times 0.95}{100mA \times 100KHz} = 2mH$$

The inductor can be allowed to decrease in value when the current sweeps from minimum to maximum value. This allows the use of smaller core sizes. The only requirement is that the ramp compensation must be adequate for the lower inductance value of the core so that there is adequate compensation at high current.

Step 4: The presence of the ramp compensation will change the dry out point, but the value found above can be considered a good starting point. Based on the amount of power factor correction the above value of L1 can be optimized after a few iterations.

Gapped Ferrites, Molypermalloy, and Powdered Iron cores are typical choices for core material. The core material selected should have a high saturation point and acceptable losses at the operating frequency.

One ferrite core that is suitable at around 200W is the #4229PL00-3C8 made by Ferroxcube. This ungapped core will require a total gap of 0.180" for this application.

OSCILLATOR COMPONENT SELECTION

The oscillator timing components can be calculated by using the following expression:

$$f_{OSC} = \frac{1.36}{R_T \times C_T} \quad (6)$$

For example:

Step 1: At 100KHz with 95% duty cycle $T_{OFF} = 500ns$ calculate C_T using the following formula:

$$C_T = \frac{T_{OFF} \times I_{DIS}}{V_{OSC}} = 1000pF \quad (7)$$

Step 2: Calculate the required value of the timing resistor.

$$R_T = \frac{1.36}{f_{OSC} \times C_T} = \frac{1.36}{100KHz \times 1000pF} \quad (8)$$

$$= 13.6K\Omega \text{ choose } R_T = 14K\Omega.$$

CURRENT SENSE AND SLOPE (RAMP) COMPENSATION COMPONENT SELECTION

Slope compensation in the ML4819 is provided internally. A current equal to $V_{CTP}/2(R18)$ is added to $I(SENSE)$ A (pin 1). This is converted to a voltage by R10, adding slope to the sensed current through T1. The amount of slope compensation should be at least 50% of the downslope of the inductor current during the off

time as reflected on pin 1. Note that slope compensation is a requirement only if the inductor current is continuous and the duty cycle is more than 50%. The highest inductor downslope is found at the point of inductor discontinuity:

$$\frac{di_L}{dt} = \frac{V_B - V_{IN DRY}}{L} = \frac{380V - 20V}{2mH} \quad (9)$$

$$= 0.18 A/\mu s$$

The downslope as reflected to the input of the PWM comparator is given by:

$$S_{PWM} = \frac{V_B - V_{IN DRY}}{L1} \times \frac{R_{T1}}{N_C} \quad (10)$$

Where N_C is the turns ratio of the current transformer (T1) used. In general, current transformers simplify the sensing of switch currents especially at high power levels where the use of sense resistors is complicated by the amount of power they have to dissipate. Normally the primary side of the transformer consists of a single turn and the secondary consists of several turns of either enameled magnet wire or insulated wire. The diameter of the ferrite core used in this example is 0.5" (SPANG/Magnetics F41206-TC). The rectifying diode at the output of the current transformer can be a 1N4148 for secondary currents up to 75mA average.

Sense FETs or resistive sensing can also be used to sense the switch current, the sensed signal has to be amplified to the proper level before it is applied to the ML4819.

The value of the ramp compensation (SC_{PWM}) as seen at pin 1 is:

$$SC_{PWM} = \frac{2.5 \times R_9}{R_{16} \times C_6 \times R_{18}} \quad (11)$$

The required value for R_{18} can therefore be found by equating:

$$SC_{PWM} = A_{SC} \times S_{PWM}$$

where A_{SC} is the amount of slope compensation and solving for R_{18} .

The value of R_9 (pin 2) depends on the selection of R_2 (pin 6)

$$R_2 = \frac{V_{IN(max)PEAK}}{I_{SINE(peak)}} = \frac{260 \times 1.414}{0.72mA} = 510K \quad (12)$$

$$R_9 > \frac{V_{CLAMP} \times R_2}{V_{IN(min)PEAK}} = \frac{4.8 \times 510K}{80 \times 1.414} \cong 22K \quad (13)$$

Choose $R_9 = 27K$

The peak of the inductor current can be found approximately by:

$$I_{LPEAK} = \frac{1.414 \times P_{OUT}}{V_{IN(min)RMS}} = \frac{1.414 \times 200}{90} = 3.14A \quad (14)$$

Selection of N_C which depends on the maximum switch current, assume 4A for this example is 80 turns.

$$R_{11} = \frac{V_{CLAMP} \times N_C}{I_{LPEAK}} = \frac{4.8 \times 80}{4} \cong 100\Omega \quad (15)$$

Where R_{11} is the sense resistor, and V_{CLAMP} is the current clamp at the inverting input of the PWM comparator. This clamp is internally set to 5V. In actual application it is a good idea to assume a value less than 5V to avoid unwanted current limiting action due to component tolerances. In this application V_{CLAMP} was chosen as 4.8V.

Having calculated R_{11} the value S_{PWM} and of R_{18} can now be calculated:

$$S_{PWM} = \frac{380V - 20}{2mH} \times \frac{100}{80} = 0.225V/\mu s$$

$$R_{18} = \frac{2.5 \times R_9}{A_{SC} \times S_{PWM} \times R_T \times C_T} \quad (16)$$

$$R_{18} = \frac{2.5 \times 27K}{0.7 \times (0.225 \times 10^6) \times 14K \times 1nF} \cong 30K$$

Choose $R_{18} = 33K$

The following values were used in the calculation:

$$R_9 = 27K \quad A_{SC} = 0.7$$

$$R_T = 14K \quad C_T = 1nF$$

VOLTAGE REGULATION COMPONENTS

The values of the voltage regulation loop components are calculated based on the operating output voltage. Note that voltage safety regulations require the use of sense resistors that have adequate voltage rating. As a rule of thumb if 1/4W resistors are available, two of them should be used in series. The input bias current of the error amplifier is approximately 0.5μA, therefore the current available from the voltage sense resistors should be significantly higher than this value. Since two 1/4W resistors have to be used the total power rating is 1/2W. The operating power is set to be 0.4W then with 380V output voltage the value can be calculated as follows:

$$R_5 = (380V)^2/0.4W = 360K \quad (17)$$

Choose two 178K, 1% connected in series.

Then R_6 can be calculated using the formula below:

$$R_6 = \frac{V_{REF} \times R_5}{V_B - V_{REF}} = \frac{5V \times 356K}{380V - 5V} = 4.747K \quad (18)$$

Choose 4.75K, 1%. One more critical component in the voltage regulation loop is the feedback capacitor for the error amplifier. The voltage loop bandwidth should be set such that it rejects the 120Hz ripple which is present at the output. If this ripple is not adequately attenuated it will cause distortion on the input current waveform. Typical bandwidths range anywhere from a few Hertz to 15Hz. The main compromise is between transient response and distortion. The feedback capacitor can be calculated using the following formula:

$$C_B = \frac{1}{3.142 \times R_5 \times BW} \quad (19)$$

$$C_B = \frac{1}{3.142 \times 356K \times 2Hz} = 0.44\mu F$$

OVERVOLTAGE PROTECTION (OVP) COMPONENTS

The OVP loop should be set so that there is no interaction with the voltage control loop. Typically it should be set to a level where the power components are safe to operate. Ten to fifteen volts above V_{OUT} seems to be adequate. This sets the maximum transient output voltage to about 395V.

By choosing the high voltage side resistor of the OVP circuit the same way as above i.e. $R_7 = 356K$ then R_8 can be calculated as:

$$R_8 = \frac{V_{REF} \times R_7}{V_{OVP} - V_{REF}} = \frac{5V \times 356K}{395V - 5V} = 4.564K \quad (20)$$

Choose 4.53K, 1%.

Note that R_5 , R_6 , R_7 and R_8 should be tight tolerance resistors such as 1% or better.

OFF-LINE START-UP AND BIAS SUPPLY GENERATION

The Start-Up Circuit in figure 12 can be either a “bleed resistor” (39KΩ, 2W) or the circuit shown in figure 13. The bleed resistor method offers the advantage of simplicity and lowest cost, but may yield excessive turn-on delay at low line.

When the voltage on pin 15 (V_{CC}) exceeds 16V, the IC starts up. The energy stored on the C10 supplies the IC with running power until the supplemental winding on T3 can provide the power to sustain operation.

6

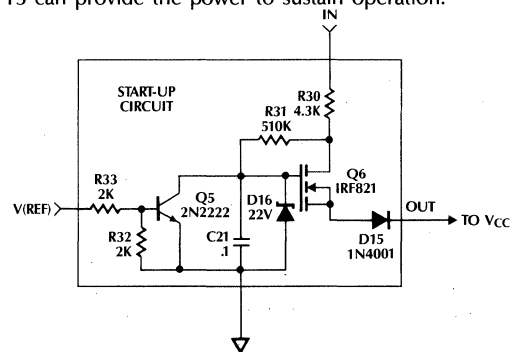


Figure 13. Start-Up Circuit

ENHANCEMENT CIRCUIT

The theory of operation of the power factor enhancement circuit (inside the dotted lines) in Figure 11 is described in APPLICATION NOTE 11 in detail. It improves the power factor and lowers the input current harmonics. Note that the circuit meets the proposed IEC 555 specifications (with the enhancement) on the harmonics with a large margin while correcting the input power factor to better than 0.99 under most steady state operating conditions.

PWM SECTION

The PWM section in figure 12 is a two switch forward converter, shown in figure 14 below for clarity. This fully clamped circuit eliminates the need for very high voltage MOSFETs. Flyback topology is also possible with the ML4819.

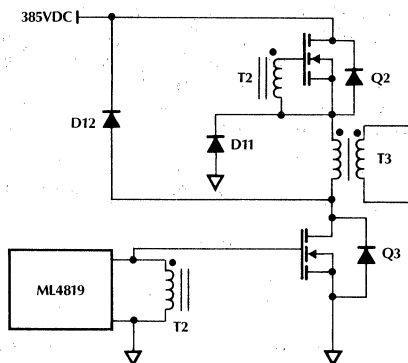


Figure 14. Two-Switch Forward Converter

This regulator (figure 12) uses current mode control. Current is sensed through R24 and filtered for high frequency noise and leading edge transient through R23 and C14. The main regulation loop is through PWM B. The TL431 (U3) in the secondary serves as both the voltage reference and error amplifier, with isolation provided by an opto coupler (U2) providing a current command signal on pin 8. Loop compensation is provided by R29 and C20. The output voltage is set by:

$$V_{OUT} = 2.5 \left(1 + \frac{R_{29}}{R_{28}} \right) \quad (21)$$

The control loop is compensated using standard compensation techniques.

Current is limited to a threshold of 2A (1V on R24). The duty cycle is limited in this circuit to below 50% to prevent transformer (T3) core saturation. The maximum duty cycle limit of 45% is set using a threshold of $V_{REF}/2$ on pin 7.

The circuit in figure 12 can be modified for voltage mode operation by utilizing the slope current which appears on pin 9 as shown in figure 15 below.

The ramp amplitude appearing on pin 9 will be

$$V_R = \frac{I_{R18}}{2} \times R(V) \quad (22)$$

where R_{18} is the slope compensation resistor. Since this circuit operates with a constant input voltage (as supplied by the PFC section) voltage feed-forward is unnecessary.

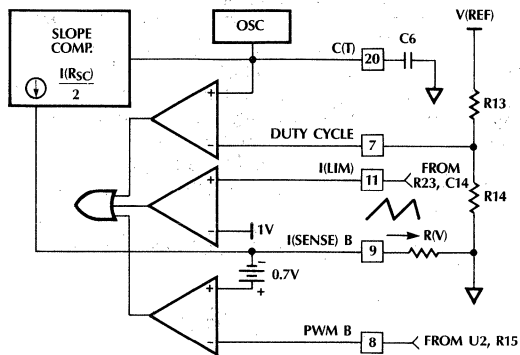


Figure 15. Voltage Mode Configuration

CONSTRUCTION AND LAYOUT TIPS

High frequency power circuits require special care during breadboard construction and layout. Double sided printed circuit boards with ground plane on one side are highly recommended. All critical switching leads (power FET, output diode, IC output and ground leads, bypass capacitors) should be kept as small as possible. This is to minimize both the transmission and pick-up of switching noise.

There are two kinds of noise coupling; inductive and capacitive. As the name implies inductive coupling is due to fast changing (high di/dt) circulating switching currents. The main source is the loop formed by Q1, D6, and C3-C4. Therefore this loop should be as small as possible, and the above capacitors should be good high frequency types.

The second form of noise coupling is due to fast changing voltages (high dv/dt). The main source in this case is the drain of the power FET. The radiated noise in this case can be minimized by insulating the drain of the FET from the heatsink and then tying the heatsink to the source of the FET with a high frequency capacitor.

The IC has two ground pins named PWR GND and Signal GND. These two pins should be connected together with a very short lead at the printed circuit board exit point. In general grounding is very important and ground loops should be avoided. Star grounding schemes are preferred.

Component Values/Bill of Materials for Figure 12

Component	Description
C1, C3	0.6 μ F, 630V Film (250 VAC)
C2	330 μ F, 400V Electrolytic
C4	6800pF, 1KV Ceramic
C5, C6	1000pF
C7	10 μ F, 35V
C8, C11, C13, C15, C16	1 μ F, Ceramic
C9, C20, C21	0.1 μ F, Ceramic
C10	1500 μ F, 25V Electrolytic
C12, C17	1 μ F, Ceramic
C14	2200pF
C18	1500 μ F, 16V Electrolytic
C19	4.7 μ F
D1–D5	1N5406
D6	MUR850
D7, D10	1N4148
D8	3V Zener diode or 4 \times 1N4148 in series
D9	MUR110
D11, D12	MUR150
D13	D83-004K
D15	1N4001
D16, D14	1N5818 or 1N5819
F1	5A, 250V, 3AG
L1	2mH, 4A I _{PEAK} Core: Ferroxcube 4229-3C8 150 Turns #24 AWG 0.150" gap
L2	10 μ H Core: Spang OF 43019 UG00 8 Turns #15AWG gap 0.05"
Q1–Q3	IRF840
Q4, Q5	2N2222
Q6	IRF821
R1	330K
R2, R31	510K
R3	5.6K

Component	Description
R4	12K
R5, R7	357K, 1%
R6	4.75K, 1%
R8	4.53K, 1%
R9	27K
R10, R18	33K
R11	91 Ω
R12, R22	10 Ω
R13, R14	4.7K
R15	4.3K
R16	15K
R17	3 Ω
R20	7.5 Ω
R21, R19	3K
R23	100 Ω
R24, R25	1 Ω
R26	1.5K
R27	1.2K
R28	8.66K, 1%
R29	2.26K, 1%
R30	2K, 1W
R32, R33	2K
T1	Spang F41206-TC or Siemens B64290-K45-X27 or X830 or Ferroxcube 768T188-3C8 N _S = 80, N _P = 1
T2	Same core as T1 N _S = N _P = 15 bifilar
T3	Core: Ferroxcube 4229-3C8 Pri. 44 Turns #18 Litz wire Sec. 4 Turns of copper strip Aux. 2 Turns #24 AWG
U2	MOC8102
U3	TL431

ML4819

ORDERING INFORMATION

PART NUMBER	TEMPERATURE RANGE	PACKAGE
ML4819CP	0°C to +70°C	Molded DIP (P20)
ML4819CS	0°C to +70°C	Molded PCC (S20W)

Power Factor Controller

GENERAL DESCRIPTION

The ML4821 provides the complete control for a "boost" type power factor correction system using the average current sensing method. Special care has been taken in the design of the ML4821 to increase system noise immunity. The circuit includes a precision reference, multiplier, average current error amplifier, output error amplifier, over-voltage protection comparator, shutdown logic, as well as a high current output. In addition, start-up is simplified by an under-voltage lockout circuit with 7V hysteresis.

In a typical application, the ML4821 controls the average current, adjusting the pulse width of the output to modulate the current so that its shape conforms to the shape of the input voltage. The reference for the current regulator is a product of the sinusoidal line voltage times the output of the error amplifier which is regulating the output DC voltage. Average line voltage compensation is provided in the multiplier to ensure constant loop gain over a wide input voltage range. This compensation includes a special "brown-out" control which reduces output power below 90V RMS input.

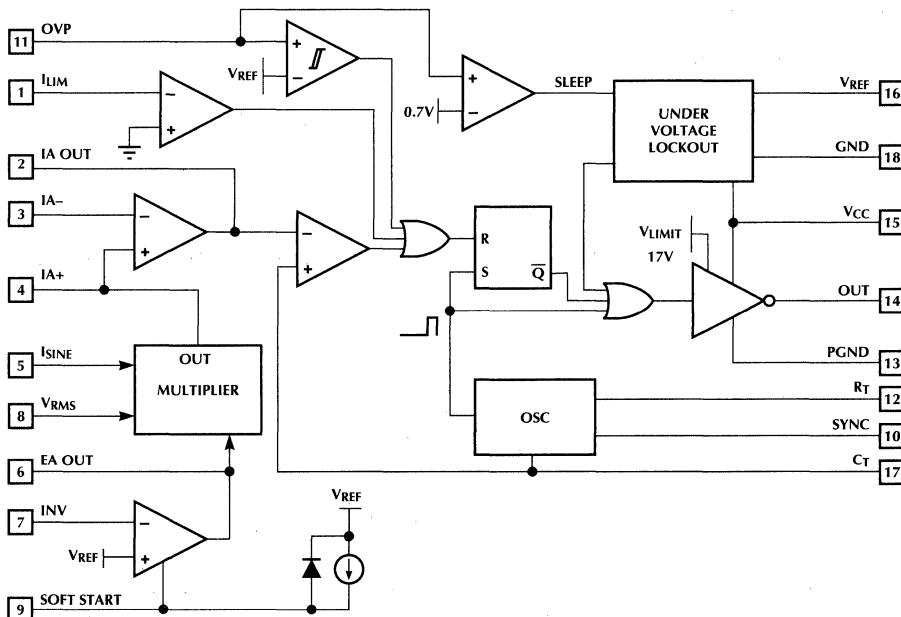
For applications information, Please see Applications Note 16.

The ML4821 uses Micro Linears bipolar array technology which allows for customization of the IC for a user's specific application. Please consult Micro Linear for semi-standard options.

FEATURES

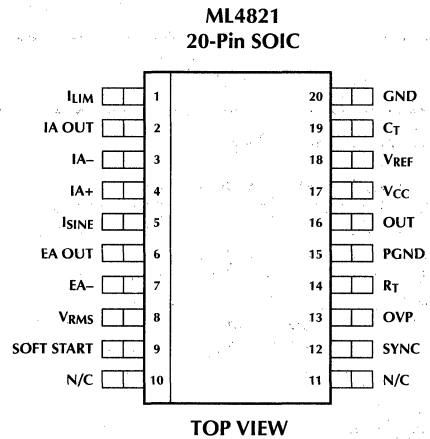
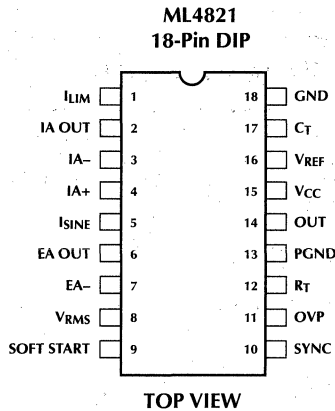
- Average current sensing for lowest possible harmonic distortion
- Average Line compensation with Brown-out control
- Precision buffered 5V Reference ($\pm 1\%$)
- Current Input Multiplier reduces external components and improves noise immunity
- 1A Peak Current Totem-Pole Output Drive
- Over-Voltage comparator eliminates output "runaway" due to load removal
- Wide common mode range in current sense comparators for better noise immunity
- Large oscillator amplitude for better noise immunity
- Output Driver internally limited to 17V
- "Sleep mode" shutdown input

BLOCK DIAGRAM



ML4821

PIN CONNECTION



PIN DESCRIPTION

PIN NO.	NAME	FUNCTION
1	I_{LIM}	Peak cycle-by-cycle Current limit input.
2	IA OUT	Output and compensation node of the average current error amplifier.
3	IA-	Inverting input of the average current error amplifier.
4	IA+	Non-Inverting input of the average current error amplifier and output of the multiplier.
5	I_{SINE}	Current Multiplier input.
6	EA OUT	Output of output voltage error amplifier.
7	INV	Inverting input to error amplifier.
8	V_{RMS}	Input for Average Line Voltage compensation
9	SOFT START	Normally connected to Soft Start Capacitor

PIN NO.	NAME	FUNCTION
10	SYNC	Oscillator synchronization input
11	OVP	Inhibits output pulses when the voltage at this pin exceeds 5V. Also, when the voltage at this pin is less than 0.7V, the IC goes into low current shut-down mode.
12	R_T	Timing Resistor for the Oscillator
13	PWR GND	Return for the High Current Totem pole output.
14	OUT	High Current Totem pole output.
15	V_{CC}	Positive Supply for the IC.
16	V_{REF}	Buffered output for the 5V voltage reference.
17	C_T	Timing Capacitor for the Oscillator.
18	GND	Analog signal ground.

ABSOLUTE MAXIMUM RATINGS

Supply Current (I_{CC})	35mA
Output Current, Source or Sink (pin 14) DC	1.0A
Output Energy (capacitive load per cycle)	5 μ J
Multiplier I_{SINE} Input (pin 5)	1.2mA
Error Amp Source Current (pin 6)	50mA
Oscillator Charge Current	2mA
Analog Inputs (pins 1,3,4, 7,8,9,10,11)	-0.3V to 5.5V
Junction Temperature	150°C
Storage Temperature Range	-65°C to +150°C
Lead Temperature (Soldering 10 sec.)	+260°C
Thermal Resistance (θ_{JA})	
Plastic DIP	65°C/W
Plastic SOIC	65°C/W

OPERATING CONDITIONS

Temperature Range	
ML4821C	0°C to +70°C

Absolute maximum ratings are those values beyond which the device could be permanently damaged. Absolute maximum ratings are stress ratings only and functional device operation is not implied.

ELECTRICAL CHARACTERISTICS

Unless otherwise specified, $R_T = 6.2K\Omega$, $C_T = 720pF$, $T_A =$ Operating Temperature Range, $V_{CC} = 15V$ (Note 2).

PARAMETER	CONDITIONS	MIN	TYP.	MAX	UNITS
OSCILLATOR					
Initial accuracy	$T_A = 25^\circ C$	90	100	110	KHz
Voltage stability	$12V < V_{CC} < 18V$		1		%
Temperature stability			2		%
Total Variation	Line, Temperature	85		115	KHz
Ramp Valley to Peak		4.7	5.2	5.6	V
R_T Voltage		4.8	5.0	5.2	V
Discharge Current (pin 12 open)	$V_{PIN17} = 2V$	7.8	8.4	9.3	mA
Sync Input Threshold		1.5	2.0	2.5	V
REFERENCE SECTION					
Output Voltage	$T_A = 25^\circ C, I_O = 1mA$	4.95	5.00	5.05	V
Line regulation	$12V < V_{CC} < 24V$		2	10	mV
Load regulation	$1mA < I_O < 20mA$		2	15	mV
Temperature stability			.4		%
Total Variation	line, load, temp	4.9		5.1	V
Output Noise Voltage	10Hz to 10KHz		50		μV
Long Term Stability	$T_A = 125^\circ C, 1000$ hrs, (Note 1)		5	.25	mV
Short Circuit Current	$V_{REF} = 0V$	-30	-85	-180	mA
VOLTAGE ERROR AMPLIFIER (EA)					
Input Offset Voltage		0		-15	mV
Input Bias Current			-50	-800	nA
Open Loop Gain	$2 < V_{PIN6} < 6V$	60	75		dB
PSRR	$12V < V_{CC} < 24V$	70	100		dB
Output Sink Current	$V_{PIN6} = 4V, V_{PIN7} = 5.5V$	300	500		μA
Output Source Current	$V_{PIN6} = 4.0V, V_{PIN7} = 4.8V$	-10	-30		mA

ML4821

ELECTRICAL CHARACTERISTICS (Continued)

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
VOLTAGE ERROR AMPLIFIER (EA) (Continued)					
Output High Voltage	$I_{PIN6} = -5mA, V_{PIN7} = 4.8V$	7.0	7.5		V
Output Low Voltage	$I_{PIN6} = 0, V_{PIN7} = 5.5V$		0	0.5	V
Unity Gain Bandwidth			1.0		MHz
Soft Start Charge Current	$V_{PIN9} = 3V$	-25	-38	-55	μA
CURRENT AMPLIFIER (IA)					
Input Offset Voltage		-1.5		+2.5	mV
Input Bias Current			-0.15	-1	μA
Input Offset Current				+400	nA
Open Loop Gain	$2 < V_{PIN6} < 7V$	80	100		dB
PSRR	$12V < V_{CC} < 24V$	65	85		dB
Output Voltage Low	$I_{OL} = 300\mu A$		0	0.5	V
Output Voltage High	$I_{OH} = -10mA$	7.0	7.5		V
Input Common Mode Range		-0.3		2.5	V
MULTIPLIER					
Gain	$V_{PIN7} = 4.8V, V_{PIN8} = 0V$	0.9	1.2	1.3	
	$V_{PIN7} = 4.8V, V_{PIN8} = 1.75V$	3.25	3.88	4.40	
	$V_{PIN7} = 4.8V, V_{PIN8} = 2.6V$	1.35	1.75	2.15	
	$V_{PIN7} = 4.8V, V_{PIN8} = 5.2V$	0.25	0.38	0.50	
Output Current	$V_{PIN7} = 5.2V, V_{PIN8} = 5.2V$		-2	-4	μA
Output Current Limit	$V_{PIN7} = 4.8V, I_{PIN5} = 500mA, V_{PIN8} = 1.75V$	370	395	410	μA
I_{LIM} COMPARATOR					
Input Offset Voltage				+15	mV
Input Bias Current			-100	-200	μA
OVP COMPARATOR					
Input Offset Voltage	Output Off	-25		5	mV
Hysteresis	Output On	85	105	130	mV
Input Bias Current			-0.3	-3	μA
Propagation Delay			150		ns
Shutdown Threshold		0.4	0.7	1.0	V
PWM COMPARATOR: I_{SENSE}					
Input Common Mode Range		0		8	V
Propagation Delay			150		ns

ELECTRICAL CHARACTERISTICS (Continued)

PARAMETER	CONDITIONS	MIN	TYP.	MAX	UNITS
OUTPUT SECTION					
Output Voltage Low	$I_{OUT} = 20\text{mA}$		0.1	0.4	V
	$I_{OUT} = 200\text{mA}$		1.6	2.2	V
Output Voltage High	$I_{OUT} = -20\text{mA}$	13	13.5		V
	$I_{OUT} = -200\text{mA}$	12	13.4		V
Output Voltage Low in UVLO	$I_{OUT} = -5\text{mA}, V_{CC} = 8\text{V}$		0.1	0.8	V
Output Rise/Fall Time	$C_L = 1000\text{pF}$		50		ns
UNDER-VOLTAGE LOCKOUT					
Start-up Threshold		14.5		16.5	V
Shut-Down Threshold		8.5		10.5	V
V_{REF} Good Threshold			4.4		V
TOTAL DEVICE					
Supply Current	Start-up, $V_{CC} = 14\text{V}, T_A = 25^\circ\text{C}$ Operating, $T_A = 25^\circ\text{C}$		0.6	1.2	mA
			26	32	mA
Internal Shunt Zener Voltage	$I_{CC} = 35\text{mA}$	25	27	35	V

Note 1: This parameter not 100% tested in production but guaranteed by design.

Note 2: V_{CC} is raised above the Start-up Threshold first to activate the IC, then returned to 15V

Note 3: Multiplier gain is defined as: $\frac{I_{OUT\text{PIN4}}}{I_{NPIN6}}$

ML4821

FUNCTIONAL DESCRIPTION

OSCILLATOR

The ML4821 oscillator charges the external capacitor (C_T) with a current equal to $2.5/R_T$. When the capacitor voltage reaches the upper threshold, the comparator changes state and the capacitor discharges to the lower threshold through $Q1$.

The Oscillator period can be described by the following relationship:

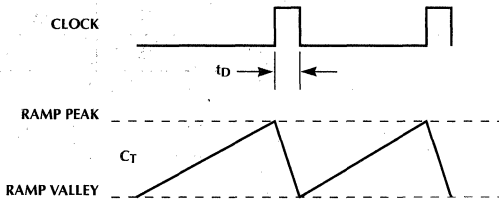
$$T_{OSC} = T_{RAMP} + T_{DEADTIME}$$

where:

$$T_{RAMP} = C(\text{Ramp Valley to Peak}) \div (I_{RT}/2)$$

and:

$$T_{DEADTIME} = C(\text{Ramp Valley to Pk}) \div (8.4\text{mA} - I_{RT}/2)$$



The ML4821 oscillator includes a SYNC input for synchronizing to an external frequency source. A positive pulse on this pin of 2V (typ) resets the oscillators comparator and initiates a discharge cycle for C_T . The R_T and C_T component values which set the ML4821 oscillator frequency should be selected to produce a lower frequency than the external frequency source.

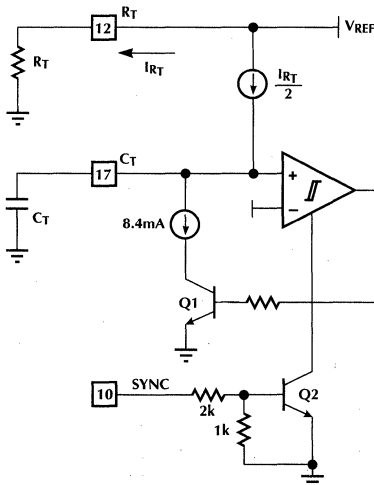


Figure 1. Oscillator Block Diagram.

ERROR AND CURRENT AMPLIFIERS

The ML4821 error amplifier is a high open loop gain, wide bandwidth, amplifier with a class A output. The soft start circuit controls the input to the error amplifier for closed loop soft start operation.

The current amplifier (IA) is similar to the error amplifier but is designed for very low offsets to allow the selection of a low value resistor for R_{SENSE} .

OUTPUT DRIVER STAGE

The ML4821 Output Driver is a 1A peak output high speed totem pole circuit designed to quickly drive capacitive loads, such as power MOSFET gates. The driver circuit's output voltage is internally limited to 17V.

MULTIPLIER

The ML4821 multiplier is a linear current input multiplier which provides high immunity to the disturbances caused by high power switching. The rectified line input sine wave is converted to a current via a dropping resistor. In this way, small amounts of ground noise produce an insignificant effect on the reference to the PWM comparator.

The output of the multiplier is a current which appears on pin 4 to form the reference for the current error amplifier and is given as:

$$I_{MUL} = K \times I_{SINE} \times (V_{EA} - 0.8)$$

where:

I_{SINE} is the current in the dropping resistor, V_{EA} is the output of the error amplifier and K is a constant determined by the $V(RMS)$ input on pin 8. K assumes a higher value for the range from 90 to 170V than in the range above 170V.

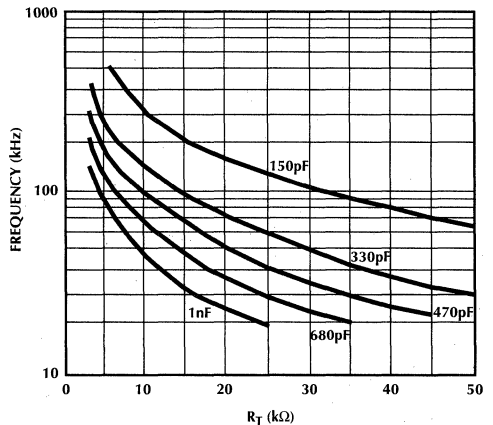


Figure 2. Oscillator Timing Resistance vs. Frequency.

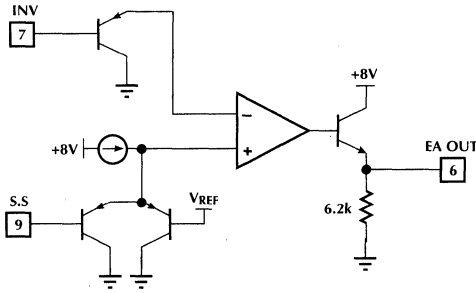


Figure 3. Error and Current Amplifier Configuration

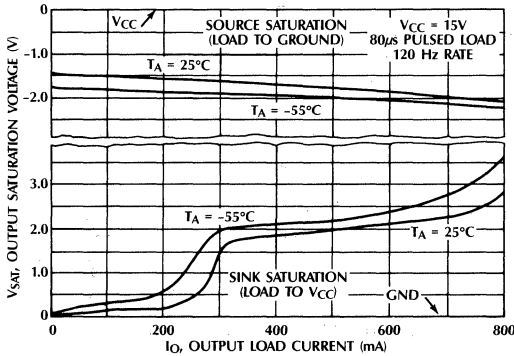


Figure 5. Output Saturation Voltage vs. Output Current.

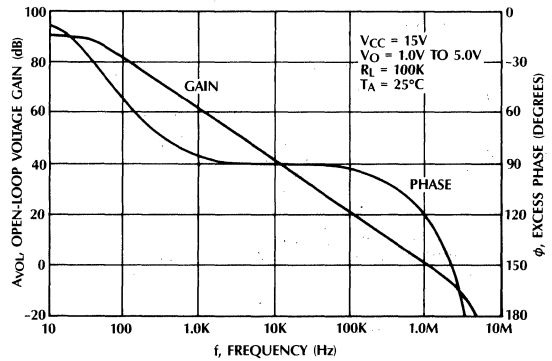


Figure 4. Error Amplifier Open-loop Gain and Phase vs. Frequency.

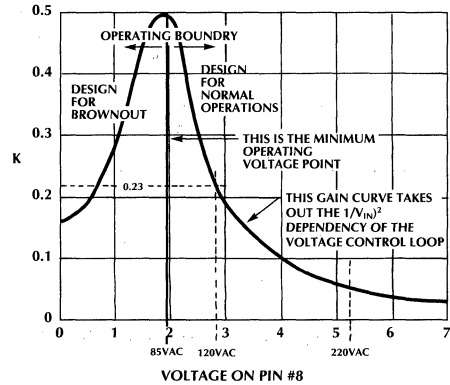


Figure 6. K-factor. Gain adjustor gain with respect to the voltage at pin #8.

The output current of the multiplier is limited to:

$$I_{MUL(MAX)} = \frac{2.5}{R_T}$$

This sets the system current limit. The multiplier output current is converted into the reference voltage for the current (IA) amplifier through a resistor to ground on pin 4, the multiplier output.

Figure 6 shows the gain adjustor (K) with respect to the voltage at pin #8. The curve has been separated in two parts. The right hand part is for operation under normal conditions in the voltage range from minimum line voltage to maximum line voltage (90VAC to 260VAC). 85VAC on the curve has been chosen to account for tolerances. Under normal operating conditions as input voltage decreases the gain increases compensating for the drop in the loop gain.

Under brownout conditions (below 85VAC) the gain decreases to limit the amount of current that is drawn from the line thus preventing an overload condition. This is a very useful feature since in many cases the load for a PFC is a constant power load. The input current has to go high to compensate for a drop in the input voltage.

UNDER VOLTAGE LOCKOUT, OVP AND CURRENT LIMIT

On power-up the ML4821 remains in the UVLO condition; output low and quiescent current low. The IC becomes operational when Vcc reaches 16V. When VCC drops below 9V, the UVLO condition is imposed. During the UVLO condition, the 5V Vref pin is "off", making it usable as a "flag" for starting up a down-stream PWM converter.

OVP, SHUTDOWN, AND IC BIAS

When the input to the OVP comparator exceeds VREF, the output of the ML4821 is inhibited. The OVP input also functions as a "sleep" input, putting the IC into the low quiescent UVLO state when the OVP pin is pulled below 0.7V.

OFF-LINE START-UP AND BIAS SUPPLY GENERATION

The circuit in Figure 11 below supplies VCC power to the ML4821. Start-up current is delivered via R10. The IC starts when pin 15 reaches 15.5V. After that time running power is delivered through the tap on L1. The configuration shown delivers a voltage proportional to VOUT.

ML4821

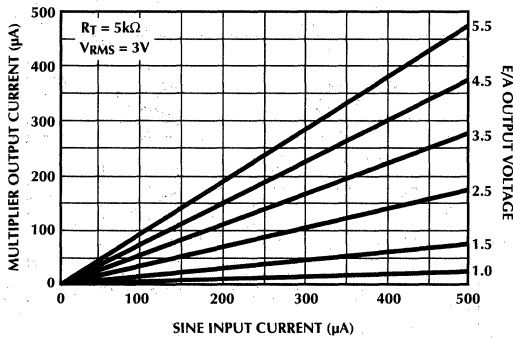


Figure 7. Multiplier Linearity.

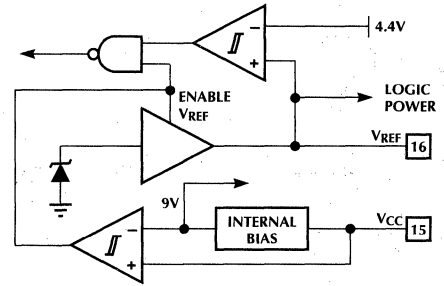


Figure 8. Under-Voltage Lockout Block Diagram.

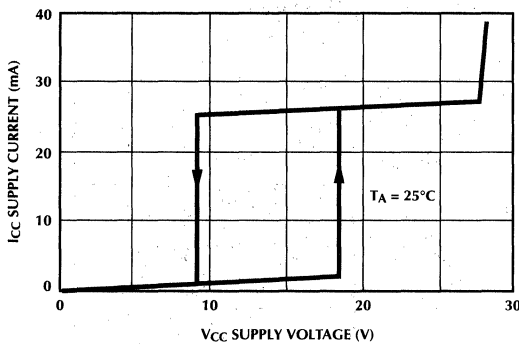


Figure 9. Total Supply Current vs. Supply Voltage.

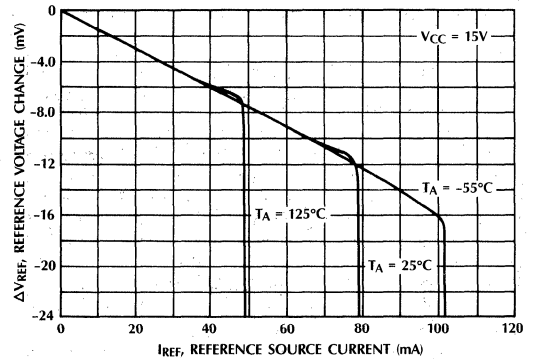


Figure 10. Reference Load Regulation.

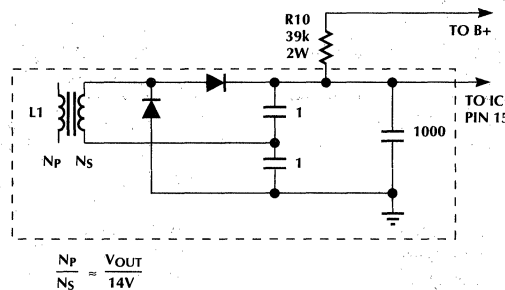


Figure 11. Bias and Start-up Circuit.

ORDERING INFORMATION

PART NUMBER	TEMPERATURE RANGE	PACKAGE
ML4821CP	0°C to +70°C	18-Pin Molded DIP (P18)
ML4821CS	0°C to +70°C	18-Pin Molded SOIC (S18W)

Note: Other packages and temperature ranges can be made available on request. Contact your local Micro Linear Representative for more information.

ML4821EVAL

Average Current PFC Controller Evaluation Kit

GENERAL DESCRIPTION

The ML4821EVAL kit provides a convenient vehicle to evaluate the ML4821 average current sense power factor correction control IC. It contains all of the necessary documentation with the evaluation board and key components to quickly evaluate the application circuit. The board is designed for a 200W universal input PFC circuit. However power components can be replaced for higher or lower power applications.

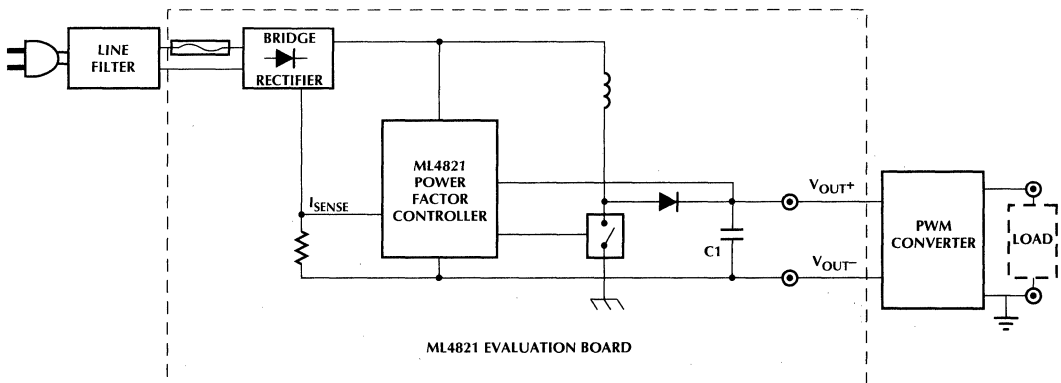
KIT COMPONENTS

- User's Guide
- ML4821 Datasheet
- Application Note 16
- ML4861 Sample
- Evaluation Board
- Key Power Semiconductor Components
- Powder Iron Toroidal Inductor
- Input and output filter capacitors
- Heatsinks

FEATURES

- Harmonic currents well below proposed IEC555-2 limits.
- Power Factor $\gg .99$
- THD $< 5\%$
- Universal Input Range (85Vac to 256Vac)
- 380V output at 200W.
- Efficiency as high as 94%.
- Auxiliary supply from additional inductor winding.
- OVP protection.

BLOCK DIAGRAM



High Frequency Power Supply Controller

GENERAL DESCRIPTION

The ML4823 High Frequency PWM Controller is an IC controller optimized for use in Switch Mode Power Supply designs running at frequencies to 1MHz. Propagation delays are minimal through the comparators and logic for reliable high frequency operation while slew rate and bandwidth are maximized on the error amplifier. This controller is designed for single-ended applications using voltage or current mode and provides for input voltage feed forward.

A 1V threshold current limit comparator provides cycle-by-cycle current limit and exceeding a 1.4V threshold initiates a soft-start cycle. The soft start pin doubles as a maximum duty cycle clamp. All logic is fully latched to provide jitter-free operation and prevent multiple pulsing. An under-voltage lockout circuit with 800mV of hysteresis assures low startup current and drives the outputs low during fault conditions.

The ML4823 is fabricated on a 40V bipolar process FB3480 Power Supply Controller Array. Customized versions of this controller are therefore easily

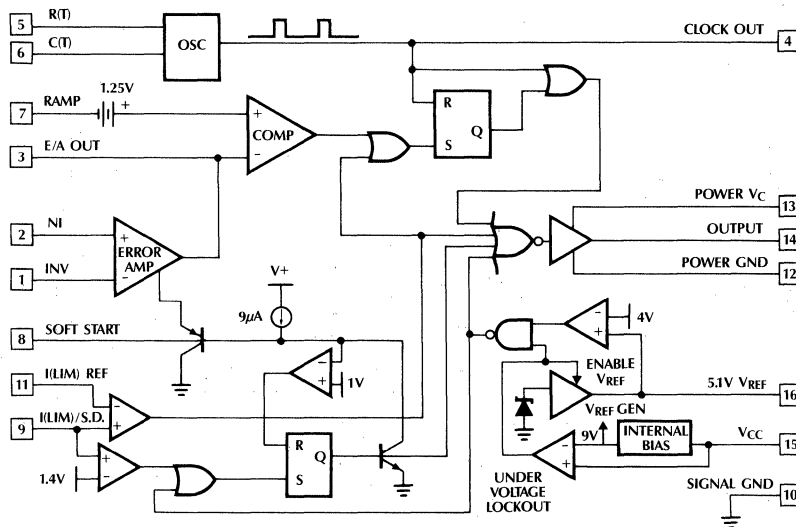
implemented. Please refer to the FB3480 datasheet for more information.

This controller is similar in architecture and performance to the UC1823 controller, however the ML4823 includes features not found on the 1823. These features are set in *Italics*.

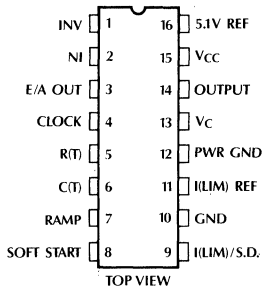
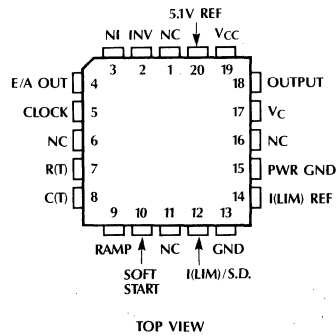
FEATURES

- Practical Operation at Switching Frequencies to 1.0MHz
- High Current (2A peak) Totem Pole Output
- Wide Bandwidth Error Amplifier
- Fully Latched Logic
- Pulse-by-Pulse Current Limiting
- Soft Start and Max. Duty Cycle Control
- Under Voltage Lockout with Hysteresis
- 5.1V, $\pm 1\%$ Trimmed Bandgap Reference
- Pin Compatible Improved Replacement for UC1823
- *Fast Shut Down Path from Current Limit to Outputs*
- *Soft Start Latch Ensures Full Soft Start Cycle*
- *Outputs Pull Low for Undervoltage Lockout*

BLOCK DIAGRAM



PIN CONNECTIONS

ML4823
16-Pin DIPML4823
20-Pin PCC

PIN DESCRIPTION

PIN #	NAME	FUNCTION	PIN #	NAME	FUNCTION
1	INV	Inverting input to error amp.	9	I(LIM)/S.D.	Current limit sense pin. Normally connected to current sense resistor.
2	NI	Non-inverting input to error amp.	10	GND	Analog Signal Ground.
3	E/A OUT	Output of error amplifier and input to main comparator.	11	I(LIM) REF	Reference input for cycle-by-cycle current limit comparator.
4	CLOCK	Oscillator output.	12	PWR GND	Return for the High Current Totem pole outputs.
5	R(T)	Timing Resistor for Oscillator — sets charging current for oscillator timing capacitor (pin 6).	13	VC	Positive Supply for the High Current Totem pole outputs.
6	C(T)	Timing Capacitor for Oscillator.	14	OUT B	High Current Totem pole output.
7	RAMP	Non-Inverting input to main comparator. Connected to C(T) for Voltage Mode operation or to current sense resistor for current mode.	15	VCC	Positive Supply for the IC.
8	SOFT START	Normally connected to Soft Start Capacitor.	16	5.1V REF	Buffered output for the 5.1V voltage reference.

ML4823

ABSOLUTE MAXIMUM RATINGS

Supply Voltage (Pins 15, 13)	30V
Output Current, Source or Sink (Pins 11, 14)	
DC	0.5A
Pulse (0.5 μ s)	2.0A
Analog Inputs	
(Pins 1, 2, 7, 8, 9)	-0.3V to 6V
Clock Output Current (Pin 4)	-5mA
Error Amplifier Output Current (Pin 3)	5mA
Soft Start Sink Current (Pin 8)	20mA
Oscillator Charging Current (Pin 5)	-5mA
Junction Temperature	
ML4823M	150°C
ML4823I, ML4823C	125°C
Storage Temperature Range	-65°C to +150°C
Lead Temperature (Soldering 10 sec.)	+260°C

Thermal Resistance (θ_{JA})

Plastic DIP	65°C/W
Ceramic DIP	65°C/W
Plastic Chip Carrier (PCC)	60°C/W

OPERATING CONDITIONS

Temperature Range

ML4823M	-55°C to +125°C
ML4823I	-40°C to +85°C
ML4823C	0°C to +70°C

Absolute maximum ratings are those values beyond which the device could be permanently damaged. Absolute maximum ratings are stress ratings only and functional device operation is not implied.

ELECTRICAL CHARACTERISTICS

Unless otherwise specified, $R_T = 3.65K\Omega$, $C_T = 1000pF$, $T_A =$ Operating Temperature Range, $V_{CC} = 15V$.

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
Oscillator					
Initial Accuracy	$T_J = 25^\circ C$, (note 1)	360	400	440	KHz
Voltage Stability	$10V < V_{CC} < 30V$, (note 1)		0.2	3	%
Temperature Stability	(note 1)			5	%
Total Variation	line, temp, (note 1)	340		460	KHz
Clock Out High		3.9	4.5		V
Clock Out Low			2.3	2.9	V
Ramp Peak	(note 1)	2.6	2.8	3.0	V
Ramp Valley	(note 1)	0.7	1.0	1.25	V
Ramp Valley to Peak	(note 1)	1.6	1.8	2.0	V
Reference Section					
Output Voltage	$T_J = 25^\circ C$, $I_O = 1mA$	5.00	5.10	5.20	V
Line Regulation	$10V < V_{CC} < 30V$		2	20	mV
Load Regulation	$1mA < I_O < 10mA$		5	20	mV
Temperature Stability	$-55^\circ C < T_J < 150^\circ C$, (note 1)		.2	.4	%
Total Variation	line, load, temp. (note 1)	4.95		5.25	V
Output Noise Voltage	10Hz to 10KHz		50		μV
Long Term Stability	$T_J = 125^\circ C$, 1000 hrs, (note 1)		5	25	mV
Short Circuit Current	$V_{REF} = 0V$	-15	-50	-100	mA
Error Amplifier Section					
Input Offset Voltage	$T_A = 25^\circ C$ Operating Temperature Range			± 15 ± 20	mV mV
Input Bias Current			.6	3	μA
Input Offset Current			.1	1	μA
Open Loop Gain	$1 < V_O < 4V$	60	95		dB

ELECTRICAL CHARACTERISTICS (Continued)Unless otherwise specified, $R_T = 3.65K\Omega$, $C_T = 1000pF$, $T_A =$ Operating Temperature Range, $V_{CC} = 15V$.

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
Error Amplifier Section (Continued)					
CMRR	$1.5 < V_{CC} < 5.5V$, $T_A = 25^\circ C$ Operating Temperature Range	75	95		dB
		65			dB
PSRR	$10 < V_{CC} < 30V$, $T_A = 25^\circ C$ Operating Temperature Range	80	110		dB
		75			dB
Output Sink Current	$V_{PIN\ 3} = 1V$	1	2.5		mA
Output Source Current	$V_{PIN\ 3} = 4V$	-5	-1.3		mA
Output High Voltage	$I_{PIN\ 3} = -0.5mA$	4.0	4.7	5.0	V
Output Low Voltage	$I_{PIN\ 3} = 1mA$	0	0.5	1.0	V
Unity Gain Bandwidth	(note 1)	3	5.5		MHz
Slew Rate	(note 1)	6	12		V/ μs
PWM Comparator Section					
Pin 7 Bias Current	$V_{PIN\ 7} = 0V$		-1	-10	μA
Duty Cycle Range	ML4823C		0	80	%
	ML4823M, ML4823I	$T_A \geq 25^\circ C$	0	80	%
	ML4823M, ML4823I	$T_A < 25^\circ C$	0	70	%
Pin 3 Zero DC Threshold	$V_{PIN\ 7} = 0V$	1.1	1.25		V
Delay to Output	(note 1)		50	80	ns
Soft-Start Section					
Charge Current	$V_{PIN\ 8} = 0.5V$	3	9	20	μA
Discharge Current	$V_{PIN\ 8} = 1V$	1			mA
Current Limit/Shutdown Section					
Pin 9 Bias Current	$0V < V_{PIN\ 9} < 4V$			± 15	μA
Current Limit Offset	$V_{PIN\ 11} = 1.1V$	0		30	mV
Pin 11 Common Mode Range	$T_A > 25^\circ C$	1.0		1.25	V
	$T_A < 25^\circ C$	1.1		1.25	V
Shutdown Threshold		1.25	1.4	1.60	V
Delay to Output	(note 1)		40	70	ns
Output Section					
Output Low Level	$I_{OUT} = 20mA$.25	.4	V
	$I_{OUT} = 200mA$		1.2	2.2	V
Output High Level	$I_{OUT} = -20mA$	13.0	13.5		V
	$I_{OUT} = -200mA$	12.0	13.0		V
Collector Leakage	$V_C = 30V$		100	500	μA
Rise/Fall Time	$C_L = 1000pF$, (note 1)		30	60	ns
Under-Voltage Lockout Section					
Start Threshold		8.7	9.2	9.6	V
UVLO Hysteresis		0.3	0.8	1.2	V
Supply Current					
Start Up Current	$V_{CC} = 8V$, $T_A \geq 0^\circ C$ $T_A \leq 0^\circ C$		1.1	2.5	mA
				3.5	mA
I_{CC}	$V_{PIN\ 1, 7, 9} = 0V$, $V_{PIN\ 2} = 1V$, $T_A = 25^\circ C$		25	33	mA

Note 1: This parameter not 100% tested in production but guaranteed by design.

FUNCTIONAL DESCRIPTION

OSCILLATOR

The ML4823 oscillator charges the external capacitor (C_T) with a current (I_{SET}) equal to $3/R_{SET}$. When the capacitor voltage reaches the upper threshold (Ramp Peak), the comparator changes state and the capacitor discharges to the lower threshold (Ramp Valley) through Q1. While the capacitor is discharging, Q2 provides a high pulse.

The Oscillator period can be described by the following relationship:

$$T_{OSC} = T_{RAMP} + T_{DEADTIME}$$

where: $T_{RAMP} = C (Ramp\ Valley\ to\ Peak) / I_{SET}$

and: $T_{DEADTIME} = C (Ramp\ Valley\ to\ Peak) / I_{Q1}$

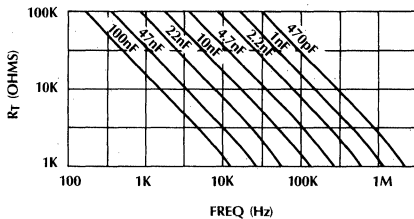
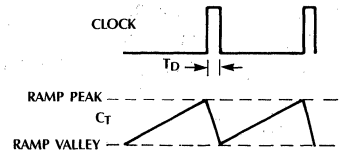
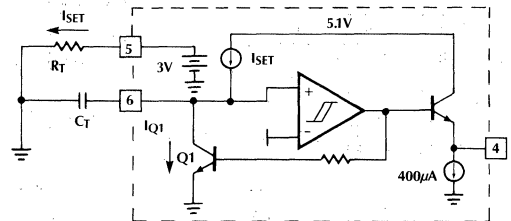


Figure 2. Oscillator Timing Resistance vs Frequency

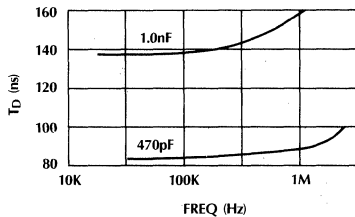


Figure 3. Oscillator Deadtime vs Frequency

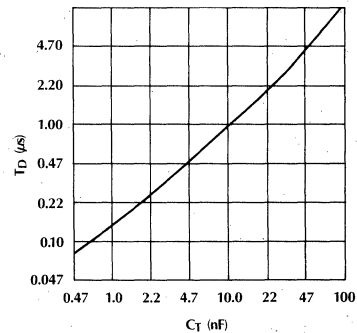


Figure 4. Oscillator Deadtime vs C(T) ($3K\Omega \leq R(T) \leq 100K\Omega$)

ERROR AMPLIFIER

The ML4823 error amplifier is a 5.5MHz bandwidth 12V/ μ s slew rate op-amp with provision for limiting the positive output voltage swing (Output Inhibit line) for ease in implementing the soft start function.

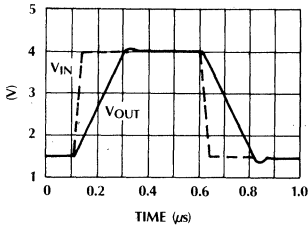


Figure 5. Unity Gain Slew Rate

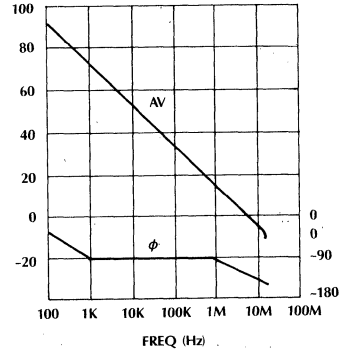


Figure 6. Open Loop Frequency Response

OUTPUT DRIVER STAGE

The ML4823 Output Driver is a 2A peak output high speed totem pole circuit designed to quickly switch the gates of capacitive loads, such as power MOSFET transistors.

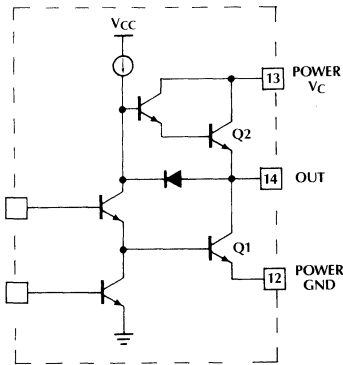


Figure 7. Simplified Schematic

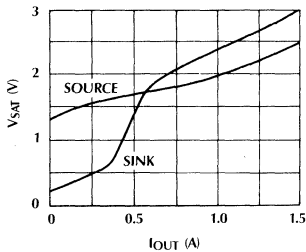


Figure 8. Saturation Curves

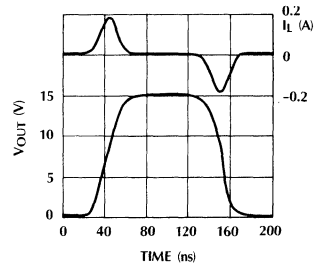


Figure 9. Rise/Fall Time ($C_L = 1000$ pF)

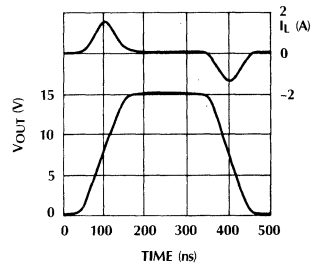


Figure 10. Rise/Fall Time ($C_L = 10,000$ pF)

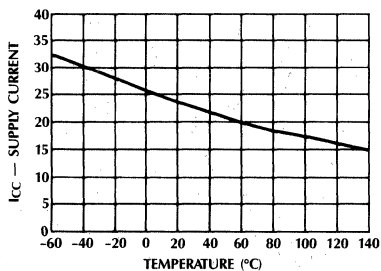


Figure 11. Supply Current vs Temperature

SOFT START AND CURRENT LIMIT

The ML4823 employs two current limits. When the voltage at pin 9 exceeds the (KLIM) REF threshold on pin 11, the outputs are immediately shut off and the cycle is terminated for the remainder of the oscillator period by resetting the RS flip flop.

If the output current is rising quickly (usually due to transformer saturation) such that the voltage on pin 9 reaches 1.4V before the outputs have turned off, a soft start cycle is initiated. The soft start capacitor (pin 8) is discharged and outputs are held "off" until the voltage at pin 8 reaches 1V, ensuring a complete soft start cycle. The duty cycle on start up is limited by limiting the output voltage of the error amplifier voltage to the voltage at pin 8.

ORDERING INFORMATION

PART NUMBER	TEMPERATURE RANGE	PACKAGE
ML4823CP	0°C to +70°C	Molded DIP (P16)
ML4823CQ	0°C to +70°C	Molded PCC (Q20)
ML4823IP	-40°C to +85°C	Molded DIP (P16)
ML4823IQ	-40°C to +85°C	Molded PCC (Q20)
ML4823MJ	-55°C to +125°C	Hermetic DIP (J16)
ML4823CS	0°C to +70°C	Molded SOIC (S16W)

High Frequency Power Supply Controller

GENERAL DESCRIPTION

The ML4825 High Frequency PWM Controller is an IC controller optimized for use in Switch Mode Power Supply designs running at frequencies to 1MHz. Propagation delays are minimal through the comparators and logic for reliable high frequency operation while slew rate and bandwidth are maximized on the error amplifier. This controller is designed to work in either voltage or current mode and provides for input voltage feed forward.

A 1V threshold current limit comparator provides cycle-by-cycle current limit while exceeding a 1.4V threshold initiates a soft-start cycle. The soft start pin doubles as a maximum duty cycle clamp. All logic is fully latched to provide jitter-free operation and prohibit multiple pulsing. An under-voltage lockout circuit with 800mV of hysteresis assures low startup current and drives the outputs low.

The ML4825 is fabricated on a 40V bipolar process from the FB3480 Power Supply Controller Array. Customized versions of this controller are therefore

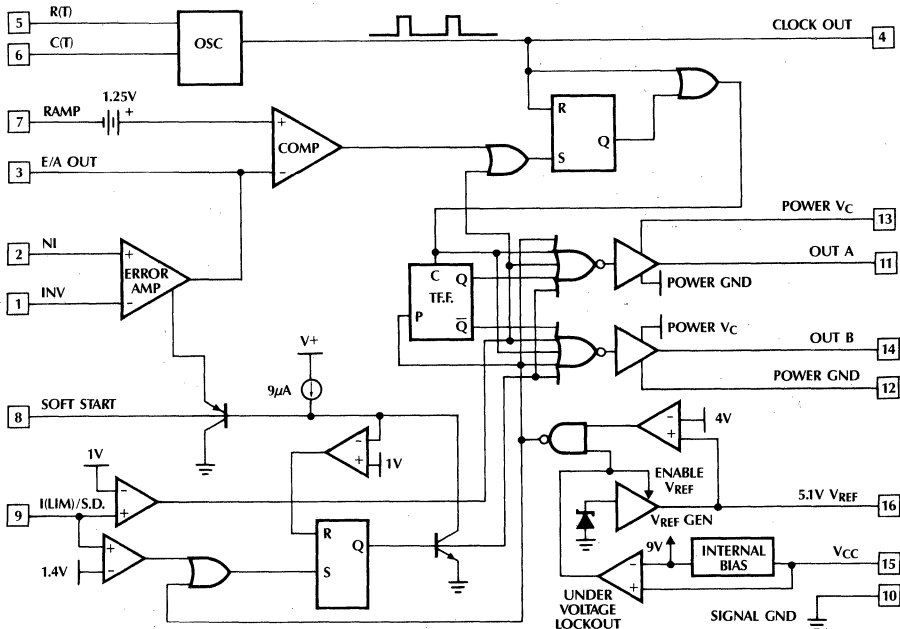
easily implemented. Please refer to the FB3480 datasheet for more information.

This controller is similar in architecture and performance to the UC1825 controller, however the ML4825 includes many features not found on the 1825. These features are set in *Italics*.

FEATURES

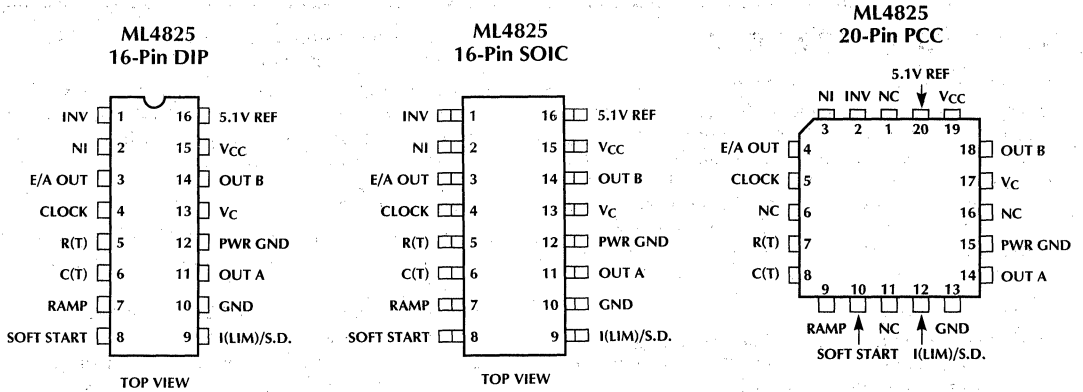
- Practical Operation at Switching Frequencies to 1.0MHz
- High Current (2A peak) Dual Totem Pole Outputs
- Wide Bandwidth Error Amplifier
- Fully Latched Logic with Double Pulse Suppression
- Pulse-by-Pulse Current Limiting
- Soft Start and Max. Duty Cycle Control
- Under Voltage Lockout with Hysteresis
- 5.1V, $\pm 1\%$ Trimmed Bandgap Reference
- Pin Compatible Improved Replacement for UC1825
- *Fast Shut Down Path from Current Limit to Outputs*
- *Outputs Preset to Known Condition After Under Voltage Lockout*
- *Soft Start Latch Ensures Full Soft Start Cycle*
- *Outputs Pull Low for Undervoltage Lockout*

BLOCK DIAGRAM



ML4825

PIN CONNECTIONS



PIN DESCRIPTION

PIN #	NAME	FUNCTION	PIN #	NAME	FUNCTION
1	INV	Inverting input to error amp.	9	I(LIM)/S.D.	Current limit sense pin. Normally connected to current sense resistor.
2	NI	Non-inverting input to error amp.	10	GND	Analog Signal Ground.
3	E/A OUT	Output of error amplifier and input to main comparator.	11	OUT A	High Current Totem pole output. This output is the first one energized after Power On Reset.
4	CLOCK	Oscillator output.	12	PWR GND	Return for the High Current Totem pole outputs.
5	R(T)	Timing Resistor for Oscillator — sets charging current for oscillator timing capacitor (pin 6).	13	V _C	Positive Supply for the High Current Totem pole outputs.
6	C(T)	Timing Capacitor for Oscillator.	14	OUT B	High Current Totem pole output.
7	RAMP	Non-Inverting input to main comparator. Connected to C(T) for Voltage Mode operation or to current sense resistor for current mode.	15	V _{CC}	Positive Supply for the IC.
8	SOFT START	Normally connected to Soft Start Capacitor.	16	5.1V REF	Buffered output for the 5.1V voltage reference.

ABSOLUTE MAXIMUM RATINGS

Supply Voltage (Pins 15, 13)	30V
Output Current, Source or Sink (Pins 11, 14)	
DC	0.5A
Pulse (0.5 μ s)	2.0A
Analog Inputs	
(Pins 1, 2, 7)	-0.3V to 7V
(Pins 9, 8)	-0.3V to 6V
Clock Output Current (Pin 4)	-5mA
Error Amplifier Output Current (Pin 3)	5mA
Soft Start Sink Current (Pin 8)	20mA
Oscillator Charging Current (Pin 5)	-5mA
Junction Temperature	
ML4825I, ML4825C	125°C
Storage Temperature Range	-65°C to +150°C
Lead Temperature (Soldering 10 sec.)	+260°C

Thermal Resistance (θ_{JA})	
Plastic DIP or SOIC	65°C/W
Plastic Chip Carrier (PCC)	60°C/W

OPERATING CONDITIONS

Temperature Range	
ML4825I	-40°C to +85°C
ML4825C	0°C to +70°C

Absolute maximum ratings are those values beyond which the device could be permanently damaged. Absolute maximum ratings are stress ratings only and functional device operation is not implied.

ELECTRICAL CHARACTERISTICS

Unless otherwise specified, $R_T = 3.65k\Omega$, $C_T = 1000pF$, $T_A =$ Operating Temperature Range, $V_{CC} = 15V$.

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS	
Oscillator						
Initial Accuracy	$T_J = 25^\circ C$, (note 1)	360	400	440	KHz	
Voltage Stability	$10V < V_{CC} < 30V$, $T_A = 25^\circ C$ (note 1)		0.2	2	%	
Temperature Stability	(note 1)			5	%	
Total Variation	line, temp, (note 1)	340		460	KHz	
Clock Out High		3.9	4.5		V	
Clock Out Low			2.3	2.9	V	
Ramp Peak	(note 1)	2.6	2.8	3.0	V	
Ramp Valley	(note 1)	0.7	1.0	1.25	V	
Ramp Valley to Peak	(note 1)	1.6	1.8	2.0	V	
Reference Section						
Output Voltage	ML4825C	$T_J = 25^\circ C$, $I_O = 1mA$	5.00	5.10	5.20	V
	ML4825I		5.05	5.10	5.15	V
Line Regulation	$10V < V_{CC} < 30V$		2	20	mV	
Load Regulation	$1mA < I_O < 10mA$		5	20	mV	
Temperature Stability	$-55^\circ C < T_J < 150^\circ C$, (note 1)		.2	.4	%	
Total Variation	ML4825C	line, load, temp (note 1)	4.95		5.25	V
	ML4825I	line, load, temp $T_A > 0^\circ C$, (note 1)	5.00		5.20	V
	ML4825I	line, load, temp $T_A < 0^\circ C$, (note 1)	4.95		5.25	V
Output Noise Voltage	10Hz to 10KHz		50		μV	
Long Term Stability	$T_J = 125^\circ C$, 1000 hrs, (note 1)		5	25	mV	
Short Circuit Current	$V_{REF} = 0V$	-15	-50	-100	mA	
Error Amplifier Section						
Input Offset Voltage	ML4825I,C	$T_A > 0^\circ C$			15	mV
	ML4825I	$T_A < 0^\circ C$			20	mV
Input Bias Current			.6	3	μA	
Input Offset Current			.1	1	μA	
Open Loop Gain	$1 < V_O < 4V$	60	96		dB	

ML4825

ELECTRICAL CHARACTERISTICS (Continued)

Unless otherwise specified, $R_T = 3.65K\Omega$, $C_T = 1000pF$, $T_A =$ Operating Temperature Range, $V_{CC} = 15V$.

PARAMETER		CONDITIONS	MIN	TYP	MAX	UNITS
Error Amplifier Section (Continued)						
CMRR	ML4825I,C	$1.5V < V_{CM} < 5.5V, T_A > 0^\circ C$	75	95		dB
	ML4825I	$1.5V < V_{CM} < 5.5V, T_A < 0^\circ C$	65	95		dB
PSRR	ML4825I,C	$10V < V_{CC} < 30V, T_A > 0^\circ C$	80	110		dB
	ML4825I	$10V < V_{CC} < 30V, T_A < 0^\circ C$	75	110		dB
Output Sink Current		$V_{PIN\ 3} = 1V$	1	2.5		mA
Output Source Current		$V_{PIN\ 3} = 4V$	-5	-1.3		mA
Output High Voltage		$I_{PIN\ 3} = -0.5mA$	4.0	4.7	5.0	V
Output Low Voltage		$I_{PIN\ 3} = 1mA$	0	0.5	1.0	V
Unity Gain Bandwidth		(note 1)	3	5.5		MHz
Slew Rate		(note 1)	6	12		V/ μs
PWM Comparator Section						
Pin 7 Bias Current	ML4825I,C	$V_{PIN\ 7} = 0V, T_A > 0^\circ C$		-1	-5	μA
	ML4825I	$V_{PIN\ 7} = 0V, T_A < 0^\circ C$			-10	μA
Duty Cycle Range	ML4825C		0		85	%
	ML4825I	$T_A > 0^\circ C$	0		80	%
	ML4825I	$T_A < 0^\circ C$	0		70	%
Pin 3 Zero DC Threshold		$V_{PIN\ 7} = 0V$	1.1	1.25		V
Delay to Output		(note 1)		50	80	ns
Soft-Start Section						
Charge Current		$V_{PIN\ 8} = 0.5V$	3	9	20	μA
Discharge Current		$V_{PIN\ 8} = 1V$	1			mA
Current Limit/Shutdown Section						
Pin 9 Bias Current	ML4825C	$0V < V_{PIN\ 9} < 4V$			+10	μA
	ML4825I	$0V < V_{PIN\ 9} < 4V$			+15	μA
Current Limit Threshold			.9	1	1.1	V
Shutdown Threshold		$T_A > 0^\circ C$	1.25	1.4	1.55	V
		$T_A < 0^\circ C$	1.25	1.4	1.60	V
Delay to Output		(note 1)		40	70	ns
Output Section						
Output Low Level		$I_{OUT} = 20mA$.25	.4	V
		$I_{OUT} = 200mA$		1.2	2.2	V
Output High Level		$I_{OUT} = -20mA$	13.0	13.5		V
		$I_{OUT} = -200mA$	12.0	13.0		V
Collector Leakage		$V_C = 30V$		100	500	μA
Rise/Fall Time		$C_L = 1000pF$, (note 1)		30	60	ns
Under-Voltage Lockout Section						
Start Threshold			8.7	9.2	9.6	V
UVLO Hysteresis			0.3	0.8	1.2	V
Supply Current						
Start Up Current	ML4825I,C	$V_{CC} = 8V, T_A > 0^\circ C$		1.1	2.5	mA
	ML4825I	$V_{CC} = 8V, T_A < 0^\circ C$			3.5	mA
I_{CC}		$V_{PIN\ 1,7,9} = 0V, V_{PIN\ 2} = 1V, T_A = 25^\circ C$		26	33	mA

Note 1: This parameter not 100% tested in production but guaranteed by design.

FUNCTIONAL DESCRIPTION

OSCILLATOR

The ML4825 oscillator charges the external capacitor (C_T) with a current (I_{SET}) equal to $3/R_{SET}$. When the capacitor voltage reaches the upper threshold (Ramp Peak), the comparator changes state and the capacitor discharges to the lower threshold (Ramp Valley) through $Q1$. While the capacitor is discharging, $Q2$ provides a high pulse.

The Oscillator period can be described by the following relationship:

$$T_{OSC} = T_{RAMP} + T_{DEADTIME}$$

where: $T_{RAMP} = C (Ramp Valley to Peak)/I_{SET}$

and: $T_{DEADTIME} = C (Ramp Valley to Peak)/I_{Q1}$

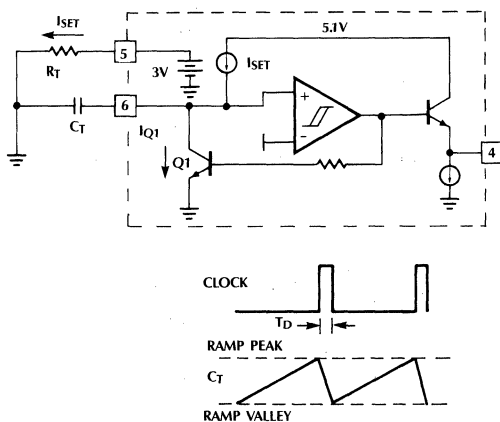


Figure 1. Oscillator Block Diagram

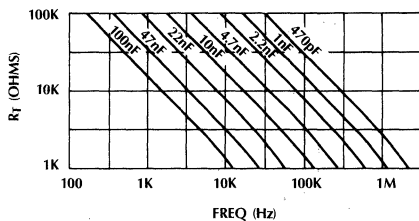


Figure 2. Oscillator Timing Resistance vs Frequency

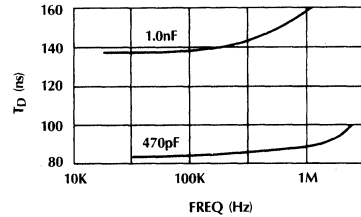


Figure 3. Oscillator Deadtime vs Frequency

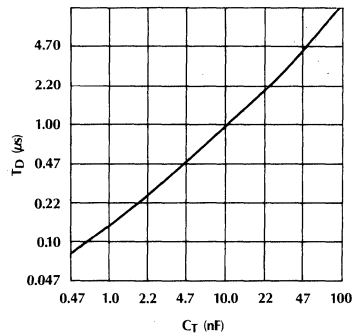


Figure 4. Oscillator Deadtime vs C_T ($3K\Omega \leq R(T) \leq 100K\Omega$)

ERROR AMPLIFIER

The ML4825 error amplifier is a 5.5MHz bandwidth $12V/\mu s$ slew rate op-amp with provision for limiting the positive output voltage swing (Output Inhibit line) for ease in implementing the soft start function.

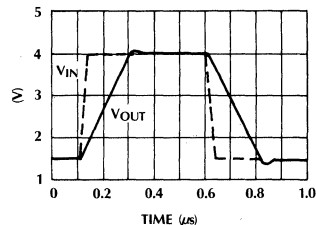


Figure 5. Unity Gain Slew Rate

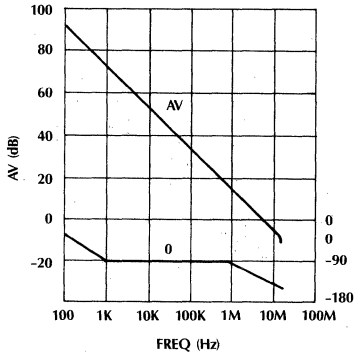


Figure 6. Open Loop Frequency Response

OUTPUT DRIVER STAGE

The ML4825 Output Driver is a 2A peak output high speed totem pole circuit designed to quickly switch the gates of capacitive loads, such as power MOSFET transistors.

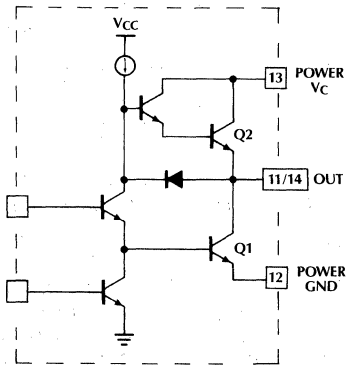


Figure 7. Simplified Schematic

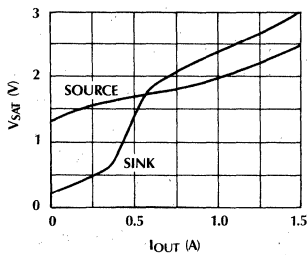


Figure 8. Saturation Curves

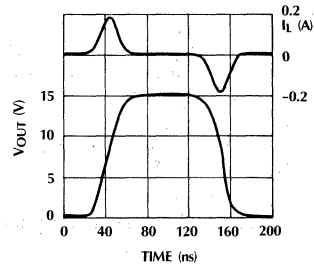


Figure 9. Rise/Fall Time ($C_L = 1000pF$)

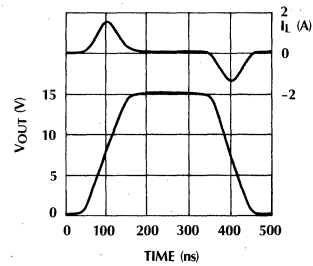


Figure 10. Rise/Fall Time ($C_L = 10,000pF$)

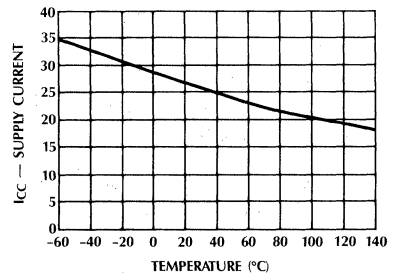


Figure 11. Supply Current vs Temperature

SOFT START AND CURRENT LIMIT

The ML4825 employs two current limits. When the voltage at pin 9 exceeds 1V, the outputs are immediately shut off and the cycle is terminated for the remainder of the oscillator period by resetting the RS flip flop.

If the output current is rising quickly such that the voltage on pin 9 reaches 1.4V before the outputs have turned off, a soft start cycle is initiated. The soft start capacitor (pin 8) is discharged and outputs are held "off" until the voltage at pin 8 reaches 1V, ensuring a complete soft start cycle. The duty cycle on start up is limited by limiting the output voltage of the error amplifier voltage to the voltage at pin 8.

ORDERING INFORMATION

PART NUMBER	TEMPERATURE RANGE	PACKAGE
ML4825CP	0°C to +70°C	Molded DIP (P16)
ML4825CQ	0°C to +70°C	Molded PCC (Q20)
ML4825IP	-40°C to +85°C	Molded DIP (P16)
ML4825IQ	-40°C to +85°C	Molded PCC (Q20)
ML4825CS	0°C to +70°C	Molded SOIC (S16W)

Electronic Ballast Controller

GENERAL DESCRIPTION

The ML4830 is a complete solution for a dimmable, high power factor, high efficiency electronic ballast. Contained in the ML4830 are controllers for "boost" type power factor correction as well as for a dimming ballast.

The Power factor circuit uses the average current sensing method with a current fed multiplier and over-voltage protection. This system produces power factors of better than 0.99 with low input current THD at > 95% efficiency. Special care has been taken in the design of the ML4830 to increase system noise immunity by using a high amplitude oscillator, and a current fed multiplier. An over-voltage protection comparator stops the PFC section in the event of sudden load decrease.

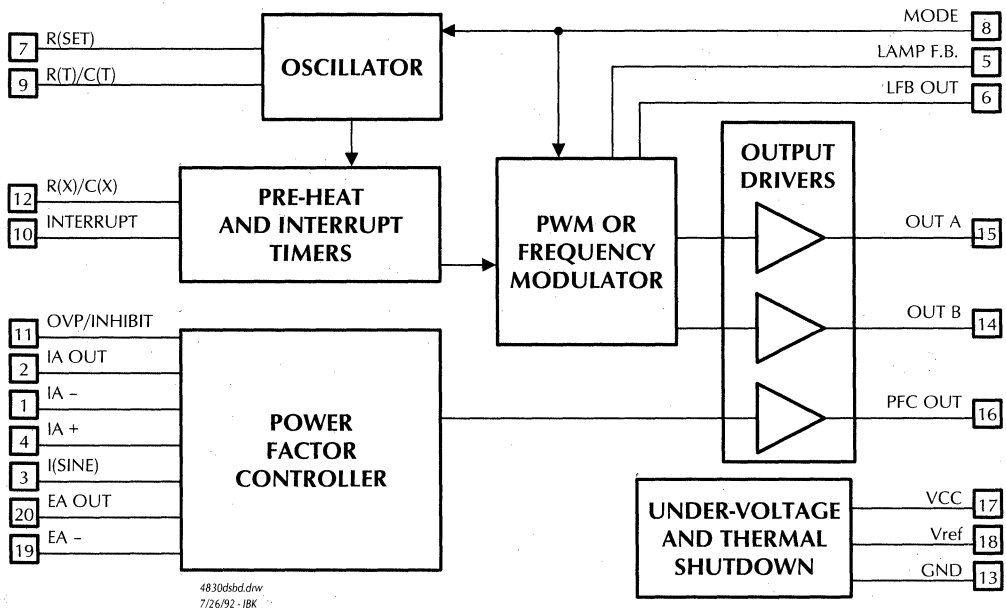
The ballast section provides for programmable starting scenarios with programmable pre-heat and lamp out-of-socket interrupt times. The IC controls lamp output through either frequency or Pulse Width control using lamp current feedback.

The ML4830 is designed using Micro Linear's Semi-Standard tile array methodology. Customized versions of this IC, optimized to specific ballast architectures can be made available. Contact Micro Linear or an authorized representative for more information.

FEATURES

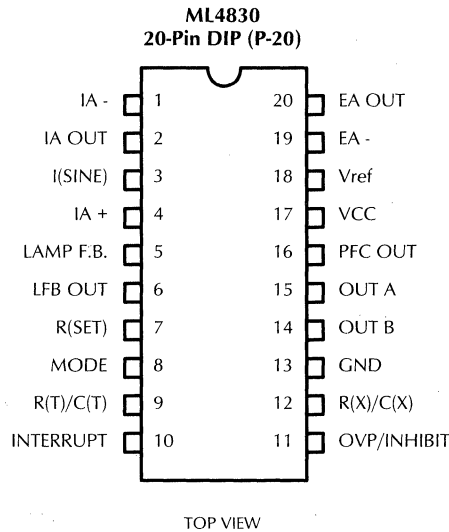
- Complete Power Factor Correction and Dimming Ballast Control on one IC
- Low Distortion, High Efficiency Continuous Boost, Average Current sensing PFC section
- Programmable Start Scenario for Rapid or Instant Start Lamps
- Lamp Current feedback for Dimming Control
- Selectable Variable Frequency or PWM dimming modes
- Programmable Restart for lamp out condition to reduce ballast heating
- Over-Temperature Shutdown feature replaces external heat sensor for safety.
- PFC Over-Voltage comparator eliminates output "run-away" due to load removal
- Large oscillator amplitude and current fed multiplier improves noise immunity

SIMPLIFIED BLOCK DIAGRAM



PAT. PEND.

PIN CONFIGURATION



PIN DESCRIPTION

PIN#	NAME	FUNCTION	PIN#	NAME	FUNCTION
1	IA -	Inverting input of the PFC average current error amplifier	10	INTERRUPT	A voltage of greater than Vref resets the chip and causes a restart after a delay of 3 times the start interval. Used for lamp-out detection and restart
2	IA OUT	Output and compensation node of the PFC average current error amplifier	11	OVP/INHIBIT	When the voltage of this pin exceeds 5V, the PFC output is inhibited. When the voltage exceeds 6.8V, the IC function is inhibited and the IC is reset. This pin can be used for a remote ballast shut-down
3	I(SINE)	PFC Current Multiplier input.	12	R(X)/C(X)	Sets the timing for the preheat, dimming lockout, and interrupt
4	IA +	Non-Inverting input of the PFC average current error amplifier and input of peak current limit comparator	13	GND	IC Ground
5	LAMP F.B.	Inverting input of an Error Amplifier used to sense (and regulate) lamp arc current. Also the input node for dimming control	14	OUT B	Ballast MOSFET drive output
6	LFB OUT	Output from the Lamp Current Error Amplifier used for lamp current loop compensation	15	OUT A	Ballast MOSFET drive output
7	R(SET)	External resistor which sets oscillator F_{MAX} Multiplier max I_{OUT} and R(X)/C(X) charging current	16	PFC OUT	Power Factor MOSFET drive output
8	MODE	Controls how the Lamp Current Error Amp and preheat timers modulate the ballast outputs. Two Variable Frequency and 1 PWM mode are available through this pin	17	VCC	Positive Supply for the IC
9	R(T)C(T)	Oscillator timing components	18	Vref	Buffered output for the 5V voltage reference
			19	EA -	Inverting input to PFC error amplifier
			20	EA OUT	PFC Error Amplifier output and compensation node

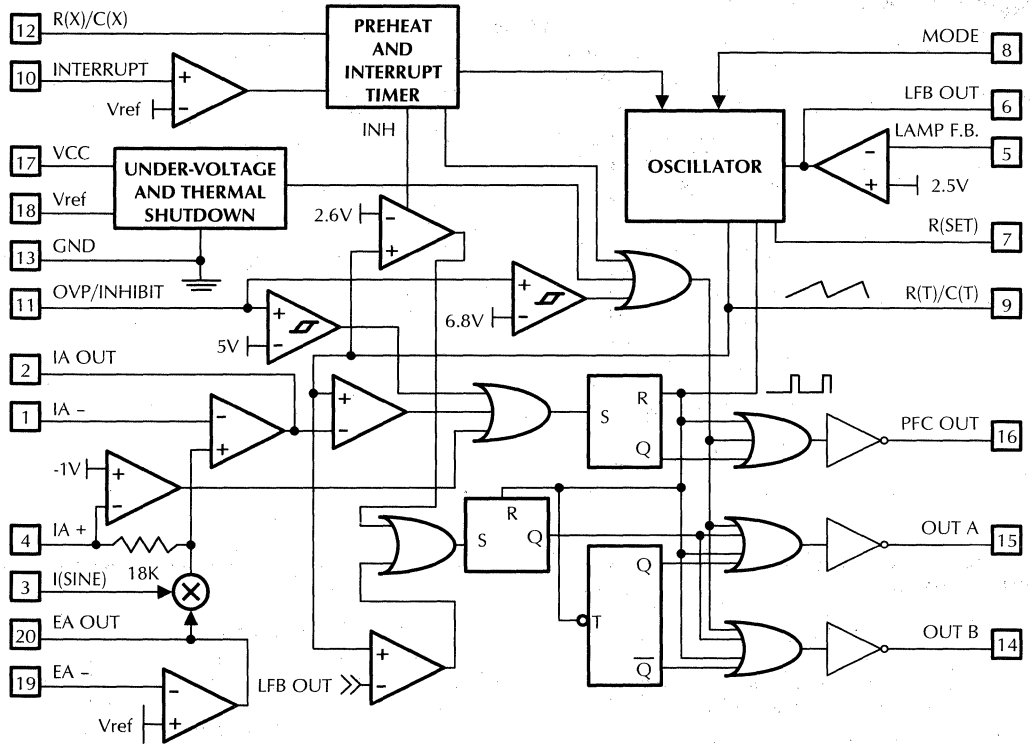


Figure 1. ML4830 Block Diagram

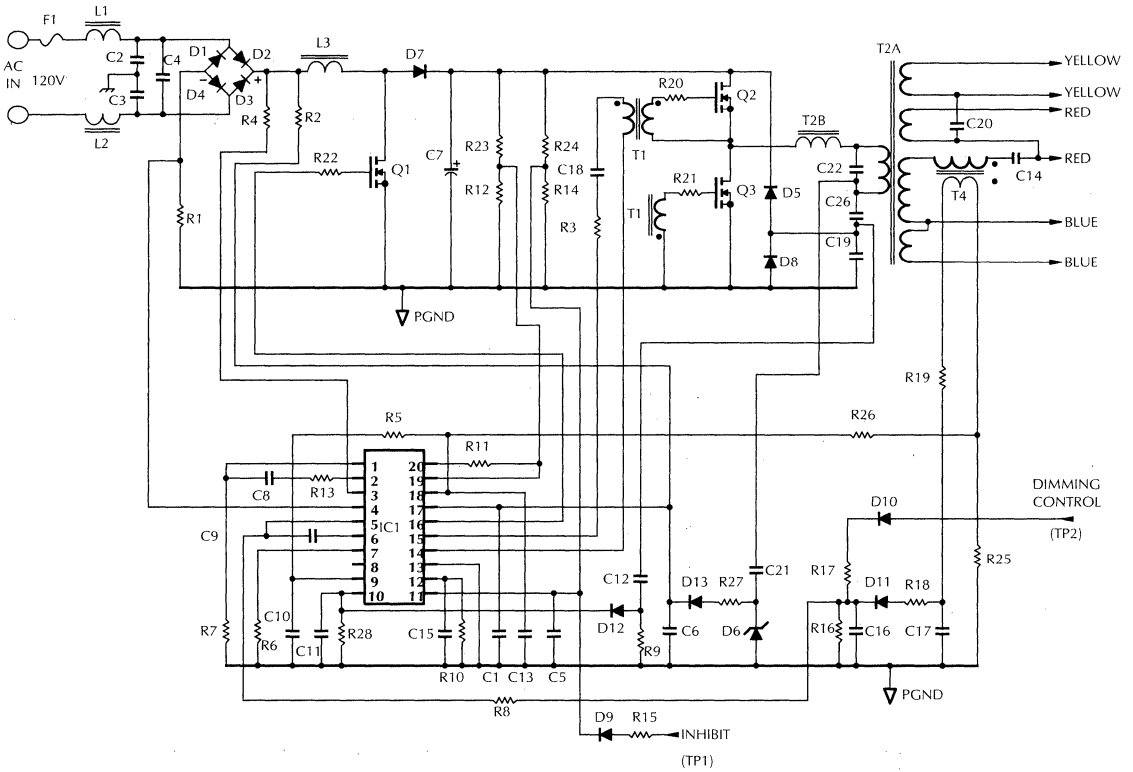


Figure 2. Typical Application: 2-Lamp Dimming Ballast

ORDERING INFORMATION

PART NUMBER	TEMPERATURE RANGE	PACKAGE
ML4830CP	0°C to +85°C	MOLDED DIP (P20)

Other packages and temperature ranges can be made available on request. Contact your local Micro Linear Representative for more information.

Low Voltage Boost Regulator

GENERAL DESCRIPTION

The ML4861 is a complete solution for DC to DC conversion for 1 to 3 cell battery powered systems. The systems requires a minimum number of external components: one inductor and one capacitor. The boost circuit of the ML4861 is capable of running with input voltages as low as 1V. Three different versions are available: a 5V output, 3.3V output, or an adjustable version.

Special care was taken in the design of the ML4861 to optimize system efficiency. Synchronous rectification eliminates the need for an external schottky diode and provides lower forward drop. The regulator operates using discontinuous boost and only starts an output pulse when the output voltage has drooped. The regulator operates as a variable frequency system and only uses power when switching.

The circuit also contains a RESET output which goes low when the IC can no longer function due to low input voltage, or when the DETECT input drops below 200mV.

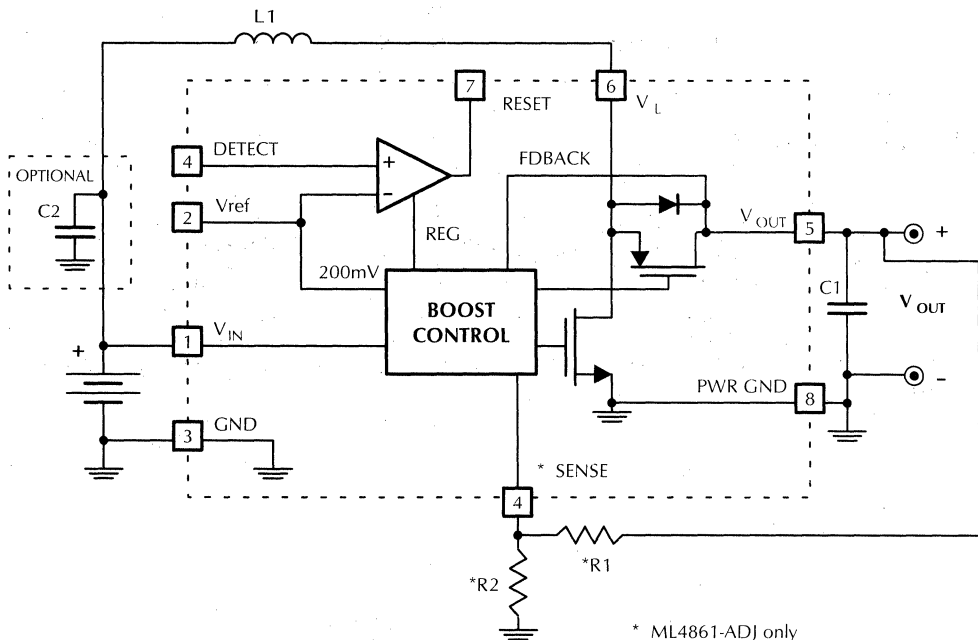
The ML4861 output is $\pm 5\%$ accurate and is specified as a complete system. Typical efficiency is 90% for a 5V output with a 2.5V input.

The ML4861 is fabricated using Micro Linear's FC3580. This allows for customizing of the IC for a user's specific application. Contact Micro Linear for more information.

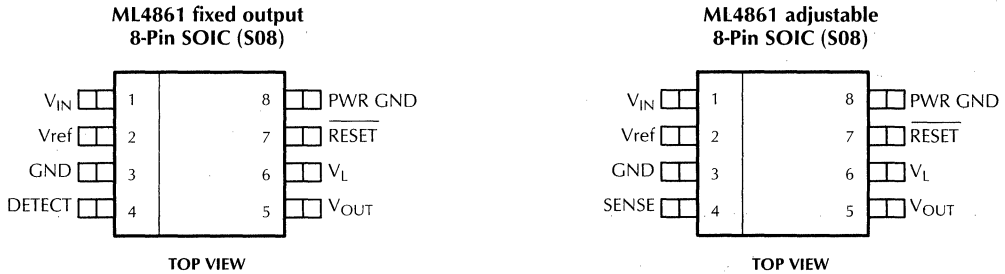
FEATURES

- Synchronous Rectification for high efficiency (90%)
- Minimal External Components : 1 Cap, 1 Inductor
- Very Low $R_{DS\ ON}$ (0.2 Ω N-Channel, 0.5 Ω P-Channel)
- Works down to 1V input
- Reset flag to Microprocessor when V_{IN} is removed
- Micro Power BiCMOS (50 μ A I_{SUPPLY})
- Low output detect comparator
- 8-Pin SOIC Package
- 5V, 3.3V and Adjustable Output Versions

BLOCK DIAGRAM



PIN CONFIGURATION



Part Number	Output Voltage	Temperature Range	Package
ML4861CS-3.3	3.3V	0°C to 70°C	SOIC (S08)
ML4861CS-5.0	5.0V	0°C to 70°C	SOIC (S08)
ML4861CS-ADJ	ADJUSTABLE	0°C to 70°C	SOIC (S08)

PIN DESCRIPTION

FIXED OUTPUT VERSIONS

PIN# NAME	FUNCTION
1 V _{IN}	Battery Input voltage.
2 V _{ref}	200mV Reference pin.
3 GND	Analog signal ground.
4 DETECT	When this input is below V _{ref} , the RESET pin goes low.
5 V _{OUT}	Output of the Boost Regulator
6 V _L	Boost Inductor Connection
7 RESET	Output goes low when regulation cannot be achieved or when DETECT is below 200mV.
8 PWR GND	Power Ground

ADJUSTABLE VERSION

PIN# NAME	FUNCTION
1 V _{IN}	Battery Input voltage.
2 V _{ref}	200mV Reference pin.
3 GND	Analog signal ground.
4 SENSE	Programming pin for the adjustable voltage version.
5 V _{OUT}	Output of the Boost Regulator
6 V _L	Boost Inductor Connection
7 RESET	Output goes low when regulation cannot be achieved
8 PWR GND	Power Ground

ABSOLUTE MAXIMUM RATINGS

Voltage on any pin	6V
Peak Inductor Current	1A
Junction Temperature	150°C
Storage Temperature Range	-65°C to +150°C
Lead Temperature (Soldering 10 Sec.)	+260°C
Thermal Resistance (θ_{JA})	
SOIC	85°C/W

OPERATING CONDITIONS

Temperature Range	0°C to +70°C
Operating Voltage Range	
V_{IN}	1.0V to $V_{OUT} - 0.2V$

Absolute maximum ratings are those values beyond which the device could be permanently damaged. Absolute maximum ratings are stress ratings only and functional device operation is not implied.

ELECTRICAL CHARACTERISTICS

Unless otherwise specified, T_A = Operating Temperature Range, $V_{OUT} = 5V$ for ML4861-5.0, -ADJ, $V_{OUT} = 3.3V$ for ML4861-3.3.

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS		
SUPPLY CURRENT							
V_{IN} Current	$V_{IN} = V_{OUT} - 0.2V$		50	70	μA		
V_{OUT} Quiescent Current			5	8	μA		
V_L Quiescent Current			1	2	μA		
REFERENCE SECTION							
Vref Output Voltage	$0 < I_{PIN2} < -5\mu A$, V_{IN} = Operating Range	190	200	210	mV		
PFM REGULATOR							
Pulse Width		9	10	11	μS		
Output Voltage	ML4861-5.0, -ADJ	Table 1, $0 < I_{OUT} < I_{MAX}$		4.75	5	5.25	V
	ML4861-3.3	Table 1, $0 < I_{OUT} < I_{MAX}$		3.135	3.3	3.465	V
RESET COMPARATOR							
DETECT Threshold		190	200	210	mV		
DETECT Bias Current			-25	-100	nA		
Pin 7 V_{OH}	$I_{OH} = -10\mu A$, $2V < V_{OUT} < 5V$	$V_{OUT} - 0.2$			V		
Pin 7 V_{OL}	$I_{OL} = 20\mu A$, $2V < V_{OUT} < 5V$			0.2	V		
V_{IN} Reset Threshold		0.65	0.85	1	V		

V_{IN}	L (μH)	EFFICIENCY % (TYP)	I_{OUT} (mA)
$V_{OUT} = 5V$			
1.0	10	75	65
1.0	20	85	40
1.5	15	77	110
1.5	30	88	60
3.0	30	81	215
3.0	60	90	120
4.5	45	81	320
4.5	90	91	180

V_{IN}	L (μH)	EFFICIENCY % (TYP)	I_{OUT} (mA)
$V_{OUT} = 3.3V$			
1.0	10	75	120
1.0	20	85	68
1.5	15	87	175
1.5	30	87	100
3.0	30	75	355
3.0	60	85	200

Table 1. Output Current Capabilities of ML4861 at Various Input Voltages

FUNCTIONAL DESCRIPTION

The ML4861 combines Pulse Frequency Modulation (PFM) and synchronous rectification to create a boost converter that is both highly efficient and simple to use. The PFM approach allows a single inductor to be charged to a fixed peak current and then completely discharged before another cycle begins, simplifying the design by eliminating the need for the conventional current limiting circuitry. Synchronous rectification is accomplished by replacing the conventional external Schottky diode with an on chip PMOS device, reducing losses and eliminating the external component.

REGULATOR OPERATION

A block diagram of the boost converter is shown in Figure 1. When V_{OUT} is at or above the desired output voltage, the circuit remains idle, drawing only $50\mu A$ from the input supply pin V_{IN} and only $5\mu A$ from the output V_{OUT} through the feedback resistors, R1 and R2. When V_{OUT} drops below the desired output level, the A1 output goes high, signaling the circuit to deliver charge to the output. Since the A2 output is normally high, the flip-flop captures the A1 set signal and creates a pulse at the gate of the NMOS transistor Q1. The NMOS will charge the inductor for $10\mu s$, resulting in a peak current given by:

$$I_{L(MAX)} = \frac{T_{ON} \times V_{IN}}{L1} \approx \frac{10\mu s \times V_{IN}}{L1}$$

For reliable operation, L1 should be chosen such that $I_{L(MAX)}$ does not exceed 1A.

When the one shot times out the NMOS releases the V_L pin allowing the inductor to fly-back and momentarily charge the output through the body diode of PMOS

transistor Q2. But as the voltage across the PMOS changes polarity, its gate gets driven low by the current sensing amplifier A2, and the PMOS shorts out the diode. The inductor then discharges into the load with minimum voltage drop through the PMOS. The output of A2 also serves to reset the flip-flop and one shot in preparation for the next charging cycle. When current in the PMOS falls to zero, A2 releases the gate and, if V_{OUT} is still low, allows the flip-flop to initiate another pulse. The inductor current and one shot waveforms are shown in Figure 2.

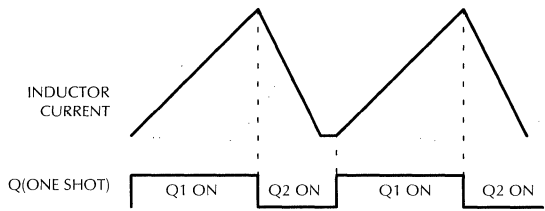
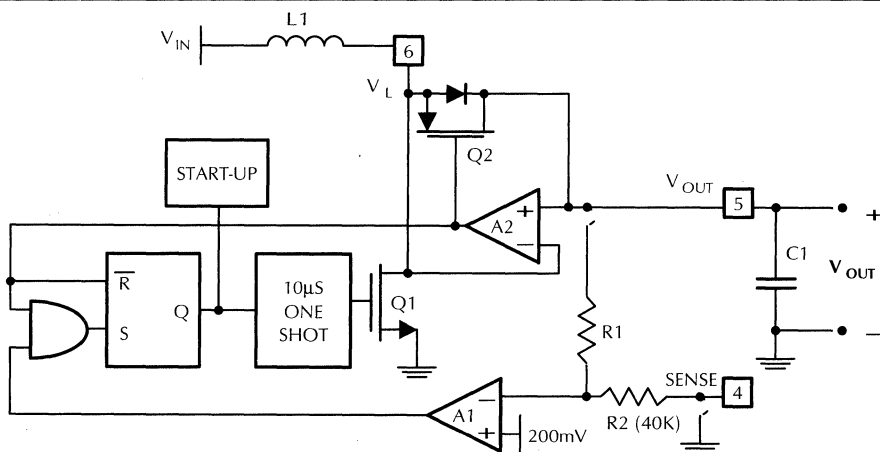


Figure 2. PFM Inductor Current Wave Forms and Timing

The maximum output current available from the regulator can be determined from the following equation:

$$I_{OUT (MAX)} \approx \frac{EFFICIENCY \times V_{IN} \times 0.45 \times I_{L(MAX)}}{V_{OUT}}$$

A trade off exists between efficiency and maximum output current. Higher efficiency can be obtained by choosing a larger inductor value. However, this will also reduce the maximum output current available.



* Jumpers not connected for ML4861-ADJ

Figure 1. PFM Regulator Block Diagram

The output filter capacitor will limit the ripple on V_{OUT} . The value can be chosen based on the acceptable change in output voltage for a single inductor discharge cycle, which is given by:

$$\Delta V_{OUT} = \frac{0.5 \times V_{IN} \times 10\mu S \times I_{L(MAX)}}{C1 \times (V_{OUT} - V_{IN})}$$

For $V_{IN} = 1.5V$, $V_{OUT} = 5V$, $I_{L(MAX)} = 0.5A$ and $C1 = 47\mu F$, the output ripple will be 23mV. Capacitor equivalent series resistance (ESR) must also be considered when choosing a capacitor, both for ripple and efficiency.

If the input voltage impedance is too high, an additional $1\mu F$ capacitor (C2) may be required from input source to keep V_{IN} from being pulled low during normal operation.

The adjustable version of the ML4861 requires an external resistor feedback string to set V_{OUT} . The output level can be determined from the following equation:

$$V_{OUT} = 0.2 \times \frac{(R1+R2)}{R2}$$

where R1 and R2 are as shown in Figure 1. R2 should be around $40K\Omega$ or less to keep base current errors low.

In the adjustable version, the feedback node on A1 is brought out to pin 4 through R2 and the jumpers shown in Fig. 1 are not connected.

RESET COMPARATOR

An additional comparator is provided to detect low V_{OUT} , low V_{IN} , or some other error condition important to the user. The inverting input of the comparator is connected to 200mV, while the non-inverting input is provided externally at the DETECT pin. The output of the comparator is the RESET pin, which swings from ground to V_{OUT} . This output is OR-ed with the internal under-voltage detect circuit, and will go low when V_{IN} is removed.

ML4861EVAL

Low Voltage Boost Regulator Evaluation Kit

GENERAL DESCRIPTION

The ML4861EVAL kit provides a convenient vehicle to evaluate the ML4861 low voltage boost regulator IC. It contains all of the necessary documentation with the evaluation board, IC and key components to quickly evaluate the application circuit. This board can be used for either fixed or adjustable versions.

The ML4861 is a complete solution for DC to DC conversion for 1 to 3 cell battery powered systems. The system requires a minimum number of external components: one inductor and one capacitor. The boost circuit of the ML4861 is capable of running with input voltages as low as 1V. Three different versions of the IC are available: a 5V output, 3.3V output, or an adjustable version. The evaluation kit is shipped with the ML4861-5.0

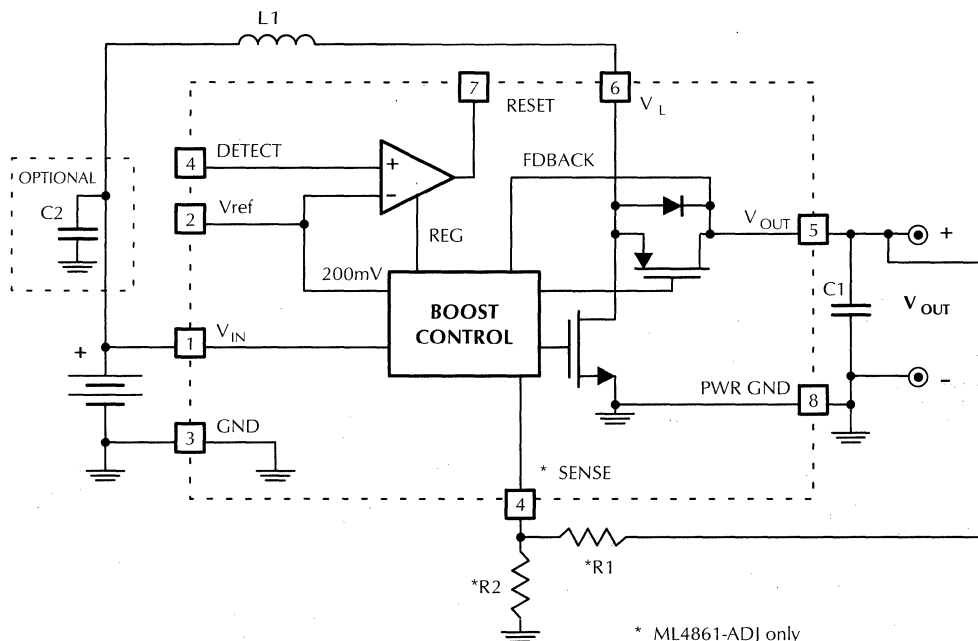
FEATURES

- Synchronous Rectification for high efficiency (90%)
- Wide Input Range (1V to 4.5V)
- Accommodates either fixed or adjustable versions
- Regulation to better than $\pm 5\%$ over line and load
- Low Output Ripple (50mV p-p)
- Regulation down to no load

KIT COMPONENTS

- User's Guide
- ML4861 Datasheet
- ML4861 Sample
- PC Board
- Toroid for Inductor (L1)

BLOCK DIAGRAM


6

Battery Power Control IC

GENERAL DESCRIPTION

The ML4862 is a complete solution for DC to DC conversion and power management in multi-cell battery powered portable computers and instruments. Several advanced techniques are incorporated in the IC for the highest possible systems efficiency and lowest possible battery drain.

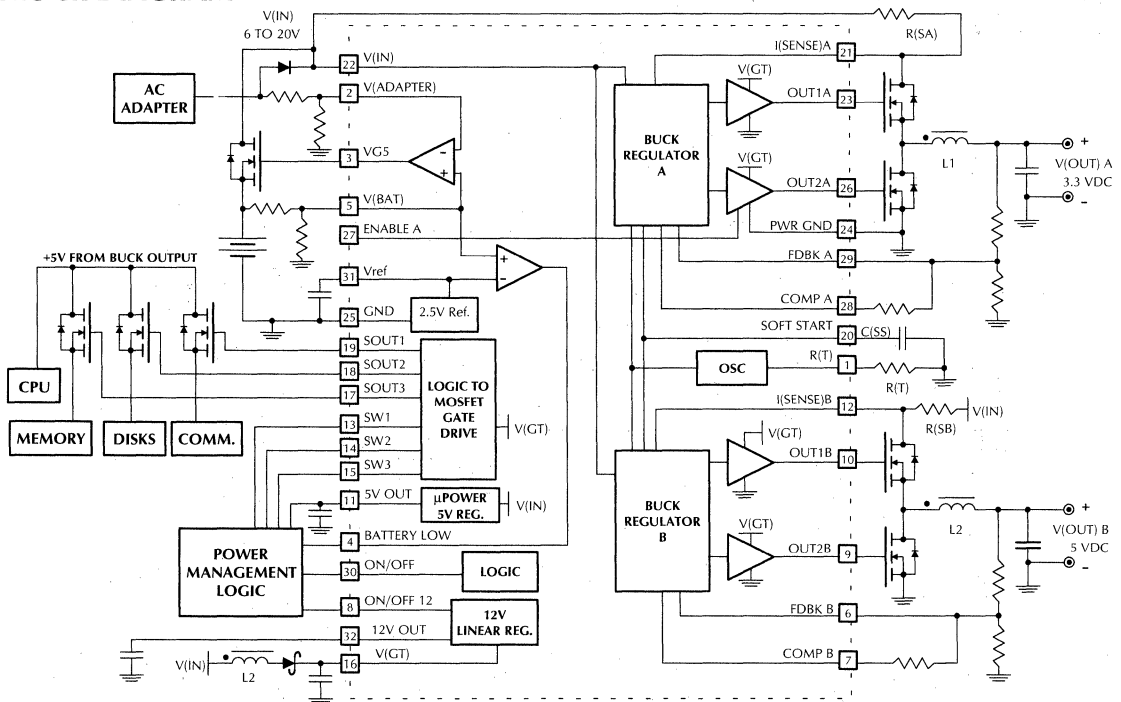
The 5.0V and 3.3V main regulators in the ML4862 each controls a Synchronous Rectified buck regulator and each drives two N-Channel MOSFETs. This allows high conversion efficiencies (90% or greater). Bias for all N-Channel MOSFETs in the system as well as the input for the 12V regulator for programming EEPROMs comes from an auxiliary winding on the buck regulator choke.

The ML4862 also contains 3 outputs to drive external N-Channel MOSFETs to power down disk drives and memory under control of external logic. Automatic switch-over to battery operation is also provided when the charger is removed. A μ Power 5V linear regulator and low battery indicator are provided for the power monitoring logic.

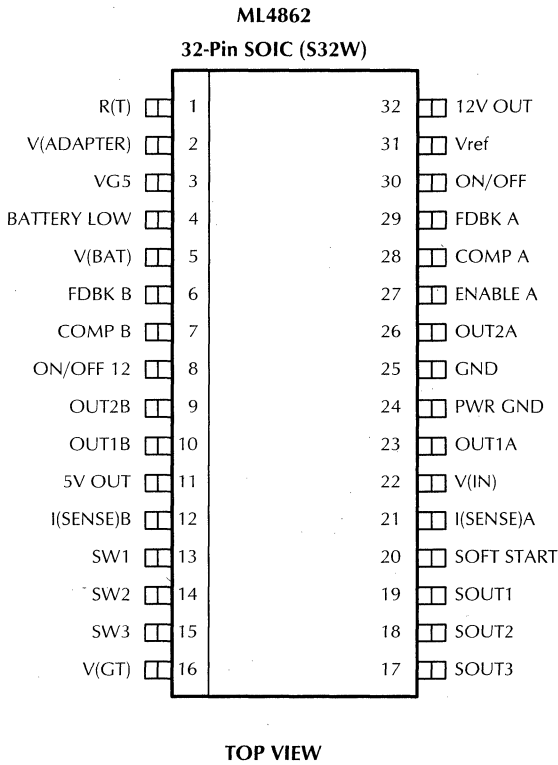
FEATURES

- Two Synchronously Rectified, 100KHz Buck Regulators: for 5V and 3.3V outputs
- Low Cost All N-Channel MOSFET Switching
- Three Logic to N-Channel Gate Drive translators for power management
- μ Power 5V standby linear regulator to run power management logic
- Output and logic for N-Channel MOSFET to disconnect battery when charger is connected
- 12V Auxiliary output available with On/Off Control for E² memory programming
- Low battery detect comparator
- Wide Input Voltage Range (5V to 20V)
- Customizable Tile Array Technology - Consult factory for additional options

BLOCK DIAGRAM



PIN CONFIGURATION



PIN	NAME	FUNCTION
9	OUT2B	5V Buck Regulator Synchronous Rectifier Output
10	OUT1B	5V Buck Regulator Switch Output
11	5V OUT	Output of the μ Power 5V regulator. Normally used to power external management circuits and logic
12	I(SENSE)B	Current Sensing for buck regulator B current limit
13-15	SW1-3	Inputs for power management MOSFET gate drivers
16	V(GT)	Boosted voltage to drive N-Channel gates and input to 12V linear regulator
17-19	SOUT3-1	MOSFET gate drive outputs for power management
20	SOFT START	Connected to a soft start capacitor
21	I(SENSE)A	Current Sensing for buck regulator A current limit
22	V(IN)	Input from Battery or AC Adapter
23	OUT1A	3.3V Buck Regulator Switch Output
24	PWR GND	Power Ground
25	GND	Logic and signal Ground
26	OUT2A	3.3V Buck Regulator Synchronous Rectifier Output
27	ENABLE A	A logic low disables Buck Regulator A's Synchronous Rectifier output
28	COMP A	Buck Regulator A frequency compensation terminal
29	FDBK A	Voltage Feedback for buck regulator A
30	ON/OFF	A low on this pin disables all IC functions except the low battery detection comparator, the linear 5V regulator and the 2.5V reference, and puts the IC into a low current consumption mode
31	Vref	Buffered 2.5V reference output
32	12V OUT	Output of the 12V linear regulator

PIN DESCRIPTION

PIN	NAME	FUNCTION
1	R(T)	Timing Resistor which sets oscillator frequency
2	V(ADAPTER)	Input to sense whether adapter is active. When this pin is above V(BAT), VG5 goes low.
3	VG5	Output to drive N-Channel MOSFET gate to switch battery out when adapter is present
4	BATTERY LOW	A logic low level indicates the voltage on V(BAT) is below 2.5V. This is an open-collector output.
5	V(BAT)	Battery Comparator input
6	FDBK B	Voltage Feedback for buck regulator B
7	COMP B	Buck Regulator B frequency compensation terminal
8	ON/OFF 12	A logic high turns on the 12V linear regulator

ABSOLUTE MAXIMUM RATINGS

Absolute maximum ratings are those values beyond which the device could be permanently damaged. Absolute maximum ratings are stress ratings only and functional device operation is not implied.

Voltage on any pin	36V
Output Current, Source or Sink (Pins 9,10, 23, 26)	
Pulsed	300mA
VG5 Source Current	-20mA
VG5 Sink Current	200mA
12V Linear Regulator Output Current	200mA
5V Linear Regulator Output Current	50mA
Logic Inputs (pins 8,13,14,15,27,30)	-0.3V to 5.5V

ISENSE Inputs (pins 12,21)	V(IN)
Comparator Inputs (pins 2, 5)	-0.3V to 5.5V
Junction Temperature	150°C
Storage Temperature Range	-65°C to +150°C
Lead Temperature (Soldering 10 Sec.)	+260°C
Thermal Resistance (θ _{JA}) Plastic SOIC	60°C/W

OPERATING CONDITIONS

Ambient Temperature Range	0°C to +70°C
V(IN) Voltage Range	5.4V to 24V
V(GT) Voltage Range	V(IN)-0.5 to 35V

ELECTRICAL CHARACTERISTICS

Unless otherwise specified, T_A=Operating Temperature Range, V(IN)=12V, V(GT)=22V, R(T)=120KΩ.

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
OSCILLATOR					
Initial accuracy		98	100	102	Khz.
Dead Time	50K ≤ R(T) ≤ 300K	600	800	1000	nS
Maximum Duty Cycle		90		94	%
Voltage Stability	6V ≤ V(IN) ≤ 20V		2		%
ERROR AMPLIFIERS					
Input Offset Voltage			2	10	mV
Input Bias Current			10	100	nA
Output High Voltage	I _{OUT} = -2mA	2.8	2.95		V
Output Low Voltage	I _{OUT} = 20μA			0.2	V
Source Current	V _{OUT} =2.5V	-5	-7		mA
Sink Current	V _{OUT} =2.5V	10	50		μA
Gain-Bandwidth Product			675		KHz
HIGH SIDE (OUT1) OUTPUTS (Pins 10, 23)					
Output High Voltage	I _{OUT} = -400mA	18.5	19.5		V
	I _{OUT} = -20mA	21	21.3		V
Output Low Voltage	I _{OUT} = 200mA		1.6	2	V
	I _{OUT} = 20mA		0.2	0.5	V
LOW SIDE (OUT2) OUTPUTS (Pins 9, 26)					
Output High Voltage	I _{OUT} = -400mA	13.5	14.1		V
	I _{OUT} = -20mA	14	14.4		V
Output Low Voltage	I _{OUT} = 200mA		1.6	2	V
	I _{OUT} = 20mA		0.2	0.5	V
SOFT START AND CURRENT LIMITS (Pins 12, 20 21)					
Voltage Threshold	From V(IN)	-250	-200	-150	mV
Bias Current	V _{I(SENSE)} =V(IN)-200mV		27	50	μA
Soft Start I _{CHARGE}		-10	-13	-16	μA
Soft Start I _{DISCHARGE}		5	6.2		mA
REFERENCE SECTION					
Output Voltage	T _A =25°C, I _{PIN31} =-1mA	2.475	2.5	2.525	V
Line regulation	5.4V < V _{IN} < 20V, I _{PIN31} ≤ -1mA		0.3	0.8	mV/V

ELECTRICAL CHARACTERISTICS

Unless otherwise specified, T_A =Operating Temperature Range, $V(IN)=12V$, $V(GT)=22V$, $R(T)=120K\Omega$.

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
12V LINEAR REGULATOR					
Output Voltage	$I_O=60mA$	11.25	12	12.75	V
Line regulation	$16V < V(GT) < 30V$.01		%/V
Drop Out Voltage ($V_{GT}-V_{OUT}$)	$I_O=10mA$		2.3	2.7	V
Load regulation	$10\mu A < I_O < 60mA$		200	400	mV
5V LINEAR REGULATOR					
Output Voltage	$I_{OUT}=1mA$	4.85	5.0	5.15	V
Input Voltage	$V_{OUT} \geq 4.85V$, $I_O=10mA$	6.2			V
Line Regulation	$5.4V < V_{IN} < 20V$, $I_O=10mA$.75	1.5	%
Load Regulation	$10\mu A < I_O < 25mA$		1.5	2	%
V(BAT) and V(ADAPTER) COMPARATORS					
Input Bias Current	$V(IN)=20V$			100	nA
Input Offset Voltage				± 10	mV
Battery Low V_{OL}	$I_{OL}=200\mu A$.4	V
Battery Low V_{OH}	$30K\Omega$ pullup to 5V	4.5	5.0		V
VG5 Source Current	$V_{PIN3}=V(GT)-3V$	-10	-15		mA
VG5 Sink Current	$V_{PIN3}=1V$	100			mA
POWER MANAGEMENT DRIVERS (Pins 13-15, 17-19)					
Source Current	$V_{SOUT}=10V$	-20	-40	-60	μA
Sink Current	$V_{SOUT}=10V$	20	42	60	μA
Output High Voltage	$I_{SOUT} = -20\mu A$	14.5	15.6		V
Output Low Voltage	$I_{SOUT} = 20\mu A$.16	.4	V
LOGIC INPUTS (Pins 8, 13-15, 30, 27)					
Logic Low (V_{IL})	$I_{IN} \geq -5\mu A$			1.1	V
Logic High (V_{IH})	$I_{IN} \geq 5\mu A$	2.5			V
SUPPLY CURRENT					
$I(IN)+I(GT)$	Sleep Mode, $T_A=25^\circ C$		115	150	μA
$I(IN)$	Run Mode, $T_A=25^\circ C$		6	10	mA
$I(GT)$	Run Mode, $T_A=25^\circ C$		4	6	mA

6

FUNCTIONAL DESCRIPTION

POWER DOWN MODES

The ML4862 operates in either a powered down mode or a run mode according to the state of the ON/OFF pin. When the ON/OFF pin is high, the IC is in the run mode and all IC sections are functioning. When the ON/OFF pin is low, the IC is in the standby mode and only the μ Power 5V linear regulator, 2.5V reference, and BATTERY LOW comparator are on. All gate drive outputs are low. The 5V linear regulator then provides the power to run the system's power management logic.

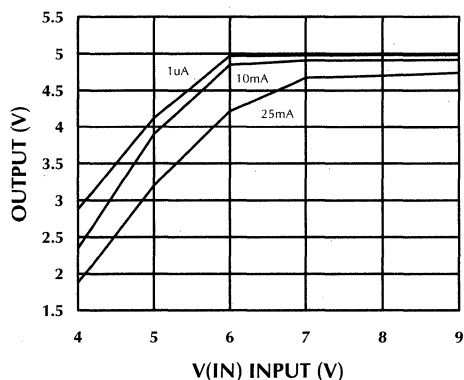


Figure 1. 5V Linear Regulator Output at low V(IN)

BUCK REGULATORS

The two buck regulators (fig. 3) are synchronously rectifying voltage mode PWM regulators capable of being used over a wide variety of loads and input voltages. The use of synchronous rectification improves system efficiency by reducing the fixed drop associated with the "freewheeling" diode in conventional regulators. These regulators also drive all N-Channel power MOSFETs, significantly improving system efficiency at a low cost. In order to drive the MOSFET gates adequately, a V(GT) supply must be provided which is higher than the battery voltage by an amount sufficient to provide full enhancement voltage to the MOSFETs. This can be generated by using a winding from the 5V buck regulator as shown below. V(GT) must not be lower than 0.5V below V(IN).

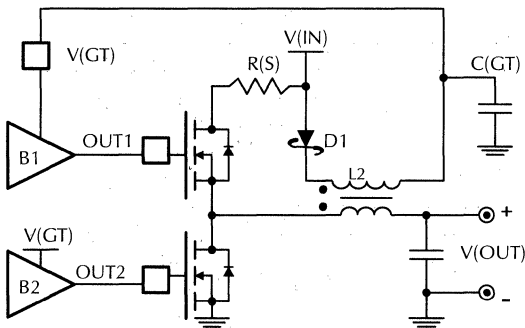


Figure 2. Generating V(GT) bias voltage

Buck Regulator A includes a pin to disable the synchronous rectifier driver (OUT2A) to prevent pulling current out from the output, allowing the inductor current to become discontinuous at light loads.

Selection of the external MOSFETs, output inductor and capacitor determine the output capabilities of the regulator. Output voltage is set by RF1 and RF2 where.

$$V_{OUT} = \frac{2.5 \times (RF1 + RF2)}{RF2} \quad (1)$$

The short circuit current limit is set by external resistor R(S).

$$I_{SHORT\ CKT} \approx \frac{0.2}{R(S)} \quad (2)$$

C(SS) is discharged when the regulator is off or when the voltage across R(S) exceeds 200mV. F2 ensures that C(SS) is fully discharged. This circuit provides reliable output short circuit protection with very little power wasted in the sensing element. The error amplifier's output voltage is limited to the voltage on the SOFT START pin. When C(SS) is discharged, the regulator's duty cycle is 0.

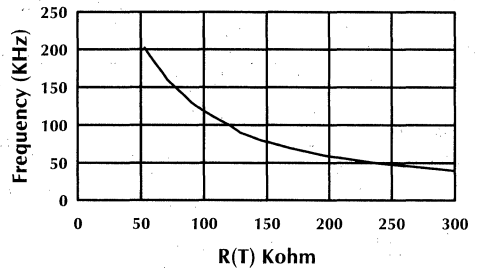


Figure 4. Oscillator Frequency vs. R(T)

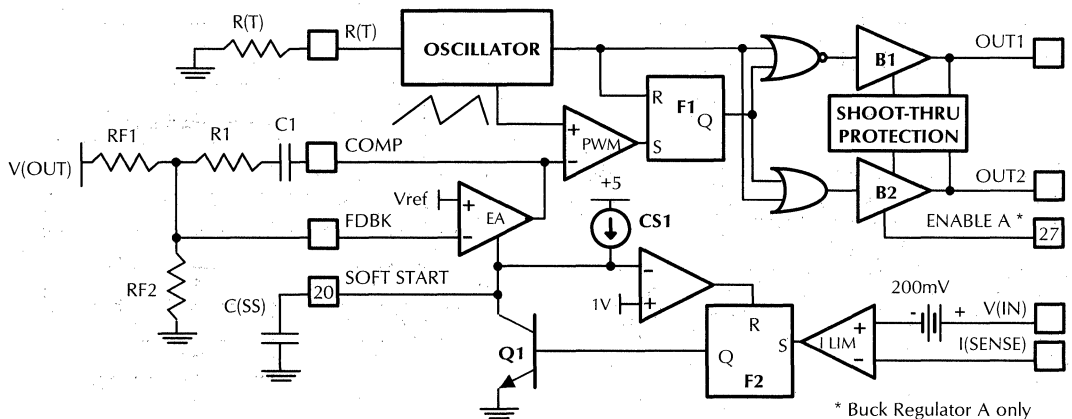


Figure 3. Buck Regulator Block Diagram

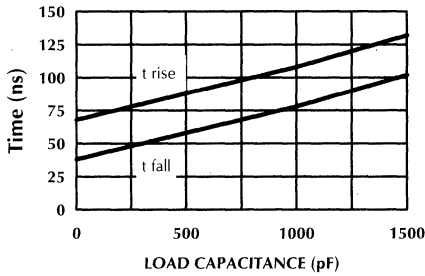


Figure 5. OUT1 Rise and Fall Time vs. Load C_{LOAD}

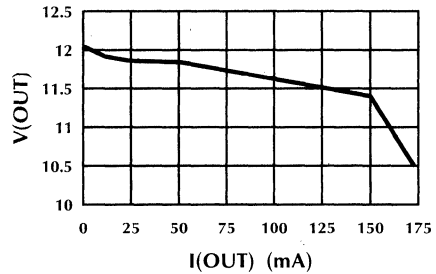


Figure 7. 12V Regulator Load Regulation

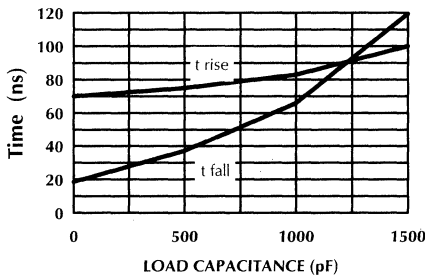


Figure 6. OUT2 Rise and Fall Time vs. Load C_{LOAD}

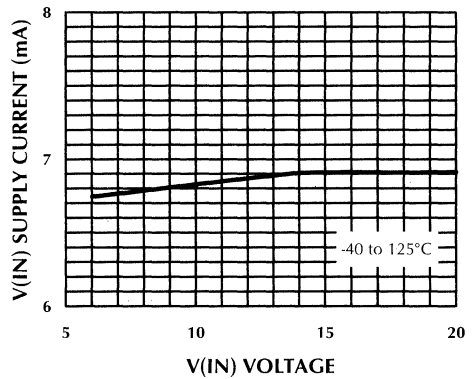


Figure 8. Supply Current (V_{IN}) vs. V_{IN} Voltage

LOGIC TO MOSFET GATE DRIVERS

This section provides a convenient translation stage for turning on low cost external N-Channel power MOSFETs for power management. SOUT sources current when SW is high and sinks current when SW is low. These outputs are limited to small currents to switch these MOSFETs slowly, reducing the transients to the main regulator output from switching in discharged capacitances. Additional slow down of the switching may be achieved by adding capacitance from the SOUT pins to GND.

BATTERY DETECTION AND ADAPTER SWITCH

These two functions are provided by comparators. The V(ADAPTER) comparator goes high when the V(BAT) pin is higher than V(ADAPTER), so that the system can run from the battery without the loss associated with a diode. When the AC adapter is plugged in, the voltage on pin 2 goes high, VG4 switches low, and the system runs from the AC adapter. A low battery comparator with an open collector output is also provided to monitor battery level.

12V LINEAR REGULATOR

The 12V regulator includes a shut-off pin. Since this regulator takes its input from V(GT) care should be taken to ensure that the regulator does not cause excessive heat in the IC when used with high values of V(IN).

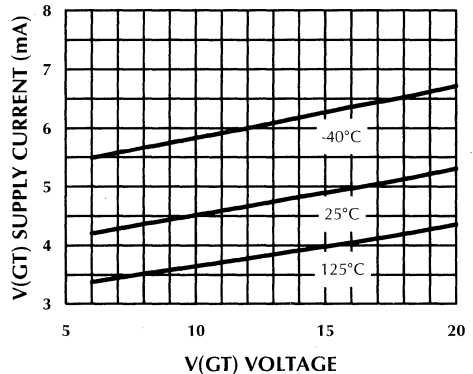


Figure 9. Supply Current (I_{GT}) vs. V_{GT} Voltage

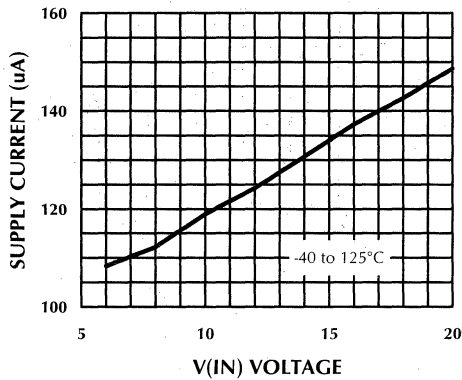


Figure 10. SLEEP mode Current (I_{IN}) vs. V_{IN} Voltage

SEMI-STANDARD OPTIONS

The ML4862 can be quickly and easily modified to suit individual customer requirements. Examples of some of the possible customizations might include:

- Reducing the number of pins, or IC cost by eliminating unwanted functions
- Replacing unwanted functions with other functions
- Putting certain external programming components on chip to save board space
- Running the 12V regulator from V(IN) instead of V(GT) (when V(IN) > 14.5V).
- Adding "Burp Mode" to increase efficiency at light loads

Please contact Micro Linear for more information on Semi-Standard options for the ML4862.

APPLICATIONS

BUCK REGULATOR INDUCTOR

Inductors are specified with three main parameters; inductance (L), maximum current (I_{OUT(MAX)}), and DC resistance. (R_L)

Inductance for a given set of requirements can be calculated with the following:

$$L \approx (V_{IN} - V_{OUT}) \frac{V_{OUT}}{V_{IN} \times F \times \Delta I}$$

Assuming that the desired ripple current:

$$\Delta I = 40\%(I_{MAX})$$

By choosing the ripple current to be 40% of maximum output current (I_{MAX}), below 20% of I_{MAX}, the inductor will actually pull current out of the output capacitor during

part of the on time of the low side MOSFET. The result of this current "shuttling" is a slight increase in losses. Buck regulator A includes the ability to disable the synchronous rectifier output to avoid current shuttling at light loads and thereby operate using conventional rectification.

To avoid inductor saturation, the maximum output current of the regulator should not exceed 80% of the current rating of the inductor, especially when using ferrites, which have a "hard" saturation characteristic. Powdered iron cores saturate more softly and may therefore be pushed closer to their rated currents.

DC resistance of the inductor sets up its conduction loss. For the same size package DC resistance decreases as inductance is decreased. It is a good rule of thumb to select the DC resistance of the inductor to be 1/4 of the sum of the on resistance of the two output MOSFETs. This sets up conduction losses evenly among the power components.

V(GT) GENERATION

A two to one secondary winding from the 5V main output inductor will provide the gate drive voltage needed for high side switching (fig. 2). Because one end of the winding is tied to the input, the secondary voltage will be 10V above the input. When specifying the inductor, the maximum current rating of the inductor is the sum of the peak main output current and turns ratio multiplication of the maximum secondary current only if both outputs draw maximum current at the same time. Depending on I(GT) load requirements, the inductor current rating may be specified for maximum main output current without including the secondary current requirement.

FREQUENCY SELECTION

Frequency is set by the resistor R(T), which establishes the charge current for the internal capacitor. Since the discharge current is a constant, the dead time of the oscillator is constant. Therefore the maximum duty cycle increases as the oscillator frequency decreases. For low input voltage applications, a lower switching frequency may be required to maintain regulation at minimum input voltage.

A lower switching frequency may also improve efficiency. Losses are comprised of:

AC Losses:

- Inductor Core
- MOSFET Switching
- MOSFET Gate Drive Current

DC Losses:

- Inductor I²R
- MOSFET I²R_{DS(ON)}
- Capacitor ESR

At lower frequency, AC losses are reduced proportionately. However the inductor's conduction losses increase for the same inductor size since inductance must increase proportionately to maintain the same amount of ripple current. Efficiency can be improved if inductor size is allowed to increase to allow more copper in the windings.

COMPENSATION

Proper compensation can be accomplished in many ways. The simplest compensation scheme (fig. 3, without C1) uses resistive feedback (R1) around the error amplifier. This reduces the gain of the op-amp which is in the overall loop's feedback path. This sets the overall loop gain to allow the loop's unity gain crossover to occur after the zero created by the ESR of the output capacitor has taken effect. This influences the two pole roll off created by the output inductor and the output capacitor providing the phase margin required to keep the loop stable under all conditions.

R1/RF1 sets the gain (A_V) of the loop. Based on the following assumptions a gain of 5 is sufficient for any range of input voltages:

1. $C_O = 100\mu\text{F}$ per 500mA of maximum output current.
2. Inductor ripple current is 40% of maximum output current.
3. Load capacitance (C_L) is not more than output capacitance (C_O). Load capacitance is capacitance that gets switched in and out when loads are switched in and out and C_O is the permanent output capacitance.
4. ESR of the total output capacitance is within the limits indicated in table 1. Maximum ESR shown is for medium input voltage range (9V to 18V). Higher input voltage range (12V to 24V) requires lower limits (about 30%) for the maximum ESR allowed. Lower input voltage range (6V to 12V) allows higher (40%) maximum ESR limits.
5. Table 1 is for 100KHz operation. At lower switching frequency maximum ESR must be lower by a linear ratio. (i.e. at half the frequency maximum ESR must also be half).

ESR (mΩ)		OUTPUT CURRENT (MAX)
MIN	MAX	
30	175	1
15	75	2.5
8	40	5

Table 1. Recommended ESR Values for 100Khz operation

Regulation can be greatly improved with a capacitor (C1) placed in series with the feedback resistor. Its value should be high enough to be no longer a factor at high frequency. C1 can be calculated as:

$$C_1 \geq \frac{\sqrt{L_O C_O(\text{MAX})}}{RF1}$$

Several of the components in Figure 11 may not be required depending on board layout and desired efficiency. $V(\text{GT})$ should not be allowed to go below $V(\text{IN})$ by more than 0.5V. D5 can be eliminated if D2 is replaced by a Schottky. D4 and D3 are required to avoid negative currents from being pulled from the substrate of the IC. These diodes can be replaced with 1N4148's on the gates of Q2 and Q4 if a damping resistor is used from the gate to the IC. D6 and D7 enhance the efficiency of the regulator by a small amount (about 1%) by preventing the MOSFET body diodes from turning on during the time when both halves of the MOSFET bridge are off. The energy which would be used for the reverse recovery of these body diodes is greatly reduced by using Schottkys. D1 can be either a conventional silicon rectifier or Schottky diode depending on efficiency vs. cost considerations.

Resistors R10 and R11 are 20mΩ resistors which can be made with a small length of wire or a PC board trace.

C15 and C14 may be necessary to overcome the inductance in the sense resistor and are typically 0.1μF.

C12 can be implemented using two 100μF tantalum capacitors in parallel or a 200μF electrolytic capacitor with a 0.1μF ceramic in parallel. C13 can be implemented in a similar fashion.

MOSFETS Q1-Q4 should be rated to withstand the maximum input voltage. Their on resistance will directly impact conduction losses and therefore efficiency. For board space considerations, the LittleFoot® series of MOSFETs provide a good trade-off between density and $R_{\text{DS(on)}}$.

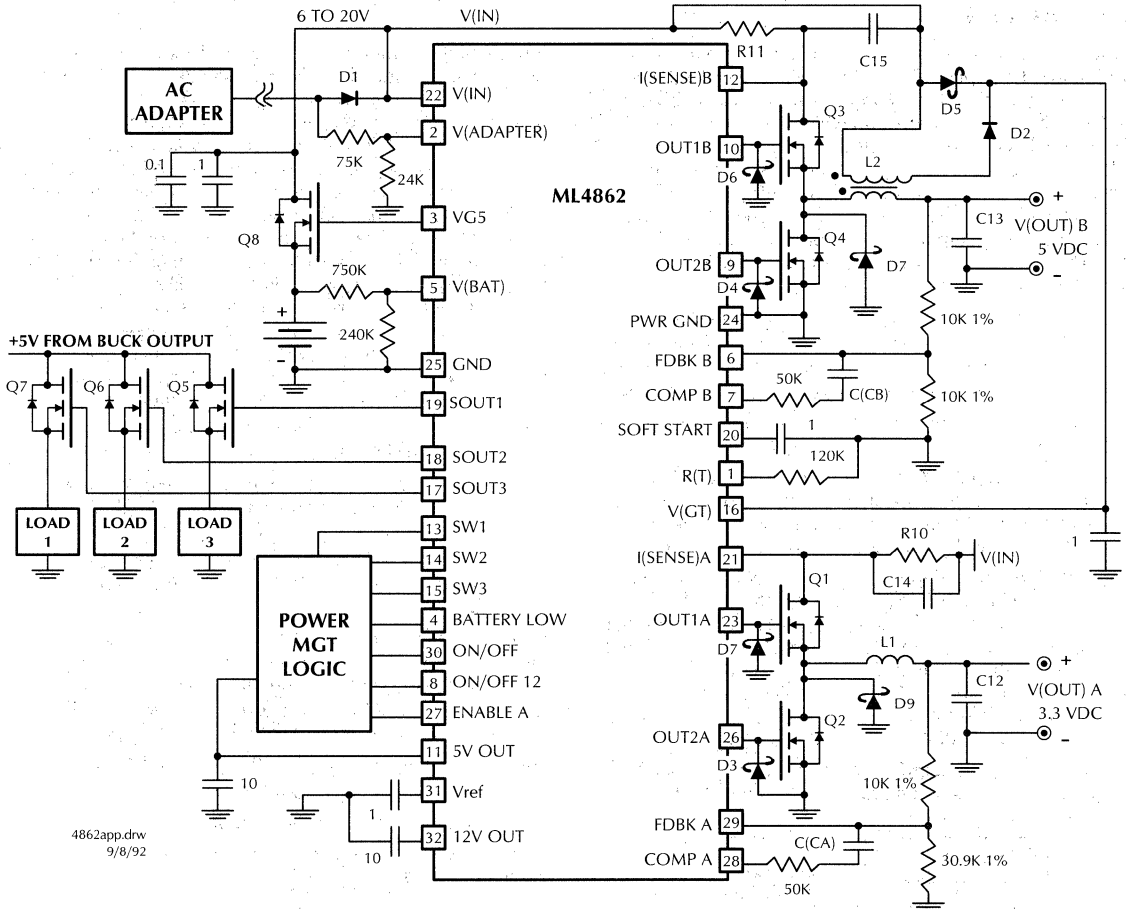
The inductor manufacturers listed above can be reached at:

Coiltronics (305) 781-8900
Sumida Electric (708) 956-0666

Careful design of the inductor can improve efficiency by trading off cost and size.

DESIGNATOR	DESCRIPTION	PART NUMBER
C14, C15	0.1uF, 50V (optional)	see text
D1	3A, 30V RECTIFIER	see text
D2	100mA, 50V (min) RECTIFIER	1N4148
D3-D7	SCHOTTKY DIODE	1N5817 or MBRS130T3
L1	47uH, 1A	SUMIDA CRD125
L2	50uH, 1.5A	COILTRONICS CTX05-11209-1
Q1-Q4	N-CHANNEL POWER MOSFET	MTD10N05E
Q5-Q7	N-CHANNEL POWER MOSFET	MMDF4N02
Q8	N-CHANNEL POWER MOSFET	Si9410
R10, R11	0.02 OHM	see text

Table 2. Circuit Values for typical application (Fig 11).



4862app.drw
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All Capacitors in μF , Schottky Diodes are 1N5817 except D6, D7 which are 1N5818 or 1N5819.

Figure 11. ML4862 Typical Application

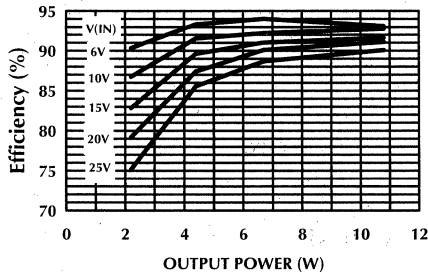


Figure 12. System Efficiency vs. Output Power

The efficiency measurements taken for figure 12 were measured with 1/3 of the output power delivered by the 3.3V regulator and 2/3 of the output power delivered the 5V output.

This circuit is available as a demonstration kit. Please contact your Micro Linear Distributor or Representative for availability and price.

ORDERING INFORMATION

PART NUMBER	TEMPERATURE RANGE	PACKAGE
ML4862CS	0°C to +70°C	32-PIN SOIC (S32N)

Other packages and temperature ranges can be made available on request. Contact your local Micro Linear Representative for more information.

6

Battery Power Controller Evaluation Kit

GENERAL DESCRIPTION

The ML4862EVAL kit provides a convenient vehicle to evaluate the ML4862 battery power control IC. It contains all of the necessary documentation with the evaluation board and key components to quickly evaluate the application circuit. The board is designed for a 11W dual output power supply. However power components can be selected for higher or lower power applications.

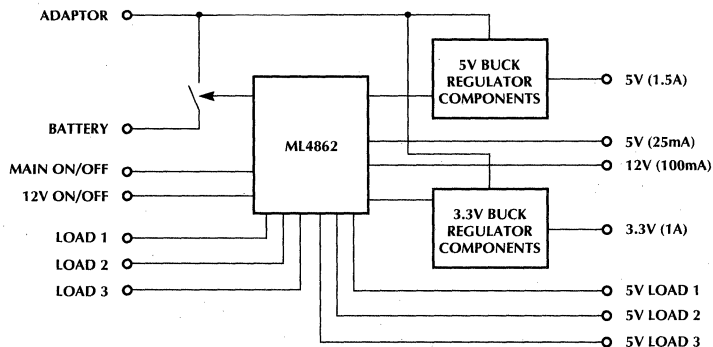
KIT COMPONENTS

- User's Guide
- ML4862 Datasheet
- Application Note
- ML4862 Sample
- Evaluation Board
- Key power semiconductor components
- SMD power inductors for both 5V and 3.3V outputs
- "Kool Mu®" toroid for inductor prototyping

FEATURES

- Wide Input Range (5.5V to 25V)
- Dual Outputs (5V at 1.5A and 3.3V at 1A)
- Efficiency as high as 95%.
- Line and Load Reg better than 5% at all conditions.
- Output Ripple to under 50mV pk-pk.
- Complete Short Circuit Protection
- Flash Memory Program Voltage (12V at 100mA)
- Load switching outputs.
- Battery connect terminals.

BLOCK DIAGRAM



USICs and Semi-Standard

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Analog and Mixed Analog/Digital Tile Arrays

Micro Linear's Tile Arrays can implement a wide range of complex circuit and performance applications. They are comprised of active and passive components arranged in organized tile patterns on an integrated circuit chip to provide maximum flexibility and ease of interconnection. After a circuit is designed and simulated, several metal patterns are created to connect the components and produce a customized analog USIC. USIC is an acronym for User Specific Integrated Circuit. USICs are integrated circuits that are specific to a customer or user and are commonly referred to as ASICs.

STANDARD AND SEMI-STANDARD PRODUCTS

The Tile Array methodology forms the foundation of many of Micro Linear's state-of-the-art product offerings. A large majority of our high volume standard products were developed using the same Tile Arrays that are used to develop high performance semi-custom solutions for customers. Standard products developed on Tile Arrays can be easily modified to meet a special need, or to gain a functional or performance advantage over a system using just standard products. This ability to change a standard product to meet a specific customer need is referred to as semi-standard products.

PROCESS TECHNOLOGIES

Micro Linear utilizes many of the latest high performance process technologies to build Tile Arrays. The highest performance and complexity is obtained with our BiCMOS process. This technology combines fast bipolar transistors with dense CMOS devices. With NPN cutoff frequencies of 4 GHz and channel lengths of 1.5 μ , this process allows the implementation of high complexity and high performance semi-custom devices. Poly resistors offer stable resistive elements with virtually no parasitics as compared to diffused resistors. The poly-metal capacitors are also very stable and offer a superior alternative over conventional metal-silicon types. The FC3500 family of general purpose tile arrays are built with the BiCMOS process.

Two different bipolar technologies are used with distinct voltage and speed characteristics. A 12 volt, 1 GHz technology and a 36 volt, 300 MHz technology. The 12 volt process is the higher speed process of the two, capable of implementing circuit band-widths up to 100 MHz. The 36 volt process is useful where higher power supply voltage operation is needed. The FB3600 family is built with the 12 volt bipolar process and the FB3400 is built with the 36 volt process.

Please refer to the Tile Array Process Technologies and Component Performance heading for more information.

GENERAL PURPOSE TILE ARRAYS

Micro Linear has three families of general purpose Tile Arrays. The arrays are configured with groupings of mini-tiles. Each mini-tile consists of selections of various types of active and passive components. Each of the Tile Arrays has a different mix and amount of mini-tiles.

The three families are:

FC3500 – 5 Volt BiCMOS
FB3600 – 12 Volt Bipolar
FB3400 – 36 Volt Bipolar

The FC3500 family of Tile Arrays, built with 5 volt, 4 GHz, 1.5 μ process, is capable of the highest performance and level of complexity. This is primarily due to the small size and high cutoff frequency (Ft) of the transistors in this technology. The FC3500 general purpose family of Tile Arrays is ideal where the circuit needs the high speed and accuracy of the bipolar transistors along with the dense logic, analog switch or FET input capability of the CMOS devices and operates from a single 5 volt power supply.

The FB3600 family offers a wide range of general purpose Tile Arrays. Ranging from the smallest, the FB3605, which can accommodate 4 analog circuit blocks and 28 digital gates to the largest pure analog, the FB3630, which can accommodate over 20 analog functional circuit blocks. The FB3635 can accommodate 9 analog circuit blocks in addition to 130 digital gates. For high current output applications the FB3623 has four 0.5A output transistors.

The FB3400 family is used when the power supply voltage in the system is greater than 12 volts. For example, common supply voltages of +/- 15 volts would require use of the FB3400 family.

APPLICATION FOCUSED TILE ARRAYS

Application focused tile arrays are optimized for a given application. Areas on the tile arrays are customized to implement the common functions required for a given application. In this way, higher performance can be achieved for these critical circuit blocks. These application focused tile arrays still maintain all the advantages of the general purpose tile arrays. All the individual components are uncommitted until connected with the two metal layers used in the final step of the wafer manufacturing process. A summary of the currently available application focused tile arrays is as follows:

USICs

APPLICATION FOCUSED TILE ARRAY

APPLICATION AREA

FC3560	Mass Storage Read Channel
FC3580	Micro Power Controller
FB3680	Electronic Ballast with Power Factor Correction
FB3651	Local Area Network (LAN) Transceiver
FB3480	Power Supply Controller
FB3490	PWM Controller/Power Factor Correction
FB3491	Resonant Mode Controller
FB3492	Phase Modulation Controller

PACKAGING

Many types of packaging are available for the Tile Arrays. Devices can be assembled, tested and produced in high volume in dual-in-line (DIP), chip carriers (PLCC, LCC), small outline, or gull wing (SOIC), as well as the latest fine pitch, very low profile quad flat packs (QFP, TQFP) and very small outline packages (SSOP). Tape and reel format shipments are available for automated assembly requirements. Please refer to the package selection guide in this section.

DESIGN METHODOLOGY

Micro Linear has extensive experience designing and producing state-of-the-art analog and mixed analog digital USICs. The key aspects of our successful methodology include: computer simulations with extremely accurate component models; in depth design reviews with the customer; expert mask layout with computer checking for accuracy; and complete packaging and testing capability.

COMPUTER SIMULATIONS

The expected performance of the finished IC is verified with computer simulations. Extensive simulations are performed with the latest workstation hardware running Micro Linear's proprietary SPICE simulator. The simulations are performed over the worst case conditions of temperature and power supply voltages as well as the expected process variations, to ensure a reliable and manufacturable device.

The accuracy of the simulations are strongly dependent on the accuracy of the component simulations models. Micro Linear device simulation models are derived from extensive characterization measurements on finished devices. They are verified by matching the characteristics the simulation program predicts versus regularly updated measurements from production devices.

DESIGN REVIEWS

A thorough design review can have a significant impact on the success of a USIC development. At Micro Linear detailed design reviews are conducted with the customer engineer and several Micro Linear engineers in attendance. All aspects of the design are reviewed including system performance, circuit schematics, simulation results, layout plan, test strategies, and package options. Upon completion of the review both the customer and Micro Linear have a high confidence for completely functional prototypes.

USIC DEVELOPMENT PROGRAM

A brief summary of the steps involved in developing a USIC with Micro Linear are:

1. FORWARD INFORMATION TO MICRO LINEAR
Customer provides block diagrams, discrete component schematics, and specification requirements on prospective USIC to Micro Linear for review.
2. TECHNICAL REVIEW AND QUOTE
Micro Linear reviews the customers block diagram, schematics, and specifications and responds with technical assessment, costs and development times.
3. FINALIZE DEVELOPMENT PLAN
Micro Linear and the customer agree on the development plan.
4. FINALIZE CIRCUIT REQUIREMENTS
Micro Linear consults closely with the customer to completely understand the functional and performance requirements.
5. CIRCUIT DESIGN AND SIMULATION
Micro Linear performs the circuit and the verification simulations.
6. DESIGN REVIEW
Customer participates in a detailed design review with Micro Linear engineers.
7. MASK LAYOUT
Micro Linear performs the mask layout for the circuit on the chosen array.
8. INTEGRATION
Micro Linear performs the integration steps including; Mask fabrication, wafer processing, wafer test, packaging, and prototype testing.
9. DELIVER PROTOTYPES
Micro Linear delivers prototypes to the customer.
10. CUSTOMER EVALUATES PROTOTYPES

Semi-Standard Products

Semi-standard analog ICs are created by modifying existing standard products. Modifications can span all the way from a functional circuit change to simple variation in the part marking. Examples of the range of possible semi-standard modifications are as follows:

CIRCUIT MODIFICATIONS

- Modify Functional Blocks
- Change Input/Output Levels
- Adjust Gain/Thresholds
- Shift Logic Levels (ECL, TTL, CMOS)
- Move Current Limit Point
- Increase/Reduce Bandwidth
- Improve Power Consumption
- Tweak Charge/Discharge Currents

OTHER SEMI-STANDARD MODIFICATIONS

- Change Part Marking
- Extend the Temperature Range
- Add Electrical Tests
- Modify the Pinout
- Change the Package

Circuit modifications are easily made to Micro Linear's standard products because they are built using our proprietary Tile Array technology. Tile Arrays are collections of active and passive components arranged in a pattern on an integrated circuit chip. Each standard product is developed by designing two layers of metal interconnect to implement the specific circuit functions required for the product.

SEMI-STANDARD DEVELOPMENT

A semi-standard circuit can be easily and quickly developed from a standard product. The customer need only define the functional, performance or physical nature of the modifications necessary to the Micro Linear standard product to meet his/her system requirements. Micro Linear will then accomplish the circuit design and complete the performance verification simulations. At this point a design review is held with the customer to make sure all aspects of the development are in accordance with the customer requirements. Micro Linear then proceeds with making the necessary layout changes, manufacturing, testing, and delivering the prototypes. This complete development flow is illustrated in Figure A.

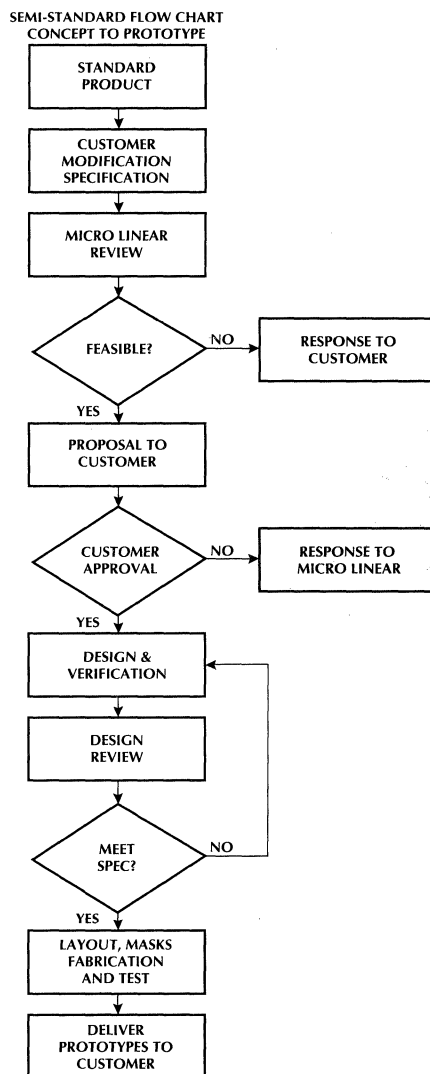


Figure A.

USICs

Micro Linear produces standard products using BiCMOS and bipolar Tile Arrays and CMOS standard cell technologies. The Tile Array technology allows the widest range functional and performance modification

possibilities. The CMOS standard cell technology, while not flexible for functional modifications, can still accommodate performance, specification, or physical changes. A summary of the types of modifications possible for each group of standard products is shown in Table B.

TABLE B

	TECHNOLOGY	FUNCTIONAL MODIFICATIONS	PERFORMANCE MODIFICATIONS	SPECIFICATION OR PACKAGING CHANGES
POWER AND MOTION CONTROL				
High Frequency PWM Control	BT	Yes	Yes	Yes
Resonant and Phase Modulation	BT	Yes	Yes	Yes
Power Factor Control	BT	Yes	Yes	Yes
Notebook/Laptop Power	BT	Yes	Yes	Yes
Low Voltage Regulators	BC	Yes	Yes	Yes
Sensorless Motor Control	BT	Yes	Yes	Yes
DATA COMMUNICATIONS				
Fiber Optic LED Drivers	BT	Yes	Yes	Yes
Fiber Optic Quantizer	BT	Yes	Yes	Yes
10Base-T Trans: AUI/MPR	BT	Yes	Yes	Yes
10Base-FL Transceivers	BT	Yes	Yes	Yes
10Base-T Physical Interface	CS	Limited	Limited	Yes
High Speed LED Drivers	BC	Yes	Yes	Yes
High Speed Quantizer	BC	Yes	Yes	Yes
AUI-MUX	BT	Yes	Yes	Yes
MLT-3 Transceiver	BC	Yes	Yes	Yes
Token Ring Physical Interface	BC	Yes	Yes	Yes
ANALOG TELECOM/NCTE				
Signal Equalizers	CS	No	Yes	Yes
Gain/Attenuators	CS	No	Yes	Yes
Sine Wave Generators	CS	No	Yes	Yes
Tone Detectors	CS	No	Yes	Yes
HARD DISK DRIVES				
Pulse Detectors	BT	Yes	Yes	Yes
Data Separators	BT	Yes	Yes	Yes
Dedicated Servo Control	BT	Yes	Yes	Yes
Voice Coil Driver	BT	Yes	Yes	Yes
Read Channel Combo	BC	Yes	Yes	Yes
Area Detection Servo	BT	Yes	Yes	Yes
Filter/Equalizer	CS	Limited	Limited	Yes
DATA CONVERSION & FILTERING				
8, 10 & 12 A/D Converters	CS	No	Limited	Yes
8-Bit D/A Converters	CS	No	Limited	Yes
Switched Cap Filter	CS	No	Limited	Yes

(Technology Code: BT = Bipolar Tile Arrays; BC = BiCMOS Tile Arrays; CS = CMOS Standard Cells)

Tile Array Process Technology and Component Performance

Micro Linear currently utilizes three advanced process technologies to manufacture Tile Array products. They are:

FC3500 — 5 volt BiCMOS

FB3600 — 12 volt Bipolar

FB3400 — 36 volt Bipolar

A comparison of some of the key parameters of the process and the components is shown in the tables below. **NOTE: All numbers are typical.**

PARAMETER	FC3500	FB3600	FB3400
Nominal Operating Voltage	5 volts	12 volts	36 volts
NPN Cutoff Frequency (f_T)	4GHz	1GHz	300MHz
PNP Cutoff Frequency (f_T)	50MHz	12MHz	5MHz
NPN Beta — Typical	100	120	140
PNP Beta — Typical	30	40	40
Resistor Matching — Best	0.5%	0.5%	2%
Resistor Matching — Typical	1-3%	2-4%	2-4%
Primary Resistor Type	Poly	Diffused	Diffused

CIRCUIT PARAMETER	FC3500	FB3600	FB3400
Amplifier Bandwidth	350MHz	100MHz	30MHz
Op-Amp Bandwidth	60MHz	20MHz	8MHz
Offset Voltages — Untrimmed	± 0.5 mV	± 1 mV	± 3 mV
Offset Voltages — Trimmed	± 0.1 mV	± 0.1 mV	± 0.1 mV
Bipolar ECL Gate Delays	0.6ns	3ns	10ns
MOS Gate Delays	1ns	N/A	N/A
Voltage Reference Accuracy — Trimmed	$\pm 0.05\%$	$\pm 0.05\%$	$\pm 0.05\%$

FC3510 General Purpose BiCMOS Tile Array

GENERAL DESCRIPTION

The FC3510 is one member of a high speed, high complexity family of General Purpose Tile Arrays using our new 5 volt, 4 GHz BiCMOS technology. These BiCMOS Tile Arrays allow high speed, high complexity, cost effective circuits to be easily integrated. The Tile Array methodology consists of simply designing and manufacturing two metal mask layers to implement the required circuitry on pre-manufactured base arrays. The base arrays contain all of the pre-defined and characterized components.

The BiCMOS process combines a fast 4 GHz analog bipolar technology with a dense 1.5 μ CMOS digital capability. This powerful combination enables a high level of integration of mixed analog and digital circuits to be achieved on a single silicon device.

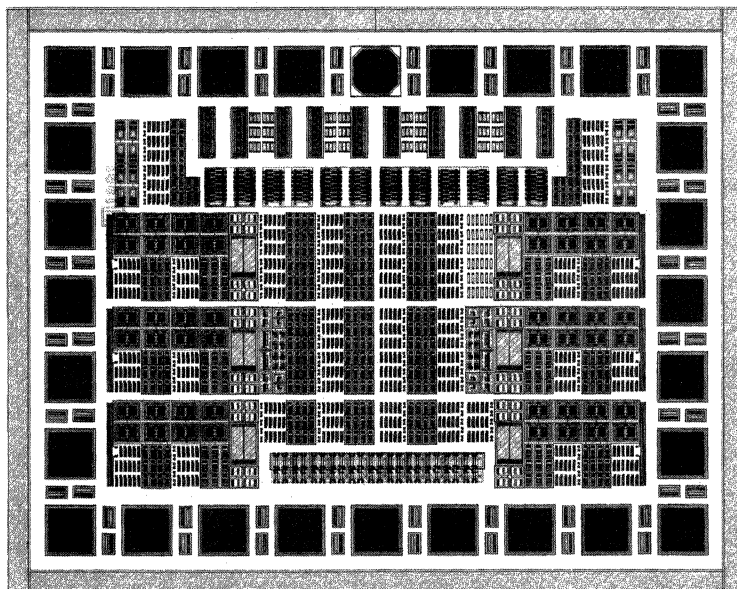
The array consists of different types of mini tiles. Each mini tile is a collection of a specific type of component such as NPNs, PNPs, NMOS or PMOS transistors, poly resistors, MOS capacitors, gates, etc. The FC3510 contains approximately 600 active devices and over 2.5M Ω of poly resistance. This array can realize up to 12 analog functional blocks combined with 22 digital gates.

FEATURES

- 5 volt, 4 GHz/1.5 μ BiCMOS Technology
- Ideal for Very High Speed, High Complexity Circuits
- Can Integrate 12 Analog Circuit Blocks with 22 CMOS Gates
- Fast, Low Risk Circuit Development and Production using the Tile Array Technology

ARRAY SUMMARY

NPN Transistors	364
PNP Transistors	96
NMOS Transistors	68
PMOS Transistors	68
Total Poly Resistance	2620K
Total MOS Capacitance	30pF
Total Components	1460
Bond Pads	28
Die Size (mils)	70 x 88.5



FC3510 BiCMOS Tile Array

FB3600 Tile Array Selection Guide

Array		FB3605	FB3610	FB3620	FB3621	FB3622	FB3623	FB3630	FB3631	FB3635
Description		Small High Frequency	Small General Purpose	Medium General Purpose	Medium High Frequency	Medium Power Schottky	Medium High Power	Large General Purpose	Large Mixed Analog/Digital	Large Mixed Analog/Digital
Mini Tile Summary										
T1	General	10	48	64	48	48	64	112	92	36
T1A	General	10								
T2	Specialized	2	6	12	8	12	12	24	12	4
T2A	Specialized	2								
T3	Power	2	4	4	2		4	4	4	2
T4	Low Noise		4	4	4		2	4		
T5	Precision			4			4	8		2
T6	NPN Intensive				8					8
T7	High Frequency				12	12			4	12
T8	Schottky Core									4
T9	ECL Logic	10							22	42
T10	ECL Logic Bias								1	1
T11	TTL Output	8			4				8	8
T12	Schottky Peripheral									2
T13	High Frequency	4								
T14	High Power NPN						4			
T15	Medium Current PNP						4			
T16	Power Schottky					14				
T17	General								1	
Array Summary										
Complexity*										
Analog		4	6	12	8	12	12	24	12	9
Digital		28							62	130
NPN Transistors		260	178	268	329	276	272	472	690	901
PNP Transistors		32	78	124	88	108	132	232	154	63
Schottky Transistors		16			8	14			16	48
Total Diffused Resistance		240K	288K	425K	495K	432K	425K	768K	850K	818K
Total Implant Resistance		816K	1563K	2048K	2064K	1920K	2048K	3584K	3928K	3064K
Total MOS Capacitance		20pF	30pF	60pF	40pF	60pF	60pF	120pF	60pF	40pF
Total Components		840	742	1092	1508	1370	1360	1944	2806	2805
Bond Pads		24	24	32	32	28	27	46	44	44
Die Size (mils)		70 × 110	82 × 102	102 × 115	102 × 115	112 × 125	115 × 122	131 × 150	142 × 156	131 × 150

* Analog complexity is in one 741 op-amp or two 339 comparator equivalents.
Digital complexity is in two input NAND gate equivalents.

(See also Power Supplies Arrays in section 6)

USICs

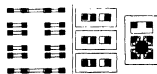
Component Performance

COMPONENT	FB3600 FAMILY
NPN Transistor	$h_{FE} = 120$ $f_T = 750\text{MHz}$ $BV_{ce0} = 14\text{V}$
NPN Large Transistor (FB3623 Only)	$h_{FE} = 100$ $f_T = 750\text{MHz}$ $BV_{ce0} = 14\text{V}$ $I_C = 0.5\text{A}$
PNP Substrate Transistor	$h_{FE} = 60$ $f_T = 24\text{MHz}$ $BV_{ce0} = 25\text{V}$
PNP Lateral Transistor	$h_{FE} = 30$ $f_T = 12\text{MHz}$ $BV_{ce0} = 25\text{V}$
Diffused Resistor	2% matching with $\pm 20\%$ absolute value
Precision Resistors	0.5% matching with $\pm 20\%$ absolute value
Implant Resistors	4% matching with $\pm 25\%$ absolute value
MOS Capacitor	$\pm 20\%$ absolute value

FB3600 Mini Tile Description

T1 Mini Tile

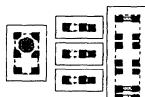
The T1 mini tile contains the components that constitute the major portion of most analog designs. It can implement many common building block functions such as current mirrors, differential gain stages and level shifters. It is the most general-purpose analog building block and has the greatest frequency of placement within the arrays.



Components	Qty.
MLC3600 minimum NPN	3
MLC3611 lateral PNP	1
MLC3620 750Ω base resistor	8
MLC3621 4KΩ implant resistor	4
MLC3622 8KΩ implant resistor	2

T1A Small T1 Mini Tile

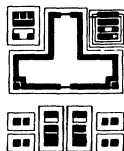
The T1A mini tile is a slightly modified T1 general purpose mini tile. It has the same active device count but has about half the resistor segments. Just like the T1 mini tile this tile is configured to implement many of the fundamental circuit blocks in integrated circuit design such as current mirrors, differential amplifiers and level shifters.



MLC3600 minimum NPN	3
MLC3611 lateral PNP	1
MLC3620 750Ω base resistor	4
MLC3621 4KΩ implant resistor	2

T2 Mini Tile

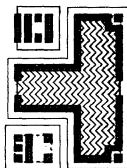
Many analog building blocks, such as op-amps, use a T2 mini tile along with multiple T1 mini tiles. This mini tile has the second greatest frequency of placement within the arrays. The MLC3630 is a silicon dioxide capacitor whose value can be programmed up to 5pF. This capacitor is often used to provide on-chip compensation for operational amplifiers.



MLC3601 circular emitter NPN	2
MLC3602 three emitter NPN	1
MLC3610 minimum vertical PNP	4
MLC3612 2X vertical PNP	1
MLC3630 5pF MOS capacitor	1

T2A Small T2 Mini Tile

This mini tile is very similar to but smaller than the T2 mini tile. This mini tile is typically used in conjunction with several T1 mini tiles. Operational amplifiers that use an on-chip compensation capacitor are built with a T2 or T2A and T1 mini tiles.



MLC3602 three emitter NPN	1
MLC3612 2X vertical PNP	1
MLC3630 5pF MOS capacitor	1

T3 Mini Tile

This mini tile contains two NPN power transistors for output stages driving up to 100mA each. These mini tiles are located in peripheral positions around the chips, in close proximity to the bonding pads.



MLC3605 power NPN	2
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T4 Mini Tile

This mini tile contains two large six emitter low noise NPN transistors. These transistors are used in circuits which require noise performance of less than $5nV\sqrt{Hz}$. These transistors can also be medium capacity power transistors.

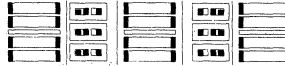


MLC3604 6X low noise NPN	2
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USICs

T5 Mini Tile

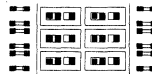
This mini tile contains six minimum geometry NPN transistors and twelve precision resistor links. These special resistors have a nominal ohmic value of 850Ω and are matched to within an accuracy of 0.5%. It is possible to construct R-2R ladders to be used in the core of an 8-bit DAC by using two T5 mini tiles.



Components	Qty.
MLC3600 minimum NPN	6
MLC3625 850Ω precision resistor	12

T6 Mini Tile

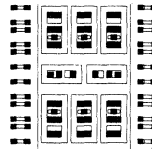
The T6 mini tile is designed for NPN intensive transistor circuit design. ECL logic can be implemented with this tile.



MLC3600 minimum NPN	6
MLC3620 750Ω base resistor	8
MLC3621 4KΩ implant resistor	4

T7 Mini Tile

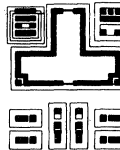
This mini tile contains six dual base contact 10mA high speed, low noise NPN transistors. Each of these small transistors has a 50Ω base resistance. Two T7 tiles can implement 100MHz cascode amplifier or a 60MHz video amplifier.



MLC3600 minimum NPN	2
MLC3606 double base NPN	6
MLC3620 750Ω base resistor	16
MLC3621 4KΩ implant resistor	8

T8 Mini Tile

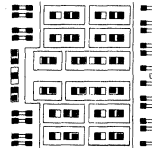
The T8 mini tile contains a mixture of schottky and other components for analog design. The schottky devices are useful for clamping signal levels and in certain high speed comparator designs.



MLC3607 minimum schottky NPN	2
MLC3602 three emitter NPN	1
MLC3612 2X vertical PNP	1
MLC3608 minimum schottky diode	4
MLC3630 5pF MOS capacitor	1

T9 Mini Tile

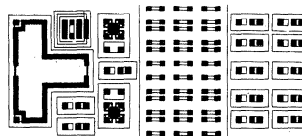
This mini tile contains one basic ECL logic cell which can implement one data latch (with set & reset). Two basic ECL logic cells can implement one edge triggered D-type flip-flop (with set and reset). An alternative usage for one basic ECL logic cell would be to implement three 2-input ECL gates, two 4-input ECL gates or one 8-input ECL gate. The actual gate can be a NAND, AND, OR or NOR gate. Two of the minimum NPNs have their collectors tied to V_{CC}.



MLC3600 minimum NPN	8
MLC3600A two minimum NPNs	2
MLC3600B minimum NPN	2
MLC3600C minimum diode	2
MLC3620 750Ω base resistor	12
MLC3621 4KΩ implant resistor	8

T10 Mini Tile

The T10 mini-tile provides the necessary temperature adjusted reference voltages for biasing the ECL logic. Unlike analog voltage references, ECL logic needs the bias reference to vary with temperature. This mini tile has its function pre-defined for most applications.



MLC3600 minimum NPN	13
MLC3611 lateral PNP	2
MLC3612 2X vertical PNP	1
MLC3620 750Ω base resistor	30
MLC3621 4KΩ implant resistor	6
MLC3630 5pF MOS capacitor	1

T11 Mini Tile

This mini tile contains components for building an output buffer for a TTL or CMOS output stage (TTL fanout of 2). The mini tile can convert the on-chip ECL logic levels to TTL or CMOS logic levels.



MLC3600 minimum NPN	2
MLC3609 6X schottky NPN	1
MLC3613 2X schottky NPN	1
MLC3620 750Ω base resistor	5
MLC3621 4KΩ implant resistor	1
MLC3624 50Ω emitter resistor	1

T12 Mini Tile

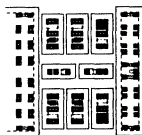
The T12 mini tile contains additional Schottky components for building TTL or CMOS input or output buffers.



MLC3607 minimum schottky NPN	2
MLC3609 6X schottky NPN	2

T13 High Frequency Mini Tile

High frequency circuit design requires specially designed components. This mini tile contains transistors with low internal base resistance and low value, small area base resistors essential for high frequency circuit design. The low value resistors are critical in high frequency design as load resistors to achieve a practical output swing with the high currents necessary to maximize the bandwidth of the transistors. The small area of these resistors is also very important because it minimizes the parasitic capacitance that limits high frequency performance. This mini tile contains minimum NPN transistors, circular NPN transistors, double base NPN transistors, base, implant, and small area base resistors.



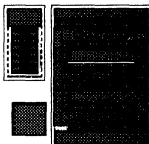
Components

Qty.

MLC3600 minimum NPN	2
MLC3601 circular emitter NPN	2
MLC3606 double base NPN	4
MLC3620 750Ω base resistor	8
MLC3621 4KΩ implant resistor	4
MLC3629 225Ω base resistor	6

T14 High Power NPN Mini Tile

The T14 mini tile contains a high current NPN power transistor and a clamp diode. These devices along with the PNP transistors on the T15 mini tile can implement a high power output stage with surge protection. The NPN power transistor can handle 0.5 Amps of current. The clamp diode is connected to protect the large NPN from transient surge voltages and currents.



MLC3605A high power NPN (0.5A)	1
MLC3609A clamp diode	1

T15 Medium Current PNP Mini Tile

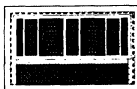
This mini tile has the PNP transistors that work in conjunction with the power devices in the T14 mini tile to implement an output stage. These PNP transistors can supply the necessary base drive current, up to 10 mA, for the large NPN transistors. Also on this mini tile is a smaller PNP that is a scaled version of the large PNP. This PNP is typically connected with the larger PNP as a current mirror with a gain of nine.



MLC3616 36X lateral PNP	1
MLC3615 4X lateral PNP	1

T16 Power Schottky Mini Tile

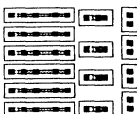
The T16 mini tile consists of one schottky NPN transistor capable of handling 120 mA. This schottky device is useful in the design of output stages that need to drive high current pulses into magnetic heads. Switching speed is enhanced over a regular power transistor because saturation, and the resulting speed degradation, is avoided with the schottky transistor.



MLC3631 large Schottky NPN	1
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T17 General Purpose Mini Tile

The T17 mini tile is used for general purpose design. It contains minimum NPN and minimum vertical PNP transistors and implant resistors.



MLC3600 minimum NPN	4
MLC3610 minimum vertical PNP	4
MLC3621 4KΩ implant resistor	12
MLC3622 8KΩ implant resistor	6



FB3605 Small High Frequency Tile Array

GENERAL DESCRIPTION

The FB3605 is a bipolar analog tile array developed for mixed analog and digital applications that require high frequency performance. This FB3600 family tile array utilizes our proprietary mini tile architecture. The mini tile approach combined with our 12 volt, 1 GHz technology allows high complexity, high speed circuits to be easily integrated.

High frequency circuits like a 90 MHz voltage controlled oscillator or other similar performance circuits can be integrated using the FB3605 tile array. In addition to this particular circuit block the array can also contain 4 full function op amps (741 type) as well as 28 gates of ECL logic and 8 digital output buffers capable of interfacing to ECL, TTL, or CMOS.

A new high frequency mini tile was designed for this array. This mini tile contains 4 double base, high frequency NPNs and 6 low value (225 ohm) base resistors along with other NPNs and resistors. The double base NPNs and the low value base resistors are the key to implementing high frequency circuits.

The FB3605 is the smallest of the FB3600 family of arrays. The small die, 70 by 110 mils, allows it to fit into a very small package. The FB3605 can be assembled in a 0.15 inch wide SOIC package for minimum board space.

FEATURES

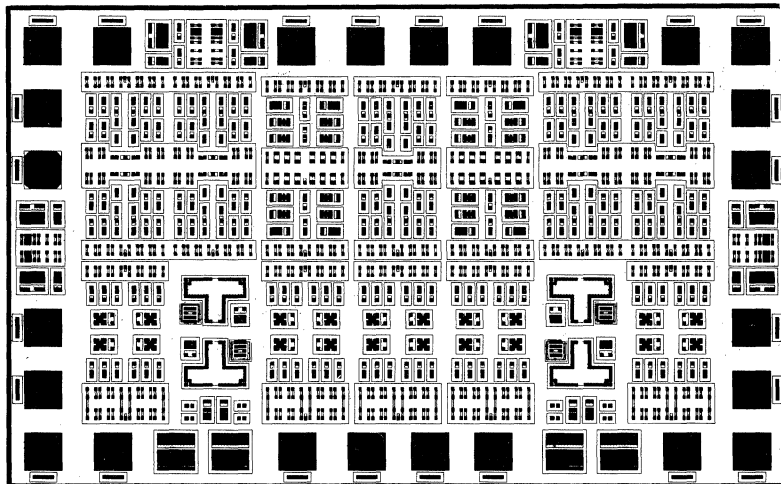
- High frequency operation
- Small die size — Fits in narrow SOIC Package
- Mixed Analog and Digital circuitry
- 5 analog circuit blocks with 28 ECL gates
- On chip MOS capacitors
- 12 volt, 1 GHz technology

ARRAY SUMMARY

NPN Transistors	260
PNP Transistors	32
Schottky Transistors	16
Total Diffused Resistance	240 K
Total Implant Resistance	816 K
Total MOS Capacitance	20 pF
Total Components	840
Bond Pads	24
Die Size (mils)	70 × 110

MINI TILE SUMMARY

T1 General	10
T1A Small T1	10
T2 Specialized	2
T2A Small T2	2
T3 Power	2
T9 ECL	10
T11 TTL Output	8
T13 High Frequency	4



FB3605 — Small High Frequency Tile Array

FB3610, FB3620, FB3630 General Purpose Tile Arrays

GENERAL DESCRIPTION

The FB3610, FB3620, and FB3630 are general purpose analog tile arrays capable of implementing a wide range of circuit functions. These FB3600 family arrays use our proprietary mini-tile architecture. The mini-tile approach combined with our 12-volt, 1 GHz technology allows high complexity, high speed circuits to be easily integrated.

Each of these general purpose arrays have the same basic structure. The difference is in the number of mini-tiles and therefore the number of total components available on each array. The different arrays can incorporate differing levels of circuit complexities. The FB3610 is the smallest and able to contain approximately six full function operational amplifiers or twelve comparators. The largest general purpose array, the FB3630, can incorporate 24 operational amplifiers or as many as 48 comparators.

Both the FB3620 and FB3630 contain precision resistor mini-tiles which allows precision circuits to be integrated on these arrays. The typical resistor match of 0.5% enables an 8-bit DAC to be implemented. All three of these arrays also contain low noise and power devices. The low noise transistors allow circuits with less than $5nV/\sqrt{Hz}$ input referred noise to be realized. The power transistors can supply up to 100mA each and can be paralleled for higher currents. Other FB3600 arrays can achieve up to 2 amps.

High performance circuits can be implemented on these arrays. Amplifiers with bandwidths up to 70 MHz and voltage controlled oscillators up to 50 MHz can be implemented on the FB3610, FB3620 or FB3630 arrays. Higher frequency performance can be achieved on other FB3600 family tile arrays.

FEATURES

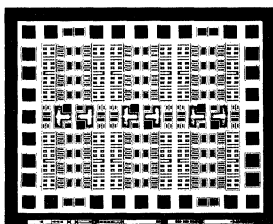
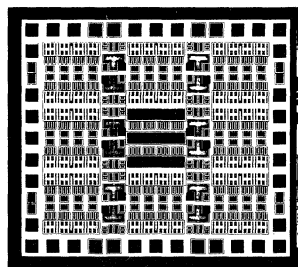
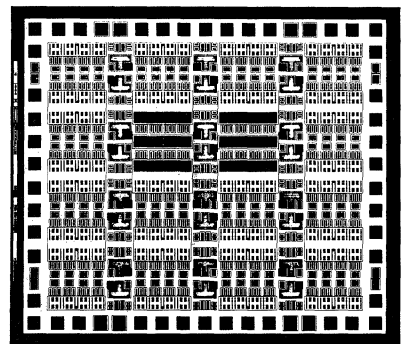
- High complexity and high performance
- Operates with supplies up to 12 volts, $\pm 10\%$
- Flexible mini-tile architecture
- Precision and high current components
- 12 volt, 1 GHz technology

ARRAY SUMMARY

	FB3610	FB3620	FB3630
NPN Transistors	178	268	472
PNP Transistors	78	124	232
Total Diffused Resistance	288K	425K	768K
Total Implant Resistance	1563K	2048K	3584K
Total MOS Capacitance	30pF	60pF	120pF
Total Components	742	1092	1944
Bond Pads	24	32	46
Die Size (mils)	82 × 102	102 × 115	131 × 150

MINI-TILE SUMMARY

	FB3610	FB3620	FB3630
T1 General	48	64	112
T2 Specialized	6	12	24
T3 Power	4	4	4
T4 Low Noise	4	4	4
T5 Precision		4	8


FB3610

FB3620

FB3630

FB3621 Medium High Frequency Tile Array

GENERAL DESCRIPTION

The FB3621 array is ideal for applications that have high frequency and low noise requirements. This FB3600 family tile array utilizes our proprietary mini-tile architecture. The mini tile approach combined with our 12-volt, 1 GHz technology allows high complexity, high speed circuits to be easily integrated.

High frequency and low noise circuits require transistors with low parasitic base resistance. This array contains a large number of transistors with dual base contacts and therefore low base resistance. In addition to these high frequency/low noise NPNs the array has a high percentage of regular NPN devices. The high overall number of NPN transistors enables a large number of high frequency circuit blocks to be implemented. ECL logic, which uses mostly NPN devices, can also be integrated on this array.

Typical types of circuit functions that can be implemented on the FB3621 array are 100 MHz cascode amplifiers, VCOs, wideband/low noise amplifiers, and high speed comparators ($T_D < 5\text{ns}$).

FEATURES

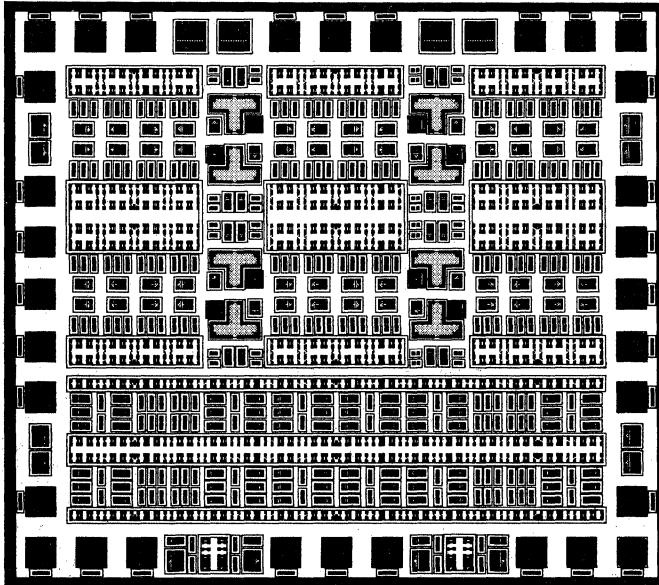
- High frequency operation
- Low noise circuits
- On-chip MOS capacitors
- 12 volt, 1 GHz technology

ARRAY SUMMARY

NPN Transistors	329
PNP Transistors	88
Schottky Transistors	8
Total Diffused Resistance	495K
Total Implant Resistance	2064K
Total MOS Capacitance	40pF
Total Components	1508
Bond Pads	32
Die Size (mils)	102 × 115

MINI TILE SUMMARY

T1 General	48
T2 Specialized	8
T3 Power	2
T4 Low Noise	4
T6 NPN Intensive	8
T7 High Frequency	12
T11 TTL Output	4



FB3621 — Medium High Frequency Tile Array

FB3622 Medium Power Schottky Tile Array

GENERAL DESCRIPTION

The FB3622 is a bipolar analog tile array developed for applications that require fast high current outputs. This FB3600 family tile array utilizes our proprietary mini tile architecture. The mini tile approach combined with our 12 volt, 1 GHz technology allows high complexity, high speed circuits to be easily integrated.

Around the perimeter of the die are fourteen large schottky NPN power transistors. These power transistors, capable of handling over 100 mA each, make this array ideal for applications which call for driving inductive loads with high currents such as magnetic write heads. Twelve high frequency tiles expand the capabilities of this array. These high frequency tiles allow high frequency circuit blocks to be included on this array.

This array can accommodate eighteen functional blocks of the approximate complexity of a 324 operational amplifier.

FEATURES

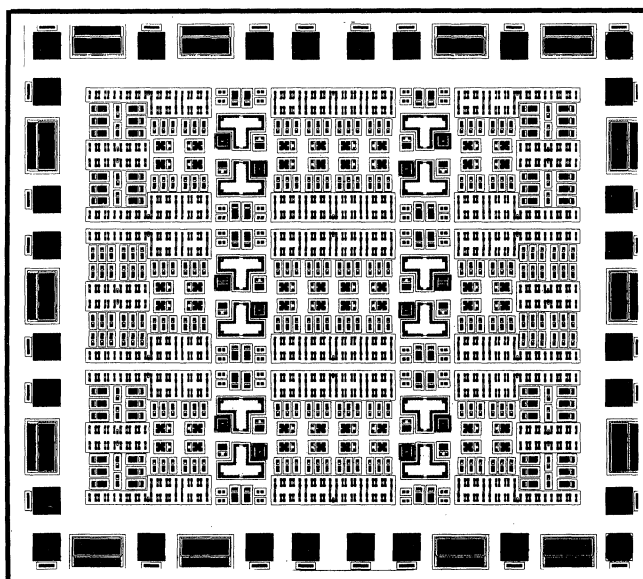
- 14 power schottky NPN transistors
- Design complexities of 18 functional blocks
- Precision and high frequency mini tiles
- On chip MOS capacitors
- 12 volt, 1 GHz technology

ARRAY SUMMARY

NPN Transistors	276
PNP Transistors	108
Schottky Transistors	14
Total Diffused Resistance	432 K
Total Implant Resistance	1920 K
Total MOS Capacitance	60 pF
Total Components	1370
Bond Pads	28
Die Size (mils)	112 × 125

MINI TILE SUMMARY

T1 General	48
T2 Specialized	12
T7 High Frequency	12
T16 Power Schottky	14



FB3622 — Medium Power Schottky Tile Array

FB3623 Medium High Power Tile Array

GENERAL DESCRIPTION

The FB3623 is a bipolar analog tile array capable of handling up to 2 amps of current. This FB3600 family tile array utilizes our proprietary mini tile architecture. The mini tile approach combined with our 12 volt, 1 GHz technology allows high complexity, high speed circuits to be easily integrated.

The array has two distinct sections, one with the power output devices and the other for general purpose circuits. The general purpose section, the larger area, is configured for circuits that can be designed using the various types of mini tiles. Op Amps, comparators, video amplifiers, voltage controlled oscillators, analog multiplexers, and mixers, are some examples of the types of circuit functions that can be realized using these mini tiles.

The other section of the array has components designed for high current output stages. Consisting primarily of four 0.5 A power NPNs, four 10 mA lateral PNPs, and four high current clamp diodes, this area can integrate output stages with a wide variety of configurations such as a four by 0.5A, two by 1 A, or 1 by 2 A for examples. The clamp diodes protect the output transistors from spurious transient signals on the output.

The two sections are separated by a diffused region which minimizes any coupling from the output transistor section back into the rest of the circuit which might cause problems due to high gain or low signal levels.

FEATURES

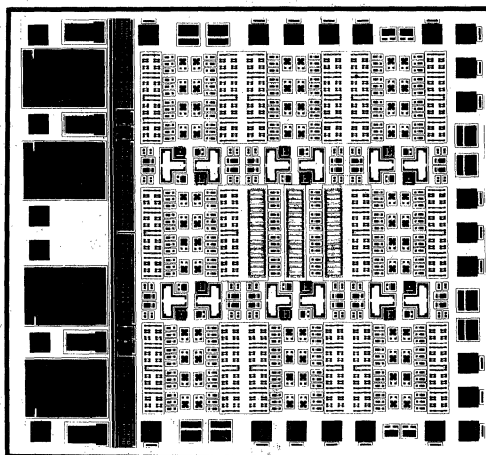
- High Current Capability — up to 2 A
- Mixed High Power and Low Level Circuits
- High complexity with High Performance
- On chip MOS capacitors
- 12 volt, 1 GHz technology

ARRAY SUMMARY

NPN Transistors	272
PNP Transistors	132
Total Diffused Resistance	425 K
Total Implant Resistance	2048 K
Total MOS Capacitance	60 pF
Total Components	1360
Bond Pads	27
Die Size (mils)	115 × 122

MINI TILE SUMMARY

T1 General	64
T2 Specialized	12
T3 Power	4
T4 Low Noise	2
T5 Precision	4
T14 High Power NPN	4
T15 High Power PNP	4



FB3623 — Medium High Power Tile Array

FB3631 Large Mixed Analog Digital Tile Array

GENERAL DESCRIPTION

The FB3631 tile array was developed for mixed analog digital applications. This FB3600 family tile array utilizes our proprietary mini tile architecture. The mini tile approach combined with our 12 volt, 1 GHz technology allows high complexity, high speed mixed signal circuits to be easily integrated.

This array is optimized for high complexity analog with a moderate amount of digital circuitry. For example, twelve 324 style operational amplifiers, sixty-six gates of logic, eight digital output buffers, and 3 other circuit blocks can all be implemented on a FB3631 tile array.

Our 1 GHz bipolar process allows us to achieve high performance circuits on our FB3600 tile arrays. Amplifiers with a bandwidth of 100 MHz and digital ECL gates with delays of 2 ns can be implemented on Micro Linear tile arrays.

FEATURES

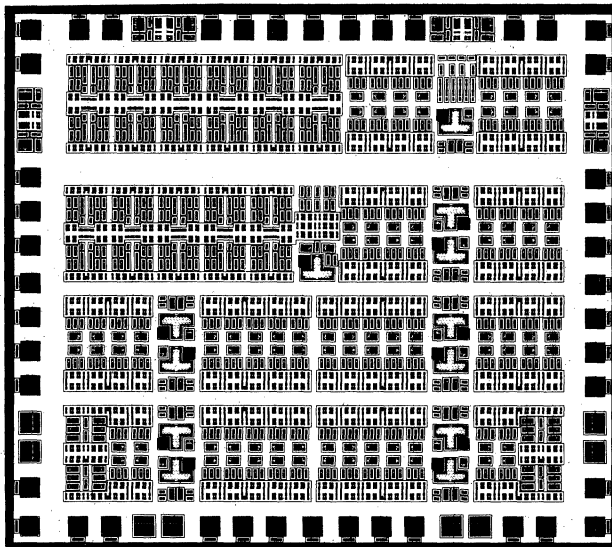
- Mixed Analog and Digital Circuitry
- 15 Analog circuit blocks with 66 ECL Gates
- 2800 Components, 44 Bond Pads
- On chip MOS Capacitors
- 12 volt, 1 GHz technology

ARRAY SUMMARY

NPN Transistors	690
PNP Transistors	154
Schottky Transistors	16
Total Diffused Resistance	850 K
Total Implant Resistance	3928 K
Total MOS Capacitance	60 pF
Total Components	2806
Bond Pads	44
Die Size (mils)	142 × 156

MINI TILE SUMMARY

T1 General	92
T2 Specialized	12
T3 Power	4
T7 High Frequency	4
T9 ECL	22
T10 ECL Bias	1
T11 TTL Output	8
T17 General	1



FB3631 — Large Mixed Analog Digital Tile Array

FB3635 Analog and Digital Tile Array

GENERAL DESCRIPTION

The FB3635 offers both analog and digital circuit design on a single tile array. The top half of the FB3635 contains components for analog circuit design. The bottom half of the array contains npn components for digital design.

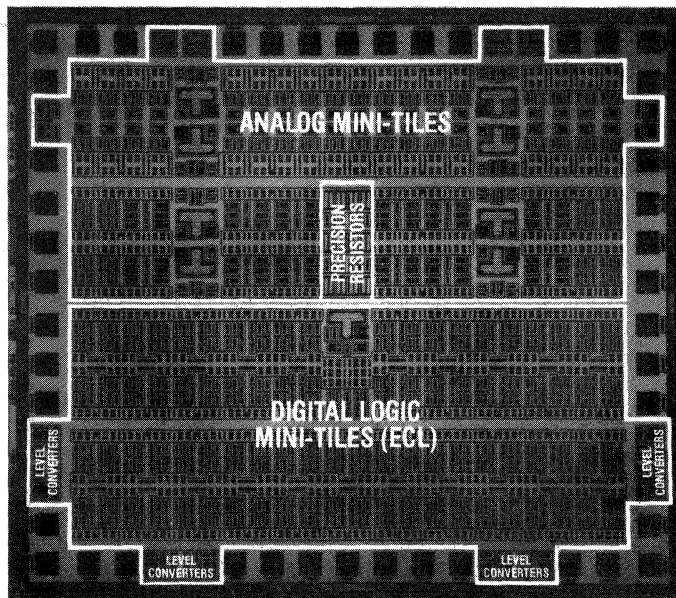
The analog section can implement eight LM324 type op amps or twelve LM339 type comparators. Many comparator applications can use a comparator with npn transistors in the input stage. In this case, the analog section of the array can be filled with eighteen comparators. A two quadrant analog multiplier, AGC circuit, analog multiplexer (switch), video amplifier or a MC1496 type modulator/demodulator can also be implemented. Each of these functions requires roughly the same number of components as one op amp. In addition, the array can also implement one 8-bit DAC, four 100MHz cascode amplifiers and a voltage reference.

The digital section of the FB3635 contains forty two digital logic cells. Each digital logic cell can implement a one bit latch (with set and reset), or three NAND gates. Two digital logic cells can implement an edge triggered D type flip-flop

FEATURES

- Mixed analog and digital tile array
- Analog section
 - nnp f_t 720MHz
 - pnp f_t 25MHz
 - Eight 5 pF MOS capacitors
 - Operates up to 12V \pm 10%
- Digital section
 - 132 NAND gates or 42 latches
 - 2 ns gate propagation delay
 - ECL logic using a single +5 volt supply
 - TTL, ECL, and CMOS compatibility

with set and reset. All logic functions are implemented using ECL logic. This provides for 2 nanosecond gate propagation delays and flip-flop toggle rates of 80MHz. The logic area can be powered off of a single 5V supply. On-chip logic level converters can convert the arrays ECL logic levels to standard TTL, CMOS or ECL logic levels.



FB3635 Tile Array

FB3400 Tile Array Selection Guide

FB3400 Family ($\pm 18V$ or up to $36V$ supply operation)

Array		FB3410	FB3420	FB3430
Description		Small General Purpose	Medium General Purpose	Large General Purpose
Mini Tile Summary				
T1	General Purpose	16	32	44
T2	Special Devices	4	12	16
T3	Power Devices	4	4	4
T4	Low Noise Devices	0	4	0
T5	Precision Resistors	0	2	3
Array Summary				
Complexity*		4	12	16
NPN Transistors		132	296	394
PNP Transistors		52	124	168
Total Diffused Resistance		176K	384K	538K
Total Implant Resistance		1600K	3200K	4400K
Total MOS Capacitance		40pF	120pF	160pF
Total Components		524	1132	1500
Bond Pads		32	46	66

* Analog complexity is in one 741 op amp or two 339 comparator equivalents. Digital complexity is in two input NAND gate equivalents.

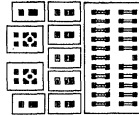
Component Performance (under typical operating conditions)

COMPONENT	FB3400 FAMILY
NPN Transistor	$h_{FE} = 120$ $f_T = 300\text{MHz}$ $BV_{ce0} = 40V$
NPN Large Transistor (FB3623 Only)	
PNP Substrate Transistor	$h_{FE} = 60$ $f_T = 20\text{MHz}$ $BV_{ce0} = 45V$
PNP Lateral Transistor	$h_{FE} = 30$ $f_T = 3\text{MHz}$ $BV_{ce0} = 45V$
Diffused Resistor	2% matching with $\pm 20\%$ absolute value
Precision Resistors	0.5% matching with $\pm 20\%$ absolute value
Implant Resistors	4% matching with $\pm 25\%$ absolute value
MOS Capacitor	$\pm 20\%$ absolute value

FB3400 Mini Tile Description

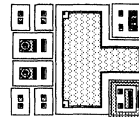
T1 Mini Tile

This is a general-purpose analog tile. It can implement many common building block functions, such as current mirrors, differential gain stages and level shifters. A T1 tile contains seven small geometry NPNs (ft 300MHz), two quad collector PNP (ft 4MHz), eleven 1k Ω resistors, and ten 10k Ω resistors.



T2 Mini Tile

The T2 tile contains a collection of specialized components. Many analog building blocks, such as op-amps, use a T1 tile and a T2 tile. The T2 tiles contain two medium size low noise NPN transistors, four small substrate PNP (ft 8MHz), one large substrate PNP (ft 8MHz), one triple emitter NPN, and a MOS capacitor whose value can be programmed up to 10pF.



T3 Mini Tile

The T3 tile contains two NPN power transistors for output stages driving up to 100mA each. These tiles are situated in peripheral positions around the chips, in close proximity to the bonding pads.



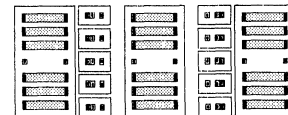
T4 Mini Tile

The T4 tile contains two low-noise NPN transistors which are used in circuits requiring low noise performance.



T5 Mini Tile

This mini tile contains ten minimum geometry NPN transistors and eighteen precision resistor links. These special resistors have a nominal ohmic value of 900 Ω and are matched to within an accuracy of 0.5%. It is possible to construct R-2R ladders used in the core of an 8-bit DAC by using 24 resistors.



FB3410, FB3420, FB3430 General Purpose Tile Arrays

GENERAL DESCRIPTION

The FB3400 family has been designed to utilize the traditional analog $\pm 15V$ signal swings and power supply rails. A single power supply of up to 36 V or split power supplies of up to $\pm 18V$ can be utilized. Significant board space and cost savings are possible with the FB3400 family of analog tile arrays.

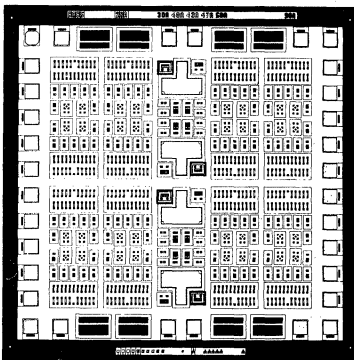
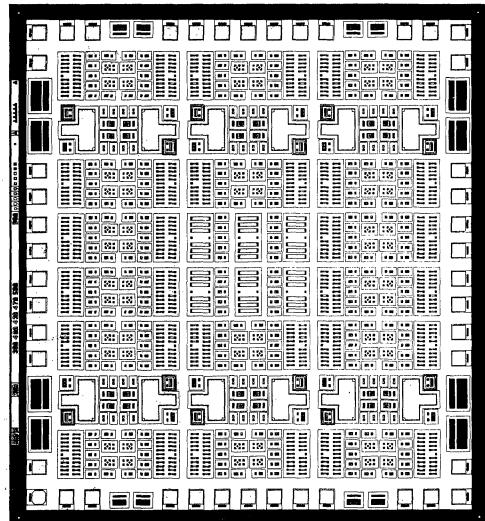
A single FB3400 Analog ASIC can typically replace ten to twenty standard analog building block components. In addition, many of the active and passive components surrounding the discrete building blocks can be incorporated on-chip.

The FB3400 family utilizes Micro Linear's new mini-tile architecture concept. The FB3400 family uses five different mini-tiles. One or more mini-tiles can be combined to implement functional blocks such as op-amps, comparators, voltage references, video amplifiers, transconductance amplifiers, modulators, demodulators, RS-232, RS-432, RS-422, V.35 drivers & receivers, D/A and pulse width modulation circuits.

FEATURES

- Optimized for up to 36V operation
- High Component Density, dual layer metal process
- 300MHz array technology
- Three high performance family members
- Design complexities of up to 16 op-amps
- On-chip precision resistors and compensation capacitors

GENERAL PURPOSE ANALOG ARRAYS


FB3410

FB3420

FC3560 Read Channel Tile Array

GENERAL DESCRIPTION

The FC3560 is an application focused tile array targeted specifically for mass storage applications. This Tile Array can implement all of the functions required in a high performance read channel.

The array consists of optimized circuit blocks for the following functions: a pulse detector, servo demodulator, data separator, frequency synthesizer, write pre-compensation, two crystal oscillators, bandgap reference, as well as 800 gates of uncommitted logic. The logic can be used to implement an encoder/decoder function, address mark generation/detection, or M & N dividers.

Typical performance of circuit blocks integrated on the FC3560 Tile Array is shown in the Tile Array summary below.

Although very specialized, the FC3560 maintains a high degree of flexibility. All of the components are uncommitted until connected with two layers of metal at the final step of the production process. The performance or function of the individual circuit functions can be modified or the architecture chosen to achieve an optimum read channel system.

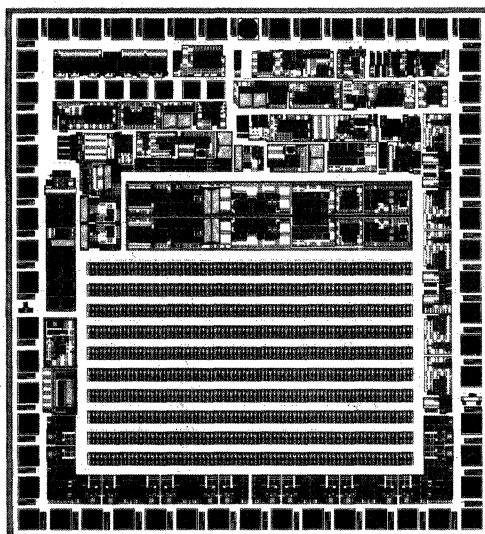
The FC3560 is fabricated using our advanced BiCMOS technology. This 4GHz, 1.5 μ process combines the advantages of high speed bipolar with dense CMOS.

FEATURES

- Application Focused Tile Array
- Optimized for Mass Storage Read Channel Functions
- Flexibility at the Functional Block and Component Level
- Advanced 4 GHz, 1.5 μ BiCMOS process

ARRAY SUMMARY

Pulse Detector	± 0.5 ns Pulse Pairing
Servo Demodulator	4 Channels
Data Separator	36 Mbits/s
Frequency Synthesizer	VCO and Charge Pump
Write Pre-Compensation	2 to 20ns
Crystal Oscillators (2)	36 MHz
Bandgap Reference	For Servo Reference
Uncommitted Logic Gates	800
Bond Pads	58
Die Size (mils)	134 x 142



FC3560 Read Channel Tile Array

FB3651 LAN Transceiver Array

GENERAL DESCRIPTION

The FB3651 is an application focused tile array intended for local area network transceiver applications. This array was developed using Micro Linear's proprietary mini tile architecture. This mini tile approach combined with our 12 volt, 1 GHz technology allows high complexity and high speed, cost effective LAN Transceiver circuits to be easily integrated.

The array consists of two distinct sections each with different component groupings, the first section, situated mostly in the middle of the array, contains custom cells to implement timer functions. These custom cells can implement up to nine timers with outputs that span from milliseconds up to 1/2 second with no external components. The long times are accomplished by generating very small yet stable currents that charge on chip capacitors. The remainder of the die area has general purpose mini-tiles for the other analog and digital circuit functions common to LAN transceiver applications.

The design of the FB3651 LAN Transceiver array is optimized for the circuit building blocks required to implement the function of a local area network transceiver. Examples of the types of circuits possible are: high speed transmitters and receivers, transmit and receive squelch, oscillators, diagnostic and fault protection circuits, LED drivers, and other similar type circuits.

This very high complexity array can realize up to 12 analog functional blocks, 6 high frequency/digital blocks, and 150 digital gates, in addition to the nine timer functions.

FEATURES

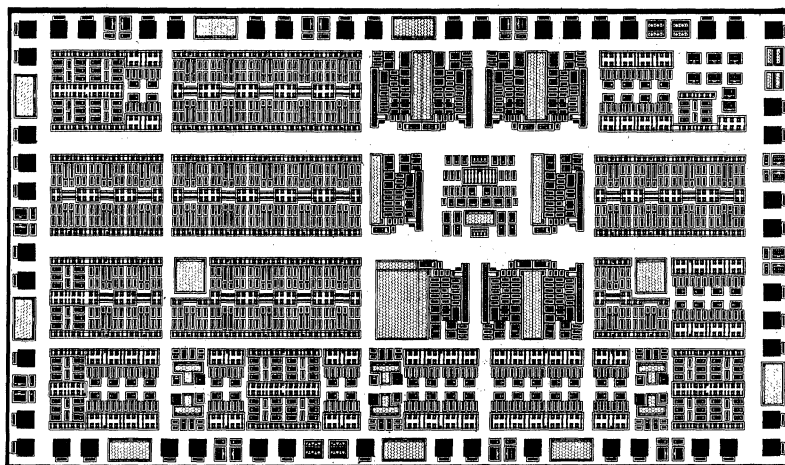
- Array optimized for local area network transceivers
- Nine independent timer functions possible for deadman, squelch, and diagnostic functions (1-500ms)
- Can implement a highly symmetrical current driven transmitter for low RFI noise and low jitter
- 18 analog circuit blocks with 150 ECL gates
- 12 volt, 1 GHz technology

ARRAY SUMMARY

NPN Transistors	1424
PNP Transistors	152
Schottky Transistors	20
Total Diffused Resistance	1020K
Total Implant Resistance	5432K
Total MOS Capacitance	310pF
Total Components	4242
Bond Pads	50

MINI TILE SUMMARY

T1 General	58
T2 Specialized	6
T7 High Frequency	17
T9 ECL Logic	50
T12 Schottky Peripheral	10
T13 High Frequency	3
Timer Cells	9
Timer Bias	1



FB3651 — LAN Transceiver Array

FC3580 Micro Power Controller Tile Array

GENERAL DESCRIPTION

The FC3580 is an application focused tile array optimized for battery powered controller applications. This Tile Array can implement all of the functions required for a DC to DC convertor for 1 to 3 cell battery powered systems. Examples of other configurations are, various boost buck regulators, MOSFET drivers, loud speaker drivers, or power amplifier applications, etc.

A synchronous rectifying boost regulator capable of running with input voltages as low as 1V and efficiencies greater than 90% can be implemented on the FC3580.

The FC3580 is fabricated using our advanced BiCMOS technology. This 4GHz, 1.5 μ process combines the advantages of high speed bipolar with dense CMOS. The bipolar devices can be used for high bandwidth, low offset and low noise amplifiers while the CMOS devices can be used for dense digital low power logic as well as for analog switches, the front ends of FET amplifiers, or power FETs. CMOS devices allow the design of power output stages that can swing rail to rail.

Some of the specialized components on the this array include; very low $R_{DS\ ON}$ (0.2 N-Channel, 0.5 P-Channel) CMOS output transistors, high quality poly resistors, stable poly-oxide capacitors, and a large amount of total resistance (>16M ohms) for low power designs.

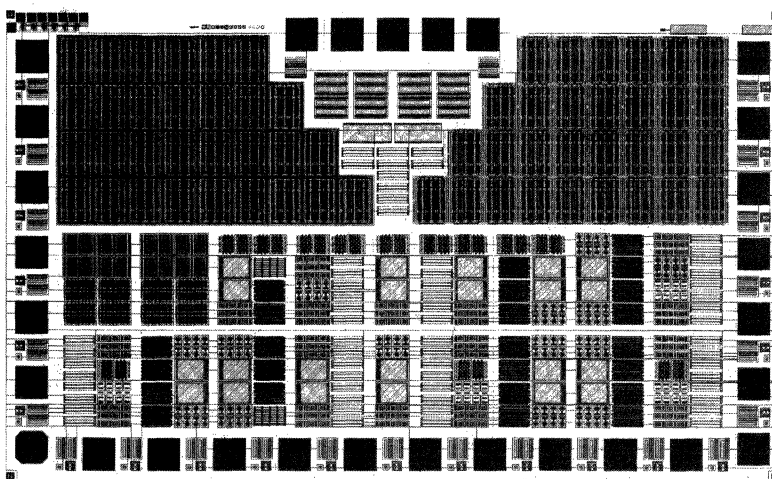
The small die size enables the FC3580 to fit in narrow (0.15") body SOIC packages.

FEATURES

- Application Focused Tile Array
- Ideal for Battery Powered Applications
- High Efficiency Down to 1V Inputs
- Rail to Rail Output Swings Possible
- Advanced 4 GHz, 1.5 μ BiCMOS Process

ARRAY SUMMARY

NPN Transistors	152
PNP Transistors	152
NMOS Transistors	42
PMOS Transistors	42
Total Poly Resistance	3300K
Total Other Resistance	13.4M
Total MOS Capacitance	77pF
Total Components	836
Bond Pads	29
Die Size (mils)	70 x 110



FC3580 Micro Power Controller Tile Array

FB3680 Electronic Ballast and Power Factor Tile Array

GENERAL DESCRIPTION

The FB3680 is an application focused tile array optimized for electronic ballast controller applications. This Tile Array can implement all of the functions required for a high efficiency low distortion, power factor corrected dimmable ballast controller.

A ballast controller with a 0.99 power factor and greater than 95% efficiency can be implemented on the FB3680.

Certain areas on the the FB3680 are customized for the specific circuit blocks required in power factor corrected ballast applications. These circuit blocks include: a 1% voltage reference, three fast, high current (50ns, 200mA) totem pole outputs, oscillator, under voltage lockout, thermal shutdown, voltage to current convertor, timer, multiplier, shunt regulator, fault protection, three op-amps (3MHz GBW), plus eight comparators and uncommitted gates of RTL logic. In total the FB3680 can implement 14 analog circuit blocks plus logic.

Although optimized for these circuit blocks, the FB3680 still maintains complete flexibility. Each component remains uncommitted until connected with the two layers of metal interconnect at the final step of the wafer manufacturing process. This allows the FB3680 to be used for a wide variety of circuit functions beyond the ones identified above.

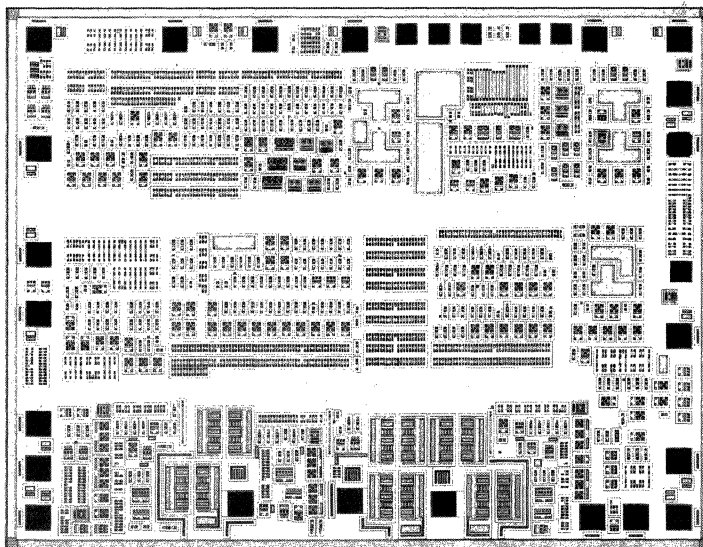
The FB3680 is fabricated using our 12 volt, 1GHz bipolar technology.

FEATURES

- Application Focused Tile Array
- Optimized for Dimming Electronic Ballast with Power Factor Correction Applications
- Can Implement a 0.99 Power Factor with 95% Efficiency
- All Components Uncommitted - Connected with Two Metal Interconnect Layers
- 12 Volt, 1 GHz, Bipolar Technology

ARRAY SUMMARY

NPN Transistors	368
PNP Transistors	207
Power NPN Transistors	12
Total Base Resistance	470K
Total Implant Resistance	2.3M
Total MOS Capacitance	50pF
Total Components	1740
Bond Pads	22
Die Size (mils)	.109 x 140



FB3680 Ballast and Power Factor Tile Array

FB3480 Power Supply Controller Array

GENERAL DESCRIPTION

The FB3480 Power Supply Controller Array is a new concept in Switch Mode Power Supply Controller (SMPS) technology. This Array is the first configurable bipolar array specifically designed for SMPS applications. The FB3480 is optimized for high performance and low design cost and time, since most of the commonly used SMPS functions have been pre-designed and characterized.

With the FB3480 a power supply designer can select his own unique controller topology and features without the need for external components. This flexibility allows compact PC board layout, minimizing interference from induced RFI/EMI, and enhancing high frequency performance.

The FB3480 contains all of the elements found in most SMPS controllers in addition to areas of uncommitted circuitry on the Array which can be configured for unique applications. The core of the array is similar in performance to the UC1825 controller and consists of an oscillator, precision voltage reference, error amplifier and two totem pole high current output stages, which are specifically optimized for high performance at high frequency. These cells can be connected with other available circuitry on the chip to form a complete SMPS controller. The "uncommitted"

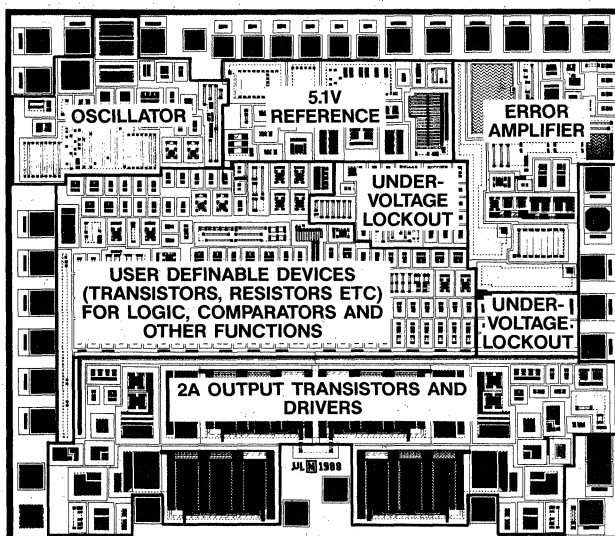
sections of the array can be configured into comparators, logic, and other functions, to implement a complete control. In addition, closely matched resistors are available to precisely control thresholds and gain settings on the chip.

The FB3480 array is used to make several state of the art standard products, including the ML4825 and ML4809. Cells from these and future Micro Linear FB3480 based standard products can be made available for customer designs.

FEATURES

- Practical operation to Switching Speeds above 1 MHz
- Precision Bandgap Reference $\pm 1\%$
- 2 A peak push-pull output stages for high speed drive of power MOSFETs
- Fast comparator to output response time ... $< 50\text{ns}$
- Available in DIP or PLCC
- Mil-Temp available
- Additional user-definable logic, comparators and other circuitry available on chip

BLOCK DIAGRAM



CIRCUIT CAPACITY

The FB3480 consists of pre-configured optimized functional cells which are commonly found in SMPS controllers. In addition the FB3480 has additional transistors and resistors available for user configuration. The Array topology is optimized to implement a full featured high performance SMPS controller utilizing both the "Dedicated Resources" (oscillator, error amp, reference and power output) and additional fully configurable resistors and transistors.

The FB3480 can be configured to fully implement the ML4825 improved pin compatible replacement for the popular UC1825 controller and still have resources on chip for additional functions.

The table below lists uncommitted resources which are still available after implementing the following controller designs completely.

The FB3480's unused elements can be configured into:

Component	ML4825		ML4823		FB3480
	Used	Unused	Used	Unused	Total
NPN Current Sink	20	9	15	14	29
NPN Transistors	27	29	21	35	56
Dual Emitter NPN	6	8	3	11	14
NPN Follower (1 Emitter)	0	8	0	8	8
NPN Follower (2 Emitter)	10	7	7	10	17
Low Voltage 4 Collector PNP	3	8	3	8	11
High Voltage 4 Collector PNP	2	2	2	2	4
PNP Current Source	3	3	2	4	6
Zeners (NPN Diodes)	2	4	0	6	6
2 A NPN Transistors	4	0	2	2	4
Implanted Resistors	37	18	29	26	55
Base Resistor	3	5	3	5	8
Bondable Pads	16	12	16	12	28

DEDICATED RESOURCES

The table below lists the dedicated functions which are available on the FB3480. Each of these blocks is described in more detail beginning on page 4.

REF #	Description	Count
OSC1	1.5 MHz. F_{MAX} R-C Oscillator	1
REF1	5.1 V Precision Reference	1
PWR1	2 A Peak Push-Pull Totem Pole Output Buffer	2
EA1	5.5 MHz Bandwidth, 12V/ μ S Slew Rate Error Amp	1
UV1	Under-Voltage Lockout Circuit	1

UNCOMMITTED RESOURCES

The FB3480 includes a large number of transistors and resistors which are used to make up the circuit functions or "cells" which are available for design and

listed beginning on page 8. The table below lists how many of these components are available and their typical performance characteristics.

TYPICAL PERFORMANCE CHARACTERISTICS OF UNCOMMITTED RESOURCES

Transistors

Ref #	Description	β @ 100 μ A	β @ 1 mA	$I_{(\beta/2)}$ (Note 1)	V_{CE0}	V_{CBO}	F_T (MHz)	Count
TN1	NPN Low Voltage		100	20	12	25	450	55
TN3	NPN Power (100 mA)		100	60	45	55	450	1
TN4	Dual Emitter NPN		100	20	45	55	450	10
TN5	NPN Emitter Follower		100	20	12	25	450	8
TN6	NPN Dual Emitter Follower		100	20	12	25	450	17
TN7	NPN Current Sinks		100	20	12	25	450	29
TP1	Lateral Low V PNP	30		1	15	25	4	8
TP2	Lateral High V PNP	30		1	45	45	4	3
TP3	Substrate PNP	60		1	45	45	20	1

Diodes

Ref #	Description	$V_{REVERSE}$	$I_{FORWARD}$ (mA)	Count
DN1	NPN Diode	6.8	2	6

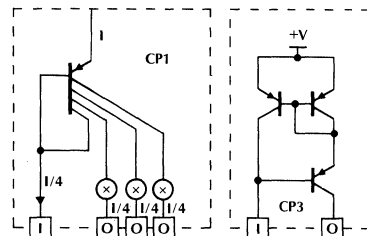
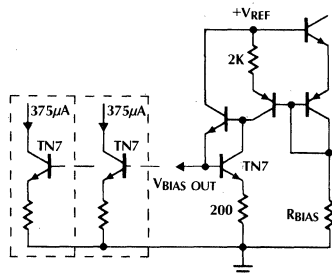
Resistors

Ref #	Type	Value	Tolerance	Ratio Matching	Count
RB1	Base	2K Ω	20%	0.5%	6
RI1	Implant	2K Ω	20%	1.0%	29
RI2	Implant	8K Ω	20%	1.0%	8
R13	Implant	30K Ω	20%	3.0%	5

Note 1: This column indicates the useful current handling capability of the transistor and is defined as the current at which the β is down to 1/2 of its nominal (100 μ A for PNP and 1mA for NPN) value.

Transistors TP1, TP2, and TP3, are constructed with four separate collectors and can be used as current sources as shown below. TN7 is a special NPN transistor which includes a 200 Ω resistor cell and is intended to be used as a current sink in conjunction with the on chip bias reference generator.

Current Sources CP1 and CP3 are examples of current sources which can be constructed with the PNP transistors. CP1 is a basic biasing current source, where the 3 output currents are equal to the input current. CP3 has an additional circuit to cancel the base current error and is more accurate and linear.



ABSOLUTE MAXIMUM RATINGS

Supply Voltage	36V
Output Current, Source or Sink (Power Outputs)	
DC	0.5A
Pulse (0.5 μ S)	2.0A
Input Voltage (logic, comparators etc.)	-0.3V to 6V
Error Amplifier Output Current	5mA
Oscillator Charging Current	-5mA
Junction Temperature	150°C
Storage Temperature Range	-65°C to +150°C
Lead Temperature (Soldering 10 sec.)	+260°C

OPERATING TEMPERATURE RANGE

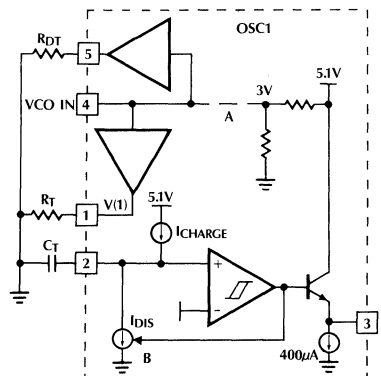
Plastic Packages (PCC or DIP)	-40°C to +85°C
Ceramic Packages	-55°C to +125°C

Absolute maximum ratings are those values beyond which the device could be permanently damaged. Absolute maximum ratings are stress ratings only and functional device operation is not implied.

FB3480 CORE CELL DESCRIPTIONS

Oscillator

The FB3480 oscillator charges the external capacitor (C_T) with a current (I_{SET}) equal to $V(1)/R_T$. When the capacitor voltage reaches the upper threshold, the comparator changes state and the capacitor discharges



to the lower threshold. While the capacitor is discharging, a high pulse is provided on (3).

Several configurations are available for the FB3480 oscillator:

1. Basic ML4825 oscillator: A fixed discharge time is provided by saturating a transistor for I_{DIS} . This discharge current is approximately 10 mA. Charge time is fixed since pin 4 is set to 3V by the internal resistor divider at node A.
2. Controlled Discharge: I_{DIS} is equal to:

$$\frac{16 \times V(\text{pin } 5)}{R_{DT}}$$
3. Voltage Controlled Oscillator. The connection at node A is open and pin 4 is brought out. The control range is from 1V to 5.5V. The voltage at pin 4 sets the charge and discharge currents (if option 2 above is implemented) thereby controlling the frequency of the oscillator.

ELECTRICAL CHARACTERISTICS (unless otherwise noted, these specifications apply for $R_T = 3.65K\Omega$, $C_T = 1000pF$, $-55^\circ C < T_J < 150^\circ C$, $V_{CC} = 15V$) fixed deadtime, A connected.

Parameter	Conditions	DESIGN LIMITS			Units
		Min	Typ	Max	
OSC1 and OSC2 Oscillator					
Initial Accuracy	$T_J = 25^\circ C$	360	400	440	KHz
Voltage Stability	$10V < V_{CC} < 30V$		0.2	2	%
Temperature Stability	$-55^\circ C < T_J < 150^\circ C$			5	%
Total Variation	line, temp.	340		460	KHz
Clock out High		3.9	4.5		V
Clock out Low			2.3	2.9	V
Ramp Peak		2.6	2.8	3.0	V
Ramp Valley		0.7	1.0	1.25	V
Ramp Valley to Peak		1.6	1.8	2.0	V
Capacitor Discharge Current		10			mA
Current Consumption			3.2		mA
Typical VCO Control Range	$1.5 < OSC2-4 < 5V$	75%		175%	f_{NOM}

FB3480

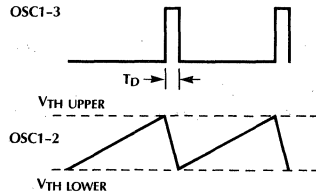
Oscillator (Continued)

The Oscillator period can be described by the following relationship:

$$T_{OSC} = T_{ramp} + T_{deadtime}$$

where: $T_{ramp} = C (V_{upper} - V_{lower}) / I_{charge}$

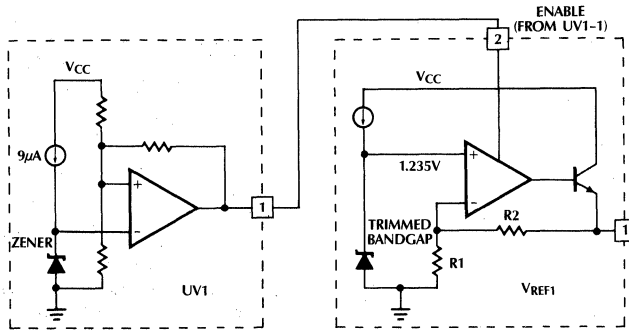
and: $T_{deadtime} = C (V_{upper} - V_{lower}) / I_{dis}$



Voltage Reference and Undervoltage Lockout

The FB3480 voltage reference is a buffered trimmed bandgap design with excellent thermal characteristics. The undervoltage lockout circuit (UV1) monitors V_{CC} and compares it to a zener voltage with hysteresis in the comparator. When the supply is sufficiently high to

allow operation of the controller, pin 1 of UV1 goes true, enabling V_{REF1} , which runs the bias circuitry for all the logic. In this way, when the V_{CC} is under voltage, the array goes into a low current consumption mode. The thresholds for UV1 can be selected.

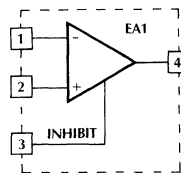


ELECTRICAL CHARACTERISTICS (unless otherwise noted, $V_{CC} = 15V$, $-55^{\circ}C < T_j < 150^{\circ}C$)

Parameter	Conditions	DESIGN LIMITS			Units
		Min	Typ	Max	
VR1					
Output Voltage	$T_j = 25^{\circ}C$, $I_O = 1mA$	5.05	5.10	5.15	V
Line Regulation	$10V < V_{CC} < 30V$		2	20	mV
Load Regulation	$1mA < I_O < 10mA$		5	20	mV
Temperature Stability	$-55^{\circ}C < T_j < 150^{\circ}C$.2	.4	mV/ $^{\circ}C$
Total Output Variation	line, temp.	5.0		5.20	V
Output Noise Voltage	10Hz to 10KHz		50		μV
Long Term Stability	$T_j = 125^{\circ}C$, 1000 Hrs		5	25	mV
Short Circuit Current	$V_{REF} = 0V$	-15	-50	-100	mA
Current Consumption			.7		mA
UV1					
Start Threshold		8.8	9.2	9.6	V
UVOL Hysteresis		.4	.8	1.2	V

Error Amplifier

The FB3480 error amplifier is a 5.5 MHz bandwidth, 12V/ μ S slew rate op-amp with provision for limiting the positive output voltage swing (Output Inhibit line) for ease in implementing the soft start function.

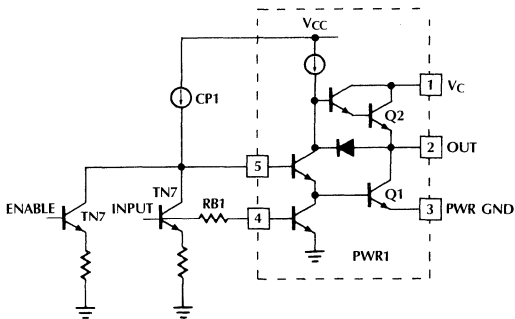


ELECTRICAL CHARACTERISTICS (unless otherwise noted, $V_{CC} = 15V$, $-55^{\circ}C < T_J < 150^{\circ}C$)

Parameter	Conditions	DESIGN LIMITS			Units
		Min	Typ	Max	
EA1					
Input Offset Voltage				10	mV
Input Bias Current			.6	3	μ A
Input Offset Current			.1	1	μ A
Open Loop Gain	$1 < V_O < 4V$	60	96		dB
CMRR	$1.5 < V_{CM} < 5.5V$	75	95		dB
PSRR	$10 < V_{CC} < 30V$	85	110		dB
Output Sink Current	$V_{EA1-1} = 1V$	1	2.5		mA
Output Source Current	$V_{EA1-1} = 4V$	-.5	-1.3		mA
Output High Voltage	$I_{EA1-1} = -0.5mA$	4.0	4.7	5.0	V
Output Low Voltage	$I_{EA1-1} = 1mA$	0	0.5	1.0	V
Unity Gain Bandwidth		3	5.5		MHz
Slew Rate		6	12		V/ μ S
Current Consumption	$I_{EA1-1} = 0$		1.5		mA

Output Driver Stage

The FB3480 Output Driver is a 2A peak output high speed totem pole circuit designed to quickly switch the gates of capacitive loads, such as power MOSFET transistors. The external translator to the left of the PWR1 cell is shown as an example of how to interface from logic signals to the output stage and is made up of uncommitted resources available on the IC.



ELECTRICAL CHARACTERISTICS (unless otherwise noted, $V_{CC} = 15V$, $-55^{\circ}C < T_J < 150^{\circ}C$)

Parameter	Conditions	DESIGN LIMITS			Units
		Min	Typ	Max	
PWR1					
Output Low Level (V_{OL})	$I_{OUT} = 20mA$		0.25	0.40	V
	$I_{OUT} = 200mA$		1.2	2.2	V
Output High Level (V_{OH})	$I_{OUT} = -20mA$	13.0	13.5		V
	$I_{OUT} = -200mA$	12.0	13.0		V
Collector Leakage	$V_C = 30V$		100	500	μ A
Rise/Fall Time	$C_L = 1000pF$		30	60	nS
Current Consumption	$I_{OUT} = 0$		7.3		mA

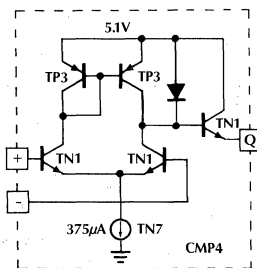
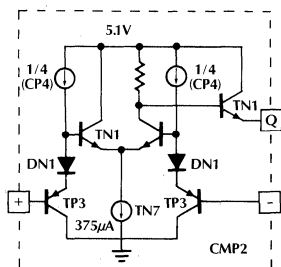
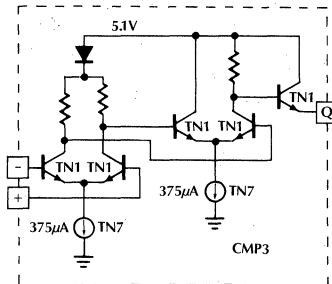
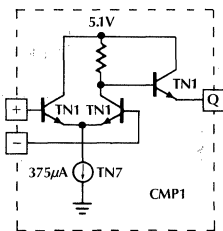
UNCOMMITTED RESOURCE CELLS

The functions listed below are pre-simulated "cells" which are available for use. These "cells" are made up from the uncommitted resources described on page 3.

Comparators

ELECTRICAL CHARACTERISTICS (unless otherwise noted, $V_{CC} = 15V$, $-55^{\circ}C < T_J < 150^{\circ}C$)

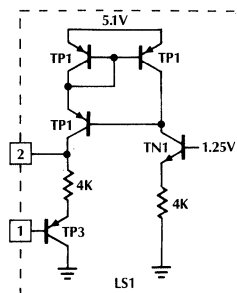
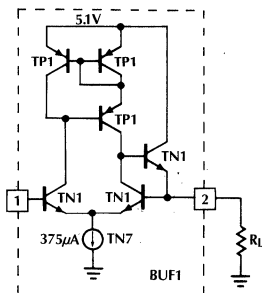
Parameter	Conditions	DESIGN LIMITS			Units
		Min	Typ	Max	
CMP1 — Simple NPN Comparator					
Output Low Level	$I_{OUT} = 1mA$		3.6	4.1	V
Output High Level	$I_{OUT} = -1mA$		4.35	4.7	V
Input Offset Voltage			10		mV
Input Common Mode Range		1		4.1	V
T_{PHL}			20		nS
T_{PLH}			20		nS
Voltage Gain			28		V/V
Quiescent Current Consumption			1		mA
CMP2 — Simple Ground Sensing PNP Input Comparator					
Output Low Level	$I_{OUT} = 1mA$		3.6	4.1	V
Output High Level	$I_{OUT} = -1mA$		4.35	4.7	V
Input Offset Voltage			20		mV
Input Common Mode Range		GND		3.1	V
T_{PHL}			25		nS
T_{PLH}			25		nS
Voltage Gain			28		V/V
Quiescent Current Consumption			1		mA



Comparators (Continued)

ELECTRICAL CHARACTERISTICS (unless otherwise noted, $V_{CC} = 15V$, $-55^{\circ}C < T_j < 150^{\circ}C$)

Parameter	Conditions	DESIGN LIMITS			Units
		Min	Typ	Max	
CMP3 — High Gain NPN Comparator					
Output Low Level	$I_{OUT} = 1mA$		3.6	4.1	V
Output High Level	$I_{OUT} = -1mA$		4.35	4.7	V
Input Offset Voltage			20		mV
Input Common Mode Range		1		4.1	V
T_{PHL}			35		nS
T_{PLH}			35		nS
Voltage Gain			700		V/V
Quiescent Current Consumption			1		mA
CMP4 — Very High Gain NPN Comparator					
Output Low Level	$I_{OUT} = 1mA$		3.6	4.1	V
Output High Level	$I_{OUT} = -1mA$		4.35	4.7	V
Input Offset Voltage			10		mV
Input Common Mode Range		1		4.1	V
T_{PHL}			500		nS
T_{PLH}			500		nS
Voltage Gain			15,000		V/V
Quiescent Current Consumption			1		mA
BUF1 — Voltage Follower (Buffer)					
Input Bias Current				4	μA
Output Voltage Range	$I_{OUT} < 1mA$	0		4.1	V
Offset Voltage				10	mV
Input Common Mode Range		1		4.1	V
Open Loop Voltage Gain			1000		V/V
Slew Rate	$C_L < 1pF$	2			$V/\mu S$
Quiescent Current Consumption			.375		mA
LS1 — 1.25V Level Shift					
Input Bias Current				12	μA
$V_{OUT} - V_{IN}$		1.1		1.4	V

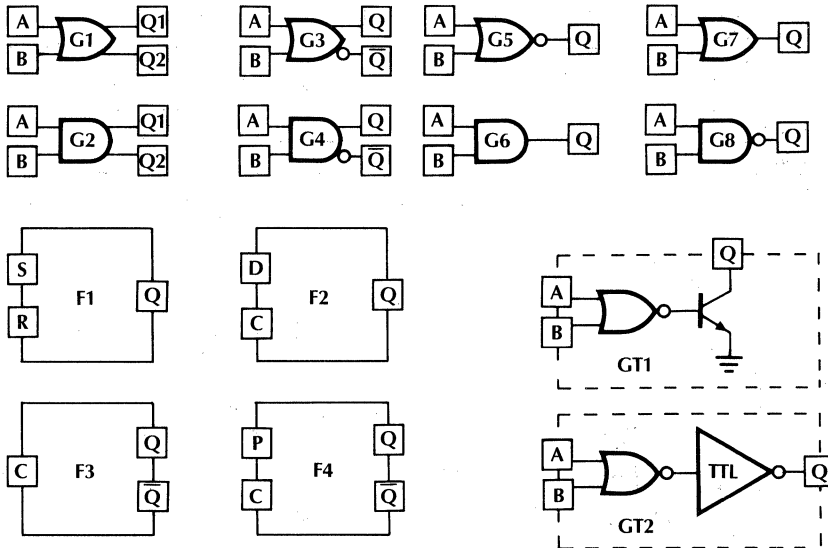


Internal Logic Macro Functions

The FB3480 logic section is a pre-characterized library made up of high speed, high noise immunity Emitter Function Logic (EFL) and Emitter Coupled Logic (ECL) functions. This logic family uses uncommitted low voltage transistors and resistors which are available to be metallized (described on page 3) to make up the logic functions below.

This family features the ability to "wire.or" the outputs as well as having a very flexible structure and fast propagation delay times. For more information on designing with EFL logic, please refer to the Application Hints on page 12.

Ref #	Description	Component Utilization							Design Limits	
		TN1	TN4	TN5	TN6	TN7	DN1	RB1	t _{PD} (nS)	I _{CC} (mA)
G1	2-Input 2-Output OR	3			1	1		1	7	.375
G2	2-Input 2-Output AND		1		1	2	1	1	5	.7
G3	2-Input Complementary Output OR/NOR	2		2		1		2	7	.375
G4	2-Input Complementary Output AND/NAND	4		2		2	2	2	7	.7
G5	2-Input NOR	3		1		1		1	7	.375
G6	2-Input AND		1	1		2	1	1	5	.7
G7	2-Input OR	3		1		1		1	7	.375
G8	2-Input NAND	4		1		2	1	1	7	.7
GT1	EFL to TTL 2-Input OR Open Collector	4				1	2	2	10	.7
GT2	EFL to TTL 2-Input OR Totem Pole	5				2	3	5	13	1.125
F1	R-S Flip Flop	2	1		1	1		1	8	.375
F2	Positive Edge Triggered D Flip Flop	4	2		2	3		1	10	1.125
F3	Positive Edge Triggered T Flip Flop	6	5			5		3	12	1.875
F4	T Flip Flop with Preset	8	1			6		4	12	2.26



APPLICATIONS INFORMATION

DESIGNING AND SPECIFYING A CONTROLLER

The FB3480's core cell architecture is designed to simplify the task of designing a PWM controller for unique needs or specific tasks. Micro Linear will design the IC for a nominal initial lot charge providing that the design uses the cells described in this datasheet. A specification for a customized FB3480 controller requires the following elements:

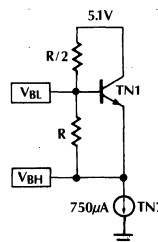
1. A block diagram describing the interconnection of the cells. This could also take the form of a modified block diagram from any of Micro Linear's standard products (such as the ML4809 or ML4825). The block diagram should be drawn in terms of the cells described in this datasheet.
2. Packaging requirements. The FB3480 has 28 bondable pads. Industrial temperature range units can be packaged in 28 pin Plastic Leaded Chip Carrier (PLCC) Plastic DIP packages from 14 to 28 pins. Military temperature range units can be packaged in Ceramic DIP packages from 14 to 28 pins.
3. Test specifications.
4. Operating Temperature Range requirements.

An accurate indication of circuit performance can be obtained by prototyping with one of Micro Linear's standard products built from the FB3480 (such as the ML4825 or ML4809) and using "off the shell" comparators and logic of similar performance to that specified in the FB3480 datasheet.

DESIGNING WITH EFL LOGIC

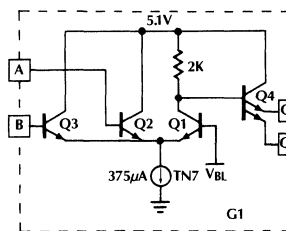
The FB3480's logic section is a collection of high frequency NPN transistors, current sink transistors, resistors and diodes which can be configured into a variety of high speed logic functions. The logic family used in the FB3480 is Emitter Function Logic (or EFL) which features speed, flexibility and simplicity. Since most of the logic is interconnected "on chip", buffering to drive PC board layout capacitances is not needed, further minimizing the number of transistors which are used to accomplish the necessary logic. In addition, the output structure lends itself to accomplishing "wired-or" functions.

The family's output voltage swings are between $(V_{REF} - V_{BE})$, logic 1, and $(V_{REF} - 2V_{BE})$, logic 0, where V_{REF} is set to 5.1V internally. Input thresholds are called V_{BH} and V_{BL} and are set by the threshold generator shown below to $(V_{REF} - V_{BE}/2)$ for a logic 1 and $(V_{REF} - 3 * V_{BE}/2)$ for a logic 0.

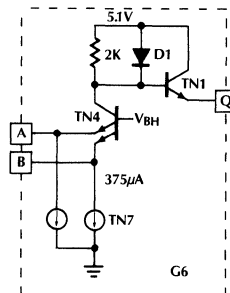


LOGIC BIAS THRESHOLD GENERATOR

In the example below, G1 is a full OR gate. When either input A or input B exceeds V_{BL} ($1.5 * V_{BE}$ down from the supply), Q1 is cut off, putting the base of Q4 at V_{REF} , which puts $V_{REF} - V_{BE}$ on the output at the emitters of Q4. When both A and B are below V_{BL} , Q1 conducts forcing the voltage on its collector to drop to $(V_{REF} - V_{BE})$. This occurs since TN7 is set so that its current ($375 \mu A$) will cause a V_{BE} drop to occur in a $2K\Omega$ resistor. The output emitters of Q4 will be at $V_{BASE} - V_{BE}$, or $(V_{REF} - 2V_{BE})$.



G6 is an AND gate using a dual emitter input. When either emitter is allowed to go below V_{BL} ($V_{BH} - V_{BE}$), TN1 conducts, which causes the output to go low. If both inputs are above V_{BL} , TN4's base sits at V_{REF} , which makes TN4's output $(V_{REF} - V_{BE})$.



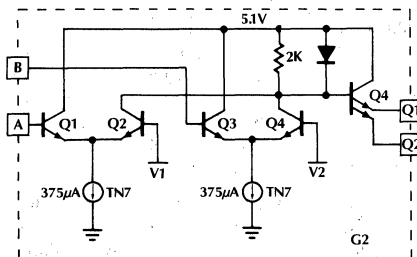
APPLICATIONS INFORMATION (Continued)

When multiple outputs are tied together, they function like an OR gate. Only one of the output emitters need to go "high" for the line to become true. This assumes that the node is loaded with one current sink.

Comparator functions with combinational logic can also be easily constructed using this family. Note that G1 has a full differential input stage. By applying a voltage (V1) on the base of Q1, the output will be true when A or B exceed V1. A similar example using an AND function is shown below.

With the FB3480, a power supply designer can select his own unique controller topology and features without the need for external components. This flexibility allows compact PC board layout, minimizing interference from induced RFI/EMI, and enhancing high frequency performance.

Inverting or complementary outputs can be obtained easily from most simple logic functions simply by moving the load resistor and output transistor base connection to the opposite collector.



FB3490 General Purpose PWM Controller Array

GENERAL DESCRIPTION

The FB3490 Power Supply Controller Array is optimized for use in Switch Mode Power Supply designs at frequencies up to 750 KHz. These "core cells" (Oscillator, Reference, Output Drivers) are optimized for high performance while retaining maximum flexibility. In addition, this array contains cells (or tiles) which can be used for logic, amplifiers, comparators and other special functions.

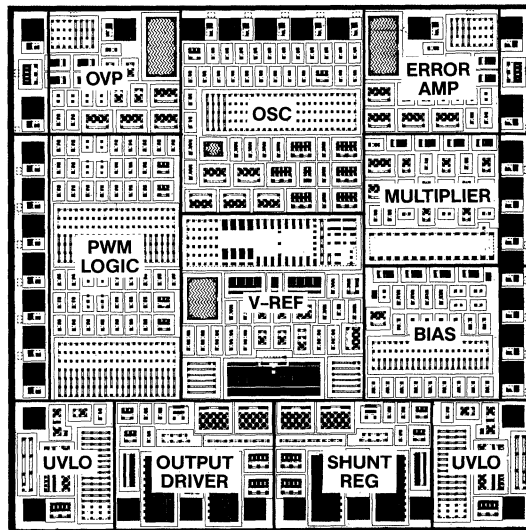
With the FB3490, a power supply designer can select a unique controller topology and feature set without the need for external components. Design and layout of a unique controller is simplified through the use of many pre-defined and pre-simulated "soft macros" which can be made available for customer designs.

This array is similar in performance to the UC1846 and was used to implement the ML4812 Power Factor Controller. Cells from these and future Micro Linear FB3490 based standard products can be made available for customer designs.

FEATURES

- Practical Operation at Switching Frequencies to 750 KHz
- Dual High Current (1A peak) Totem Pole Outputs
- $\pm 0.5\%$ Trimmed Bandgap Reference
- Multiple Error Amp systems possible
- Extensive library of "soft macro" building block functions available for user design
- 40V bipolar dual layer metal process

BLOCK DIAGRAM



FB3491 Resonant Mode Controller Array

GENERAL DESCRIPTION

The FB3491 is an application focused tile array intended for resonant mode power supply controller applications. This array, built on our 40 volt technology, consists of customized and general purpose groupings of components to implement the various power supply control circuit blocks. Certain areas, such as the power output section and the oscillator, are customized to obtain a higher level of performance for these critical circuit functions.

The array has four high current (2A peak) output transistors to implement two high current, low cross conduction, totem pole type output drivers. The high current capability allows the quick charge and discharge of the gate capacitance of external power MOSFET devices.

High speed emitter function logic circuits can be implemented to achieve fast current sense circuits. A propagation delay of less than 50ns from current sense to output shutdown can be achieved. An oscillator that is capable of operating up to 3MHz and a precision reference with an accuracy of $\pm 1\%$ and a temperature stability of 50 ppm/ $^{\circ}\text{C}$ are examples of the level of performance that can be achieved.

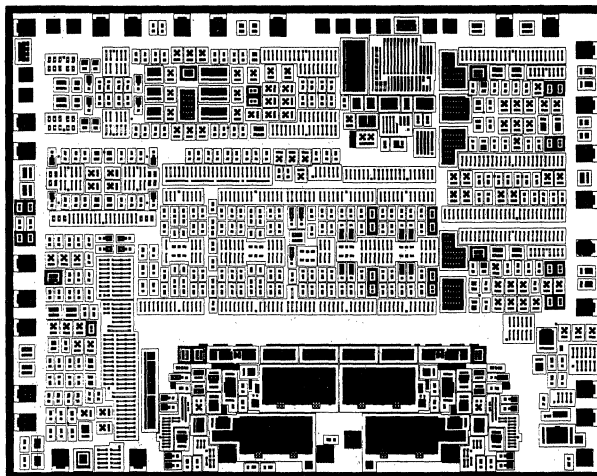
Standard products built on this tile array can be easily modified to create a semi-standard version optimized for a specific customers application.

FEATURES

- Array Optimized for Resonant Mode Power Supply Control Circuits
- High Current (2A) Output Transistors for Fast Output Drivers
- Can Implement all the Circuit Blocks for a High Performance Resonant Mode Controller
- 6 Analog Circuit Blocks and 40 Gate Complexity
- 40 Volt, 400 MHz Technology

ARRAY SUMMARY

NPN Transistors	315
PNP Transistors	126
Power NPN Transistors	4
Schottky Transistors	33
Total Diffused Resistance	570K
Total Implant Resistance	2600K
Total MOS Capacitance	34pF
Total Components	1248
Bond Pads	30
Die Size (mils)	140 x 181



FB3491 — Resonant Mode Controller Array

FB3492 Phase Modulation Controller Array

GENERAL DESCRIPTION

The FB3492 is an application focused tile array intended for phase modulated power supply controller applications. This array, built on our 40 volt technology, consists of customized and general purpose groupings of components to implement the various power supply control circuit blocks. Certain areas, such as the power output section, the oscillator, and the voltage reference are customized to obtain a higher level of performance for these critical circuit functions.

The array has eight high current (2A peak) output transistors to implement four high current, low cross conduction, totem pole type output drivers. The high current capability allows the quick charge and discharge of the gate capacitance of external power MOSFET devices.

High speed emitter function logic circuits can be implemented to achieve fast current sense circuits. A propagation delay of less than 50ns from current sense to output shutdown can be achieved. An oscillator that is capable of operating up to 3MHz and a precision reference with an accuracy of $\pm 1\%$ and a temperature stability of 50 ppm/ $^{\circ}\text{C}$ are examples of the level of performance that can be achieved.

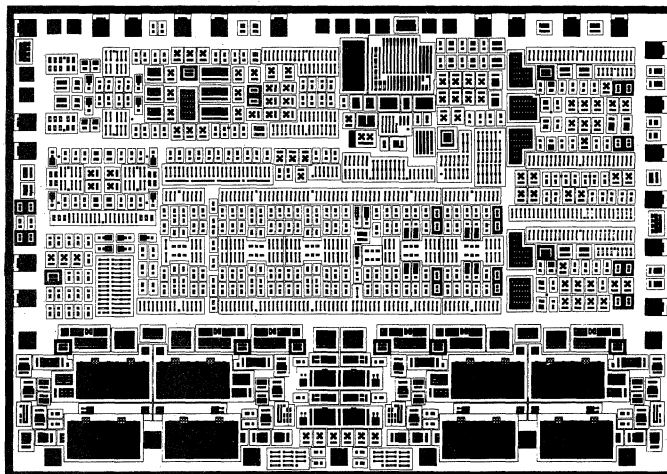
Standard products built on this tile array can be easily modified to create a semi-standard version optimized for a specific customers application.

FEATURES

- Array Optimized for Phase Modulated Power Supply Control Circuits
- High Current (2A) Output Transistors for Fast Output Drivers
- Can Implement all the Circuit Blocks for a High Performance Phase Modulation Controller
- 9 Analog Circuit Blocks and 60 Gate Complexity
- 40 Volt, 400 MHz Technology

ARRAY SUMMARY

NPN Transistors	323
PNP Transistors	131
Power NPN Transistors	8
Schottky Transistors	39
Total Diffused Resistance	580K
Total Implant Resistance	2500K
Total MOS Capacitance	34pF
Total Components	1290
Bond Pads	29
Die Size (mils)	140 × 201



FB3492 — Phase Modulation Controller Array

Package Selection Guide

Package	FC3510	FC3580	FC3560	FB3605	FB3610	FB3620	FB3621	FB3622	FB3623	FB3630	FB3631	FB3635	FB3680	FB3651	FB3410	FB3420	FB3430	FB3480	FB3490	FB3491	FB3492	
CDIP-8	X	X		X	X																	
CDIP-14	X	X		X																		
CDIP-16	X	X		X	X	X	X	X	X				X		X				X			
CDIP-18	X	X		X	X	X	X	X	X				X		X				X			
CDIP-20	X	X		X	X	X	X						X									
CDIP-24	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X
CDIP-28	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X
CDIP-40	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X
LCC-16	X	X		X	X	X	X	X	X				X		X				X	X		
LCC-20	X	X	X	X	X	X	X	X	X				X		X				X	X		
LCC-28	X	X	X	X	X	X	X	X	X	X	X	X	X		X	X			X	X	X	
LCC-32	X	X	X	X	X	X	X	X	X	X	X	X	X		X	X	X	X	X	X	X	X
LCC-44	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X
PDIP-8	X	X		X	X	X	X	X	X				X		X				X	X		
PDIP-14	X	X	X	X	X	X	X	X	X	X		X	X		X				X	X	X	X
PDIP-16	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X
PDIP-18	X	X	X	X	X	X	X	X	X	X		X	X		X				X	X	X	X
PDIP-20	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X
PDIP-22	X	X		X	X	X	X	X	X				X		X				X			
PDIP-24	X	X	X	X	X	X	X	X	X	X	X	X	X		X	X			X	X	X	
PDIP-28	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X
PDIP-40	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X
PDIP-48	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X
PLCC-20	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X
PLCC-28	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X
PLCC-32	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X
PLCC-44	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X

Package	FC3510	FC3580	FC3560	FB3605	FB3610	FB3620	FB3621	FB3622	FB3623	FB3630	FB3631	FB3635	FB3680	FB3651	FB3410	FB3420	FB3430	FB3480	FB3490	FB3491	FB3492	
PLCC-68	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X
PQFP-44	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X
PQFP-52	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X
SDBR-14	X	X		X	X	X	X	X	X				X		X			X	X			
SDBR-16	X	X		X	X	X	X	X	X				X		X			X				
SDBR-18	X	X		X	X	X	X	X	X				X		X			X	X			
SDBR-20	X	X		X	X	X	X	X	X				X		X			X	X			
SDBR-24	X	X	X	X	X	X	X	X	X	X	X	X	X		X	X		X	X	X		
SDBR-28	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X
SDBR-40	X	X	X	X	X	X	X	X	X	X	X	X	X		X	X		X	X	X		
SOIC-8	X	X		X																		
SOIC-14	X	X		X																		
SOIC-16	X	X	X	X	X	X	X	X	X	X		X	X		X			X	X			
SOIC-18	X	X	X	X	X	X	X	X	X	X	X	X	X		X	X		X	X	X	X	
SOIC-20	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X
SOIC-28	X	X	X	X	X	X	X	X	X	X	X	X	X		X	X	X	X	X	X	X	X
SOIC-32	X	X	X	X	X	X	X	X	X	X		X	X		X			X	X	X		
SSOP-20	X	X	X	X	X	X	X	X	X	X		X	X		X			X	X			
SSOP-24	X	X	X	X	X	X	X	X	X	X		X	X		X			X	X			
SSOP-28	X	X		X	X	X	X	X	X				X		X			X	X			
TQFP-32	X	X	X	X	X	X	X	X	X	X	X	X	X		X			X	X			
TQFP-44	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X
TQFP-52	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X
TQFP-64	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X	X

X – Indicates the array is available in this package

Package Descriptions

- SDBRZ = Sidebrazed DIP
- LCC = Leadless Chip Carrier
- CDIP = Ceramic DIP
- PDIP = Plastic Dip
- PCC = Plastic Leaded Chip Carrier
- SOIC = Small Outline
- SSOP = Shrink Small Outline Package
- TQFP = Thin Quad Flat Pack

Quality and Reliability

Section 8

Quality and Reliability	8-1
Appendix A — Failure Rate Calculation	8-11
Appendix B — Radiation Hardness of 12V Bipolar Process	8-13



Quality and Reliability

Micro Linear is dedicated to excellence in its people and products. By adopting a policy of continuous improvement, we pledge to provide defect free products and services which meet or exceed our customers' expectation.

Total Quality Management

At Micro Linear we are committed to total quality management by building quality into every step of the manufacturing process from design to product qualification; from receiving to shipping. The Total Quality Management program at Micro Linear Corporation is a detailed program involving engineering and manufacturing and is designed to produce the highest quality linear integrated circuits available.

Wafer Inspection

Emphasis is placed on statistical analysis, electrical measurements on specially designed Process Control Monitors in accordance with MIL-STD-414 (sampling by variables), visual inspection, and film measurements. Potential reliability hazards are investigated and detected early by utilizing diagnostic and device structures on each wafer and periodic SEM analysis.

Assembly Inspection

Comprehensive receiving inspection for all materials and piece-parts is performed in accordance with the strictest quality assurance procedures. To assure conformance and control to specifications, documented quality control checks and monitors are performed on-line.

Testing

Micro Linear has invested in the latest "state-of-the-art" analog testers to achieve the most complete and thorough parametric testing of integrated circuits in the industry. Data sheets provide the customer a precise listing of parameters which are 100% tested. The calibration system is in compliance with MIL-STD-45662.

Traceability

All units are marked with unique lot numbers. These lot numbers provide complete traceability all the way to wafer fab as well as assembly and test.

Micro Linear considers traceability to be essential for good engineering control and additional insurance for its customers.

ESD (Electro Static Discharge)

Products are fully characterized to MIL-STD-883C, Method 3015 and strict controls on handling and packaging are observed. A full ESD program, from design through manufacturing, incorporates training of all employees who handle Micro Linear products.

Major Change Control

Major change controls are in place to notify customers in accordance with MIL-M-38510. Micro Linear reviews all process, product, and package changes. All changes with possible impact are submitted for a re-qualification which may include electrical, mechanical, and/or thermal characterization. If applicable, reliability requalification is performed.

Process Control

Process monitors and gate inspections insure that all devices are properly tested and that the required sample tests are performed prior to shipment. Inspection records and reports concerning monitors and inspection data are used to status the quality level of products through the final test operations. Statistical sampling plans insure the quality of the product.

Micro Linear welcomes OEM quality system surveys. Micro Linear is qualified by a number of customers to MIL-Q-9858A and MIL-I-45208 for military programs.

Failure Analysis and Reporting of Customer Returns

A formal program exists to record, analyze, and take appropriate corrective action on all returns. A Corrective Action Committee reviews all discrepancies and assigns responsibility to implement solutions or improvements on a weekly basis. A report is generated and sent to the customer stating our findings and corrective action.

Quality and Reliability

Document Control

All records providing product traceability are maintained in accordance with MIL-M-38510. All company documents for procedures, specifications, drawings, travelers, flow charts, schematics, etc. that define customer requirements, raw material requirements, design, manufacture, and testing of products are controlled by a Document Control organization within the Micro Linear's Quality Assurance group.

Audits

Critical manufacturing areas are audited by a quality inspector at specified intervals. The audit verifies adequacy of operator training, correct revisions, the procedures, proper data entry, and record maintenance. In addition, weekly audits include an ESD program, particle count, calibration, and document control programs.

Supplier Control

Control of the quality of the incoming material is critical to the success of Micro Linear. Under the TQM philosophy, Micro Linear has an audit program of its suppliers of Class 1 material defined by those directly associated with the final products. Such audit is carried out on a defined frequency and performed by both manufacturing and quality personnel. Information gathered from the audit is reviewed with the supplier to incorporate programs to improve the quality of the material provided to Micro Linear.

Quality Indices Report

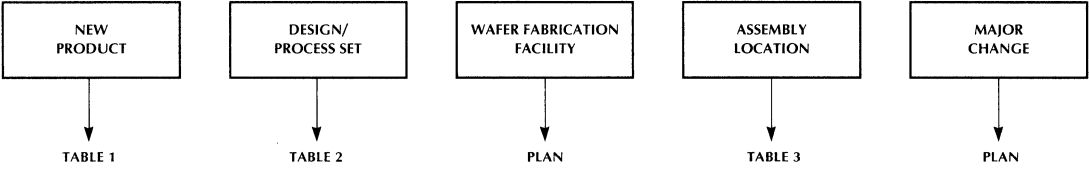
Various quality data are collected each month and summarized in a report presented to management for a review. This Quality Indices Report contains such information as process capability indices from wafer fab processing and assembly, on-going reliability data by process types and ppm data from in-line electrical testing as well as the AOQ data.

Reliability Program

Micro Linear's Reliability Program consistent with those of other semiconductor manufacturers utilizes various accelerated life tests as tools for establishing reliability status and progress. These tests are undertaken to identify infant mortality and wearout failure mechanisms for specific or generically similar device families.

Micro Linear's Reliability program has three components: Qualification, Quality Conformance and Reliability Audit. Each design/process technology set, each wafer fabrication facility, and each assembly location by package type is initially qualified. Periodic re-evaluation (Quality Conformance Testing) is performed, thereafter, to evaluate the on-going reliability of products and processes. In addition, Micro Linear has the third component, Reliability Audit named ACT (Advance Conformance Testing) to ensure the reliability of products shipped to customers. ACT defines auditing of samples from each process and wafer fab facility and subjecting them to an accelerated life testing. Plan for each program is illustrated in Figures 1 through 3 and detailed in Tables 1 through 4.

Micro Linear's product reliability is monitored closely and we have an extensive reliability data base for both hermetic and molded devices. This data is published on a quarterly basis.



NOTE: "Plan" are the appropriate stresses and tests determined by a qualification committee.

Figure 1. Qualification Testing

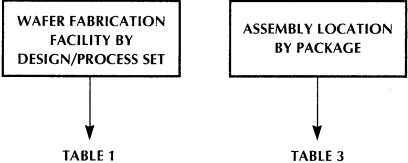


Figure 2. Quality Conformance Testing

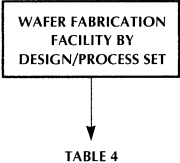


Figure 3. Reliability Audit (ACT)

Quality and Reliability

Stress/Test	Method	Condition	Quantity
Life Test	1005	1000 hrs @ 125°C	77
HAST		50 hrs @ 130/85% R.H.	45
ESD	3015		3

Table 1.

Stress/Test	Method	Condition	Quantity
Life Test	1005	1000 hrs @ 125°C	77
HAST		50 hrs @ 130/85% R.H.	45
Autoclave	QAP36004	168 hrs @ 121°C, 15 psi	45
Temperature Cycle	1010	1000 cyc. @ -55 to +125°C	45
Thermal Shock	1011	200 cyc. @ -55 to +125°C	45
High Temp Storage	QAP36005	1000 hrs. @ 150°C	45

Table 2.

Stress/Test	Method	Condition	Quantity
Life Test	1005	1000 hrs @ 125°C	77
HAST		50 hrs @ 130/85% R.H.	45
Autoclave*	QAP36004	168 hrs @ 121°C, 15 psi	45
Temperature Cycle	1010	1000 cyc. @ -55 to +125°C	45
Thermal Shock*	1011	200 cyc. @ -55 to +125°C	45
High Temp Storage	QAP36005	1000 hrs. @ 150°C	45
Physical Dimensions	2016		15
Solderability	2003	22 leads	3
Resistance to Solder Heat	QAP36002	260°C, 10 sec	32
Resistance to Solvents	2015		4
External Visual Inspection	QAP34001		15
Lead Integrity	2004		3

Note * Qualification testing only.

Table 3.

Stress/Test	Method	Condition	Quantity
Life Test	1005	48 hrs @ 125°C	125

Table 4.

Process Control/Quality Conformance

Reliability evaluations provide a snapshot of the product at a particular point in time. Process control is necessary to insure the picture obtained is accurate. Process control provides consistency and hence, predictability.

Defect-free material is a pre-requisite to shipping cost-effective products which conform to specified requirements. The system for doing this is shown in Figure 4. The focal point to this system is the Vendor Qualification Board comprised of representatives from Manufacturing Engineering, Quality, Reliability, and Purchasing. In addition, a Corrective Action Committee with representatives from the above disciplines meets weekly to evaluate all discrepant material reports. These reports are the result of any non-conformance both internal and external (vendors) to Micro Linear.

Because of the extreme sensitivity of wafer fabrication on product reliability, special care is taken to evaluate wafer process control. This is shown in Figure 5.

Wafer Fab Process Control

FRONT-END PROCESS

- Initial Oxidation Thickness
- Buried Layer Mask CD
- Buried Layer Resistivity
- Epi Resistivity and Thickness
- Epi Oxidation Thickness
- Iso Mask CD
- Iso Diffusion Resistivity
- Iso Oxide Thickness
- Base Mask CD
- Implant Oxide Thickness
- Base Resistivity
- Base Oxide Thickness
- Implant Mask CD
- Emitter Mask CD
- Emitter Resistivity
- Emitter Oxide Thickness
- Cap Mask CD
- Cap Oxide Thickness
- Emitter Oxide Thickness
- Field Oxide Thickness
- Contact Mast CD
- Metal Thickness

BACK-END PROCESS

- Dielectric Oxide Thickness
- Metal Thickness
- Passivation Thickness
- Defect Inspection
- SEM Inspection

ADDITIONAL CHECKS IN DIFFUSION

- CV Plots of Oxide and Drive Tubes
- Etch Rate Monitors
- Temperature Profiling
- Quartzware Cleaning
- Particle Counts in Tubes

MASKING

- Photoresist Thickness
- Incoming Mask CD Measurement
- Etch Rate Control
- Exposure Intensity
- Hard Bake Temperature
- Resist Pinhole Check
- Exposure Monitor
- CD Monitor

Assembly Process Control

- DI Water Resistivity
- Wafer Saw/Wash Monitor
- 2nd Optical Inspection
- Die Shear Monitor
- Bond Wire Pull Monitor
- Ball Bond Shear Monitor
- 3rd Optical Inspection
- Mold Temperature
- Deflash Inspection
- Solderability Test
- Plating Thickness/Composition
- Marking Permanency
- Coplanarity Inspection
- Final Visual Inspection
- Air Monitors (temperature, humidity, particle)

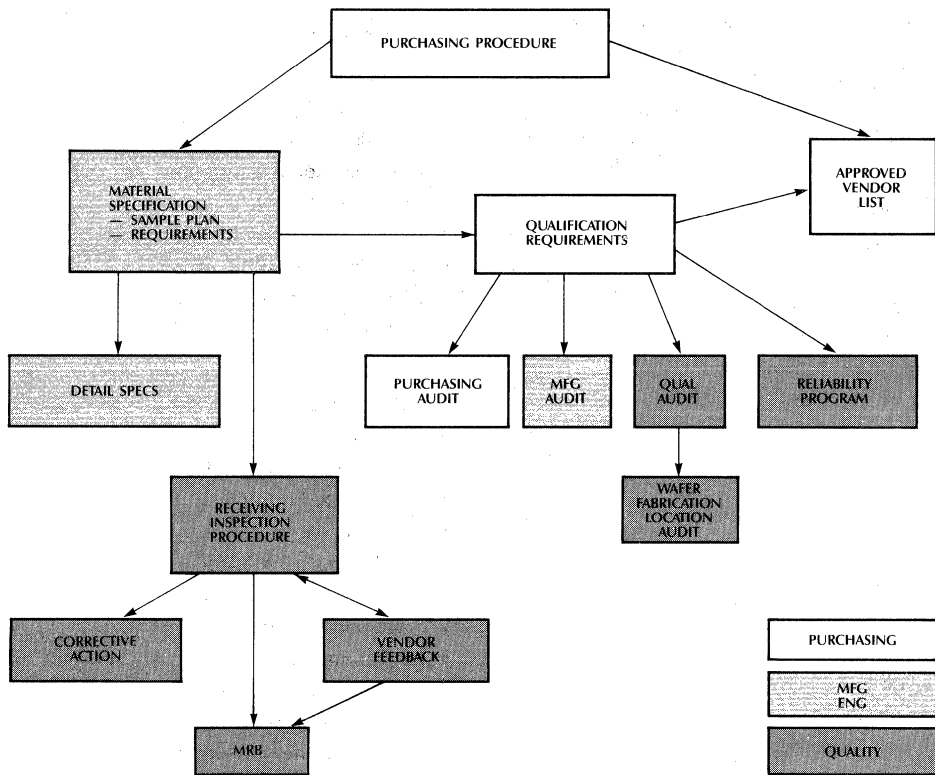


Figure 4. Vendor Control System

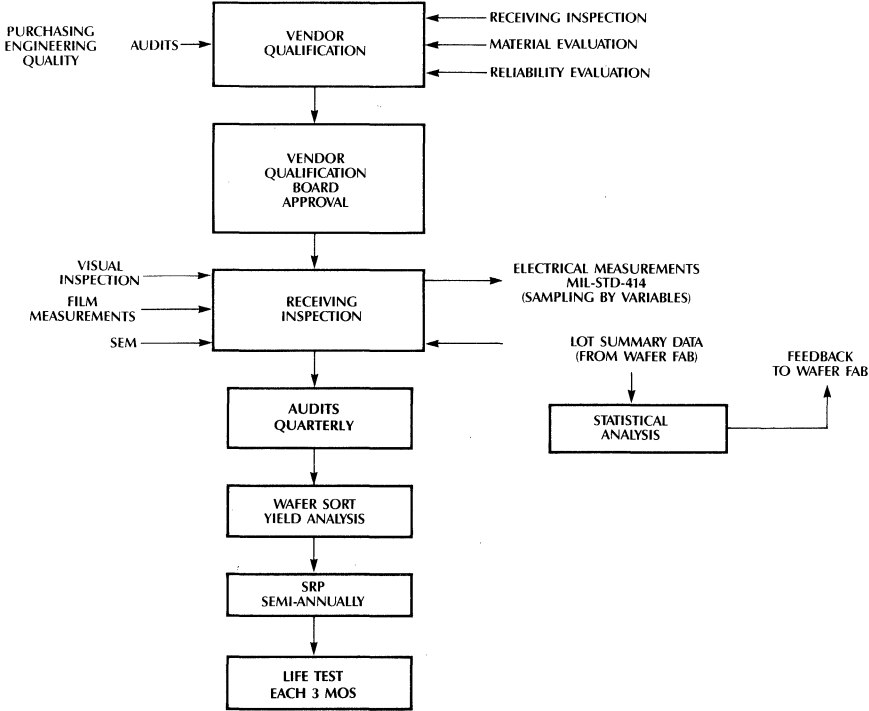


Figure 5. Process Control System

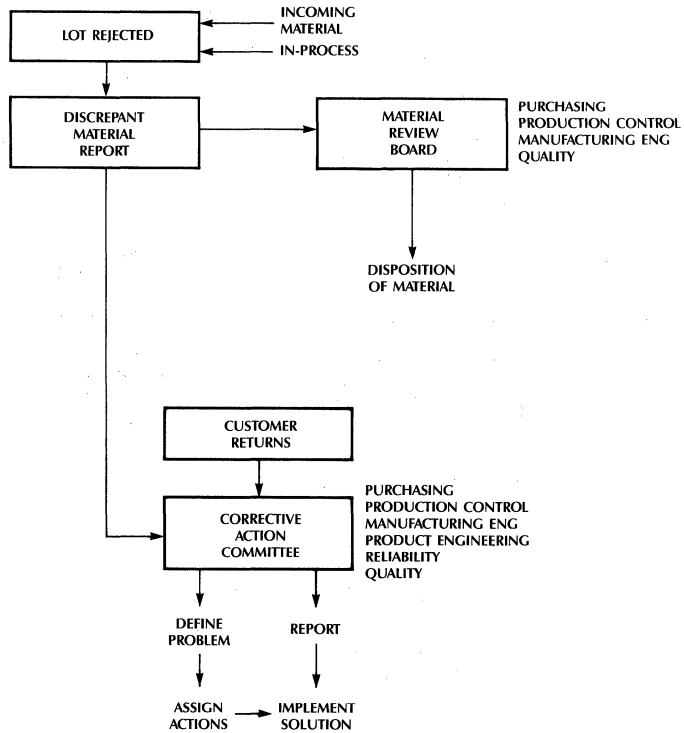
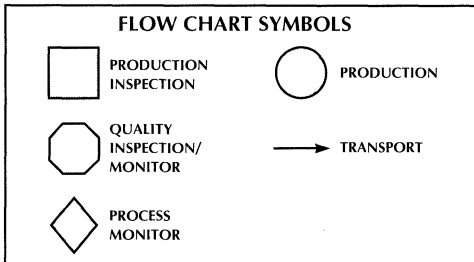
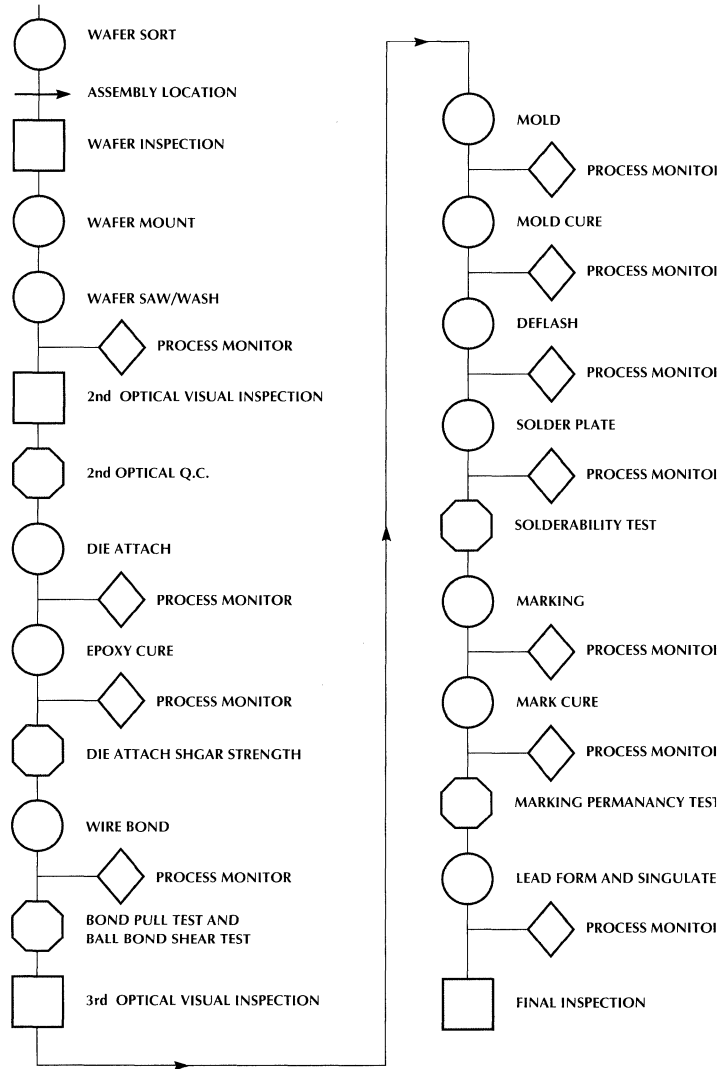


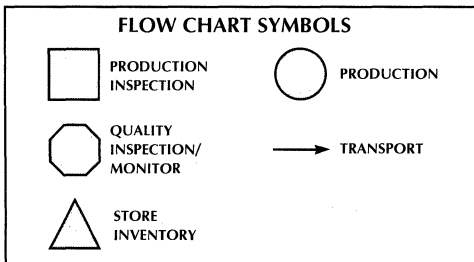
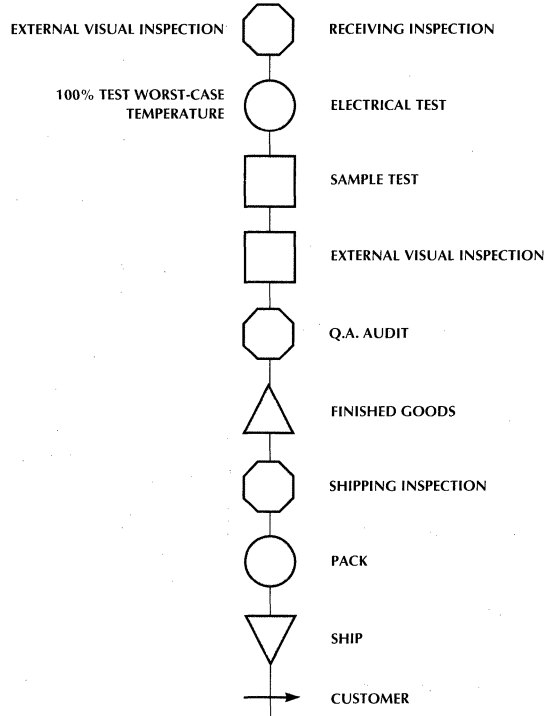
Figure 6. Corrective Action Program

Typical Molded Package Assembly Flow



Quality and Reliability

Typical Test Flow



APPENDIX A. Failure Rate Calculations

In order to predict the rate at which product will fail, it is necessary to accelerate the life of the product. This is most commonly done by a temperature and/or voltage stress, a process known as burn-in. The equation for both stresses is exponential, hence large acceleration factors can be achieved. In our studies, only temperature was used in the acceleration equation; the devices were biased at nominal voltages. The equation is shown below. It is known as the Arrhenius Reaction Rate Equation, named for the man who modeled the relationship between temperature and the chemical reaction property of materials.

Arrhenius Reaction Rate Equation

$$Af = \text{Exp} \left[\frac{Ea}{K} \left(\frac{1}{T1} - \frac{1}{T2} \right) \right]$$

Af: Acceleration Factor

Ea: Activation Energy (in electron volts)

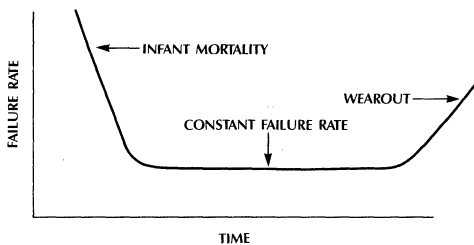
K: Boltzmanns Constant (8.62×10^{-5})

T1: Temperature of System Operation (°K)

T2: Temperature of Life Test (°K)

Burn-in when run for 1000 hours, is called "life test". Interim readouts normally occur at 168 and 500 hours. The hypothesis is that a "bathtub curve" will result. This curve, shown below, illustrates a device's failure rate versus time. Certain manufacturing defects have a tendency to cause failures early in the life of a device (infant mortality). The failure rate associated with these defects can be accelerated by applying stresses, such as temperature and voltage, which do not appreciably affect the normal failure rates or wear out mechanisms.

Bathtub Curve



Activation Energies

In order to calculate the acceleration factor, the activation energies for various failure modes encountered in the semiconductor industry are required. Initially, failure modes are assumed based on industry experience. As failures occur, they are rigorously analyzed and the failure modes then used to determine which activation energies are appropriate for determining failure rates. The following table describes the most common failure modes and their activation energies.

Table 1.

Failure Mechanism	Ea	Stress
Oxide Defects	0.3 eV	High Voltage Op Life
Contamination	1.0 eV	High Voltage Bias
Silicon Defects	0.5 eV	High Voltage
Metal Line Electromigration	0.5 eV	High Voltage Op Life
Contact Electromigration	0.9 eV	High Voltage Op Life
Masking Defects Assembly Defects	0.5 eV	Hi Temp. Storage Op Life
Microcracks	N/A	Temperature Cycling
Short Channel Charge Trapping	-.06 eV	Low Voltage Hi Vol Op Life

Acceleration Factors

Once the activating energy is determined for a given failure mechanism, the acceleration factor can be calculated using the Arrhenius equation. The following table lists some of the common activation energies and its associated acceleration factors between different ambient temperature.

Table 2.

Est. T ₁ Accelerated Temperature	Estimated T ₂ Typical Application Temperatures				Activation Energy (eV)
	25°C	40°C	55°C	70°C	
125°C	132	52	22	10	0.5
150°C	313	123	53	24	
125°C	1522	376	106	33	0.75
150°C	5530	1367	384	121	
125°C	6587	1231	268	67	0.9
150°C	30994	5793	1262	314	

Quality and Reliability

Failure Rates

At Micro Linear, failure rate are generally stated at 60% confidence level using Chi square statistic per the following formula.

$$\lambda_{\max} = \frac{\chi^2_{1-\alpha} [\text{with } df = 2(r + 1)]}{2t}$$

where:

- λ_{\max} = maximum failure rate
 - χ^2 = chi square distribution
 - r = number of failures
 - df = degree of freedom
 - t = total number of test hours
 - α = statistical error expected in estimate.
- For 60% confidence level, $\alpha = 0.4$ or $1-\alpha = 0.6$

Selected values of Chi Square distribution are listed in Table 3.

Table 3. Percentiles of the Chi Square Distribution.
(Values of χ^2 corresponding to certain selected probabilities)

		60% Confidence Level	90% Confidence Level
Probability in %		60.0	90.0
$1 - \alpha$		0.60	0.90
df	Total Failures		
1		0.708	2.71
2	0	1.830	4.61
3		2.950	6.25
4	1	4.040	7.78
5		5.130	9.24
6	2	6.210	10.60
7		7.280	12.00
8	3	8.350	13.40
9		9.410	14.70
10	4	10.500	16.00
11		11.500	17.30
12	5	12.600	18.50
13		13.600	19.80
14	6	14.700	21.10
15		15.700	22.30
16	7	16.800	23.50
17		17.800	24.80
18	8	18.900	26.00
19		19.900	27.20
20	9	21.000	28.40

Failure rate may be expressed a number of ways. Table 4 compares various ways of expressing failure rates.

Table 4. Failure Rates

NO. OF FAILURES PER DEVICE HOURS	FAILURE RATE	% PER 1000 HOURS	PPM (HOURS)	FITS	MTBF (HOURS)
$1/1 \times 10^9$	0.000000001	0.0001	0.001	1	1×10^9
$1/1 \times 10^8$	0.00000001	0.001	0.01	10	1×10^8
$1/1 \times 10^7$	0.0000001	0.01	0.1	100	1×10^7
$1/1 \times 10^6$	0.000001	0.1	1	1000	1×10^6
$1/1 \times 10^5$	0.00001	1	10	10000	1×10^5
$1/1 \times 10^4$	0.0001	10	100	100000	1×10^4
$1/1 \times 10^3$	0.001	100	1000	1000000	1×10^3

APPENDIX B. Radiation Hardness of 12V Bipolar Process

The Micro Linear 12V bipolar process has demonstrated selective hardness to radiation exposure. The components most commonly used in the 12V process which are described in table 1, were exposed up to

10^6 Rads total dose ionizing radiation. A second group of the same components were exposed to non-ionizing radiation of up to 10^{14} fluence neutrons/sq cm. Neither group was exposed to both types of radiation.

DESCRIPTION	BIAS DURING IRRADIATION	POST IRRADIATION MEASUREMENTS	FIGURES
Minimum Geometry NPN	$V_{CES} = +5V$	Δh_{fe}	1, 4
Lateral Quad Collector PNP	$V_{CES} = -5V$	Δh_{fe}	2, 5
Vertical PNP	$V_{CES} = -5V$	Δh_{fe}	3, 6
45Ω N+ Resistor	No bias	ΔR	3, 6
850Ω P+ Resistor	No bias	ΔR	3, 6
10KΩ Implanted P Resistor	No bias	ΔR	3, 6
10pF Capacitor	No bias	ΔI_L	3, 6

Table 1. Components

Figures 1 through 3 show the results of the ionizing radiation tests. Figures 4 through 6 show the results of the non-ionizing radiation tests.

- * The resistors and capacitors were not significantly altered by exposure to these radiation levels. They are not included in the figures.

Summary

The h_{fe} of the NPN transistors degrade by approximately 50% at 10^5 Rads and 80% at 10^6 Rads. The PNPs degrade more severely by approximately 80% at 10^5 Rads and reach unity at 10^6 Rads. Degradation vs. neutron fluence is similar but less severe.

Micro Linear circuits, exclusively using NPN devices and passive components, can be designed to perform in a high radiation environment.

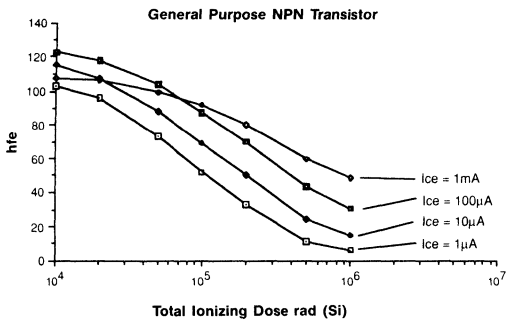


Figure 1.

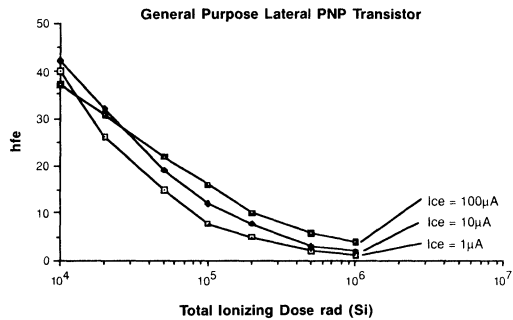


Figure 2.

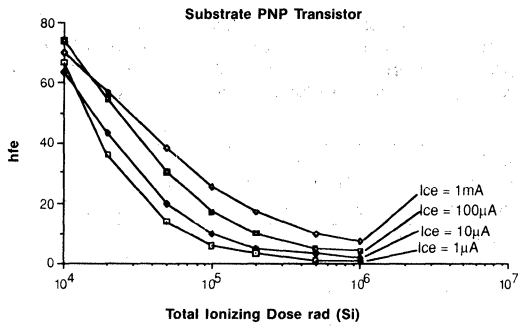


Figure 3.

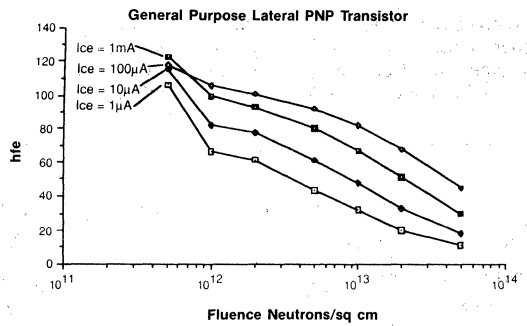


Figure 4.

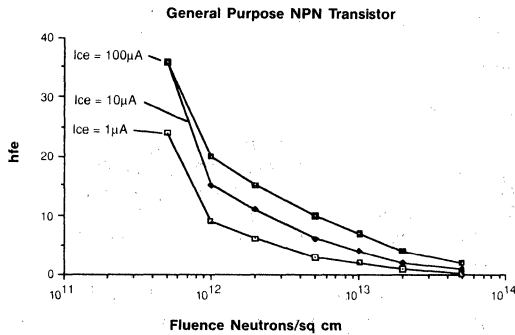


Figure 5.

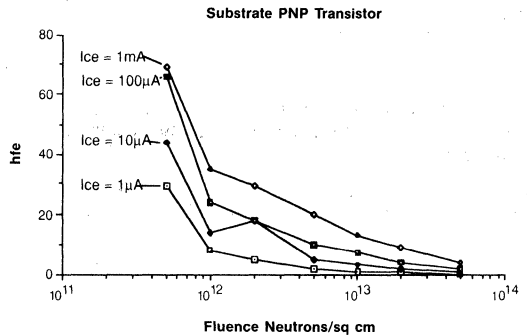


Figure 6.

/M8 Military Product Flow

Section 9

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/M8 Product Screening Flow	9-2
/M8 Manufacturing Flow	9-3
/M8 Product Qualification	9-5
Ordering Information	9-6





Military Product Flow

A specification to establish the general test methods and procedures for purchase of integrated circuits to military quality and reliability assurance requirements.

Micro Linear is committed to supplying the military marketplace with service, as well as, quality and reliable components second to none. The Micro Linear /M8 program is designed to provide off-the-shelf high integration linear integrated circuits with extended screening and testing utilizing the methods of MIL-STD-883C, Class B as its reference documentation.

The Quality and Reliability Assurance program at Micro Linear Corporation is a wide ranging program involving engineering and manufacturing designed to produce the highest quality linear integrated circuits available.

Wafer Inspection

Emphasis is placed on statistical analysis, electrical measurements on specially designed Process Control Monitors in accordance with MIL-STD-414 (sampling by variables), visual inspection, and film measurements. Potential reliability hazards are investigated and detected early by utilizing diagnostic and device structures on each wafer and periodic SEM analysis.

Assembly Inspection

Comprehensive receiving inspection for all materials and piece-parts is performed in accordance with the strictest quality assurance procedures. To assure conformance and control to specifications, documented quality control checks and monitors are performed on-line.

Testing

Micro Linear has invested in the latest "state-of-the-art" analog testers to achieve the most complete and thorough parametric testing of integrated circuits in the industry. Data sheets provide the customer a precise listing of parameters which are 100% tested. The calibration system is in compliance with MIL-STD-45662.

Traceability

For complete traceability to the wafer fab and assembly lot, a mark is placed on all units giving information on a unit-by-unit basis.

Micro Linear considers traceability to be essential for good engineering control and additional insurance for its customers. The information provided exceeds the seal week control required by MIL-M-38510.

ESD (Electro Static Discharge)

Products are fully characterized to MIL-STD-883C, Method 3015 and strict controls on handling and packaging are observed. A full ESD program, from design through manufacturing, incorporates training of all employees who handle Micro Linear products.

Major Change Control

Major change controls are in place to notify customers in accordance with MIL-M-38510. Micro Linear reviews all process, product, and package changes. All changes with possible impact are submitted for a re-qualification, which may include electrical, mechanical, and/or thermal characterization. Reliability requalification is performed if applicable.

Quality Assurance

Process monitors and gate inspections insure that all devices are properly tested and that the required sample tests are performed prior to shipment. Inspection records and reports concerning monitors and inspection data are utilized to status the quality level of products going through final test operations. Statistical sampling plans ensure the quality of the product.

Micro Linear welcomes OEM quality system surveys. Micro Linear is qualified by a number of customers to MIL-Q-9858A and MIL-I-45208 for military programs.

Failure Analysis and Reporting of Customer Returns

A formal program exists to record, analyze and take appropriate corrective action on all returns. A Corrective Action Committee reviews on a weekly basis all discrepancies and assigns responsibility to implement solutions or improvements. A report is generated and sent to the customer stating our findings and action.

/M8 Military Product Flow

/M8 Product Screening Flow

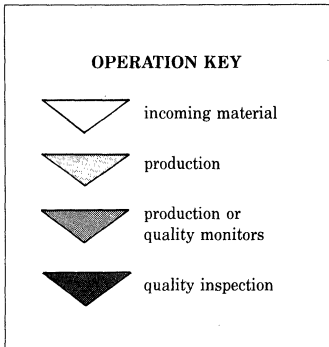
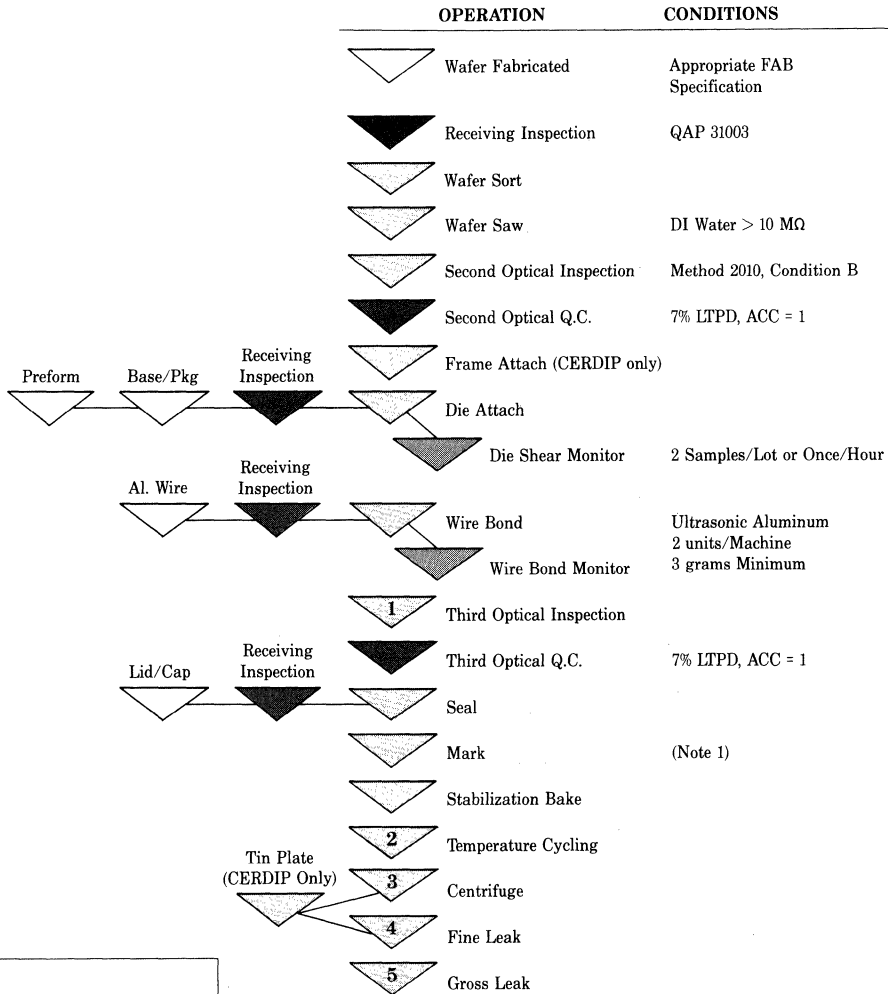
Micro Linear's standard extended screening process outlined below utilizes the methods of MIL-STD-883C, Class B as it's reference documentation.

Despite lower cost and faster delivery of the following standard /M8 flow, there are cases where a custom or special flow is required. Micro Linear is ready to discuss and accomodate custom flows to meet design or other mandatory requirements.

	OPERATION	CONDITION
1	100% Internal Visual	Method 2010, Condition B
2	100% Temperature Cycling	Method 1010, Condition C
3	100% Constant Acceleration, Y1 Orientation Only	Method 2001, Condition E
4	100% Seal Fine Leak	Method 1014, Condition A
5	100% Seal Gross Leak	Method 1014, Condition C
6	100% Pre Burn-In Electrical, 25°C	Data Sheet, 100% Noted Parameters
7	100% Burn-In, 160 Hrs at 125°C	Method 1015
8	100% Post Burn-In Electrical, 25°C	Data Sheet, Parameters Noted 100% Tested
9	Percent Defective Allowable Calculation	PDA = 5%
10	100% Final Electrical Test, -55°C and +125°C	Data Sheet Parameters Noted 100% Tested
11	QA Sample 116/0 Electrical Test, -55°C, +25°C, and +125°C	Group A, Subgroups 1, 2, 3, 4, 5, 6, 7, 8, 9, 10, 11 Data Sheet Parameters Noted 100% Tested
12	Quality Conformance Inspection Test Sample Selection	Method 5005.11 Group B
A	Resistance to Solvents	Method 2015
B	Solderability, Soldering Temperature of 245 ± 5°C	Method 2022 or 2003
C	Bond Strength, Ultrasonic	Method 2011, Condition C or D
13	Sample LTPD = 2, C = 1 Seal Fine Leak	Method 1014, Condition A
14	Sample LTPD = 2, C = 1 Seal Gross Leak	Method 1014, Condition C
15	100% External Visual	Method 2009
SHIP		




















/M8 Manufacturing Flow

Screening to the /M8 flow is part of the manufacturing flow shown below. The numbered steps correspond to the operations of the /M8 product screening flow.

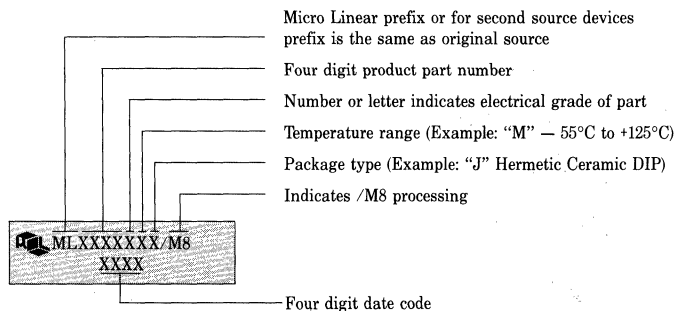
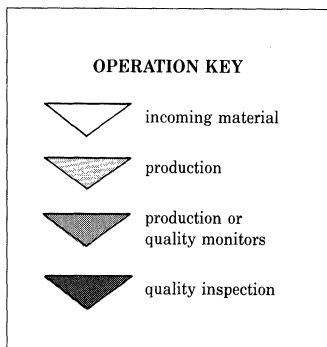


/M8 Military Product Flow

/M8 Manufacturing Flow (Continued)

OPERATION	CONDITIONS	
	Marking Permanency	
	Lead Trim	
	Visual Inspection	
	Quality Monitor	7% LTPD, ACC = 1
	Pack and Ship	
	Receiving Inspection	PUR 34001 (Note 2)
	Electrical Test	
	Burn-in	
	Electrical Test	
	PDA Calculation	
	Electrical Test, Min. and Max. Temperature	
	Q.A. Sample	
	Quality Conformance Inspection	
	Q.A. Fine Leak Sample	
	Q.A. Gross Leak Sample	
	External Visual	
	Finished Goods	
	Shipping Inspection	QAP 35030
	SHIP	

Note 1: Marking of product screened to /M8 test methods and procedures is as follows:



Note 2: Country of origin may be United States, Korea, Hong Kong or Thailand.

/M8 Product Qualification

Generic data can be provided for the following qualification conditions or methods.

Actual qualification on a given lot can be performed on a customer lot basis. Contact your Micro Linear sales representative for any additional price adder and delivery information.

Group C Die-Related Tests

Test	Condition	Quantity/Accept No.
1. Steady-State Life (Burn-In Circuit Available Upon Request)	Method 1005 1000 Hr at 125°C or equivalent	LTPD 5
End Point Life Test Electricals 25°C	Data Sheet, 100% Noted Parameters	

Group D Package-Related Tests

1. Physical Dimensions	Method 2016	LTPD 15
2. a. Lead Integrity	Method 2004	LTPD 15
b. Seal	Method 1014	
Fine	Condition A	
Gross	Condition C	
3. a. Thermal Shock	Method 1011 Test Condition B 15 Cycles	LTPD 15
b. Temperature Cycling	Method 1010 Test Condition C 100 Cycles	
c. Moisture Resistance	Method 1004	
d. Seal	Method 1014	
Fine	Condition A	
Gross	Condition C	
e. Visual Examination	Method 1004 Method 1010	
f. End Point Electricals 25°C	Data Sheet 100% Noted Parameters	
4. a. Mechanical Shock	Method 2002 Condition B	LTPD 15
b. Vibration, Variable Frequency	Method 2007 Condition A	
c. Constant Acceleration	Method 2001 Condition E Y1 Orientation	
d. Seal	Method 1014	
Fine	Condition A	
Gross	Condition C	
e. Visual Examination		
f. End Point Electricals 25°C	Data Sheet 100% Note Parameters	

/M8 Military Product Flow

REFERENCES

Government documents and specifications.

FED-STD-2090	Clean Room and Work Station Requirements, Controlled Environment.
MIL-M-38510	Micro Circuits, General Specification for.
MIL-Q-9858	Quality Program Requirements.
MIL-STD-414	Sampling Procedures and Tables for Inspection by Variables for Percent Defective.
MIL-STD-883	Test Methods and Procedures for Microelectronics.
MIL-STD-11331	Parameter to be Controlled for the Specification of Microcircuits.
MIL-STD-45662	Calibration Systems Requirements.

ORDERING INFORMATION

Product processed to the /M8 flow is ordered by adding /M8 to the standard product part number.

Example:



All /M8 product are shipped with a certificate of conformance. Information with regard to non-standard electrical testing or preconditioning, and wafer traceability may be obtained by contacting your Micro Linear sales representative.

Application Notes

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FB3600 Digital Logic Design

Micro Linear's Bipolar ASIC Technology allows the mixture of both analog and digital circuitry on an integrated circuit. Micro Linear has combined the advantages of TTL, and ECL logic on our FB3600 family of bipolar tile arrays. Our +5 volt version of ECL can interface to the outside world at standard TTL, CMOS or 10K ECL levels. It requires the use of only a standard +5 volt power supply. On-chip, gate propagation delay times as low as 2 nanoseconds are possible. High density ECL digital components occupy fifty percent of Micro Linear's FB3635 tile array. In addition, a certain amount of digital logic can be implemented on all of the FB3600 tile arrays.

Traditionally, 10K ECL logic uses -5.2 volts supply. This additional supply is only needed for applications requiring an external ECL logic interface. Our FB3635 and FB3621 tile arrays contain schottky components. These components are often useful for implementing high speed TTL & CMOS output drivers. On-chip ECL Logic requires a voltage reference which changes over temperature. Normally, on-chip voltage references are designed to be stable over variations in temperature. The schematic diagram for this circuit has been provided.

This application note has been designed to aid a design engineer using a workstation with Micro Linear's analog ASIC design libraries. The circuits provide basic building blocks which can be integrated on our FB3600 family of tile arrays. The circuitry and discussion provided in this application note provide a starting point for the design engineer's own workstation circuit design and simulations.

Two Input ECL NOR Gate

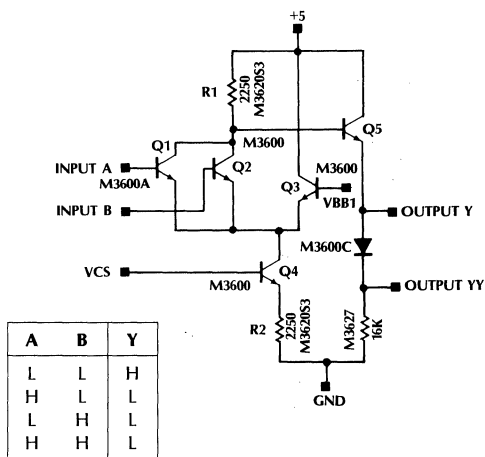
One of the major advantages of ECL logic is that the transistors never saturate. This plus the small signal swings reduce the propagation delay time through the gate. The propagation delay can be adjusted by changing the current level used by the circuit. The gate propagation delay decreases as the operating current level increases.

The ECL NOR gate, shown in figure 1, is designed for conventional +5 volt power supply operation. VCS is a preset bias voltage of 1.35 volts. This develops a voltage drop of 750mV across R2. The two ECL inputs (Input A & Input B) have a logic high (true) value of 4.25 volts and a logic low (false) value of 3.50 volts. VBB1 is a preset bias voltage which is about half way between the logic high and logic low voltages.

Micro Linear's single +5V operation is different from traditional 10K ECL logic which uses a single -5.2 volt supply. This establishes the 10K ECL logic high level in between -.810 and -.960 volts and a logic low level is in between -1.650 and -1.850 volts.

The NOR gate operates on the current flow from Q4. All the current from Q4 will be steered through either the Q3 leg or the Q1/Q2 leg of the circuit. If either Input A

Figure 1. NOR Gate



or Input B logic voltage is high, all the current will flow up the Q1/Q2 leg of the circuit. This occurs because either or both transistors (Q1, Q2) have an input voltage which is above Q3 input voltage. Current flowing up the Q1/Q2 leg will cause a 750mV voltage drop to occur across R1 (same resistance as R2). This also results in Output Y being set at 5 volts minus 750mV minus 750mV (Q5 base to emitter voltage drop). Thus, Output Y is set at a logic low level (3.05 volts).

Both inputs need to be logic low for Output Y to have a logic high result (4.4 volts). In this case, the voltages on both Q1 and Q2 bases are less than the voltage on the base of Q3. This will cause all the current from Q4 to flow up the Q3 leg. The base of Q5 will be about 5 volts since the voltage drop across R1 is close to zero.

It is important to note that R1 always equals R2 and that the voltage drops (typically 750mV) across the base emitter will change with temperature. Since all the transistors on this IC are about the same temperature, they and the ECL voltage references will all track together with temperature. Thus the ECL logic works well over variations in temperature. The absolute values of the voltages stated above will change slightly with temperature. The values of resistors R1, R2, and R3 are adjusted for the desired speed vs power tradeoffs. The values shown in the NOR gate (figure 1) are typical values.

Figure 1 also shows an Output YY terminal. Some ECL logic gates need to have an extra diode voltage drop for its output. We will call this the "bias level B" output/input. The Output Y terminal does not have this extra diode voltage drop. Thus, we will call this the "bias level A" output/input.

FB3600 Application Note

Two Input ECL NAND/AND Gate

The basic operation of this gate's differential pair and the two output stages is very similar to the NOR gate discussion. The NAND/AND gates input stage requires Input A to be a "bias level A" input and Input B to be a "bias level B" input. A "bias level A" input needs to be driven by a "bias level A" output. Similarly, a "bias level B" input needs to be driven by a "bias level B" output.

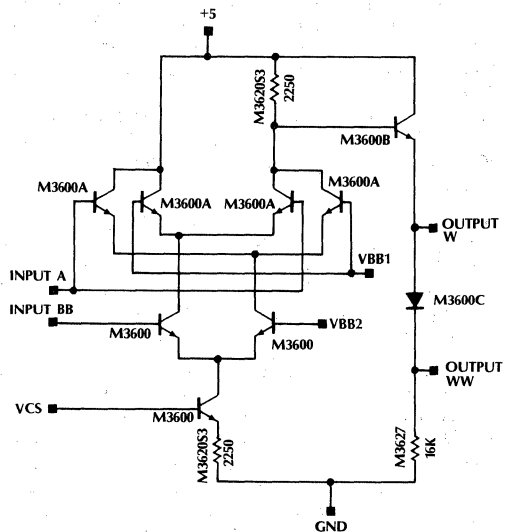
The NAND/AND gate shown in figure 2 has two output sections. The NAND output uses section A output stage. The AND output uses section B output stage. This gate can have either output stages omitted.

The NAND gate has its "bias level A" result on Output X and its "bias level B" output on Output XX. Similarly, the AND gate has its "bias level A" result on Output W and "bias level B" output on Output WW.

Two Input ECL XOR Gate

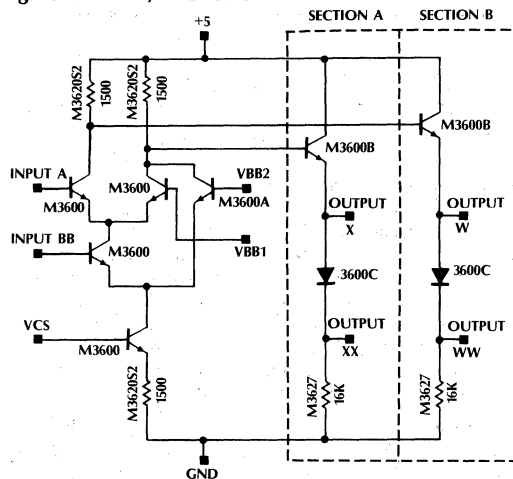
The basic operation of this gate's two differential pair and the output stages is very similar to the NAND/AND gate discussion. Figure 3 contains a circuit diagram for this gate. It uses one "bias level A" (Input A) input and one "bias level B" input (Input BB). The gates output is available as "bias level A" (Output W) and as "bias level B" (Output WW).

Figure 3. XOR Gate



A	B	W
L	L	L
H	L	H
L	H	H
H	H	L

Figure 2. NAND/AND Gate



A	B	X	W
L	L	H	L
H	L	L	L
L	H	L	L
H	H	L	H

ECL Data Latch

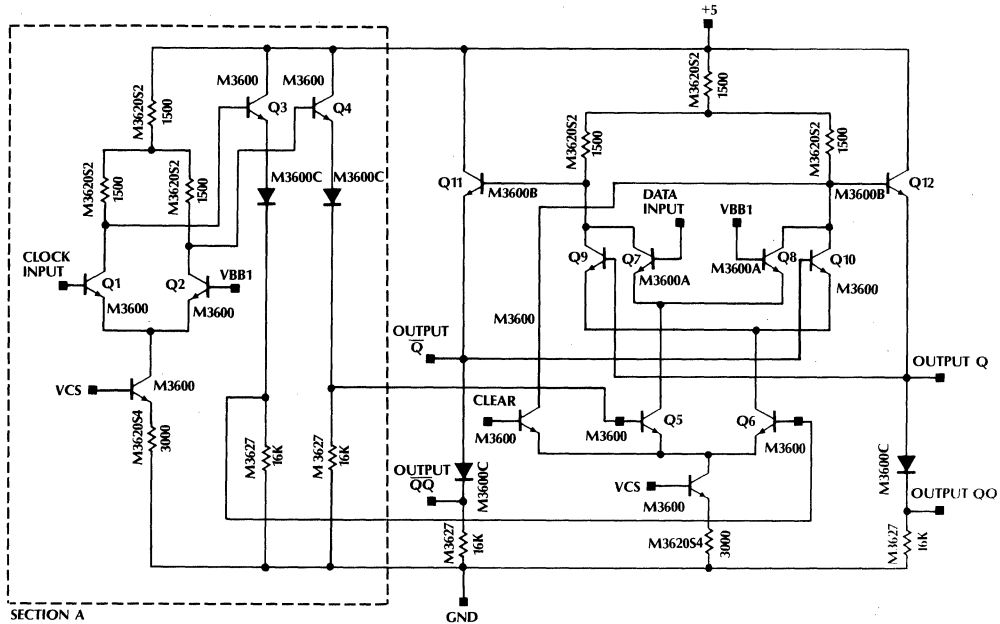
The circuit diagram for a single bit ECL data latch is shown in figure 4. As long as the Clock Input is logic high, the data latch will pass the data from the Data Input through to the output. If the input data changes, the output will track the change. This is called the pass through mode of operation. The pass through mode will end as soon as the Clock Input signal changes to logic low. When this transition occurs, the current value of input data will latch. The data latch will remain fixed as long as the clock remains low. Should the Clock Input return to the high state, the data latch will return to the pass through mode of operation. The data latch is level triggered instead of edged triggered.

Both the Clock Input and the Data Input are "bias level A" inputs. The "bias level A" outputs are Q and \bar{Q} . The "bias level B" outputs are QQ and $\bar{Q}\bar{Q}$.

When the Clock Input is high, the current value of the Data Input will be present at the Q and QQ outputs. An inverted version of Data Input will be present at the \bar{Q} and $\bar{Q}\bar{Q}$ outputs. When the Clock Input is low, the latched value of the previous Data Input will be present at the Q and QQ outputs.

The data latch circuit contains circuitry to adjust the Clock Input signal. The circuitry shown in section A contains a circuit for converting a "bias level A" logic input into two "bias level B" output signals. The two output signals reflect the input signal and a complement of the input signal.

Figure 4.



The “bias level A” clock input signal drives the base of Q1. Transistor Q1 and Q2 form a differential pair. The base of Q2 is driven by a reference voltage which is midway between logic high and logic low. When the Q1 input signal is high, the current will flow only through the Q1 leg of the differential pair. This will cause the collector of Q1 to have a voltage of about 4.25 volts and the collector of Q2 to have a voltage of about 5 volts. Substantial current will now flow through Q3. Thus, a “bias level A” logic low is present at the emitter of Q3. The diode in series with Q3 emitter shifts the output voltage to a “bias level B” output. This “bias level B” output will have a logic low value. Note that the Q3 outputs represent the complement of the Clock Input signal. Thus, a low Clock Input signal will result in a logic high output at the emitter of Q3.

A buffered version of the Clock Input signal is provided. This output will have the same logic level as the Clock Input signal. A “bias level A” version of the Clock Input signal is available at the emitter of Q4. The diode in series with Q4 emitter shifts the output voltage to a “bias level B” output.

The buffered Clock Input signal and its buffered complement will drive the bases of Q5 and Q6, respectively. When the data latch is in the data pass through mode (Clock Input high), transistor Q5 is turned on and transistor Q6 is turned off. If the Data Input is logic high, all of the current in the differential pair (Q7 & Q8) will flow in the Q7 leg. The current flow through the

resistor in the Q7 leg will produce a 750mV drop. This sets the collector of Q7 at 4.25 volts. This will cause Q11 emitter to be at “bias level A” logic low (output Q). Output $\overline{Q\overline{Q}}$ will be “bias level B” logic low. The lack of current flow in the Q8 leg will cause the collector of Q8 to be at about 5 volts. This will cause the emitter of Q12 to be logic high (“bias level A”). Output QQ will be at “bias level B” logic high.

When the Data Input is logic low, then all of the current will flow through the Q8 leg. Transistor Q11 emitter will now be at logic high. Output $\overline{Q\overline{Q}}$ will be at “bias level B” logic high. Transistor Q12 emitter will now be at logic low and the output QQ will be at logic low.

The data latch will store the current state of the output when the Clock Input signal changes to logic low. This will cause transistor Q5 to turn off and transistor Q6 to turn on. The base of Q9 gets its input from output Q. The base of Q10 gets its input from the output \overline{Q} . Since it takes a few nanoseconds for Q11 and Q12 to change state after the input data changes, the data latch is now getting its input data from the previous output data. This feedback loop causes the data latches output to remain fixed.

The data latch also contains a CLEAR input. This input should normally be logic low (“bias level B”). A logic high will reset the data latch to logic low. As long as the CLEAR input is logic high, the data latch will remain reset.

FB3600 Application Note

One Bit ECL Register or Flip Flop

The circuit shown in figure 5 can be used as a single bit positive edge triggered register or as a flip flop. We shall first review its operation as a one bit register. This circuit latches the Input Data upon the Clock Input changing from logic low to logic high. The data will remain latched until the next time the Clock Input changes from logic low to logic high. Similar to the data latch circuit, the Clock Input and Data Input signal are both "bias level A" inputs. The register has four outputs. The outputs are available in both "bias level A" (Q and \bar{Q}) and "bias level B" (QQ and \overline{QQ}). The register's stored value (Q and \bar{Q}) and its complement value (Q and \bar{Q}) are also provided.

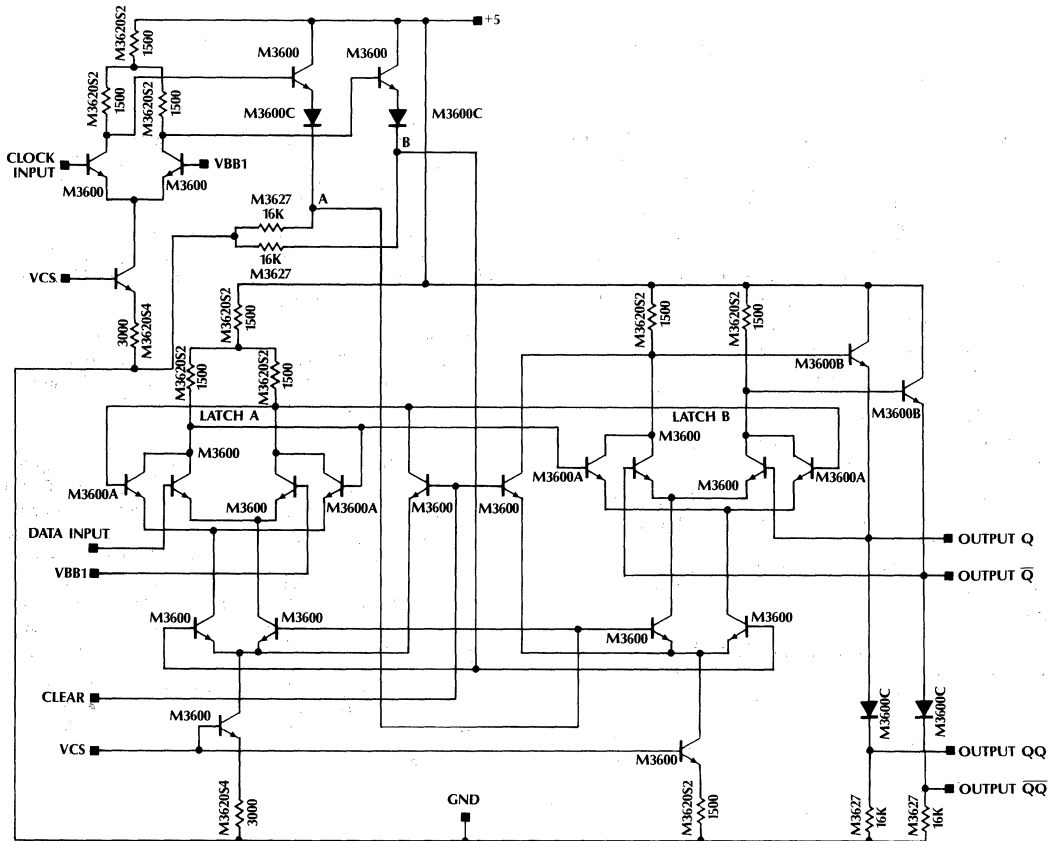
This circuit is simply two data latches in series. Both data latches use a common clock. When one latch is in the data pass through mode, the other latch is latched. When the Clock Input signal is high, data latch A is latched and data latch B is in the pass through mode. When the Clock Input signal is low, data latch A is in the data pass through mode and data latch B is latched. If the Clock Input signal

changes from low to high, latch A will latch its current input data and data latch B will pass data latch A output values directly to its output. This can change the data register's output. If the Clock Input signal changes from high to low, the output data will not change since latch B will latch itself using data provided by latch A previous outputs. Note that a flip flop can be implemented by connecting the register's Q output to the register's Data Input.

The register also contains a CLEAR input. This input should normally be logic low ("bias level B"). A logic high will reset the register to logic low. As long as the CLEAR input is logic high, the register will remain reset.

The Clock Input circuit has two "bias level B" outputs (point A and B). If these output connections are switched, the register will latch upon a logic high to low transition. This will cause it to be in pass through mode whenever the Clock Input is low. The data register will latch its data whenever the Clock Input is high.

Figure 5. Register or Flip Flop

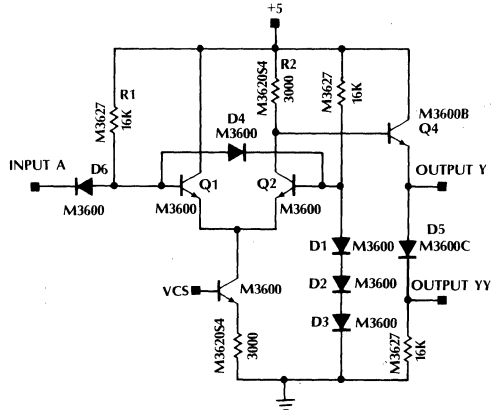


TTL and CMOS Input Interface Circuit

This circuit, shown in figure 6, converts a TTL or CMOS logic level input into an on-chip ECL level input. Output Y is a "bias level A" version of Input A. Output YY is a "bias level B" version of Input A. Transistors Q1 and Q2 forms a differential pair. The string of three diodes (D1, D2, and D3) sets the base of transistor Q2 at 2.25 volts. Given a high TTL/CMOS logic level drive at Input A, resistor R1 will bias the base of Q1 to be above the base Q2. This will cause all of the current to flow through the Q1 leg of the differential pair. Since the voltage drop across R2 will be about zero, the base of Q4 will be at 5 volts. The emitter of Q4 will be "bias level A" logic high. A "bias level B" version of this output will be produced at Output YY.

When Input A is driven by a TTL/CMOS low logic level, the base of Q1 will be biased below the base of Q2. Now all of the current will flow through the Q2 leg of the differential pair. The current flow will cause a 750mV voltage drop to occur across R2. The emitter of Q4 (Output Y) is now set at a "bias level A" logic low. Diode D5 produces a "bias level B" version of this output.

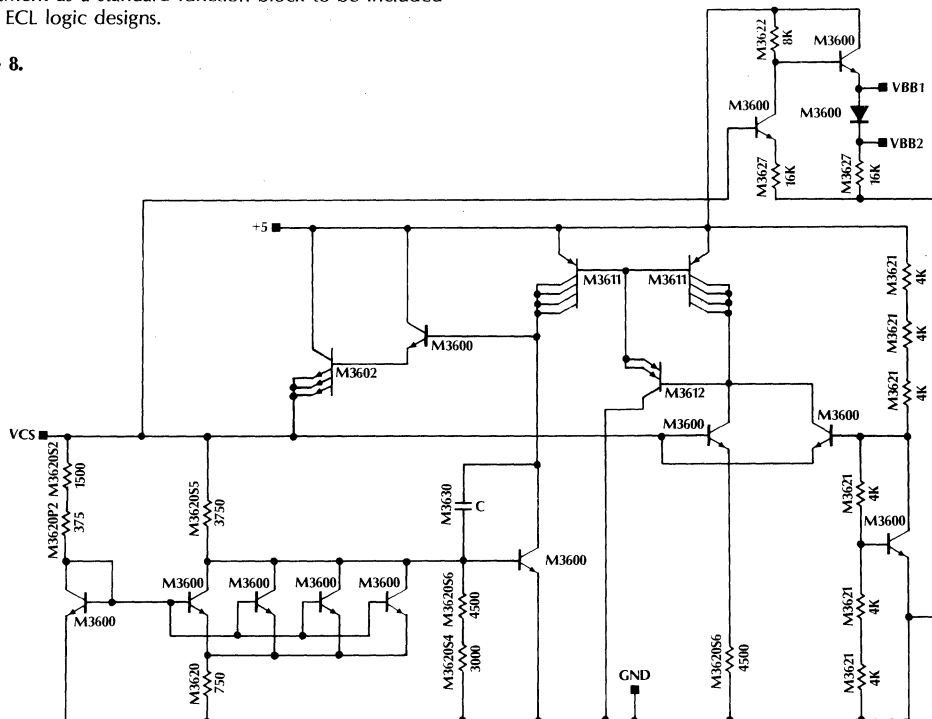
Figure 6. Input Interface



Voltage Reference for FB3600 ECL Logic

The circuit shown in figure 8 supplies the necessary reference voltages for our ECL logic. It has been designed to vary the output voltage with temperature. This block has been designed by Mirco Linear's engineering department as a standard function block to be included on all ECL logic designs.

Figure 8.



On-chip ECL to TTL or CMOS Output Interface

The circuit shown in figure 7 takes an on-chip "bias level A" ECL input and produces a TTL/CMOS compatible output. If the input is logic high (true) then the output will be logic high (true). The circuit simply buffers and shifts the logic voltage level from on-chip ECL voltage levels to TTL/CMOS voltage levels.

Section A contains circuitry similar to the CLOCK input for the data latch circuit. It converts the input into two signals (buffered version and a complement buffered version). If the input is driven by a high logic level, Q1 will be turned on and Q2 will be turned off. This causes the base of Q3 to be 4.25 volts and the base of Q4 to be 5 volts. Transistors Q3 and Q4 drives the bases of Q5 and Q6 respectively. Since the base voltage of Q4 is greater than the base voltage of Q3, the base voltage of Q6 will be greater than the base voltage of Q5. This will cause all of the current in the Q5 & Q6 differential pair to flow in the Q6 leg. If Input A is driven by a logic low level, all of the current in the Q5 & Q6 differential pair will flow in the Q5 leg.

The collector of Q6 drives the final output circuitry in section B. When Input A is at logic high, the collector of Q6 will be at .9 volts. The voltage drop across the base and emitter (.75 volts) of Q7 and Q8 will result in the bases of Q11 and Q12 being driven by about .9 volts. This will turn on Q11 and Q12. With Q11 turned on, Q13 will be turned off and Q12 will be switched on. The output voltage will be about .75 volts.

If Input A is at logic low, the collector of Q6 will be at 1.5 volts. The voltage drop across the base and emitter of Q7 and Q8 will result in the bases of Q11 and Q12 being driven by about .9 volts. This will turn on Q11 and Q12. With Q11 turned on, Q13 will be turned off and Q12 will be switched on. The output voltage will be about .75 volts.

We have just reviewed how section A circuitry drives the differential pair of Q5 & Q6. We have also reviewed how the collector of Q6 drives the output drive circuitry contained in section B. Next, we will examine how the Q5 & Q6 differential pair have been biased.

Section D contains a circuit known as a base emitter voltage multiplier. The voltage at the collector of Q15 will be determined by the following equation,

$$\text{Voltage at collector of Q15} = [1 + (R1/R2)] \times .75$$

The value of R1 and R2 is 10K Ω and 4K Ω respectively. The .75 represents the typical voltage drop across a transistors base to emitter. This sets voltage at the collector of Q15 at 2.6 volts. The voltage drop across the base and emitter of Q14 will set the collector voltage of Q5 at 2.6 - .75 = 1.85 volts.

When all of the current flows in the Q5 leg of the differential pair (Q5 & Q6), there will not be a voltage drop across the circuitry in section E (no current flow). The collector of Q6 is now set at 1.85 volts. If all of the current flows in the Q6 leg of the differential pair, there will be a .75 voltage drop across the diode. This sets the collector of Q6 at 1.1 volts.

The circuitry in section F provides a bias current for a current mirror. Resistor R3 value was chosen for a .5mA current flow with a 4.4 voltage drop across it. This input bias current generates the base to emitter voltage for Q16 which drives the bases of Q17, Q18, Q9 and Q10. Each of these transistors will sink .5mA.

Transistors Q19, Q21 and Q20 also form a current mirror. Transistor Q20 and its 8K emitter resistor have been added for stability. The purpose of the current mirror is to keep the current flowing through Q3 and Q4 approximately equal.

Transistors Q11 and Q12 are schottky clamped transistors. They consist of a npn transistor with a schottky diode connected between the base and the collector. The function of this diode is to limit the current flowing into the base. This prevents the transistor from saturating. The schottky diode sends the excess base current into the collector. This limits the voltage drop across the collector and the emitter to about 200mV. The typical base emitter voltage drop is .75 volts. These devices can be replaced with regular npn transistors if the logic's switching rates are low (a few MHz). Saturated transistors have much slower switching times than non saturated transistors.

Voltage Reference for FB3600 ECL Logic

The circuit shown in figure 8 supplies the necessary reference voltages for our ECL logic. It has been designed to vary the output voltage with temperature. This block has been designed by Mirco Linear's engineering department as a standard function block to be included on all ECL logic designs.

Trimming Analog Bipolar Arrays

High performance analog integrated circuits are becoming a necessity in the design of state of the art analog/digital systems. With standard analog IC's this requires the designer to specify premium performance parts. These same premium performance circuit functions are not typically available in semicustom arrays. By utilizing trimming techniques, though, improved performance can still be obtained. Trimming analog bipolar arrays is a very viable, cost effective approach for improving the key parameters of a circuit.

If tighter specifications are required than can be obtained using good design techniques the circuit may be trimmed at the wafer level by a technique of selectively shorting zener diodes. This is known commonly as zener zapping. This technique can be used to trim the input offset voltage of an op amp or the output voltage of a precision reference. For example, the offset voltage of our MLC350⁽¹⁾ operational amplifier can be trimmed from a maximum of 7mV to less than 1mV. The MLC340 voltage reference can be trimmed to an accuracy of better than 1%. Many types of parameters may be trimmed within the limitations of the technique as described below.

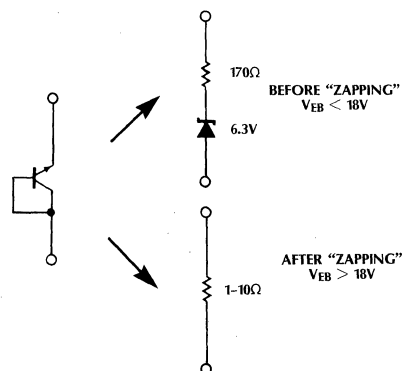
Although there are other ways in which a bipolar integrated circuit may be trimmed, zener zapping has become well established because it does not require extra processing steps and can be implemented at the wafer level. Unlike laser trimming, the technique is not limited to altering a resistive element, and does not require a large capital investment. Fusible links, another well-established method, requires currents in the ampere range in order to blow the standard 1 micron thick aluminum, resulting in a questionable blown connection. A thinner link would require additional wafer processing steps.

The Zener Zapping Technique

This process is called zener zapping because the emitter-base diode of a bipolar transistor is permanently shorted by passing a relatively large current through it while in the reverse breakdown avalanche mode. It produces a reliable 1-10 ohm link between the emitter and base pads. (See Fig. 1) This is a very reliable connection because of the double short which actually occurs. The first short is caused by the destruction of the pn junction. In addition, the presence of a large electric field during thermal runaway causes metal to migrate across the silicon surface beneath the oxide layer, producing a second short.

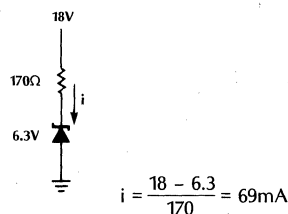
This set of events occurs when the voltage across the emitter-base junction is increased beyond the 6.3V

Figure 1.



avalanche breakdown point, to above 18V. At about 18V the instantaneous power dissipation exceeds 1.2W (figure 2) and an oscillatory, thermal runaway condition occurs. In less than a second the junction is destroyed leaving a 1-10 ohm short. The current required is less than 300mA, so remote probe pads (the bonding pads) can be used without damage to the pads or traces.

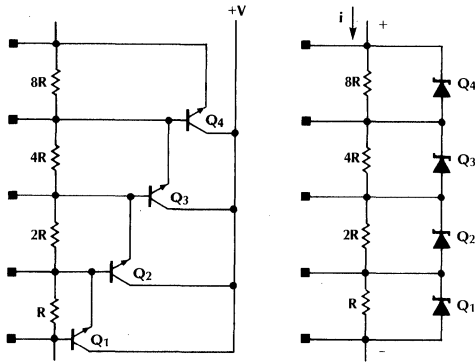
Figure 2.



$$\text{Power Dissipation} = iV = (69\text{mA})(18\text{V}) = 1.24\text{W}$$

The circuit in figure 3 illustrates a simple implementation of this technique to alter the total resistance of a circuit path. Before any of the zeners are blown the total resistance equals 15R. This value can be altered to equal any integer multiple of R from 1R to 15R by selectively blowing only four zeners in a binary fashion. This is possible due to the binary arrangement of the resistor values. For example, to obtain a resistance of 5R, Q₂ and Q₄ should be shorted resulting in 4R + 1R = 5R. Note that 5 equals 0101 in binary which is represented by Q₄, Q₃, Q₂, Q₁ with shorts being 0's and opens being 1's.

Figure 3.



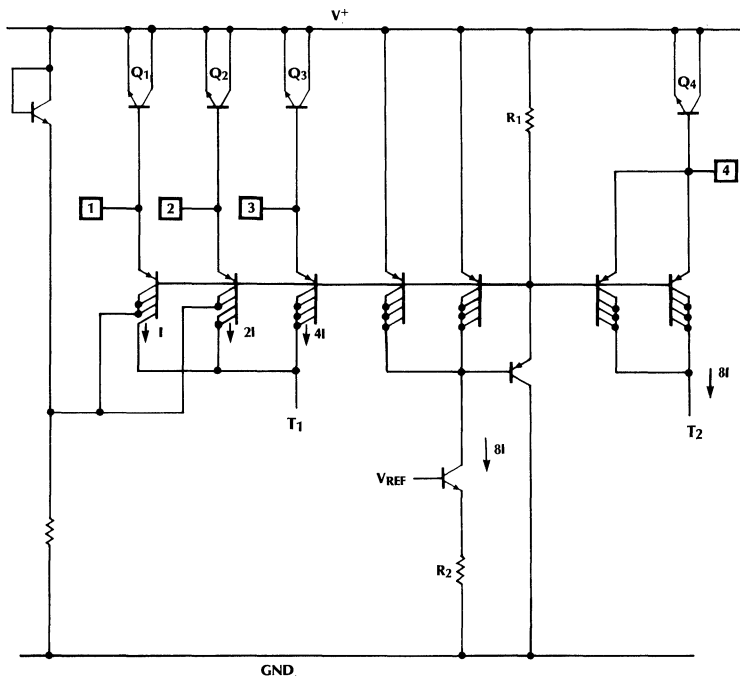
This type of circuit arrangement has two restrictions of which the designer should be aware. During normal operation of the circuit, the current through the resistor string should not be allowed to flow opposite to the direction indicated in the drawing. This would forward bias the base-emitter junction of the transistors, and effectively short out a resistor intended to be used. In

addition, the forward voltage drop across each of the resistors should not exceed the zener breakdown voltage, about 6.3V. This would allow current to flow out of the resistor string and through the zener, altering the intended operation of the circuit.

The preceding example illustrates the use of zener zapping to alter a resistive element in a circuit. In many cases modifying a current source is a more useful way of trimming a design. Figure 4a shows trimmable current sources used to reduce the input offset voltage of an op amp.

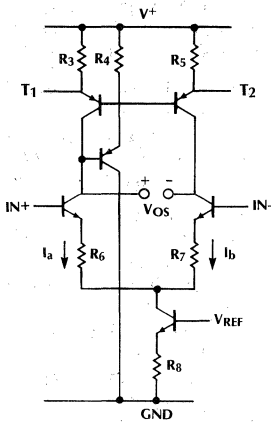
In this example the balance of current in two circuit paths is altered using zener zapping. This technique is particularly useful for reducing the input offset voltage of an operational amplifier which has added emitter degeneration in the input stage in order to improve slew rate (figure 4b). The emitter resistors used in this circuit, R_6 and R_7 , will contribute significantly to the offset voltage of the input stage. By modifying the balance of current between I_A and I_B the increased offset voltage V_{OS} can be compensated. In this example there are again 4 bits of trimming resolution with the 3 least significant bits controlling one current path and the most significant bit controlling the other. With this configuration the balance of current can be altered in either direction. In other words, the current in T_1 can be increased or decreased relative to T_2 .

Figure 4a.



Application Note 2

Figure 4b.



To increase the current through T_1 relative to T_2 you simply short Q_1 , Q_2 , and Q_3 in a binary fashion to get increments of I from $1I$ to $7I$. To increase the current through T_2 relative to T_1 you short Q_4 which increase the current through T_2 by $8I$. Then if you want less current, short Q_1 , Q_2 , and Q_3 in a binary fashion to offset the increase in T_2 by increments of I down to $1I$.

The source of current in the trim circuit should be of the same type as the circuit to be trimmed so their temperature coefficients will match. In this case the V_{REF} in both circuits should be the same, and resistors R_2 and R_8 should be of the same type.

Shorting zeners is an irreversible process. Thus, it is important to check the results of a trim bit pattern before actually destroying the junctions. This is done by shorting the probe pads externally in the desired pattern through relays. In this way all combinations can be tried and the best results can be chosen and implemented.

Although these two examples both use 4 bits (4 zeners) to trim a circuit, any number can be used to get more or less resolution. The designer should be aware though of the practical limitations of each circuit to be trimmed. Other error terms like temperature coefficients will eventually become significant, and additional trimming beyond this point would be fruitless. In addition, the more zeners you use, the more probe pads are required. In a full custom circuit where minimum die size is the ultimate goal, the additional die area required for the diodes, pads and trim circuitry may become significant. An array, however, typically has unused components available for the trim circuitry, and you only have to be concerned with the number of bonding pads available. If all of the pads are already being used for pinouts then a larger array would be required.

These examples illustrate the usefulness and flexibility of zener zapping. There are many other potential applications for this technique though, and with a good understanding of the basic diode shorting process the design engineer can be creative in its application.

⁽¹⁾ The MLC350 is one of the circuits in Micro Linear's library of macrocells. Performance details of this circuit and other macrocells can be found in the FB300 Macrocell and Component Library booklet.

Design Techniques for Low Input Bias Current

Analog systems often require high impedance inputs to accommodate the demand for higher accuracy. Measurement systems which interface to photodetectors or high impedance transducers require devices with low offset voltage and low input bias current. This is necessary to receive and amplify the signal without introducing any significant errors. Under this constraint, the designer will often select a FET as the primary input device. Although a FET input stage may be appropriate in a discrete circuit design, there are other all bipolar techniques which are just as effective and better suited to an analog array. In some cases, these techniques will out perform the FET alternative.

This application note describes three alternatives for obtaining low input bias currents. The design techniques described can be applied to many different types of circuits from simple emitter followers to complex amplifiers. For example, a typical all bipolar operational amplifier can achieve input bias currents of about 100nA with an offset voltages of about 1mV⁽¹⁾. Unfortunately these characteristics are still not good enough for many of the applications previously mentioned. The input bias current can be minimized by using one of the following design techniques, 1) reducing the collector current 2) using a Darlington configuration 3) employing current cancellation techniques. This document will briefly describe the first two methods but will provide a detailed analysis of the cancellation technique as it provides the best performance trade-off and is the most involved.

Reducing the Collector Current

The simplest approach to achieve low input bias current is to reduce the collector current of the input transistors. Since the base current tracks the collector current by a factor of beta, reducing the collector current of the input transistors will reduce the input bias current into the bases. Beta will degrade at lower collector currents (figure 1), however, setting a practical limitation on this technique at about 50pA base current. If the circuit does not require a high slew rate or high gain bandwidth, this may be an acceptable method.

The Darlington Configuration

Figure 2 shows a differential Darlington configuration which will reduce the input bias requirements by a factor of beta. It will also double the offset voltage and reduce the voltage gain by 2. The offset voltage doubles due to the additional mismatching of the added devices, while the voltage gain suffers because only one-half of the input signal appears across the inner pair of transistors. A higher slew rate and gain bandwidth, though, can be achieved with this technique, over simply reducing the collector current, but it requires more components.

Figure 1. Current Gain vs. Collector Current

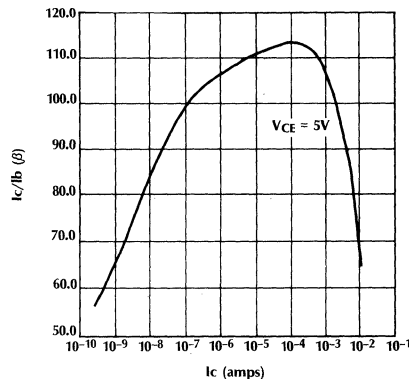
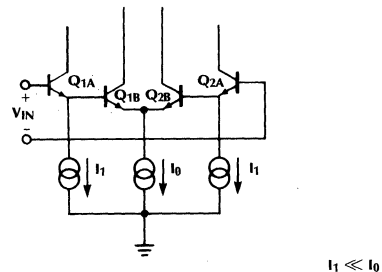


Figure 2.



The Cancellation Technique

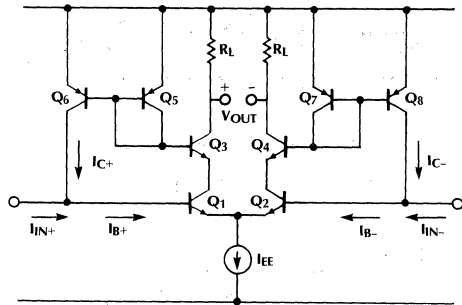
An all bipolar solution to low input bias current with low offset voltage while maintaining high collector currents for noise, slew rate or bandwidth reasons requires a technique called Input Bias Current Cancellation.

Input bias current cancellation is a circuit design technique which measures the input current and forces an equivalent amount back into the input nodes (figure 3). Ideally, this results in perfect cancellation of the input current. In the circuit in figure 3, the base currents into Q_3 and Q_4 duplicate the base currents into Q_1 and Q_2 . These currents are then sensed by Q_5 and Q_7 and equivalent currents are fed back, via Q_6 and Q_8 , into the input nodes. The total current at each input is thus,

$$I_{IN} = I_B - I_C$$

Application Note 3

Figure 3.



Assuming all PNP betas (β_P) are equal, all NPN betas (β_N) and all base-emitter voltage drops are equal,

$$I_C = I_B \frac{\beta_N}{(1 + \beta_N)} \frac{\beta_P}{(2 + \beta_P)}$$

$$I_{IN} = I_B \left(1 - \frac{\beta_N}{(1 + \beta_N)} \frac{\beta_P}{(2 + \beta_P)} \right)$$

If all betas are very high,

$$\begin{aligned} I_C &= I_B \\ \text{so} \\ I_{IN} &= 0 \end{aligned}$$

The main contributor to cancellation errors in this circuit is the low beta of the PNP devices. This sets a practical limitation on this technique at about 5-10% of the uncanceled current, as shown by the following example.

Assumptions: $\beta_N = 100$, $\beta_P = 30$, $I_B = 70\text{nA}$

$$I_{IN} = 70 \left(1 - \frac{100}{1 + 100} \frac{30}{2 + 30} \right)$$

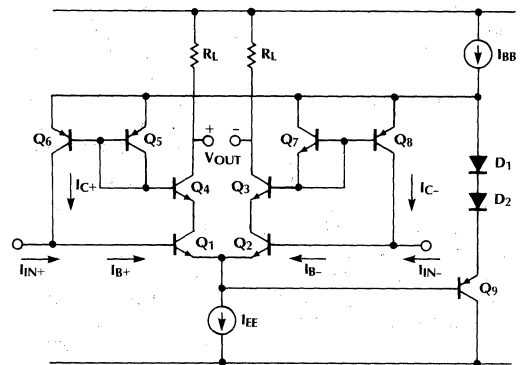
$$I_{IN} = 70 (.0718)$$

$$I_{IN} = 5\text{nA}$$

This technique does not reduce the input offset current. In fact, the additional circuitry, with its additional mismatches, increases the offset current by a factor of about 3. The input bias current can be reduced to about the same value as the offset current, setting the limitation on this technique at about 1-10nA.

In the equations above, the betas of PNPs in the current mirror were assumed to be equal. To enhance the viability of this assumption, the V_{CE} of each PNP should be kept equal. With the cancellation circuitry tied to the positive supply the V_{CE} of Q_6 and Q_8 will change with the input supply voltage, while the V_{CE} of Q_5 and Q_7 will remain constant. This further aggravates any beta mismatch which already exists. To reduce this effect the circuit can be self-biased using current source I_{BB} , diodes D_1 and D_2 , and transistor Q_9 , as shown in figure 4.

Figure 4.



This circuit keeps the voltage across the cancellation circuitry fixed as the input common mode voltage changes, which in turn keeps the beta of each device constant.

These techniques for reducing input bias current demonstrate the reality of achieving levels sometimes thought only possible with JFETs or MOSFETs. Circuits being considered for analog array integration which contain discrete FETs or FET input op amps should not be categorized as not possible. Rather, each individual circuit should be analyzed for its critical parameters, keeping in mind the trade-offs described above. If none of the bipolar solutions is adequate an external FET can always be used as an input buffer.

⁽¹⁾ The offset voltage can be reduced by making use of wafer trimming techniques. At Micro Linear a process called zener zapping is used. For more information about this process see the application note titled "Trimming Bipolar Arrays".

High Frequency Complex Filter Design Using the ML2111

Charles Yager
Carlos Laber

1.0 Introduction

Switched capacitor filters have been growing in popularity because of their advantages over active filters. Switched capacitor filters don't require external precision capacitors like active filters. Their cutoff frequencies have a typical accuracy of $\pm 0.3\%$, and they are less sensitive to temperature changes. This allows consistent, repeatable filter designs. Another distinct advantage of switched capacitor filters is that their cutoff frequency can be adjusted by changing the clock frequency. Switched capacitor filters offer higher integration at a lower system cost.

Until the introduction of the ML2111, commercially available switched capacitor filters were limited to about 20 KHz center frequencies. The ML2111 uses the versatile architecture of the MF10 with enhanced performance to reach center frequencies of up to 150 KHz with Q values up to 20.

Designing high frequency, high order filters using the ML2111 is the main topic of this application note. Particular attention is focused on mode 1c, which has the advantage of operating at high frequencies while allowing the center frequency to clock ratio to vary based on external resistors. A flexible building block is introduced which implements all the necessary types of bi-quads to realize high order complex filters. Finally an example is given which illustrates the design of an eighth order Elliptic bandpass filter with a center frequency of 90 KHz and a passband from 81 KHz to 100 KHz.

The first part of the application note covers a variety of issues: layout, how fast the system clock can be changed for sweeping filters, and some differences between continuous and sampled data filters. For the reader who is already familiar with sampled data filters, section 2 on Effects of Sampling, Aperture, Aliasing, and Signal Reconstruction may be skipped.

2.0 Effects of Sampling

Since the ML2111 is a switched capacitor filter, it behaves as a sampled data system. Switched capacitor filters, as opposed to digital filters, are analog sampled data systems. The signal remains in the analog domain, as the charge on a capacitor. Whether using an analog or digital sampled data system, the effects of sampling the signal must be considered.

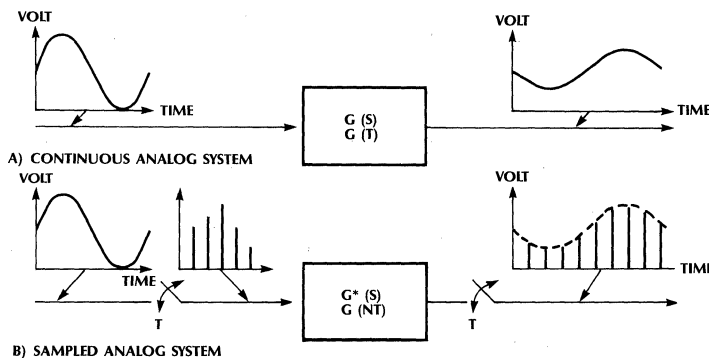
Figure 1 shows a time domain input and output signal of an analog sampled data system. In the ideal case, the sampled data system, samples the input signal instantaneously, or with an impulse function. The amplitude of each sample is equal to the instantaneous amplitude of the input signal. The output is a series of narrow pulses, each separated by time T, the sampling period.

2.1 Aperture

Since an impulse function in the time domain corresponds to a flat spectrum in the frequency domain, the input spectrum is exactly reproduced in the frequency domain, however, in reality the sampling signal is periodic and has a finite pulse width. When convoluting a finite pulse width with an input spectrum $F(j\omega)$ with unity amplitude, the result is found to be:

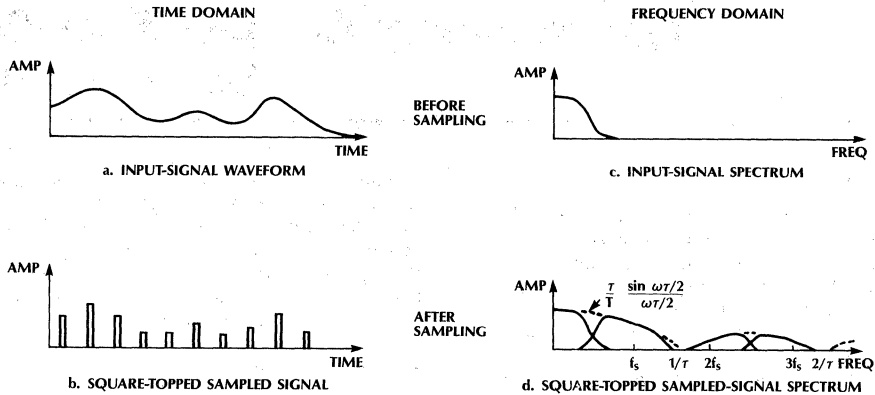
$$F_{st}(j\omega) = \frac{T}{T} \frac{\sin(\omega T/2)}{\omega T/2} \sum_{n=-\infty}^{\infty} F[j(\omega - n\omega_s)] \quad (1)$$

Figure 1: Signal Processing Systems



Application Note 4

Figure 2: Analysis of a Sampled Signal



From this equation, the gain is a continuous function of frequency defined by $(\tau/T) (\sin(\omega\tau/2)/(\omega\tau/2))$ where τ is the sample pulse width in seconds, T is the sample period in seconds, and ω the frequency in radians per second.

The time and frequency domain plots for the finite pulse width sampled signal are shown in figure 2. Figure 2 is a plot of the previous equations where the frequency spectrum is formed around multiples of the sampling frequency. As long as the adjacent spectra do not overlay (aliasing distortion), the continuous signal can be reconstructed from the discrete samples.

To evaluate the amplitude distortion caused by having a finite pulse width, one can simply solve equation 1. Since the ML2111 has a zero-order hold τ/T is unity. Assuming a 7.5 MHz sampling frequency and a bandwidth of 150 KHz, the amplitude distortion or attenuation is 5.7×10^{-3} dB.

The equation shows that when the sampling frequency is 40-50 times greater than the bandwidth, the aperture effects are negligible.

2.2 Aliasing

Another potential source for distortion in a sampled data system is aliasing. Aliasing distortion occurs when the input frequency to a sampled data system contains frequency components above one half the sampling frequency. These higher frequency components beat with the sampling frequency and are reflected back into the baseband causing aliasing distortion.

The additional spectral components caused by sampling the input signal are the sum and differences of the input frequencies with multiples of the sampling frequency. For example, assume the input to a sampled data system is a sine wave with a frequency of 100 KHz (f_i) sampled at 250 KHz (f_s), as shown in figure 3a. The first few spectral

components will be at: ($f_i = 100$ KHz; original signal, $f_s - f_i = 150$ KHz, $f_s + f_i = 350$ KHz, $2f_s - f_i = 400$ KHz, $2f_s + f_i = 600$ KHz, . . .). Now assume f_i has a second harmonic, which would be at 200 KHz, the spectral components are shown in figure 3b. If our bandwidth of interest were from DC to $f_s/2$, then the $f_s - 2f_i$ component interferes with the original signal. If we were to reconstruct the original signal by lowpass filtering it, we could not separate the aliased component, $f_s - 2f_i = 50$ KHz, from the original signal.

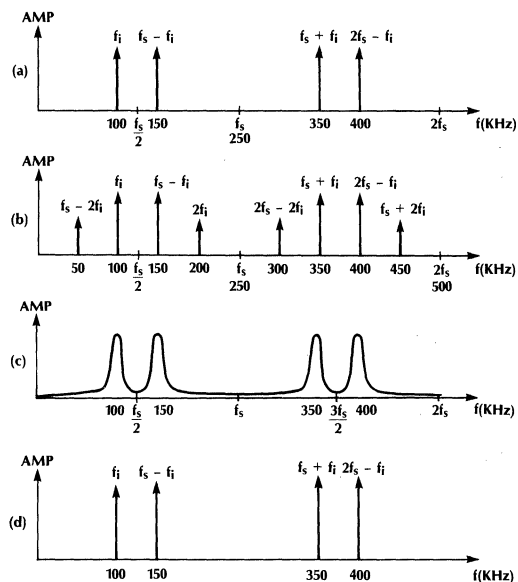
If our bandwidth of interest is a bandpass, the aliased component may not interfere. For example, if the ML2111 were to be used as a four pole bandpass filter with a center frequency at 100 KHz and a $Q = 10$ as shown in figure 3c, then the aliasing components in the above example would be filtered out as shown in figure 3d. But if the ML2111 were to be used as a low pass filter, then the $f_s - 2f_i$ aliased component would not be filtered out by the ML2111, and an anti-aliasing filter would be needed.

If the input signal is not band-limited, and the aliasing components fall within the bandwidth of interest, then a lowpass filter or anti-aliasing filter must be placed in front of the ML2111. This filter must be a continuous filter rather than a sampled data filter, however, the complexity of this filter is typically much less than the ML2111 filters, and its frequency response is less critical allowing for relaxed component tolerances.

Since no frequency component can be totally eliminated, one must determine the acceptable amplitude of the aliasing components that will *not* impact the Signal to Noise ratio of the system.

The higher the ratio of sampling frequency to input bandwidth, the lower the requirements on the anti-aliasing filter. Figure 4 shows the effects of sampling rate

Figure 3. Aliasing Distortion Using Sample Data Filters



rate on the separation of sampled signal spectra. Note the amount of overlap increases as the sampling frequency is decreased for a fixed input signal bandwidth. In general, the higher the sampling frequency, the less aliasing distortion. Since the ML2111's sampling frequency is typically either 50 or 100 times greater than the input bandwidth, the aliasing distortion may be negligible.

2.3 Signal to Noise Ratio and Aliasing Distortion

To determine whether aliasing distortion could be a problem, one must first determine the Signal to Noise Ratio of the overall system. Aliasing distortion less than the signal to noise ratio is of no concern.

The data sheet specifies noise based on Q and bandwidth. From these specs one can deduce the S/N ratio of one bi-quad in the ML2111. Using a simplified example, a bandpass filter with a Q = 10 and a system clock to center frequency ratio of 50:1 has noise that is 262 μ Vrms over a 750 KHz bandwidth; taken from the specs in the data sheet. To determine the maximum input signal amplitude, one must consider the slew rate spec. The typical value is 2 V/ μ sec, however a comfortable safety margin is 1.495 V/ μ sec for the commercial temperature range and 1.256 V/ μ sec for the military temperature range. The slew rate = $2\pi fA$, where f is the maximum input frequency, and A is the peak amplitude in volts. Therefore $A = 1.495E6 / (2 * \pi * 100E3) = 2.3$ Volts; and the S/N = 78 dB.

Based on a 100 KHz bandpass filter with a Q = 10, $f_{CLK}:f_0 = 50:1$, and a signal to noise ratio of 78 dB, what sort of anti-aliasing filter would be sufficient? One must first look at the spectrum of the input signal, particularly in the 4.895 MHz to 4.905 MHz frequency range since this is the range that will be reflected back into the bandwidth of interest, 95 KHz to 105 KHz. If the frequency components in the 4.895 MHz to 4.905 MHz are below 78 dB, they will have a minimum impact on the signal to noise ratio. Let's assume that these frequency components are down only 20 dB. Then the anti-aliasing filter will have to attenuate the frequencies in the 4.895 MHz to 4.095 MHz range by $78 - 20 = 58$ dB, and pass the frequencies in the 95 KHz to 105 KHz frequency range with no attenuation. A simple two pole Butterworth filter with a cutoff frequency of 170 KHz will be sufficient, however there will be an attenuation of about 0.5 dB at 100 KHz due to this filter.

Figure 5a shows a Sallen-Key active filter capable of implementing two poles, and figure 5b shows a Rauch filter also implementing two poles. These two active filters are good examples to use for anti-aliasing and reconstruction filters. Using the Rauch filter for the above example, $C_5 = 400$ pF, $C_8 = 90$ pF, and $R = R_4 = R_6 = R_7 = 5$ K Ω . Fortunately the cutoff frequency for the antialiasing and reconstruction filters are not critical since capacitors can vary 5% and resistors can vary 1%. Taking into account component tolerance for our example, the cutoff frequency can vary worst case from 152 KHz up to 178 KHz.

The important aspects to note are that one must first determine the signal to noise ratio in the bandwidth of interest. Based on this bandwidth, are there any frequencies that will be reflected back into the bandwidth of interest, and if so how much will they need to be attenuated? Remember that frequency components reflected back outside of the bandwidth of interest, will be filtered by the ML2111. Since the ratio of the sampling frequency to the center frequency is large on the ML2111, most designs will not need an anti-aliasing filter, and if they do, a simple two pole butterworth should suffice.

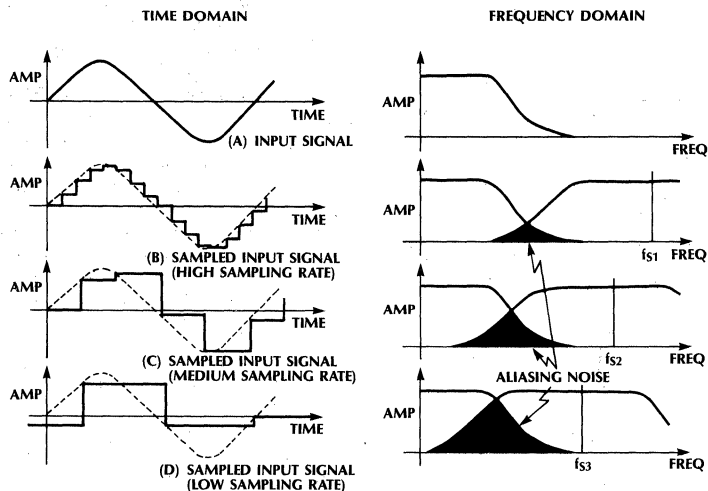
2.4 Signal Reconstruction

The output signal of a switched capacitor filter contains higher frequency components since it is a sampled signal. Many systems can accommodate these higher frequency components; however, if they interfere with the system's performance, then a signal reconstruction filter can be employed.

A time domain and frequency domain plot of the output from the ML2111 is given in figure 6. The output signal changes amplitude every clock period. These sharp transitions elicit high frequency components in the output signal. Once again, the fact that the ratio of the sampling frequency to the input bandwidth is high, reduces these distortion effects. As a result of the $\sin(x)/x$ envelope, the higher frequency components are attenuated. For example, assuming the input bandwidth is 100 KHz and

Application Note 4

Figure 4: Effects of Sampling Rate on Aliasing Noise



the sampling rate is 5 MHz, the frequencies around 4.9 MHz are down 34 dB, and they degrade towards zero as the frequency reaches 5 MHz. A single pole reconstruction filter with a cutoff frequency at 200 KHz would add an additional attenuation of 27 dB at 4.9 MHz but would attenuate the output by 1 dB at 100 KHz. A two pole Butterworth as in figure 5a or 5b would yield 58 dB of attenuation at 4.9 MHz and only 0.5 dB at 100 KHz.

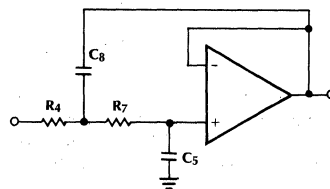
3.0 Layout Considerations

The layout of any board with analog and digital circuitry combined mandates careful consideration. The most important steps in designing a low noise system are:

1. All power source leads should have a bypass capacitor to ground on each printed circuit board (PCB). At least one electrolytic bypass capacitor (50 μ F or more) per board is recommended at the point where all power traces from the ML2111 join prior to interfacing with the edge connector pins assigned to the power leads.
2. Layout the traces such that analog signal and capacitor leads are far from the digital clock.
3. Both grounds and power supply leads must have low resistance and inductance. This should be accomplished by using a ground plane where ever possible. Either multiple or extra large plated through holes should be used when passing the ground connections through the PCB.

4. Use a separate trace for clock ground, and connect it to the edge connector board ground.
5. Use ground plane on both sides of PC board.
6. All power pins on ICs should have 0.1 μ F and a 0.01 μ F capacitors in parallel tied to ground, and as close to the power pins as possible.
7. Stray capacitance, lead lengths, and traces, on pin 4 and 17, the negative input of the op amp, should be kept to a minimum, particularly for high frequency filters which are more sensitive.

Figure 5a. Sallen-Key Filter

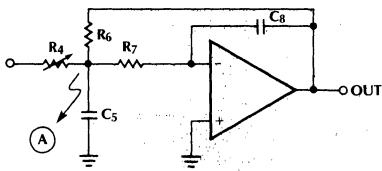


$$\frac{1}{s^2B + sC + 1}$$

$$B = R_4 R_7 C_5 C_8 = 1/\omega_0^2 \quad R_4 = R_7$$

$$C = C_5 (R_4 + R_7) = 1/Q\omega_0$$

Figure 5b. Rauch Filter



DC Gain: $\frac{R_6}{R_4} = H(0)$ [Minimize Parasitic C at Node (A)]

Transfer function: $\frac{H(0) (1/B)}{s^2 + s(C/B) + 1/B} = \frac{H(0) \omega_0^2}{s^2 + s(\omega_0/Q) + \omega_0^2}$

$$B = R_6 R_7 C_8 C_5$$

$$C = \frac{R_6 R_7 C_8 + C_8 (R_6 + R_7) R_4}{R_4}$$

Choose Butterworth response for example:

$$\omega_0 = 2\pi[170 \text{ KHz}] \approx 1.06 \times 10^6 \text{ rad/s}$$

$$Q = .707$$

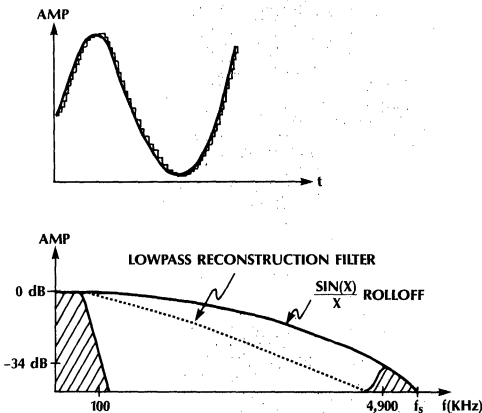
$$\text{say } R_4 = R_7 = R_6 = R \Rightarrow 1/\omega_0^2 = R^2 C_8 C_5$$

$$C = RC_8 + C_8(R_6 + R_7) = RC_8 + 2C_8R = 3RC_8$$

$$\Rightarrow \frac{C}{B} = \frac{3RC_8}{R^2 C_8 C_5} = \frac{3}{RC_5} = \frac{\omega_0}{Q}$$

$$\Rightarrow \begin{cases} C_5 = \frac{3Q}{R\omega_0} & \text{say } R = 5 \text{ K}\Omega \\ & \therefore C_5 = 400 \text{ pF} \\ & C_8 = 90 \text{ pF} \\ C_8 = \frac{1}{\omega_0^2 R^2 C_5} \end{cases}$$

Figure 6: Signal Reconstruction



3.1 Clocks and Output Loading

It is important to properly terminate the clock input to prevent overshoot. Each pin has protection diodes for Electro-Static Discharge (ESD), and any overshoot of more than 0.3 to 0.5 volts will be injected directly into the ML2111's ground and/or supplies. Matching the characteristic impedance of the line will prevent any ringing thus reduce clock noise.

When operating with high clock frequencies, the output of the op amp and integrators should be properly loaded. Ideally these outputs—LP, BP, and N—like to drive a total of 2 to 3 mA of peak current each. Assuming the output voltage swing is ± 2 volts, the sum of R_5 and R_6 , in mode 1c for example, should be 2 V/2 mA or about 1,000 ohms; assuming no other resistors are connected. Sometimes this is difficult to do if the ratios and loading cannot simultaneously be achieved. In this case an additional loading resistor placed as close as possible to the output pin will serve the purpose of properly loading the outputs.

4.0 Sweeping Filters

One particularly nice feature of sampled data filters is the fact that the center frequency of a filter is directly related to the clock frequency. For a lowpass filter, increasing the clock frequency increases the cutoff frequency. Even though the center frequency increases proportionally with the clock, Q stays constant. Therefore in a bandpass filter, increasing the clock frequency increases the center frequency as well as the bandwidth. Table 2 in the data sheet illustrates this relationship. (Note that there is some Q deviation as the system clock goes beyond a certain value. Refer to figure 2E in the data sheet for a graph of this phenomenon)

A good rule of thumb for the maximum rate a filter can be swept is that the Sweep Rate should be less than the square of the bandwidth of the filter. This will reduce attenuation of the passband as a result of sweeping the filter. The theoretical derivation of this approximation is as follows.

Assume we have a bandpass filter with an in-band signal that starts at $t = 0$. The output of the filter will exponentially increase until it reaches the steady state gain of the passband. After 4 time constants (τ), the output sine wave will be at 98% of its final amplitude.

Sweeping a filter is analogous to keeping the filter constant and sweeping the input frequency. To prevent the filter from attenuating the sweeping input signal by more than 2% or 0.16 dB:

$$\text{Sweep Rate} < BW/4\tau \quad (2)$$

but the time constant can be approximated by:

$$\tau \approx Q/2\pi f_0 \quad (3)$$

and,

$$Q = f_0/BW \text{ or } BW = f_0/Q \quad (4)$$

substituting τ and BW into equation (2) results in:

$$\text{Sweep Rate} < \pi BW^2/2 \quad (5)$$

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5.0 High Frequency Operation

There are three basic modes for the ML2111 — mode 1, 2 and 3. Within each mode there are several variations as shown in the table below.

Mode 1*	High Frequency Mode
1, 1a, 1d 1b, 1c	f_0 up to 150 KHz; Q up to about 20** f_0 up to 100 KHz; Q up to about 30
Mode 2	Flexible for Notches
2, 2a, 2b	f_0 up to 30 KHz; Q up to about 30
Mode 3	Most Flexible/Low Component Count
3, 3a	f_0 up to 30 KHz; Q up to about 30

* Q and f_0 have an inverse relationship. This table is only an approximation. Actual performance depends on board layout and stray capacitance.

** 15% or less Q deviation. Higher Q's can be realized with greater deviation.

Mode 1 is the only mode which has the input amplifier outside the resonant loop. This is important because the input amplifier reduces the bandwidth potential of the filter. *Only Mode 1 can achieve filters with resonant frequencies up to 150 KHz.*

Inserting an ML2111 into an MF10, LMF100, or LTC1060 socket and increasing the clock frequency does not automatically increase the bandwidth potential up to 150 KHz. If these pin-compatible parts were designed using Mode 1, the bandwidth improvements would be realized; however if they were used in another mode, there would be limited bandwidth improvements.

Complex, high order filters usually have pole pairs with different center frequencies; Elliptical and Chebyshev filters are two examples. To realize two pole pairs in one ML2111 with different center frequencies, one must either use two different clocks, or use a mode which allows the center frequency to be modified by external resistors.

Using different clock frequencies to realize poles with different center frequencies is not recommended. Besides the additional expense of providing more than one clock, the two system clocks may beat with each other and possibly result in side tones that falls within the passband of the filter. Additionally if anti-aliasing is needed, separate anti-aliasing filters would be needed for each stage.

Looking at tables 1 and 2 in the ML2111 data sheet, one can see the modes that allow the center frequency to be modified by external resistors. These modes each have an additional coefficient multiplied by $f_{CLK}/100(50)$. From the block diagrams one can see that the modes which allow the center frequency to be modified, feedback the LP output using a resistor divider. The modes that restrict the ratio to 50 or 100 have a unity gain LP feedback.

If the coefficient multiplied by $f_{CLK}/100(50)$ is greater than or equal to 1, as in Mode 1b, then the ratio of f_{CLK} to f_0 can be less than 50 or 100. Whereas if this coefficient is less than or equal to 1, then the ratio of $f_{CLK}:f_0$ can be greater than or equal to 50 or 100. Reducing the ratio of f_{CLK} to f_0 to less than 40 to 50 is not recommended. As the ratio of the sampling frequency to the center frequency is reduced, the approximation of a sample data filter to a continuous filter is reduced. Aperture effects increase, aliasing effects may increase, harmonics in the output increase, and the warpage between the discrete and the continuous filter increase. 40 to 50:1 is the minimum recommended ratio of f_{CLK} to f_0 .

Based on the above arguments one might conclude that 100:1 is better than 50:1. In general this is true for switched capacitor filters, but not for the ML2111. The specifications in the data sheet show that a 50:1 ratio provides a more accurate Q than a 100:1, and a 50:1 ratio allows higher frequency filters.

Mode 3 is the most flexible since the center frequency can be greater than or less than $f_{CLK}/100(50)$ by selecting R_2 and R_4 . Its also the most efficient since it has the lowest component count. However mode 3 can only work up to 30 to 40 KHz or Qs up to the 10 to 30 range; higher f_0 can be obtained with lower Qs. Sometimes a small capacitor (C_4) across R_4 can compensate the filter response and offer less Q deviation. The value should be selected by setting C_4 equal to $1/2\pi R_4 BW$ where BW is approximately equal to 2 to 4 MHz.

Another reason mode 3 can only be used at lower frequencies is that there is a true sample and hold at the positive input of the summer. This sample and hold adds a 72 degree delay at the center frequency when using a 50:1 ratio ($360^\circ/50$). By using a higher ratio this delay is lowered. Since the ML2111 allows a higher system clock than other competing devices, this delay can usually be made smaller for similar center frequencies.

In conclusion, for high frequency filters use Mode 1. For complex filters with various center frequencies use Mode 1c. In most cases one should choose 50:1 over 100:1 ratio for more accurate Q's and center frequencies.

5.1 A Flexible Building Block

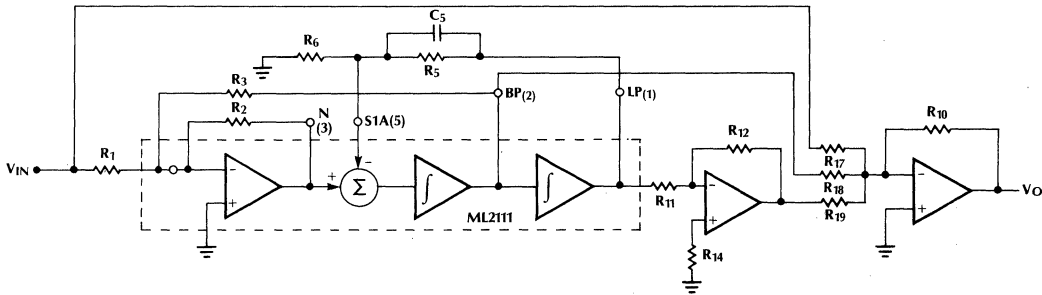
Figure 7 shows the block diagram of a second order section which includes both a complex pole pair and a complex zero pair. The poles are provided by the ML2111 and the zeros realized by one and sometimes two external op amps. This building block uses mode 1c which allows the poles to have a center frequency based on external resistors as well as the clock, plus it can be used in higher frequency filters since the op amp is outside of the resonant loop. The same feedforward circuit can be used on other modes as well, but for high

frequency filters, where each complex pole pair has a different center frequency, mode 1c is the best choice. As mentioned before, only when Butterworth filters are

desired, use mode 1 to achieve higher frequencies and a higher dynamic range. The transfer function for the flexible building block is given below.

$$\frac{V_O}{V_{IN}} = \left(-\frac{R_{10}}{R_{17}} \right) \left[\frac{s^2 + \frac{R_2}{R_3} \left[1 - \frac{R_{17} R_3}{R_{18} R_1} \right] s\omega_1 + \left[1 + \frac{R_2 R_{12} R_{17}}{R_1 R_{11} R_{19}} \left(1 + \frac{R_5}{R_6} \right) \right] \frac{R_6}{R_5 + R_6} \omega_1^2}{s^2 + \frac{R_2}{R_3} s\omega_1 + \frac{R_6}{R_5 + R_6} \omega_1^2} \right] \quad (6)$$

Figure 7: Flexible Building Block



At least one and sometimes two external op amps are required to realize the zeros. The first op amp serves as an inverter, while the second one sums the input signal with the lowpass and bandpass outputs. A fast op amp should usually be used with greater than 10 MHz bandwidth to minimize signal phase shifts. Depending on the application, sometimes a slower amplifier will suffice. In some cases no external op amp is necessary and the second op amp in the ML2111 if not being used will suffice. This was done in figure 34 in the data sheet.

A handy set of equations to convert pole and zero locations given in rectangular coordinates to f_0 and Q values is as follows:

$$\text{Complex Pole} = \sigma + j\omega; \quad f_0 = \frac{\sqrt{\sigma^2 + \omega^2}}{2\pi} \quad Q = \frac{1}{2} \sqrt{1 + (\omega/\sigma)^2} \quad (7)$$

By cascading several of these building blocks, complex high frequency Elliptical filters can be realized.

With the Flexible Building Block a lowpass, highpass, notch, and allpass section can be realized by properly positioning the zero locations. Zero locations are chosen by selecting the appropriate resistors. The difference between the lowpass output provided by the ML2111 in mode 1c and the lowpass function realized by the flexible building block is that in mode 1c the response is monotonically decreasing, while the Flexible Building Block has a complex zero pair which inserts a ripple in the stop band and flattens out at high frequency.

5.2 Lowpass

For a lowpass design with a notch, the zeros should be placed on the $j\omega$ axis at frequencies greater than the poles' center frequency. In the numerator of the transfer function for equation 6, the coefficient for $s\omega_1$ should be set to zero; setting $\frac{R_3}{R_1} = \frac{R_{18}}{R_{17}}$.

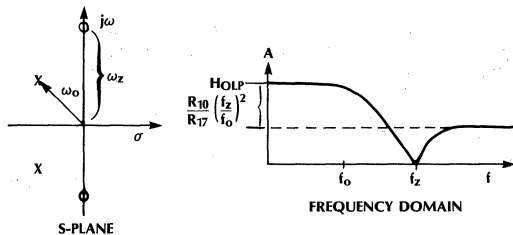
Since $\frac{R_6}{R_5 + R_6} \omega_1^2 = \omega_0^2$, the coefficient

$1 + \frac{R_2 R_{12} R_{17}}{R_1 R_{11} R_{19}} \left(1 + \frac{R_5}{R_6} \right)$ determines the center frequency of the zero. In this form it is always greater than one, therefore the center frequency of the zero is always greater than the center frequency for the poles; hence a lowpass filter. The pole/zero location and the frequency response are shown below.

$$f_0 = \frac{f_{CLK}}{100(50)} \sqrt{\frac{R_6}{R_5 + R_6}}$$

$$Q = \frac{R_3}{R_2} \sqrt{\frac{R_6}{R_5 + R_6}}$$

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Equations for the lowpass configuration:

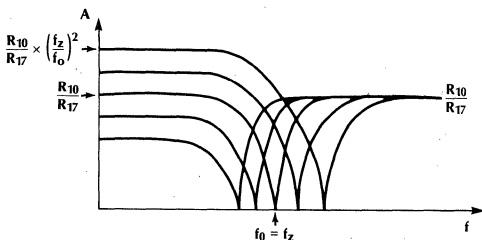
$$\frac{R_{17}}{R_{19}} = \frac{\left(\frac{f_z}{f_0}\right)^2 - 1}{\frac{R_2}{R_1} \left(1 + \frac{R_5}{R_6}\right)} \quad f_z = f_0 \sqrt{1 + \frac{R_2 R_{12} R_{17}}{R_1 R_{11} R_{19}} \left(1 + \frac{R_5}{R_6}\right)}$$

$$\text{DC Gain} = H_{\text{OLP}} = \left(-\frac{R_{10}}{R_{17}}\right) \left(1 + \frac{R_2 R_{12} R_{17}}{R_1 R_{11} R_{19}} \left(1 + \frac{R_5}{R_6}\right)\right)$$

The ratio of the zero to the pole frequency determines the DC to high frequency attenuation.

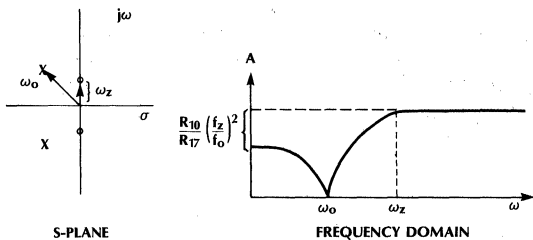
When the zeros are at the same frequency as the poles the bi-quad becomes a notch, and there is no difference between the high frequency and low frequency gain. The larger the difference between the pole and zero frequencies, the greater the rejection. Figure 8 illustrates the relationship between pole/zero location and gain.

Figure 8: Varying f_z and Keeping f_0 and Q Constant



5.3 Highpass

For a highpass filter the zeros must be less than the center frequency for the poles. The pole/zero plot and the frequency plot are shown below.



To place the zeros at a lower frequency than the poles the coefficient $1 + \frac{R_2 R_{12} R_{17}}{R_1 R_{11} R_{19}} \left(1 + \frac{R_5}{R_6}\right)$ must be less than one. This can be done by removing the inverter in figure 7, which makes the sign of R_{19} negative. To place the zeros on the $j\omega$ axis, once again $\frac{R_3}{R_1} = \frac{R_{18}}{R_{17}}$. Equations for the highpass configuration:

$$f_z = f_0 \sqrt{1 - \frac{R_2 R_{12} R_{17}}{R_1 R_{11} R_{19}} \left(1 + \frac{R_5}{R_6}\right)}$$

$$H_{\text{OHP}} = -\frac{R_{10}}{R_{17}}$$

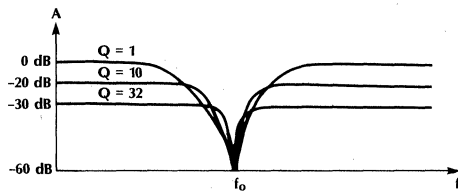
5.4 Notch

Even though mode 1c provides a notch output, the notch realized by the flexible building block achieves 0 dB of gain at DC and at high frequencies regardless of the Q value. The problem with the notch in mode 1c is that

$$H_{\text{ON1}}(f \rightarrow 0) = H_{\text{ON2}}(f \rightarrow f_{\text{CLK}}/2) = \sqrt{\frac{R_6}{R_5 + R_6}} \cdot \frac{1}{Q}$$

As Q increases $H_{\text{ON1,2}}$ must decrease otherwise the bandpass output node, BP pin 2 or 19, will saturate. The restriction is that $H_{\text{OHP}} = 1 = -R_3/R_1$. Let's take a simple case when $R_5 = 0$, then $H_{\text{ON1}} = H_{\text{ON2}} = 1/Q$. The plot below shows the notch for different Q 's in mode 1c.

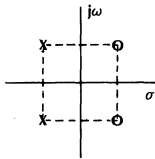
Figure 9: Mode 1c Notch when $R_5 = 0$



To realize the notch using the Flexible Building Block the zeros must be placed on the $j\omega$ axis at the same resonant frequency as the poles. Therefore from equation 6, $\frac{R_3}{R_1} = \frac{R_{18}}{R_{17}}$ and $R_{19} \rightarrow \infty$. Setting R_{19} equal to infinity means removing it from the circuit; which saves an op amp and a few resistors. H_{OHP} still must equal one; however the gain at DC and $f_{\text{CLK}}/2$ is independent of Q ; $H_{\text{ON1}}(f \rightarrow 0) = H_{\text{ON2}}(f \rightarrow f_{\text{CLK}}/2) = -R_{10}/R_{17}$. Tuning R_{18} adjusts the depth of the notch. See figure 34 in the data sheet for an example.

5.5 Allpass Equalizer

An allpass filter is used to linearize the filter's phase response. A linear phase response results in a constant group delay. An allpass filter keeps the gain constant and just shifts the phase. To keep the gain constant and only shift the phase, the poles and zeros must be equal but on opposite sides of the s-plane as shown below.



S-plane representation of 2nd order Allpass Filter

The Flexible Building Block can function as an allpass when $R_{19} \rightarrow \infty$ and $\frac{R_{17}R_3}{R_{18}R_1} = 2$. The Transfer function for the allpass is:

$$\frac{V_O}{V_{IN}} = \left(-\frac{R_{10}}{R_{17}} \right) \frac{s^2 - \frac{R_2}{R_3} \omega_1 s + \frac{R_6}{R_5 + R_6} \omega_1^2}{s^2 + \frac{R_2}{R_3} \omega_1 s + \frac{R_6}{R_5 + R_6} \omega_1^2} \quad (8)$$

5.6 Frequency Compensation

In some cases it is possible to improve the Q accuracy and minimize Q deviation by adding a capacitor (C_5) in parallel with R_5 in figure 7. This capacitor serves as compensation for a pole at around 2.4 MHz in the output of LP. The zero location should be placed at around 2.4 MHz, where the internal pole is. Unfortunately C_5 adds a pole as well as a zero to this branch. If this pole is too close to the zero, the benefit of C_5 is diminished. The zero location is: $f_Z = 1/2\pi R_5 C_5$ and the pole location is: $f_P = 1/2\pi(R_5 \parallel R_6)C_5$, ($R_5 \parallel R_6$ is the parallel equivalent resistance). The larger the ratio of the pole frequency to the zero frequency, the better this capacitor will serve.

The highest center frequency attained is when R_5 equals zero. (Note: Practically speaking R_5 should never be zero, to allow fine tuning of f_0 .) Unfortunately C_5 cannot properly compensate the 2.4 MHz internal pole with a negligible value for R_5 . To overcome this problem, compensation can be achieved at high frequencies using an op amp in the LP feedback branch as shown in figure 10.

The center frequency in mode 1c is calculated by the following equation:

$$\frac{f_{CLK}}{f_0} = \frac{50}{\sqrt{k}} \text{ where } k = \text{Transfer function}$$

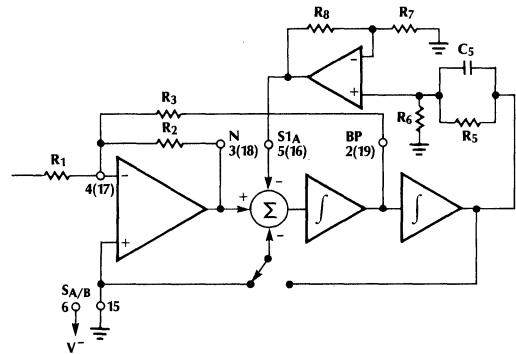
With a passive feedback loop using R_5 and R_6 , $k = \frac{R_6}{R_5 + R_6}$.

However when using the op amp configuration as in

figure 10, $k = \left(\frac{R_6}{R_5 + R_6} \right) \left(1 + \frac{R_8}{R_7} \right)$. When $k = 1$ the ratio is 50.

Using active feedback in mode 1c has the unique advantage of allowing the ratio of clock to center frequency to be less than 50 by setting k greater than 1. It is not recommended to use ratios less than 40-50, however this feature does allow more freedom in tuning the center frequency of the pole above or below the ratio of 50. If the circuit uses a crystal for f_{CLK} , and the pole needs to be tuned, R_8 could be a potentiometer to allow tuning of the pole. For this compensation to work $\frac{R_8}{R_7}$ should be 4-9 to provide phase lead before phase lag.

Figure 10: Compensation Using Active Feedback for High Frequency Poles



$C_5 = 33\text{-}66\text{pF}$ (Depends on board's parasitics)
 $R_8 = 1800\Omega$; $R_7 = 200$
 $R_6 = 100\Omega$, $R_5 = 900\Omega$

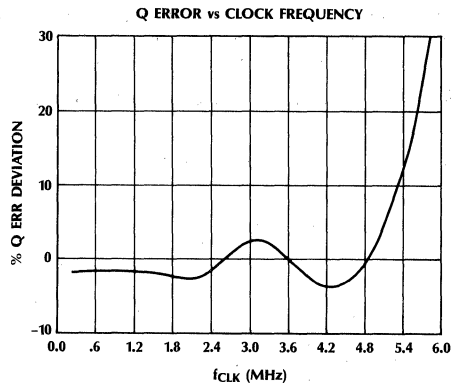
Using mode 1 instead of mode 1c as configured in figure 7, is a better solution for high frequency poles; however there are certain cases where mode 1 cannot be used. For example, if one of the two bi-quads in the ML2111 is already used in mode 1c, then the other one must also operate in mode 1c. It would be less expensive to add an op amp to the second bi-quad of an existing ML2111 than to add an additional ML2111 just to use one bi-quad operating in mode 1.

Figure 11a shows the Q accuracy vs. clock frequency in mode 1c using passive feedback for a Q approximately equal to 10. Q inaccuracy dramatically increases just beyond 100 KHz center frequency. Figure 11b shows Q accuracy vs. center frequency in mode 1c using active feedback with a DC transfer function of 1. The op amp used for this measurement was an AD5539, where

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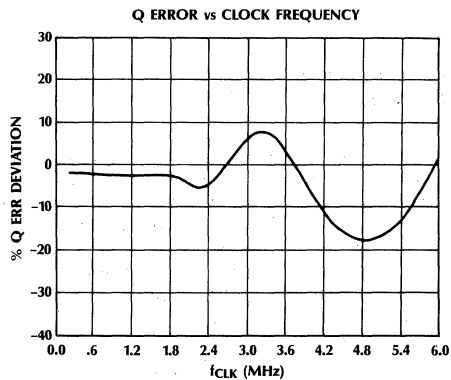
$\frac{R_8}{R_7} = \frac{R_5}{R_6} = 9$. This op amp is a good choice because it has a wide bandwidth, 220 MHz, and is low cost. The figure shows that Q deviation does not dramatically increase until well beyond 120 KHz; therefore for higher frequency operation and high Q, the use of mode 1c with active feedback is recommended.

Figure 11a: Mode 1c with Passive Feedback



$$\frac{R_3}{R_2} = 10; R_5 = 0; R_1 = R_3 = 20K$$

Figure 11b: Mode 1c with Active Feedback



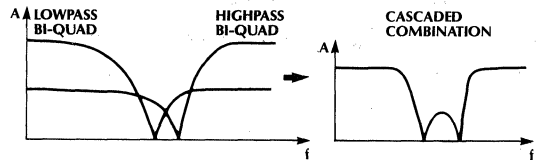
$$C_5 = 33pF; \frac{R_3}{R_2} = 10; R_5 = 0; R_1 = R_3 = 20K$$

6.0 Design Methodology for Complex Filters

The previous section described how to use the Flexible Building Block to implement lowpass, highpass, notch, and allpass second order sections. Higher order filters are achieved by cascading these second order sections. For example an Elliptical notch is accomplished by cascading lowpass and highpass sections as shown in figure 12.

An Elliptical bandpass is also a combination of highpass and lowpass sections, except for a bandpass filter the cutoff frequency for the highpass bi-quads are lower than the cutoff frequency for the lowpass.

Figure 12: Fourth Order Elliptic Notch



Once the pole and zero location have been determined for the filter desired, the next step is to choose the proper mode of operation and translate the center frequency and Q values for each pole and zero into resistor values. If the pole and zero locations are given in real and imaginary values, they can be converted to f_0 and Q by using equation 7.

For center frequencies between 0 and 20 KHz, either mode 3 or mode 1c can be used. Sometimes mode 3 or mode 3a will result in a lower component count. However mode 3 should be used with caution since high Qs and high parasitic capacitance on pin 4 and 17 can lead to oscillations. This can usually be compensated by using a capacitor across R4, which provides some phase lead, and low value resistors such as 1-2 Kohms.

For center frequencies between 20 to 100 KHz, where each pole has a different center frequency, mode 1c should be used. This range can be extended up to 120 KHz with active compensation in the LP feedback path as shown in figure 11b. The combination of high Qs (20 to 30), and high frequencies (above 80 to 100 KHz), and parasitic capacitance across R6, can lead to oscillations. This can be dealt with by placing a capacitor C5 across R5, or by using active compensation. Additionally the signal swing should be limited to about 1 to 1.4 volts peak-to-peak.

For filters where f_0 is the same for all pole locations, such as Butterworth, Lowpass or Highpass. High order filters with cutoff frequencies up to 150 KHz can be realized using mode 1. In this case the signal level can be increased to 2.82 volts peak-to-peak.

7.0 Design Example

The following is an example an eighth order Elliptic bandpass filter with a center frequency of 90 KHz and a bandwidth of 19 KHz. This filter was designed built and tested on its own printed circuit board. A print of the masks for the PCB, and a photograph of the performance of the filter is included at the end of this section.

In general, high Q filters (Elliptic and Chebyshev) will have higher sensitivity to component and temperature variations and higher noise than low Q filters such as Butterworth and Bessel.

- a) 8th Order Elliptic Bandpass with the following Filter characteristics:

Amax: 0.5 dB (peak to peak passband ripple)

Amin > 50 dB (stopband attenuation)

(f_1, f_2) Passband: 81,000 to 100,000 Hz
(geometrically symmetric) $\Rightarrow f_c^2 = f_1 \times f_2$

(f_c) Center: 90,000 Hz

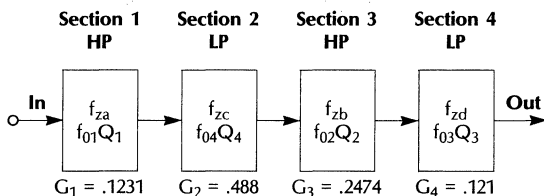
Stopband: 70.5 KHz to 115 KHz

- b) Obtain:

$$\text{Poles} \begin{cases} f_{01} = 80839 \text{ Hz} & Q_1 = 30.2 \\ f_{02} = 85820 \text{ Hz} & Q_2 = 10.86 \\ f_{03} = 94383 \text{ Hz} & Q_3 = 10.86 \\ f_{04} = 100200 \text{ Hz} & Q_4 = 30.2 \end{cases}$$

$$\text{Zeros} \begin{cases} f_{za} = 69185 \text{ Hz} \\ f_{zb} = 50082 \text{ Hz} \\ f_{zc} = 117080 \text{ Hz} \\ f_{zd} = 161733 \text{ Hz} \end{cases}$$

- c) After considering a few pole-zero pairing combinations the following (not necessarily optimum) combination was adopted.



Note: G_i are the high frequency gains. ($= R_{10}/R_{17}$)

Because of difficulty in solving equations first order equation were calculated and final values found by using potentiometer.

- d) Choose $f_{CLK} =$ highest $f_0 = 100200$
@ 50:1 $\Rightarrow 50f_0 = 5.01$ MHz. Choose $\sim 10\%$ higher $f_{CLK} = 5.5$ MHz. It's better to choose a slightly larger f_{CLK} to be able to adjust R_5 .

- e) Design Procedure.

Section 1. *) Want a ratio $= \frac{5.5 \text{ MHz}}{80839 \text{ Hz}} \cong 68 = R_1$

in Mode 1c $R_1 = 50 \times \sqrt{1 + \frac{R_5}{R_6}} \Rightarrow 1 + \frac{R_5}{R_6} = 1.8516$

Assume $R_5 + R_6 = 1000 \Omega$

then $1 + \frac{R_5}{R_6} = \frac{1000}{R_6} = 1.8516$

$\Rightarrow R_6 \cong 540 \Omega$ (fixed R)

$\Rightarrow [R_5 = R_6 (1.8516 - 1) \cong 460 \Omega]$ [1000 Ω trim pot]

- *) Want $Q = 30.2$ use following approximation:

$$\frac{R_3}{R_2} \approx \frac{Q \sqrt{1 + \frac{R_5}{R_6}}}{1 + Q(f_0/f_x)} \quad ; \text{ where } f_x = 2.4 \text{ MHz} \quad (\text{internal pole})$$

$\cong 20.4 \Rightarrow$ assume $R_2 = 2000 \Omega$

and $R_3 \approx 40.7 \text{ K}\Omega$ (100 K trim pot)

and initially assume $R_1 = R_3 = 40.7 \text{ K}\Omega$

- *) Zero. Use the following approximation.

$f_{za} = 69185 \text{ Hz}$

$$\frac{R_{17}}{R_{19}} \cong \frac{1 - (f_z/f_0)^2}{\frac{R_2}{R_1} \left(1 + \frac{R_5}{R_6}\right)} = \frac{1 - (69185/80839)^2}{1/20 \times 1.8516} = 2.89$$

Since R_{19} loads the LP output then assume $R_{19} > 50000 \Omega$. Also since later we will fine tune the gains this relationship will slightly change. Thus, initially assume a higher R_{17} which can be change later if needed.

Choose $R_{17} \cong 30 \text{ K}\Omega$ (fixed R)

and $R_{19} \approx 10 \text{ K}\Omega$ (20 K Ω trim pot)

choose $R_{18} = R_{17}$ (initially)

and $R_{10} = R_{17} \times G_1 = 30\text{K} \times .123 = 3690 \Omega$ (fixed R)

(Note: This is a first order approximation that underestimates the value of Q, whose final value will be tuned in later in the breadboard stage.)

- f) 1. By looking at the bandpass output adjust R_5 until the peak frequency is f_0 , in this case 80839 Hz
2. Then adjust R_3 until $Q = 30.2$
3. Then change R_1 until the peak of the bandpass or lowpass output (larger of the two) is about 0 dB. R_1 does not need to be a trim pot.
4. Now by looking at the output of the section adjust R_{19} to place the zero at the correct frequency (in this case 69185 Hz)

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5. Adjust R_{18} to obtain a deeper notch. Sometimes R_{18} is not needed at all and can be removed from the circuit.
6. Check the high frequency gain so that it is $G_1 = .1231$
7. Design the rest of the sections the same way
8. Keep R_{10} of first section as a trim pot to slightly trim gain of the whole filter (if important in the application).

For this design these are the final values:

Section 1.

$R_1 = 94.5 \text{ K}\Omega$	$R_{10} = 3.83 \text{ K}\Omega$
$R_2 = 2 \text{ K}\Omega$	$R_{17} = 32.4 \text{ K}\Omega$
$R_3 = 66.2 \text{ K}\Omega$ (100 K pot)	$R_{18} = 15 \text{ K}\Omega$
$R_5 = 452 \Omega$ (1 K Ω pot)	$R_{19} = 4.65 \text{ K}\Omega$
$R_6 = 540 \Omega$	$H_{LP \text{ PEAK}} = 1.1512$ (+1.22 dB)
	$H_{BP \text{ PEAK}} = .846$ (-1.45 dB)

Section 2.

$R_1 = 65 \text{ K}\Omega$	$R_{10} = 14.34 \text{ K}\Omega$
$R_2 = 2 \text{ K}\Omega$	$R_{17} = 28.7 \text{ K}\Omega$
$R_3 = 47.4 \text{ K}\Omega$ (100 K pot)	$R_{18} = \infty$
$R_5 = 170 \Omega$ (500 Ω pot)	$R_{19} = 2.9 \text{ K}\Omega$
$R_6 = 830 \Omega$	$H_{LP \text{ PEAK}} = 1.12$ (+.984 dB)
	$H_{BP \text{ PEAK}} = .972$ (-.247 dB)

Section 3.

$R_1 = 31.5 \text{ K}\Omega$	$R_{10} = 9.05 \text{ K}\Omega$
$R_2 = 2 \text{ K}\Omega$	$R_{17} = 40 \text{ K}\Omega$
$R_3 = 23.3 \text{ K}\Omega$ (50 K pot)	$R_{18} = 16.86 \text{ K}\Omega$
$R_5 = 389 \Omega$ (1 K Ω pot)	$R_{19} = 6.35 \text{ K}\Omega$
$R_6 = 600 \Omega$	$H_{LP \text{ PEAK}} = 1.074$ (+.62 dB)
	$H_{BP \text{ PEAK}} = .834$ (-1.58 dB)

Section 4.

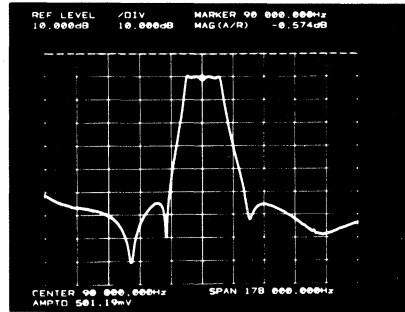
$R_1 = 25 \text{ K}\Omega$	$R_{10} = 12 \text{ K}\Omega$
$R_2 = 2000 \Omega$	$R_{17} = 99.97 \text{ K}\Omega$
$R_3 = 21.4 \text{ K}\Omega$ (50 K pot)	$R_{18} = \infty$
$R_5 = 279 \Omega$ (500 Ω pot)	$R_{19} = 6 \text{ K}\Omega$
$R_6 = 732 \Omega$	$H_{LP \text{ PEAK}} = 1.12$ (+.924 dB)
	$H_{BP \text{ PEAK}} = .953$ (-.418 dB)

Note: All R's are 1% metal film 1/4W
Trim pots are 25 turns. 1/2W

When placing resistors in and out of the ML2111 filter circuit, specifically R_3 , the filter will oscillate at f_0 due to the Q going to infinity. Also when designing high frequency high Q filters, such as $f_0 = 100 \text{ KHz}$ and $Q = 30$ like pole #4, high voltage swings may cause nonlinear operation provoking oscillations. Changing f_{CLK} momentarily to a much lower value will restore the filter to a stable linear operation. Thus it is important for high frequency, high Q filters to limit the input signal swing to about 500-700 mV peak.

7.1 Performance Measurements, Schematics and PCB Layout

Figure 13: Frequency Response of Eighth Order Elliptic Filter.



The center frequency is at 90 KHz with the lower cutoff at 81 KHz and the upper cutoff at 100 KHz. The stopband is down -55 dB at 70.5 KHz and 115 KHz.

Figure 14: Passband of Filter Showing 0.5 dB Ripple.

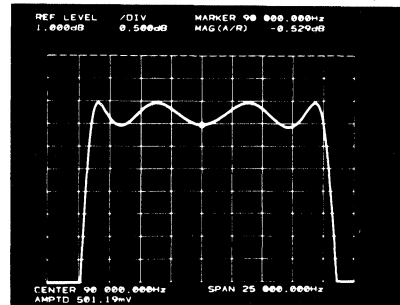
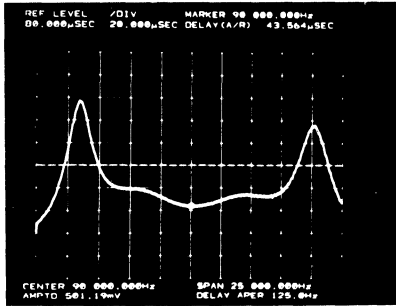
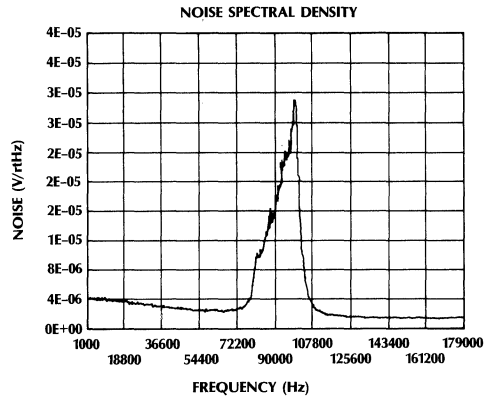


Figure 15: Group Delay.



A constant group delay can be achieved by adding allpass equalizer sections to this filter.

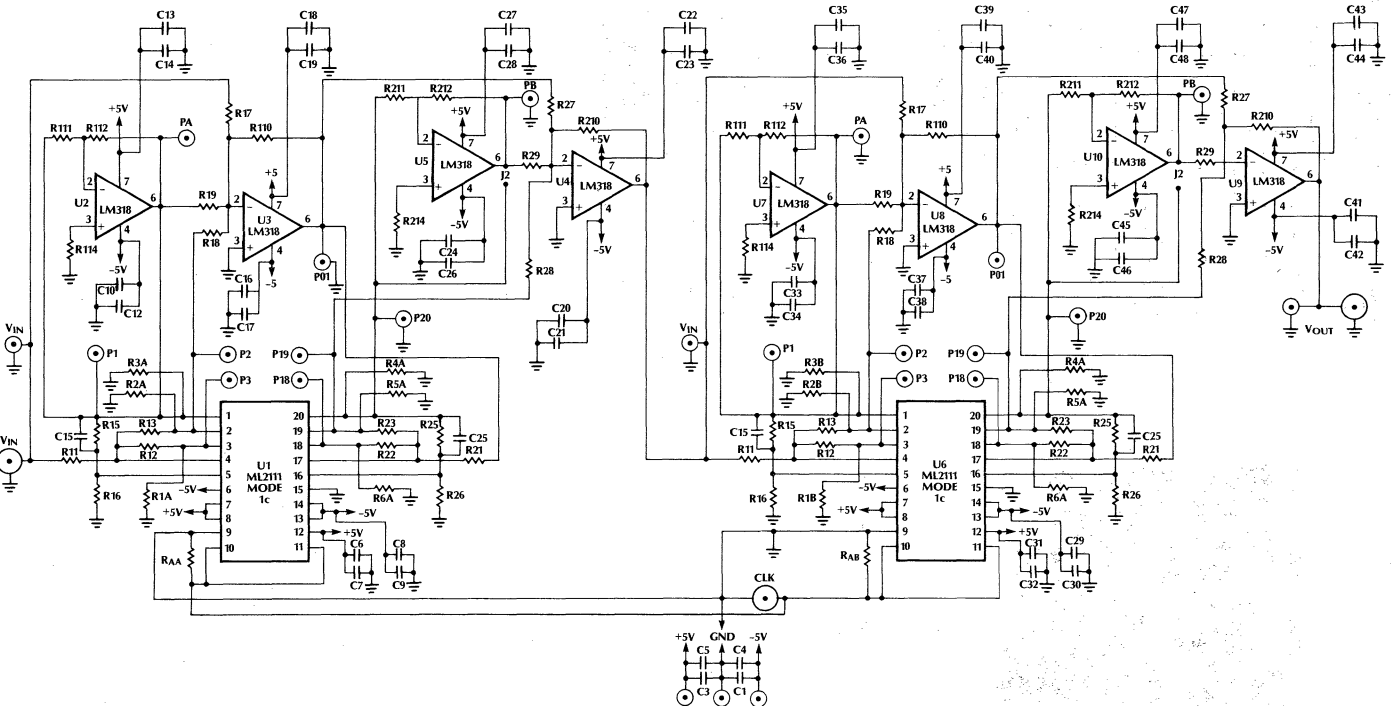
Figure 16: Power Spectral Density of the Noise



This plot shows that the pole/zero pairing and order of the bi-quad sections chosen was not optimum as far as noise is concerned. The plot shows that the upper band edge of noise is higher than the lower band edge. A different combination of pole/zero pairing and order pairing would have yielded a flatter noise response and possibly a lower noise value; which would have then improved the S/N ratio. The current design yields S/N of about 40 dB assuming a noise bandwidth from 1 KHz to 179 KHz. Input voltage = 353 mV_{rms}, output noise voltage = 3.14 mV_{rms}.

Application Note 4

Figure 17: Schematic Diagram of Printed Circuit Board



(Note that U2 and U7 are not used in this design. This is a general purpose development board and not a minimal component count design).

Figure 18a: PCB Layout Component Side

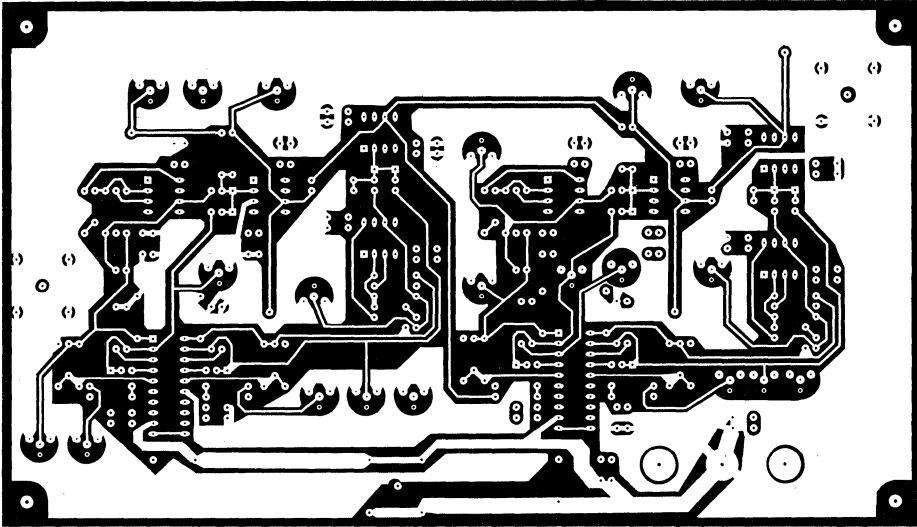
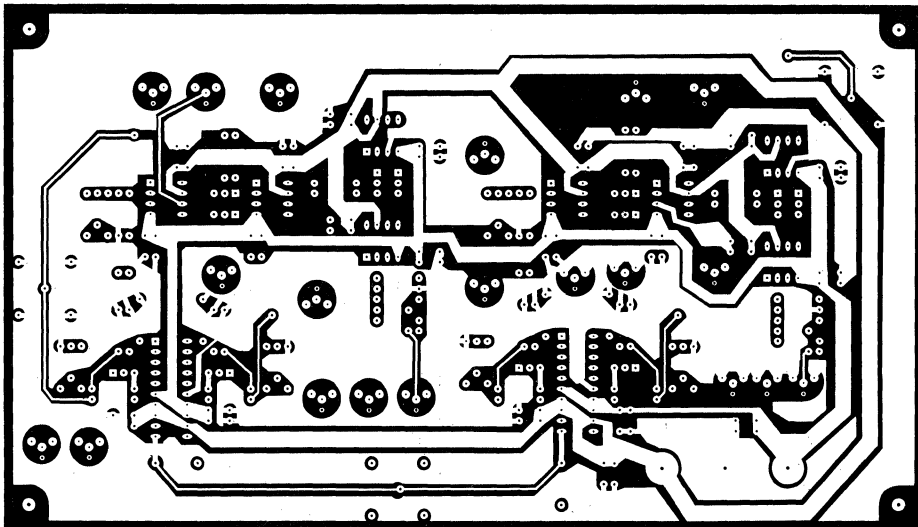


Figure 18b: PCB Layout Solder Side



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Application Note 4

ML2111 Application Board Parts List

Part #	Value	Note
Resistors		
R1A	1 K Ω	
R2A	500 Ω	
R3A	1 K Ω	
R4A	1 K Ω	
R5A	500 Ω	
R6A	1 K Ω	
R11A	94.5 K Ω	
R12A	2 K Ω	
R13A	66.2 K Ω	100 K Ω Pot
R15A	452 Ω	1 K Ω Pot **
R16A	540 Ω	
R17A	32.4 K Ω	
R18A	15 K Ω	
R19A	4.65 K Ω	10 K Ω Pot*
R110A	3.83 K Ω	10 K Ω Pot*
R111A	OPEN	
R112A	OPEN	
R114A	OPEN	
RAA	100 Ω	
RAB	100 Ω	
R210A	14.3 K Ω	
R211A	5 K Ω	
R212A	5 K Ω	
R214A	2.5 K Ω	
R21A	65 K Ω	
R22A	2 K Ω	
R23A	47.4 K Ω	100 K Ω Pot
R25A	170 Ω	500 Ω Pot**
R26A	830 Ω	
R27A	28.7 K Ω	
R28A	OPEN	
R29A	2.9 K Ω	10 K Ω Pot*
R1B	1 K Ω	
R2B	500 Ω	
R3B	1 K Ω	
R4B	1 K Ω	
R5B	500 Ω	
R6B	1 K Ω	
R110B	9.05 K Ω	
R111B	OPEN	
R112B	OPEN	
R114B	OPEN	
R11B	31.5 K Ω	
R12B	2 K Ω	
R13B	23.3 K Ω	50 K Ω Pot
R15B	389 Ω	1 K Ω Pot**
R16B	600 Ω	
R17B	40 K Ω	
R18B	16.86 K Ω	
R19B	6.35 K Ω	50 K Ω Pot*

Part #	Value	Note
Resistors (Continued)		
R210B	12 K Ω	
R211B	5 K Ω	
R212B	5 K Ω	
R214B	2.5 K Ω	
R21B	25 K Ω	
R22B	2 K Ω	
R23B	21.4 K Ω	50 K Ω Pot
R25B	279 Ω	500 Ω Pot**
R26B	732 Ω	
R27B	100 K Ω	
R28B	OPEN	
R29B	6 K Ω	100 K Pot*
Capacitors		
C15A	OPEN	
C25A	OPEN	
C15B	OPEN	
C25B	OPEN	
C1	100 μ F	bypass
C3	100 μ F	bypass
C4	0.1 μ F	bypass
C5	0.1 μ F	bypass
C6	0.1 μ F	U1 bypass
C7	0.01 μ F	U1 bypass
C8	0.1 μ F	U1 bypass
C9	0.01 μ F	U1 bypass
C10	OPEN	U2 bypass
C12	OPEN	U2 bypass
C13	OPEN	U2 bypass
C14	OPEN	U2 bypass
C16	0.1 μ F	U3 bypass
C17	0.01 μ F	U3 bypass
C18	0.1 μ F	U3 bypass
C19	0.01 μ F	U3 bypass
C20	0.1 μ F	U4 bypass
C21	0.01 μ F	U4 bypass
C22	0.1 μ F	U4 bypass
C23	0.01 μ F	U4 bypass
C24	0.1 μ F	U5 bypass
C26	0.01 μ F	U5 bypass
C27	0.1 μ F	U5 bypass
C28	0.01 μ F	U5 bypass
C29	0.1 μ F	U6 bypass
C30	0.01 μ F	U6 bypass
C31	0.1 μ F	U6 bypass
C32	0.01 μ F	U6 bypass
C33	OPEN	U7 bypass
C34	OPEN	U7 bypass
C35	OPEN	U7 bypass
C36	OPEN	U7 bypass
C37	0.1 μ F	U8 bypass
C38	0.01 μ F	U8 bypass

ML2111 Application Board Parts List (Continued)

Part #	Value	Note
Capacitors (Continued)		
C39	0.1 μ F	U8 bypass
C40	0.01 μ F	U8 bypass
C41	0.1 μ F	U9 bypass
C42	0.01 μ F	U9 bypass
C43	0.1 μ F	U9 bypass
C44	0.01 μ F	U9 bypass
C45	0.1 μ F	U10 bypass
C46	0.01 μ F	U10 bypass
C47	0.1 μ F	U10 bypass
C48	0.01 μ F	U10 bypass
Jumpers		
J1A	IN	
J2A	OUT	
J1B	IN	
J2B	OUT	

Part #	Value	Note
ICs		
U1	ML2111CCP	
U2	OPEN	
U3	LM318H	
U4	LM318H	
U5	LM318H	
U6	ML2111CCP	
U7	OPEN	
U8	LM318H	
U9	LM318H	
U10	LM318H	
Miscellaneous		
20	scope probe sockets	
3	BNC connectors	
3	female banana plugs	
2	20 pin low profile sockets	

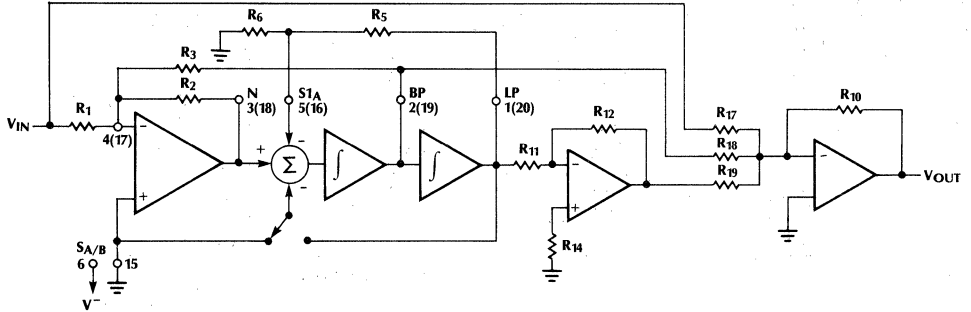
* Gain and zero frequency adjustment. May not be needed if application can tolerate slight variations in stop band.

** R5 - In most cases R5 can be replaced by a 1% resistor after trimming has been done.

Application Note 4

Appendix A. Flexible Building Block Summary

Lowpass —



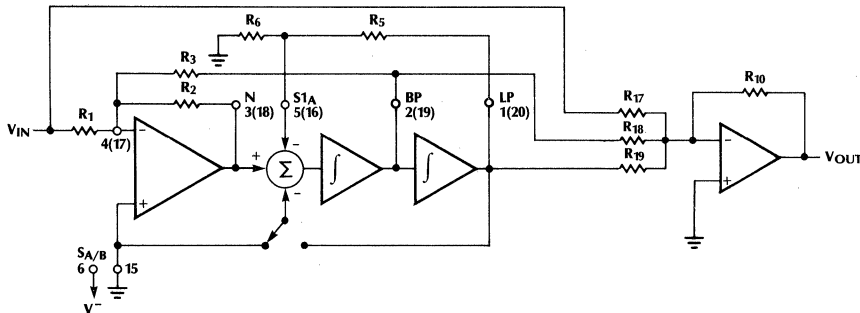
$$f_0 = \frac{f_{CLK}}{100(50)} \sqrt{\frac{R_6}{R_5 + R_6}}; Q = \frac{R_3}{R_2} \sqrt{\frac{R_6}{R_5 + R_6}}$$

$$f_Z = f_0 \sqrt{1 + \frac{R_2 R_{12} R_{17}}{R_1 R_{11} R_{19}} \left(1 + \frac{R_5}{R_6}\right)}$$

$$H_{OLP} = \left(-\frac{R_{10}}{R_{17}}\right) \left(1 + \frac{R_2 R_{12} R_{17}}{R_1 R_{11} R_{19}} \left(1 + \frac{R_5}{R_6}\right)\right)$$

$$\frac{R_3}{R_1} = \frac{R_{18}}{R_{17}}$$

Highpass —

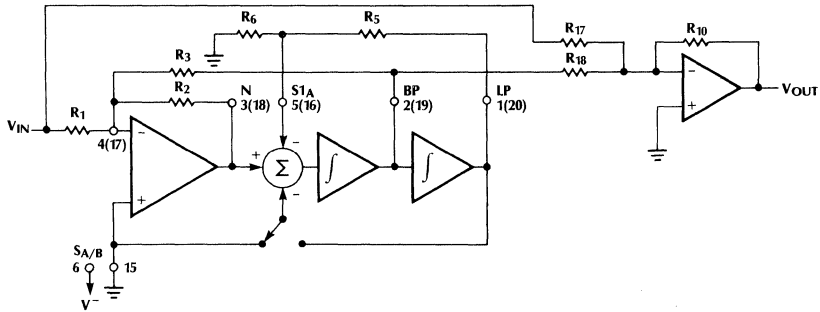


$$f_0 = \frac{f_{CLK}}{100(50)} \sqrt{\frac{R_6}{R_5 + R_6}}; Q = \frac{R_3}{R_2} \sqrt{\frac{R_6}{R_5 + R_6}}$$

$$f_Z = f_0 \sqrt{1 - \frac{R_2 R_{12} R_{17}}{R_1 R_{11} R_{19}} \left(1 + \frac{R_5}{R_6}\right)}; H_{OHP} = -\frac{R_{10}}{R_{17}}$$

$$\frac{R_3}{R_1} = \frac{R_{18}}{R_{17}}$$

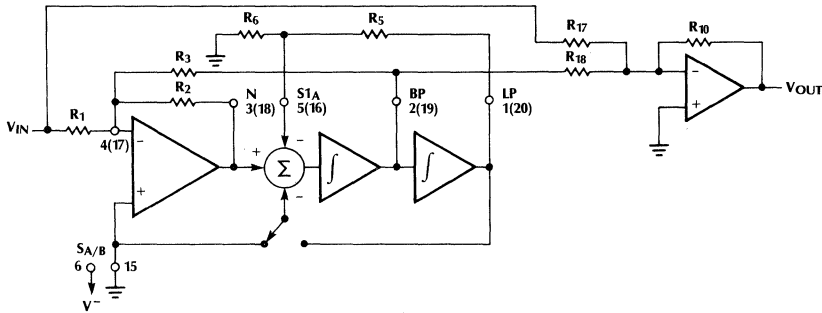
Allpass —



$$f_0 = f_z = \frac{f_{\text{CLK}}}{100(50)} \sqrt{\frac{R_6}{R_5 + R_6}}; Q = \frac{R_3}{R_2} \sqrt{\frac{R_6}{R_5 + R_6}}$$

$$H_{\text{OAP}} = -\frac{R_{10}}{R_{17}}; \frac{R_{17}R_3}{R_{18}R_1} = 2$$

Notch —



$$f_N = \frac{f_{\text{CLK}}}{100(50)} \sqrt{\frac{R_6}{R_5 + R_6}}; Q = \frac{R_3}{R_2} \sqrt{\frac{R_6}{R_5 + R_6}}$$

$$H_{\text{ON1}}(f \rightarrow 0) = H_{\text{ON2}}(f \rightarrow f_{\text{CLK}}/2) = -\frac{R_{10}}{R_{17}}; \frac{R_3}{R_1} = \frac{R_{18}}{R_{17}}$$

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Application Note 4

Appendix B. Derivation of Flexible Building Block Transfer Function

$$V_N = -\frac{R_2}{R_1} V_{IN} - \frac{R_2}{R_3} V_{BP}$$

$$V_{BP} = \frac{\left(V_N - \frac{R_6}{R_5 + R_6} V_{LP} \right)}{s} \omega_1$$

$$V_{LP} = \frac{V_{BP} \omega_1}{s}$$

$$\frac{s V_{BP}}{\omega_1} = -\frac{R_2}{R_1} V_{IN} - \frac{R_2}{R_3} V_{BP} - \frac{R_6}{R_5 + R_6} \frac{V_{BP} \omega_1}{s}$$

$$V_{BP} \left[\frac{s}{\omega_1} + \frac{R_2}{R_3} + \frac{R_6}{R_5 + R_6} \frac{\omega_1}{s} \right] = -\frac{R_2}{R_1} V_{IN}$$

$$\frac{V_{BP}}{V_{IN}} = \left(-\frac{R_2}{R_1} \right) \frac{\frac{R_2}{R_3} s \omega_1}{s^2 + \frac{R_2}{R_3} s \omega_1 + \frac{R_6}{R_5 + R_6} \omega_1^2}$$

$$V_O = -\frac{R_{10}}{R_{17}} V_{IN} - \frac{R_{10}}{R_{18}} V_{BP} + \frac{R_{12} R_{10}}{R_{11} R_{19}} V_{LP}$$

$$V_O = -\frac{R_{10}}{R_{17}} V_{IN} - \frac{R_{10}}{R_{18}} \left(-\frac{R_2}{R_1} \right) \frac{s \omega_1 V_{IN}}{s^2 + \frac{R_2}{R_3} s \omega_1 + \frac{R_6}{R_5 + R_6} \omega_1^2} + \frac{R_{12} R_{10}}{R_{11} R_{19}} \left(-\frac{R_2}{R_1} \right) \frac{\omega_1^2 V_{IN}}{s^2 + \frac{R_2}{R_3} s \omega_1 + \frac{R_6}{R_5 + R_6} \omega_1^2}$$

$$V_O = \left(-\frac{R_{10}}{R_{17}} \right) \left[\frac{s^2 + \frac{R_2}{R_3} s \omega_1 + \frac{R_6}{R_5 + R_6} \omega_1^2 - \frac{R_{10} R_2 R_{17}}{R_{18} R_1 R_{10}} s \omega_1 + \frac{R_{12} R_2 R_{17}}{R_{11} R_1 R_{19}} \omega_1^2}{s^2 + \frac{R_2}{R_3} s \omega_1 + \frac{R_6}{R_5 + R_6} \omega_1^2} \right]$$

$$\frac{V_O}{V_{IN}} = \left(-\frac{R_{10}}{R_{17}} \right) \left[\frac{s^2 + \frac{R_2}{R_3} \left(-\frac{R_{10}}{R_{17}} \right) \left[\frac{R_{10} R_3}{R_{18} R_1} - \frac{R_{10}}{R_{17}} \right] s \omega_1 + \left[1 + \frac{R_2 R_{12} R_{17}}{R_1 R_{11} R_{19}} \left(1 + \frac{R_5}{R_6} \right) \right] \frac{R_6}{R_5 + R_6} \omega_1^2}{s^2 + \frac{R_2}{R_3} s \omega_1 + \frac{R_6}{R_5 + R_6} \omega_1^2} \right]$$

$$\frac{V_O}{V_{IN}} = \left(-\frac{R_{10}}{R_{17}} \right) \left[\frac{s^2 + \frac{R_2}{R_3} \left[1 - \frac{R_{17} R_3}{R_{18} R_1} \right] s \omega_1 + \left[1 + \frac{R_2 R_{12} R_{17}}{R_1 R_{11} R_{19}} \left(1 + \frac{R_5}{R_6} \right) \right] \frac{R_6}{R_5 + R_6} \omega_1^2}{s^2 + \frac{R_2}{R_3} s \omega_1 + \frac{R_6}{R_5 + R_6} \omega_1^2} \right]$$

$$f_1 = \frac{f_{CLK} 2\pi}{50} \quad H_{OBP} = -\frac{R_3}{R_1}$$

$$f_0 = \sqrt{\frac{R_6}{R_5 + R_6}} \frac{f_{CLK} 2\pi}{50} \quad Q = \frac{R_2}{R_3} \sqrt{\frac{R_6}{R_5 + R_6}}$$

$$\omega = 2\pi f$$

ML2200, ML2208 Software Driver

1.0 Introduction

This application note presents a very simple software driver for the ML2200/ML2208 Data Acquisition Peripheral. As mentioned in the data sheet, under section 6.0 "Methods of Data Transfer to the Microprocessor", there are several ways to handle the A/D converted data output from the ML2200/ML2208; 1) Data on Demand, 2) Polling; 3) Interrupt, or 4) DMA. This application note presents a driver for Data on Demand.

An application using Data on Demand requires the A/D converted data at arbitrary times, as opposed to the other three methods of data transfer which requires the microprocessor to periodically read the data. The ML2200/ML2208 operating in a Data on Demand mode is not running continuously. Data on Demand would be more characteristic of a data

acquisition application rather than a signal processing application which would need to sample a signal periodically in order to be able to reconstruct it.

The driver is written in pseudo code, which is no particular language but should be easily translatable to any computer language. It is a step-by-step process of reading and writing values to ML2200/ML2208 registers.

Four modules are covered: Initialization, Activate Conversion and Read Data, Self Test Diagnostic, Self Calibration Diagnostic, Power Down and Power Up Modules. Initialization covers power-up procedures and optionally may call Self Test and Self Calibration Diagnostic modules. Activate Conversion is the steady state module that is called each time the A/D data is desired. Power Down and Power Up are used only if this capability is desired.

Initialization Mode (Power-On Initialization)

- | | |
|--|---|
| 1) Power on | |
| 2) Write (40H) to Control Register | ; Reset |
| 3) Write (80H) to Control Register | ; Set Calibration |
| 4) Wait 16,520 external clocks | |
| 5) Read Status Register | |
| 6) Is CLCP = 1, Yes: continue, No: go back to step 5 | |
| 7) Write (40H) to Interrupt Acknowledge Register | ; CLCPAK |
| -) Call (Self Calibration Diagnostic Module) | ; OPTIONAL |
| | ; Use Program shown on last page figure 1 |
| 8) Write (88H) to Index Register | ; Point to first |
| 9) Write (08H) Window High Reg | ; instruction RAM use |
| 10) Write (26H) Window Low Reg | ; auto increment |
| 11) Write (01H) Window High Reg | |
| 12) Write (26H) Window Low Reg | |
| 13) Write (02H) Window High Reg | |
| 14) Write (26H) Window Low Reg | |
| 15) Write (03H) Window High Reg | |
| 16) Write (26H) Window Low Reg | |
| 17) Write (04H) Window High Reg | |
| 18) Write (26H) Window Low Reg | |
| 19) Write (05H) Window High Reg | |
| 20) Write (26H) Window Low Reg | |
| 21) Write (06H) Window High Reg | |
| 22) Write (26H) Window Low Reg | |
| 23) Write (07H) Window High Reg | |
| 24) Write (26H) Window Low Reg | |

Application Note 5

- 25) Write (0AH) to Control Reg ; set MSTR bit so
; that pulse goes out
; each conversion and
; put in DMA mode to
; facilitate reading
; data.
 - 26) Write (0BH) to Control Reg ; Set Run bit
 - 27) Call (Self Test Diagnostic Module) ; OPTIONAL
- ; end of initialization module.

Activate Conversion and Read Data Module

(Called each time A/D converted data is desired)

- 1) Read status register
- 2) Is ISQ = 1? Yes: continue, No: go back to step 1
- 3) Write (10H) to Interrupt Acknowledge Reg ; acknowledge ISQ
- 4) Wait ($8 \times 31.4 \mu s = 251.2 \mu s$)
- 5) Read status register
- 6) DBR = 1? Yes: continue; No: go back to step 5
- 7) Read Window Low Register save as High Byte ; DMA mode allows
Read Window Low Register save as Low Byte ; μP to read High and
; Low bytes at same
; address
- 8) Go back to step 7 seven more times
- 9) Return

Self Test Diagnostic Module

(Assumes the program in figure 1 is already loaded in the Instruction RAM as performed in the initialization module. When the SLFT bit is set, the diagnostic program is the same one as shown on page 22 of the data sheet. This module sets the SLFT bit, starts a conversion, then checks the data for the results.)

- 1) Read Control Register
- 2) Or (20H) ; Set SLFTST in
; Control Register
- 3) And (7FH) ; don't set CAL bit
- 4) Write back into control register
- 5) Call (ACTIVATE CONVERSION AND READ DATA MODULE)
- 6) Check selftest data ; Data 0 = 0
; Data 1 = +1
; Data 2 = -1
; Data 3 = 0

(Note: these values may not be exact due to the potential noise in the system)

- 7) Read Control Register
- 8) And (5FH) ; clear SLFT bit
- 9) Write back into control register
- 10) Return

Self Calibration Diagnostic Module

(This can be used to verify that the part is properly calibrated. It should be called between steps 7 and 8 in the initialization module. This test is not necessary since each production part is fully tested before it is shipped.)

- 1) Write (08H) to Index Register ; Point to the first
- 2) Write (88H) to Window High Register ; Instruction
- 3) Write (60H) to Window Low Register ; Load with RDCAL
- 4) Write (01H) to Control Register ; Set RUN bit
- 5) Read Status Register ; Wait for ISQ
- 6) Is ISQ = 1? Yes: continue, No: go back to step 5
- 7) Write (10H) to Interrupt Acknowledge Register ; Start Program
- 8) Read Status Register
- 9) Is DBR = 1? Yes: continue, No: go back to step 8 ; Wait for Data
- 10) Write (00H) to Index Register ; Point to Data
- 11) Read Window Low Register
- 12) Is Data = 0FFH? Yes: Failed Calibration, No: continue
- 13) Write (00H) to Control Register ; Take out of Run Mode
- 14) Return

Power Down Module

- 1) Read Control Register ; Clear the Run Bit
- 2) And with (7EH)
- 3) Write Control Register
- 4) Read Status Register and Process any conditions
- 5) Write (0FFH) to Interrupt Acknowledge Register ; Clear all Interrupt
- 6) POWER DOWN (PDN pin goes low) ; Conditions
- 7) Return

Power Up Module (Coming from a Power Down State)

- 1) POWER UP (PDN pin goes high)
- 2) Wait (10 msec)
- 3) Read Control Register ; Set the Run Bit
- 4) Or with (01H)
- 5) And with (7FH)
- 6) Write to Control Register
- 7) Return

	Last	ALRMEN	Mode	CHAN	Cycle	Gain	REF
SEQ0	0	0	Intra Sequence Pause	CH0	13	1	Internal
SEQ1	0	0	Immed Execute	CH1	13	1	Internal
SEQ2	0	0	Immed Execute	CH2	13	1	Internal
SEQ3	0	0	Immed Execute	CH3	13	1	Internal
SEQ4	0	0	Immed Execute	CH4	13	1	Internal
SEQ5	0	0	Immed Execute	CH5	13	1	Internal
SEQ6	0	0	Immed Execute	CH6	13	1	Internal
SEQ7	1	0	Immed Execute	CH7	13	1	Internal

Figure 1. ML2208 Program Used in Driver

Application Note 5

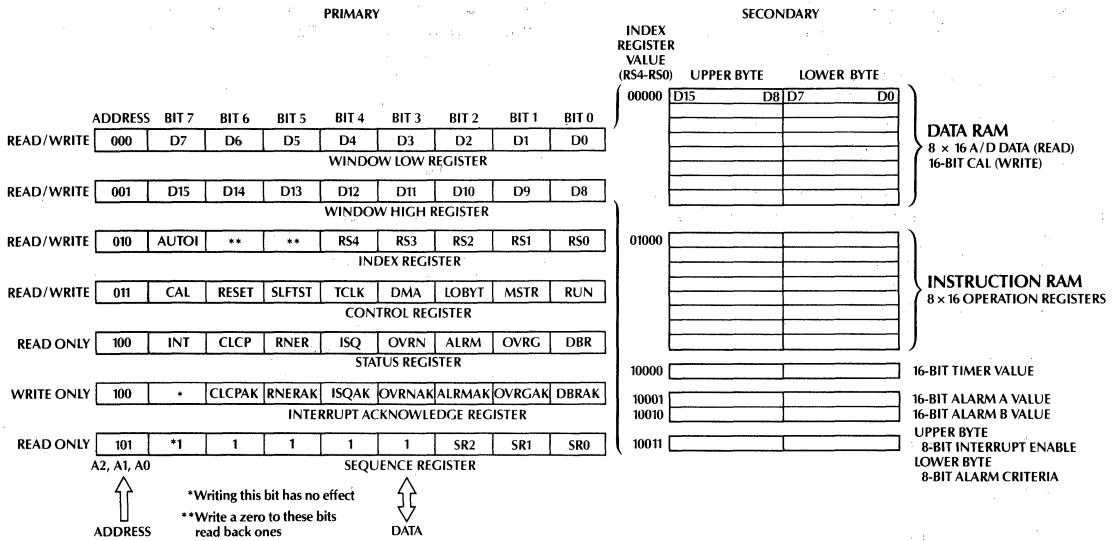


Figure 2. ML2200/ML2208 Registers

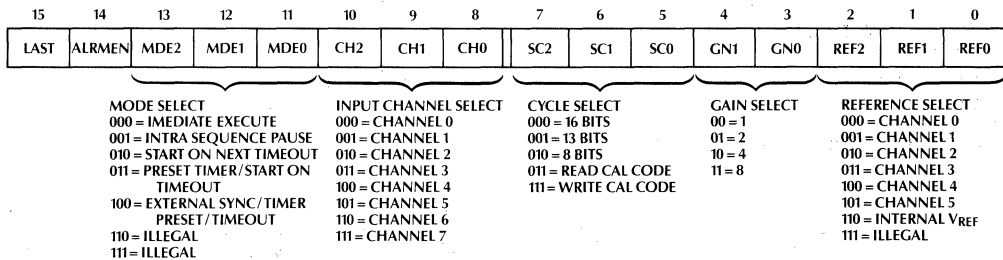


Figure 3. ML2208 Bit Map of Instruction RAM

Fiberoptics

Introduction

Although fiberoptic technology has been around for some time, its cost and the lack of standardization has hindered its widespread application, until recently. The introduction of new integrated circuits developed specifically for fiberoptic systems has lowered the costs, making fiberoptic links more competitive. Applications in Telephony, LANs, WANs, and point to point high speed interfaces, have helped make fiberoptics one of the fastest growing segments in the electronics market.

Micro Linear's fiberoptic products can be used to implement a range of different fiberoptic interfaces. Data rates up to 100 Megabaud which are compatible with ECL or TTL are achievable using a single 5 volt supply. Most of the applications for these products require bandwidths above 1MHz, where the quality of the interface can be compromised with a poor implementation. With this in mind, having a thorough understanding of fiberoptics will significantly contribute to the success of a circuit design. This application note will address the transmit and receive circuits, some important PC board layout techniques, and will conclude with a sample circuit and board layout.

Fiberoptics

Fiberoptic systems have several key advantages over their wire equivalents, which account for the continued effort to make them practical in more applications. The most significant advantage is the low level of attenuation

seen with high frequency signals. This feature allows a higher degree of multiplexing than is achievable using wire. This is exactly what is needed for long distance telephone lines and computer networks. Other attractive features include a lack of RFI radiation and a low sensitivity to EMI noise. These characteristics make it easier to meet FCC regulations and increase the security of data transmissions.

In a fiberoptic system (figure 1) digital data is coded into a serial bit stream represented by bursts of light from a laser diode or LED. This light is channeled by the fiber to a PIN photodiode at the receiver which is sensitive to the frequency of the light transmitted. Because light effects the reverse current flow through a PIN photodiode, a transimpedance amplifier is required to convert this current to a voltage and boost the low level signal to something usable. A quantizing circuit usually follows because variable fiber lengths and conditions will distort the signal. The Quantizer squares the signal and conforms to standard interface levels (ECL or TTL).

Fiberoptics is not a perfect interface, though. The signal level can be attenuated by insertion loss at the transmitter and receiver, connector loss, and transmission loss. These losses limit the maximum length of the fiber and affect the requirements of the transmitter and receiver. In order to accommodate a worst case situation, a flux budget should be developed so that minimum circuit performance levels can be ascertained.

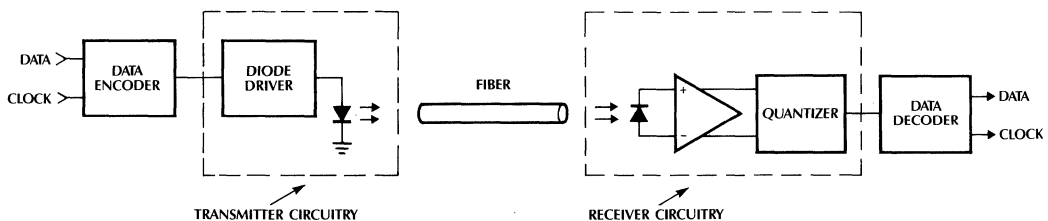


Figure 1.

Application Note 6

Defining a Flux Budget

$$10 \log \left(\frac{\phi_T}{\phi_R} \right) = \alpha_{OL} + \alpha_{TC} + \alpha_{CR} + n\alpha_{CC} + \alpha_M$$

A flux budget is a mathematical representation of the optical power in a fiberoptic system. It accounts for connector losses, attenuation due to fiber length, and a safety margin defined by the designer. Defining this budget is one of the first things that should be done when designing a fiberoptic link.

Each of the terms is defined as follows:

- ϕ_T is the flux (μW) available from the transmitter
- ϕ_R is the flux (μW) required by the receiver
- α_{OL} is the fiber attenuation constant (dB/km)
- L is the fiber length (km)
- α_{TC} is the transmitter-to-fiber coupling loss (dB)
- α_{CC} is the fiber-to-fiber loss (dB) for in-line connectors
- n is the number of in-line connectors
- α_{CR} is the fiber-to-receiver coupling loss (dB)
- α_M is the safety margin (dB)

A graphical representation of the flux budget is shown in figure 2. Option (a) depicts the use of in-line connectors. Option (b) is without them.

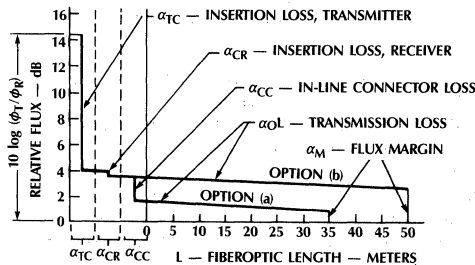


Figure 2.

To keep power consumption at a minimum, the appropriate starting point is the minimum acceptable signal level at the receiver. This minimum received power level, summed with several interface losses gives the minimum output power of the LED. If the fiber length can vary in a given system then the dynamic range of the receiver is important and the maximum received power must also be calculated.

Dynamic Range

The dynamic range of the receiver must be large enough to accommodate all the variables a system may present. Figure 3 shows an example dynamic range calculation for transmission distances ranging from 10 meters to 1000 meters with 12.5dB/km cable, and up to two in-line connectors.

α_{LED} = LED output variation	= 7.0dB
α_{LDC} = LED driver variation	= 2.2dB
α_{OL} = 1km \times 12.5dB/km	= 12.5dB
$n\alpha_{CC}$ = 2 \times 2dB	= 4.0dB
α_M	= 3.0dB
Thermal Variations	= 1.0dB
Dynamic Range	<u>29.7dB</u>

Figure 3.

The Circuit Design

The combination of low receiver input sensitivity with significant dynamic range requires the receiver to have two important features: amplitude control and AC coupling.

An offset voltage in the receiver will reduce its sensitivity by not allowing low level signals to trigger the digital output circuit. The circuit in figure 4 contains both AC coupling between the transimpedance and limiting amplifiers, and a DC restoration loop around the limiting amplifiers. These two features keep the offset voltage through the receiver to an absolute minimum, thus maximizing sensitivity.

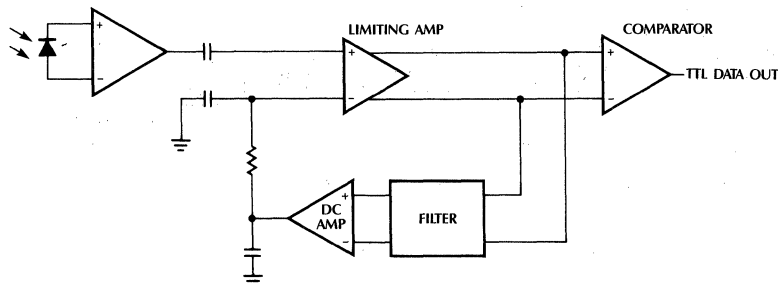


Figure 4.

In order to handle a wide dynamic range, like 50dB, special attention must be given to the receiver circuit design. Applying a large input signal to a typical amplifier can cause the transistors in the signal path to saturate resulting in pulse width distortion and reduced bandwidth. Some technique of controlling the amplitude must be incorporated in order to protect the signal integrity.

Amplitude control can be achieved with either an AGC circuit or with the use of limiting amplifiers. An AGC circuit keeps the transistors out of saturation by reducing the gain of the circuit as the signal amplitude increases. A limiting amplifier simply limits the signal amplitude to a point before saturation. This technique results in a simpler, higher bandwidth design and so was chosen by Micro Linear.

Data Format

The data format is important since it affects the bandwidth and duty cycle which the interface must accommodate. There are many ways to code data in a serial format. Some codes allow unlimited consecutive symbols while others do not. Those that do not are called Run-Length-Limited (RLL) codes. A fiberoptic interface which incorporates AC coupling to increase sensitivity can only pass RLL type codes. The particular run-length-limited code chosen must be considered carefully since it will affect the bandwidth of the system.

Manchester code is popular in AC coupled systems because it has a 50% duty cycle and can be encoded and decoded with relatively simple circuits. In Manchester code two symbols are used for each bit transmitted. This doubles the fundamental frequency which the interface must handle. A more efficient RLL code is 4B5B. This code uses 5 symbols to send 4 bits. This represents an increase in efficiency from 50% (Manchester) to 80% (4B5B). A fiberoptic interface which will transmit 40 Megabits per second using 4B5B coding must accommodate 50 Megabaud (symbols per second). Since there are always 2 symbols per cycle the minimum system bandwidth is 25 Megahertz. If Manchester code was used to transmit 40 Megabits per second the interface would have to handle 80 Megabaud or a minimum bandwidth of 40 Megahertz.

Bandwidth

From the example just described you can see how the code chosen effects the minimum bandwidth of the fiberoptic interface to be designed. The optimum system bandwidth is actually somewhat higher though due to four conflicting concerns: noise, intersymbol interference, power, and bit error rate.

If an interface were designed with a 3dB bandwidth equal to the minimum bandwidth as described above, level transitions would be smooth, like a sine wave.

This is not desirable for a digital signal. Also, smooth rise and fall times will cause interference between adjacent symbols resulting in a distortion of the output signal. This is known as intersymbol interference. A fiberoptic interface with a higher bandwidth will have faster rise and fall times and less intersymbol interference. On the other hand, a higher bandwidth will increase the noise on the output signal. When you combine these two opposing effects with the desire for low power and a low BER (Bit Error Rate) (which also conflict), an optimum bandwidth can be derived. The curve in figure 5 indicates the optimum bandwidth is about 50% higher than the minimum bandwidth.

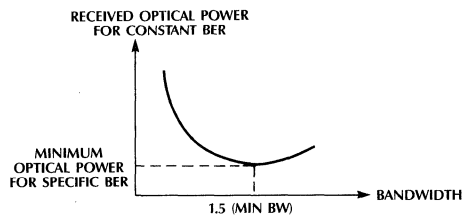


Figure 5.

Transmitter Design

The light source can be either a laser diode or an LED. Because a laser diode has such a narrow spectrum of radiant light it is called a Single Mode light emitter. Multi Mode light emitters radiate a wider spectrum of light. LEDs are Multi Mode and as such suffer from a higher level of chromatic dispersion, caused by different propagation velocities for light of different wavelengths. This is the dominant bandwidth limiting factor for LED driven fiberoptic links. Light emitting diodes have an emission spectrum on the order of 40 to 60nm full width at half maximum amplitude centered at 820nm. On the other hand, LEDs are much cheaper than laser diodes and can be modulated in the 100MHz range, making them suitable for short to medium distance communications such as LANs and point-to-point computer interfaces.

LEDs are current driven devices so a current modulation circuit is needed to use the LED as a data transmitter. In applications where the data rate is less than 10MHz a circuit similar to figure 6 will be adequate to drive the LED without a significant amount of pulse width distortion. Unfortunately, LEDs do not turn on or off linearly nor are their rise and fall times equal. For data rates above 10MHz these characteristics need to be considered in order to get the best possible performance.

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Application Note 6

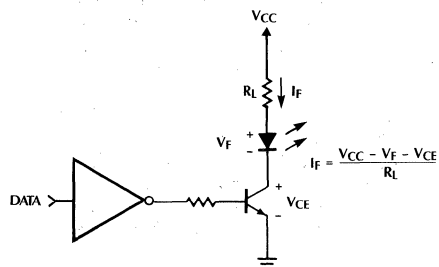


Figure 6.

Two techniques which can be used to improve the turn-on time of an LED are "pre-bias" and "drive current peaking". Pre-bias is a small forward voltage applied to the LED in the "off" state. This voltage prevents the junction and parasitic capacitances from discharging completely when the LED is in the "off" state, thus reducing the amount of charge that the driver must transfer to turn the emitter back on. Drive current peaking is a momentary increase in LED forward current that is provided by the driver during the rising and falling edges of the current pulses that are used to modulate the emitter. The time constant of this peaking circuit needs to be equal to the minority carrier lifetime of the emitter so that the rise and fall times will be improved without causing excessive overshoot of the optical pulses. Figure 7 shows the problems which can result from excessive peaking.

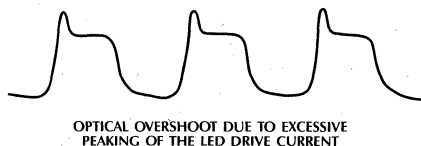


Figure 7.

The circuit in figure 8 implements both the pre-bias and peaking techniques described above. When the DATA signal is low the voltage divider created by R_1 and R_2 can be set-up so the voltage between R_1 and R_2 is slightly less than the LED turn-on voltage. This pre-bias voltage prevents the LED capacitance from discharging completely which allows the LED to turn on faster because less time is required to completely charge the junction capacitance. The time to completely charge the LED can be reduced further by increasing the amount of current flowing in the LED

during turn-on. The capacitor in this circuit has the effect of connecting R_3 in parallel with R_2 for a short time during level transitions. This momentary condition allows additional current to flow through R_3 and the LED. By matching the R_3C time constant to the minority carrier lifetime of the LED, peak performance is achieved.

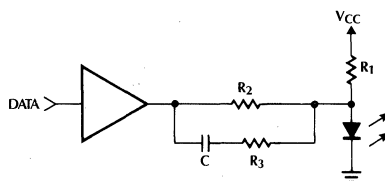


Figure 8.

LEDs are characteristically harder to turn off than to turn on. This phenomenon is commonly referred to as the long-tailed response, and is depicted in figure 9 as it relates to transmitted optical power. Circuits such as the one in figure 6 exhibit this problem because there is no low impedance path to dissipate the stored charge in the LED when turning off. In order to compensate for this an active pull down configuration should be used. For the circuit in figure 8, this can be achieved by using an input buffer with a totem pole output structure. When the DATA signal is low, the lower transistor of the totem pole is active. Acting as a current sink, this device provides a low impedance path for the charge stored in the LED junction, reducing pulse-width distortion and the magnitude of the long tail.

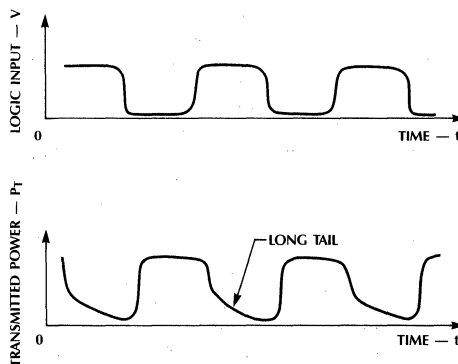


Figure 9.

Receiver Design

For optimum performance the receiver needs to combine a wide dynamic range (about 50dB), high sensitivity (down to $1\mu W$), high bandwidth (50MHz) and compatibility with standard digital interfaces (ECL or TTL). Another feature which is required in some fiberoptic systems is a Link Monitor. This circuit monitors the input level and sets a flag and/or disables the digital output when the input falls below a predetermined point.

The four major functional blocks of a receiver are the optical to current conversion, the current to voltage conversion, the analog to digital conversion, and the Link Monitor. A PIN photodiode and a transimpedance amplifier can be used to perform the first two functions while the third and fourth require several discrete standard devices and a significant amount of design effort or one of Micro Linear's Quantizer products.

There are several manufacturers of discrete PIN photodiodes and transimpedance amplifiers which are suitable for this application. Some of these manufacturers offer both functions in a single module compatible with fiberoptic connectors. These modules, like the Hewlett Packard HFBR-24X6, isolate the most noise sensitive section of the receiver, the PIN photodiode to transimpedance amplifier connection, and protect it from outside influences. In addition, they are relatively low cost, and eliminate the need to design the fiberoptic connector hardware.

The output of these receiver modules is a low level analog voltage which is directly proportional to the incident optical power. This signal needs to be amplified, squared-off, and appropriately level shifted (for ECL or TTL outputs). As described earlier, a limiting amplifier can be used in this application to accommodate a wide input dynamic range while maintaining a high bandwidth.

Building a high speed analog to digital conversion circuit which must perform over a wide dynamic range with low offsets using off-the-shelf components is difficult. Furthermore, a worst case analysis may be impossible because some of the parameters critical to a fiberoptic receiver design, such as input offset and input referred noise are not always included in the discrete component specifications. The typical bandwidth of these devices may be known but the minimum is not always guaranteed, yet this is required for a worst case analysis.

The ML4621 Quantizer

The ML4621 Quantizer eliminates these problems by providing a monolithic solution. This product includes a limiting amplifier front end, a comparator output section and a Link Monitor. The differential data path between

the amplifier section and the comparator section is available to the user for filtering or wave shaping. In addition, both ECL and TTL outputs are available, and the Link Monitor peak detector can be controlled with an external current source or the value of the peak detector capacitor.

Input Amplifier Section

The ML4621 has a two stage limiting amplifier with a DC restoration feedback loop. Figure 10 shows this input circuitry in detail. The two input coupling capacitors C_1 and C_2 perform two important functions. First they eliminate any offset voltage created by the transimpedance amplifier, and in addition they create a high pass filter at the input of the Quantizer. This filter establishes the low corner frequency, f_L , of the Quantizer's 3dB bandwidth.

$$f_L = \frac{1}{2\pi \cdot 8000 \cdot C} \quad (C = C_1 = C_2) \quad (1)$$

8000 represents the parallel combination of the DC bias setting resistors 10K and 35K. Using a $0.1\mu F$ capacitor for C_1 and C_2 establishes a corner frequency at about 200Hz. C_4 and C_5 control the high corner frequency, f_H .

$$f_H = \frac{1}{2\pi \cdot 425 \cdot C} \quad (C = C_4 = C_5) \quad (2)$$

425 represents the internal impedance at nodes CF1 and CF2. Using a 20pF capacitor for C_4 and C_5 establishes a corner frequency at about 19MHz. If CF1 and CF2 are left open the high corner frequency will be $>50\text{MHz}$ for the ML4621. Equation 2 applies when 2 capacitors are tied between CF1 and CF2 to the ground. If one capacitor is used between CF1 and CF2, the value derived for C should be divided by two.

The bandwidth of the receiver, as defined by $f_H - f_L$, can be adjusted to the particular needs of different systems. The high pass filter not only eliminates DC offsets but also reduces any low frequency power supply noise picked-up in the transimpedance amplifier and associated traces. The low pass filter reduces high frequency noise which directly effects the signal to noise ratio and thus the sensitivity of the receiver. Since these circuits were designed for maximum bandwidth, some band limiting should be used as indicated in figure 5 to maximize sensitivity.

Although the input is AC coupled, the offset voltage *within* the limiting amplifiers will be present at V_{OUT+} and V_{OUT-} . This is represented by V_{OS} in figure 11. In order to reduce this error a DC feedback loop is incorporated. First, the DC component of V_{OUT+} and V_{OUT-} is developed through an RC filter of 25K and 10pF. Then a difference amplifier circuit with a gain of 10 is used to provide a single ended signal, stored in C_3 , which can be fed back into the inverting input terminal. This negative feedback loop nulls the offset voltage, forcing V_{OS} to be zero.

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Application Note 6

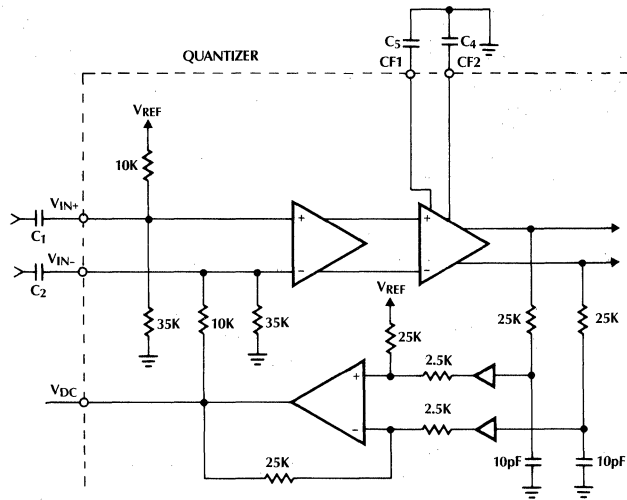


Figure 10.

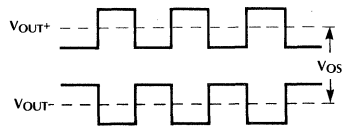


Figure 11.

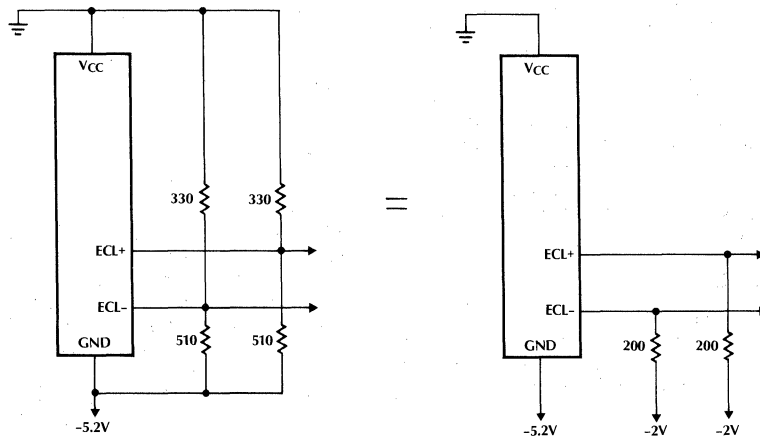


Figure 12.

The limiting amplifiers have a maximum output voltage swing of about 700mV_{p-p}. Since the gain of the amplifiers is 75, input signals greater 9mV will be clipped at about 2.7V and 3.4V. Typically this signal is connected directly to the comparator inputs. If some filtering or wave shaping is desired between the amplifier output and the comparator input, the ML4621 should be used since these nodes are brought out to pins. If AC coupling is involved, the DC bias must be reestablished between (GND + 2V) and (V_{CC} - 1V). Also, the loading on V_{OUT+} and V_{OUT-} should be kept below 3mA, and be aware that CMP+ and CMP- will sink about 25μA.

Output Comparator Section

The ML4621 has both ECL and TTL outputs. If the ECL output is to be used, the power to the TTL output section can be removed by connecting V_{CC} TTL and GND TTL to V_{CC}. This will reduce the V_{CC} supply current by 5 to 10mA. The Quantizer can be powered by -5.2V (V_{CC} = 0V and GND = -5.2V), which produces standard ECL output levels, or +5V (V_{CC} = +5V and GND = 0V), providing raised ECL levels. The ECL outputs on the ML4621 can source up to 10mA, so a 200Ω load tied to -2V (below V_{CC}) can be accommodated. If a -2V supply is not available, connecting the ECL output to GND through a 510Ω resistor, and to V_{CC} through a 330Ω resistor will provide the same voltage swing as 200Ω tied to -2V (see figure 12).

The output comparator is gated with the CMP ENABLE pin which is active low. When CMP ENABLE is high, ECL+ is held high, ECL- is held low, and TTL OUT is held high. If the Quantizer is powered by +5V and ground then any external TTL compatible signal can be used to control this pin. If a -5.2V supply is used, the signal should be appropriately level shifted. In either case, the TTL LINK MON pin can be used to drive the CMP ENABLE pin directly. The TTL LINK MON is an output signal from the Minimum Signal Discriminator circuit providing the Link Monitor function.

Link Monitor Section

The TTL LINK MON and ECL LINK MON pins both provide an output signal indicating when the input data signal is below a user defined acceptable level. Under normal operating conditions this output will be low, indicating the data is of acceptable amplitude. The voltage levels on the TTL LINK MON pin are TTL compatible if the power supply is +5V. With a -5.2V supply the ECL LINK MON output pin will provide single ended ECL levels. The TTL LINK MON pin can also be used to drive an LED, providing a visible link status indicator. This pin can sink up to 10mA.

The Minimum Signal Discriminator circuit contains a peak detector, a comparator, and output level shift circuitry (figure 13). The droop rate of the peak detector is:

$$\frac{dV}{dt} = \frac{I_{SET}}{C_6} \quad (3)$$

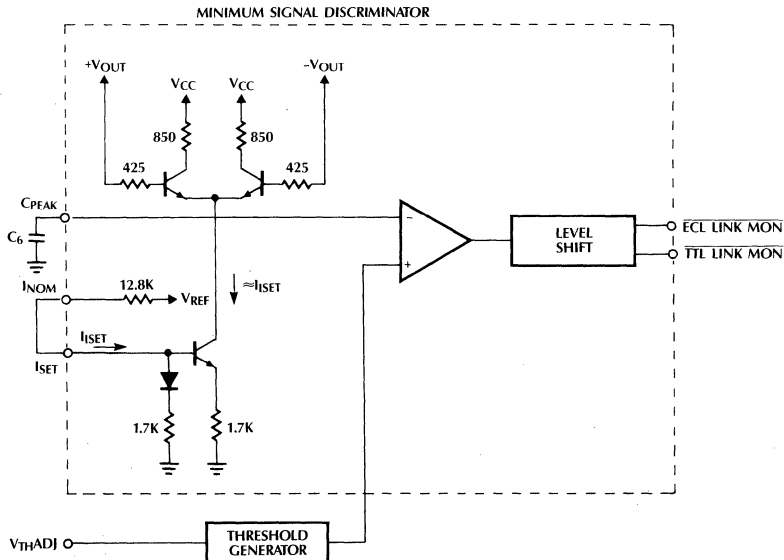


Figure 13.

Application Note 6

The peak detector droop rate can be controlled adjusting either the value of C_6 at the C_{PEAK} pin or I_{SET} at the I_{SET} pin. If I_{NOM} is connected to I_{SET} , I_{SET} will be $125\mu A$. The droop rate for this product can be adjusted with C_6 . The ML4621 has these extra pins, which allows the user to set I_{SET} with an external resistor, R_{EXT} , tied between I_{SET} and V_{CC} . I_{SET} would then be:

$$I_{SET} = \frac{V_{CC} - 0.7}{R_{EXT} + 1700} \quad (4)$$

The output of the peak detector is a DC voltage proportional to half the peak-to-peak voltage between V_{OUT+} and V_{OUT-} . If this signal is larger than the voltage provided by the Threshold Generator circuitry the TTL LINK MON and ECL LINK MON pins will both be low.

The Threshold Generator level shifts the reference voltage at V_{THADJ} through a circuit which has a temperature coefficient matching that of the limiting amplifiers. This improves the accuracy of the Link Monitor over temperature. The relationship between V_{THADJ} and V_{TH} (the minimum voltage at the input which will trigger the Link Monitor) is:

$$V_{THADJ} = 600V_{TH} + 0.7 \quad (5)$$

In this equation V_{TH} is the *peak* value of the input signal. The operating range over which these equations apply is indicated by the graphs in figure 14. The on-chip reference voltage, V_{REF} , can be tied directly to

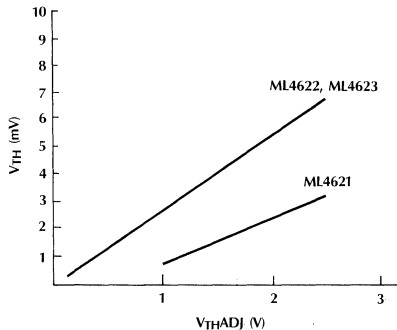


Figure 14.

V_{THADJ} to set the threshold level. This 2.5V low impedance source will set the threshold at its maximum allowable level as indicated in the graph. A lower threshold level can be set by dividing down V_{REF} with a resistor string, as in figure 15. The V_{THADJ} voltage can be calculated as follows:

$$V_{THADJ} = V_{REF} \frac{R_2}{R_1 + R_2} \quad (6)$$

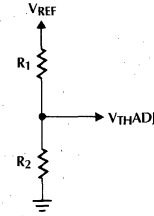


Figure 15.

If, for example, you were using the ML4621 and you wanted the Link Monitor to trigger when the received optical power went below $1\mu W$ ($-30dBm$), you first need to calculate the resultant voltage at V_{IN+} and V_{IN-} . If you were using the HFBR-24X6 Fiberoptic Receiver with a responsivity of $8mV/\mu W$, the peak-to-peak voltage would be:

$$1\mu W \times 8mV/\mu W = 8mV_{P-P} \quad (7)$$

So the Link Monitor should trigger at some point slightly lower than 4mV peak, say 3mV. The reference voltage at V_{THADJ} should then be:

$$V_{THADJ} = 600(.003) + 0.7 = 2.5V \quad (8)$$

This is a convenient value since the reference voltage supplied by the Quantizer, V_{REF} , is 2.5V. Thus, shorting V_{REF} to V_{THADJ} on the ML4621 will set the minimum input signal level at about 3mV.

The Link Monitor has about 0.4mV (peak) hysteresis built-in. V_{THADJ} in equation 5 is the *high* threshold level (the trigger point when the input voltage is rising). The *low* threshold level (the trigger point when the input voltage is falling) is about 0.4mV less than the levels given in these equations. More hysteresis can be induced by connecting a resistor between TTL LINK MON and V_{THADJ} creating a positive feedback loop.

A Sample Circuit

The circuit in this section (figure 16) is a point-to-point fiberoptic interface designed to pass 20MBd data over 1 kilometer of 62.5/125 μm fiber cable. The main components are the ML4632 LED driver, the HFBR-1414 LED, the HFBR-2416 Receiver, and the ML4621 Quantizer. Choosing $-30dBm$ for the minimum received optical power makes equations 7 and 8 applicable and allows V_{REF} to be used to set the Link Monitor.

Applying figure 5 to the 20MBd data rate yields an optimal bandwidth of about 15MHz. Using equation 2, the value for C_5 is derived by setting f_H equal to 15MHz and solving for C. The lower corner frequency, f_L , should be chosen so that, at least, any 60Hz line noise is filtered out. Choosing 0.1 μF for C_1 and C_2 will set the lower corner at 200Hz (equation 1).

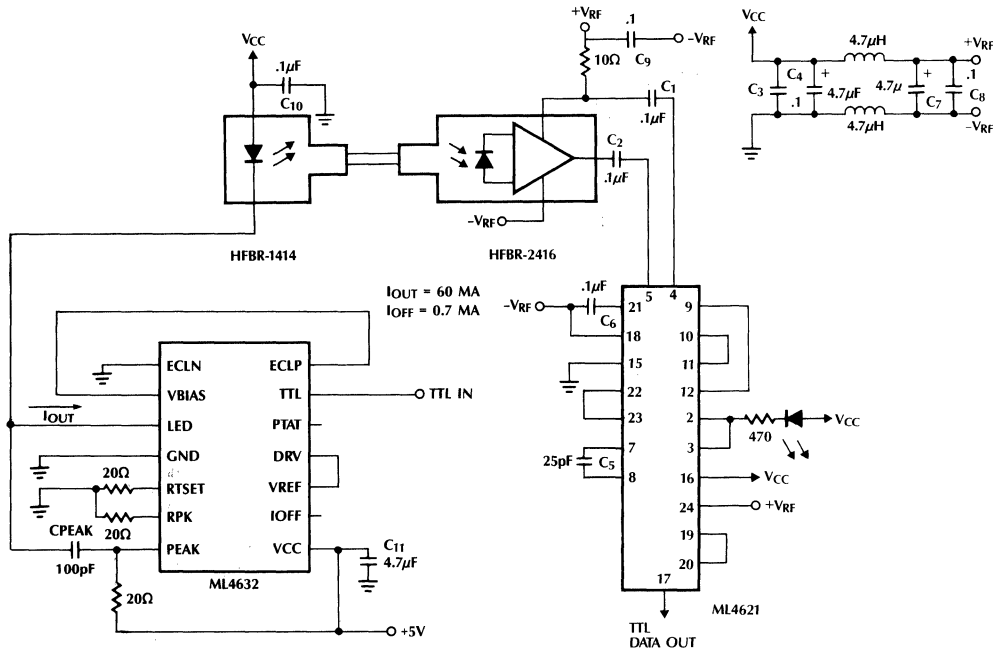


Figure 16.

The only external component left to calculate is C_6 . Since the baud rate is 20Mbd, the time between peaks in the Link Monitor's peak detector is 50ns. If no more than 0.1% droop is acceptable under the worst case conditions (when the input signal is the smallest), and the internal current source I_{SET} is used, then C_6 is calculated as follows:

- Smallest input signal = 3mV
- Gain through amplifiers = 75
- Smallest voltage at V_{OUT+} and V_{OUT-} = $3mV \times 75 = 225mV$
- Smallest acceptable droop voltage = $0.1\% \times 225mV = 225\mu V$
- Slowest acceptable droop rate = $\frac{225\mu V}{50ns} = 4.5V/\mu s$
- Smallest $C_6 = \frac{125\mu A}{4.5V/\mu s} = .028\mu F$

A 0.1 μF capacitor is a convenient acceptable value for this application.

Now that the minimum received power is known, a flux budget can be developed for the interface, and the required optical power from the LED can be derived. Since the output power of the HFBR-1414 is specified out of a short length of fiber attached to the LED unit, no α_T term is required in the flux budget. If no in-line connectors are used and the remaining terms are:

- $\phi_R = 1\mu W$
- $\alpha_O = 10dB/km$
- $L = 1km$
- $\alpha_{CR} = 0.2dB$
- $\alpha_M = 3.0dB$

solving the flux budget equation for ϕ_T yields:

$$10 \log \frac{\phi_T}{\phi_R} = \alpha_O L + \alpha_{CR} + \alpha_M \quad (9)$$

$$10 \log \frac{\phi_T}{1\mu W} = 10(1) + 0.2 + 3 = 13.2dB$$

$$\log \phi_T - \log 1\mu W = 1.32$$

$$\log \phi_T + 6 = 1.32$$

$$\log \phi_T = -4.68$$

$$\phi_T = 20.9\mu W \text{ (-16.8dBm)}$$

The HFBR-1414 has a minimum Peak Output Optical Power of 31.6 μW (-15dBm) when coupled to a 62.5/125 μm fiber cable, and when a forward current (I_f) of 60mA is applied. Since the optical output power vs. forward current relationship of the LED is approximately linear, the forward current required to get the minimum output power, 31.6 μW , can be calculated as follows:

$$\frac{20.9\mu W}{31.6\mu W} \times 60mA = 40mA \quad (10)$$

Application Note 6

To program the ML4632 RTSET can vary from 12 to 30 ohm. This will allow and drive current of 40mA to 100mA. For more detailed information refer to the ML4632 Data Sheet.

Now, to make sure there is no chance of saturating the receiver with too much power, a dynamic range calculation is in order. Since there are no in-line connectors and the cable length is fixed, the only dynamic range components are the thermal variations (α_T), the user defined system margin (α_M), the LED output power tolerance (α_{LED}), and the LED drive circuit tolerance (α_{LDC}):

$$\alpha_{LDC} = 10 \log \frac{\frac{100}{60} (-12\text{dBm})}{\frac{40}{60} (-12\text{dBm})} = 3.97\text{dB} \quad (11)$$

$$\alpha_{LED} = 10 \log \frac{-9\text{dBm}}{-16\text{dBm}} = 7.0\text{dB} \quad (12)$$

α_{LDC}	=	3.97dB
α_{LED}	=	7.0dB
α_M	=	3.0dB
α_T	=	1.0dB
Dynamic Range:		14.98dB

So the maximum input power (ϕ_{TMAX}) the HFBR-2416 Receiver will see is:

$$10 \log \frac{\phi_{TMAX}}{1\mu\text{W}} = 14.98\text{dB} \quad (13)$$

$$\phi_{TMAX} = 31.47\mu\text{W}$$

This is well below the 150 μW maximum spec for the HFBR-2416, and the resultant output voltage is

$$31.47\mu\text{W} \times 8\text{mV}/\mu\text{W} = 251.76\text{mV}_{P-P} \quad (14)$$

This is well below the 1.4V maximum input voltage of the ML4621.

The Board Layout

Refer to Application Note 15.

Harlan Ohara & Vince Cardinale

Expanding the ML2200 Input Multiplexer

1.0 Introduction

If the four channel differential input multiplexer on the ML2200 is insufficient for your application, it can be expanded using one of three methods described in this document. An expanded input multiplexer will greatly enhance the processing power of the ML2200 but will restrict some of its flexibility. The limitations of each circuit are discussed at the end of this application note.

The first circuit controls up to 64 differential inputs but restricts the ML2200 to always run eight operations. The second circuit controls up to 128 single ended inputs and again restricts the ML2200 to always run eight operations. The third circuit is limited to eight differential inputs but is fully programmable in the number of operations. Each circuit is fully synchronized with the ML2200 and can be built with off-the-shelf components.

Although this application note discusses only the ML2200, the ML2208 can be used as well. Only minor operational issues within the ML2208 are affected.

2.0 General Theory of Operation

An external counter (74LS163) is used to control the additional multiplexer devices (DG506 or DG507). The counter is incremented by the SYNC pin of the ML2200, which must be programmed as an output. The SYNC pin is suitable for this purpose since it always signifies the start of a new operation.

Synchronization with the ML2200 is achieved by utilizing the DBR pin to load input channel #1 into the counter. Since the DBR signal comes out after the sequence of operations are complete and the next sequence is started, it is too late to correctly synchronize the counter prior to the beginning of the next sequence. Synchronizing on the "1" count, however, will always reset the counter to the proper value if synchronization is ever lost.

Due to the above described behavior, a "boundry" problem exists in the very first sequence of operations and the first operation of the second sequence after the chip is started. In order to get out of this problem, the RUN bit of the control register is decoded and duplicated in these circuits. The RESET signal is also brought in. These signals force the counter to predetermined states and relieves the "boundry" problem.

3.0 A 64-Channel Differential Input Circuit

This circuit, shown in figure 1, provides up to 64 additional input channels. Two 74LS163 counters, U5 and U6, are used to develop the address for each channel. Only 6 of the 8 available counter bits are needed to control the channel selection. The 3 LSBs are used to drive the multiplexer (DG507) address bus and the 3 MSBs are decoded and used to drive each multiplexer's enable pin. A 74LS138 (U7) is used to decode the MSBs, providing control for 8 multiplexer chips. The 2 unused counter bits could be used to address additional multiplexers, providing control for up to 256 channels.

A D-type flip-flop (U4A, 74LS74) is used to reset both counters whenever the ML2200 is reset or not in RUN mode. This is accomplished by presenting an active low signal at the synchronous clear inputs of the counters whenever either the RESET or HALT condition exists. When the ML2200 is placed in RUN mode, the first SYNC pulse resets the counters to zero instead of incrementing them. At the same time, the first SYNC pulse clears the resetting condition by clearing the flip-flop.

The other flip-flop in this circuit, U4B, is used to trap the 0 to 1 transition of the DBR signal. This transition causes the output of U4B to go low, which sets up the counters for a load cycle. The next SYNC pulse (which should correspond to the first operation of the ML2200) will then load the counter with a "1", forcing the counters to be synchronized to the sequence. As before, the same pulse that performs the load operation also clears the load operation.

U1 (74LS138) and U2 (74LS379) form the logic that decodes the RUN bit from the microprocessor bus to develop one of the conditions that resets the counters. U1 simply decodes the address of the control register within the ML2200 and U2 latches the status of the RUN bit.

This circuit requires operations to be done in groups of eight and each operation within the group must have the same characteristics.

1

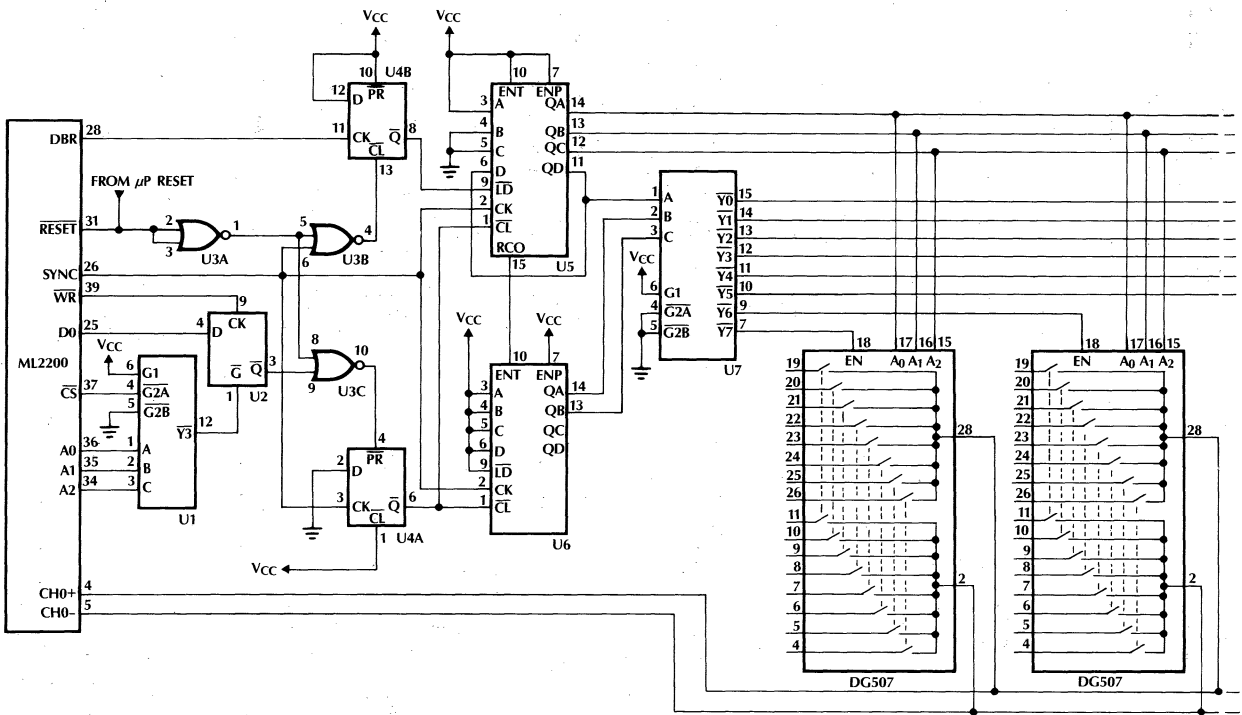


Figure 1.

4.0 A 128-Channel Single Ended Input Circuit

This circuit, shown in figure 2, is identical to the differential circuit with three exceptions:

1. Four address bits are used to drive each multiplexer, since each multiplexer chip now contains 16 channels instead of 8.
2. The decoder chip (U7) is shifted one bit up on the counter output. This makes room for the extra bit needed for multiplexer addressing.
3. DG506 multiplexers are used because this is a single ended application.

This circuit requires operations to be done in groups of eight and each operation within the group must have the same characteristics.

5.0 An 8-Channel Differential Input Circuit

This circuit, shown in figure 3, is very similar to the two previous circuits. It still uses two D-type flip-flops to load and reset the counters. The difference here is that register U7 (74LS379) is provided at address location 6 within the 8 byte ML2200 address space and is used to store a count equal to the number of operations programmed by the microprocessor in the ML2200.

Address location 6 and 7 are spare locations within the ML2200 address space. A 74LS85 (U8) four bit comparator is used to reset the counter to zero when the maximum count is reached. A single DG507 provides the eight input channels.

This circuit is not restricted to performing operations in groups of eight. Each channel can be programmed and initiated individually. This flexibility is maintained at the expense of limiting the number of differential inputs to eight.

6.0 Circuit Limitations

In order to maintain synchronization with the ML2200, these circuits contain several inherent limitations which are described below:

1. The SYNC pin is limited to use as an output.
2. These designs allow less settling time for external input circuits, such as instrumentation amplifiers, than otherwise would be possible.
3. The capability to do random reference selection is lost.
4. The capability to do random input channel selection is lost. Input channels must be scanned sequentially.

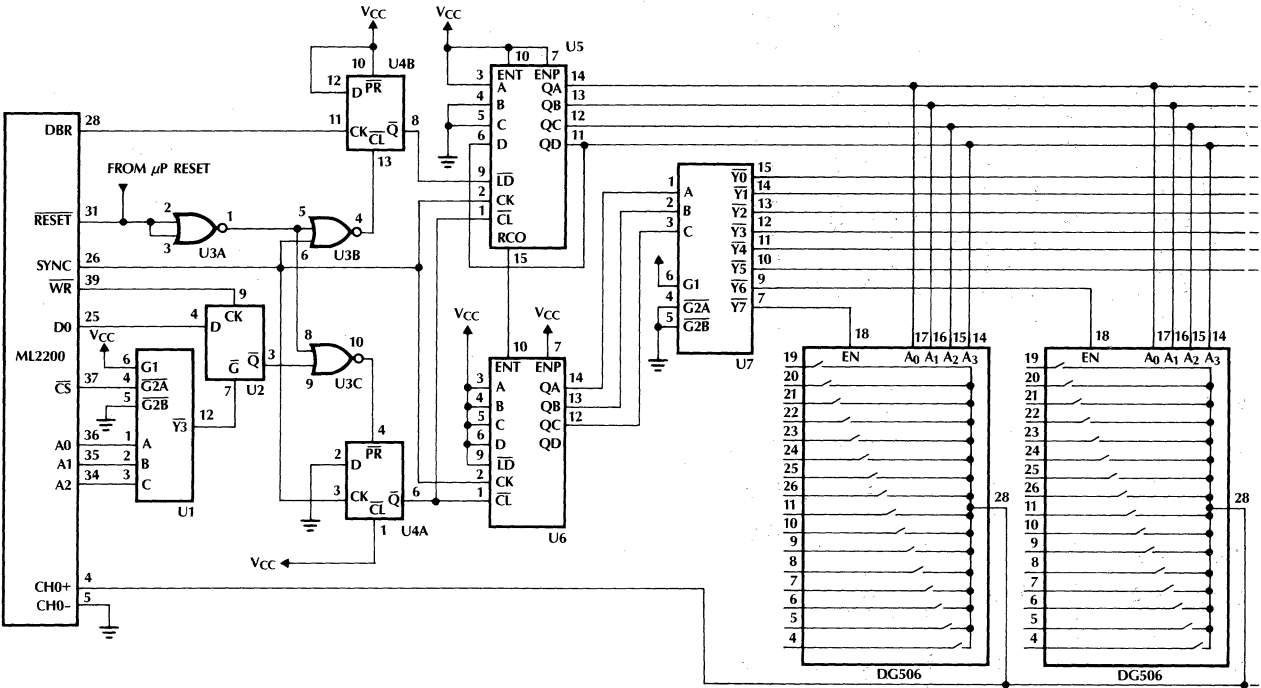


Figure 2.

Harlan Ohara

One Pin Crystal Oscillators

1.0 Introduction

Micro Linear has frequently used a one pin oscillator design in its CMOS chips. The concept of a one pin oscillator may seem peculiar to some at first, but the design topology has been around for many years, dating back to vacuum tube days. This topology is shown in Figure 1 and is commonly known as the Colpitts oscillator. The only difference compared to previous implementations is that an MOS transistor is used as the active element.

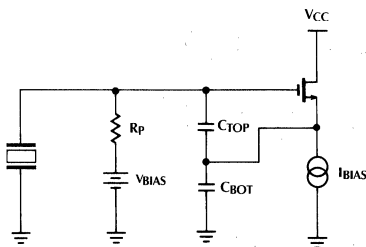


Figure 1.

There are two important advantages to using this particular topology versus the more common two pin design (which is called a Pierce oscillator):

1. Only one pin is required, leaving the extra pin for maximum functionality. This is increasingly important as chips become more complex in function.
2. No external components are usually required except for the crystal. If extremely high frequency accuracy is required, then an external capacitor in parallel with the crystal can be used to trim the frequency.

All is not free, however, there are some disadvantages:

1. This design is less tolerant of external parasitics to ground than the two pin design. This is not usually a problem since the designs used in Micro Linear's circuits have been provided with sufficient margin to handle typical printed circuit board parasitics.

2. Flexibility in terms of user adjustment of design parameters is less in this design. Again, this is not seen to be a problem for two reasons:

- a) Board level designers rarely adjust the two pin design parameters.
- b) Enough margin is provided so that adjustment is not needed.

2.0 Theory of Operation

Exact circuit analysis of the oscillator is a complex procedure for several reasons:

1. In a practical design situation, the system equations are a minimum of 5th order. Exact hand calculations are difficult at best.
2. Final oscillation conditions are not only based on small signal analysis, but very dependent on large signal non-linear situations.
3. The crystal model is generally a simple case in small signal analysis but element parameters can change with excitation level.

In this section, no attempt is made to provide a complete exact analysis. An alternate approach is taken in which hand calculations can closely approximate the small signal solution. This approach is also much more heuristically satisfying in that effects of design parameter changes can be seen without applying a lot of math. This also serves to provide the user with a way to calculate an approximate frequency of oscillation if more exact frequency tolerances are required other than just plugging the crystal in.

For the more technically inclined, an exact small signal sample design procedure is presented in Appendix A using the MathCAD™ program on an IBM-compatible PC. This is possible due to the presence of a root solver capability in MathCAD. Users of HP calculators or numerical analysis computer programs can also perform this analysis. Appendix B contains an example procedure to estimate final oscillation amplitude while Appendix C goes through the procedure to calculate the closed loop root locus plot which is then used to estimate oscillator startup time.

2.1 Crystal Model

The typical crystal model is shown below:

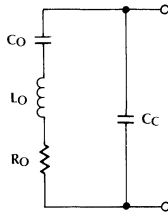


Figure 2.

Typical values for the 12.352 megahertz crystal used in some of Micro Linear's telecom chips are:

$$\begin{aligned} L_0 &:= 8.005814 \cdot 10^{-3} \text{ henries} \\ C_c &:= 5.10 \cdot 10^{-12} \text{ farad} \\ C_0 &:= 20.7558 \cdot 10^{-15} \text{ farad} \\ R_0 &:= 15 \text{ ohms} \end{aligned}$$

The admittance of the crystal is:

$$Y_{XTL}(s) := C_c \cdot s + \frac{1}{L_0 \cdot s + \frac{1}{C_0 \cdot s} + R_0}$$

Plot 50 points versus frequency:

$$x := 1 \dots 50$$

Define radian frequency values:

$$\omega_x := 12.346 \cdot 10^6 \cdot 2 \cdot \pi + x \cdot 120$$

Plot real (resistive) part:

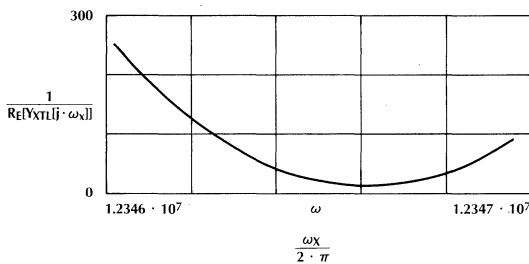


Figure 3.

Plot susceptance:

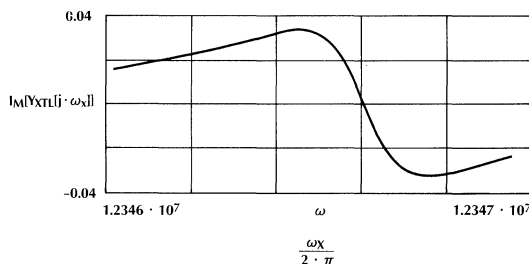


Figure 4.

Note that at the series resonance frequency of:

$$\frac{1}{2 \cdot \pi \cdot \sqrt{L_0 \cdot C_0}} = 1.2346608 \cdot 10^7 \text{ Hz}$$

the susceptance is zero and the resistance is: $R_0 = 15$.

Oscillators that operate the crystal in the series mode use this characteristic as part of a feedback loop in which the loop gain is maximum at this frequency. Above this frequency the susceptance is inductive. Oscillators such as this one-pin design (and the two-pin Pierce) operate the crystal in the parallel mode, "using it as an inductor."

2.2 Simplified Hand Calculation of Loop Gain:

In this section, a simplified (but approximate) method of calculating the loop gain is shown. This method also demonstrates in a more heuristic way how loading affects the oscillator and how one may choose the crystal characteristics, especially the series resistance R_0 . The calculation for the approximate frequency of oscillation is also shown. This calculation is quite accurate.

Before we proceed with the calculations, two general principles will be presented that are used to make the calculations. Derivations of these principles can be found in reference [1]:

1. In Figure 5, it is seen that an RLC circuit with series loss can be represented by a circuit with parallel loss (resistance). This applies when the circuit Q is high (>10).

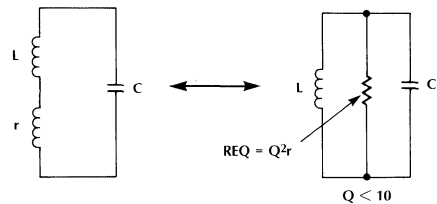


Figure 5.

2. In Figure 6, an RLC circuit with series capacitors can be classified as a "parallel resonant transformer" circuit. Again, this is accurate only when the Q is high (this is easily satisfied with crystal circuits, with Q's in the ten to hundreds of thousands). In this case, the two capacitors act like a transformer with a turns ratio of:

$$n := \frac{C_1}{C_1 + C_2}$$

Hence any resistance can be reflected by the square of the turns ratio.

This capacitance in parallel with the crystal inductance just so happens to produce a resonant frequency which is very close to the frequency of oscillation:

$$\omega_O := \frac{1}{\sqrt{L_O \cdot C_{TOT}}} \quad \omega_O = 7.762 \cdot 10^7$$

$$\frac{\omega_O}{2 \cdot \pi} = 1.2353724 \cdot 10^7$$

Note that this frequency is .014% higher than the specified 12.352 Megahertz. This is because this crystal was ground with a specified capacitive load of 18pF across its terminals. In our case, we only have about 13pF.

The circuit Q is now calculated:

$$Q := \frac{1}{\omega_O \cdot R_O \cdot C_{TOT}} \quad Q = 4.143 \cdot 10^4$$

The equivalent parallel resistance across the crystal inductance is now calculated using principle 1) above:

$$R_{PX} := Q^2 \cdot R_O \quad R_{PX} = 2.574 \cdot 10^{10}$$

This is now reflected to the oscillator pin using principle 2) above:

$$R_{PXR} := R_{PX} \cdot \left[\frac{C_O}{C_O + C_{PTOT}} \right]^2 \quad R_{PXR} = 3.415 \cdot 10^4$$

This is now combined with the parallel resistance present at the oscillator pin:

$$R_{PTOT} := \frac{R_{PXR} \cdot R_P}{R_{PXR} + R_P} \quad R_{PTOT} = 2.546 \cdot 10^4$$

Note that this can be reflected again through the "capacitive transformer" of C_{TOP} and C_{BOT} :

$$R_{PTOTR} := R_{PTOT} \cdot \left[\frac{C_{TOP}}{C_{TOP} + C_{BOT}} \right]^2$$

$$R_{PTOTR} = 6.364 \cdot 10^3$$

We now combine this with the resistance present at the transistor source:

$$R_L := \frac{R_{LT} \cdot R_{PTOTR}}{R_{LT} + R_{PTOTR}} \quad R_L = 5.895 \cdot 10^3$$

This is the load resistance. This multiplied by the transconductance will give us our gain up to the source of the transistor from the input.

$$g_m R_L = 9.433$$

Note that the input to our circuit is the voltage applied across the gate to source of the MOS transistor, or in other words, the voltage across C_{TOT} . Using the "capacitive transformer principle" we can determine that the loop gain is:

$$A_L := \frac{C_{BOT}}{C_{TOP}} \cdot g_m \cdot R_L \quad A_L = 9.433$$

This gives us our loop gain, which hopefully is more than 1 to allow oscillations. Note that this compares favorably with the exact analysis given in Appendix A of 9.41.

This is a good time to pause and reflect on what the above analysis tells us:

1. It is seen that the oscillation frequency depends on the total capacitance at the oscillator pin in series with the crystal capacitor C_O . Since C_O is very small (typically 10's of femto-farads) external capacitance has a small effect on the oscillation frequency. This provides a means of "tweaking" the frequency to an exact value with a trimmer capacitor placed from the oscillator pin to ground.
2. Using the "capacitive transformer" principle, it is seen that large values of capacitance at the oscillator pin reduces the loop gain since the crystal resistance is now reflected into a smaller value and hence the product $g_m \cdot R_L$ is smaller. Oscilloscope probes can contribute a significant amount of parasitic and should be used carefully when debugging this circuit. If frequency trimming is employed by placing a parallel adjustment capacitor to ground, it must be done carefully so that the loop gain is not made too small.
3. Lossy components at the oscillator pin also reduces the product $g_m \cdot R_L$. This is especially important at higher crystal frequencies, where printed circuit board material or socket material becomes more and more lossy. The value of the on-chip bias resistor varies with frequency from about 1M Ω to about 100k Ω (over a 1 to 20 MHz range). Note that this is a fairly high impedance which is easily affected by external parasitics. Oscilloscope probes can be particularly lossy at these frequencies.

2.3 Three More Important Criteria for Consideration:

Three more important items need to be covered. These items may or may not be OK even if the loop gain calculation is adequate (more than 1):

1. Oscillator phase margin.
2. Nyquist criterion.
3. Final oscillation amplitude

The theory for the above criteria is too lengthy to cover in an application note; only a brief qualitative explanation will be given. The reader is encouraged to consult references [3], [4], and [5]. The exact analysis given in Appendix A will cover the phase margin and Nyquist criterion.

1

Application Note 8

Oscillator Phase Margin:

Figures 11A through 11D in Appendix A show the open loop transfer characteristics of the oscillator. 11A shows the overall magnitude over a wide range of frequencies.

The crystal characteristic is not visible since it occurs over a very narrow range. 11B shows the phase characteristic.

These curves show that the circuit produces a single pole rolloff of 6dB per decade and a final phase shift of 90 degrees. Examining the circuit in Figure 10, we see that the loop transfer function starts out at 180 degrees out of phase. This is because the output is developed across C_{BOT} and the input is taken across C_{TOP} . At DC, the voltage across C_{TOP} is of opposite phase to that of C_{BOT} , since the top side of C_{TOP} has a DC path to ground. The single pole rolloff is primarily due to (but not exactly) the combination of R_{IT} and the total capacitance seen at the source of the MOS transistor to ground. The crystal inductance combined with the circuit capacitances can almost provide another 180 degrees of phase shift. This, combined with the 180 degrees from the active element will provide ALMOST, but not quite the 360 degrees needed for oscillation. This is where the single pole rolloff comes in. Examining 11C shows that the loop gain peaks first then dips. This is due to primarily a complex pole pair and a complex zero pair. The peak is a result of the crystal resonating with all circuit capacitances:

$$\omega_p := \frac{1}{\sqrt{L_O \cdot C_{TOT}}} \quad \omega_p = 7.7620737 \cdot 10^7$$

The complex zero pair comes about when the crystal resonates with all capacitances except for the C_{TOP} and C_{BOT} combination:

$$\omega_z := \sqrt{L_O \cdot \left[\frac{C_p \cdot C_o}{C_p + C_o} \right]} \quad \omega_z = 7.7656489 \cdot 10^7$$

The phase shift at the complex pole pair passes through zero and the amplitude peaks to provide the oscillation point. The phase then goes past the point needed for oscillation and then passes through zero again at the complex zero location, returning to the 90 degree point where it started. The amount that the phase shift passes the point necessary for oscillation is called the phase margin. This depends on:

1. The proximity of the complex zero and pole pair, which is determined by the difference in value of the C_{TOP} and C_{BOT} combination relative to the external circuit capacitances. Large parasitics decrease the distance between the pole and zero pair, degrading the phase margin.
2. The circuit "Q," which is a function of the reflected crystal resistances. Applying the reflection algorithm described above shows that large capacitive parasitic values produce a lower "reflected" crystal resistance and thus a lower "Q." Additionally low values of parasitic loss resistances present at the oscillator pin will have the same effect.

If both of the above situations exist, the phase may not cross the zero point at all and oscillations will not start. The exact analysis procedure in Appendix A gives a quantitative description of this situation.

Nyquist Criterion:

In Figures 11C and 11D in Appendix A, note that the loop gain falls to below unity at the complex zero point. A situation can exist where perhaps, if g_M is large, the loop gain will remain above unity, even at the complex zero frequency. This represents a violation of the Nyquist criterion for oscillation in that the Nyquist plot never encircles the -1,0 point. This can happen with crystals at the lower frequencies around 1MHz or so. Appendix A gives a quantitative analysis of this situation, and reference [5] goes in detail for the theory.

Final Oscillation Amplitude:

This section outlines a qualitative explanation of the final oscillation amplitude. For a complete analysis, refer to references [1] and [4]. Appendix B gives an example analysis using the numerical methods of the MathCAD software on an IBM PC. If the user is interested at a higher level, please contact Micro Linear for design parameters.

Given the loop gain at the frequency where the phase shift crosses zero, oscillations start and then increase in amplitude. The waveforms across C_{TOP} and C_{BOT} in Figure 10 are close to sinusoidal due to the high Q of the circuit. The drain current of the MOS transistor, however, is a square law version of the gate to source voltage. Thus, at large signals, the effective g_M of the transistor is reduced by a factor which is related to only the first harmonic of the drain current. This is the only component which is fed back around the loop due to the high Q. For a given small signal loop gain, the amount of oscillation amplitude necessary to maintain the large signal loop gain at one will determine the final oscillation amplitude. For a typical Micro Linear design, this usually falls within the power supplies. Occasionally, then the loop gain is high, the amplitude may exceed the power supplies but is "clamped" by the input static protection diodes present on the oscillator pin. This forward biases the substrate, but the input protection structure of the oscillator pin prevents any harmful effects from this phenomenon.

3.0 Design Parameters

The following section outlines some parameters necessary to perform the hand calculation analysis described above and the exact analysis in Appendix A for various Micro Linear chips at the time of this application note:

ML2200, ML2208, ML2230, ML2233, ML2221:

These designs are restricted to 3–7MHz only; no frequency trim. Provide minimum parasitic from pin to ground possible.

C_{TOP} : 10pF
 C_{BOT} : 12pF
 Typical g_m : 500 μ A/V
 Typical R_{LT} : 8k Ω
 R_P at 3MHz: 240k Ω (see Appendix D)
 R_P at 7MHz: 140k Ω

ML2031, ML2032:

These designs are restricted to 3–15MHz only; frequency trimming with capacitor allowed if desired.

C_{TOP} : 16pF
 C_{BOT} : 16pF
 Typical g_m : 1.6mA/V
 Typical R_{LT} : 80k Ω
 R_P at 3MHz: 220k Ω (see Appendix D)
 R_P at 15MHz: 100k Ω

ML2035, ML2036:

These designs are restricted to 2–18MHz only; frequency trimming with capacitor allowed if desired.

C_{TOP} : 18pF
 C_{BOT} : 18pF
 Typical g_m : 1.6mA/V
 Typical R_{LT} : 80k Ω
 R_P at 2MHz: 290k Ω (see Appendix D)
 R_P at 15MHz: 100k Ω

In addition, the package pin capacitance is needed along with any stray capacitance due to the bond wire, etc. These values vary from device to device, but an approximate value would be about 1–3pF.

4.0 Crystal Specifications

For most situations, standard microprocessor type crystals will work fine in these circuits. If more precise frequency tolerance or unusual frequencies are desired, a special grind will have to be ordered from a crystal manufacturer.

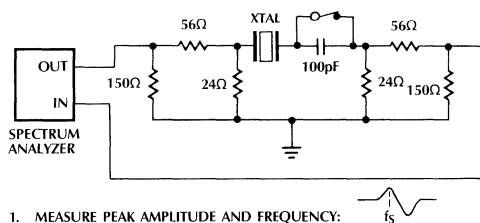
1. Calculate what capacitance will be seen by the crystal in your board, then specify this to the crystal manufacturer.
2. An approximation of the series resistance tolerable can be made using the above analysis or the exact analysis in Appendix A. One fact which is rarely known is that crystal resistances on startup can be much higher than when the crystal is being excited. Specify both a low level maximum series resistance and an operating level series resistance (e.g., 10nW to 1 μ W startup level and 1 μ W to 200 μ W operating level). An equation to calculate crystal dissipation is given in Appendix A.

3. Frequency tolerances of about .005% are common, tighter tolerances are available.
4. Frequency stability over temperature (0–70°C) of about .005% are common; special order for extended temperature range or tighter tolerance.
5. Frequency stability is typically dominated by the crystal itself. Temperature coefficients of the parasitic capacitances come into play and can be calculated using the equations described above. Variation of the oscillator g_m and internal capacitance values versus temperature has a very minor role in stability (1–5ppm or so over 0–70°C).

5.0 Board Level Design Verification

Some simple tests can be performed during the debugging process to verify that the crystal being used and the parasitics present are acceptable for manufacture:

1. Measure the crystal parameters. A procedure to do this is described below. This is a procedure described in reference [6].



1. MEASURE PEAK AMPLITUDE AND FREQUENCY: PEAK AMPLITUDE GIVES R_O VALUE.
 2. OPEN SWITCH, MEASURE PEAK FREQUENCY (THIS SHIFTS UP!)

$$\Delta f = \frac{1}{8\pi^2 f_s C_T L_O} \quad C_T = 100\text{pF} + C_C \text{ (CASE CAPACITANCE)}$$
 SOLVE FOR L_O .
 3. $f_s = \frac{1}{2\pi\sqrt{L_O C_O}}$, SOLVE FOR C_O
2. Observe crystal startup at the high temperature/low power supply specification of your system. Crystal startup is the most stringent test of the design. Often times the series resistance of the crystal at low levels is many times that at operation. Do this over a wide sample range of the intended crystal to be used.

Application Note 8

3. Be sure that the oscillator amplitude is at least 2 volts peak to peak at the high temperature/low supply case. This is to insure that the buffer that squares the sine wave up remains operational. If less than 2 volts, consider using a crystal with a lower series resistance or decrease parasitic capacitance on the oscillator pin. Do not use long lead lengths or traces from the oscillator pin to the crystal.
4. Observe crystal startup times. This is a good indication of available loop gain. Crystal startup time is a function of the real part of the closed loop transfer function. Appendix C provides a sample of how to calculate a root locus plot versus varying g_m 's.
5. When observing the oscillator pin, use a FET probe or use a standard probe in series with a 1pF capacitor to prevent loading the pin with excessive parasitic. When observing frequency stability, use a spectrum analyzer with an antenna wire pickup to minimize parasitic effects. Alternately, if a buffered output of the oscillator is available, measure the frequency at this point.

6.0 References

- [1] Clarke-Hess, "Communications Circuits: Analysis and Design," Addison-Wesley Publishing Company, 1971.
- [2] P.R. Gray and R.G. Meyer, "Analysis and Design of Analog Integrated Circuits," Second Edition, John Wiley and Sons, 1984.
- [3] J.T. Santos and R.G. Meyer, "A One Pin Crystal Oscillator for VLSI Circuits," *IEEE Journal of Solid State Circuits*, Vol SC-19 No. 2, April 1984.
- [4] R.G. Meyer and D.C.F. Soo, "MOS Crystal Oscillator Design," *IEEE Journal of Solid State Circuits*, Vol SC-15, No. 2, April 1980.
- [5] M.A. Unkrich and R.G. Meyer, "Conditions for Start-Up in Crystal Oscillators," *IEEE Journal of Solid State Circuits*, Vol SC-17, No. 1, February, 1982.
- [6] N.J. Watson, "Crystal Testing Using Spectrum Analyser TF 2370," Application Note No. 14, Marconi Instruments.

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APPENDIX A

ONE PIN OSCILLATOR DESIGN

This document is an exact small signal analysis of the one pin oscillator using the MathCAD software package.

*****define units first:

$$\begin{aligned} \text{rad} &\equiv 1 \\ \text{deg} &\equiv \pi \cdot \frac{\text{rad}}{180} \\ \text{TOL} &\equiv 10^{-6} \quad \text{sets MathCAD tolerance} \end{aligned}$$

*****Define Parameters:

$$\begin{aligned} \text{Ctop} &\equiv 16 \cdot 10^{-12} & \text{Cbot} &\equiv 16 \cdot 10^{-12} & \text{Cc} &\equiv 5 \cdot 10^{-12} & (\text{xtal case cap}) \end{aligned}$$

$$\begin{aligned} \text{Ro} &\equiv 15 & \text{Lo} &\equiv 8.005813989 \cdot 10^{-3} & \text{Co} &\equiv 20.7558457 \cdot 10^{-15} \end{aligned}$$

$$\text{****Serial Resonant Frequency } \omega_s := \frac{1}{\sqrt{\text{Lo} \cdot \text{Co}}} \quad \omega_s = 1.2346594 \cdot 10^7 \text{ } 2 \cdot \pi$$

$$\text{****Crystal Q: } \text{Qxtal} := \frac{1}{\omega_s \cdot \text{Ro} \cdot \text{Co}} \quad \text{Qxtal} = 4.14 \cdot 10^4$$

****CAPACITOR DIVIDER:

$$\text{****Equivalent C of Capacitor Divider: } \text{Ceq} := \text{Ctop} \cdot \frac{\text{Cbot}}{\text{Ctop} + \text{Cbot}}$$

$$\text{Ceq} = 8 \cdot 10^{-12}$$

$$\text{****Transformer Ratio for Capacitive Divider: } n := \frac{\text{Ctop}}{\text{Ctop} + \text{Cbot}}$$

$$n = 0.5$$

****Impedance Reflection Ratio for Capacitive Divider:

$$\text{zc} := n^2 \quad \text{zc} = 0.25$$

Application Note 8

****Transistor Characteristics:

$$g_m := 1.6 \cdot 10^{-3} \quad R_{lt} := 80 \cdot 10^3$$

****Pin Characteristics:

**External pin parasitics--this is the real and imaginary parts of any external parasitic at the oscillation frequency):

$$R_{ext} := 100 \cdot 10^3 \quad C_{ext} := 5 \cdot 10^{-12}$$

****Total Parallel Capacitance:

$$C_{ptot} := C_c + C_{ext} + C_{eq}$$

$$C_{ptot} = 1.8 \cdot 10^{-11} \quad C_{pt} := C_{ptot} - C_{eq}$$

$$C_{pt} = 1 \cdot 10^{-11}$$

****Equivalent Capacitance for Pole Pair:

$$C_{pol} := C_o \cdot \frac{C_{ptot}}{C_o + C_{ptot}} \quad C_{pol} = 2.073 \cdot 10^{-14}$$

****Approximate Frequency of Oscillation:

$$W_{pol} := \frac{1}{\sqrt{L_o \cdot C_{pol}}} \quad F_{pol} := \frac{W_{pol}}{2 \cdot \pi} \quad F_{pol} = 1.235371 \cdot 10^7$$

****Ratio of the Capacitive Divider Between Crystal and Oscillator:

$$n_x := \frac{C_o}{C_o + C_{ptot}} \quad n_x = 0.001$$

****Impedance Reflection Ratio:

$$z_x := n_x^2 \quad z_x = 1.327 \cdot 10^{-6}$$

$$****Q with Cload: \quad Q := \frac{W_{pol} \cdot L_o}{R_o} \quad Q = 4.143 \cdot 10^4$$

****Equivalent Parallel R for Crystal Only:

$$R_{px} := Q^2 \cdot R_o \quad R_{px} = 2.574 \cdot 10^{10}$$

****Reflected Value of the Crystal Resistance at the Oscillator:

$$R_{xtal} := R_{px} \cdot z_x \quad R_{xtal} = 3.415 \cdot 10^4$$

Total parallel resistance at the oscillator pin:

$$G_{sum} := \frac{1}{R_{ext}} + \frac{1}{R_{xtal}} \quad R_{sum} := \frac{1}{G_{sum}} \quad R_{sum} = 2.546 \cdot 10^4$$

****Reflected Parallel R at Oscillator to Transistor Source:

$$R_{ref} := R_{sum} \cdot z_c \quad R_{ref} = 6.364 \cdot 10^3$$

****Total Load Seen at Transistor Source:

$$R_l := R_{ref} \cdot \frac{R_{lt}}{R_{ref} + R_{lt}} \quad R_l = 5.895 \cdot 10^3$$

****Loop Gain Calculated with Reflection Method:

$$A_l := g_m \cdot R_l \cdot \frac{C_{bot}}{C_{top}} \quad A_l = 9.433$$

**in dB: $20 \cdot \log(A_l) = 19.493$

*****EXACT CALCULATIONS:*****

**Polynomial Coefficients for Zeros:

$$n_0 := 1$$

$$n_1 := (C_o \cdot R_o + C_o \cdot R_{ext} + C_{pt} \cdot R_{ext})$$

$$n_2 := (C_o \cdot L_o + C_o \cdot C_{pt} \cdot R_o \cdot R_{ext})$$

$$n_3 := (C_o \cdot C_{pt} \cdot L_o \cdot R_{ext})$$

guess first zero location:

$$\omega := \frac{-1}{R_{ext} \cdot (C_{pt} + C_o)} \quad \omega = -9.9792871 \cdot 10^5$$

$$s := \omega$$

$$z_1 := \text{root} \left[n_3 \cdot s^3 + n_2 \cdot s^2 + n_1 \cdot s + n_0, s \right]$$

$$z_1 = -9.9792906 \cdot 10^5$$

Application Note 8

guess complex zero pair location:

$$\omega := \sqrt{\frac{1}{\text{Lo} \cdot \frac{\text{Co} \cdot \text{Cpt}}{\text{Co} + \text{Cpt}}}} \quad \omega = 7.7656404 \cdot 10^7$$

s := $\omega \cdot i$

$$z2 := \text{root} \left[n3 \cdot s^3 + n2 \cdot s^2 + n1 \cdot s + n0, s \right]$$

$$z2 = -1.9722866 \cdot 10^3 + 7.7656391 \cdot 10^7 i$$

***** Zeros:

$$z1 = -9.9792906 \cdot 10^5$$

$$z2 = -1.9722866 \cdot 10^3 + 7.7656391 \cdot 10^7 i$$

*****Poles:

**Coefficients for Poles:

$$d0 := 1$$

$$d1 := (\text{Co} \cdot \text{Ro} + \text{Co} \cdot \text{Rext} + \text{Ctop} \cdot \text{Rlt} + \text{Ctop} \cdot \text{Rext} + \text{Cbot} \cdot \text{Rlt} + \text{Cpt} \cdot \text{Rext})$$

$$d2 := \left[\begin{array}{l} \text{Co} \cdot \text{Lo} + \text{Co} \cdot \text{Ctop} \cdot \text{Ro} \cdot \text{Rlt} + \text{Co} \cdot \text{Ctop} \cdot \text{Ro} \cdot \text{Rext} + \text{Co} \cdot \text{Ctop} \cdot \text{Rlt} \cdot \text{Rext} \dots \\ + \text{Co} \cdot \text{Cbot} \cdot \text{Ro} \cdot \text{Rlt} + \text{Co} \cdot \text{Cbot} \cdot \text{Rlt} \cdot \text{Rext} + \text{Co} \cdot \text{Cpt} \cdot \text{Ro} \cdot \text{Rext} \dots \\ + \text{Ctop} \cdot \text{Cbot} \cdot \text{Rlt} \cdot \text{Rext} + \text{Ctop} \cdot \text{Cpt} \cdot \text{Rlt} \cdot \text{Rext} + \text{Cbot} \cdot \text{Cpt} \cdot \text{Rlt} \cdot \text{Rext} \end{array} \right]$$

$$d3 := \text{Co} \cdot \text{Ctop} \cdot \text{Lo} \cdot \text{Rlt} + \text{Co} \cdot \text{Ctop} \cdot \text{Lo} \cdot \text{Rext} + \text{Co} \cdot \text{Cbot} \cdot \text{Lo} \cdot \text{Rlt} + \text{Co} \cdot \text{Cpt} \cdot \text{Lo} \cdot \text{Rext} \dots \\ + \text{Co} \cdot \text{Ctop} \cdot \text{Cbot} \cdot \text{Ro} \cdot \text{Rlt} \cdot \text{Rext} + \text{Co} \cdot \text{Ctop} \cdot \text{Cpt} \cdot \text{Ro} \cdot \text{Rlt} \cdot \text{Rext} \dots \\ + \text{Co} \cdot \text{Cbot} \cdot \text{Cpt} \cdot \text{Ro} \cdot \text{Rlt} \cdot \text{Rext}$$

$$d4 := \text{Co} \cdot \text{Ctop} \cdot \text{Cbot} \cdot \text{Lo} \cdot \text{Rlt} \cdot \text{Rext} + \text{Co} \cdot \text{Ctop} \cdot \text{Cpt} \cdot \text{Lo} \cdot \text{Rlt} \cdot \text{Rext} \dots \\ + \text{Co} \cdot \text{Cbot} \cdot \text{Cpt} \cdot \text{Lo} \cdot \text{Rlt} \cdot \text{Rext}$$

guess first pole location:

$$\omega := \frac{-1}{(\text{Rext}) \cdot (\text{Ctop} + \text{Cbot} + \text{Cpt})} \quad \omega = -2.3809524 \cdot 10^5$$

s := ω

$$p1 := \text{root} \left[d4 \cdot s^4 + d3 \cdot s^3 + d2 \cdot s^2 + d1 \cdot s + d0, s \right]$$

$$p1 = -2.492 \cdot 10^5$$

guess second pole location (this is dominant pole)

$$\omega := \frac{-1}{\text{Rlt} \cdot \left[\text{Cbot} + \frac{\text{Ctop} \cdot \text{Cpt}}{\text{Ctop} + \text{Cpt}} \right]} \quad \omega = -5.6423611 \cdot 10^5$$

s := ω

$$p2 := \text{root} \left[d4 \cdot s^4 + d3 \cdot s^3 + d2 \cdot s^2 + d1 \cdot s + d0, s \right]$$

$$p2 = -8.6971683 \cdot 10^5$$

guess complex pole pair location:

$$\omega := \frac{-1}{\sqrt{\text{Lo} \cdot \frac{\text{Co} \cdot (\text{Cpt} + \text{Ceq})}{\text{Co} + \text{Cpt} + \text{Ceq}}}} \quad \omega = -7.7620652 \cdot 10^7$$

s := i · ω

$$p3 := \text{root} \left[d4 \cdot s^4 + d3 \cdot s^3 + d2 \cdot s^2 + d1 \cdot s + d0, s \right]$$

$$p3 = -1.3566881 \cdot 10^3 - 7.7620647 \cdot 10^7 i$$

*****Summary:*****

$$tp := 2 \cdot \pi$$

$$z1 = -9.9792906 \cdot 10^5$$

$$\frac{z1}{tp} = -1.5882534 \cdot 10^5$$

$$z2 = -1.9722866 \cdot 10^3 + 7.7656391 \cdot 10^7 i$$

$$\frac{z2}{tp} = -313.899161 + 1.2359398 \cdot 10^7 i$$

1

Application Note 8

$$p1 = -2.492351 \cdot 10^5$$

$$p2 = -8.6971683 \cdot 10^5$$

$$p3 = -1.3566881 \cdot 10^3 - 7.7620647 \cdot 10^7 i$$

$$\frac{p1}{tp} = -3.9666998 \cdot 10^4$$

$$\frac{p2}{tp} = -1.3841973 \cdot 10^5$$

$$\frac{p3}{tp} = -215.9236169 - 1.235371 \cdot 10^7 i$$

*****System Function:*****

$$H(s) := \frac{gm \cdot Rlt \cdot [n3 \cdot s^3 + n2 \cdot s^2 + n1 \cdot s + n0]}{d4 \cdot s^4 + d3 \cdot s^3 + d2 \cdot s^2 + d1 \cdot s + d0}$$

*****Plot System Function*****

$$x := 1 \dots 80$$

$$l := \frac{x}{10}$$

$$\omega_x := \left[\frac{1}{tp \cdot 10^x} \right]$$

$$20 \cdot \log \left[\left| \left[H \left[i \cdot \omega_x \right] \right] \right| \right]$$

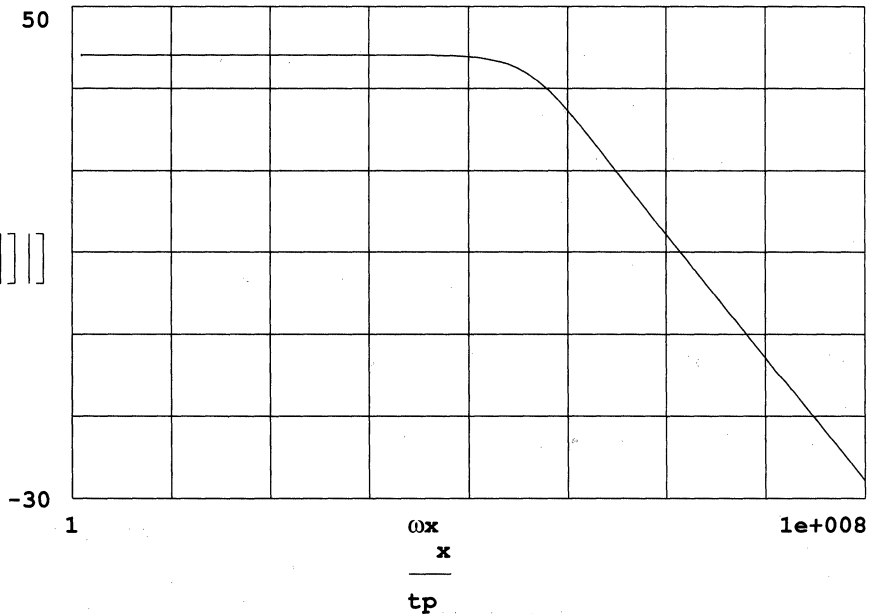


Figure 11A

plot phase:

$$\text{Ph} := \left[\frac{\arg(H(i \cdot \omega x)) \cdot 180}{\pi} \right]$$

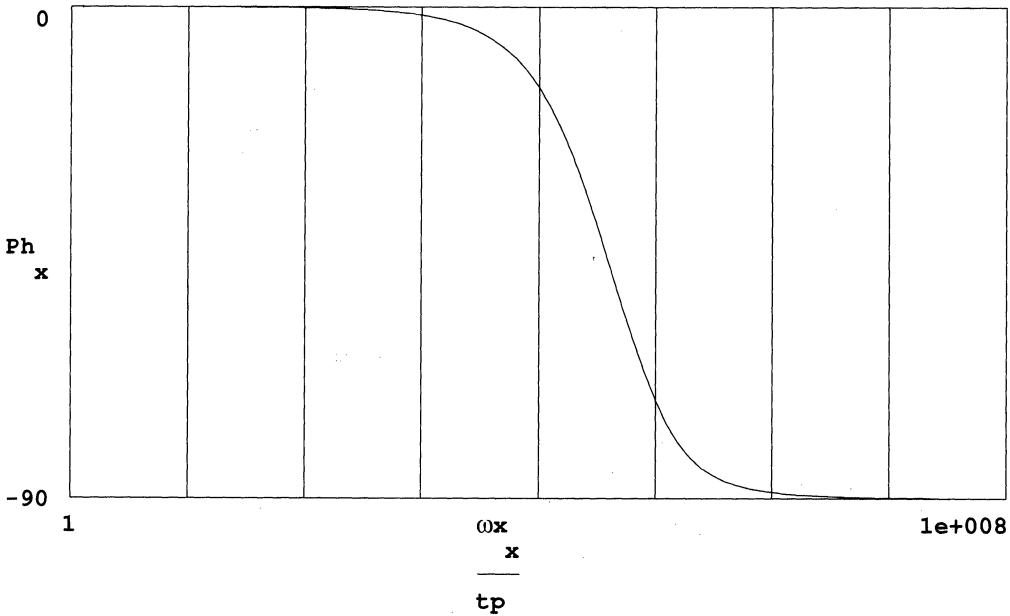


Figure 11B

plot an expanded area around the complex pole/zero pair:

$n := 0 \dots 100$

$\text{range} := (|\text{Im}(z2)| - |\text{Im}(p3)|)$ $\text{range} := \text{range} \cdot 2$

$\text{lo} := |\text{Im}(p3)| - \text{range}$

$\text{hi} := |\text{Im}(z2)| + \text{range}$

$\omega x_n := \left[(\text{hi} - \text{lo}) \cdot \frac{n}{100} + \text{lo} \right]$

1

Application Note 8

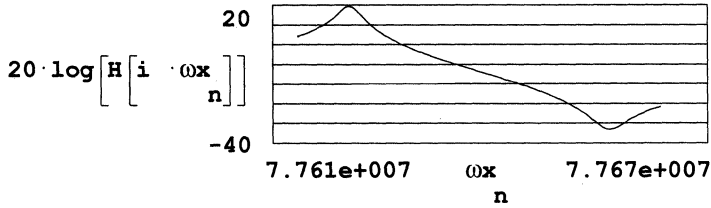


Figure 11C

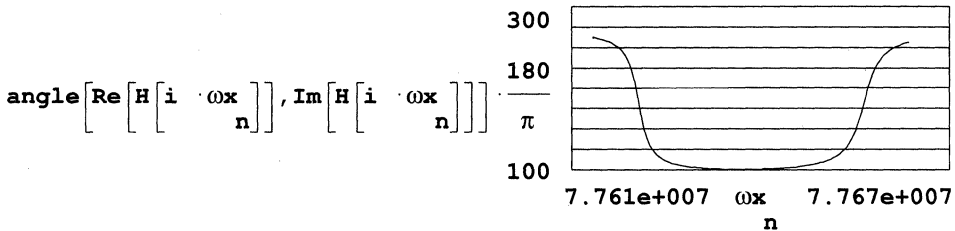


Figure 11D

Nyquist plot:

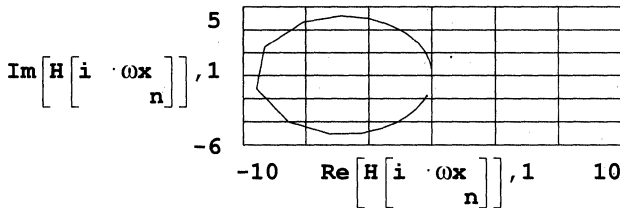


Figure 11E

$$\omega_g := i \cdot (\text{Im}(p_3) + 10) \quad \frac{\omega_g}{\text{tp}} = -1.2353708 \cdot 10^7 i$$

Find exact frequency of phase zero crossing due to complex pole pair:

$$\omega_{\text{osc}} := \text{root}(\text{Im}(H(\omega_g)), \omega_g)$$

check:
 $\arg(H(\omega_{\text{osc}})) = 180 \cdot \text{deg}$

Exact zero phase frequency and amplitude:

$$\frac{\omega_{\text{osc}}}{\text{tp}} = -1.2353722 \cdot 10^7 i \quad |H(\omega_{\text{osc}})| = 9.41$$

Find exact zero phase and amplitude due to complex zero pair:

guess first:

$$\omega_g := i \cdot (\text{Im}(z_2) - 100) \quad \frac{\omega_g}{\text{tp}} = 1.2359383 \cdot 10^7 \text{ i}$$

$$\omega_{\text{zer}} := \text{root}(\text{Im}(H(\omega_g)), \omega_g)$$

check:

$$\arg(H(\omega_{\text{zer}})) = -179.999 \cdot \text{deg} \quad |H(\omega_{\text{zer}})| = 0.02 \quad \text{amplitude}$$

$$\frac{\omega_{\text{zer}}}{\text{tp}} = 1.2359386 \cdot 10^7 \text{ i} \quad \text{frequency}$$

Find phase margin for oscillator (maximum phase between complex pole and zero pair):

$$\omega_x := - \left[(|\omega_{\text{zer}}| - |\omega_{\text{osc}}|) \cdot \frac{n}{105} + |\omega_{\text{osc}}| + 100 \right]$$

$$\pi - \max \left[\text{angle}(\text{Re}(H(i \cdot \omega_x)), \text{Im}(H(i \cdot \omega_x))) \right] = -79.361 \cdot \text{deg}$$

Application Note 8

APPENDIX B

This goes through an example calculation of the reduction in small signal gm due to a certain amplitude across Ctop or the gate of the MOS device. The analysis can be carried out in a more general manner and graphs can be plotted out for the purpose of providing a graphical solution to ascertain the final oscillation amplitude given an initial set of bias conditions. The procedure is for example purposes only. If the reader requires more specific information, please contact Micro Linear directly.

This analysis was carried out in MathCAD.

define some numbers:

MOS transistor threshold voltage: $v_t := .926$

MOS transistor k factor: $\beta := \frac{700}{4.2} \cdot 10^{-6}$

$\omega := 1$ here, frequency doesn't matter, so make it 1

$v_{bq} := 1.158$ dc bias value of gate to source

simple MOS equation for drain current

$id(v_{gs}) := \text{if} \left[v_{gs} > v_t, \left[\frac{1}{2} \cdot \beta \cdot (v_{gs} - v_t)^2 \right], 0 \right]$

$idq := id(v_{bq})$ calculate dc bias current

$idq := 2.112 \cdot 10^{-4}$

$v_{in}(t, v_b) := v_b + a \cdot \cos(\omega \cdot t)$ define the gate to source excitation

$gm_q := \sqrt{2 \cdot \beta \cdot id(v_{bq})}$ small signal gm at dc bias

$gm_q = 0.0018173$

Find the average bias voltage of vgs which MUST equal the dc bias current:

$TOL := 10^{-8}$ guess $v_b := v_{bq} - .2$

Given

$\frac{1}{2 \cdot \pi} \int_0^{2 \cdot \pi} id(v_{in}(t, v_b)) dt \approx idq$

$newvb := \text{Find}(v_b)$

$newvb = 0.898$ $v_{bq} = 1.158$ compare to new bias

check:

$$\left[\frac{1}{2 \cdot \pi} \int_0^{2 \cdot \pi} id(vin(t, newvb)) dt \right] = 2.112 \cdot 10^{-4}$$

This represents a shift of:

$$vbq - newvb = 0.26 \quad \text{volts in vgs bias}$$

for a sine wave amplitude of:

$$a \equiv .5 \quad \text{volts peak of vgs (note: this is NOT the output voltage!)}$$

This gives a steady state vgs of:

$$vg(t) := vin(t, newvb)$$

The first harmonic of the drain current is:

$$id1 := \frac{1}{\pi} \int_0^{2 \cdot \pi} id(vg(t)) \cdot \cos(\omega \cdot t) dt$$

$$id1 = 3.624 \cdot 10^{-4} \quad \text{The large signal gm, or GM is now calculated:}$$

$$GM := \frac{id1}{a} \quad GM = 7.247 \cdot 10^{-4}$$

The normalization of GM/gm is now shown:

$$\frac{GM}{gmq} = 0.399 \quad \text{This shows a reduction of the small signal gm.}$$

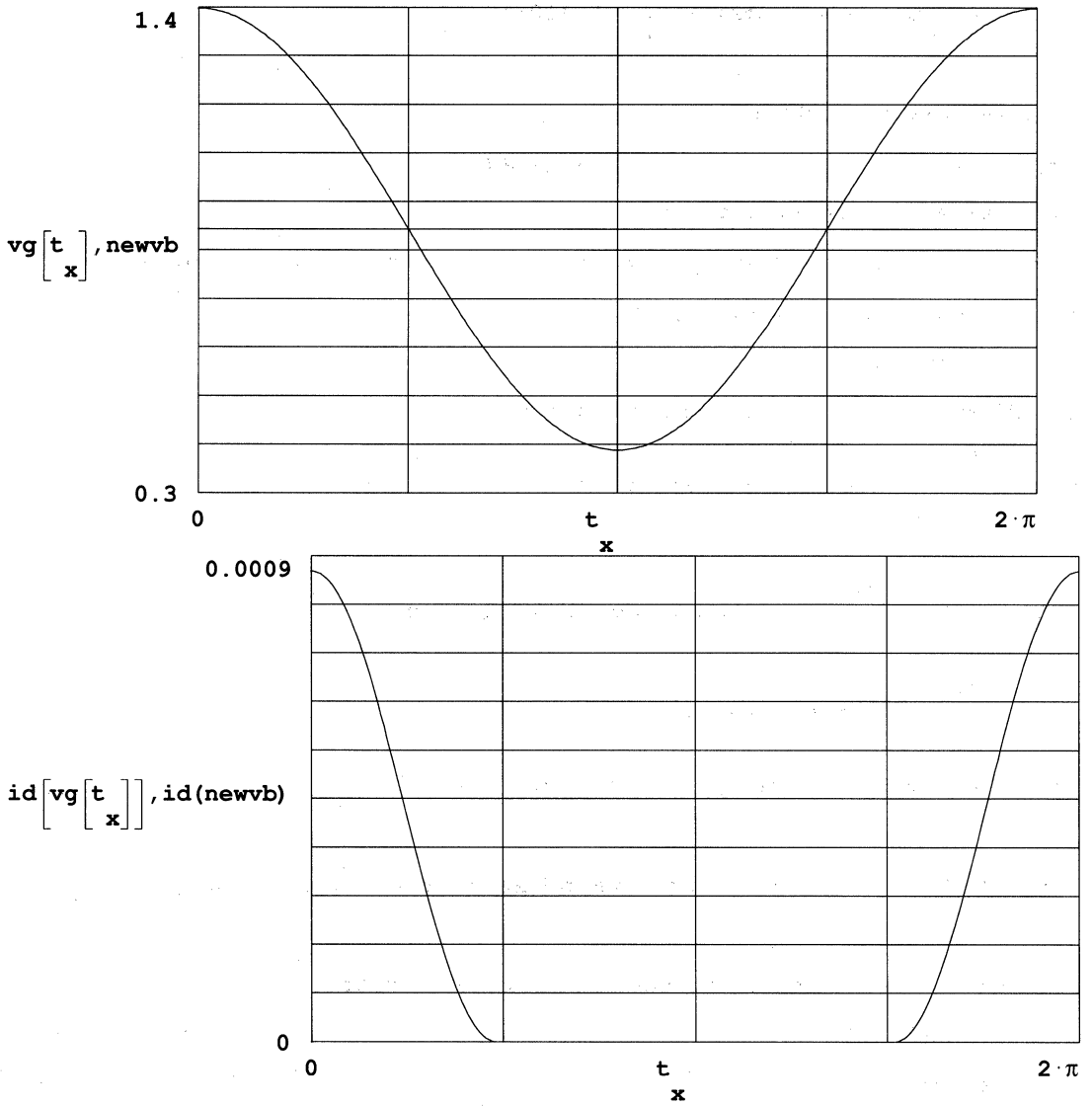
Plot gate voltage and drain current in steady state:

$$x := 0 .. 255$$

$$t := \frac{2 \cdot \pi \cdot x}{255}$$

10

Application Note 8



APPENDIX C

Oscillator Root Locus Calculation

This calculates the real and imaginary parts of the closed loop transfer function. The real part is the time constant of the initial exponential startup transient.

define a range of gm values to calculate over:

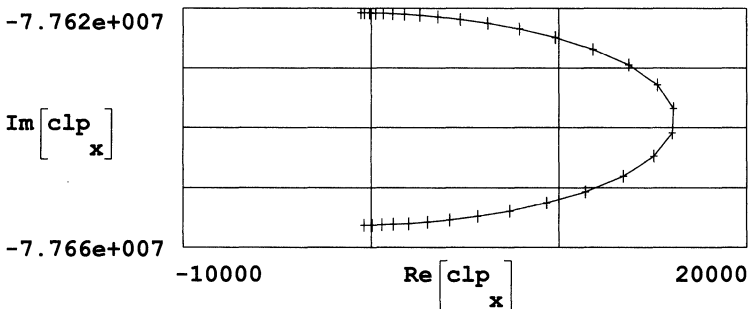
x := 40 ..10

$$gm_x := 10 \cdot \left[\frac{x}{10} \right]^{-6} \quad TOL \equiv 10^{-6}$$

Calculate root locus for gm. These are the roots of the characteristic equation of the closed loop transfer function. See Appendix A for the definitions of the coefficients shown below:

$$clp_x := \text{root} \left[\begin{array}{l} d4 \cdot s^4 + \left[d3 + n3 \cdot gm_x \cdot rmb \right] \cdot s^3 + \left[d2 + n2 \cdot gm_x \cdot rmb \right] \cdot s^2 + \left[d1 + n1 \cdot gm_x \cdot rmb \right] \cdot s + d0 + n0 \cdot gm_x \cdot rmb \end{array} , s \right]$$

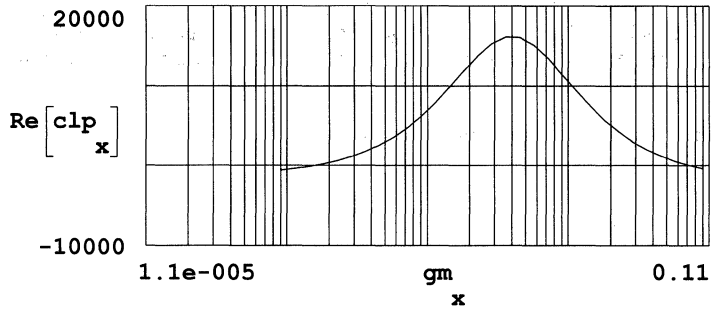
Root Locus Plot:



10

Application Note 8

Real part of closed loop transfer function versus gm:



Note in the above plot that there is a limited range of g_m 's for which the poles of the closed loop transfer function remain in the right half plane. In other words, too low a g_m creates too low a gain. However, too high g_m values also violate the Nyquist Criterion.

APPENDIX D

This shows how to calculate the real part of the on chip bias resistor Rp for inclusion in the loop gain calculations. This was done on MathCAD.

This is a calculation of the Real part of a distributed RC network with one end shorted to ground. A parallel R part is calculated.

ORIGIN := 1 m := 1 ..3

Following DC resistance values are for the ML2031, ML2035, and ML2200. The total capacitance is also calculated for the network.

$$r0 := \begin{bmatrix} 240 \cdot 10^3 \\ 640 \cdot 10^3 \\ 6 \\ 10 \end{bmatrix} \quad c0_m := r0_m \cdot 100 \cdot .08 \cdot \frac{10^{-15}}{4 \cdot 10^3}$$

Define a range of frequencies to use: n := 1 ..40

$$f_n := n \cdot \frac{10^6}{2} \quad \omega := (f \cdot 2 \cdot \pi)$$

Calculate the admittance of the distributed network:

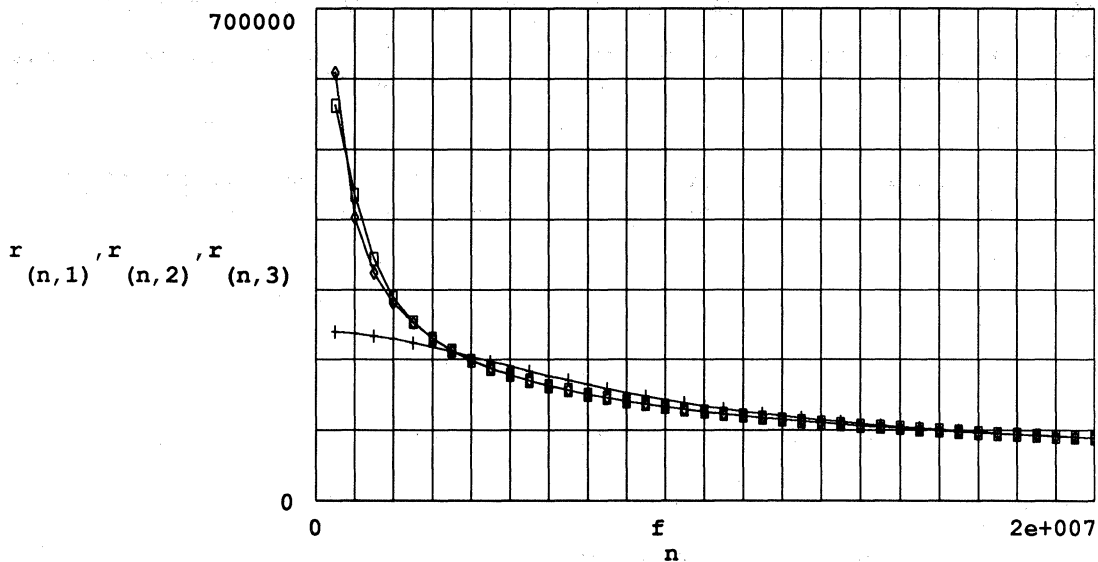
$$\Gamma_{n,m} := \sqrt{\frac{i \cdot \omega_n \cdot c0_m \cdot r0_m}{n \cdot m \cdot m}} \quad Z0_{n,m} := \frac{r0_m}{\sqrt{\frac{i \cdot \omega_n \cdot c0_m}{n \cdot m}}} \quad y1 := \frac{1}{Z0 \cdot \tanh(\Gamma)}$$

Extract the real part and invert to get resistance:

$$r := \frac{1}{\text{Re}(y1)}$$

Application Note 8

Plot the real part versus frequency:



diamonds=ML2200, boxes=ML2035, pluses=ML2031

An Improved Method of Load Fault Detection

High frequency supply designs pose unique problems in fault detection. A typical method of output fault detection in most standard controllers is to provide a cycle-by-cycle current limit (V_{TH1}) to limit the peak current in the output switch. In addition, these controllers have a second current limit (V_{TH2}), which is typically set 40% higher than the cycle by cycle limit. Crossing V_{TH2} on the I_{SENSE} input resets the Soft Start circuit and allows current in the output to decay before re-starting the system.

In theory, by the time the power output stage can begin to turn off from having crossed V_{TH1} , the output current will have exceeded V_{TH2} and a soft start reset will be performed. This technique works well if leakage inductance is low and turn-off delay is high enough to cause enough energy to transfer to the output inductor, causing the current to build up in subsequent cycles (figure 2a). This current build up takes place when the output is short circuited because the output inductor has almost no voltage across it and therefore a very shallow discharge slope. If, however, energy transfer is low due to fast turn-off of the outputs (which is desirable to minimize switching losses) energy transfer to the output inductor will be minimized, resulting in the supply continually running at the cycle by cycle current limit to a short circuit with no reset occurring (figure 2b).

High frequency controllers are designed to minimize T_{PD} and turn off the output MOSFET gates quickly. This implies that the event which triggers soft start reset will not persist for very long if it is detected at all. The short persistence of the triggering event requires that Q1 discharge C1 in a very short time, typically resulting in a partial discharge and an inadequate reset. A solution to this problem (figure 3) is implemented in all of Micro Linear's PWM IC's. Flip-flop (F2) and comparator (A4) are added to the circuit, to ensure a full reset. If desired, a delay (as implemented in the ML4809 and ML4811) can also be added before restart, which lowers the system's effective duty cycle allowing the supply to cool down.

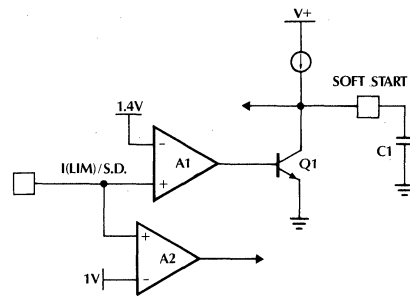


Figure 1. Typical Two Threshold Current Limit/Fault Detect

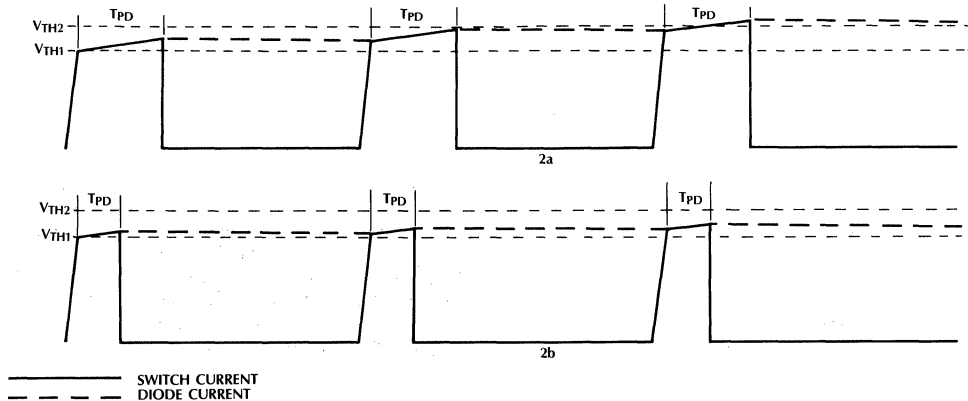


Figure 2. Current Waveforms During Output Short Circuit Slow Output Turn-Off (2a) and Fast Turn-Off (2b)

Application Note 10

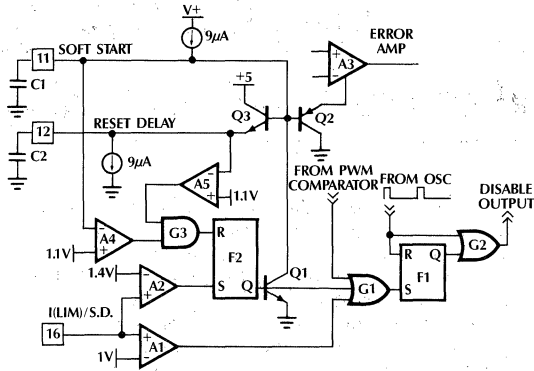


Figure 3. Improved Soft Start Reset with Delay — ML4809

Integrating Fault Detection

The “two threshold” detection technique described above limits the system designer’s freedom to optimize his magnetics and minimize switching time. Since detecting the fault relies on building inductor current up on successive cycles, propagation delay cannot be minimized (as shown in figure 2) for this technique to work. Since these two parameters are important terms in high frequency supply efficiency, the need to compromise due to inadequacies in fault detection presents a problem.

A method of circumventing this problem involves “counting” the number of times the controller terminates the PWM cycle due to the cycle by cycle current limit.

Figure 4 shows the Integrating Fault Detect circuit. When the I_{LIM} signal (switch current) crosses the 1.1V threshold A1 signals the F1 to terminate the cycle and sets F3, which is reset at the beginning of each PWM cycle. The output of F3 turns on a current source to charge C2. When, after several cycles, C2 has charged to 3V, A5 turns on F2 to discharge soft start capacitor C1. Charge is continually bled from C2 by R1. If a current surge is short lived (for instance a disk drive start-up or a board being plugged in to a live rack) the control can “ride out” the surge (figure 5a) with the switch protected by the cycle by cycle limit. R1 and C1 can be selected to track diode heating, or to ride out various system surge requirements as required.

If the high current demanded is caused by a short circuit (figure 5b), the duty cycle will be short and the output diodes will carry the current for the majority of the PWM cycle. C2 charges fastest for low duty cycles (since F3 will be on for a longer time) providing for quicker shutdown during short-circuit when the output diodes are being maximally stressed.

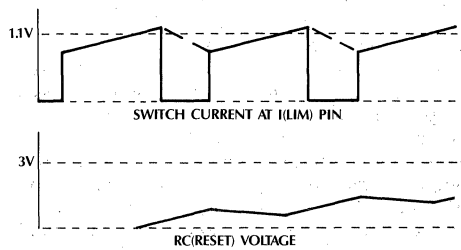


Figure 5a. Integrating Fault Detection Response to Load Surge

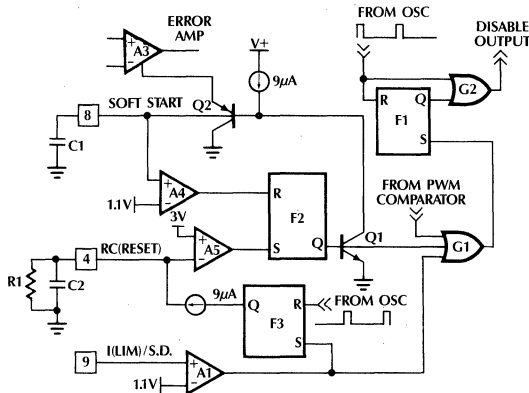


Figure 4. Integrating Soft Start Reset Circuit with Delay (ML4811)

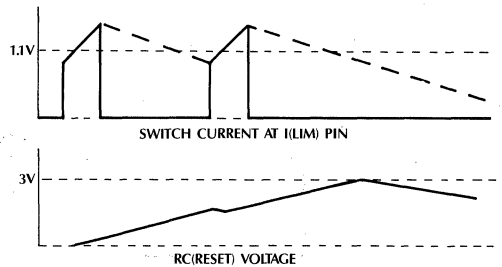


Figure 5b. Response to Load Fault (Short Circuit)

The Integrating Fault Detection circuit allows reliable detection of output faults independent of supply magnetics and propagation delays. Additionally, this method of fault detection is inherently noise immune, programmable, and can distinguish between load surges and load faults (short circuits).

Mehmet K. Nalbant

Power Factor Enhancement Circuit

A simple enhancement circuit for the ML4812 is described. The circuit which will be called the power factor enhancement circuit greatly improves the power factor while reducing the total harmonic distortion of the current waveform.

The circuit details for implementing the power factor enhancement circuit are given below. Figure 1 shows the schematic diagram of the circuit. The circuit operates by generating a small DC current bias and injecting it into the I_{SINE} (pin 6) input of the ML4812. This current injection has the net effect of improving the zero current crossover distortion. It does this by lifting the shoulders of the current waveform around the zero crossover areas.

The circuit in the dotted lines in Figure 2, shows the details of the implementation. The circuit automatically adjusts the amount of the injected DC bias as a function of the line voltage. The reason behind the variable amount of DC current injection is that at lower input voltages, the amount of DC bias that is required is less.

Based on experience, the amount of bias required at 220 VAC is approximately four times higher than at 120 VAC. The proper scaling can be adjusted by choosing appropriate values for the various resistors used and the zener diode voltage. The amount of bias that is required is a function of the boost inductor value and the ramp compensation. For best performance the value of the inductor should be chosen as high as possible which in turn will necessitate a small amount of ramp compensation.

One way to find the required bias currents is summarized below: the first step is to find the optimum bias at the nominal operating point, for example, at 120 VAC. This is done by connecting a variable resistor to the reference output of the IC. The initial value of the resistor is selected such that, the bias current equals the peak to peak ramp compensation voltage when the duty cycle is at its maximum. After the optimum value at the nominal operating conditions is found the input voltage is increased to 220 VAC and the same procedure above is repeated to find the optimum value of the resistor at the 220 VAC nominal operating conditions. The bias currents corresponding to the two resistor values above can be used to calculate the values of the components in the enhancement circuit. The formulas for calculating the various components are given below:

$$V_{C3(VIN)} = \frac{0.9V_{IN(RMS)} R6}{R1 + R2 + R6} \quad (1)$$

$$I_{SINE(VIN)} = \frac{V_{C3} - V_{BE} - V_Z - V_{ISINE}}{R7} \quad (2)$$

$$r = \frac{I_{SINE(220 VAC)}}{I_{SINE(120 VAC)}} \quad (3)$$

$$r = \frac{V_{C3(220 VAC)} - V_{BE} - V_Z - V_{ISINE}}{V_{C3(120 VAC)} - V_{BE} - V_Z - V_{ISINE}} \quad (4)$$

Where:

$I_{SINE(VIN)}$ = Bias current into the I_{SINE} input as function of the input voltage.

V_{BE} = Base emitter voltage of Q3 (0.7V nominal).

V_{ISINE} = Voltage at I_{SINE} input, typically 0.7V.

r = Ratio of bias current at 220 VAC input to the bias current at 120 VAC input.

By choosing a value for $V_{C3(220 VAC)}$ the value of $V_{C3(120 VAC)}$ is also found. These two values can be substituted to the equation above to calculate the required value for V_Z . The value of $R7$ can be found by using (2). The values of the remaining components can be calculated by using (1).

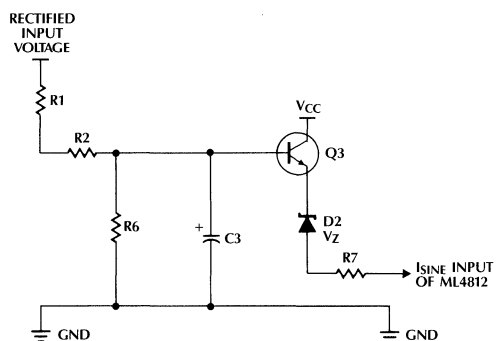


Figure 1. The Enhancement Circuit

Application Note 11

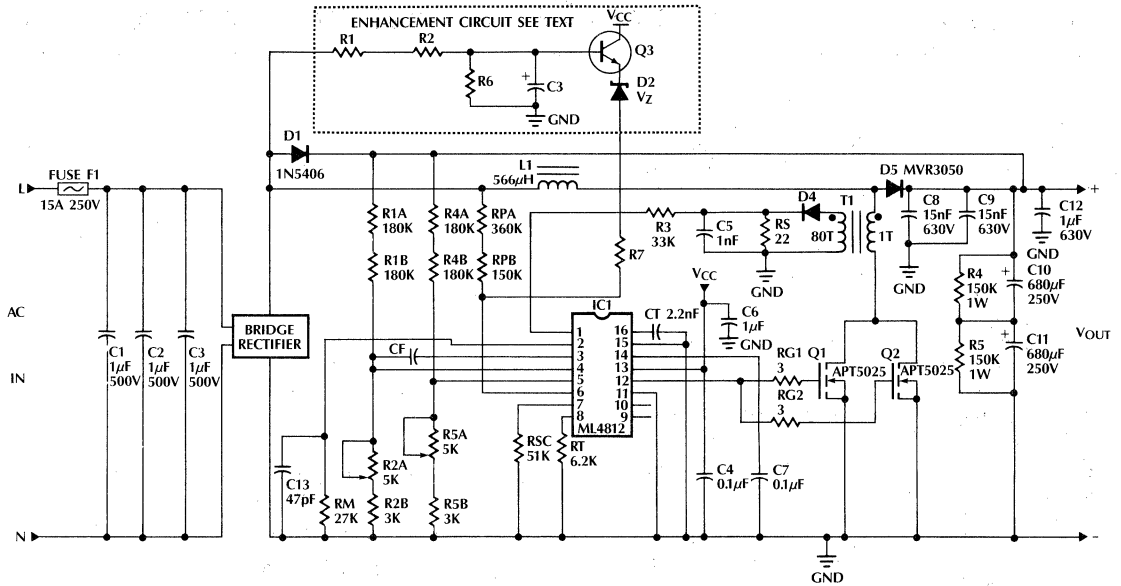


Figure 2. 1KW Enhanced Power Factor Correction Circuit

The circuit of Figure 2, is a 1KW input, power factor regulator. For this circuit the values of the enhancement circuit components were as follows:

- R1 + R2 = 330K
- R6 = 22K
- R7 = 22K
- D2, Vz = 3.5V
- Q3 = 2N2222 or any equivalent small signal transistor.
- C3 = 10µF electrolytic cap

Table 1 shows the performance of the power factor regulator with and without the enhancement circuit.

Table 1. Effect of Enhancement Circuit on Power Factor

Input Voltage (VAC)	Input Power (W)	Power Factor	
		With Enhancement	Without Enhancement
120	742	.998	.991
	365	.994	.976
220	706	.996	.976
	352	.969	.940

Generating Phase Controlled Sinewaves with the ML2036

Vince Cardinale

Introduction

The 16-bit resolution of the ML2036 combined with its Inhibit feature makes it a powerful tool for generating precision sinewaves. It can produce frequencies from DC to 50kHz in 1Hz increments with -40dB harmonic distortion *and* has the control to stop the output at any given time or at the next zero crossing, *with no external components*.

Precise phase control can also be obtained with the addition of a few external devices. With the addition of phase control two or more ML2036 sinewave generators can be synchronized at any angle from 0 to 360 with better than 1 degree resolution.

Inhibit Mode

In order to place the ML2036 in Inhibit mode three conditions must occur simultaneously. The three-level P_{DN}-INH input pin must be at the V_{SS} voltage (-5V), the shift register must be loaded with all zeros, and the LATI pin must be a logic "1" (+5V). Once these three conditions are met the output continues to operate until it reaches V_{OS} + |V_X| if the next zero crossing is positive going, or V_{OS} - |V_X| if the next zero crossing is negative going, and then holds this level (see figure 1). The output will stay at this voltage until a new frequency is loaded into the data latch, at which point the output will continue where it left off. If the output stopped at zero after approaching from below 0V then it will start-up going positive. If it stopped after approaching from above 0V then it will start-up going negative.

Initialization

In order to synchronize the ML2036 you must first initialize it so it will start up at a known point in the sinewave. By using the Inhibit mode you can stop the part at 0V but you can't be sure from which direction it approached zero, or more importantly which direction it will start-up. If you can guarantee that it stopped while approaching from below 0V then you can be sure it will start-up going positive. This can be done if the LATI pin is not allowed to be high when the output is above ground. The circuit in figure 2 and the following procedure demonstrate how this can be implemented.

Initialization Procedure

- 1) Power up
- 2) Set LAT high
- 3) Set INH low
- 4) Load MSB: **0001 0000 0000 0000** :LSB
- 5) Set LAT low
- 6) Wait at least 1 output cycle time
- 7) Load all 0s
- 8) Set INH high (INH must go high before LAT by at least a NAND gate delay)
- 9) Set LAT high
- 10) Wait at least 1.5 output cycle times
Output stops at 0V going high
- 11) Load desired frequency
- 12) Set LAT low
Output begins at 0V going high
- 13) Set INH low

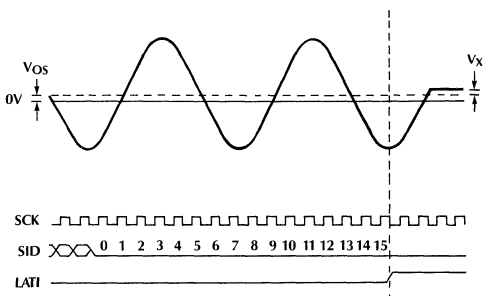


Figure 1. Inhibit Mode

$$|V_x| = \frac{V_{PEAK}}{256}; \text{ For } f_{OUT} \leq \frac{f_{CLK}}{2048}$$

$$|V_x| \leq \frac{V_{PEAK}}{256} + V_{PEAK} \text{ Sine} \left(\frac{8\pi f_{OUT}}{f_{CLK}} + \frac{\pi}{512} \right)$$

$$\text{For } f_{OUT} > \frac{f_{CLK}}{2048}$$

Application Note 12

Synchronization

At the completion of step 10 the part is initialized. Its output is stable at about 0V and will start up going positive. If you want to synchronize the output with some external event you can load the shift register with the desired frequency (step 11) and then set LAT low (step 12) synchronously with the external event. If you want to synchronize two ML2036 sinewave generators together initialize them both as described, and then set LAT low (step 12) on both circuits simultaneously.

Precise phase control between two parts can be achieved by initializing both parts, starting one and then waiting a known time before starting the other. For example, to produce two 5kHz sinewaves with 90° phase shift you should wait 50μs between starting each circuit. Since the ML2036 uses a 3MHz reference clock to update the output (assuming a 12MHz clock is used to drive CLK_{IN}) the phase resolution will be 0.6°. This resolution will vary from 0.0012° for two 10Hz signals to 6° for two 50kHz signals.

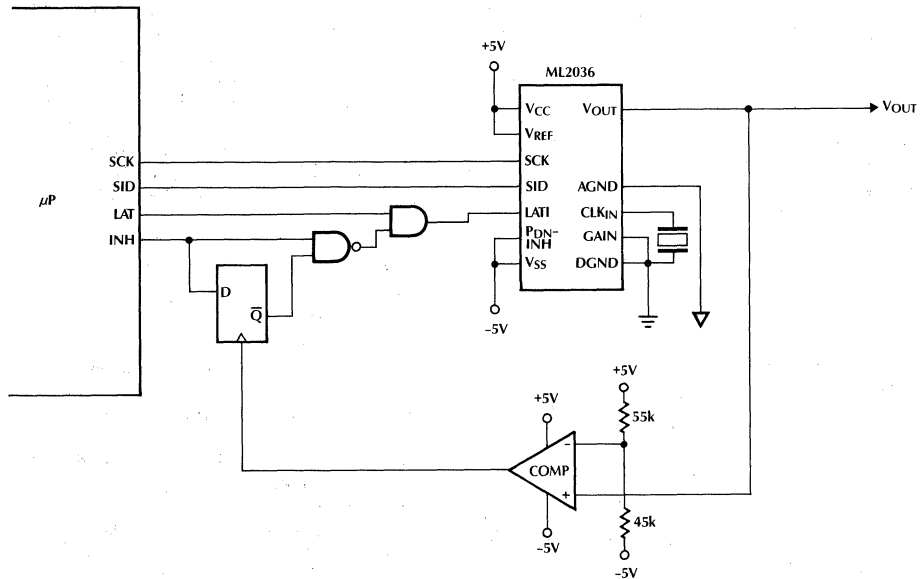


Figure 2.

William Cho

Designing with 10Base-T Transceivers

Micro Linear's family of 10Base-T transceivers offer highly integrated solutions for internal and external MAUs (Media Attachment Units) as well as HUB MAUs. These chips offer a high performance current drive transmitter with very low jitter and RFI noise. The ML4652 and ML4658 are 10Base-T transceivers that provide an AU interface for internal and external DTE MAUs while ML4654 and ML4655 provide TTL and ECL interfaces suited for Multiport Repeaters.

The following application note will cover some of the design issues that arise when designing either type of Media Attachment Units for Local Area Networks based on 10Base-T.

Internal (Embedded) MAUs with Shared AUI Port

Figure 1 shows a detailed schematic for an internal MAU design with a shared AUI (Attachment Unit Interface) port. The optional port requires additional circuitry as defined in the IEEE 802.3 specifications for proper termination and protection at the serial interface chip (or Manchester Encoder/Decoder) and the 10Base-T transceiver chip connected to an AUI port.

An AUI connection requires termination impedance of 78Ω on the receive end of the transmission lines (DI and CI). As such R1 and R2 in parallel with R3 and R4 provide the proper termination. This also applies to the receive output pins 4 and 5 of the transceiver chip. The 357Ω resistors for R3 and R4 was chosen to properly bias the driver circuitry (see section on AUI driver output). The $2k\Omega$ values for R7 and R8 were chosen to provide the BIAS voltage for Tx+ and Tx- inputs. This also will not load down the 78Ω transmission line when the AUI port is connected and the transceiver chip is tri-stated. The output AUI drivers of the transceiver chip must be tri-stated to not load down the transmission lines when the AUI port is connected and the twisted pair port is disconnected. Powering down the chip will tri-state the outputs.

The transceiver can be powered down by switching V_{CC} to GND as shown or by switching the ground connection to open condition. A logic level MOSFET with an "on" resistance ($R_{DS(ON)}$) of 0.5Ω or less can be used by connecting to the ground pin of the chip to power down the chip. When switching the ground off one must also include the ground connections of the driver resistors of COL and Rx outputs (R3, R4, R11, and R12). Another method of powering down the transceiver is to use an external mechanical switch as shown.

The isolation transformer is required for protection of the transceiver chip from 16V with respect to the system ground at the AUI interface during a fault condition as specified in 74.1.6. and 74.2.6 sections of the IEEE 802.3 standards for both the driver and the receiver.

If a shared AUI port is not required, then the design becomes simpler. Figure 13 of the datasheet shows AC coupling between the serial interface and the transceiver. This is to block DC bias voltage of the serial interface chip that may not match that of the transceiver. Micro Linear's transceivers require the input bias to be between 2.5V and 4.5V for CI and DI. If the two chips are compatible one can eliminate the AC coupling capacitors and bias resistors. By using a DC coupled interface, biasing the driver outputs is all that is needed for proper operation (Figure 2).

Twisted Pair Interface

The twisted pair connection to 10Base-T requires additional filtering and isolation components. The output structure of the twisted pair drivers are of the current drive type. This poses several very significant advantages when driving the twisted pair medium. Because the drivers are current driven, the differential outputs are well matched for a balanced signal transmission. Balanced transmission is crucial for meeting tight regulations on signal shapes. Also current driven outputs produce lower common-mode voltages for a lower EMI radiation. This can be a very significant issue when facing FCC regulations. Another advantage to current mode is that output drive, can be easily adjusted to compensate for losses in the transformer or output filter. RTSET will set the level of output drive current by the relationship:

$$RTSET = (RL/100) * 220$$

where RL is the characteristic impedance of the twisted pair cable.

The twisted pair differential output will see an effective resistance of 50Ω from the parallel combination of the two 200Ω resistors and reflected secondary AC line impedance of 100Ω for unshielded twisted pair. By driving 42mA to the 50Ω complex load, the differential signal voltage will swing $\pm 2.5V$ peak around the 5V bias point when taking transients into consideration.

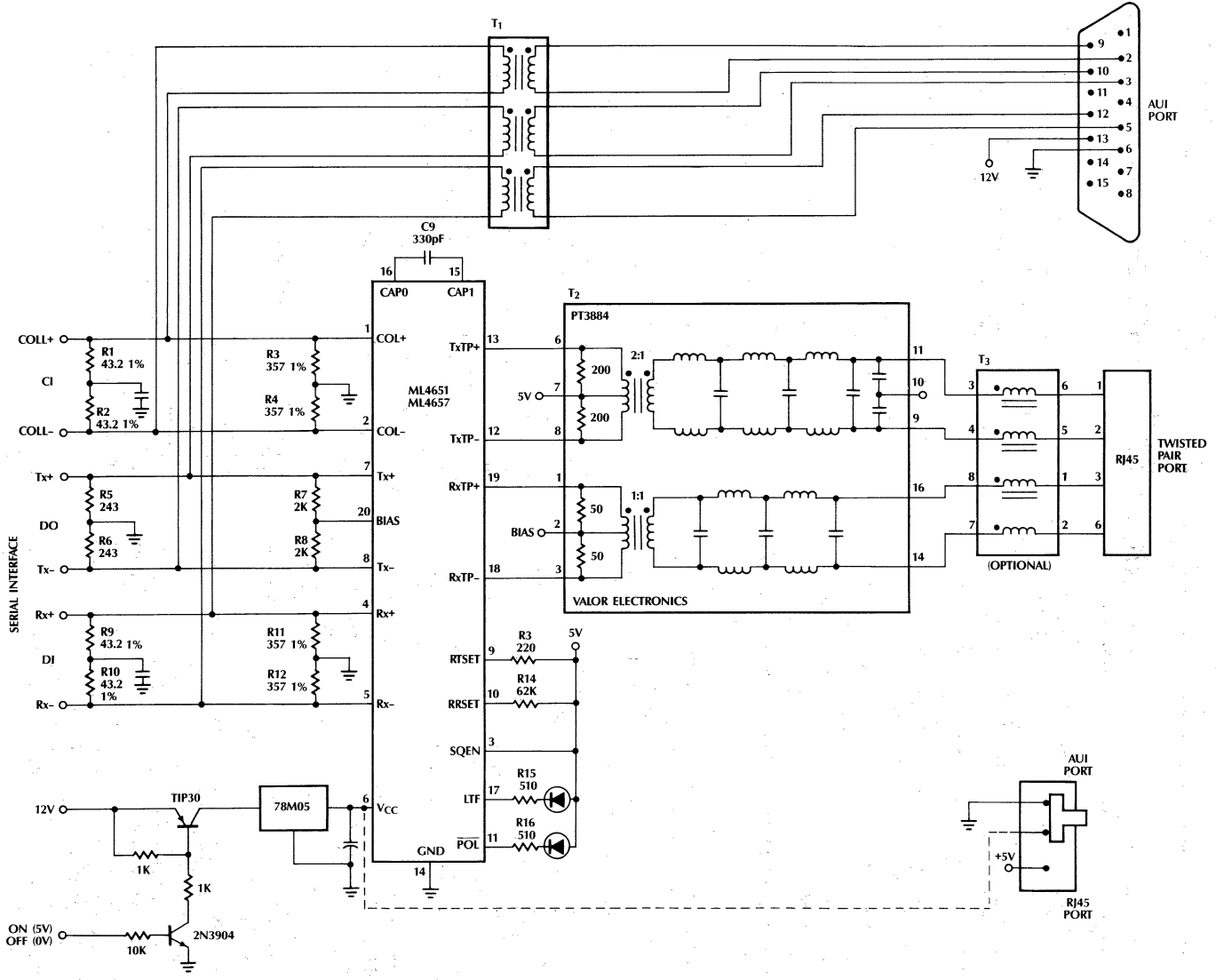


Figure 1. Internal MAU with Shared AUI Port

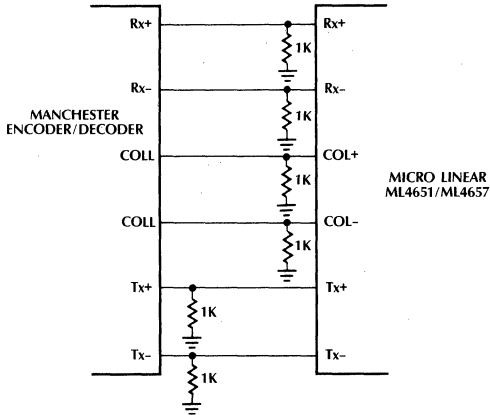
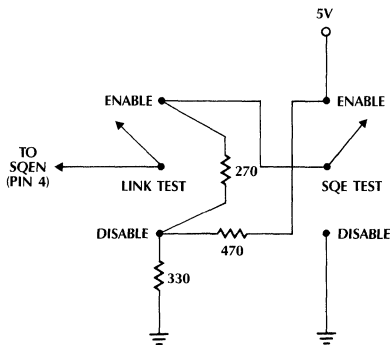


Figure 2. DC Coupled Interface for DTE Card Application

The isolation and filter components for both transmit and receive lines can all be integrated into one dip package style module. One such product can be obtained from Valor Electronics in San Diego. There are several other manufacturers who have these products available (refer to Figure 13 in the datasheet).

The output chokes shown in Figure 1 will pass any differential signal but block common mode voltages. Because Micro Linear's 10Base-T transceivers have very low common mode output voltage, this extra filtering choke may not be needed. Good board layout will also help.



SQE Test	Link Test	SQEN Pin 4
Enable	Enable	5V
Enable	Disable	3.3V
Disable	Enable	0V
Disable	Disable	1.2V

Figure 3. Mode Selection Circuit

External MAU

An external MAU design typically adds more LED outputs for status indication and adds circuitry for configuring the chip for SQE and Link Test options (See Figure 12 of datasheet). The selection of SQE and Link Test circuitry can be implemented in various ways. One such option is to use two SPDT switches to produce the proper voltage levels (Figure 3). The selected voltage to the SQEN input pin (pin #4 for ML4652 and ML4658) will internally configure the chip for the option to activate SQE test or Link test.

AUI Driver Output

The output structure of the driver stage connecting to the AUI is an open emitter type. The output is biased at typically 4.2V when high and 3.6V when low. That is a differential voltage of about $\pm 0.6V$ across a 78Ω load which calculates to about 7.7mA output current during transmission. A 360Ω resistor at the output pin sets its current at 11.7mA when high and 10mA when low. In the case when the positive output is high, the current (I_O) flowing out of its drive transistor is the sum of 7.7mA and 11.7mA (Figure 4). That means the current flowing out of the negative output is 10mA minus 7.7mA. It then becomes apparent that the termination resistance must be low enough as to not shut off either of the output drive transistors but not too low as to saturate the transistor and dissipate excessive power.

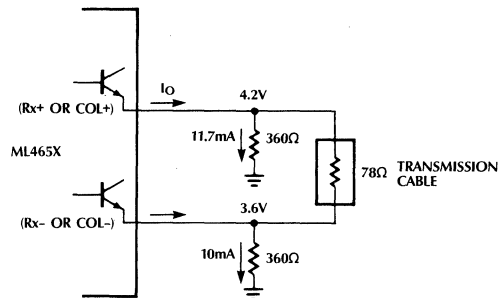


Figure 4. AUI Driver Circuitry

Jon Klein

Generating Fixed Frequency Sine Waves with ML2035

The ML2035 Programmable Sine Wave Generator is a convenient solution for generating accurate sine waves. Often an accurate fixed sine wave reference is required in applications such as Uninterruptable Power Supplies. Normally a microcontroller or microprocessor is used to program the ML2035's output frequency.

Most power supplies do not incorporate a microprocessor. This application note will show several alternate methods for programming the ML2035 for 50 or 60Hz sine wave output using off-the-shelf components.

The circuit in Figure 1 programs the ML2035 for 60Hz. The circuit shifts in the decimal number 141 to get to 60Hz with a NTSC color burst crystal (3.579545 MHz).

The 'HC4060 counter is used as an oscillator and timer. Q5 stays high for 16 clock cycles. During the first 8, the 'HC165 shifts out the codes on A thru H which are the complements of 141 (Binary 1000 1101) LSB to MSB respectively. During that time, S1 on the 'HC165 shift register is also high. That means that for the next 8 clock cycles the Q BAR output will be low, loading in 0's for the most significant 8 bits. SCK on the ML2035 and CLK on the shift register are run from complementary phases of the oscillator, since the 'HC165 changes data on the rising edge of its CLK and the ML2035 latches the same data on the rising edge.

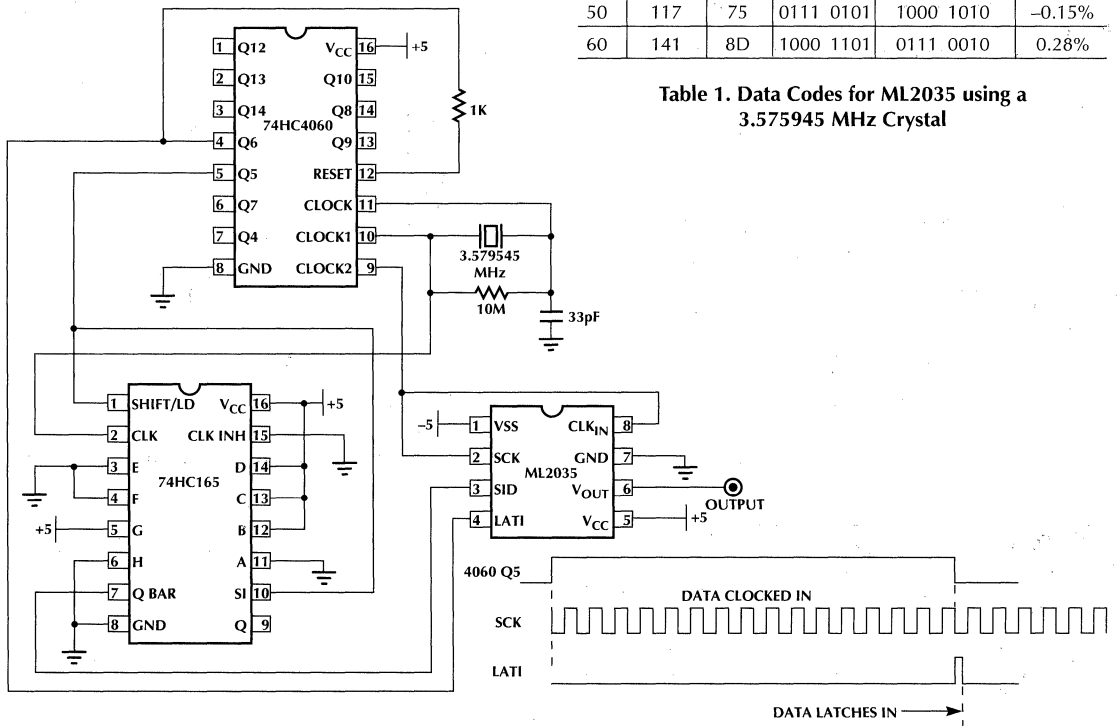


Figure 1. Programming the ML2035 for 60Hz. output using NTSC color burst crystal.

Application Note 14

When Q5 goes low again, Q6 goes high providing a reset for the counter and also a short pulse for the LATI input. The 1K resistor between Q6 and the 4060 Reset line delays the reset slightly, effectively stretching the LATI pulse to 50ns.

This circuit could also be run from +10V and GND by creating an "artificial ground" at 50% of the 10V line (two 1K resistors and a 10µF capacitor).

For 50 Hz output use the code shown in the table 1.

The circuit in Figure 1 allows the use of an inexpensive and readily available crystal, but has the disadvantage of not being "single pin" programmable for 50/60 Hz. The circuit in Figure 2 requires a non-standard, more expensive crystal frequency but has the advantages of being pin programmable for 50/60 Hz and eliminating the shift register. A 2.4576 MHz crystal is required to generate exactly 50 and 60Hz with no error.

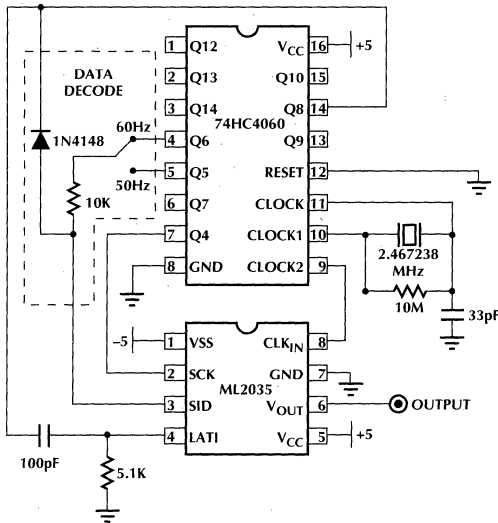


Figure 2. Single Jumper 50 or 60 Hz selection

These lower frequency (below 3.5MHz) crystals, however tend to be larger in size and significantly more expensive.

F _{OUT}	D(dec)	Hex Value	Binary Value	Frequency Error
50	170	AA	1010 1010	-0.39%
60	204	CC	1100 1100	-0.39%

Table 2. Data Codes and error terms for ML2035 using a 2.4576 MHz standard crystal

Table 3 shows the codes necessary to generate 50 and 60 Hz sine waves with the circuit in Figure 1 from various standard crystal frequencies. Note that for the highest accuracy, the 4.194304 MHz crystal yields both 50 and 60Hz sine waves with no frequency error.

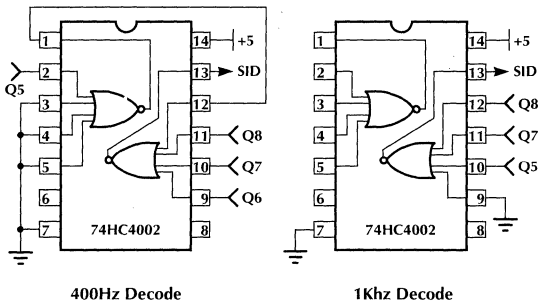
F _{CRYSTAL} (MHz)	F _{OUT}	D(dec)	D(hex)	74HC165 Code ABCD EFGH	Error
4.00	50	105	69	1001 0110	0.14%
4.00	60	126	7E	1000 0001	0.14%
4.194304	50	100	64	1001 1011	0.00%
4.194304	60	120	78	1000 0111	0.00%
6.00	50	70	46	1011 1001	0.14%
6.00	60	84	54	1010 1011	0.14%
8.00	50	52	34	1100 1011	-0.82%
8.00	60	63	3F	1100 0000	0.14%

Table 3. Shift register values and frequency errors for various standard crystal values (fig 1).

To generate 400Hz and 1KHz tones, the circuit of Figure 2 must be modified to shift in 0 for the first 8 clock pulses and the data in table 4 for the last 8 clocks. This can be accomplished by replacing the "Data Decode" blocks of Figure 2 with the decoding shown below.

F _{CRYSTAL} (MHz)	F _{OUT}	D(dec)	D(hex)	74HC165 Code ABCD EFGH	Error
6.5536	400	512	200	1111 1101	0.00%
6.5536	1000	1280	500	1111 1010	0.00%

Table 4. Generating 400Hz and 1Khz sine waves



Suggested Crystal Manufacturers:

Nymph/Saronix: (415) 855-6829
Pletronics: (206) 776-1800

Manijeh Fadaee

Designing an IEEE 802.3 FOIRL Transceiver

Introduction

The ML4661 FOIRL Transceiver with the ML4621/ML4622 fiber optic quantizer construct both an internal and external Fiber Optic Inter-Repeater Link (FOIRL) described in the IEEE 802.3 standard. The ML4661 through its standard 802.3 AU interface can be connected to an AUI cable, Ethernet Manchester Encoder/Decoder or Hub controller. The following topics will be discussed in this Application Note:

- 1) ML4661 Features
- 2) ML4621/ML4622 Features
- 3) Filtering Power & Ground
- 4) Attachment Unit Interface
- 5) Interfacing ML4661 to National DP83950 Repeater Interface Controller
- 6) Interfacing ML4661 to AT&T T7201 Multi-Port Repeater
- 7) FOIRL system specification
- 8) 10Base-FL (ML4662)
- 9) Layout Considerations
- 10) Initial debug of the FOIRL/10Base-FL board

ML4661 Features

The ML4661 integrates many of the functions needed for an FOIRL transceiver. The Attachment Unit Interface complies with the IEEE standard offering Transmit, Receive, and Collision pair signals. Data transmission includes transmit squelch, a 1MHz idle signal, and the jabber function. The receiver accepts ECL compatible levels from the ML4621 or ML4622 quantizer, passing through the receive squelch and onto the AUI. The complete FOIRL state machine is also incorporated including collision detect, loopback, and low light conditions.

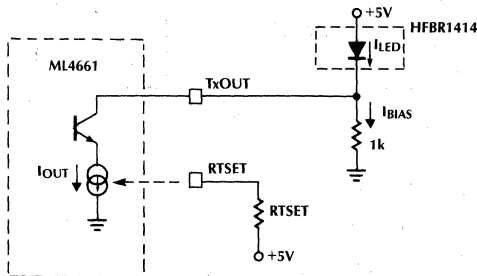


Figure 1.
 $I_{LED} = I_{BIAS} + I_{OUT}$

One of the features is the capability to disable or enable the SQE function. This allows an FOIRL transceiver to be connected to a DTE as well as a repeater. When connecting to a DTE, SQE must be enabled. When connecting to a repeater, SQE is disabled. The forthcoming 10Base-FL specification includes SQE test, however, most FOIRL compatible equipment today supersedes the standard by including the SQE feature.

The ML4661 transmitter consists of a current driver output (TxOUT) which directly drives an HP fiber optic LED transmitter (HFBR1414). The output structure of the driver stage (TxOUT) is an open collector (NPN) acting as a current source as shown in Figure 1.

The 1KΩ pull-down resistor on the TxOUT pin prebiases the LED by applying a small forward current while the LED is in the "off" or low light state. The prebias current prevents the junction and parasitic capacitances from discharging completely when the LED is in the "off" state, thus reducing the amount of charge that the driver must transfer to turn the diode back on.

The resistor on the RTSET pin controls the amount of current driven by this pin (TxOUT). RTSET and the pull-down resistor together set the extinction ratio. To calculate RTSET value for a certain sink current (maximum 80mA) at TxOUT use this equation:

$$RTSET = (52\text{mA}/I_{OUT}) 162\Omega$$

The ML4661 Transmitter has been designed to drive the cathode of the LED (note: higher optical power transmitted corresponds to the low logic state). Current flowing through the LED corresponds to a logic low. When TX- > TX+ (DO on AUI), current flows and this will be a logic low. When TX+ > TX- (DO on AUI), no current flows (except bias) corresponding to a logic high.

The receiver inputs are ECL levels (the ML4661 receiver inputs) coming from the ML4621 or the ML4622. When LMON_{IN} (TTL) coming from the ML4621 or the ML4622 is low, and the level of the received signal exceeds the receive squelch requirement, the receive data is buffered and transmitted out to DI pair.

The five LED status outputs are active low, open collector outputs. They provide a visible status of the link as follows:

\overline{XMT} : is on when transmission is taking place.

\overline{RCV} : is on when the transceiver is receiving a frame from the ML4621 or the ML4622 (V_{IN+} and V_{IN-}).

COLSN: is on when the transmitter and the receiver are active at the same time (collision). When a collision takes place a 10MHz \pm 15% square wave with a 50% \pm 10% duty cycle will be sent out to COL+ and COL- output pins.

JAB: turns on when the transmitter is on for more than 20 to 150 Msec. When the jabber is on, in order to not bring down the network, the jabber logic disables the transmitter and turns on the collision signal COL+ and COL-.

LMON: is on when LMON_{IN} coming from the ML4621 or the ML4622 is low and there are signal transitions on RX+ and RX-.

ML4621/ML4622 Features

The ML4621/ML4622 data quantizer is used for signal recovery applications in Fiber Optic systems. The ML4621/ML4622 has a wide bandwidth and large gain which makes it capable of accepting an input signal from a fiber optic receiver as low as 2mV. This analog input gets converted to digital outputs at the TTLOUT pin or ECL+ and ECL- output pins.

The TTL output has been disabled in this application Note by pulling up TTL_{VCC} and TTL_{GND} since ECL output levels are required by the ML4661. The 3K pulldown resistors on the ECL outputs of the ML4621 keep the outputs biased in their operating range. Due to internal pulldown resistors in the ML4622, the external pulldown resistors are not required if the ECL output are the only outputs being used. The ML4661 inputs present a minimal load to the ML4621/ML4622 ECL outputs.

ML4621/ML4622 include a two stage input limiting amplifier with a DC restoration feedback loop. The bandwidth of the ML4621 can be adjusted to the particular needs of the application with the capacitor across pin 7 and 8 (CF1 and CF2 for high corner frequency). The .1 μ F input capacitors on V_{IN+} and V_{IN-} set the low corner frequency. (The output source impedance of the fiber optic receiver must be kept low "about 50 Ω " to make the input capacitors on V_{IN+} and V_{IN-} effective).

Since the logic condition for the 802.3 FOIRL is as follows:

"light" = 0
"dark" = 1

The received signal has to get inverted before it goes to the RXIN+ & RXIN- inputs of the ML4661. When the output of the Fiber Optic receiver (HFBR2416) is connected to the V_{IN+} of the ML4621/ML4622, the signal is non-inverted at the ECL+ & ECL- outputs. Therefore, the ECL+ must be connected to the RXIN- of the ML4661 and the ECL- to the RXIN+ of the ML4661. The output of the Fiber Optic receiver (HFBR2416) may be connected to the V_{IN-} of the ML4621 to invert the ECL outputs of the ML4621/ML 4622. If this is done, the following connections must be made.

ECL+ \rightarrow RXIN+
ECL- \rightarrow RXIN-

The Link Monitor function is implemented by the minimum signal discriminator and the threshold generator circuits. The TTL_{LINKMON} and ECL_{LINKMON} outputs both indicate when the input data signal is less than a user defined acceptable level. This is done by monitoring the input signal and peak detecting the output of the limiting amplifier and comparing this level with the voltage at V_{THADJ}. V_{THADJ} is set by the user as specified in the data sheet. To set the minimum input signal of the ML4621 to 3mV, V_{REF} can be tied directly to V_{THADJ} to provide 2.5V at V_{THADJ}.

$$\text{ML4621: } V_{\text{THADJ}} = 600 V_{\text{INTH(P)}} + .7V \quad (1)$$

$$\text{ML4622: } V_{\text{THADJ}} = 500 V_{\text{INTH(P-P)}} \quad (2)$$

In these equations V_{INTH} is the peak or peak to peak value of the input signal. The receiver sensitivity can be calculated when the Hewlett Packard HFBR2416 with a typical responsivity of 6mV/ μ W is being used.

$$\begin{aligned} \text{ML4621: } V_{\text{INTH(P)}} &= 3\text{mV(Peak)} & (V_{\text{THADJ}} &= V_{\text{REF}}) \\ \text{Received Power} &= 6\text{mV(P-P)}/(6\text{mV}/\mu\text{W}) \\ &= 1\mu\text{W} = -30\text{dBm (PEAK)} \end{aligned}$$

$$\begin{aligned} \text{ML4622: } V_{\text{INTH(P-P)}} &= 5\text{mV(Peak to Peak)} & (V_{\text{THADJ}} &= V_{\text{REF}}) \\ \text{Received Power} &= .833\mu\text{W} = -30.7\text{dBm (PEAK)} \end{aligned}$$

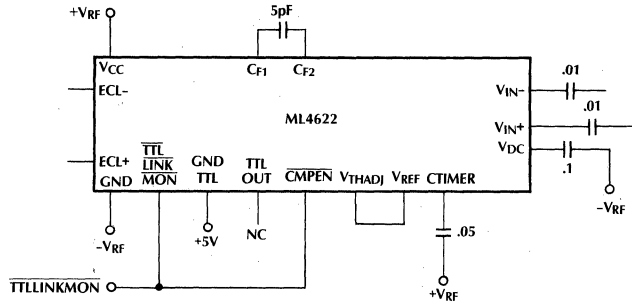
Note: Peak Power = Average Power + 3dBm

This meets the IEEE802.3 FOIRL receiver specifications. A lower threshold level can be set by dividing down V_{REF} with a resistor divider, as shown in Figure 2b. By choosing 1K and 140 Ω , the V_{THADJ} will be 2.2 volt in the ML4621 which will set the minimum power level at about 2.4mV peak and minimum launch power at -33dBm (considering worst responsivity of 4.5mV/ μ W). However due to a better stability of the link monitor in the ML4622, both standards will be met considering the worst conditions by tying the V_{THADJ} to V_{REF}. For more detailed information refer to Application Note 6 and Application Brief 1.

In the case of oscillation at TTL_{LINKMON}, hysteresis can be added to the ML4621 in two different ways as follows:

- 1) Adding a feedback resistor from the TTL_{LINKMON} output to the V_{THADJ} (this will only work if a resistor divider is being used to arrive at V_{THADJ}).
- 2) A capacitive feedback can be implemented by connecting a capacitor from the TTL_{LINKMON} to the ISET pin on the minimum signal discriminator (this will apply if V_{REF} is tied to the V_{THADJ}).
Note: Adding a 300 ohm to 600 ohm pull-up resistor at the TTL_{LINKMON} to +5Volt may be needed for the stability.

Based on the layout, the value of the hysteresis resistor and capacitor change. Adding 7pF capacitor in the ML4661EVAL board (Fig. 10) will serve this purpose. Since hysteresis has been added internally to the ML4622 to increase the stability, the external hysteresis components are not required.



NOTE: IF TTILOUT IS USED, TIE GNDTTL TO UNFILTERED GROUND AND REMOVE L1. IF TTILOUT AND ECL OUTPUTS ARE BOTH USED, ADD 3k PULLDOWN RESISTORS AT ECL OUTPUTS.

Figure 2A. Fiber Optic Receive Circuit A

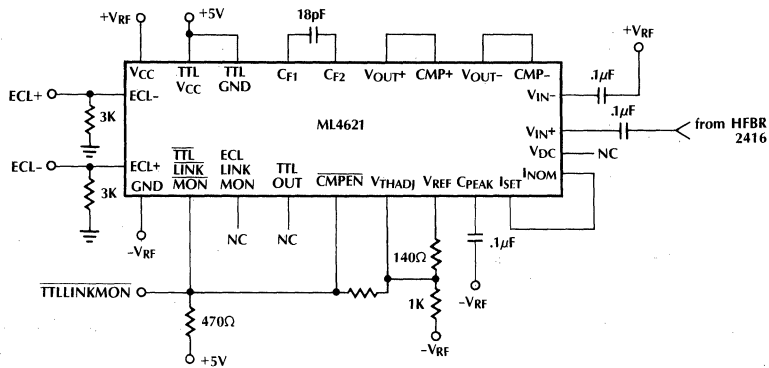
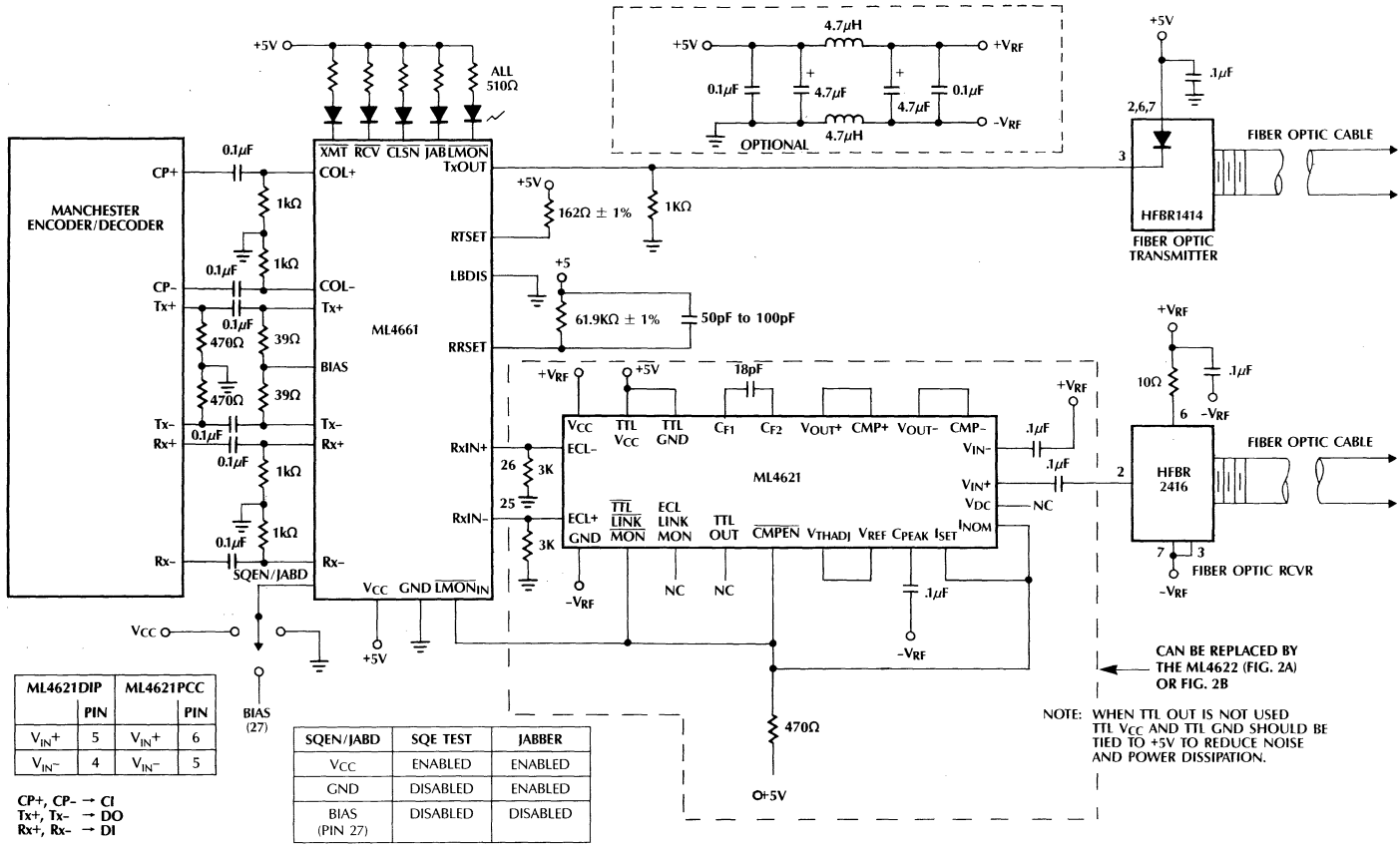


Figure 2B. Fiber Optic Receive Circuit B

Figure 3. Internal MAU with AC Coupling



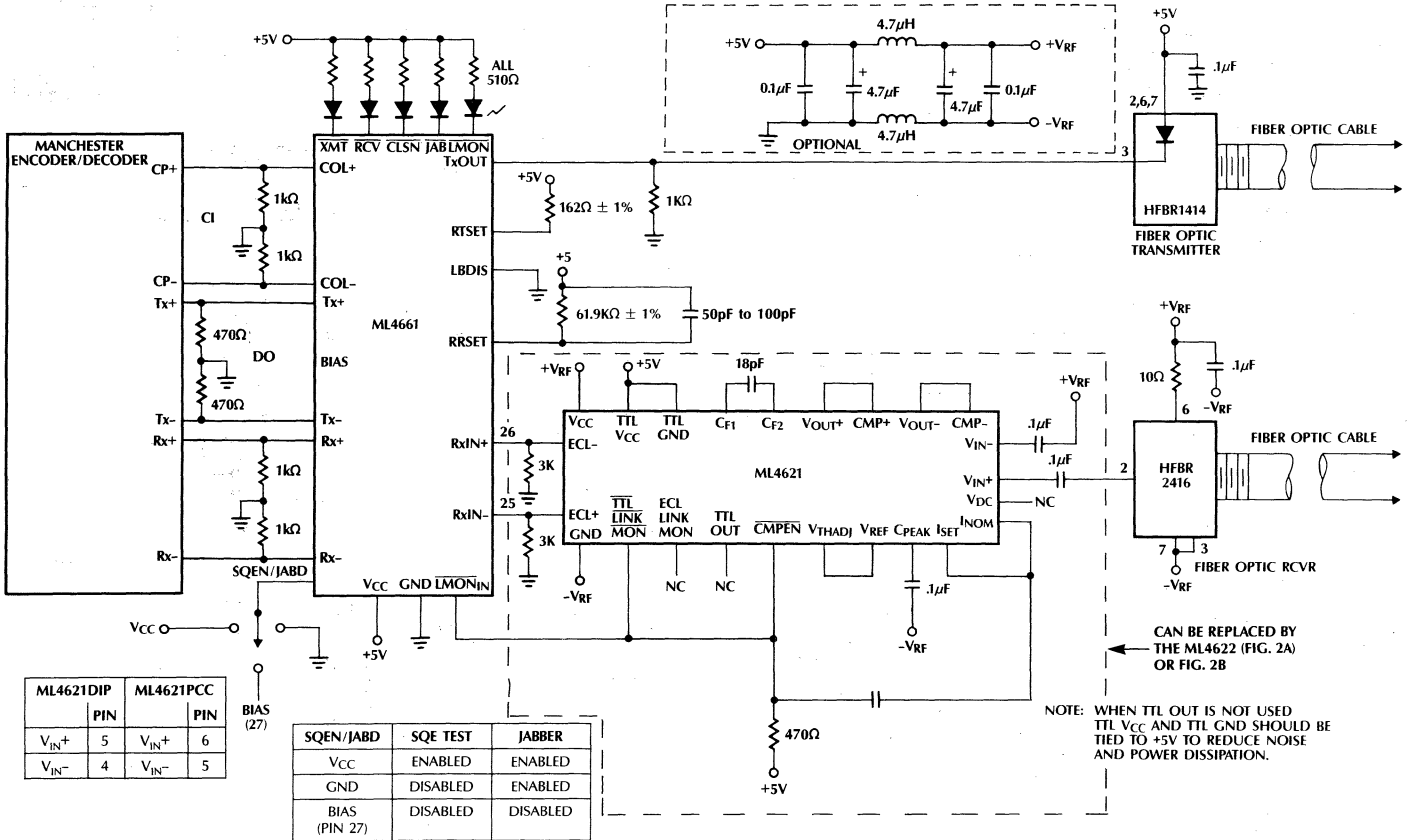


Figure 4. Internal MAU with DC Coupling

ML4621DIP	PIN	ML4621PCC	PIN
V _{IN+}	5	V _{IN+}	6
V _{IN-}	4	V _{IN-}	5

SQEN/JABD	SQE TEST	JABBER
VCC	ENABLED	ENABLED
GND	DISABLED	ENABLED
BIAS (PIN 27)	DISABLED	DISABLED

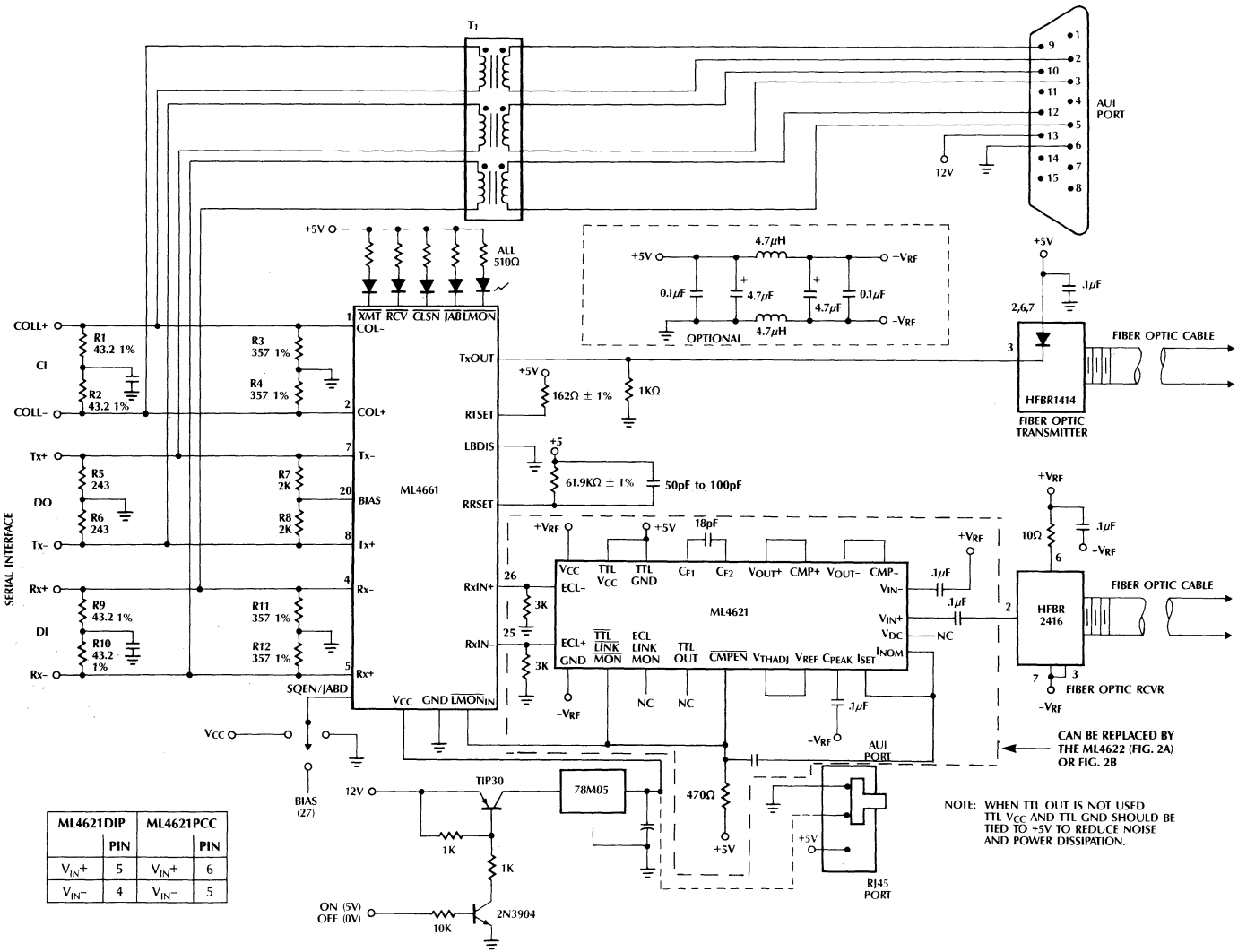


Figure 5. Internal MAU with Shared AUI Port.

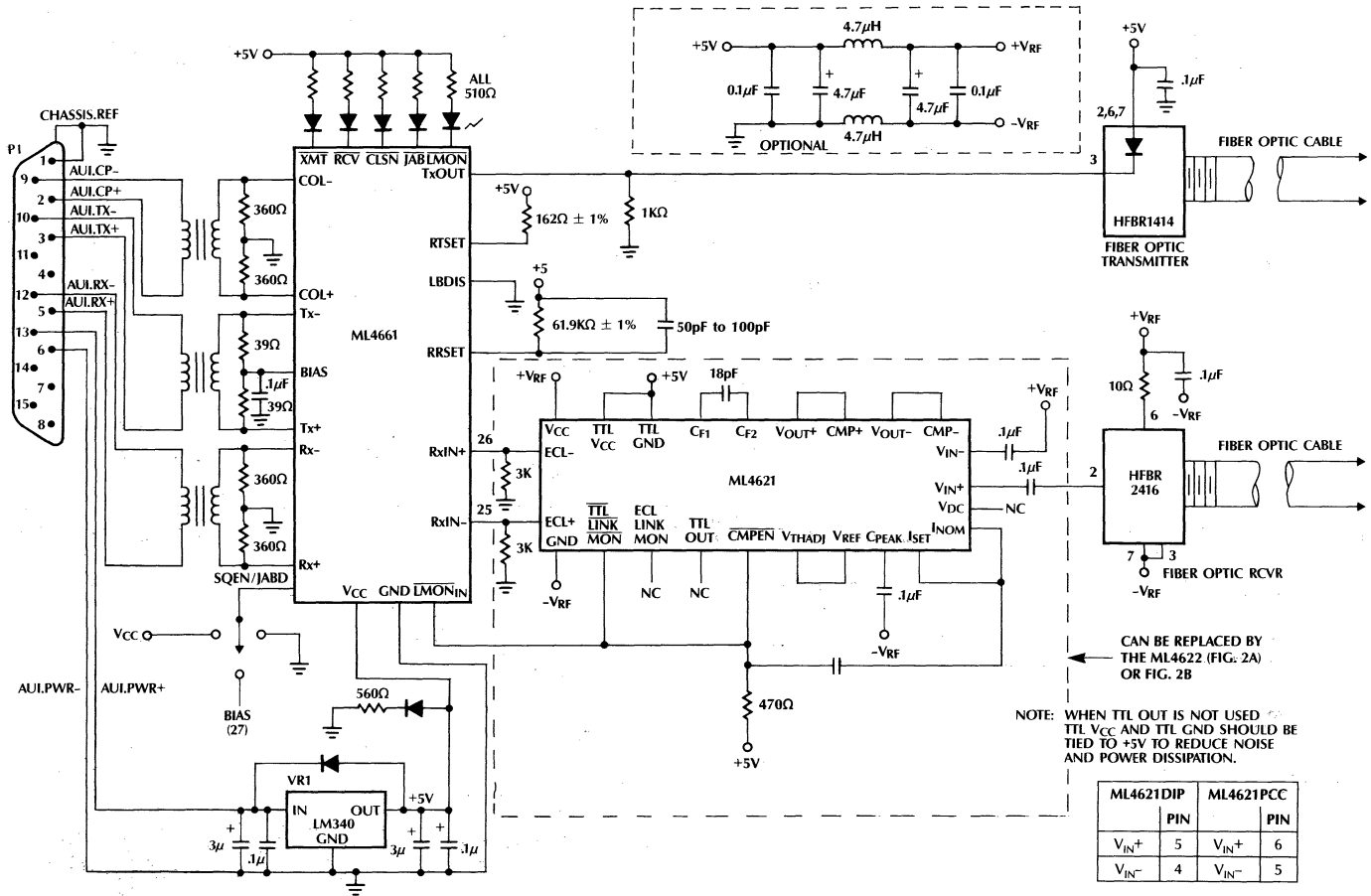


Figure 6. External MAU

Filtering Power and Ground

Filtering is necessary since unintended feedback through the power supply system (the metallic conductive path of the +5V power line or the Ground reference line for the receiver) can create sustained oscillations or degrade the sensitivity of the receiver. Filters in the power supply for the post-amplifiers, comparator stages and the receiver, prevent noise generated by the quantizer output from being conducted back through the power system to the input amplifier stage.

The quantizer inputs are sensitive low level inputs. V_{CC} and Ground Decoupling are necessary. Disabling the TTL output (by connecting TTL_{GND} and TTL_{VCC} to V_{CC}) of the quantizer will also reduce noise.

Attachment Unit Interface

The ML4661 and the ML4621 can be used as an internal MAU with the option of having a shared AUI port or as an external MAU. Figures 3, 4, 5, and 6 show a detailed schematic for these three configurations.

Internal MAU: The AU interface may be AC coupled through $.1\mu F$ capacitors across DO, DI, CI pair (Figure 3). They may be DC coupled if the DC levels DO (2V to $V_{CC} - .5V$), DI (3.6V to 4.5V), CI (3.6V to 4.5V) of the manchester encoder/decoder & the ML4661 are the same (Figure 4). (If DC coupling is used, the BIAS pin is not connected and the 39 Ω resistors are not needed.) If AC coupling interface is used, DO which is an input must be DC biased (shifted up in voltage) through the BIAS pin for the proper common mode input voltage.

DI and CI are emitter follower outputs which need external 1K Ω or greater (depending on the particular manchester encoder/decoder) pulldown resistors to ground.

By using 1K pulldown resistors we can minimize power dissipation by not having to drive the 78 Ω AUI cable.

Internal MAU with Shared AUI Port: The AU interface is AC coupled through isolation transformers T1 (Figure 5). This is to protect the transceiver chip, from 16V with respect to the system ground at AUI interface during a fault condition (as specified in 74.1.6 and 74.2.6 section of the IEEE 802.3 standards for both the driver and the receiver). In addition, it blocks the DC offset voltage of the AUI port that may not match that of the transceiver. An AUI connection requires termination impedance of 78 Ω $[(R1+R2) \parallel (R3+R4)]$ on the receive end of the transmission lines (DI and CI). The 357 Ω resistors for R3 and R4 are chosen to properly bias the driver circuitry. The 2K resistors on TX+ and TX- provide common mode bias input voltage for the ML4661.

The 243 Ω resistors (R5, R6) drive the DO pair (either the DO pair of the AUI port of the ML4661). The output AUI drivers of the transceiver must be tri-stated in order to not load down the transmission lines when the AUI port is connected and the FIBER OPTIC port is disconnected. Powering down the chip will cause the outputs to be in tri-state (refer to Application Note 13). This power down circuitry is shown in Figure 5.

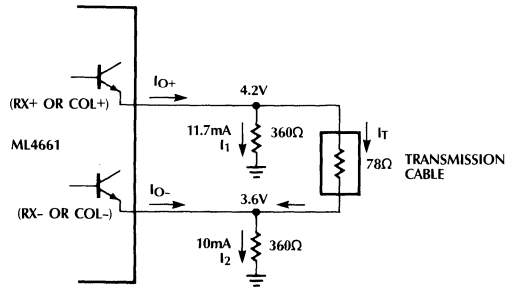


Figure 7. AUI Driver Circuitry.

External MAU: The AU interface is AC coupled through isolation transformers (Figure 6). An AUI connection requires termination impedance of 78 Ω . Two 39 Ω , 1% resistors tied to DO pair provide impedance matching (78 Ω) as well as the proper common mode input voltage to the ML4661.

DI and CI pairs are emitter follower outputs. The output structure of the driver stage (RX+, RX-, COL+, COL-) is open emitter (Figure 7). The output is biased at typically 4.2V when high and 3.6V when low. That is a differential voltage of about +.6V across a 78 Ω load.

The pulldown resistors have to be chosen such that during transmission, a minimum of 2.0mA can be sourced by RX- or COL-. By using a 360 Ω pulldown resistors the RX- or COL- source 2.3mA and the RX+ or COL+ source 19.4mA as follows:

$$I_T = .6V/78\Omega = 7.7mA$$

$$I_{O+} = I_1 + I_T$$

$$I_{O+} = (4.2V/360\Omega) + 7.7mA$$

$$I_{O+} = 11.7mA + 7.7mA$$

$$I_{O+} = 19.4mA$$

$$I_{O-} + I_T = I_2$$

$$I_{O-} = I_2 - I_T$$

$$I_{O-} = (3.6V/360\Omega) - 7.7mA$$

$$I_{O-} = 10mA - 7.7mA$$

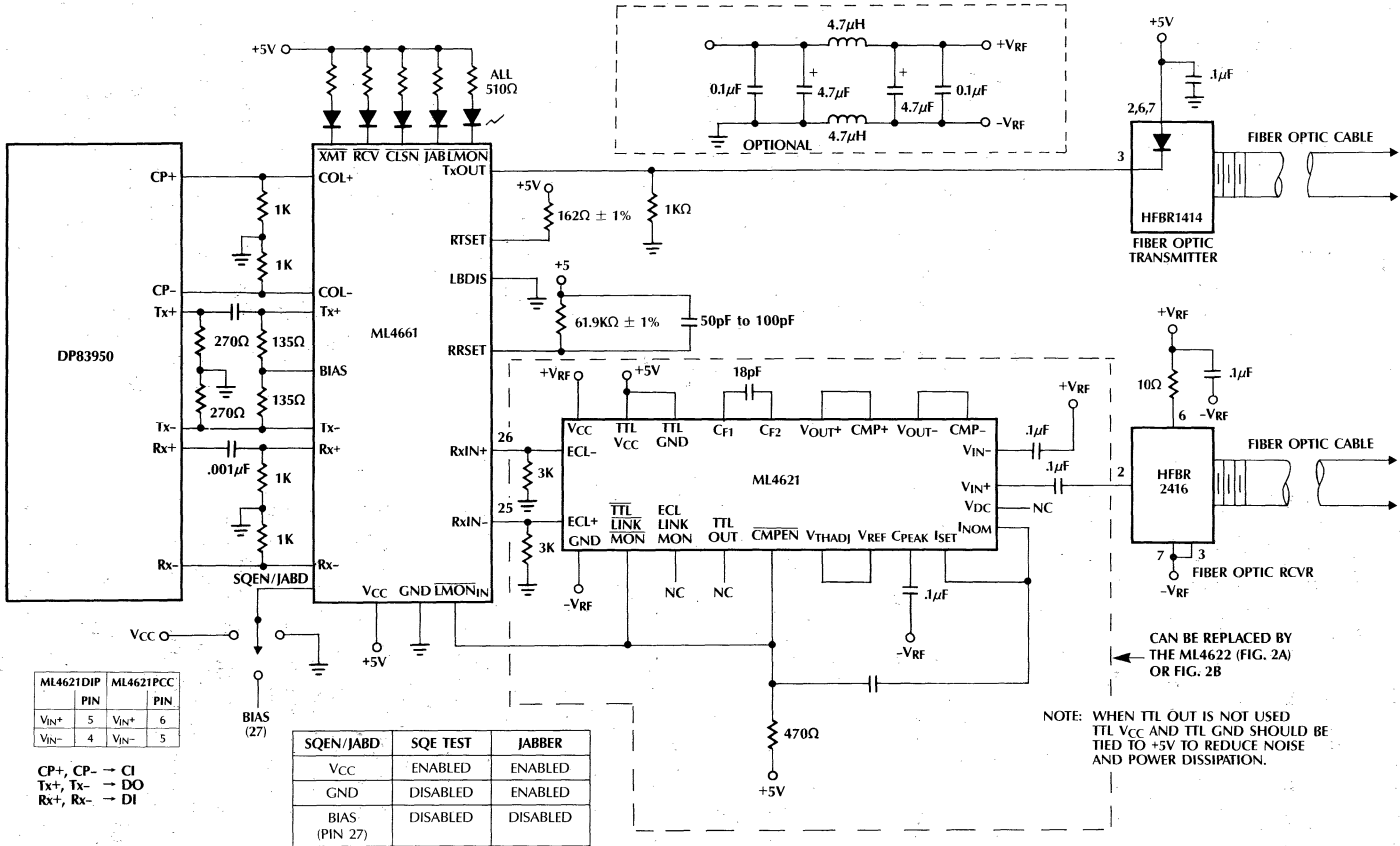
$$I_{O-} = 2.3mA$$

The termination resistance must be low enough (minimum 200 Ω) to not shut off either of the output drive transistors, but not too low in which case the output transistors could saturate.

Interfacing ML4661 to DP83950

The ML4661 and ML4621 can be used in a HUB application. Figure 8 shows the interface between the ML4661 and the National Semiconductor DP83950 Repeater Interface Controller (RIC). The DI and CI pairs are DC coupled but DO is AC coupled with $.1\mu F$ in each lead. The 1K Ω pulldown resistors on the DI & CI pairs provide the necessary source current to drive DI & CI pairs.

Figure 8. Interfacing the ML4661 to DP83950



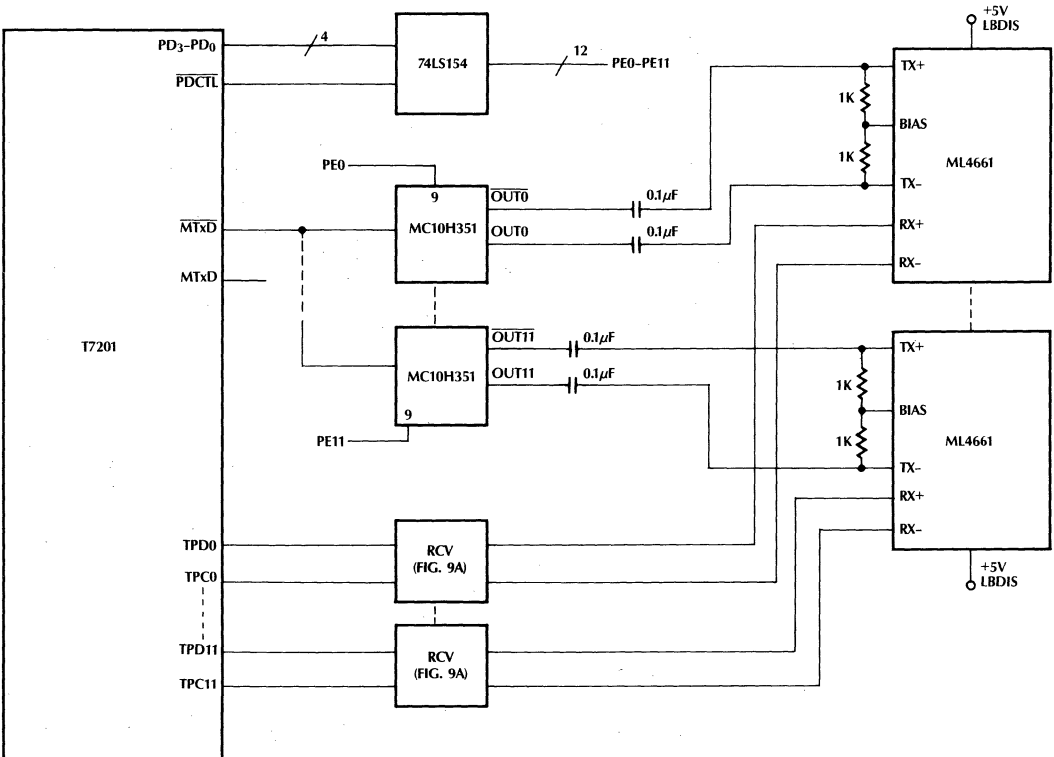


Figure 9. Interfacing ML4661 to T7201.

Application Note 15

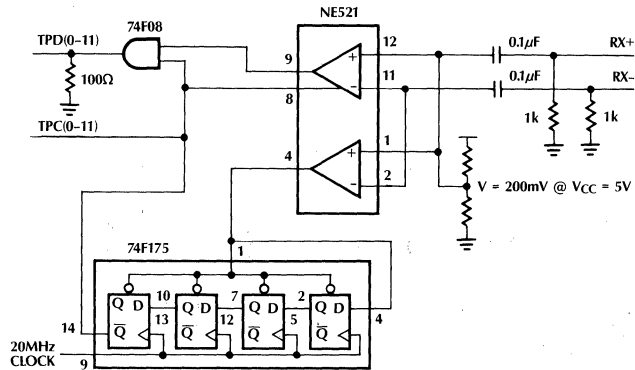


Figure 9a. RCV

Interfacing ML4661 to T7201

The T7201 is AT&T Multi-Port Repeater. The analog front-end of the T7201 interface is shown in Figure 9A. It consists of two main sections, Transmit and Receive.

Transmit Section: The T7201 to line driver interface consists of the six signals from the T7201 (MXTD, MTXD, PD3-PD0, PDCTL).

The T7201 HUB controller distributes the signal to all ports with the exception of the one that transmitted the data. The 74LS154 decodes the PD3-PD0 address lines. The output of the 75LS154 then selects and disables the port that the data was received from. The MC10H351 that is selected (disabled) is the only output that is not transmitting data to its respective transceiver.

The manchester data is provided by the T7201 on the MTXD and $\overline{\text{MTXD}}$ lines. The $\overline{\text{MTXD}}$ transmit output pin of the T7201 gets converted to the ECL differential pair by the MC10H351 and will drive the transmit pair (TX+ and TX-) of the ML4661.

Receive Section: The ECL differential receiver output (RX- & RX+) of ML4661 must be converted to a TTL single ended signal. This is accomplished with the NE521.

In addition, the incoming receive signal is fed into a gated line receiver RCV which is controlled by the squelch circuitry.

Squelch Circuit: The squelch circuit differentiated noise from valid incoming data on the receive pair. It does this by detecting signals that are above a preset voltage threshold level for a sufficient period of time. When there is no signal on the receive pair, the squelch circuit disables the line receiver, and deasserts the TPC0-TPC11 signal to the T7201. When a signal above the threshold arrives, TPC0-TPC12 is asserted, and the line receiver is enabled. The squelch circuit ensures that the receive circuits in the T7201 are operating only

during packet reception. The circuit shown in Figure 9 uses a high-speed comparator (NE521) with an offset threshold. The output of this comparator is fed to a retriggerable timing circuit that controls the TPC0-TPC11 signal to the T7201. To ensure recognition of the Idle (end of packet signal), and to prevent midpacket deassertion of TPC0-TPC11, the timing circuit should be set to detect positive pulses between 1.5 and 2.0 bit times (200ns). The timing circuit can be implemented by using a quad flip-flop (74F175) clocked from a 20MHz clock generator. The 74F175 activates TPC0-TPC11 for the pulses greater than 200mV, within 150ns to 200ns timing window (four 20MHz clock cycles).

FOIRL System Specifications

Some of the key parameters required by the IEEE 802.3 FOIRL Standard are listed below:

Transmitter Specifications

- 1) Peak Emission Wavelength = 790 to 860nm.
- 2) Spectral Width < 75nm. This is determined by measuring the Full Spectral Width at Half Maximum Amplitude (FWHM) of the LED optical emission. This parameter must be measured at the maximum temperature at which the LED will be operated.
- 3) Minimum Extinction $\geq 13\text{dB}$
Extinction = $|P_{T(\text{ON})} - P_{T(\text{OFF})}|$, where P_T = peak transmitted optical power measured in dBm.
- 4) Optical Rise/Fall time T_R and $T_F < 10\text{ns}$.
- 5) $|T_R - T_F| \leq 3\text{ns}$.
- 6) Transmitter Jitter $\leq 2\text{ns}$.
- 7) $P_{T(\text{MAX})} = -9\text{dBm}$ (PK)
Launched power into 62.5/125 fiber with NA = .275
- 8) $P_{T(\text{MIN})} = -15\text{dBm}$ (PK)
Launched power into 62.5/125 fiber with NA = .275 at the beginning of the LED lifetime.
- 9) $P_{T(\text{MIN})} = -18\text{dBm}$ (PK)
Launched power into 62.5/125 fiber with NA = .275 at the end of the LED lifetime.

Receiver Specifications

- 1) Overdrive limit = -9dBm (PK) maximum.
- 2) Sensitivity $\geq -27\text{dBm}$ (PK) minimum.
- 3) The data output of the receiver must be inhibited before the Bit Error Ratio of the fiber optic link degrades to greater than 10^{-10}

Total Link Specifications

- 1) Jitter at receive out $\leq 4\text{ns}$.
- 2) Transmitter/Receiver must be compatible with fibers having the following core/cladding diameters: 50/125, 62.5/125, 85/125 and 100/140.
- 3) Link must operate at a maximum length of 1Km with each type of fiber.

10Base-FL

The IEEE 802.3 10Base-F task force has proposed a new fiber optic Ethernet standard that will update the original 1987 FOIRL standard. This proposed standard known as 10Base-FL describes a fiber optic link segment with enhanced performance and backwards compatibility with FOIRL. One of the enhancements for 10Base-FL, SQE test, is already included in the ML4661. This allows a FOIRL MAU to connect to a HUB or a DTE. Other modification have to do with extending the fiber optic cable length from 1,000 meters to 2,000 meters, and internal state machine modifications.

The ML4662 is a 10Base-FL transceiver which is pin compatible with the ML4661 and meets the 10Base-FL standard. This new part will plug into an existing ML4661 socket and be pin for pin compatible. Note that since the fiber optic cable length has been extended for 10Base-FL, the Receive Sensitivity specification has been improved from -27dBm peak for FOIRL to -32.5dBm average for 10Base-FL. An FOIRL design using the ML4661 today may want to target a receive sensitivity of -32.5dBm average so that the product can be upgraded to 10Base-FL without a redesign.

To upgrade a FOIRL MAU to 10Base-FL, the following steps should be taken:

- 1) Replace the ML4661 with the ML4662.
- 2) Remove 50pF to 100pF capacitor across RRSET (pin 13 of the ML4661).
- 3) Set the receive sensitivity for -32.5dBm (average).

Layout Considerations

The fiber optic transceiver consisting of the ML4661 transceiver and the ML4621/ML4622 fiber optic quantizer are simple to implement from a data point of view. Electrically, the quantizer is resolving 2 mv signals in a logic environment that has an abundance of 5 volt signals. The fiber optic receiver and the quantizer require careful layout, attention to noise coupling, and very clean power supply busses. The following recommendations should be considered while laying out the printed circuit artwork.

Power Supply

- 1) Isolate and filter the power and ground to the ML4621/ML4622 (analog portion) and HP receiver (to ensure that noise is not coupled into the low level receiver inputs). This can be accomplished with a pi filter that has a 4 to $7\mu\text{H}$ inductor in both the power lead as well as the ground lead.
- 2) Make sure that adequate decoupling is used on both sides of the pi filter, on each chip, and at the fiber optic transmitter and receiver. The fiber optic receiver should be decoupled from the +5 Volt Filtered bus with a 10 ohm resistor and decoupling capacitor. Allow room for large (.47 μF) decoupling capacitors and determine if they can be reduced during testing of the prototypes.

Ground Plane

- 1) The printed circuit board should be a 4 layer board with the +5V. and ground each providing a shielding plane on the inner layers. The fiber optic receiver and the ML4621 analog front ends should have its own power supply planes separate from the +5 Volt and Ground planes for the remaining circuitry. These planes should be separated by an air gap physically and electrically by the power supply pi filter from the logic +5 volt and ground.
- 2) Connect unused pins of HP receiver to the low level receiver Ground.

Transmitter

- 1) The transmitter output (TXOUT) traces should be as short as possible and make them wide to lower their characteristic inductance.
- 2) Keep RRSET and RTSET traces (of the ML4661) and resistors away from each other.

General Layout

- 1) The physical layout for the receiver should be in a straight line to minimize the trace lengths and potential for noise coupling between the logic signals at the output of the quantizer and the low level signals on the inputs.
- 2) The trace from the output of the HP fiber optic receiver to the ML4621/ML4622 (V_{IN+} , V_{IN-}) should be as short as possible and shielded if possible.
- 3) Because of the high gain low level input circuitry in the ML4621/ML4622, parasitic feedback from the high-level logic-compatible output must be kept to a minimum in order to prevent undesired oscillations. This is accomplished with a layout which physically separates the receiver inputs (V_{IN+} , V_{IN-}) and outputs (ECL+, ECL-, TTLOUT, CMPEN, TTL_{LINKMON}, ECL_{LINKMON}).
- 4) If the TTL outputs of the ML4621/ML4622 are not used, Connect GNDTTL and VCCTTL to +5 volt (this will disable the TTL driver).

Initial Debug of the FOIRL/10Base-FL Board

- 1) AUI is connected but Fiber Optic cable is not connected.
 - a) Look for 1MHz Idle signal at pin 18 of the ML4661. If there is no Idle signal, verify the following:
 - i) Ground and +5V to the ML4661.
 - ii) RRSET must be 61.9k (1%) at pin 13 of the ML4661 to +5V
 - iii) RTSET should be 162 ohm at pin 12 of the ML4661 to +5V to set the current driven by the TxOUT to 52mA.
 - b) The LMON LED must be OFF. If the LMON LED is ON, check the TTL_{LINKMON}. If it is low, measure the noise level at filtered power and ground, V_{IN+} and V_{IN-} of the ML4621/ML4622. If the peak noise level at the input of the ML4621/4622 is greater than the minimum V_{INTH} (Equations 1 and 2), the TTL_{LINKMON} gets activated (low).
 - c) The RX LED must be OFF. Otherwise there must be transitions on RX+ and RX- (pins 25 and 26 of the ML4661) less than 3μsec apart because of one of the following reasons:
 - i) Too much noise at the inputs of the ML4621 (not filtering properly).
 - ii) Feedback between the inputs and outputs of the ML4621/ML4622 (poor layout).
 - iii) Crosstalk between TxOUT (pin 18 of the ML4661) and inputs of the ML4621/ML4622 (poor layout).
- 2) Connect Receive Fiber (HFBR2416) to Fiber Optic LED transmitter (HFBR1414) which is sending active idle signal. (Either from another MAU's LED transmitter or from the same MAU's LED transmitter with loopback disabled.)
 - a) The LMON LED must go ON and the RX LED must be OFF. If the LMON LED is OFF, verify the following steps:
 - i) The Receive Power must be within the FOIRL/10Base-FL standard range.
 - ii) Verify the idle signal at V_{IN+} or V_{IN-} of the ML4621/ML4622.
 - iii) TTL_{LINKMON} (pin 2 of the ML4621) must be low.
 - b) If the RX LED is ON as well as the LMON LED, check step 1c.
- 3) Start to transmit. The LMON and TX LEDs must be ON. The RX and CLSN LEDs must be OFF in the transmitting MAU if two MAU's are being used. In this case, if the RX and CLSN LEDs are on, check step 1c.
- 4) Disconnect the Fiber Optic cable from the HFBR2416. The LMON LED must go OFF. If the LMON LED stays on, check step 1b. The RX LED must be OFF. Otherwise check step 1c.
- 5) After successfully completing the initial debug of the FOIRL board, verify that the board meets FOIRL/10Base-FL specifications.

Figure 10 is the schematic of the FOIRL evaluation board (ML4661EVAL) which meets the FOIRL standard. This board incorporates all the above critical points of the layout as shown in Figures 11 through 16. The ML4661EVAL is available for purchasing.



Figure 11.

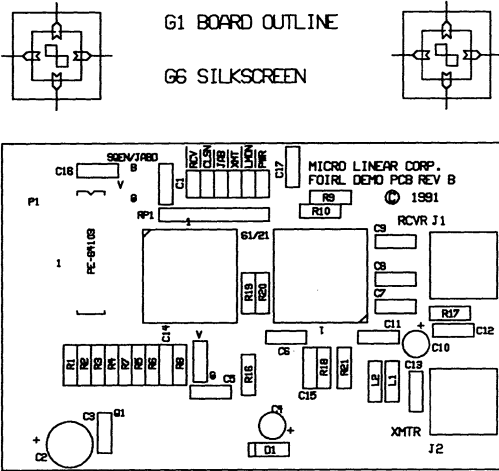
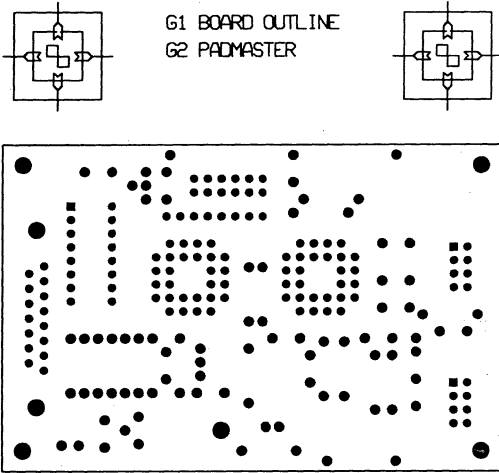


Figure 12.

Figure 13.

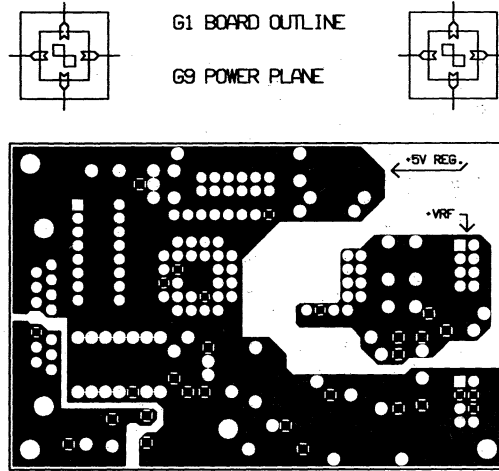
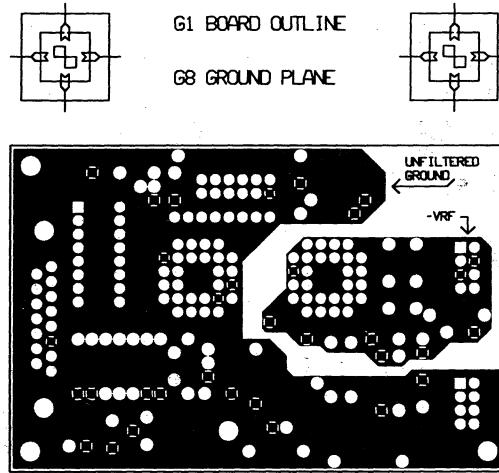


Figure 14.

Figure 15.

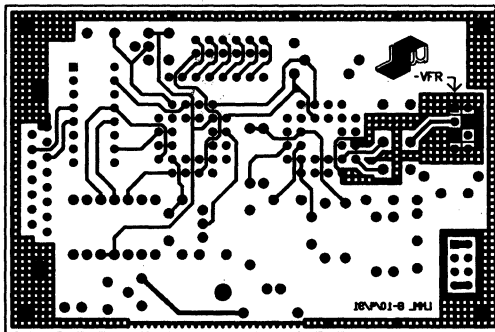
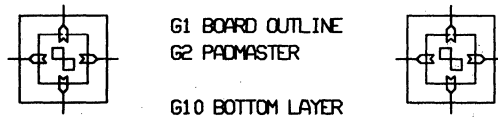
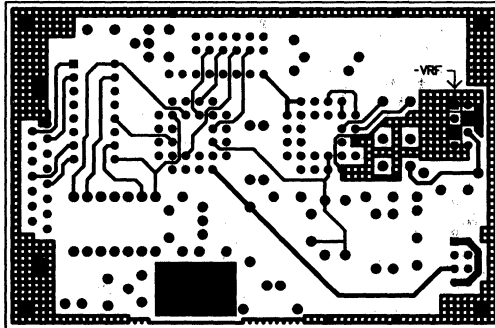
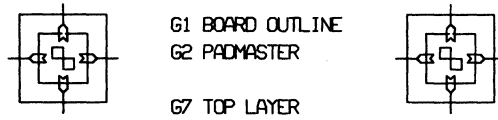


Figure 16.

Mehmet K. Nalbant
 William Cho

Theory and Application of the ML4821 Average Current Mode PFC Controller

I. THEORY OF OPERATION

The ever increasing importance of power factor correction has prompted the design and availability of many power factor controller Integrated Circuits. Power factor correction requires special control circuits that are able to force the input current waveshape to be sinusoidal and in phase with the input sinusoidal voltage.

There are several ways that this can be accomplished. One method is the average current mode controlled boost topology power factor correction circuit, using the ML4821 dedicated average current mode controller IC. This paper is going to present enough theoretical background information along with practical examples to enable the design of such circuits.

Average current mode control can produce a high quality input sinusoidal current waveform. Although it can be used with many different power supply topologies, it excels when it is used with the continuous inductor current, boost topology.

Power Factor Correction

What is power factor? Enough has been said and written in the past couple years about this question. Therefore we are not going to elaborate on it. Instead we are going to look into how a power factor correction circuit operates.

Figure 1, shows the simplified block diagram of a power factor correction circuit. The circuit functions by monitoring the input full wave rectified line voltage as well as the output voltage. The two feedback signals are combined to set up the current trip points that shape the input current waveform to be sinusoidal and yet still regulate the output over line and load variations.

From now on we are going to use the acronym PFC instead of power factor controller.

Figure 2 shows the basic circuit diagram of a PFC with all necessary connections made. The heart of the circuit is the current modulator. The modulator consists of a linear multiplier, a current amplifier, and a PWM comparator. These three functional blocks enable the circuit to force the input current to be sinusoidal.

A current that is proportional to the input full wave rectified voltage is produced with the help of resistor R_L . We will call this the **reference**. The reference is applied to one of the inputs of the multiplier. The other input of the multiplier is the output of the error amplifier. For the time being we are going to assume that the output of the error amplifier changes slowly compared to the line frequency. This is in general true since the bandwidth of this amplifier is set low by its feedback components.

The multiplier is a current input type. This enables the multiplier to have greater ground noise immunity. When there is a current at its input, its terminal voltage is a diode drop between 0.7V and 1V. In fact it is part of a current mirror. **Therefore a voltage source of low impedance should never be applied to this input.**

The output of the multiplier is current that is the product of the reference current and the output of the error amplifier that monitors the output voltage. This output current is applied to resistor R_c (see Figure 2). This voltage subtracts from the sensed voltage across R_s and is applied to the current error amplifier. Under closed loop control the current error amplifier will try to keep this voltage differential close to zero volts. This forces the voltage

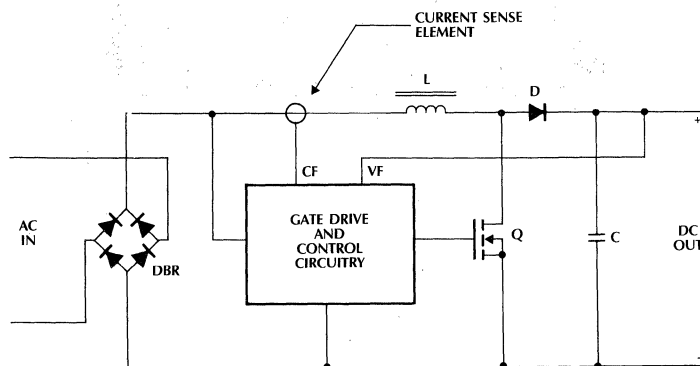


Figure 1. Top level block diagram of the power factor controller

Application Note 16

switching frequency. 100KHz is a good trade-off point between inductor size and circuit efficiency. Switching losses in the circuit will include major losses in the MOSFET, output diode, and the inductor. Because the MOSFET is charged to the output voltage at every turn-off, switching losses will be significant at any input voltage and output current. The output diode must reverse recover high current with the full output voltage. Core losses in the inductor will not be as significant because the 100KHz AC ripple current is relatively low compared to the almost DC 120Hz rectified sinusoidal current.

For the purposes of this next analysis we are going to ignore the current ripple in the inductor. Lets also assume that somehow the MOSFET duty cycle is such that the inductor is forced to carry a current that has a full wave rectified sinusoidal waveshape. From the operation of the boost circuit there needs to be equilibrium throughout the entire 50Hz or 60Hz cycle. Large signal equations describing the operation of the boost circuit should hold.

$$V_{OUT} = \frac{V_{IN}}{1 - D_{ON}} \quad (1)$$

and

$$D_{ON} = \frac{V_{OUT} - V_{IN}}{V_{OUT}} \quad (2)$$

Since $V_{IN} = v_{IN}(t) = \sqrt{2}V_{INRMS}|\sin(\omega t)|$ we get

$$D_{ON} = d_{ON}(t) = \frac{V_{OUT} - \sqrt{2}V_{INRMS}|\sin(\omega t)|}{V_{OUT}} \quad (3)$$

Also by definition

$$\hat{i}_l(t) = \sqrt{2} \frac{P_{OUT}}{V_{INRMS}} |\sin(\omega t)| \quad (4)$$

The MOSFET current is the inductor current chopped at high frequency with the above duty cycle. The diode current on the other hand is the inductor current chopped at high frequency with duty cycle $(1 - d_{ON}(t))$. By substituting we can get an expression for the average current that passes through diode D (i.e. $I_D = i_d(t)$).

Note that the little hat on top of the variables denotes average value.

$$\hat{i}_d(t) = \hat{i}_L(t)(1 - d_{ON}(t)) \quad (5)$$

By substituting (3) and (4) into (5),

$$\begin{aligned} \hat{i}_d(t) &= 2 \frac{P_{OUT}}{V_{OUT}} \sin^2(\omega t) \Rightarrow \\ \Rightarrow \hat{i}_d(t) &= \frac{P_{OUT}}{V_{OUT}} - \frac{P_{OUT}}{V_{OUT}} \cos(2\omega t) \end{aligned} \quad (6)$$

As can be seen from the above equation the diode current consists of two parts. It has an average value consistent with the output power and output voltage (first term). Also it has an AC component with a peak value equal to that of the average value. The DC part of this current is simply

the output load current. It flows through the output load. The AC part however flows through the output capacitor C. Consequently it may become a parameter when determining the value of this capacitor.

Now lets get back to the loops of the PFC. Earlier we mentioned that there are two control loops; an inner high bandwidth current loop and a much slower outer voltage loop. Figure 3 shows the two loops in block diagram form. First we are going to examine the two loops separately. Then we are going to see the criteria for proper connection.

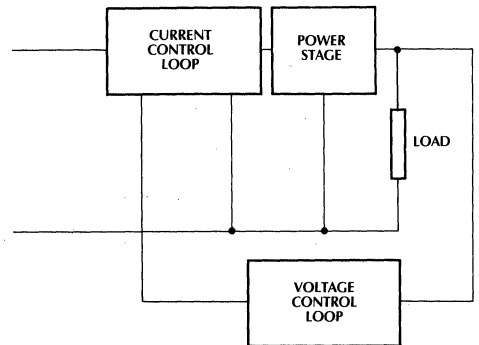


Figure 3. The two loops of a PFC. Inner current loop and outer voltage loop.

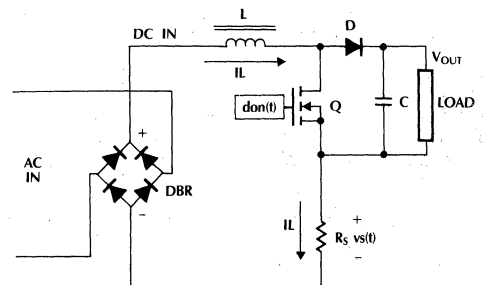


Figure 4. The power stage.

The Current Control Loop

The current control loop constitutes the inner loop and its job is to force the input current waveshape to follow the shape of the input voltage. It does this by modulating the duty cycle of the MOSFET in the power stage. The input voltage is a full wave rectified sine wave. Thus it is harmonically rich. The current control loop along with the power stage has to have enough bandwidth to follow this full wave rectified waveform. It can be shown that a bandwidth of a few KHz is sufficient. In order to proceed we need to derive expressions that give the responses of both the power stage and of the current loop.

As can be seen from Figure 4 the input of the power stage is the duty cycle output of the current pulse width modulator. Therefore we can describe the power stage as a functional block that has as its input the duty cycle information and as output the sensed voltage across the sense resistor R_s . The average current that flows through this resistor is equal to the average current that flows at the input of the PFC.

We can define

$$G_{PS}(s) = \frac{V_S(s)}{D_{ON}(s)} \quad (7)$$

as the gain of the power stage. The response can be found by assuming that the output voltage is constant and by using the state space averaging technique. The response shows a single pole roll off and is given by

$$I_L(s) = \frac{V_{OUT} D_{ON}(s)}{sL} \quad (8)$$

Since $V_S(s) = R_S I_L(s)$,

$$G_{PS}(s) = \frac{V_{OUT} R_S}{sL} \quad (9)$$

The above expression gives the small signal gain of the power circuit in the complex s -domain. It is the ratio of the sensed current waveform voltage to the incremental changes in the duty cycle. We can go one step further and incorporate in the above power stage gain the gain of the pulse width modulator. To do this we first have to find the gain of the modulator itself. For that we have to know the amplitude of the applied ramp to the noninverting input along with the allowable voltage swing range at its inverting input. The gain of the modulator is

$$G_{PWM} = \frac{\Delta D_{ON}}{\Delta V^-} \quad (10)$$

where ΔV^- is the voltage at the inverting input of the PWM comparator.

For the ML4821 the amplitude of the oscillator is 5.2V peak to peak. Therefore when the voltage at the (-) of the PWM comparator changes by 5.2V the duty cycle goes

from zero to full duty cycle. If we assume that the deadtime is very small (normally around 5%) the gain of the PWM stage becomes

$$G_{PWM} = \frac{1}{5.5} \quad (11)$$

Now we can combine the gains of the power and PWM stages to get the following

$$G_{PST} = \frac{V_S(s)}{V^-(s)} = \frac{V_{OUT} R_S}{5.5sL} \quad (12)$$

Note that in actuality $V^-(s) = V_{IA OUT}(s)$. Therefore the overall current loop response will be determined by the responses of the current amplifier and the power stage. The overall response will be dictated by the required current loop bandwidth. For good waveform quality the total response should have a bandwidth of a few KHz.

Determination of the Current Loop Bandwidth

There is a theoretical upper limit for this bandwidth and is given by the following equation

$$f_{CLCO} = \frac{f_s}{6} \quad (13)$$

Thus for an operating frequency of 100KHz the maximum allowable current loop bandwidth is approximately 16KHz

The Gain Adjustor

Analysis of the voltage control loop shows that as the RMS AC input voltage goes up, the system gain increases by V_{RMS} . The gain increases with input voltage since the input voltage drives the one input to the multiplier. The second term is because the di/dt on the inductor increases in proportion to the input voltage.

Since the gain varies with V_{RMS} , it then follows that the unity gain crossover frequency of the loop will change with a 1:8 ratio as the line changes from 90VAC to 260VAC. This complicates the loop design since the wide variation in crossover frequency would require the low line crossover frequency to be set very low while the high line crossover frequency would be set high.

The ML4821 cancels the square law dependency by adjusting the gain of the multiplier as a function of the RMS input voltage. The multiplier gain is equal to:

$$\frac{1}{kV}$$

Where k assumes one of two values in the active region (active region is the voltage range that appears on pin #8 that corresponds to the desired operating input voltage range). Voltage on pin #8 is a scaled down average of the rectified input AC voltage. Below we are going to see ways for designing an appropriate network that will accomplish this task.

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Lets take a look at Figure 5. This curve shows the gain adjustor gain with respect to the voltage at pin #8. The curve has been separated in two parts. The right hand part is for operation under normal conditions in the voltage range from minimum line voltage to maximum line voltage (90VAC to 260VAC). 85VAC on the curve has been chosen to account for tolerances. Under normal operating conditions as input voltage decreases the gain increases compensating for the drop in the loop gain.

Under brownout conditions (below 85VAC) the gain decreases to limit the amount of current that is drawn from the line thus preventing an overload condition. This is a very useful feature since in many cases the load for a PFC is a constant power load. The input current has to go high to compensate for a drop in the input voltage.

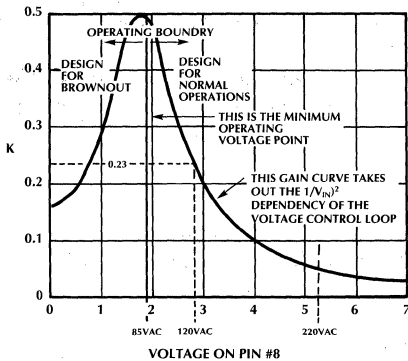


Figure 5. K-factor. Gain adjustor gain with respect to the voltage at pin #8.

Figure 6, shows the way pin #8 should be connected to the input line. The network consists of R8, C3, R9, R15 and C7 to form a two stage RC low pass filter and voltage divider. To calculate the values of the components, we must first select the minimum operating voltage point. Then we correspond this to the start of the brownout condition. From Figure 5, this is 85VAC which corresponds to approximately 2V at pin #8. On the same axis 2.8V corresponds to 120VAC. In other words when the input voltage is 120VAC the voltage at pin #8 must be 2.8V. Therefore the output voltage of the below filter/divider network should be 2.8V.

The RMS value of the input sinewave is equal to the RMS value of the full wave rectified sinewave after the full bridge rectifier. The average value of the full wave rectified sinewave on the other hand is proportional to its RMS value and they are related as follows

$$V_{AVG} = \frac{2\sqrt{2}V_{INRMS}}{\pi} \quad (14)$$

The average voltage at pin #8 is given by

$$V_{PIN\#8} = \frac{R15}{R8 + R9 + R15} V_{AVG} \quad (15)$$

Assuming:

$$R8 = 910K$$

$$R9 = 91K$$

R15 can be found by equating the above equation to 2.8V and solving it.

This yields an R15 value of 27K. The values of C3 and C7 are chosen for good attenuation at 120Hz and minimum delay. Typical values are as follows

$$C3 = 0.1\mu F$$

$$C7 = 0.47\mu F$$

For most applications these values are good even though the values of the resistors may change to accommodate different brownout or operating conditions. The values are output power independent.

With the gain adjustor functional, the multiplier output current is given by

$$I_{MO} = K \times I_{SINE} \times (V_{EAOUT} - 0.8) \quad (16)$$

where

I_{SINE} = reference current through pin #5.

K = gain adjustor gain (this quantity is dependent on the voltage present at pin #8). This is related to k from previous discussion. But is not equivalent.

V_{EAOUT} = output voltage of the error amplifier.

The maximum value of the multiplier output current is limited by the value of the timing resistor and it is given by

$$I_{MOMAX} = \frac{2.5}{R_T} \quad (17)$$

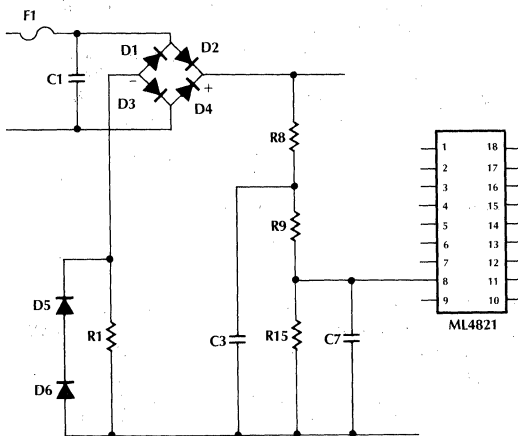


Figure 6. V_{RMS} sensing network.

Typical value for R21 for 100KHz operation is 6.2K, in which case $I_{MO\ MAX} = 400\mu A$.

It is a good idea to limit the maximum output current of the multiplier below the current limit point, but high enough to get maximum output power.

Output Capacitance

This is a good point to talk about the output capacitance. The parameters that affect its choice are listed below

1. Hold-up time capability, usually 20msec for computer power supplies.
2. Ripple current handling capability.
3. Allowable third harmonic distortion.

The hold-up time capability is the amount of time at rated output power that will take the capacitor voltage to discharge to a minimum operating voltage. The start point of the dropout should be the minimum operating output voltage, for this type of PFC this is usually less than the nominal value of 380VDC.

$$C_{OUT} = \frac{2P_{OUT}t_{HLD}}{V_{OUT\ MIN}^2 - V_{OP\ MIN}^2} \quad (18)$$

where:

C_{OUT} = output capacitance.

P_{OUT} = output power.

t_{HLD} = hold-up time, normally 20msec.

$V_{OUT\ MIN}$ = minimum value of the output regulated voltage, normally happens at full load.

$V_{OP\ MIN}$ = minimum input voltage of the driven load, usually a switching power supply.

The chosen capacitor should be able to handle the ripple current that will flow through it. The peak value of this ripple current, as it was found earlier is equal to the output DC current. The RMS ripple current through the capacitor is

$$I_{COUT\ RMS} = \frac{I_{OUT\ DC}}{\sqrt{2}} \quad (19)$$

The third consideration in the determination of the output capacitor is the output ripple voltage which can be found using the following

$$|V_{OUT\ RIPPLE}|_{PEAK} = I_{OUT\ DC} \sqrt{\left(\frac{1}{4\pi f_L C_{OUT}}\right)^2 + ESR^2} \quad (20)$$

where:

f_L = line frequency.

ESR = ESR of the of the output capacitor.

Depending on the amount of the output capacitor the contribution of the ESR on the output ripple voltage may not be ignored.

The output ripple voltage will contribute to the third harmonic distortion of the input current. The actual amount will depend on the value of the output ripple voltage and the gain of the error amplifier at 120Hz.

The Voltage Control Loop

The inner current control loop can be modelled as a controlled current source. This simplifies the analysis of the voltage control loop.

Typical loads for a PFC are switching power supplies which are essentially constant power loads. These kinds of loads exhibit negative resistance at their input terminals. An increase in the input voltage causes a drop in the input current. It is therefore important that the voltage control loop error amplifier is correctly compensated. The two other types of loads are the constant resistance and the constant current.

Before we proceed with the design of the voltage control loop we have to analyze the loop to find out what parameters affect its dynamics. Earlier we mentioned that this loop has a very low bandwidth. If the bandwidth of this loop is high, excessive amount of the second harmonic component present at the output will be injected in the control loop causing third harmonic distortion of the input current.

Typical values for this loop are between 10Hz and 20Hz. To find the open loop voltage gain we have to calculate the change in the output voltage of the error amplifier that produces the required maximum output power change. This can be calculated by using the following expression

$$\Delta V_{EAOUT} = \frac{P_{IN} \times R_S \times R_L}{R_{MO} \times K \times V_{RMS}^2} \quad (21)$$

where:

P_{IN} = maximum input power.

R_S = current sense resistor.

R_L = input voltage sense resistor that connect to pin #5.

R_{MO} = resistor at the output of the multiplier.

K = gain adjustor gain at V_{RMS} , from the curve of Figure 5, K at 120VAC is 0.23.

V_{RMS} = input RMS voltage, this voltage is normally 120VAC.

Thus the open loop gain can be found to be

$$|G_{V.O.L.}|_{dB} = 20 \log \frac{P_{IN}}{2\pi f C_{OUT} V_{OUT\ DC} \Delta V_{EAOUT}} \quad (22)$$

where:

$G_{V.O.L.}$ = open loop response for the voltage error amplifier.

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The above expression gives the response of the magnitude with respect to frequency. The response has a -20dB/decade slope and a constant phase lag of 90 degrees.

A suitable error amplifier configuration is shown in Figure 7.

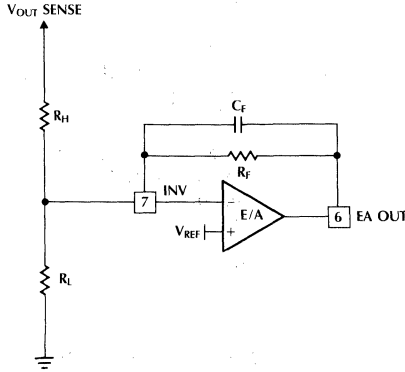


Figure 7. Error amplifier configuration.

Calculation of the output voltage sense resistors R_h and R_l : We have to pick a value for the output voltage under full load. The load regulation of the PFC can be expected to be 15 to 30V. Therefore if we pick a minimum output that is high under full load, there is the danger that under no load conditions the output voltage will be over 400V. Normally a minimum value of 370V will result at a high value of less than 400V. The reason for this seemingly poor regulation is the configuration of the error amplifier with a feedback resistor R_f that is close in value or less than R_h . Much better DC regulation can be obtained by using a blocking capacitor in series with R_f , but it will degrade the transient response of the circuit introducing a bounciness to the input current under transient conditions.

The output voltage of the error amplifier should be designed for 4 to 5 volts maximum at full load. Higher voltage gives better noise immunity and dynamic range. However that means the output voltage will have a larger variation due to its influence on the output voltage dividers. A good place to start is 4.4V. A value for R_h is picked that is normally between 680K Ω and 1M Ω . In this case we are going to pick 825K Ω . The value of the feedback resistor is found based on the loop design criteria. With these two values and minimum output voltage defined, R_l can be calculated using the following formula.

$$R_l = \frac{5R_h R_f}{R_f (V_{OUT\ MIN} - 5) - 0.6R_h} \quad (23)$$

See Figure 8 for definitions of the parts.

With the above chosen and calculated values one now can calculate the maximum output voltage under no load conditions as follows

$$V_{OUT\ MAX} = R_h \left(\frac{4.3}{R_f} + \frac{5}{R_l} \right) + 5 \quad (24)$$

Calculation of the OVP components: The sense resistors for the OVP circuit are easier to calculate. The voltage at which point the OVP circuit will act is being determined in part by the maximum tolerable voltage at the output before damage due to overvoltage that can occur. A good rule of thumb which may not be applicable in all cases is to set a voltage that is 10 to 15V higher than $V_{OUT\ MAX}$ as calculated by the above expression. Therefore $V_{OVP} = V_{OUT\ MAX} + 10V$.

OVP protection is facilitated by connecting a voltage divider to pin #11. The high side of this divider is connected to the output terminals of the PFC and the low side to ground. For the time being we are going to call these two resistors R_{OVPh} and R_{OVPl} . We are going to assume a value for the high side resistor and calculate the value of the low side. For that purpose one can use the following formula

$$R_{OVPl} = \frac{5R_{OVPh}}{V_{OVP} - 5} \quad (25)$$

The OVP pin on ML4821 is a multifunction pin. Pin #11 is also used for remote shutdown. When this pin is pulled to ground the IC shuts down. The pin can be pulled to ground using a small signal FET or bipolar transistor such as the 2N2222. Due to this multifunctionality, the pin should be biased higher than 1.0V whenever the part needs to be operated without the input power applied. **Extreme care should be exercised however when input power is applied.** It should be made sure that the voltage on this pin reflects the correct divided down output voltage for safe operation.

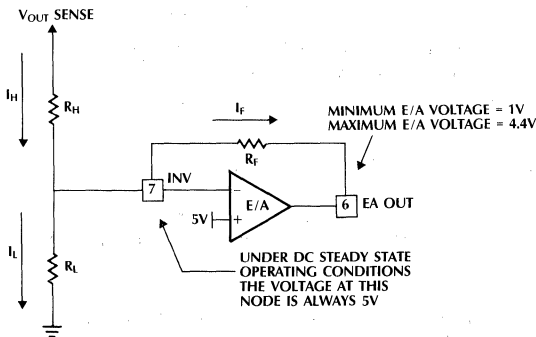


Figure 8. DC part of the feedback circuit. It is used for sense component calculations

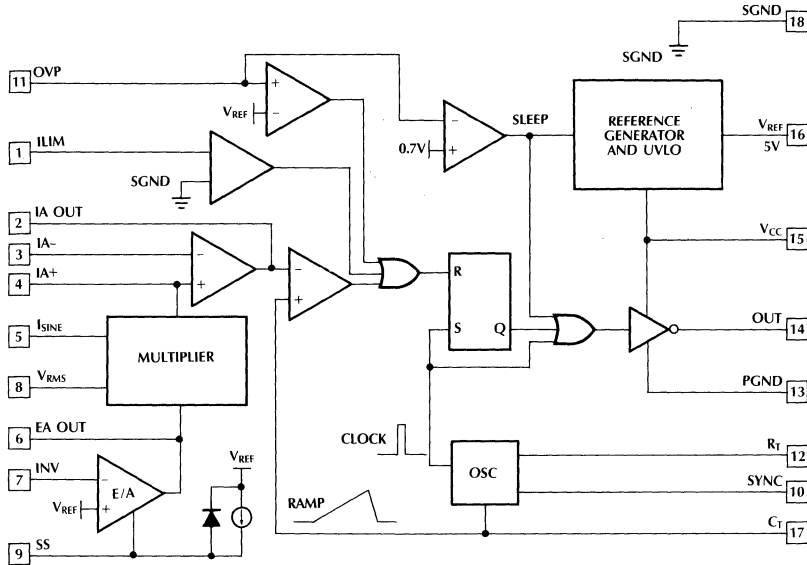


Figure 9. Block diagram of the ML4821

Block Diagram of the ML4821

Figure 9, shows the block diagram of the ML4821. So far we have covered most of the functions of this IC. But there are additional functions to extend its usefulness in various applications.

One of them is the SYNC function that enables the IC to frequency lock to an external oscillator. Pin #10 is reserved for this function. A positive pulse on this pin of 2V or higher resets the oscillator's comparator and initiates a discharge cycle for the timing capacitor connected to pin #17. For proper operation however the ML4821 oscillator should be set to operate at a frequency that is roughly 10-15% lower than that of the external driving source.

The rectangular block labeled multiplier combines both the multiplier function and the gain adjustor functions that we mentioned earlier.

Another useful function is the Soft Start function. This may be a useful function in some applications where controlled output voltage rise is desired. To use this function effectively however it is necessary to have an auxiliary bias power supply able to supply and maintain power to the control circuitry while the output voltage of the PFC is rising up slowly. The amount of the Soft Start capacitance required is a function of the delay in the output voltage rise time and the internal charging current.

The undervoltage lockout function of the IC can be used for off-line start-up as shown in Figure 10.

The Boost Inductor

One of the key components in the PFC is the boost inductor. The value of this inductor affects many other

design parameters. Most of the current that flows through this inductor is at low frequency (assuming low percentage ripple). This is particularly true at the lowest input voltage where the input current is highest.

Normally the acceptable level of ripple current is between 10 and 20%. For operation at 100KHz the following formula will produce acceptable results

$$L = \frac{300}{P_{OUT}} \text{ mH} \quad (26)$$

The peak to peak ripple current for any input output voltage combination can be found by using the following formula

$$\Delta I_{L-P-P} = \frac{V_{IN}(V_{OUTDC} - V_{IN})}{f L V_{OUTDC}} \quad (27)$$

where:

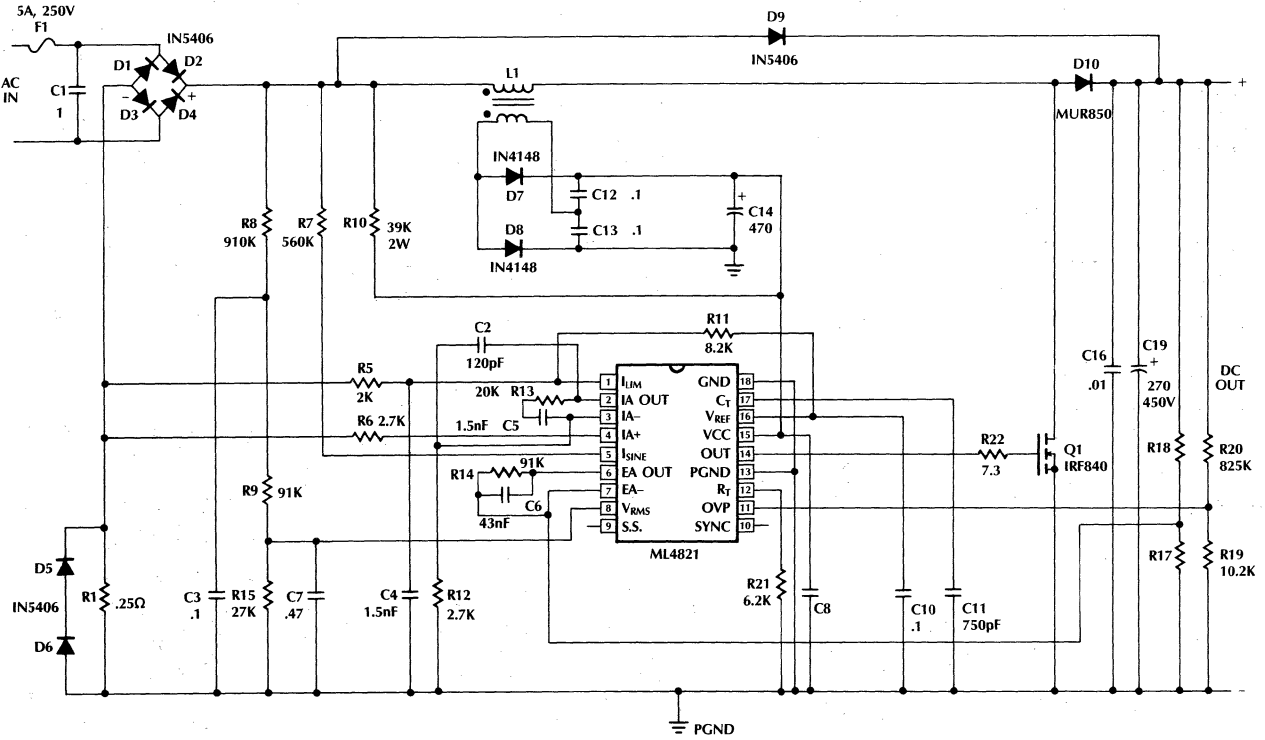
V_{IN} = peak value of the input full wave rectified waveform.

f = operating frequency.

For an output voltage of 380V the maximum peak to peak ripple happens when the input voltage is 134VAC and its value can be calculated using

$$\Delta I_{L-P-P \text{ MAX}} = \frac{V_{OUTDC}}{4fL} \quad (28)$$

This ripple current which has a triangular shape will produce a ripple voltage at the switching frequency and its harmonics on the input impedance.



NOTES: 1. CAPACITOR VALUES IN μ F.
 2. PC BOARD OF THIS CIRCUIT AVAILABLE. CONTACT YOUR LOCAL MICRO LINEAR DISTRIBUTOR (ML4821-EVAL).

Figure 10. Schematic diagram of the 200W PFC

The Input High Frequency Bypass Capacitor

This capacitor which should normally be at the line side of the input bridge rectifier helps to bypass the high frequency ripple current. Its impedance is few Ohms at the switching frequency. Therefore there is a need for additional filtering at the input, if differential conducted noise specifications are to be met. Note that this capacitor has to be an approved across the line type (an X-type capacitor).

II. DESIGN OF A 200W, 100KHz PFC

The complete schematic diagram of a 200W PFC is shown in Figure 10. We are going to cover in detail the pin by pin design of this PFC. We will use the formulas and the procedures presented in the previous sections. Although the design is at 200W it can easily be extended for power levels above that.

We are going to start the design by choosing the main components such as the boost inductor L1 and output capacitance.

The Boost Inductor: We can use (26) for the calculation of the inductor value

$$L = \frac{300}{200} \text{ mH} = 1.5 \text{ mH} \quad (29)$$

The maximum peak current that this inductor will see is simply the peak of the input sinusoid plus 1/2 the ripple current due to the switching action at 85VAC. Assuming 90% efficiency the input power will be 222W. This results in an input RMS current of approximately 2.6Arms with corresponding peak value of 3.7A. The peak to peak ripple current at 85VAC is 540mA. Therefore the peak inductor current will be 4.0A. The choice for the core material should be such that the inductance value will not change when this current passes through the winding.

It is important to remember that the higher the inductance, the lower the ripple current, which in turn means less filtering required on the input line to meet line conducted noise requirements. It also means lower core losses. The cost is more number of turns.

Good candidates for core materials are:

- Powder Iron Cores
- Mollypermalloy Cores
- Gapped Ferrite Cores, provided that the gap is not excessive.

Normally for any reasonable core material, core loss is not an issue due to the large number of turns required for such an inductor. The critical parameter is the change in their permeability under high current excitation, and a large number of turns. Therefore a careful analysis should be made to determine suitability of a core material for the given application.

For the present application we are going to choose a powder iron core of toroidal form. The core material is from Micrometals Inc., and the part number is T184-40 and it will contain 102 Turns. The inductor will maintain

approximately 80% of its zero current inductance at 4.0A. Therefore the inductance value will drop to 1.2mH at 4.0A. This will be the value that should be used to recalculate the peak to peak ripple current when time comes to design an input filter for the PFC. Also because the ripple current will increase, the current limit point should be set higher to account for this variation.

The Output Storage Capacitor: At an earlier section we mentioned the criteria for the selection of this capacitor. Lets assume that we need a hold-up time of 20msec and the output voltage is allowed to drop from 370V to 330V before regulation is lost in the driven switching power supply. We can use (18) to calculate a capacitance value

$$C_{OUT} = \frac{2 \times (200W) \times (20\text{msec})}{(370V)^2 - (330V)^2} = 285\mu\text{F} \quad (30)$$

The rated voltage of this capacitor should be at least 450V. The closest standard value offered by UNITED CHEMI-CON is 270μF type SMG with a voltage rating of 450V (a 330μF can also be used). Note that two capacitors of lower voltage rating can also be connected in series (i.e., 250V) provided that shunt ballasting resistors are also used.

The Output Diode: The output diode (D10) should be an ultra fast type capable of supporting the peak input current for a couple of milliseconds. Power dissipation is the limiting factor. For this design an MUR850 was chosen. Note that various manufacturers may be working on diodes with better reverse recovery characteristics.

Surge Bypass Diode: This diode labeled D9 on Figure 10 helps to bypass surges at the input line during start-up. This prevents possible saturation of inductor L1.

Output Circuit Very High Frequency Bypass: Capacitor C16 serves this purpose. It is used to control the output di/dt loop. It can be a high voltage high frequency ceramic type of 0.01μF.

Oscillator Circuit: Pins #12 and 17.

Timing resistor (R21): The choice for this resistor sets both the charging current for the timing capacitor, and some other internal currents. One of them is the maximum multiplier current (see formula 17). For a PFC operating at 100KHz a typical value for this resistor is 6.2K.

Timing capacitor (C11): For details on how to calculate its value refer to data sheet. For this application its value is 720pF.

Gate Drive: Pin# 14

The gate driver of the IC can directly drive power MOSFETs, normally a series resistor is used to damp any oscillations that may arise due to parasitic trace inductances and the gate capacitance. Its value should be chosen such that it will not result in excessive switching losses. If two paralleled MOSFETs are driven then their gates should be decoupled using two individual gate resistors.

For this example a gate resistor of 10Ω was used.

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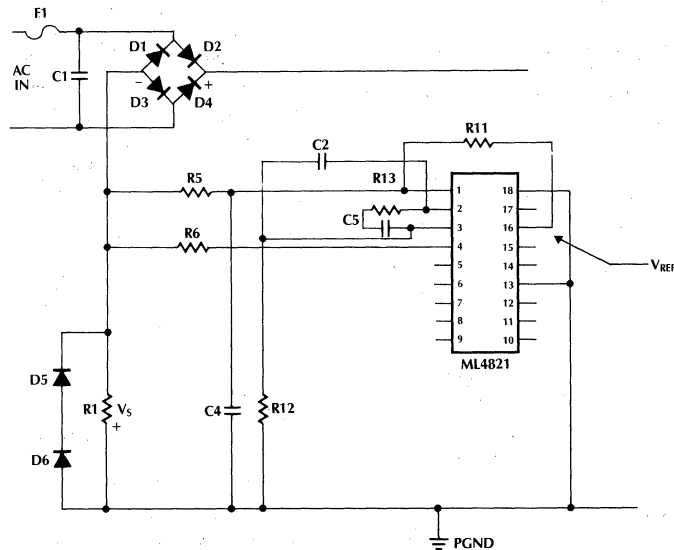


Figure 11. Current limit and current sense circuit connections

Depending on the layout a Schottky diode may be necessary across the gate drive to ground due to substrate current injection which can produce unpredictable behavior. The cathode should be connected to pin #14 and anode to ground. It should be placed as close to the IC as possible. Substrate current injection occurs when an output pin is forced more than about 0.5V below ground.

Power and Signal Grounds: Pins #13 and 18

These two grounds should go to ground plane and they should be connected together with the shortest possible trace length.

V_{REF}: Pin #16

The V_{REF} pin of the IC should be decoupled very well with a high quality ceramic capacitor. A typical value is 1μF. For higher power levels (P_{OUT} > 500W) additional capacitance may be required for proper operation.

Overvoltage Protection: Pin #11

This protects against accidental increases of the output voltage. As soon as a voltage higher than the set limit is detected the IC stops sending pulses to the MOSFET, until the voltage has dropped to safe limits. In a boost regulator if the voltage loop ceases to operate the only way to limit the output voltage from rising to destructive levels is the overvoltage protection circuit.

Also due to the low bandwidth of the voltage control loop there may be situations that the voltage may rise to destructive levels such as sudden removal of the output

load. Under those conditions the OVP circuit will activate preventing further rise.

Formula (25) is used to set the activation limit of the OVP circuit.

From Figure 10, assuming the value of R₂₀ = 825K, R₁₉ can be found as follows

$$R_{19} = \frac{5 \times (825 \times 10^3)}{400 - 5} = 10.2K \quad (31)$$

V_{CC}: Pin #15

This is the supply pin of the IC. Normally a quality ceramic capacitor should be connected to this pin as close as possible to the body of the IC for effective decoupling. For low power applications (< 500W) 1μF may be sufficient, but for higher power applications experience shows that two individual capacitors could be necessary.

In order to facilitate off-line start-up the IC has a large Under Voltage Lockout hysteresis. For bootstrapped operation a reservoir capacitor (C14) is charged with a small current through R10 which is connected to the input high voltage line. When the voltage on this capacitor reaches 16V the IC “wakes-up”. A winding on L1 (see Figure 10) “steals” part of the energy to supply the current requirements of the IC. This way the circuit continues to operate.

The time that it takes initially for the voltage to reach 16V, and therefore for the circuit to start, is a function of the resistor R10. This is a power resistor and for as long as power is applied it wastes power, usually about 2W.

The value of C14 is being determined by the current requirements of the circuit. C14 has to be sufficiently large for the circuit to bootstrap. However it should not be too large because it will take a long time initially to charge it to 16V, and turn on the IC.

Current Limit: Pin #1

As in every switching regulator, there is a need for current limit in the PFC as well. To understand its operation better, let's look at Figure 11.

The current limit point is set by R11 and R5. The value of R5 can be selected first. Then the value of R11 is calculated using the following. In this case R5 = 2.0K. Assuming a current limit voltage of 1.2V across the sense resistor.

$$R11 = \frac{V_{REF}}{V_S} R5 = \frac{V_{REF}}{I_{IN} R1} R5 \quad (32)$$

where:

I_{IN} = current at which limiting action should start.

R1 = sense resistor.

$$R11 = \frac{5}{1.2} (2.0K) = 8.2K$$

For proper operation, the sense voltage across R1 that triggers the current limiting action should be greater than the sense voltage produced at low line and full load. In this case current limiting action starts when the sense voltage is 1.2V. Therefore at full load and at low line the sense voltage should be less than 1.2V. A 15% less voltage corresponds to 1.0V.

Current Sense Circuit: Pins #3 and 4.

These two pins are used to sense the return current of the power circuit. The average value of this current is forced to follow the sinewave shape as being determined by the IC. Pin #4 is at the same time connected to the output of the multiplier. As you recall the maximum value of the multiplier output current was set to be 400µA by the timing resistor R21.

In the current limit section we said that at maximum power and low line the sense voltage should be 1.0V. Therefore R6 should be chosen such that it will produce 1.0V at 400µA.

$$R6 = \frac{1.0V}{400\mu A} \cong 2.7K \quad (33)$$

Earlier we calculated that the expected maximum peak current is 4.0A. Using this value R1 can now be calculated

$$R1 = \frac{1.0V}{4.0A} \cong 0.25\Omega \quad (34)$$

The value of R12 which is a feedback resistor for the current amplifier is chosen to be equal to R6. This is set to cancel out the input bias current of the current amplifier. Hence R12 = 2.7K.

Design of the Current Loop Amplifier Components: Pins #2, 3, and 4.

The design of the current loop is one of the most critical tasks in the overall design. To do that we have to have knowledge of the open loop response of the power stage. Equation (12) gives this response which is plotted in Figure 12.

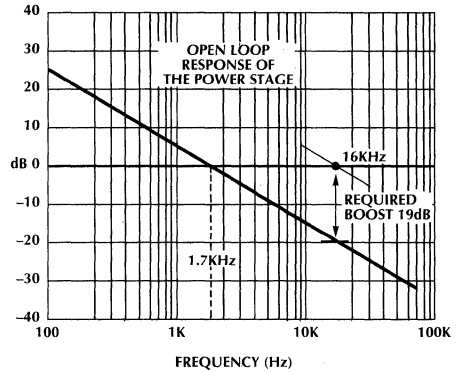


Figure 12. Open loop response of the power stage, and required boost for 16KHz loop crossover.

The required gain boost for unity gain crossover at 16KHz, as set by (13), can be calculated by using (12).

$$\frac{380 \times 0.24}{2 \times \pi \times 5.5 \times (16KHz) \times (1.5mH)} = 0.11 \text{ or } -19dB \quad (35)$$

An appropriate current amplifier response that will accomplish this is shown in Figure 13. The equations that give the asymptotic gain response in each one of the three regions are given below.

$$[1] \dots \dots \dots |G|_{dB} = 20 \log \frac{1}{2 \pi f R12 C5} \quad (36)$$

$$[2] \dots \dots \dots |G|_{dB} = 20 \log \frac{R13}{R12} \quad (37)$$

$$[3] \dots \dots \dots |G|_{dB} = 20 \log \frac{1}{2 \pi f R12 C2} \quad (38)$$

In order to complete the design of the current control loop, we have to calculate the feedback component values. Using (37) the value of R13 is found as

$$20 \log \frac{R13}{R12} = 19dB \Rightarrow R13 \cong 20K \quad (39)$$

$$C5 = \frac{1}{2 \times \pi \times f \times R12 \times 10^{\frac{19}{20}}} \Rightarrow C5 \cong 2.3nF \quad (40)$$

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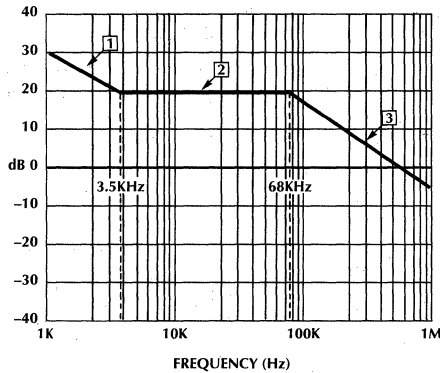


Figure 13. Desired current error amplifier response.

where:

$$f = 3.5\text{KHz}$$

$$R12 = 2.7\text{K}$$

$$\text{Gain boost} = 19\text{dB}$$

$$C2 = \frac{1}{2 \times \pi \times f \times R12 \times 10^{\frac{19}{20}}} \Rightarrow C2 \cong 120\text{pF} \quad (41)$$

where:

$$f = 68\text{KHz}$$

With the values of the feedback components now calculated we can plot the overall closed loop response of the inner current loop. Keep in mind that logarithmic slopes and gain values just need to be added to get the overall response. The result is shown in Figure 14.

Calculation of R7, the I_{SINE} Resistor: Pin# 5.

In a previous calculation we have assumed that the maximum voltage of the error amplifier is 4.4V. The minimum voltage under normal operating conditions is about 0.8V. That necessitates a change in the output voltage of the error amplifier of 3.6V from no load to full load. Note that due to feedforward compensation the output of the error amplifier will not change for line variations. It is important also to note that the amplifier output is capable of going to 7.5V.

Equation (21) can be solved for R7.

$$R7 = \frac{\Delta V_{EAOUT} \times K \times V_{RMS}^2 \times R6}{R1 \times P_{IN}} \quad (42)$$

Substituting the known values in the above we get

$$R7 = \frac{3.7 \times 0.23 \times 120^2 \times 2200}{210 \times 0.24} = 535\text{K} \quad (43)$$

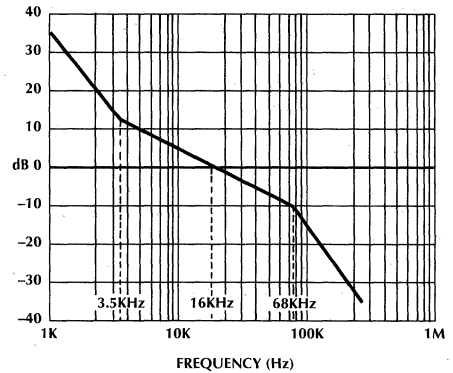


Figure 14. Overall closed current loop response.

The nearest standard value for R7 is 560K. With this resistor value, the I_{SINE} current should be calculated to check the input current range corresponding to the line input voltage range. The multiplier requires that the I_{sine} current be less than 500μA as suggested in Fig 8 of the data sheet.

Design of the Voltage

Loop Amplifier Components: Pins #6 and 7.

Equation (22) gives the magnitude of the open loop gain. The response has a -20dB/decade slope with constant 90 degree phase lag. To proceed with the design of the error amplifier feedback components we have to pick the unity gain crossover frequency. In this application we are going to crossover the 0dB line at 10Hz.

The frequency where the open loop response crosses over the 0dB (unity gain) line can be found by solving (22) for *f*.

Figure 15, shows the open loop along with desired amplifier responses.

$$f = \frac{P_{IN}}{2 \pi C_{OUT} V_{OUTDC} \Delta V_{EAOUT}} \quad (44)$$

$$f = \frac{210}{2 \times \pi \times 270 \times 10^{-6} \times 380 \times 3.7} = 88\text{Hz} \quad (45)$$

Now we can complete the design of the error amplifier feedback components. For unity gain crossover at 10Hz the amplifier needs to have an attenuation of 19dB at 10Hz.

$$20 \log \frac{R14}{R18} = -19\text{dB} \Rightarrow R14 = 92\text{K} \quad (46)$$

Then we calculate the value of C6. For that we look up the asymptotic break point of the response curve, which in this case is 40Hz.

$$C6 = \frac{1}{2 \pi f R14} = \frac{1}{2 \times \pi \times 40 \times (92\text{K})} = 43\text{nF} \quad (47)$$

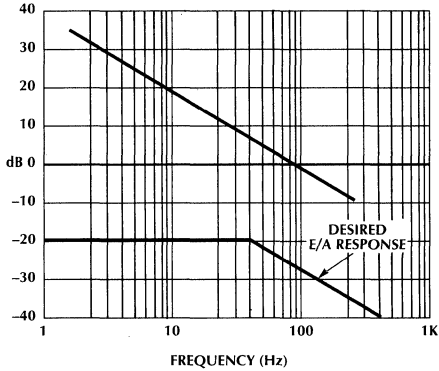


Figure 15. Open loop and desired error amplifier response.

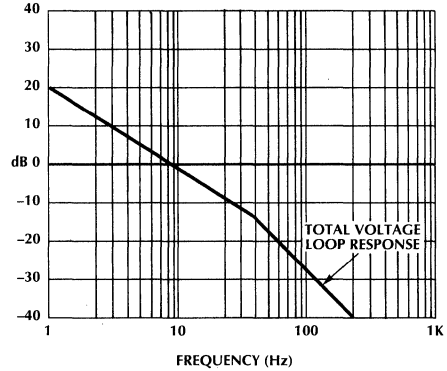


Figure 16. Total voltage closed loop response.

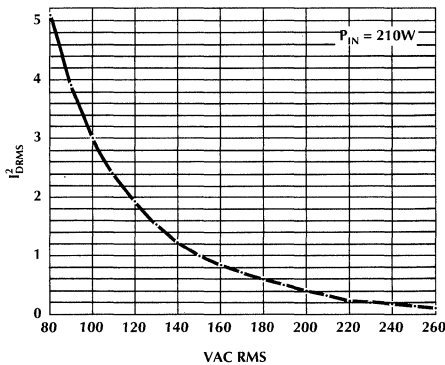


Figure 17. Square RMS drain current versus input RMS voltage.

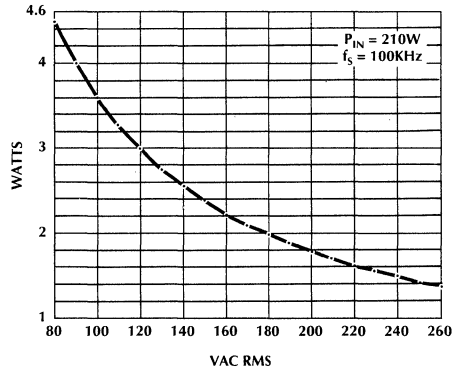


Figure 18. Switching power loss for a 200W PFC operating at 100KHz versus the input voltage.

Figure 16, shows the overall closed voltage loop response.

Voltage sense resistor calculation:

R18 was already assumed to be 825K. R17 can be calculated using (23).

$$R17 = \frac{5 \times (825K) \times (91K)}{91K(370 - 5) - 0.6(825K)} = 10.38K \quad (48)$$

Losses In The Power MOSFET Q1:

There are three kind of losses in the power MOSFET, these are listed below:

1. Conduction Losses, due to the conduction of the drain current.
2. Capacitive Losses, due to the charge and discharge of the total drain source capacitance. This is a switching loss.
3. Turn-On and Turn-Off Losses, these are also switching losses.

The conduction losses can be calculated by using (49). Equation (49) gives the RMS value of the drain current which can be used to calculate the conduction losses. As can be expected it is a function of the input power and input and output voltages.

$$I_{DRMS} = \frac{2P_{IN}}{V_{INRMS}} \sqrt{\frac{1}{4} - \frac{2\sqrt{2}V_{INRMS}}{3\pi V_{OUT}}} \quad (49)$$

The assumption made in the derivation of the above is that the ripple to average current ratio is very small which is normally true for this kind of PFC. It happens in the low input voltage range. In reality the nonzero ripple will increase the value of the calculated RMS current by a small amount.

Figure 17 gives the value of the square of the RMS drain current with respect to the input RMS voltage. One can use this graph to calculate the power loss due to conduction in the MOSFET. This is simply

$$P_{CMOSFET} = I_{DRMS}^2 \times R_{D-SON} \quad (50)$$

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For our design example the minimum line was 85VAC. Using the graph above that corresponds to a value of 4.5. Using an IRF840 type MOSFET and assuming that its ON resistance at the operating temperature will be 50% higher than its 25°C resistance, the conduction power loss will be 5.4W. At 120VAC the power loss will be 2.2W, and at 220VAC will be 0.34W! As you can see it is quite difficult to have an optimum circuit at the same time with wide input voltage range. For specific applications where the input voltage range is narrow it is more advantageous to have the output voltage closer to the maximum peak of the input voltage.

Lets give an example with respect to the last statement made in the paragraph above. Lets assume that the power level is still the same but that the operating voltage range is 85VAC to 135VAC and that the output voltage is 200VDC. With these operating parameters (50) yields a value of 2.99, which corresponds to power loss of 3.6W, Compare this to 5.4W in the above example.

Normally efficiency measurements are made at nominal operating voltages i.e., 120VAC. However the PFC should be able to function without failure at low line conditions. That necessitates careful selection of components and thermal design for good reliability.

Capacitive losses in the MOSFET are due to discharge of the total drain source capacitance. We use the term total because the drain gate capacitance contributes to this loss too. The losses can be calculated using (51).

If we assume a total capacitance of 350pF then (51) yields 2.5W.

$$P_{CAPD-S} = \frac{1}{2} C_{D-S} V_{OUT}^2 f_S \quad (51)$$

The calculation of switching losses are a little more difficult since they are a function of many things such as gate drive conditions that may include the physical layout. In any case an equation that can be used to give some indication of these losses is given below:

$$P_{SWITCHING} = \frac{2\sqrt{2}V_{OUT} f_S P_{IN} t_{TR}}{\pi V_{INRMS}} \quad (52)$$

where:

t_{TR} = transition time

f_S = switching frequency (= 100KHz in this case).

Assuming waveform symmetry during both turn-on and turn off transitions and ignoring possible secondary effects we can use (52) to calculate the switching losses. Our example was designed at 120VAC and 210W input with a reasonable value for the transition time of 50nsec. The resulting losses are approximately 3W.

As a final step lets add up all the losses in the MOSFET for 120VAC. The resulting total loss is 7.7W. If this yields an unacceptable efficiency, an optimized MOSFET switch should be used. Loss calculations can be made using the three equations. Note that the derivation of these equations is rather long and tedious requiring careful modelling.

III. EVALUATING A PFC CIRCUIT

This part of the application note will give practical information that may be useful when trying to get the bread board up and running to meet required specifications. It will show that measuring power factor, harmonic current content, and efficiency may impose new challenges to even experienced power supply design engineers. Also it will contain some performance data that may serve as a reference point.

Waveforms

Operating waveforms are shown in the following figures. They are taken with output at 200W and input at 120VAC. Figure 19 shows the power factor corrected input current waveform. Upper waveform is current at 1A per division. Voltage and current are in phase and identical. Figure 20 shows the inductor waveform. The shaded portion of the upper waveform indicates the ripple current riding on top of a rectified sinusoidal current. The lower waveform is an expanded view of the upper waveform. Figure 21 shows the current limit waveform on pin 1. As the input current increases, the valleys of this waveform approach zero volts. However, because the multiplier current is limited to 400µA, the current waveform will sag before the current reaches the current limit level. Current limit level is reached during a transient condition when the inductor current increases rapidly before the voltage loop can compensate for it. Figure 22 shows the output of the current amplifier (pin #4). It sets up the trip points of the PWM comparator.

REMINDER: The OVP pin requires at least 0.7V for the chip to begin operation.

Layout

Board layout is critical in this application as it is in any power control circuits. One must pay close attention to the high current circulating paths. The control circuitry and it associated ground plane should be away from the high current power paths as much as possible. Current should be steered away from the high impedance nodes such as the input to both error amplifiers as well as to both current limit and OVP comparators. Also magnetic fields generated by the magnetics components as well the switching power components can inject noise into the high impedance nodes such as that to the current limit comparator. The heat sink should either be grounded or at least AC coupled to ground by a high frequency ceramic capacitor and kept as far away from the IC as possible.

Power Factor

Input power factor, harmonic current content, and waveshape are all used when describing the performance of a power factor circuit. It is important to keep in mind that regulatory specifications such as the IEC555 for Europe will require that just the harmonic current levels meet certain limits. The proposal currently sets these limits (for Class D) as a function of power level up to 300W. Above 300W, the limits are absolute. Thus even a low power factor number at high input line voltage can easily meet the limits since the input current level is

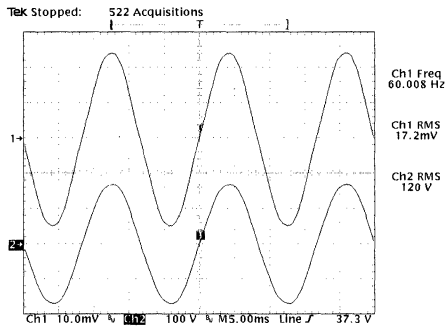


Figure 19. Input current and waveform.

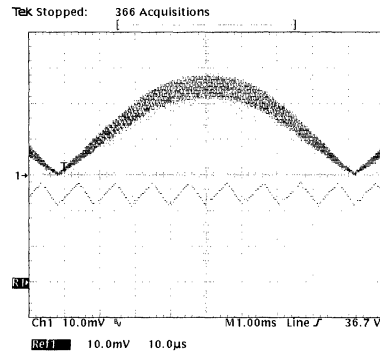


Figure 20. Inductor current magnified 1K times.

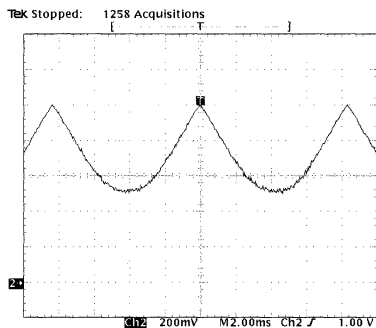


Figure 21. Current limit comparator.

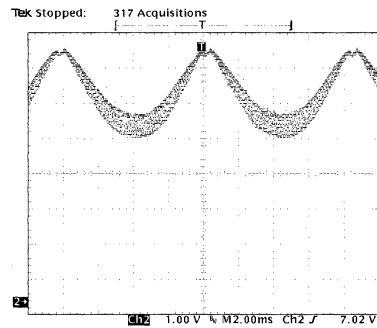


Figure 22. Current amp output.

proportionally low. Nevertheless power factor is a good parameter for measuring the capabilities of a power factor circuit.

To measure power factor, one must have a very reliable power meter that accurately measures both apparent power (product of RMS voltage and RMS current) and true average power. Some of the older model meters measure power factor by determining the phase angle between current and voltage waveform. Obviously this will not work for our purposes.

True average power for distorted AC waveforms can only be determined when the current and voltage are simultaneously sampled which is then multiplied and integrated. There are several meters on the market that are possible candidates for this purpose. But none that we have looked at seem to do the job as well as the one offered by Voltech. The Voltech PM1000 measures true power by sampling the waveform and analyzing the analog signal using digital methods. It uses DSP to filter, multiply, and integrate both voltage and current simultaneously.

A study was done by comparing the results of a power factor measurement with the Voltech to that of another well known meter manufacturer who perform the multiplication and integration of the power signal VIA

analog methods. Both meters gave nearly identical measurements when evaluated against a calibrated standard reference unit. However when the units were set up in the lab to measure power factor of an actual switching power circuit, only the Voltech gave the expected measurement readings. For whatever the reason the analog meter gave grossly false readings. Noise in the switching circuit is prime suspect. This does not suggest that meters with analog circuits can not be designed to reject noise and work effectively whatever the waveshape and environment. But one should be very careful about which meters are trustworthy when it comes to measuring power with non-sinusoidal and noisy waveforms.

Harmonic current

Harmonic current content can be measured using the Voltech PM1000. It gives a percentage of the fundamental up to the 13th harmonic. Results of this method was compared to that of the HP spectrum analyzer 3585A. The results were quite close. The Voltech PM3000 which is a three phase power meter measures harmonic currents to the 99th. Results show that harmonic current level beyond the 13th harmonic remain low throughout the spectrum. The proposed IEC555 specification is expected to require harmonic current content conformity to the 40th.

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Efficiency

The three power devices add up to almost all of the losses. The inductor, MOSFET, and diode were selected to meet desired performance specifications at relatively low cost. An IRF840 seemed to provide the best performance at these power levels. A larger MOSFET (APT5025) did improve efficiency at low input voltage. However its larger output capacitance increased switching losses significantly at 100KHz. A very common 8A ultra-fast rectifier (MUR850) was used for the output diode. This oversizing was done because larger diodes have lower reverse recovery times for a given current level. New diodes have been introduced with lower reverse recovery times. A low cost ferrite EC core was used with AWG21 wires. There was very little temperature rise in the magnetic material. However the wires heated up at low input voltage where RMS current is significant. A low cost powdered iron with AWG20 magnet wire was also tried. Efficiency decreased about one percent.

Efficiency measurement requires accurate measurement of the output average power as well as the input average power. Because the output voltage as well as output current is DC, one might just measure the two readings using typical lab bench top DMM. However for one

reason or another, this output power measurement did not match the measurement results of the average power reading of the Voltech power meter. The output power measurement was off by a factor of almost 6%. The bench top DMM used was Fluke's 8050A. To get truly accurate efficiency measurements, the same power meter should be used to measure both the output power as well as the input power.

Efficiency measurements for this application note was done on a Voltech PM3000 three phase power meter. This meter allows the connection of both the input power as well as the output power into one metering unit. Efficiency can be easily determined by measuring the input and output power with a push of a button. Accuracy however does not come easy even with this approach. The output power reading randomly varied up to 3% under steady state conditions.

Two different model meters from the same manufacturer gave current reading that was off by two to three percent. This might indicate that even if one was to go out of the way to obtain two same meters from the same manufacturer, efficiency measurements as well as other parameters may not be guaranteed to be as accurate as one may require.

Table 1. Performance Data

	PO = 50W					PO = 100W					PO = 150W					PO = 200W				
	V _O	P _O	P _i	PF	n	V _O	P _O	P _i	PF	n	V _O	P _O	P _i	PF	n	V _O	P _O	P _i	PF	n
	I _{FUNDAMENTAL}					I _{FUNDAMENTAL}					I _{FUNDAMENTAL}					I _{FUNDAMENTAL}				
%HARMONIC	3rd	5th	7th	9th	11th	3rd	5th	7th	9th	11th	3rd	5th	7th	9th	11th	3rd	5th	7th	9th	11th
	13th	15th	17th	19th	21st	13th	15th	17th	19th	21st	13th	15th	17th	19th	21st	13th	15th	17th	19th	21st
V _{IN} = 90VAC	395	54	62	0.99	0.87	387	105	117	0.99	0.90	379	153	166	0.99	0.92	371	197	218	0.99	0.90
	I _{FUND} = 0.67A					I _{FUND} = 1.3A					I _{FUND} = 1.8A					I _{FUND} = 2.4A				
	3.5	1.7	0.04	0.32	0.30	3.1	1.6	0.02	0.23	0.21	2.9	1.5	0.05	0.22	0.24	2.4	1.6	0.20	0.27	0.37
	0.06	0.05	0.05	0.01	0.11	0.04	0.03	0.06	0.01	0.01	0.09	0.18	0.18	0.12	0.13	0.24	0.24	0.24	0.15	0.15
V _{IN} = 120VAC	393	52	61	0.99	0.85	386	105	115	0.99	0.91	378	150	161	0.99	0.94	369	194	212	0.99	0.91
	I _{FUND} = 0.51A					I _{FUND} = 0.96A					I _{FUND} = 1.3A					I _{FUND} = 1.8A				
	3.7	2.2	0.12	0.46	0.37	3.2	1.9	0.05	0.29	0.25	3.1	1.9	0.01	0.21	0.26	2.9	1.9	0.09	0.15	0.12
	0.14	0.05	0.08	0.05	0.23	0.10	0.07	0.11	0.05	0.03	0.07	0.05	0.09	0.07	0.03	0.09	0.03	0.03	0.01	0.02
V _{IN} = 180VAC	394	52	60	0.96	0.87	385	105	114	0.99	0.92	378	151	163	0.99	0.93	369	190	207	0.99	0.92
	I _{FUND} = 0.34A					I _{FUND} = 0.64A					I _{FUND} = 0.87A					I _{FUND} = 1.2A				
	4.6	2.3	1.6	2.1	2.7	4.7	2.2	0.06	0.51	0.49	4.5	2.0	0.03	0.39	0.34	4.4	1.7	0.08	0.31	0.28
	1.9	1.0	0.87	0.95	0.66	0.18	0.17	0.12	0.12	0.08	0.11	0.12	0.12	0.10	0.04	0.11	0.09	0.09	0.04	0.02
V _{IN} = 220VAC	393	53	61	0.93	0.87	383	102	114	0.97	0.90	374	150	162	0.99	0.93	365	192	207	0.99	0.93
	I _{FUND} = 0.28A					I _{FUND} = 0.52A					I _{FUND} = 0.71A					I _{FUND} = 0.95A				
	5.4	2.4	0.21	1.0	1.0	5.3	2.8	0.02	0.77	0.51	4.9	2.3	0.08	0.56	0.35	4.7	2.3	0.06	0.47	0.33
	1.8	2.8	2.7	2.0	1.0	0.41	0.47	0.38	0.25	0.15	0.21	0.25	0.23	0.17	0.06	0.15	0.21	0.18	0.18	0.04
V _{IN} = 260VAC	394	53	62	0.87	0.85	387	103	15	0.95	0.90	379	153	164	0.97	0.93	372	195	208	0.98	0.94
	I _{FUND} = 0.24A					I _{FUND} = 0.44A					I _{FUND} = 0.61A					I _{FUND} = 0.82A				
	6.2	4.5	0.59	0.76	1.3	5.3	3.4	0.94	0.47	1.7	4.9	2.8	0.27	0.63	0.58	4.7	2.4	0.16	0.51	0.34
	0.96	1.1	0.36	1.5	2.3	1.4	0.80	1.0	1.4	1.0	0.26	0.35	0.53	0.24	0.10	0.24	0.26	0.38	0.14	0.01

The following are specifications to two inductors that may be considered for the 200W PFC converter.

Magnetics, Inc.
(412) 282-8282
OP44317 (Ferrite, EC)
 $N_p = 118$ turns, AWG21
 $N_s = 5$ turns, AWG30
gap = 1.7mm

Micrometals
(714) 630-7420
T184-40 (Powdered Iron, Toroid)
 $N_p = 102$ turns, AWG20
 $N_s = 3$ turns, AWG30

Performance Data

Table 1 shows the results obtained from an application circuit. Pertinent power and power factor measurements were taken as well as the harmonic current content as a percentage of the fundamental. It is intended to be a typical reference point in which to judge new designs. It is not unlikely that the performance can be improved and optimized via various methods.

Application Note 19

by Mehmet K. Nalbant

Phase Modulated PWM Topology with the ML4818

INTRODUCTION

One of the biggest goals in power supply design is to get the maximum amount of output power while maximizing efficiency and minimizing both the cost and size of the respective power supplies. There are many conflicting parameters in trying to do so. For example, let's examine how the size of a power supply is a strong function of the size of the passive storage elements. It is well known that the size of inductors and capacitors greatly depends on the operating frequency. The higher the frequency the smaller the inductors and capacitors necessary. Increasing the operating frequency is one thing, implementing it is another. It is also a well known fact that increased operating frequencies result in lower efficiencies in PWM controlled switched mode power supplies.

Increasing the frequency of PWM controlled power supplies was one solution in the reduction of the passive elements. That posed some limitations due to the nature of operation. Simultaneous conduction of high currents in the presence of high voltage during turn-on and turn-off times at high frequencies resulted in high switching losses. Thus violating one of the most important parameters of the switching power supply design that is the efficiencies were now lower than in lower frequency operation.

This application note will introduce the "Phase Modulated PWM Topology" that overcomes many of the shortcomings of conventional PWM topologies at high operating frequencies.

LOSSES IN SWITCHING POWER SUPPLIES

The typical losses in switching power supplies can be divided in two classes conduction losses and switching losses. The most common switching element used in modern switching power supplies is the power MOSFET. This device when enhanced for conduction has a finite channel resistance named $R_{DS(ON)}$. When current passes through this device conduction losses result which are proportional to:

$$P_C = I_{DS(RMS)}^2 R_{DS(ON)}$$

In addition to the conduction losses, due to the switching action of the device in the presence of high currents and high voltages, there are also switching losses. These losses can be further subdivided in turn-on, turn-off and capacitive discharge losses. Figure 1 shows how turn-off losses can result during switching in a simplified way.

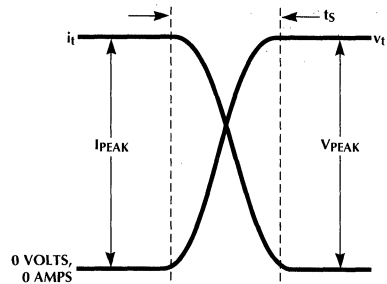


Figure 1. Waveforms in the switching element in a switching power supply during turn-off, assuming linear rise and fall.

The energy in turn-off instance can be found by integrating the product of the voltage and current waveforms over the complete switching interval i.e.,

$$W_{tOFF} = \int_0^{t_s} v_{t_i} i_t dt$$

Assuming linear waveforms and symmetry the above integral can be simplified to the following:

$$W_{tOFF} = \frac{1}{2} I_{PEAK} V_{PEAK} t_s$$

The total power lost therefore can be found by multiplying the above with the repetition rate, that is the switching frequency hence

$$P_{OFF} = \frac{1}{2} I_{PEAK} V_{PEAK} t_s f$$

To give an example suppose that the power switch switches 10 amps at 380 volts for 50nsec at 100KHz. The resulting power loss due to just this event would be 9.5 Watts. At 200KHz it would be 19 Watts and so on. This power loss must be dissipated by the switching element and poses a problem for the adequate removal of the generated heat. One can appreciate the losses at even higher frequencies.

Turn-on switching losses result in a similar manner. Assuming again symmetry the turn-on losses can be found by integrating the current and voltage waveforms over the switching interval.

$$P_{ON} = \frac{1}{2} I_{PEAK} V_{PEAK} t_s f$$

We are going to see later that the switching event is more complicated than what is depicted above due to the presence of parasitic inductive and capacitive components such as leakage and lead inductances and drain source capacitances.

To reduce the switching losses several methods were used such as dissipative and non dissipative snubbers. As the name implies dissipative snubbers dissipate their energy as heat, whereas the non-dissipative snubbers return their energy back to the input line, thus tending to be more efficient. In any case even with the use of non dissipative snubbers there remain problems that limit the maximum operating frequency.

Several respected institutions along with many manufacturers tried to find a way around the above problems. This resulted in the proliferation of several new power supply topologies with each one claiming to be the solution for operation at high frequencies.

Resonant power supplies thought to be a possible solution had their own share of problems. Although not a new technology they found a home in some applications. They were never really widely accepted by the industry. Part of the reason was the absence of analytical tools for the analysis and design and suitable controllers. Thanks to the efforts of many people they are better understood today but most of the manufacturers are still reluctant to put products on the market based on this technology.

Resonant power supplies encompass a wide range of topologies and they can be subdivided into three major subclasses. These are as follows:

- 1) Current resonant or zero current switching ZCS.
 - 2) Voltage resonant or zero voltage switching ZVS.
- and
- 3) Multi-resonant, in the majority of which both the current and voltage is resonant.

It is beyond the scope of this application note to give an exhaustive explanation for each type of the resonant conversion techniques. For more information, the interested reader can draw on the vast amount of technical papers published over the last few years. It suffices to say that among the resonant conversion techniques one that is of particular interest for high frequency operation is the zero voltage switching, or ZVS.

Switches such as power MOSFETs have a drain-source capacitance of several hundred picofarads. When this capacitance charges and discharges, energy is lost that results in power loss. Figure 2, shows a typical power switch configuration.

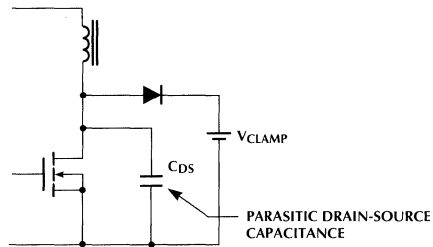


Figure 2. Typical switch stage of a switching power supply, with the parasitic drain-source capacitance shown explicitly.

The amount of power lost can be calculated by using the following formula

$$P = \frac{1}{2} C_{DS} V_{CLAMP}^2 f$$

As an example consider the following case $C_{DS} = 500\text{pF}$, $V = 380\text{ volts}$, $f = 500\text{KHz}$ a circuit with these parameters results in a power loss of 18 watts. Therefore one can see the importance of capacitance C_{DS} .

In switching power supplies as we mentioned earlier it is often advantageous to use an external drain-source capacitance in the form of a snubber. This takes some of the burden of the switching loss from the switching device and puts it on the snubber circuit. The use of such capacitors further compounds the problem of capacitive discharge losses. If a way could be found to discharge the total drain-source capacitance non-dissipately then that would represent a solution to the problem. Figure 3, shows this concept, for the time being we are not going to discuss the actual implementation of such a circuit. It is evident from the diagram that the switching loss can be reduced to zero if the voltage were also zero. From these diagrams it is evident that turn-on and turn-off power losses will be zero. Total switching times are in the order of 100nsec.

Zero Voltage Switching techniques represent such a solution. There are some limitations and shortcomings even to these techniques. When ZVS is accomplished through resonance of the voltage waveform then the design and analysis of such power supplies is more complicated. As a rule of thumb the operating drain currents are also higher than in PWM controlled power supplies.

To summarize, the ideal power supply would be the one that doesn't have operating frequency limitations because of switching losses, would be easy to design and manufacture and will be cost effective utilizing each one of its components to their fullest extent. In the next section we will discuss such a topology that has many of these desirable characteristics.

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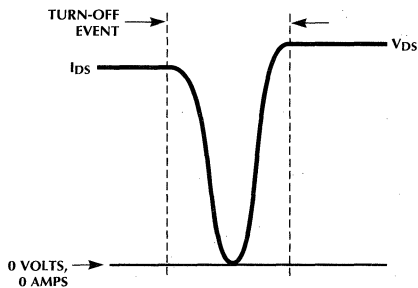


Figure 3a. Ideal turn-off waveforms of a ZVS switching element.

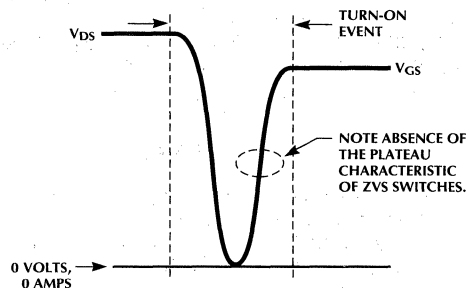


Figure 3b. Ideal turn-on waveforms of a ZVS switching element.

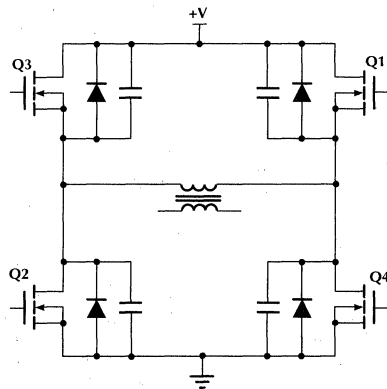


Figure 4. Typical H bridge power switch configuration as it is used in medium to high power switching power supplies.

PHASE MODULATED PWM TOPOLOGY (PMPT)

Phase modulated PWM Topology although not a panacea has many of the favorable characteristics mentioned above. It is a promising topology for medium to high power systems. Basically it is a full bridge topology with appropriately coordinated gate drive waveforms for each one of the MOSFETs in the H bridge. Its operating waveforms are very close to ideal. Turn-on and turn-off switching losses are almost eliminated. Operating drain currents are almost equivalent to those of a regular full bridge PWM topology, thus it does not require the use of expensive large die area MOSFET switches. The only difference is how the two topologies handle their respective switching events.

The analysis and design of the power circuit of the PMPT topology is identical to that of the classical PWM topology. Having said that, there are special set of considerations associated with the design of a high frequency high power transformer used in the PMPT.

The key idea behind the PMPT is that the voltage across the MOSFET is allowed to swing to "zero volts" just before the start of the next conduction cycle in the respective switches.

Figure 4, shows a typical H bridge configuration, the diodes and the capacitors across the MOSFETs are the intrinsic parasitic components present in these components. Typical values for the capacitors range anywhere from 100pF to 500pF for the larger devices. The reverse recovery time of the body diodes are in the range of 100nsec. In the figure snubber circuitry has not been shown.

The power switch section of the PMPT is identical to the one shown in Figure 4. To achieve PMPT operation the switches must be driven differently. In the regular PWM topology, gate drive is applied to the two diagonal switches based on the required duty cycle, then there is a period during which all switches are OFF (deadtime) and then gate drive is applied to the opposing diagonal switches.

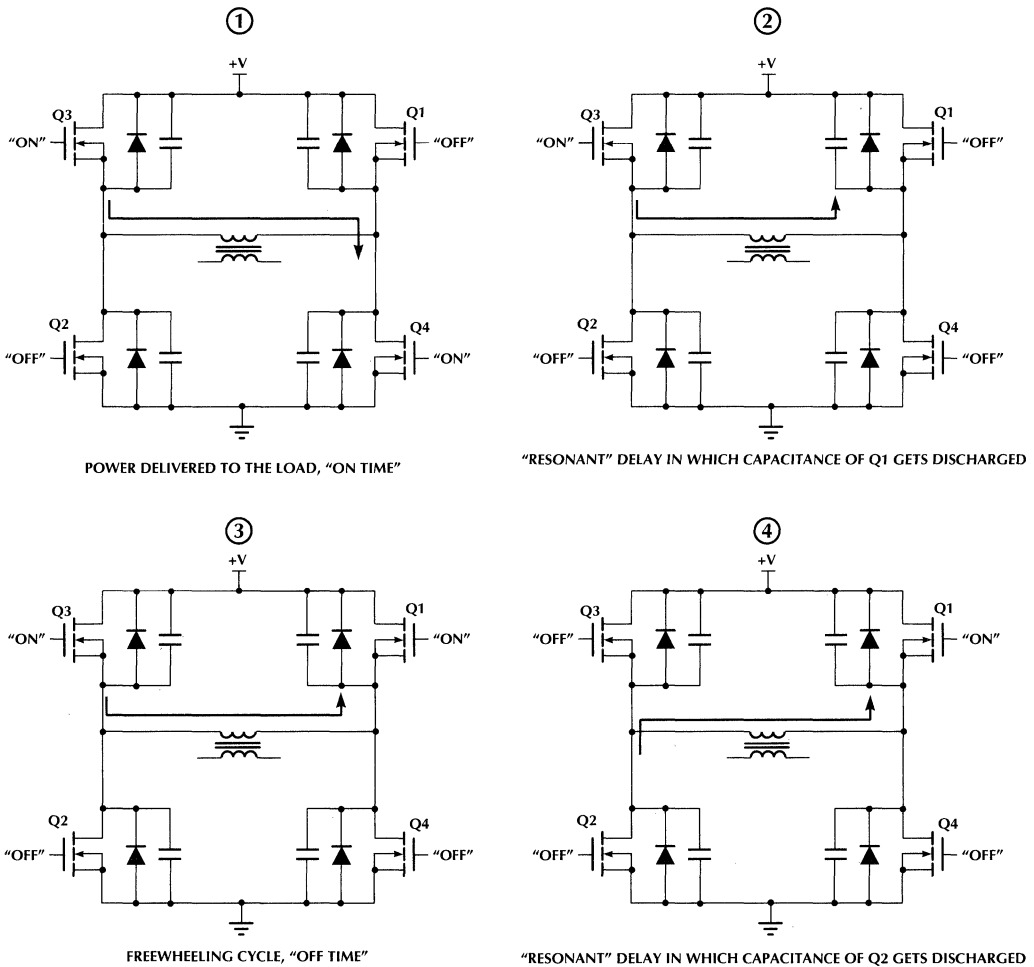


Figure 5a. PMPT power circuit cycles.

In the PMPT in order to accomplish ZVS, the leakage, and magnetizing inductances of the power transformer are utilized along with the drain-source capacitances of the power MOSFETs. The body diode of the MOSFETs also serves to clamp positive and negative going voltages. Thus the parasitic components of the MOSFETs are put to good use with this topology.

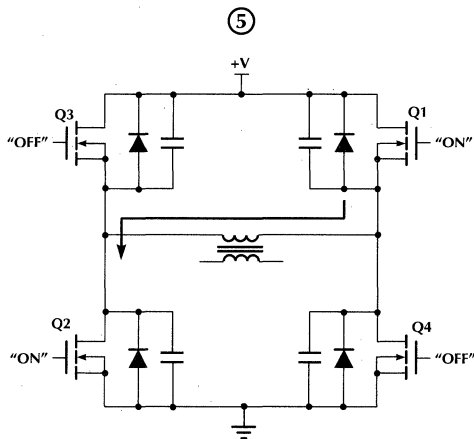
Sometimes in order to further reduce the turn-off losses additional capacitance may be necessary across the drain-source of each MOSFET.

The operation of the PMPT is best understood by examining one full cycle of events in the power circuit. For the time being one can assume that the power transformer magnetizing and leakage inductances will behave as current sources. Figures 5a and 5b show the power stage of the PMPT through one complete cycle.

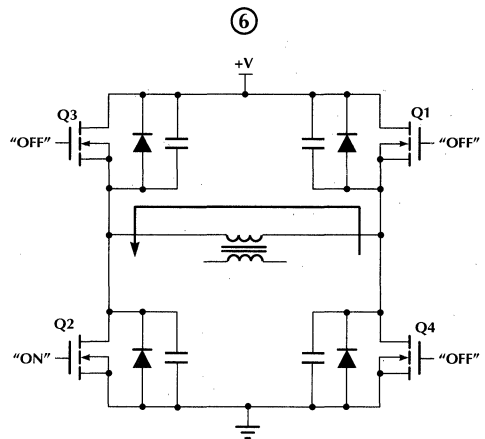
1. The two diagonal MOSFETs are conducting, power is delivered through the transformer to the load. The primary load current is flowing through the leakage inductance of the transformer. The total primary current is equal to the load current plus the increasing magnetizing current of the transformer. The magnetizing current is of importance here since when the output load is very light there is very little reflected load current to complete the ZVS action.

2. Q4 turns off, the capacitance across Q1 was charged to +V when Q4 was ON with the turning OFF of Q4 the current through the transformer inductances starts to charge the drain-source capacitance of Q4 while at the same time discharges the capacitance of Q1. This action continues until the body diode of Q1 turns ON to clamp the voltage across Q1 to approximately -0.7V. The current through the transformer is sustained in the upper half of the power circuit as shown in the figure of phase (2).

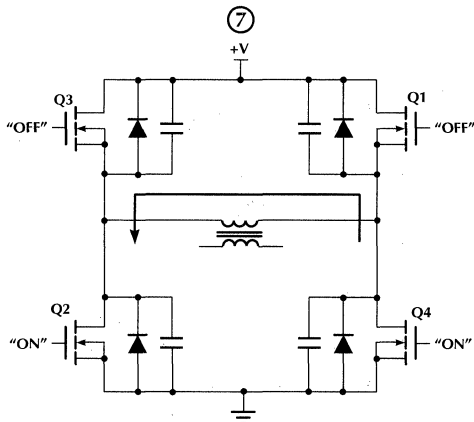
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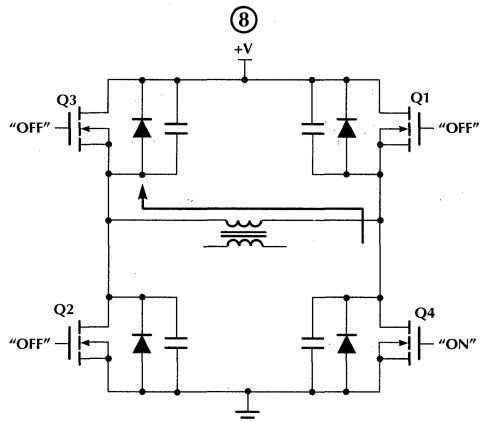
ZERO VOLTAGE SWITCHED "ON TIME,"
POWER DELIVERED TO THE LOAD.



"RESONANT" DELAY IN WHICH CAPACITANCE OF Q4 GETS DISCHARGED



FREEWHEELING CYCLE "OFF TIME"



"RESONANT" DELAY IN WHICH CAPACITANCE OF Q3 GETS DISCHARGED

Figure 5b. PMPT power circuit cycles.

3. When the voltage across Q1 reaches approximately "0 volts" Q1 turns ON. The time that is required for the capacitance of Q4 and Q1 to reach the desired voltage is programmed as delays in the gate drive waveforms of the controller. This delay is programmable with an external resistor for complete flexibility. The current in this phase is circulating through the conduction channels of Q3 and Q1.

4. Q3 turns off, the transformer current now starts to charge and discharge the capacitances of Q3 and Q2 respectively. It requires again a finite amount of time for the drain voltage of Q2 to reach "0 volts" this time is consistent with the programmed delay at the outputs of the controller. It is the presence of this delay that makes ZVS possible. When the voltage across Q2 reaches "0 volts" then Q2 will be turned ON with no voltage across it. Thus accomplishing our goal of non-dissipative turn ON switching.

5. With the complete discharge of its drain source capacitance Q2 now is ready to turn ON. Power is delivered to the load through the conducting path of Q1 and Q2 for an amount of time that is determined by the control circuit. The product of this time, times two, times the operating frequency of the oscillator gives the duty cycle of the converter as in regular PWM converters.

$$\text{Duty Cycle} = 2t_{\text{ON}}f$$

Thus the calculation of output voltages or of the required transformer turns ratio becomes a task that is quite similar to that of the regular PWM converters. The difference is that in regular PWM converters the magnetizing inductance is maximized in order to get the minimum amount of magnetizing current. In the PMPT magnetizing current has to be at certain level to facilitate ZVS when the reflected load

current to the primary is insufficient to do so. This magnetizing current adds to the reflected load current thus requiring the use of a lower R_{DS} MOSFET. This is the penalty paid by using the PMPT. In any case the requirements are far less if conventional resonant technique were to be used.

6. Following the power transfer of the above diagonal pair, Q1 turns OFF. The voltage across Q4 starts to decrease, and when this voltage reaches "0 volts" the next phase starts.

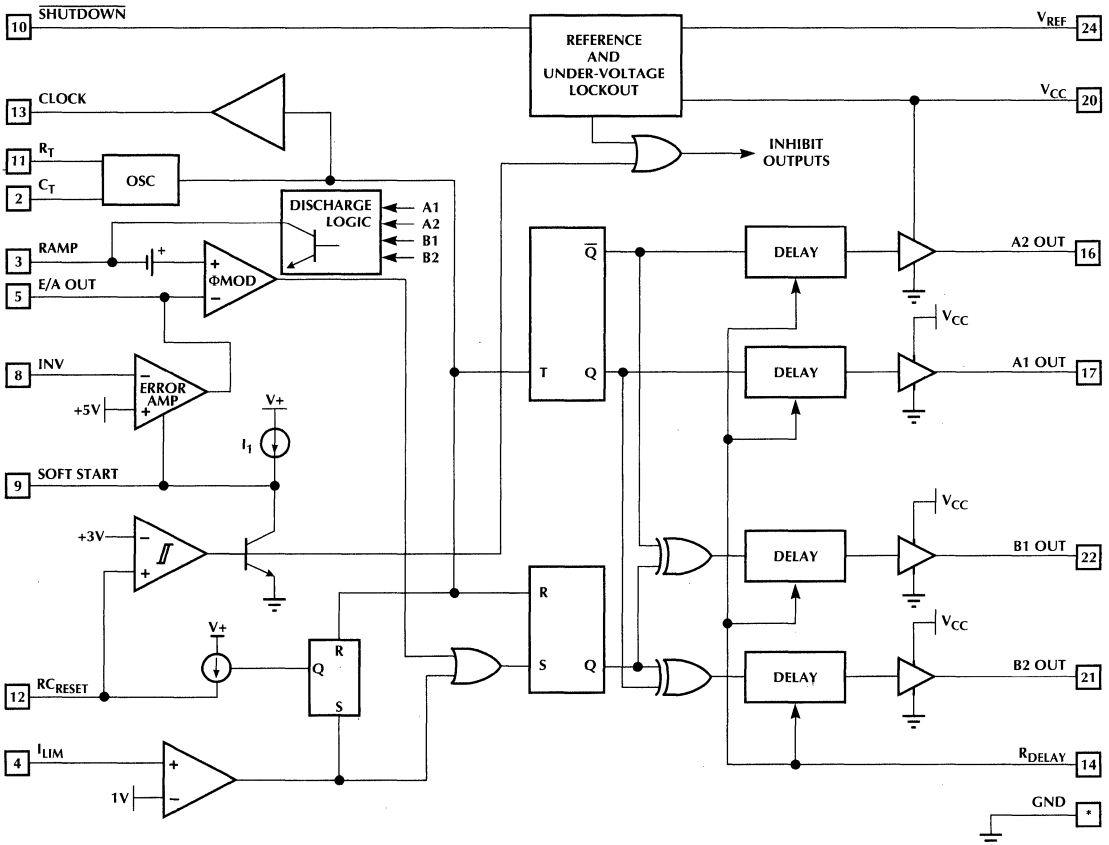
7. In this phase Q4 turns on and the primary current circulates in the conduction channels of the lower pair.

8. Q2 turns OFF and the current starts to charge and discharge the capacitances of Q2 and Q3 respectively. When the voltage across Q3 has reached "0 volts" then Q3 turns ON non-dissipatively and the complete cycle repeats itself from phase (1).

CONTROL CIRCUIT CONSIDERATIONS

The ML4818 PMPT controller has been designed to generate all the necessary timing and gate waveforms, and it contains logic circuitry for effective fault management that is of paramount importance in high power switching power supplies.

The control method involved instead of attempting to change the pulse width of the switches, changes the pulse width of the power pulse. Each of the switches operates under constant duty cycle that approaches 50%. In actuality the duty ratio is in the range of 40% to 45%. The remaining 5% to 10% of the time is being used for the ZVS action to take place. The above percentages may change with various operating frequencies. The effective maximum power pulse width can be much closer to 50% for optimum performance



*PINS 1, 6, 7, 15, 18, 19 AND 23 ARE GND

Figure 6. Functional Block diagram of the ML4818.

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at lower frequencies (for example at 100KHz). A demonstration board is available from Micro Linear that operates at about 500KHz. Probing this board is a very good way of learning about the various operating modes of this very important class of PWM topologies.

Lets now discuss how phase modulation is accomplished. Earlier we mentioned that each of the four switches in the power bridge circuit has its own gate drive waveform. That requires four individual gate drive signals to be generated by the control circuit. This is exactly what the ML4818 does. The controller has four outputs that can directly drive the four MOSFETs. Figure 6 shows the internal block diagram of this controller.

In order to fully understand the control mechanism, we will look into the heart of the controller. Figure 7 shows the phase modulator core of the controller. All fault and supervisory circuitry has been left out. Also both of the complementary outputs and all driver and delay blocks have been left out for further simplicity. By examining how the

other two outputs behave, one is able to grasp the basic operating principles of the phase modulator.

As can be seen from Figure 7, the core of the circuit is quite simple. Assuming that the two comparators are inactive for the time being the only stimulus that the circuit receives is from the clock pulse. Under this assumption the circuit reduces to the one shown in Figure 8.

From Figure 8, clearly if the set input of FFB is always logic "0" then the "Q" output of FFB will be reset or logic "0". From the operation of the exclusive OR gate then the "B" output will be equal to the inverted output of FFA i.e., outputs A and B will be 180 degrees out of phase from each other. Figure 9, shows the relevant waveforms.

If now we assume that a periodic stimulus is present at the set Input "S" of FFB occurring at some instance other than the clock pulse instance then the resulting waveforms will be different as is evident by examining Figures 8 and 9. The resulting timing diagram is shown in Figure 10.

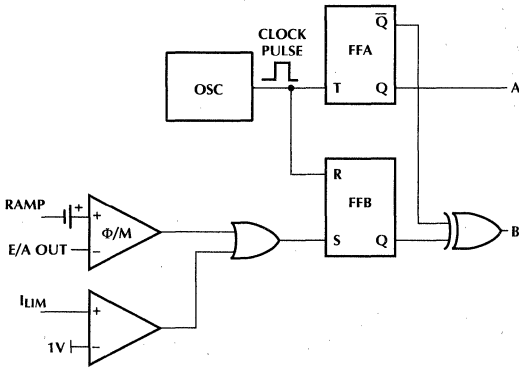


Figure 7. Phase modulator of the ML4818.

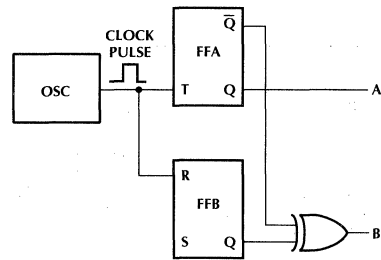


Figure 8. Phase modulator control logic.

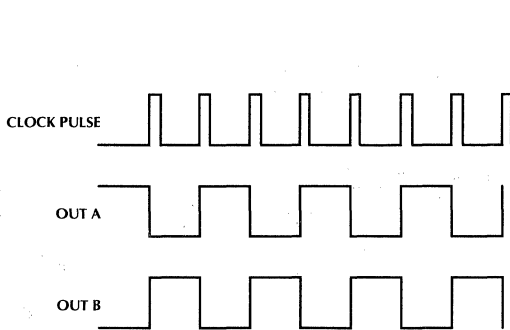


Figure 9. Timing waveforms of the basic phase modulator at the absence of stimulus other than the clock pulse.

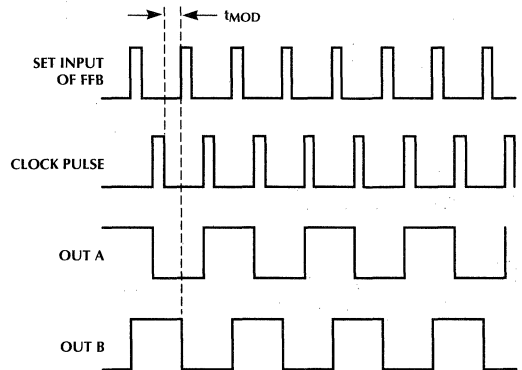


Figure 10. Resulting timing waveforms when there is a periodic stimulus at input "S" of FFB with period T equal to the period of the clock pulse.

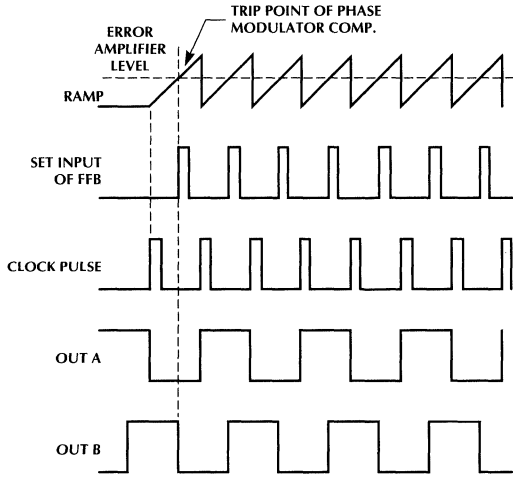


Figure 11. Generation of the set pulse for FFB. Ramp shape may be different in actual application.

Output "A" is free toggling whenever the clock pulse is present, whereas output "B" is the exclusive OR'ed output that is the result of outputs FFA, Q and FFB. The exclusive OR gate here functions as a controlled inverter. The other two outputs of Figure 6, behave in similar way with the only difference that they are 180 degrees out of phase with the ones described above.

By controlling time t_{MOD} then we are able to control the phase shift between outputs "A" and "B". The set input pulse for FFB is normally generated by either the phase modulator comparator of Figure 12, or by the current limit comparator. As in normal PWM regulators one input of the phase modulator comparator is the output of the error amplifier which sets the trip level and the other input is either a voltage ramp or the sensed primary (or secondary) current waveform of the power circuit. Figure 11, shows the generation of the set pulse and the resulting timing waveforms. When the output of the error amplifier changes then the trip level changes thus it is possible to continuously control time t_{MOD} by changing the trip level. In switching power supplies it is also necessary to limit the power pulse width whenever the primary load current exceeds certain predetermined value. The second comparator in Figure 7, serves that purpose. Thus the output of the phase comparator and of the current limit comparator are logic OR'ed to produce the set pulse for FFB. Figure 12, shows the complete logic diagram of the phase modulator with four outputs labeled A1, A2, B1, and B2 respectively.

The timing waveforms for all outputs can be derived from the above diagram. Figure 13, shows the relationship of the outputs with respect to each other.

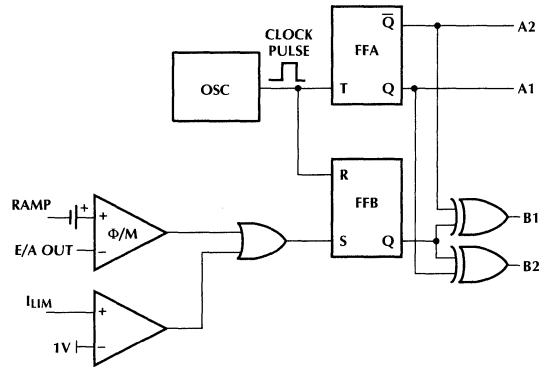


Figure 12. Complete logic diagram of the phase modulator with all four outputs shown.

DELAY OF THE GATE DRIVE WAVEFORM

So far we saw how controllable phase shifted outputs could be generated. We also saw that in order to have ZVS in the bridge circuit in the transition phase between conduction of the opposing legs, one of the MOSFETs is ON and the remaining are OFF. It is during this time that the drain source capacitance of the device next to turn ON is discharged to "zero volts". In the traditional H-bridge either two of the devices are ON or all of them are OFF.

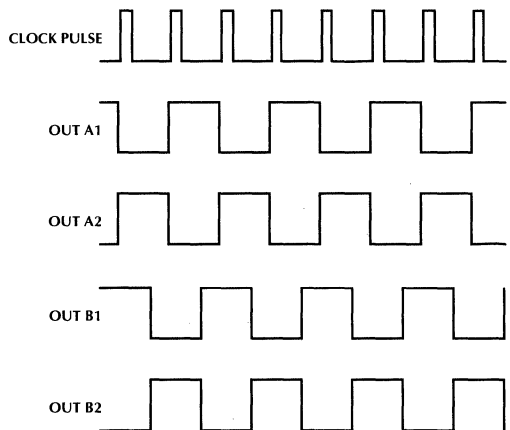


Figure 13. Timing diagram showing the waveform present on all four outputs of the phase controller.

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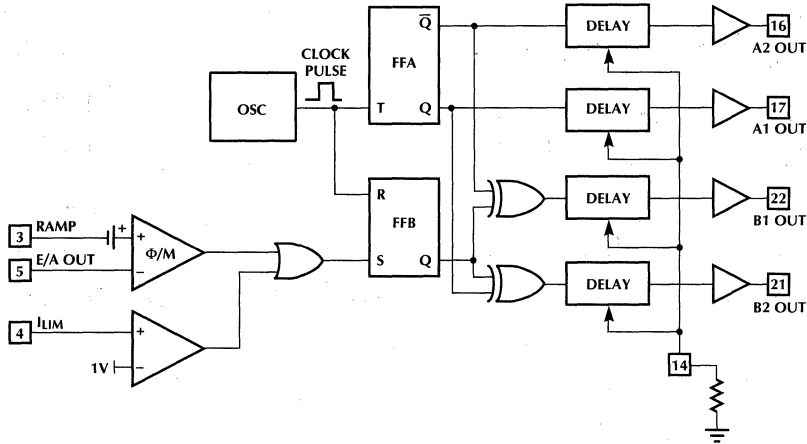


Figure 14. Phase modulator with delay and output driver blocks shown.

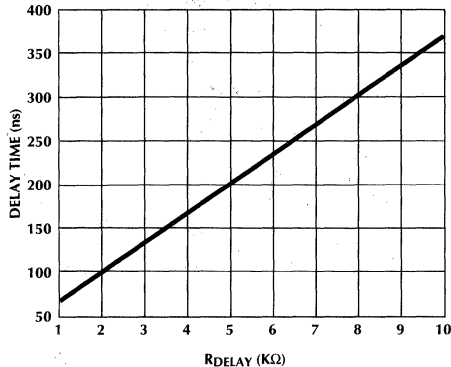


Figure 15. Chart for the determination of the external delay resistor.

In order to get this transition period it is necessary to modify the gate drive waveforms slightly. The modification consists of adding a predetermined amount of delay to the leading edge of the gate drive waveforms. Figure 14, shows the phase modulator along with the delay and output driver blocks. The delay blocks contain all the necessary electronics for the generation of the delay with the use of a single external resistor. The timing capacitor is integrated into the part. The value of the delay resistor R_{DELAY} can be found by using the chart of Figure 15, or it can be calculated by using the formula below

$$R_{DELAY} = \frac{DELAY(ns) - 33.34}{33.33} K\Omega$$

The individual gate drive waveforms are restricted to less than 50%. The resulting waveforms are shown in Figure 16. The shaded area in the leading edges shows the reduction in the pulse width. Note that the delays are only present at the leading edges of the waveforms. Hence the less than 50%

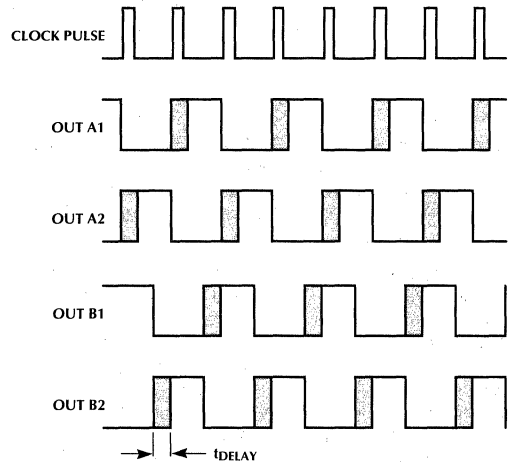


Figure 16. Leading edge delay of the drive waveforms necessary for ZVS operation.

duty cycle. Because all the drive signals have this delay complementary waveform symmetry is preserved. Although the delay time can be adjusted using an external resistor, it should be kept in mind that resulting actual delays in the power circuit may differ, this will be due to slew of the drive waveforms, also it requires a finite time to charge and discharge the gate capacitances of the MOSFETs.

The amount of time that is required to complete ZVS will vary as the load current reflected to the primary varies. The power transformer magnetizing inductance has to be able to develop enough current during "ON time" for ZVS to complete its cycle. That will require careful design of the power transformer. In most cases the transformer will have to be gapped. A side effect of the gapping will be the stabilization of the magnetizing inductance.

The idealized power transfer cycles of the PMPT are shown in Figure 17. Power transfer takes place during the hatched periods. The width of these periods depends on the phase relationship between the "A" and "B" outputs. A good way to observe the phase modulation action is to watch the oscilloscope traces of "A1" and "B1" outputs.

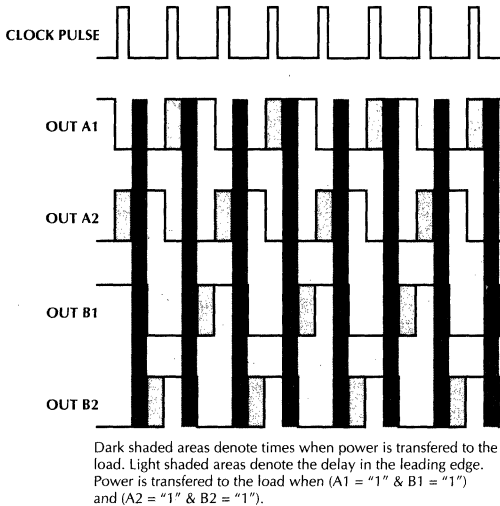


Figure 17. Power transfer cycles of the PMPT.

DESIGNING FOR ZERO VOLTAGE SWITCHING

So far we saw that it is possible to do non-dissipative switching. The ML4818 power supply controller with its operational flexibility is able to provide all the necessary waveforms needed for such power supply. The most important thing left to be done is the design of the power transformer. Zero Voltage Switching properties greatly depend on this component.

Figure 18, shows the simplified equivalent model of a typical power transformer. The model is indeed very simple, actual transformers are very difficult to model due to the presence of primary and secondary effects such as saturation and inter-winding capacitances.

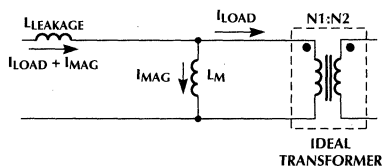


Figure 18. Simplified model of a real transformer.

It is obvious from Figure 18, that the current that will flow through the MOSFETs will be the sum of both the reflected secondary current which we call the load current and of the current that builds up in the magnetizing inductance of the transformer.

The load current is a function of the output load and can change anywhere from zero to its full rating as it reflects to the primary. The magnetizing current on the other hand is a function of the ON time and of the primary applied voltage. The output filter inductor where the load current flows through is normally very large. For all practical purposes the reflected current of the primary can be assumed constant during the intervals of interest.

$$I_{MAG} = \frac{V_{IN} t_{ON}}{L_M}$$

It is also important to remember that inductors can be approximated as current sources. With all this in mind lets now look into what happens when a power MOSFET switches OFF. Figure 19, shows one leg of the bridge circuit with the parasitic drain source capacitances. Assuming that the lower device was conducting current just prior to turning off the following events may happen.

If the gate drive waveform is fast enough and drops to zero volts before the drain source capacitance can charge to any significant voltage then the turn off event will be non-dissipative. This is the principle also with snubbers, where a large amount of external capacitance diverts the current from the channel for non-dissipative switching. Where the PMPT excels is that just before turn ON the energy stored in the snubber or drain source capacitances of the MOSFETs is returned back to the source as we saw earlier. This way there is not a penalty for using additional snubber capacitance. And since the turn ON is at zero voltage the switching event is lossless even at very high frequencies i.e., 500KHz and above.

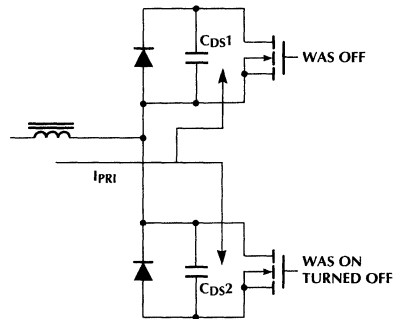


Figure 19. MOSFET switching in PMPT, and charge and discharge of the drain source capacitances.

In Figure 19, in order to discharge the drain source capacitances there needs to be a certain amount of current stored in the inductances of the transformer. Since the total charging current is the sum of the load currents and the magnetizing current, whenever the load current is low the charging has to be done by the current that the magnetizing inductance was charged. Hence the importance of the magnetizing inductance.

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One can get quite complicated in trying to calculate the required amount magnetizing inductance necessary for ZVS. But the following simple procedure could be followed with some experimentation to get familiar with the technique.

The required maximum duty cycle is the first parameter to consider. Along with the delay that will be introduced by the delay circuit of the ML4818, in high frequency conversion the effective duty cycle will be reduced by the charging effect of the leakage inductance. Here we will assume that the reduction of the duty cycle by the charging of the leakage inductance is negligible (not true in most of the cases).

Lets assume that the required duty cycle results in a delay time t_D . In other words the time available to complete the charging of the drain source capacitance is t_D . The charging configuration of the bridge circuit changes depending on its state just prior to ZVS. In Figure 19, the bridge circuit was delivering power to the secondary, so it was in a power transfer cycle just prior to the next ZVS. Under this condition the drain source capacitances will be charged linearly by a current equal to

$$I_{PRI} = I_{LOAD} + I_{MAG}$$

This is not the worst case. The worst case happens after the above there is a period of freewheeling of the current before the next device turns ON with ZVS. This corresponds to phases (3) and (4) in Figure 5a. During this period the output diodes effectively short the transformer and the only inductance in the circuit is the leakage inductance and carries a current slightly less than at the end of the power transfer cycle. Therefore all the calculations are based on this case.

During phases (2) and (3) the capacitances charge linearly with the reflected load current. During phases (3) and (4) the capacitances resonate with the leakage inductance and charge in a resonant fashion. The worst case is when the load current is very close to zero. Under that condition the freewheeling current in the leakage inductance is equal to the magnetizing current at the end of the power transfer cycle.

The assumption we are going to make is that the delay time represents one quarter of the resonant cycle determined by the leakage inductance and twice the drain source capacitance of the MOSFETs, since there are two MOSFETs per side.

The procedure for calculating the necessary amount of the leakage and magnetizing inductances is shown below. An important note here is that in most of the cases the leakage inductance will be determined by the actual transformer construction and it will be a given. In that case the equation below could be used first to calculate the required amount of the delay time.

1. Calculate the required amount of either delay or leakage inductance value when one of them is the given value

$$\frac{1}{4t_D} = \frac{1}{T} = \frac{1}{2\pi\sqrt{2C_{DS}L_{LEAKAGE}}} \Rightarrow$$

$$L_{LEAKAGE} = \frac{2}{C_{DS}} \left(\frac{t_D}{\pi} \right)^2$$

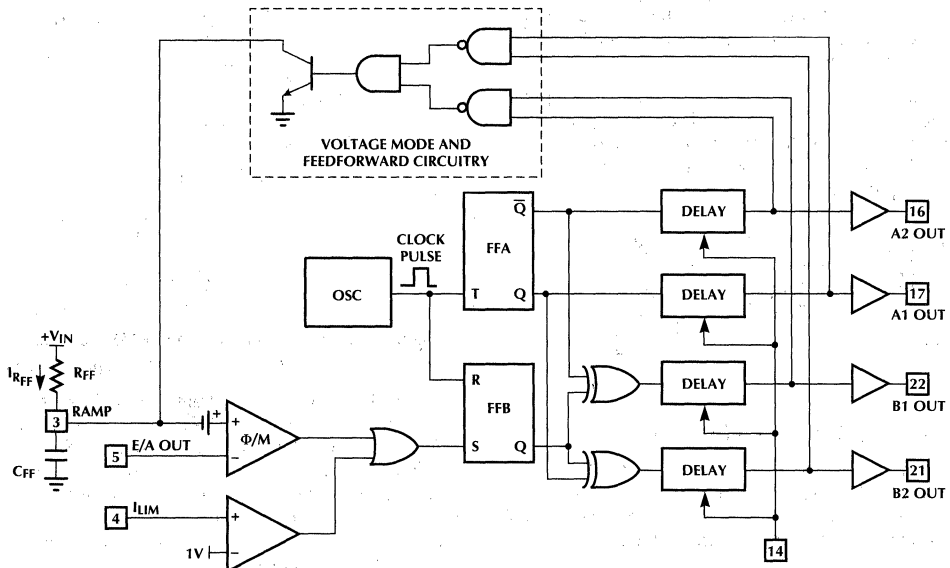


Figure 20. Voltage mode and feed-forward circuitry of the ML4818.

or

$$t_D = \pi \sqrt{\frac{C_{DS} L_{LEAKAGE}}{2}}$$

2. Calculate the minimum current for ZVS from the energy required to reach $V_{IN(MAX)}$. Note that twice the energy necessary to swing one leg of the bridge is required to be stored by the leakage inductance. This is because we have two ZVS actions during one complete power transfer cycle.

$$\frac{1}{2} L_{LEAKAGE} I_{MAG(min)}^2 = 4 \left(\frac{1}{2} C_{DS} V_{IN(max)}^2 \right) \Rightarrow$$

$$L_{LEAKAGE} I_{MAG(min)}^2 = 4 C_{DS} V_{IN(max)}^2 \Rightarrow$$

$$I_{MAG(min)} = \sqrt{\frac{4 C_{DS} V_{IN(max)}^2}{L_{LEAKAGE}}}$$

3. Calculate the value of L_{MAG} .

$$L_{MAG} = \frac{V_{IN} t_{ON}}{I_{MAG(min)}}$$

Example: Lets suppose that the following are given

$$L_{LEAKAGE} = 15 \mu H$$

$$C_{DS} = 225 pF$$

$$V_{IN(max)} = 370 \text{ Volts}$$

$$t_{ON} (@ V_{IN(max)}) = 1 \mu sec$$

The calculations for the above example are as follows:

$$T_D = 3.14 \sqrt{\frac{(225 pF)(15 \mu H)}{2}} = 129 nsec$$

$$I_{MAG(min)} = \sqrt{\frac{4(225 pF)(370 V)^2}{15 \mu H}} = 2.86 A$$

$$L_{MAG} = \frac{(370 V)(1 \mu sec)}{2.86 A} = 95.85 \mu H$$

The value of the magnetizing inductance found above is somewhat low. This is because we made the assumption that there will be ZVS down to very light loads. If we do not allow the output load to go to very light loads or if we are willing to live with NZVS (Near Zero Voltage Switching) then the resulting magnetizing inductor values will be higher resulting in lower conduction losses. The value of magnetizing inductance that was chosen for the typical PMPT power supply of Figure 23, was 400 μH , the value of the leakage inductance was approximately 15 μH .

MORE ABOUT THE ML4818 CONTROLLER, VOLTAGE OR CURRENT MODE OPERATION

The ML4818 controller is able to control a PMPT converter operating either voltage mode or current mode. It contains special logic circuitry for that purpose. That same circuitry allows for voltage feed-forward in voltage mode operation. Figure 20, shows the logic circuit internal to the IC that enables the above.

Pin #3 is pulled to ground at the end of the power cycle and is kept at ground until the start of the next power transfer cycle. Thus the discharge of the feed-forward capacitor is enabled. An important note here is that the PMPT operating in voltage mode required the internal logic of Figure 20. It is not possible to operate voltage mode by only connecting pin #3 to the oscillator ramp. This is unlike conventional PWM regulators.

Feed-forward voltage mode operation provides automatic line correction without the need for the voltage control loop to change the duty cycle. The correction takes place within a single cycle. The current through resistor R_{FF} is proportional to input voltage therefore the charging time of the C_{FF} capacitor is proportional to input line voltage and consequently the time that it takes to reach the threshold set by the error amplifier.

$$I_{R_{FF}} = \frac{V_{IN}}{R_{FF}}$$

$$t_{ON} = \frac{V_{E/A} C_{FF}}{I_{R_{FF}}} = \frac{V_{E/A(max)} C_{FF} R_{FF}}{V_{IN(min)}}$$

The necessary values for any given application can be calculated using the above equations which can be solved for any of the unknown values. The resulting ramp will affect the open loop gain of the voltage control loop. The open loop gain will be independent of the variations in input voltage. The open loop gain for the voltage mode controlled case can be calculated by using the following.

$$G_{o.l.} = \frac{C_{FF} R_{FF} f_{OSC}}{N}$$

where N = Primary to secondary turns ratio

f_{OSC} = oscillator frequency

$G_{o.l.}$ = open loop voltage gain

For the circuit in Figure 23, the open loop voltage gain was calculated to be 7 or 16.9db, below the corner frequency of the output filter. In this case the corner frequency is 4.1KHz.

A TYPICAL PMPT POWER SUPPLY

Figure 23, shows a typical off-line PMPT supply, as it can be seen from the schematic diagram only a handful of components are required to build a fully functional power supply. Specifications for this supply are shown in Table 1.

The power supply was not optimized for any particular application. It is important to note that higher efficiencies

Application Note 19

can be obtained by using lower loss magnetic materials and lower ON resistance MOSFETs.

The controller in this design is located at the primary side. The voltage feedback is accomplished with the use of an optocoupler. The current transfer ratio for this optocoupler is almost linear for a limited operating range, this enabled the use as it is shown in Figure 23. For more realistic applications an error amplifier at the secondary would probably be required. If an auxiliary supply is available then the controller itself can be situated in the secondary. Demonstration board that implements this power supply is available from Micro Linear Corp., it is a useful tool in gaining familiarity with this topology.

The control loop design for the voltage mode version of a PMPT supply is identical to that of a regular full bridge PWM converter. For simplicity in this case the loop is stabilized relying on the ESR of the output capacitor. The feedback components are calculated to give enough margin for loop stability.

The gate drive for the four MOSFETs is accomplished by using two drive transformers (T2 and T3). There are two secondary windings on each transformer to be able to drive the two MOSFETs in each leg of the bridge Figure 21, shows this configuration. The primary of the transformers connect to outputs A1, A2 and B1, B2 through a DC coupling capacitor to prevent drive core saturation under abnormal conditions. The windings of the transformers are trifilarly wound to minimize leakage inductances. A toroid of 0.5" outside diameter with 10 turns for each windings seem to

function fairly well. The wires should be insulated to provide isolation.

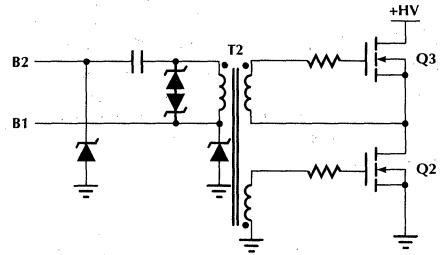


Figure 21. Gate drive scheme for fail-safe operation.

The power circuit consists of a DC blocking capacitor in series with the power transformer to prevent core saturation under unbalanced and abnormal conditions. The current transformer in series senses the current for cycle by cycle current limit. If the current limit persists then capacitor C8 charges to 3V triggering an internal comparator and shutting down the power supply. The resistor connected across C8 helps to discharge this capacitor and the power supply tries to soft start.

For high power applications where large size MOSFETs are used it may be necessary to use external gate drivers to

Table 1.

Input Voltage Range	90VAC to 140VAC or 180VAC to 260VAC
Output Voltage	15V adjustable
Output Voltage Adjustment Range	12.6V to 20V (@ 120VAC and I _{out} =1.3A)
Output Current	13.3A
Line Regulation (90VAC to 140VAC)	<30mV
Load Regulation (10% to 100% @ 20VAC)	<300mV
Current limit set to approximately	15A
Output Power	200W
Efficiency at full load and 120VAC	82%
Output Voltage Ripple	250mV (without additional filtering)
Voltage ramp time at turn on (0V to 15V)	<8msec
Oscillator Frequency	500KHz
Average power under short circuit	<10W
Short Circuit Capability	Indefinite
Short Circuit Protection Method	Hiccup

Typical PMPT Power Supply Parts List

PART#	VALUE
Resistors	
R1, R2	240K, 1/4W
R3	82K, 2W
R4	39, 1/4W
R5, R20	1K, 1/4W
R9, R10, R11, R12	5.1, 1/4W
R6	4.3K, 1/4W
R7	240K, 1/4W
R8	7.5K, 1/4W
R13	510, 1/4W
R16	1K, 1/4W
R14	1K, 1/4W, POT
R15	100K, 1/4W
R17	330K, 1/4W
R19	100K, 1/4W
R21	5.1K, 1/4W

PART#	VALUE
Capacitors	
C1, C2	680 μ F, 200V ELECTROLYTIC
C3	200 μ F, 25V, ELECTROLYTIC
C4	0.1 μ F, CER.
C5	680pF, PRECISION
C6	470pF, CER.
C7, C10, C15, C12, C16	1 μ F, CER
C8	56nF, CER
C9	0.33 μ F, 630V, POLYPROPYLENE
C11	100 μ F, 25V, ELECTROLYTIC
C14	0.01 μ F, 1KV, CER.
C18	1000pF, CER.
C20	220pF, CER.
C22	120pF, CER
C21	470pF, CER.
C23, C24	10nF, 1KV

divert the power dissipation from the ML4818 controller to the external drivers. The drivers can be simple NPN-PNP pairs Figure 22, shows such a configuration.

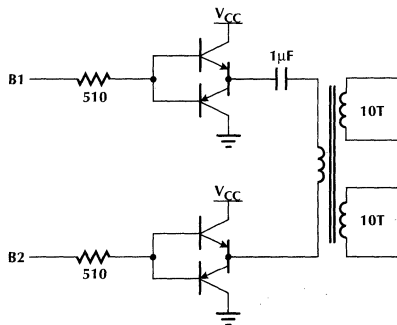


Figure 22. External gate drivers using NPN-PNP pairs.

PART#	VALUE
Diodes	
D1, D2	MUR150
D3, D4, D5, D6	1N5406, 3A, 600V
D9, D10, D11, D12	1N4148
D13	MBR3045PT, 30A, 45V, SCHOTTKY
D14, D15, D7, D8	1N5818, SCHOTTKY
D16, D17, D18, D19	1N5248, 18V ZENER

PART#	VALUE
MOSFETS	
Q1, Q2, Q3, Q4	IRF840

PART#	VALUE
IC's	
IC1	ML4818, PHASE-MOD. IC
OP1	MOC8102, OPTOCOUPLER

PART#	VALUE
Inductors	
L1	200 μ H, 0.3A FILTER CHOKE
L2	15 μ H, LITZ WIRE, 15A FILTER CHOKE

PART#	VALUE
Transformers	
T1	45T/2X4T/2X4T, Lmag = 400 μ H OBTAINED BY GAPPING, PRIMARY AND SECONDARIES ARE LITZ WIRE, Lleakage = 15 μ H, POT CORE, HIGH FREQUENCY MATERIAL
T2, T3	10T/10T/10T, GATE DRIVE TRANSFORMER, WOUND TRIFILAR ON 0.5" O.D. TOROID WITH INSULATED WIRE
T4	1T/80T, SAME CORE AS ABOVE, 80T IS AWG #28 MAGNET WIRE

PART#	VALUE
Fuses	
F1	5A, 250V FUSE

Application Brief 1

How to Set the Sensitivity of the ML4621, ML4622, ML4624

SECTION 1

HOW TO ADJUST THE SENSITIVITY OF THE ML4622/ML4624:

The sensitivity of ML462X is adjustable by changing the voltage level at V_{THADJ} pin. The sensitivity should be set at a point which guarantees error free operation with minimum signal level and the maximum noise level on the quantizer inputs. The first step is to determine the input threshold at which errors begin to occur. To determine this the following steps are recommended.

- 1) Tie V_{THADJ} to ground on the receiving station to find the maximum sensitivity.
- 2) Transmit $*10^9$ bits of data from transmitting station and verify that receiving station receives all the bits without any error.
- 3) If receiving station does not receive the data correctly, go to step 4). Otherwise measure the received power and attenuate the receive signal more. Then go to step 2.
- 4) At this point you know the maximum sensitivity of the receiving station before it receives any error. The minimum level must be at least -32.5dBm average to meet 10BASE-FL standard or -27dBm peak to meet FOIRL standard. Now you can set the sensitivity of the receiving station to any level you want as long as it is greater than the maximum sensitivity and less than minimum sensitivity (when $V_{THADJ} = V_{REF}$) of the ML4622/ML4624. The Link Mon signal will then turn off (high) before receiving any errors.

5) Now you should determined the proper voltage at V_{THADJ} which will meet 10BASE-FL standard with minimum signal level and the maximum noise level.

A) Signal Level: The responsivity of the HFBR2416 can be as low as $4.5\text{mV}/\mu\text{W}$ and as high as $11.5\text{mV}/\mu\text{W}$. So we calculate minimum $V_{IN(P-P)}$ at the input of the quantizer when receive power is -29.5dBm peak.

$$R_p \text{ (MV/UW)} = \text{Responsivity of the HFBR2416}$$

$$P_R \text{ (dBm)} = \text{Average receive power}$$

$$V_{IN(P-P)} = \text{Input peak to peak voltage at the input of ML4622, ML4624}$$

$$-29.5\text{dB} = 1.122\mu\text{W}$$

$$V_{IN(P-P)} = 1.122\mu\text{W} \times 4.5\text{mV}/\mu\text{W}$$

$$V_{IN(P-P)} = 5.049\text{mV} \quad (0)$$

Thus V_{THADJ} for the minimum signal level can be calculated as follow:

$$V_{THADJ} = 500 V_{IN(P-P)} \text{ (IN mV)}$$

$$V_{THADJ} = 500 (5.049) = 2.52\text{V} \quad (1)$$

B) Noise Level: The maximum random noise ($V_{N(MAX)}$) of the HFBR2416 is 0.7mV which it occurs at a responsivity of $R_p = 8.2\text{mV}/\mu\text{W}$. This input signal will be attenuated by the internal low pass filter of the quantizer. If capacitor across CF1 and CF2 is 5pF , the attenuated noise voltage will be:

$$f = 1/2\pi \times 800 \text{ (C+4)} = 22.1\text{MHZ}$$

$$\text{(ML4622/ML4624)}$$

$$N = \left(\sqrt{\frac{22.1\text{MHZ}}{125\text{MHZ}}} \right) (.7\text{mV}) = .294\text{mV}_{P-P}$$

As shown in figure 1, the signal to noise ratio required at the fiber optic receivers comparator is:

$$S/N @ \text{BER of } 1 \times 10^{-10} = 12.8 \quad (\text{FOIRL})$$

$$S/N @ \text{BER of } 1 \times 10^{-9} = 12 \quad (\text{10BASE-FL})$$

We can calculate the signal level at the input of the quantizer.

$$S = V_{IN(P-P)}$$

$$V_{IN(P-P)} = (12.8)(.294) = 3.76\text{mV}_{(P-P)} \quad (\text{FOIRL})$$

$$V_{IN(P-P)} = (12)(.294) = 3.52\text{mV}_{(P-P)} \quad (\text{10BASE-FL})$$

The link monitor threshold of the ML4622/24 should be set to reject the output voltage of the HFBR2416 when it is 3.52mV_{P-P} for the 10BASE-FL or 3.76mV_{P-P} for the FOIRL. The V_{THADJ} for this signal level can be calculated as follows:

$$V_{THADJ} = 500 (4.74) = 2.37\text{V} \quad (\text{ML4622})$$

$$V_{THADJ} = 417 (4.74) = 1.97\text{V} \quad (\text{ML4624}) \quad (2)$$

Setting the V_{THADJ} at 2.5V (tie to V_{REF}) will set the input threshold greater than the maximum noise level of both specifications. This will allow the quantizer to reject the worst case noise levels and meet both specifications for minimum signal level of 5.049mV_{P-P} .

* 10^9 bits to meet 10BASE-FL standard and 10^{10} bits to meet FOIRL standard.

Application Brief 1

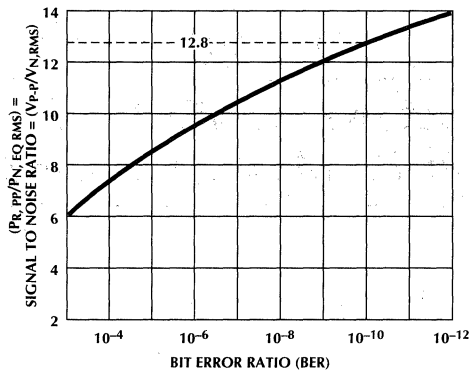


Figure 1.

SECTION 2

HOW TO SET V_{THADJ} FOR GREATER SENSITIVITY:

Lower voltage at V_{THADJ} gives you more sensitivity. By adding a resistor divider as shown in Figure 2 you can lower the voltage at V_{THADJ} to have sensitivity more than -29.5dBm peak. You can calculate the resistors' value as follow:

- Find the V_{THADJ} for the sensitivity you want at the input of the ML4622/ML4624.

$$V_{THADJ} = 500 V_{IN(P-P)}$$

$$V_{THADJ} = V_{REF} (R1/R1 + R2)$$

$$V_{REF} = 2.5V$$

$$V_{THADJ} = 2.5V (R1/R1 + R2) \quad (3)$$

- Set $R1 = 1k$ and solve EQU. 3 for $R2$.

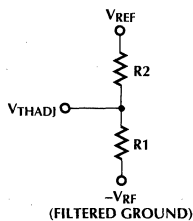


Figure 2.

SECTION 3

HOW TO SET V_{THADJ} IN THE ML4621 TO MEET 10BASE-FL STANDARD:

To determine proper voltage at V_{THADJ} we must calculate V_{THADJ} for the minimum signal level and the maximum noise level.

A) Signal Level: As we calculated in Section 1, the signal at the input of the ML4621 can be as low as 5.049mV peak to peak. Thus, V_{THADJ} for the ML4621 can be calculated as follow:

$$V_{THADJ} = 0.7V + 600 V_{IN(PEAK)} \text{ (IN mV)}$$

$$V_{THADJ} = 0.7V + 600 (5.049/2)$$

$$V_{THADJ} = 2.21V$$

B) Noise Level: As explained in section 1, we can calculate the input signal level at the ML4621 for the maximum random noise level of .7mV as follow:

$$f = 1/2\pi RC = 37.4\text{MHZ}$$

$$(C=10\text{pF across CF1 \& CF2})$$

$$N = \left(\sqrt{\frac{37.4\text{MHZ}}{125\text{MHZ}}} \right) (.7\text{mV}) = .382\text{mV}_{P-P}$$

$$V_{IN(P-P)} = (12.8)(.382) = 4.88\text{mV}_{P-P} \quad (\text{FOIRL})$$

$$V_{IN(P-P)} = (12)(.382) = 4.58\text{mV}_{P-P} \quad (\text{10BASE-FL})$$

Since $V_{IN(P-P)}$ is less than 5.49mV (10BASE-FL requirement), we can set the V_{THADJ} to 2.21V by using a resistor divider from V_{REF} . The resistors can be calculated as follow:

$$V_{THADJ} = V_{REF} (R1/R1+R2)$$

$$2.21V = 2.5V (R1/R1+R2) \quad (4)$$

Choose $R1 = 1K$ and solve equation 4 for $R2$.

$$2.21 = 2.5 (1000/1000+R2)$$

$$R2 = 140\Omega$$

Application Brief 2

ML4632 Verses A Voltage Driven Output

INTRODUCTION

This Application Brief covers one of the issues which must be considered to meet IEEE 802.5, IEEE 802.4 and IEEE 802.3 FOIRL and 10BASE-FL. The Optical Power at the output of the LED transmitter (Launched Power) can violate these standards. One way to improve this parameter is to use a current source to drive a Fiber Optic LED transmitter.

ML4661/ML4662, FOIRL and 10BASE-FL transceivers, are the best solution for 802.3 applications (refer to Application Note 15). However if ML4661/ML4662 is not being used in a 802.3 application, ML4632 is the second option to be used.

On the other hand ML4632 can be used for the IEEE 802.4 and IEEE 802.5 (4Mbps) applications.

ANALYSIS

ML4632 is a Fiber Optic LED driver with a programmable current driven output. This will allow user to program the output current with the accuracy of $\pm 10\%$ through a resistor. The ML4632 can be used to drive a Fiber Optic LED transmitter like HP LED transmitter (HFBR1414). The ML4632 regulates the current through the HFBR1414 regardless of power supply variations or variations in V_f between LED transmitters. This will result in a more precise launch power at the output of the transmitter. However if a voltage driven output is used such as CMOS or Schotky gates, the variation in forward voltage (V_f) with forward current (I_f) and the power supply variation must be taken into account. These variations cause a wider range of Launched Power which will violate the standards.

LAUNCHED POWER

Table 1 shows the HFBR1414 Peak Launched Power measured out of 1m of cable and table 2 indicates Launched Power range for the FOIRL and 10 BASE-FL Standards.

TABLE 1

PARAMETERS	PEAK LAUNCHED POWER (dBm)			CONDITIONS	
	MIN	TYP	MAX	$T_A(^{\circ}\text{C})$	I_f
62.5/125 μm Fiber Cable NA = 0.275	-15	-12	-10	25	60mA
	-16		-9	-40 to +85	55mA
	-15.5		-10.5	25	
	-16.5		-9.5	-40 to +85	

TABLE 2

802.3	PEAK LAUNCHED POWER (dBm)	
	MIN	MAX
FOIRL	-18	-9
10BASE-FL	-17	-9

Note: Peak launch power = Average launch power +3dBm.

Table 3 shows the Launch Power if ML4632 is used to drive HFBR1414. We choose 55mA as forward current to meet the HFBR1414's current condition.

TABLE 3

CONDITION	I_f	$P_{T(55\text{mA})}$ (Note 1)		$P_{T(I_f)}$ (Note 2)		P_R (dBm) (Note3)
		MIN	MAX	MIN	MAX	
High	60mA	-16.5	-9.5	-16.2	-9.2	+0.3
Nominal	55mA	-15.5	-10.5	-15.5	-9.5	0.0
Low	50mA	-16.5	-9.5	-16.9	-9.9	-0.4

Note 1: $P_{T(55\text{mA})}$: Optical Power of the HFBR1414 when I_f is 60mA.

Note 2: $P_{T(I_f)}$: Optical Power of the HFBR1414 for $I_f = 55\text{mA} \pm 10\%$.

Note 3: P_R : is relative power ratio in dBm ($P_{T(I_f)} - P_{T(55)}$).

Application Brief 2

Table 4 shows the Launched Power if a voltage source is used to drive the HFBR1414. To calculate I_F at high and low end, we can calculate the resistor value at nominal condition:

$$R = (V_{CC} - V_F)/I_F$$

$$R = V_R/I_F$$

$$R = (5 - 1.7)/60\text{mA} = 55\Omega$$

So I_F can be calculated for the low and high conditions.

CONCLUSION

Due to the HFBR1414's Optical Power range, Using a voltage source as the LED driver can violate the FOIRL and 10 BASE-FL standards. However using the ML4632 to drive the HFBR1414 meets both standards in the worst condition.

TABLE 4

CONDITION	V_F (V)	V_{CC} (V)	V_R (V)	I_F (mA)	$P_{T(60\text{mA})}$ (Note 1)		$P_{T(I_F)}$ (Note 2)		P_R (dBm) (Note3)
					MIN	MAX	MIN	MAX	
High	1.48	5.5	4.02	73.1	-16	-9	-15.2	-8.2	+0.8
Nominal	1.7	5.0	3.3	60	-15	-10	-15	-10	0.0
Low	2.09	4.5	2.41	43.8	-16	-10	-17.8	-11.8	-1.8

Note 1: $P_{T(60\text{mA})}$: Optical Power of the HFBR1414 when I_F is 60mA.

Note 2: $P_{T(I_F)}$: Optical Power of the HFBR1414 for different I_F .

Note 3: P_R : is relative power ratio in dBm ($P_{T(I_F)} - P_{T(60)}$).

Packaging Information

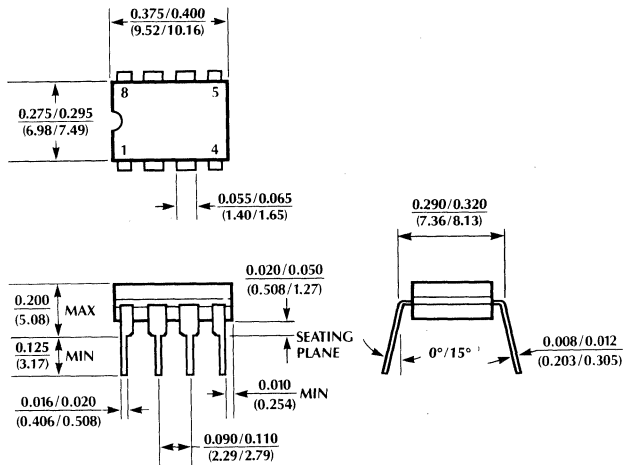
Section 11

Package: J08	8-Pin Hermetic DIP	11-1
Package: P08	8-Pin Molded DIP	11-1
Package: S08N	8-Pin SOIC (Narrow)	11-2
Package: J14	14-Pin Hermetic DIP	11-2
Package: P14	14-Pin Hermetic DIP	11-3
Package: S14N	14-Pin SOIC (Narrow)	11-3
Package: J16	16-Pin Hermetic DIP	11-4
Package: P16	16-Pin Molded DIP	11-4
Package: S16N	16-Pin SOIC (Narrow)	11-5
Package: S16W	16-Pin SOIC	11-5
Package: J18	18-Pin Hermetic DIP	11-6
Package: P18	18-Pin Molded DIP	11-6
Package: S18W	18-Pin SOIC	11-7
Package: J20	20-Pin Hermetic DIP	11-7
Package: P20	20-Pin Molded DIP	11-8
Package: Q20	20-Pin Molded Leadless PCC	11-8
Package: R20	20-Pin SSOP	11-9
Package: S20W	20-Pin SOIC	11-9
Package: P22	22-Pin Molded DIP	11-10
Package: J24W	24-Pin Hermetic DIP	11-10
Package: J24N	24-Pin Hermetic Ceramic DIP (Narrow)	11-11
Package: P24W	24-Pin Molded DIP	11-11
Package: P24N	24-Pin Molded DIP (Narrow)	11-12
Package: R24	24-Pin SSOP	11-9
Package: S24W	24-Pin SOIC	11-12
Package: J28W	28-Pin Hermetic DIP	11-13
Package: P28W	28-Pin Molded DIP	11-13
Package: Q28	28-Pin Molded Leadless PCC	11-14
Package: R28	28-Pin SSOP	11-9
Package: S28W	28-Pin SOIC	11-14
Package: H32	32-Pin TQFP	11-15
Package: Q32	32-Pin Molded Leadless PCC	11-15
Package: S32W	32-Pin SOIC	11-16

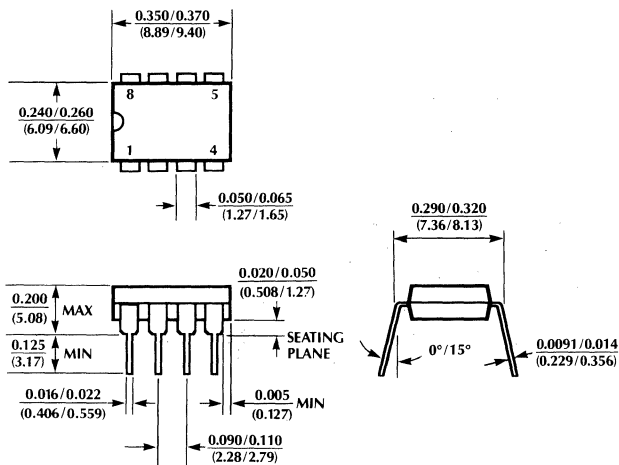
Package: J40	40-Pin Hermetic DIP	11-16
Package: P40	40-Pin Molded Plastic DIP	11-17
Package: G44	44-Pin PQFP	11-17
Package: Q44	44-Pin Molded Leaded PCC	11-18
Package: H44	44-Pin TQFP	11-18
Package: H48	48-Pin TQFP	11-15
Package: G52	52-Pin PQFP	11-17
Package: H52	52-Pin TQFP	11-18
Package: H64	64-Pin TQFP	11-18

PHYSICAL DIMENSIONS inches (millimeters)

Package: J08
8-Pin Hermetic DIP (CERDIP)

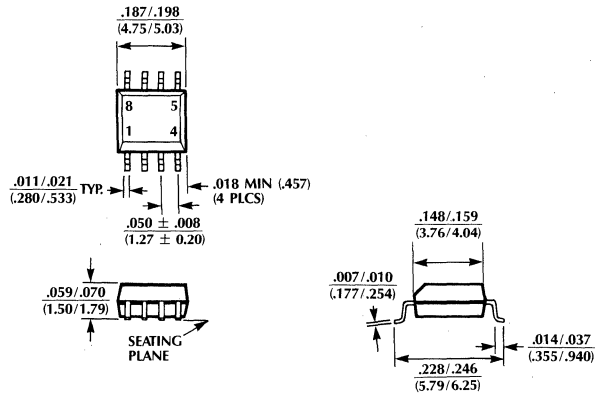


Package: P08
8-Pin Molded DIP



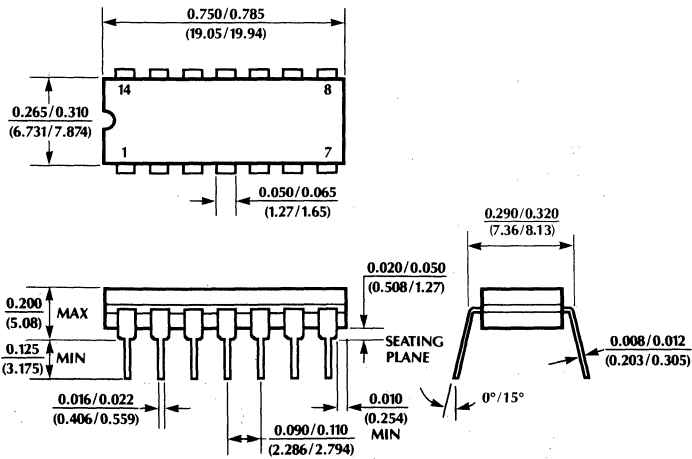
PHYSICAL DIMENSIONS inches (millimeters)

Package: S08N 8-Pin SOIC (Narrow)



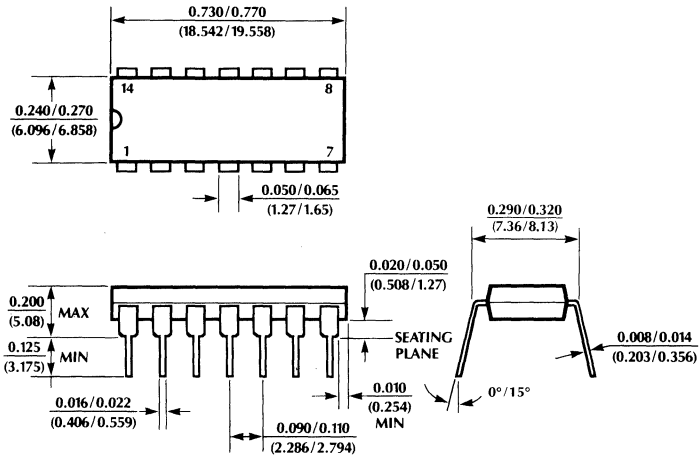
NOTE 1: SEATING PLANE LEAD COPLANARITY $.005$ (0.127) (BOTTOM OF LEADS).

Package: J14 14-Pin Hermetic DIP (CERDIP)

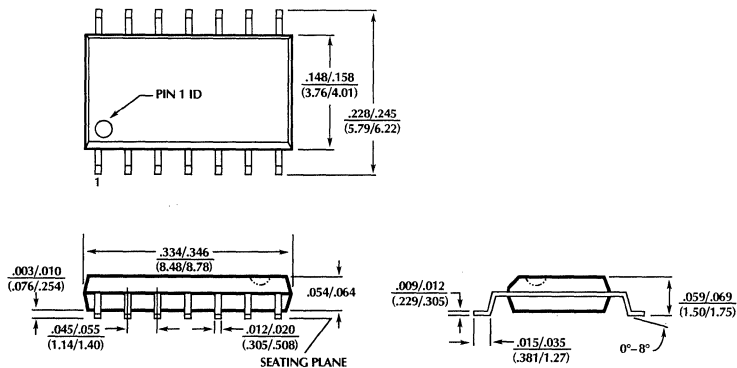


PHYSICAL DIMENSIONS inches (millimeters)

Package: P14 14-Pin Molded DIP



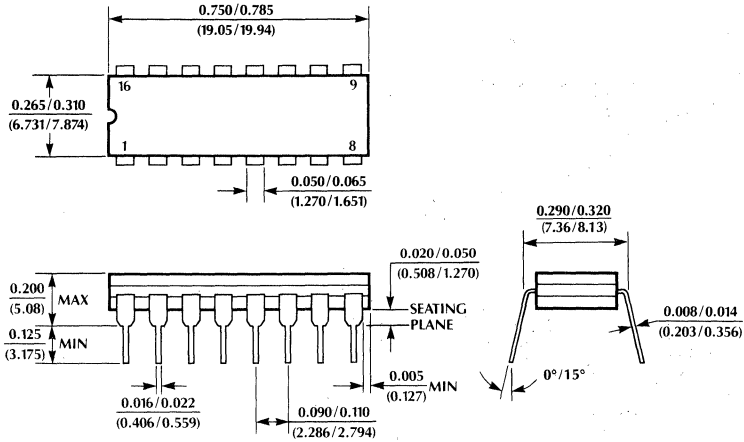
Package: S14N 14-Pin SOIC (Narrow)



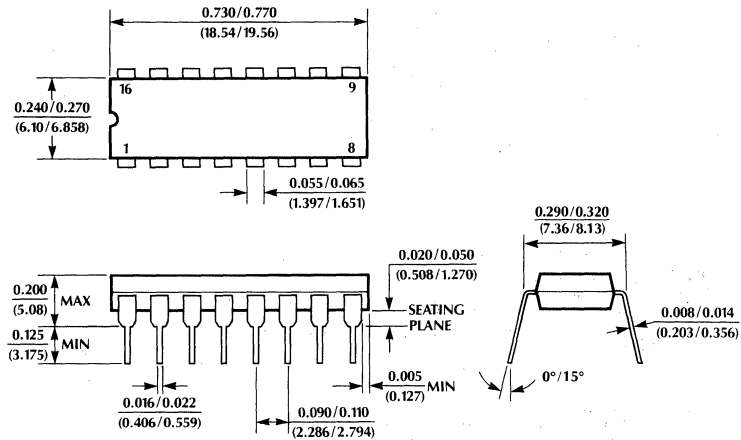
- NOTES:
1. Dimensions are in inches (millimeters)
 2. Seating plane coplanarity $\pm .005$ (bottom of leads after forming)
 3. Skewed or bent leads not to exceed $.005^\circ$ from its true position

PHYSICAL DIMENSIONS inches (millimeters)

Package: J16
16-Pin Hermetic DIP (CERDIP)

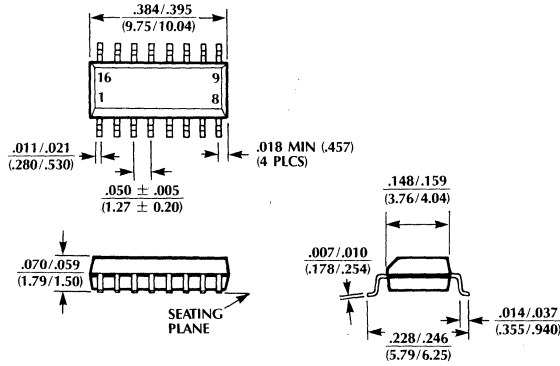


Package: P16
16-Pin Molded DIP



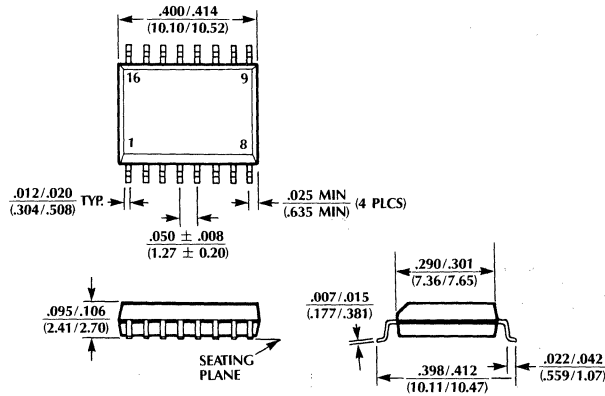
PHYSICAL DIMENSIONS inches (millimeters)

Package: S16N 16-Pin SOIC (Narrow)



NOTE 1: SEATING PLANE LEAD COPLANARITY .005 (0.127) (BOTTOM OF LEADS).

Package: S16W 16-Pin SOIC (Wide)

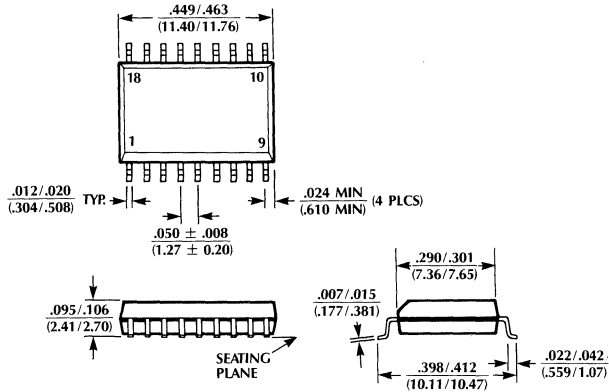


NOTE 1: SEATING PLANE LEAD COPLANARITY .005 (0.127) (BOTTOM OF LEADS).

1

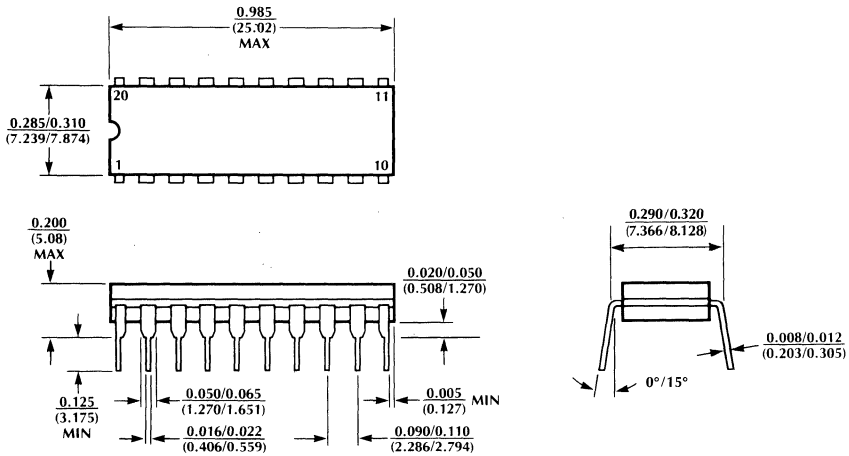
PHYSICAL DIMENSIONS inches (millimeters)

Package: S18W
18-Pin SOIC (Wide)



NOTE 1: SEATING PLANE LEAD COPLANARITY .005 (0.127) (BOTTOM OF LEADS).

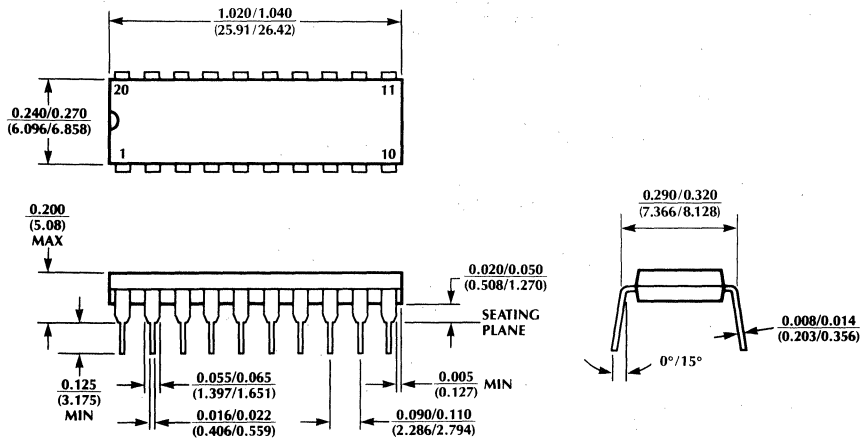
Package: J20
20-Pin Hermetic DIP (CERDIP)



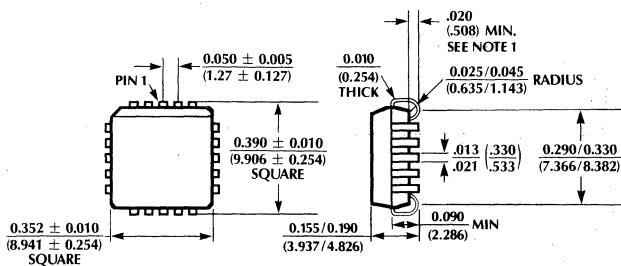
1

PHYSICAL DIMENSIONS inches (millimeters)

Package: P20
20-Pin Molded DIP



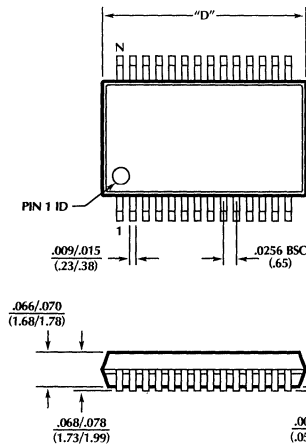
Package: Q20
20-Pin Molded Leaded PCC



NOTE: SEATING PLANE LEAD COPLANARITY .005".

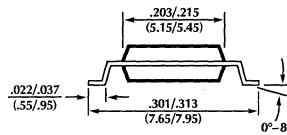
PHYSICAL DIMENSIONS inches (millimeters)

Packages: R20, R24, R28
20-Pin SSOP, 24-Pin SSOP, 28-Pin SSOP

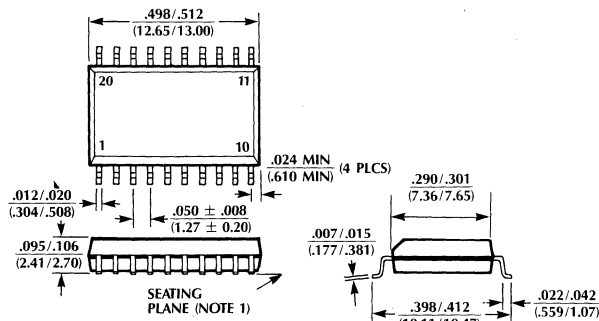


20L, 24L, 28L SSOP

"N" LEADS	inches	DIM. "D" (mm)	PACKAGE I.D.
20	.274/.294	(6.95/7.45)	R20
24	.313/.333	(7.95/8.45)	R24
28	.392/.412	(9.95/10.45)	R28



Package: S20W
20-Pin SOIC (Wide)

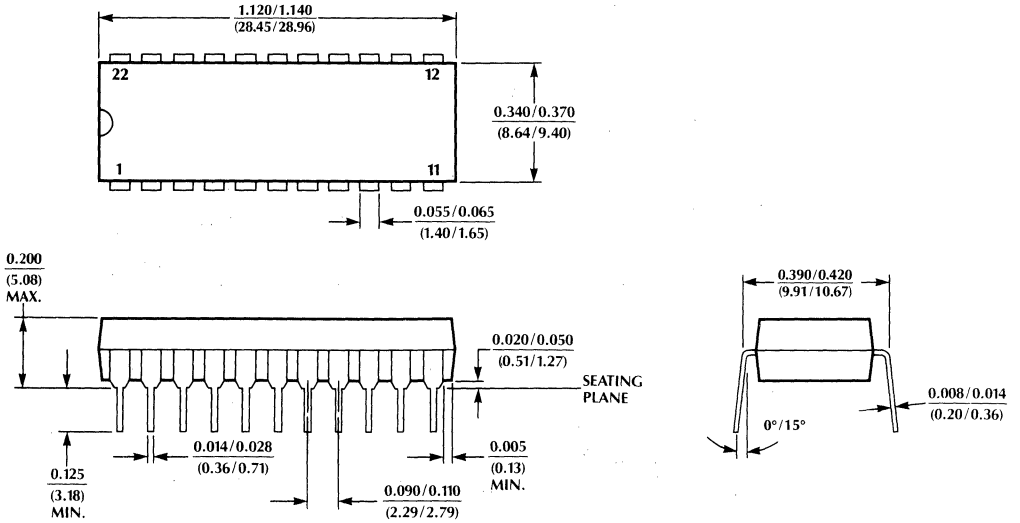


NOTE 1: SEATING PLANE LEAD COPLANARITY .005 (0.127) (BOTTOM OF LEADS).
NOTE 2: NOMINAL DIMENSIONS ±0.005 (0.127) UNLESS OTHERWISE SPECIFIED.

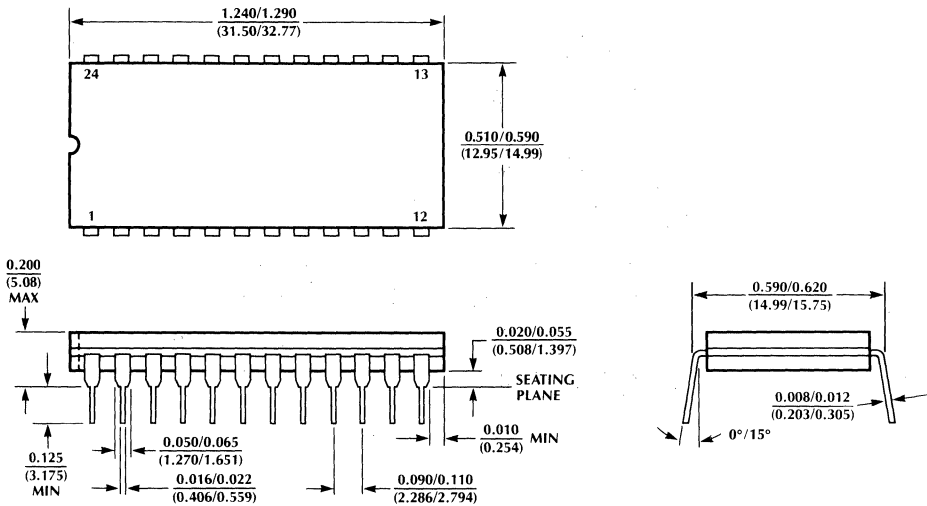
1

PHYSICAL DIMENSIONS inches (millimeters)

Package: P22
22-Pin Molded DIP

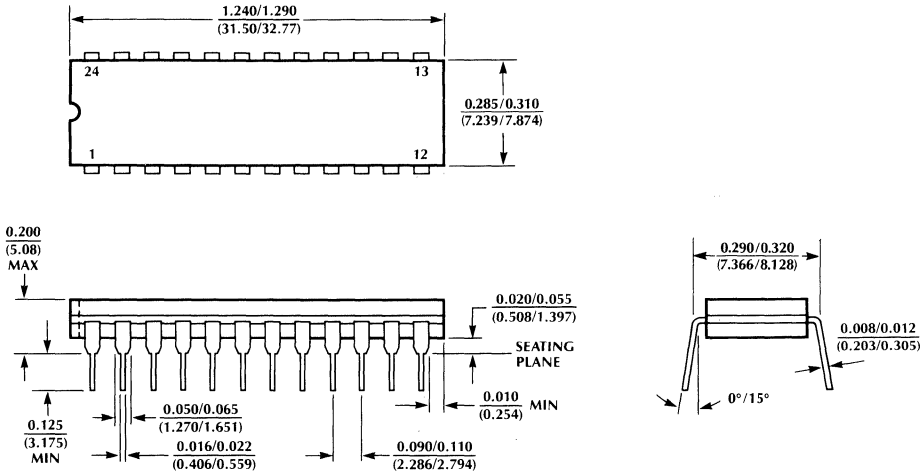


Package: J24W
24-Pin Hermetic DIP (CERDIP)

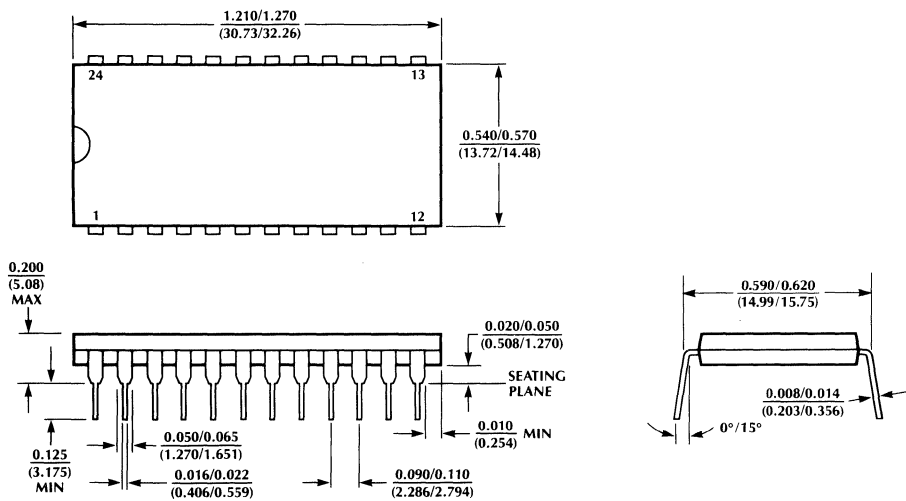


PHYSICAL DIMENSIONS inches (millimeters)

Package: J24N
24-Pin Hermetic Ceramic DIP (Narrow)

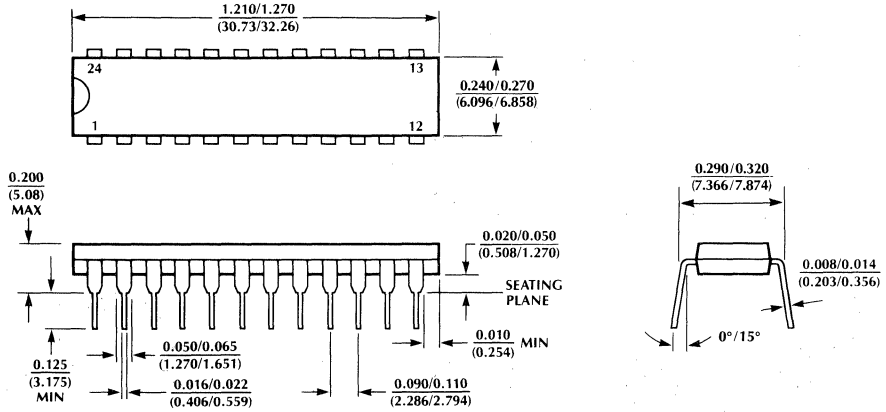


Package: P24W
24-Pin Molded DIP (Wide)

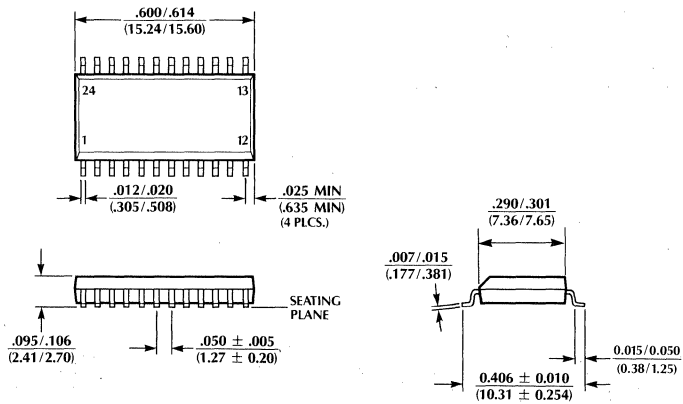


PHYSICAL DIMENSIONS inches (millimeters)

Package: P24N 24-Pin Molded DIP (Narrow)



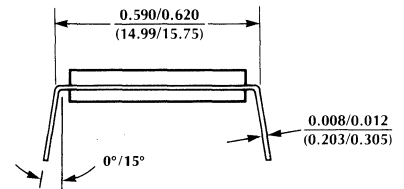
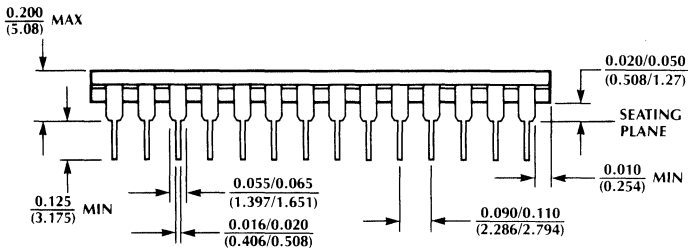
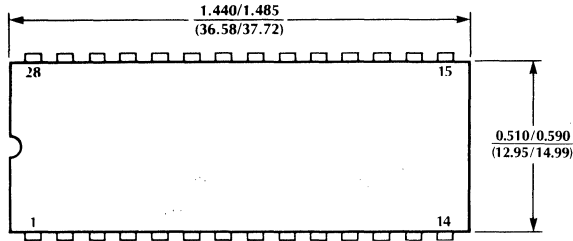
Package: S24W 24-Pin SOIC



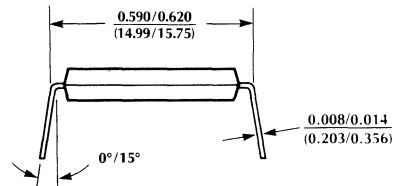
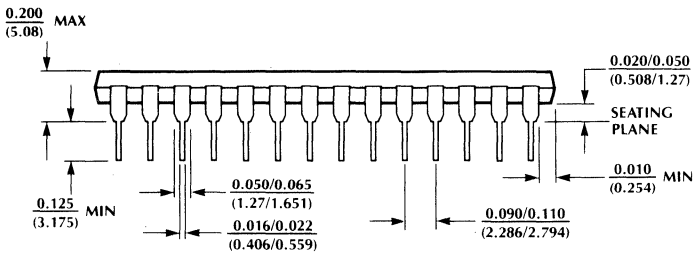
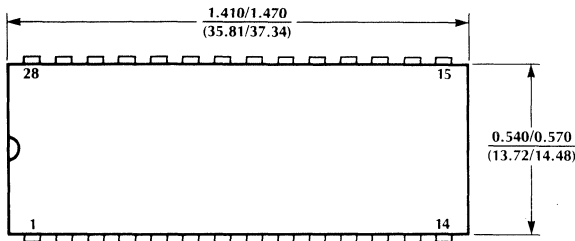
NOTE 1: SEATING PLANE LEAD COPLANARITY .005 (0.127) (BOTTOM OF LEADS).

PHYSICAL DIMENSIONS inches (millimeters)

Package: J28W
28-Pin Hermetic DIP (CERDIP)

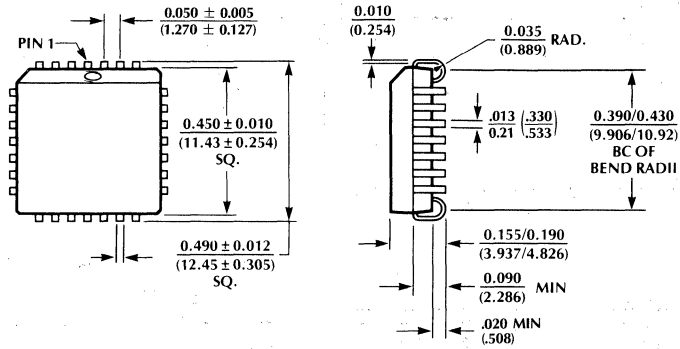


Package: P28W
28-Pin Molded DIP



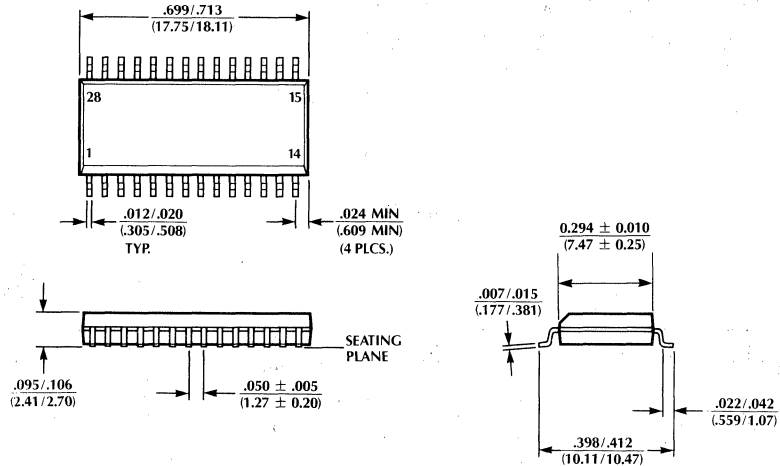
PHYSICAL DIMENSIONS inches (millimeters)

Package: Q28 28-Pin Molded Ledged PCC



NOTE: SEATING PLANE LEAD COPLANARITY 0.005.

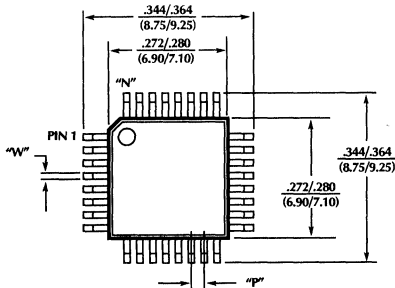
Package: S28W 28-Pin SOIC



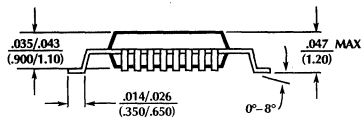
NOTE 1: SEATING PLANE LEAD COPLANARITY .005 (0.127) (BOTTOM OF LEADS).

PHYSICAL DIMENSIONS inches (millimeters)

Packages: H32, H48
32-Pin TQFP, 48-Pin TQFP

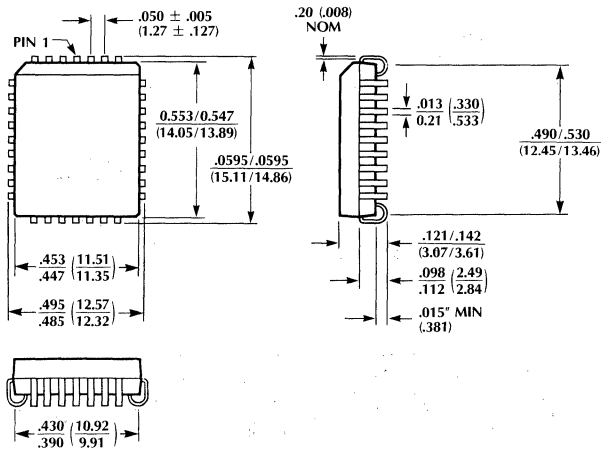


"N"	32L	48L
"p"	.315 BASIC (.80)	.0197 BASIC (.50)
"W"	.009/.015 (.23/.38)	.008/.012 (.20/.30)
PACKAGE ID	H32	H48



NOTES: 1) Lead frame copper thickness .005/(0.127)
2) Max .004/(.102) lead coplanarity

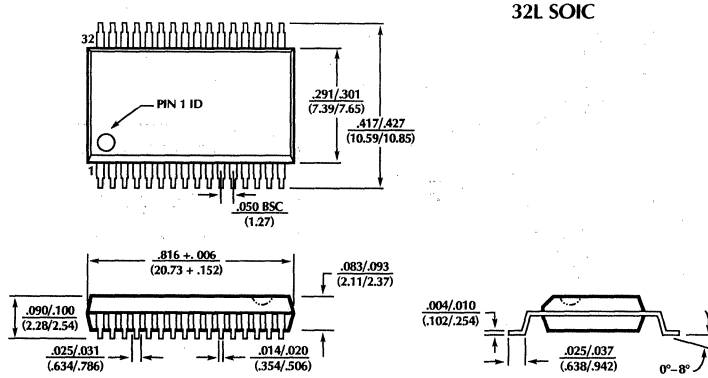
Package: Q32
32-Pin Molded Leaded PCC



1

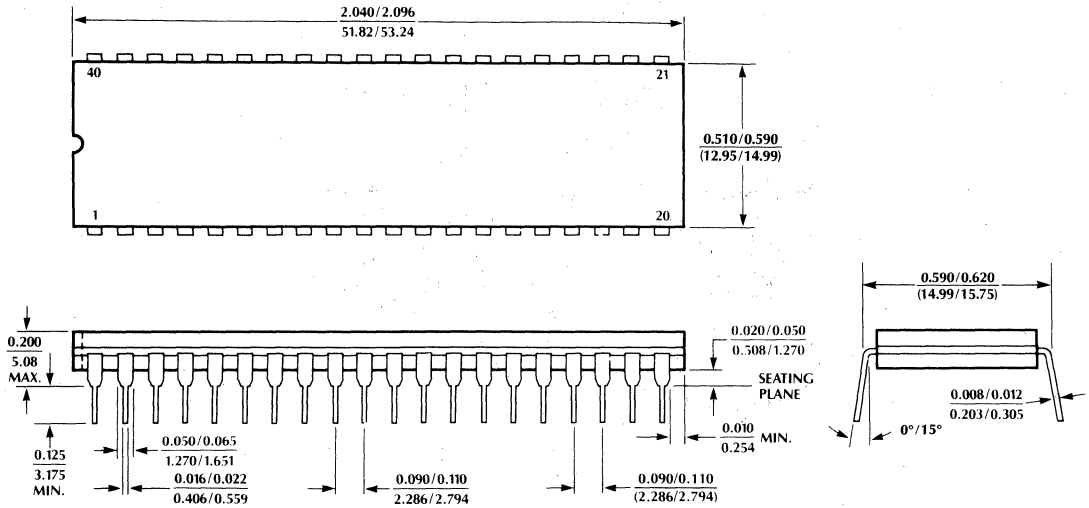
PHYSICAL DIMENSIONS inches (millimeters)

Package: S32W 32-Pin SOIC



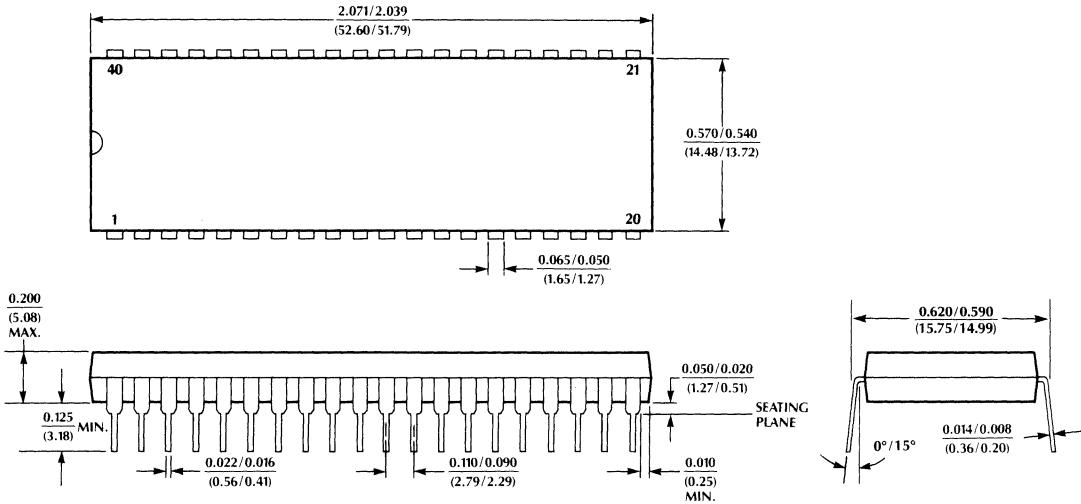
- NOTES:
1. Dimensions are in inches (millimeters)
 2. Lead coplanarity to be within .005" (.127mm)
 3. Max lead skew .005" (.127mm) from its true position

Package: J40 40-Pin Hermetic DIP (CERDIP)



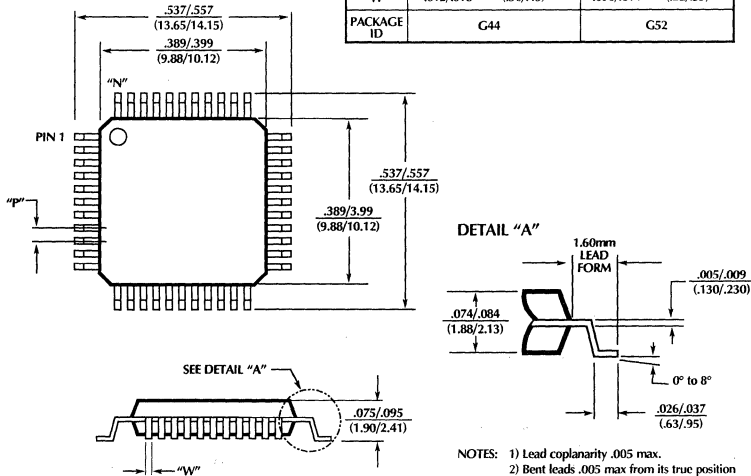
PHYSICAL DIMENSIONS inches (millimeters)

Package: P40
40-Pin Molded Plastic DIP

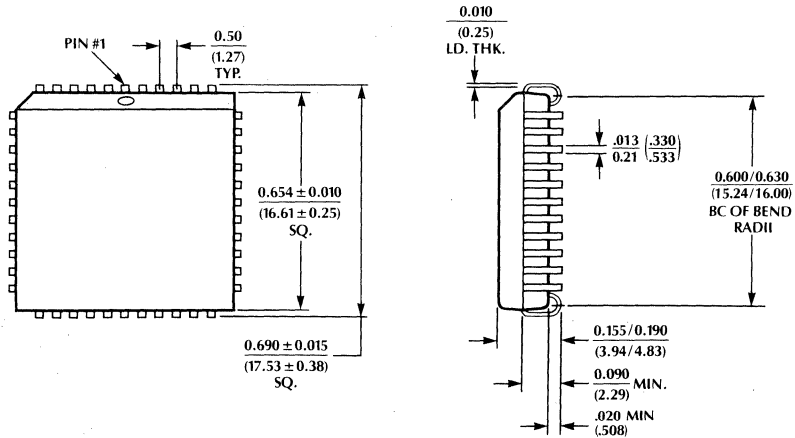


Packages: G44, G52
44-Pin PQFP, 52-Pin PQFP

"N"	44L	52L
"P"	.0315 (.80) BSC	.0265 (.65) BSC
"W"	.012/.018 (.30/.45)	.090/.014 (.22/.35)
PACKAGE ID	G44	G52



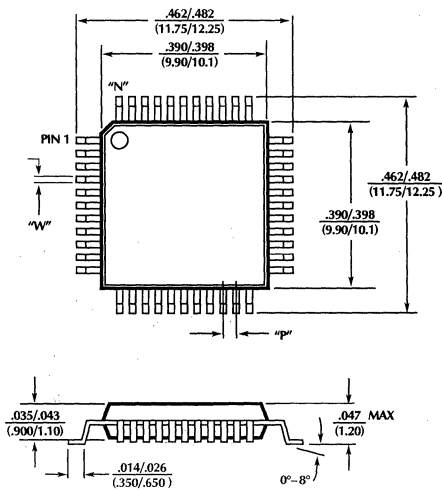
Package: Q44
44-Pin Molded Leaded PCC



NOTE:
 1. SEATING PLANE LEAD COPLANARITY ±.005

Packages: H44, H52, H64
44-Pin TQFP, 52-Pin TQFP, 64-Pin TQFP

44L, 52L, 64L TQFP
10 x 10 x 1.0mm BODY



"N"	44L	52L	64L
"p"	.80 BASIC (.0315)	.65 BASIC (.0256)	.50 BASIC (.0197)
"W"	.18/.32 (.007/.013)	.23/.37 (.009/.015)	.18/.32 (.007/.013)
PACKAGE ID	H44	H52	H64

NOTES:
 1) Lead frame copper thickness .005/(0.127)
 2) Max .004/(.102) lead coplanarity



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