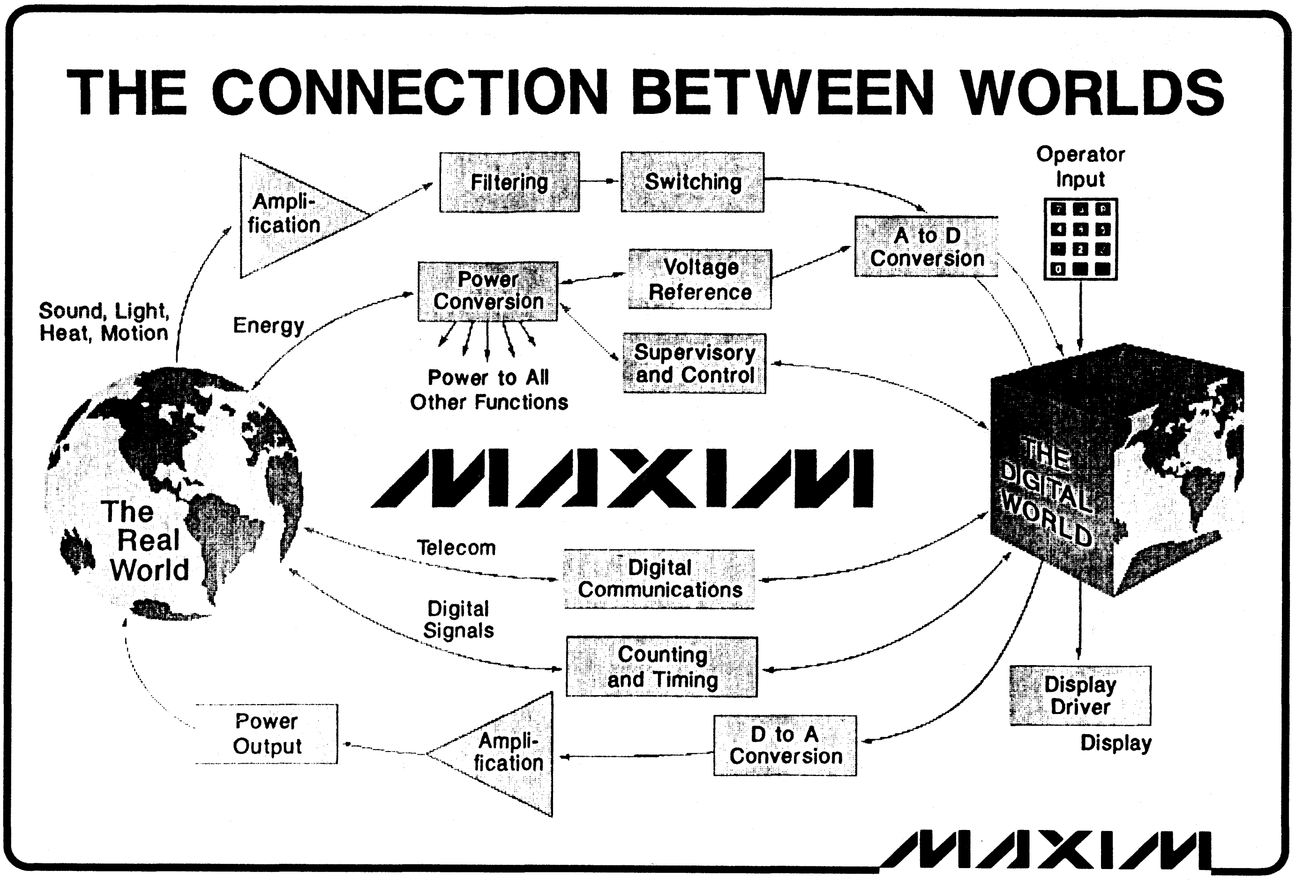
MAXIM

1988/89 Seminar Applications Book

A/D Converters
D/A Converters
Analog Switches
Analog Multiplexers
Active Filters
Power Supply Circuits
Video Products
RS-232/Interface Circuits
Op Amps/Buffers
Voltage References
Display Drivers
Timers/Counters



Maxim's product line covers a wide spectrum of functions involving amplification, measurement, and processing of analog signals and the interface between the analog and digital worlds. Maxim's expertise in both worlds makes it a complete source for analog, data conversion, and interface products as well as power conversion and control ICs. In solving real world problems, the analog and digital worlds really become one domain. Our goal is to do all that is needed to develop products that provide complete solutions for our customers applications.

Table Of Contents

COMPANY UPDATE

- Founded – April 1983
- Headquarters in Sunnyvale, California
- Public Offering – February 1988
- Mil Std 883 Qualification – July 1988
- 9 Successive Profitable Quarters
- Over 75% Annual Growth Rate

MAXIM

MIL-STD-883 REV C IS HERE NOW

- MAXIM CAN SUPPLY:
  - /883 Class B, Rev C
  - /HR Devices
  - Standard Military Drawings (SMDs)
  - Source Control Drawings (SCDs)
- Maxim is not presently supplying QPL devices (38510/ Sheets)

MAXIM

MIL-STD-883 REV C COMPLIANCE

- Manufacturing Facility
- Controlled Environment
- ESD Control
- Calibration System
- Configuration Control

MIL-STD-883 DEVICE TESTING

- 100% Screening Methods
- Sample Testing Methods

MAXIM

MILITARY QUALIFICATION

/883 REV C IS HERE NOW! The first qualified units are now in box stock.

For each device certified to 883 REV C, new internal data sheets are generated per 883 and DESC guidelines. Electrical test programs and burn-in procedures are verified for compliance to these data sheets.

The following device/package combinations are currently available:

AD581SH/883	AD584TH/883
AD581TH/883	OP07J/883
AD584TH/883	OP07AJ/883

Many more products are soon to follow.

Compliance rules govern both Maxim and its subcontractors. These must be audited to government standard specifications by Maxim QA. Maxim QA is responsible for audits and certifications which must be maintained and retained. Reaudits and audit reports are required annually.

MIL-STD-883 REV C COMPLIANCE RULES:

Manufacturing Facility	Mil-Std-976
Clean Room and Work Station Controlled Environment	Fed-Std-209
Electrostatic Discharge Control	Mil-Hnbk-217
Calibration System	Mil-Std-45662
Configuration Control – EngineeringChanges, Deviations, and Waivers	DOD-Std-480

Mil-Std-883 Class B Qualification testing is required to demonstrate that the product meets or exceeds the screening and sample test specifications of Mil-Std-883, Methods 5004 and 5005.

MIL-STD-883 DEVICE QUALIFICATION TESTING:

100% Screening	Method 5004
Sample Testing	Method 5005

WHAT MAKES MAXIM UNIQUE?

- Introduced more new products than any analog IC company over the past 5 years
- One of the most rapidly growing analog IC companies
- Provides military quality ICs to the commercial market
- Sales presence in all major worldwide markets
- Successful design and management teams that include many industry founders

MAXIM

DESIGN INNOVATION

- World class designers responsible for:
  - 110 successful designs prior to joining Maxim
  - Resulting in over **$500** million in marketplace sales
  - 85 Successful designs while at Maxim
  - More analog products introduced in last 5 years than any other company

MAXIM

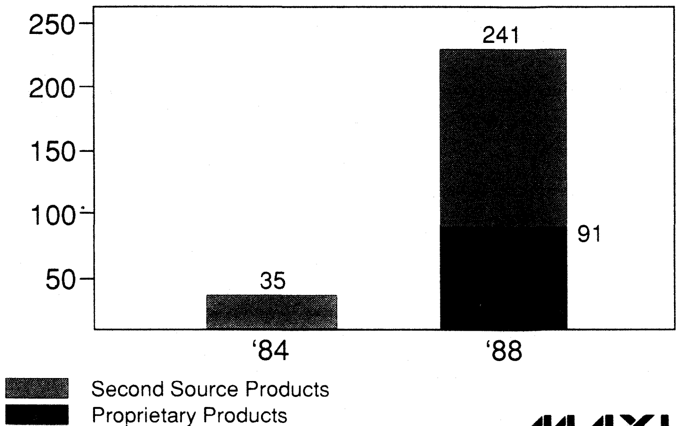
OVER 30 SIGNIFICANT INDUSTRY FIRSTS FOR MAXIM PERSONNEL

Including:

- First CMOS Analog IC
- First Mixed Digital/Analog IC
- First +5V RS-232 Interface IC
- First CMOS Chopper Amplifier
- First CMOS LED Display Drivers
- First Combined Video Amp/Multiplexer
- And 24 More!

MAXIM

GROWTH OF MAXIM PROPRIETARY PRODUCTS



MAXIM'S PRODUCT FAMILIES

- A/D and D/A Converters
- Power Supply Circuits
- Analog Switches, Multiplexers
- Analog Filters
- Op Amps, Buffers, Video Products
- Voltage References
- Timers, Counters, Display Drivers

NEW PRODUCT DEVELOPMENT

AREAS OF EMPHASIS:

- Increased speed
- Increased precision
- More system integration
- Serial interfaces
- Opto-isolated interfaces
- Battery operated systems

Engineering Solutions in Silicon

MAXIM QUALITY LEADS THE ANALOG INDUSTRY

- Extensive testing and quality control assures military quality for the commercial market at commercial prices
- All of Maxim's commercial products are fully traceable
- Maxim has received 883 certification

RESULT: 5.7 FAILURES PER BILLION OPERATING HOURS

MAXIM

Maxim is dedicated to providing its customers with the best available product at a reasonable price. At Maxim, the "best" product is judged by its functionality, conformance to specifications, performance over time, delivery, and acceptance by the customer. These attributes can be generalized by the terms "VALUE" and "QUALITY". Maxim's Quality Assurance Group oversees, and has explicit control of, all company operations.

Product quality is characterized at several different levels. We strive at Maxim to achieve the highest attainable level for our customers and ourselves. We believe that this is the only approach to developing mutually beneficial long term business relationships. This simple strategy has been instrumental in Maxim's growth as a leading supplier of quality products for the analog market.

QUALITY ASSURANCE AT MAXIM

- Review of Proposed Designs
- Qualify Raw Material Suppliers
- Qualify New Packages
- Monitor Manufacturing
- Document Control
- Personnel Training
- Failure Analysis

MAXIM

FAILURE ANALYSIS AT MAXIM

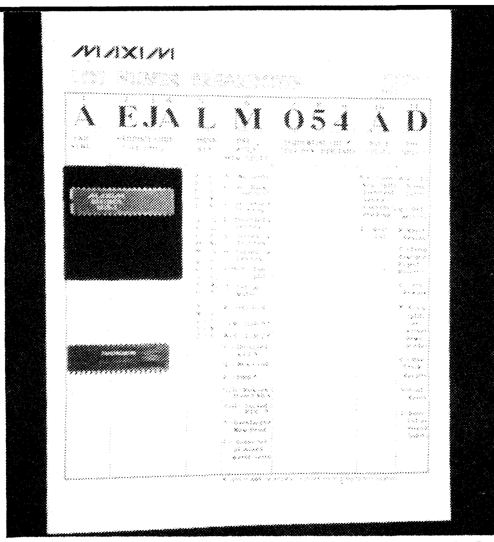
TO IDENTIFY	– Manufacturing Errors – Testing Inadequacies – Customer Design Problems – Design Weaknesses
TO ADVISE	– Customer Corrective Action
TO FIX	– Manufacturing Problems – Test Inadequacies – Design Flaws

MAXIM

VISUAL INSPECTION

Returned devices first undergo a visual inspection to determine:

1) Are they the correct part number? Manufacturer?
2) Are they adequately packaged? Antistatic?
3) Are the devices physically damaged?
4) Is there foreign material between or on the leads?
5) Are they fit for automatic testing?

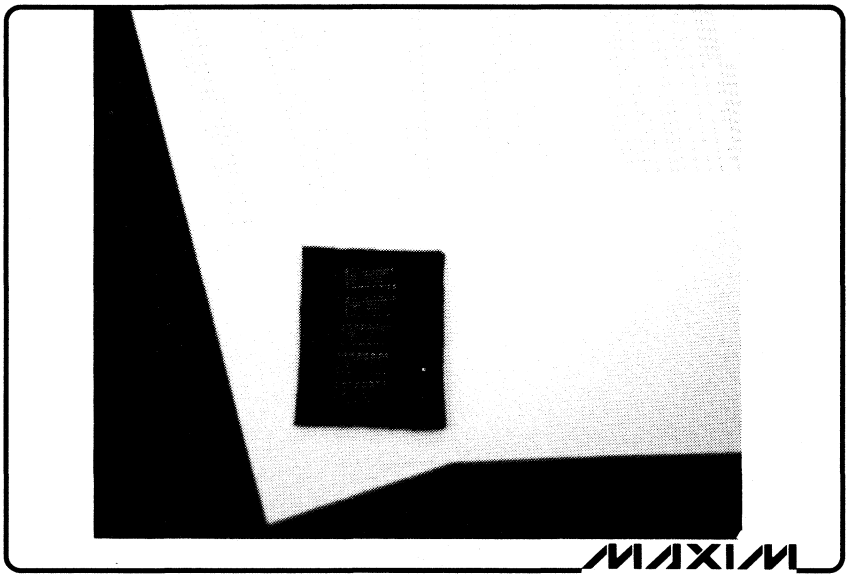


PACKAGE MARKING

Maxim's standard packaging includes two labels, a top mark and a back mark. The back mark traces to the manufacturing lot number. The lot number identifies the mask set, the product revision, and even the assembly location. All without opening the device.

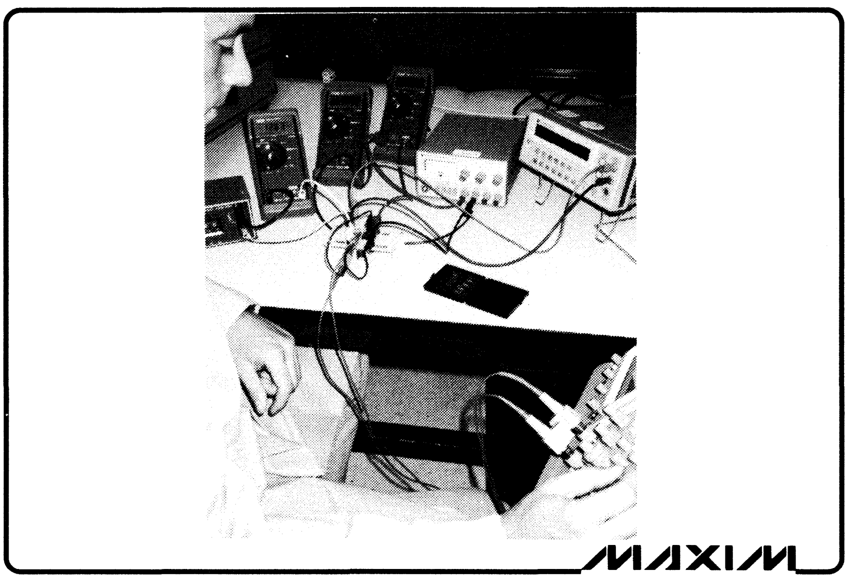
ELECTRICAL TEST

The first electrical test in the failure analysis procedure mimics the final production test for outgoing product at 25°C. This is done first (if possible) to check for gross device failures.



TEST RESULTS

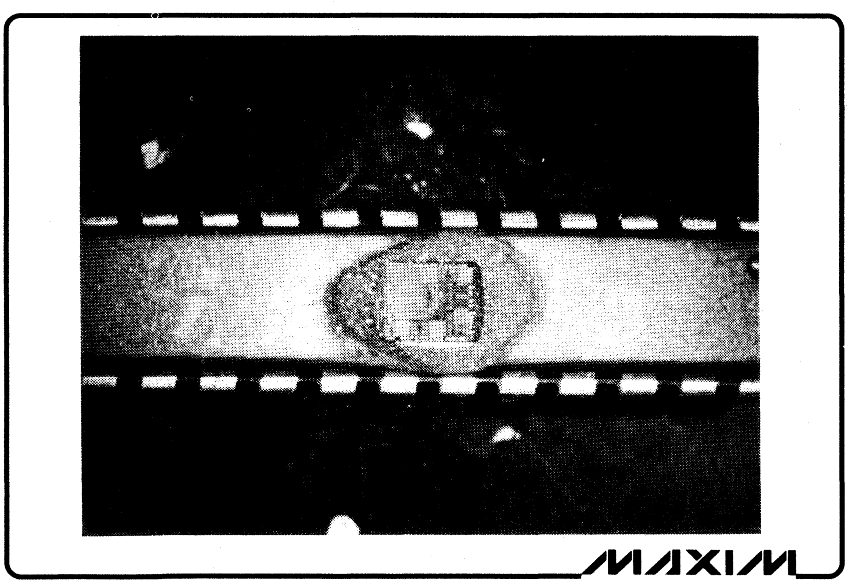
The test results are logged so that the performance of individual devices can be tracked.



BENCH TEST

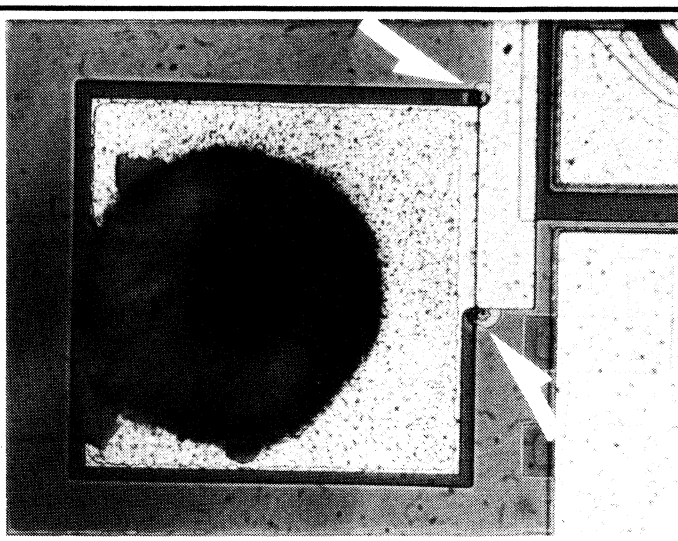
If a solution cannot be found in the automatic test results, the next step is a thorough bench check of the device(s). This may involve:

1) Verification of automatic test results.
2) Other tests that are impractical on automatic testers.
3) Duplication of customer circuitry.



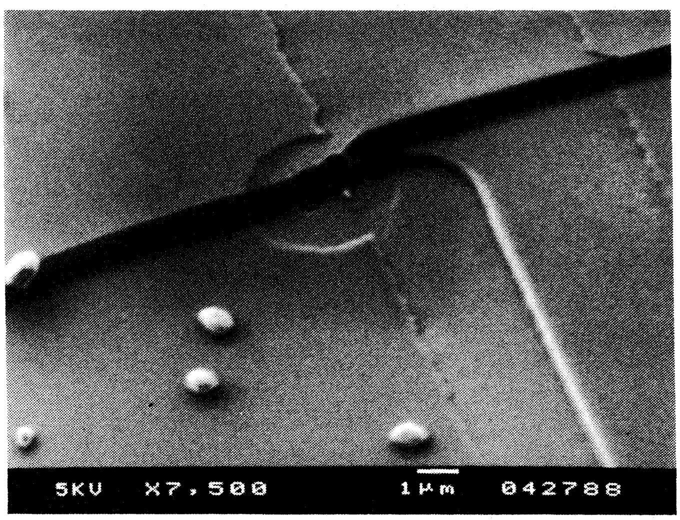
PACKAGE DECAP

After all electrical tests have been completed, the device is decapsulated for a detailed inspection of the die. For plastic packages, a controlled jet-etching process dissolves plastic to expose the die.



DIE EXAMINATION

The visual appearance of the damaged die area can provide clues as to the cause, or causes, of the failure: thermal stress, ESD, input overvoltage, etc. The location of damage on the die: analog input, digital input, output, supply metalization, etc., can also provide valuable information on the cause of trouble.



SEM INSPECTION

A SEM (Scanning Electron Microscope) examination reveals even more detail on the condition of the device.

THE REPORT

- Background
- Conclusion
- Verification
- Discussion
- Correction

FAILURE ANALYSIS REPORT

At the conclusion of the analysis, a detailed report of the findings is issued to the customer. The report contains five major sections:

Background - Customer's Conditions

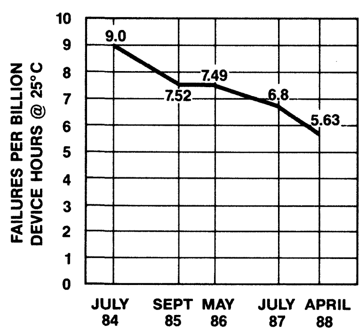
Conclusion - Summary of Report Facts and Results

Verification - Verify the Problem

Discussion - Failure Cause, Mode, Mechanism

Corrective Action - By Maxim and/or Customer

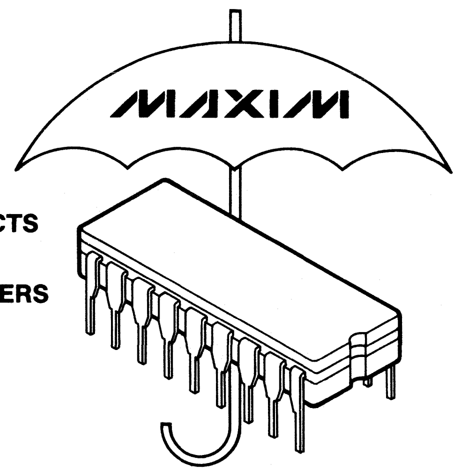
MAXIM'S RELIABILITY HISTORY



Maxim's "FIT" (Failures In Time) rate is continually improving. What is more significant is that the rate has decreased in the face of dramatically increasing production volume over the same period of time.

FIT rates are calculated from data collected during high temperature accelerated life tests. This data is collected from *every manufacturing lot*. Predicted failures at 25°C are correlated to measured data at 150°C using a calculated acceleration factor. Refer to Maxim publication RR-1D "Product Reliability Report" for further details on FIT measurements.

μP/MEMORY PROTECTION INTERFACE



- NEW MAX690 FAMILY PRODUCTS
- NEW RS-232 DRIVERS/RECEIVERS

MAX690 – MAX697 MICROPROCESSOR SUPERVISORY CIRCUITS

- Power Up Reset
- Brownout (Vcc < 4.5V) Reset
- Power Fail Warning
- Battery Switchover for CMOS RAM
- CMOS RAM and EEPROM Write Protect
- Watchdog Timer (Software Monitor)

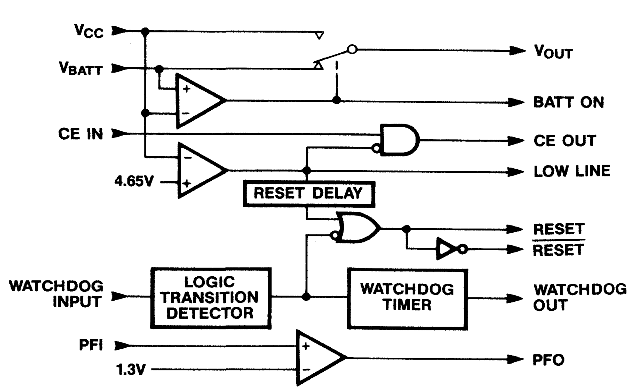
In industrial microprocessor-based systems, such as controllers and intelligent instruments, precise power monitoring and reliable reset signals are needed to maintain data integrity throughout various power faults and failures. Supervisory circuitry must ensure that critical data is not lost when power goes down. The requirements are much more strict than with consumer electronics products where a supervisory circuit might consist of nothing more than a capacitor connected to the microprocessor's reset input.

In high performance systems there is an increasing need for software monitoring. A watchdog timer does this by looking for activity on an I/O pin. If the processor fails or a software bug causes the system to hang up, the watchdog generates a reset signal. The MAX690-697 family of µP supervisory circuits is designed to perform the above functions with far fewer components and higher reliability than discrete circuitry.

THE MAX690 FAMILY FLEXIBLE OPTIONS

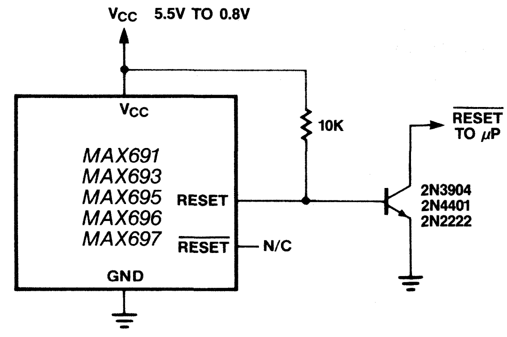
- 4.4V, 4.65V, or Adjustable Reset Threshold
- 50ms, 200ms, or Adjustable Reset Duration
- Automatic Battery Switchover/Memory Protect
- 8 or 16 pin DIP, Small Outline, Mil-Temp
- 160µA I_Q Version: MAX697
- 1µA Drain in Battery Backup Mode

MAX691 BLOCK DIAGRAM



The MAX690, MAX691, and other parts in this series, combine several popular microprocessor supervisory and memory protection functions on a single chip. In the past, getting this much performance on a printed circuit board meant designing in dozens of analog components and performing several adjustments.

VALID RESET TO 0.8V V_{CC}



The MAX690 family will operate, maintaining a valid reset, with V_{CC} as low as 2 volts. Some newer microprocessors, however (including the Motorola 68HC11), are somewhat prone to overwriting their own internal EEPROMs at very low voltages, so a 2V minimum is sometimes not low enough to ensure that data won't be lost. Maxim devices with an active high RESET output (MAX691/3/5/6/7), and with the aid of an external transistor, will generate a reset output that is valid as long as V_{CC} is 0.8V or greater.

Supervisory Power Supply Circuits

uP Reset, Power Fail Detector, Battery Switchover, and Watchdog Timer

Part Number	Pins	Reset Level (Volts)	Supply Current (mA)	Reset Delay (ms)	Battery Switch	RAM Protect	Low Line In	Low Line Out
MAX690	8	4.65	4	50	Yes	No	No	No
MAX691	16	4.65	4	50*	Yes	Yes	No	Yes
MAX692	8	4.40	4	50	Yes	No	No	No
MAX693	16	4.40	4	50*	Yes	Yes	No	Yes
MAX694	8	4.65	4	200	Yes	No	No	No
MAX695	16	4.65	4	200*	Yes	Yes	No	Yes
MAX696	16	Adj.	4	50*	Yes	No	Yes	Yes
MAX697	16	Adj.	.16	50*	No	Yes	Yes	Yes

* default; also adjustable

Voltage Detectors

Part Number	Description	Supply Voltage Range	Supply Current (typ/max)	Threshold Accuracy
MAX8211	Single Channel, Non-inverting voltage detector	2.0V to 16.5V	5µA/15µA	±3.5%
MAX8212	Single Channel, Inverting voltage detector	2.0V to 16.5V	5µA/15µA	±3.5%
ICL7665	Two Channels: one inverting, one non-inverting	1.6V to 16.0V	2.5µA/10µA	±7.5%
ICL7665A	Two Channels: one inverting, one non-inverting	2.0V to 16.0V	2.5µA/10µA	±2%

RS232 TOPICS

- MAX230-241 Features
- MAX232 Block Diagram
- The Extended RS232 Family
- Up to 8 Drivers/10 Receivers – MAX244-249
- Isolated RS-232 – MAX250/251/252

MAX230-MAX241 FEATURES

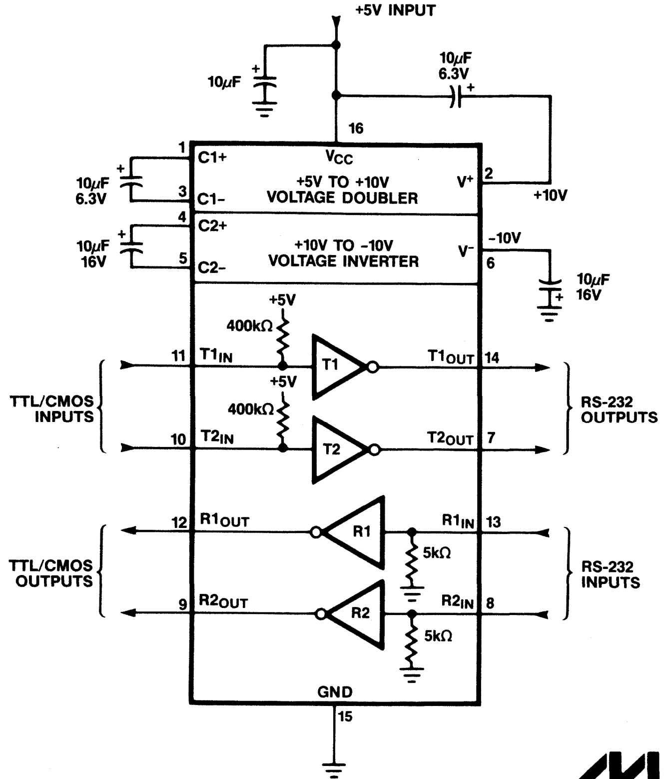
- Single +5V Supply (Except MAX231 and MAX239)
- Up to 5 Transmitters and 5 Receivers in One Device
- Low Power Shutdown (MAX230, 235, 236, 240 and 241)
- On-Chip Capacitors (MAX233, MAX235)
- Surface Mount (Except MAX233, MAX235)

The MAX23X and MAX24X families' key feature is that line drivers supply true RS-232 output levels even though the chips operate from a single +5V logic supply. The plus and minus supplies needed to meet the RS-232 minimum output specification of ±5V into 3kΩ are internally generated by an on-chip charge pump. As shown in the selector table, new versions have been designed to meet a variety of system requirements.

The MAX232 Family

Part Number	Power Supply	No. of RS-232 Drivers	No. of RS-232 Receivers	Ext. Caps.	Shut-down	Receivers Output 3-State	RS-232 Driver 3-State	Pins	Features
MAX230	+5V	5	0	4	Y	–	–	20	
MAX231	+5V and +7.5V to 13.2V	2	2	2	–	–	–	14	
MAX232	+5V	2	2	4	–	–	–	16	
MAX232A	+5V	2	2	4	–	–	–	16	12V/µs Slew Rate
MAX233	+5V	2	2	0	–	–	–	20	No Ext. Capacitors
MAX234	+5V	4	0	4	–	–	–	16	
MAX235	+5V	5	5	0	Y	Y	–	24	No Ext. Capacitors
MAX236	+5V	4	3	4	Y	Y	–	24	
MAX237	+5V	5	3	4	–	–	–	24	
MAX238	+5V	4	4	4	–	–	–	24	1488/1489 Repl.
MAX239	+5V and +7.5V to 13.2V	3	5	2	–	Y	–	24	No Ext. Capacitors
MAX240	+5V	5	5	4	Y	Y	–	44	
MAX241	+5V	4	5	4	Y	Y	–	28	
MAX242	+5V	2	2	4	Y	Y	Y	18	
MAX243	+5V	2	2	4	–	–	–	16	1 Rcvr Has Neg Threshold
MAX244	+5V	8	10	4	–	–	–	44	
MAX245	+5V	8	10	0	Y	Y	Y	40	1 Rcvr Always Active
MAX246	+5V	8	10	0	Y	Y	Y	40	1 Rcvr Always Active
MAX247	+5V	8	9	0	Y	Y	Y	40	1 Rcvr Always Active
MAX248	+5V	8	8	4	–	Y	–	44	2 Rcvr and Drvr Enable Inputs
MAX249	+5V	6	10	4	–	Y	–	44	2 Rcvr and Drvr Enable Inputs
MAX250	+5V	2	2	–	Y	Y	Y	14	Isolated RS-232 Chip Set
MAX251	+5V	2	2	–	Y	Y	Y	14	Isolated RS-232 Chip Set
MAX252	+5V	2	2	0	Y	Y	Y	40	Complete Isolated RS-232
MAX1080	+5V	2	2	4	Y	–	Y	18	LT1080 Equivalent

MAX232 BLOCK DIAGRAM



The MAX232 generates ± supplies with two flying-capacitor charge pumps. The first charge pump converts the +5V input to a nominal +10V at V^+. The second converts +10V to -10V. The ±10V supplies then power the RS-232 transmitters. With two drivers and two receivers, the MAX232 is suitable for RS-232 connections where both the DTE (Data Terminal Equipment) and the DCE (Data Communications Equipment) use one data line and one control line.

The MAX232 was introduced by Maxim in 1985. In 1986, 11 more devices were added to the family. These parts have additional drivers and receivers needed to implement many types of full RS-232 ports. Especially notable are the MAX233 and MAX235, which contain a patented scheme using internal charge pump capacitors that requires no external components.

THE EXTENDED RS232 FAMILY

- MAX232A: High Speed (12V/µs into 2500pF/3K Ω)
- **MAX240:** Transmitter Shutdown Three-State Receiver Output
- MAX243: High Speed, Negative Threshold on 1 Receiver
- MAX1080: 18 pin, Transmitter Shutdown

THE MAX244-MAX249

- Up to 8 Transmitters per Device
- Up to 10 Receivers per Device
- No External Capacitors (MAX245/246/247)
- Low Power Shutdown With One Receiver Always Active (MAX245/246/247)
- Two Transmitter, Two Receiver Enable Pins (MAX248/249)
- Two Complete PC-AT Serial Ports (MAX249)

MAX232A

The RS-232 family has been further expanded with the addition of the MAX232A. This device features a faster slew rate than the MAX232 while still meeting RS-232's 30V/µs maximum slew rate requirement to minimize ringing in long cables. In addition, transmitter outputs go to a high impedance state when the chip is powered down.

MAX243 – NEGATIVE THRESHOLD

The new MAX243 is pin compatible with the MAX232A, and differs only in that RS-232 cable fault protection is removed on one of the two receiver inputs. This means that control lines such as CTS and RTS can either be driven or left floating, and communication will not be interrupted. Different cables are not needed to interface with different pieces of equipment.

The input threshold of the receiver without cable fault protection is -0.8V rather than +1.4V. Its output goes positive *only* if the input is connected to a control line that is actively driven negative. If not driven it defaults to the 0 or "OK to send" state. Normally, the MAX243's other receiver (+1.4V threshold) is used for the data line (TD or RD), while the negative threshold receiver is connected to the control line (DTR, DTS, CTS, RTS, etc.).

Other members of the MAX232 family implement the optional cable fault protection as specified by RS-232 specifications (now the EIA-232D spec). This means that a receiver output goes high whenever its input is driven negative, left floating, or shorted to ground. The high output tells the serial communications IC to stop sending data. To avoid this, the control lines must either be driven or connected with jumpers to an appropriate positive voltage level.

3-STATE TRANSMITTERS

Several new devices also provide 3-state drivers for use in network applications. A logic level input controls this function. See the Selector Guide on page 13 for parts with this feature.

TWO PC SERIAL PORTS

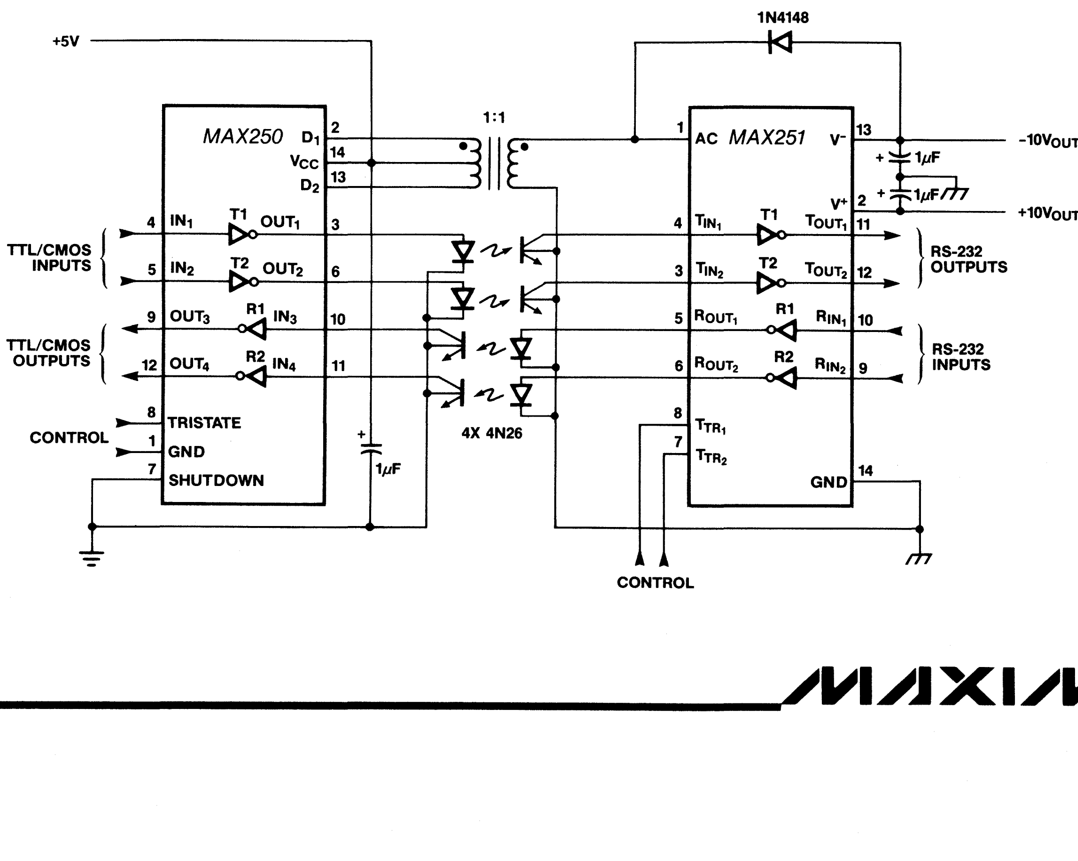
The MAX244 through MAX249 can implement two full "PC" compatible serial ports. The MAX245, MAX246, and MAX247 contain internal charge pump capacitors and need no external components.

ADVANCE NOTICE

MAX250, MAX251, AND MAX252 ISOLATED RS-232

- MAX250/251: 2 Chip Set Uses External Opto-Isolators and Transformer
- MAX252: Complete 1500V Isolated RS-232, Two Transmitters, Two Receivers, No External Components

MAX250-251 ISOLATED RS-232 INTERFACE



OTHER MAXIM INTERFACE PRODUCTS

Counters and Timers

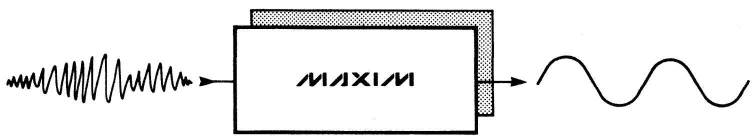
Part Number	Description	Maximum Count	Output	Speed (MHz max)	Supply Voltage	Supply Current	Features
ICM7217	4 Digit Up/Down	9999	C.A. LED	2	4.5V to 5.5V	350mA typ	Equals and Zero outputs, counter preset and pre-determining register set by thumb-wheel switches
ICM7217A	4 Digit Up/Down	9999	C.C. LED	2	4.5V to 5.5V	100mA typ	
ICM7217B	4 Digit Up/Down	5959	C.A. LED	2	4.5V to 5.5V	200mA typ	
ICM7217C	4 Digit Up/Down	5959	C.C. LED	2	4.5V to 5.5V	100mA typ	
ICM7224	4-1/2 Digit	19,999	LCD	15	3V to 16V	25µA max	
ICM7225	4-1/2 Digit	19,999	C.A. LED	15	3V to 6V	25µA max	
ICM7240	8 Bit Binary	1-255	open drain	15	2V to 16V	500µA max	RC oscillator or ext. clock
ICM7242	Fixed 8 Bit	128/256	CMOS	15	2V to 16V	500µA max	Programmable Time outs
ICM7250	2 Digit BCD	1-99	open drain	15	2V to 16V	500µA max	RC oscillator or ext. clock
ICM7260	2 Digit Timer	1-59	open drain	15	2V to 16V	500µA max	RC oscillator or ext. clock
ICM7555	CMOS 555 Timer		CMOS	0.5	2V to 18V	120µA max	
ICM7556	CMOS 556 Timer		CMOS	0.5	2V to 18V	240µA max	
MM74C945	4 Digit Up/Down	9,999	LCD	3	3V to 6V	60µA max	
MM74C947	4 Digit Up/Down	9,999	LCD	3	3V to 6V	60µA max	

NOTE:
C.A. LED = Common Anode LED Display
C.C. LED = Common Cathode LED Display

LCD and LED Display Drivers

Part Number	Input Formula	Display Formats	4-Digit	8-Digit	10-Digit	4-Char	5-Char	LCD	LED
MAX7231	6-bit Parallel	Hex, BCD, or Code B, plus 16 annunciators		X				X	
MAX7232	Bit Serial	Hex, BCD, or Code B, plus 20 annunciators			X			X	
MAX7233	6-bit Parallel	Upper Case ASCII				X		X	
MAX7234	Bit Serial	Upper Case ASCII					X	X	
ICM7211	µP and Muxed 4-bit	Hex, BCD and Code B	X					X	
ICM7212	µP and Muxed 4-bit	Hex, BCD and Code B	X						X
ICM7218/ ICM7228	8-bit Parallel	Hex, BCD, Code B, plus No Decode		X					X

ACTIVE FILTERS



- MAX260 SERIES PROGRAMMED FILTERS
- WHAT'S A SWITCHED CAPACITOR?
- STATE VARIABLE FILTERS
- FILTER DESIGN SOFTWARE

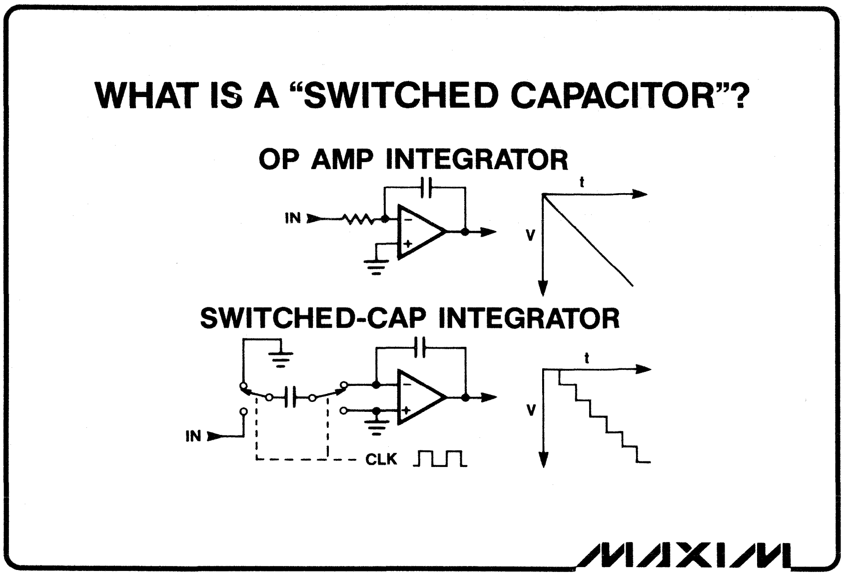
MAXIM'S FAMILY OF FILTER PRODUCTS

Typ fo Limit	2nd Order Universal µP Prog.	Pin Prog.	Res Prog.	Bandpass Pin Prog.	Lowpass Zero VOS
7.5kHz	MAX260				
20kHz					MAX280
30kHz			MF10		
57kHz	MAX261	MAX263	MAX265	MAX267	
140kHz	MAX262	MAX264	MAX266	MAX268	

MAXIM

Switched Capacitor Filters

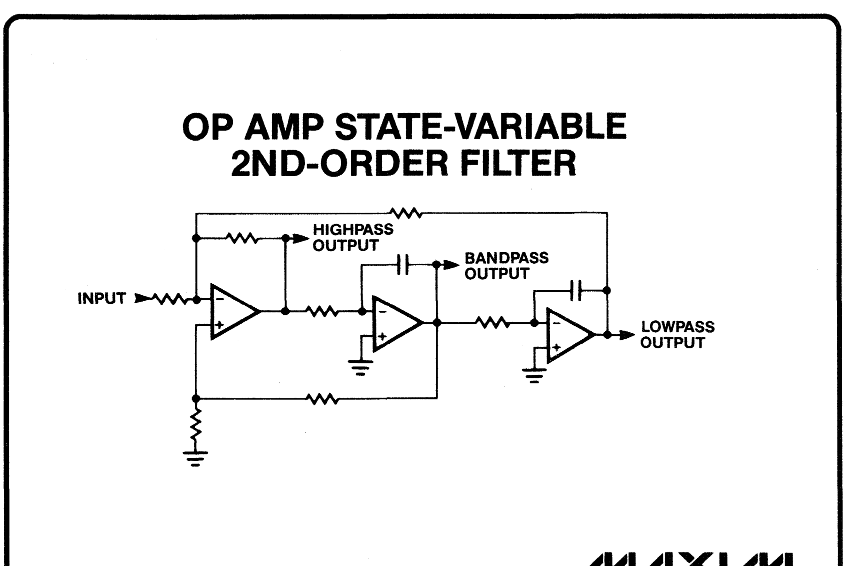
Part Number	Description	Analog Frequency Range	Features
MAX260	Universal Filter	0.01Hz to 7.5kHz	Microprocessor Interface
MAX261	Universal Filter	0.40Hz to 57.0kHz	Microprocessor Interface
MAX262	Universal Filter	1.0Hz to 140.0kHz	Microprocessor Interface
MAX263	Universal Filter	0.4Hz to 57.0kHz	Pin Strap Interface
MAX264	Universal Filter	1.0Hz to 140.0kHz	Pin Strap Interface
MAX265	Universal Filter	0.4Hz to 57.0kHz	Resistor and Pin Strap Interface
MAX266	Universal Filter	1.0Hz to 140.0kHz	Resistor and Pin Strap Interface
MAX267	Bandpass Filter	0.4Hz to 57.0kHz	Pin Strap Interface
MAX268	Bandpass Filter	1.0Hz to 140.0kHz	Pin Strap Interface
MAX280/ LT1062	Lowpass Filter	0Hz to 20kHz	0V DC Offset Error
MF10	Universal 2nd Order Filter	0.1Hz to 30.0kHz	Resistor Programmed



The integrator used in Maxim's switched-capacitor filters actually "simulates" a continuous op-amp integrator. The upper circuit is termed "continuous" because it operates in continuous time as opposed to the switched-cap circuit which samples in time. This circuit in fact only approximates the continuous case, but the approximation is usually better than 1% when the sample rate is high.

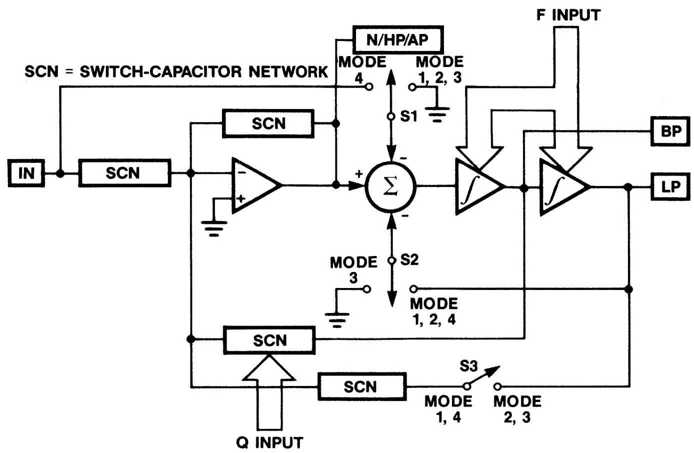
The input resistance of the continuous integrator is simulated in the lower circuit by the clocked (switched) input capacitor. The time constant of this integrator is directly proportional to the rate at which the switched-capacitor is clocked. The clock/center frequency ratio in Maxim's filters is also programmed by changing the integrator's time constant with different feedback capacitors. These are accurately programmed because precise capacitor ratios are fabricated on the monolithic filter chip.

The output of the switched-cap circuit is a staircase waveform which appears as a sampled version of the output of the continuous integrator. The clock is typically run at high speed with respect to the input signal (approximately 100 times) so that the staircase amplitude is small compared to the input.



The basic element in MAX26X series filters is the 2nd-order state-variable filter, implemented here with continuous integrators. A key advantage of this configuration is relatively independent control of filter parameters (hence the name "state variable") such as center frequency and Q. Other advantages include multiple outputs (highpass, lowpass, bandpass, notch and allpass) and flexible feedback configurations for different operating modes.

SWITCHED-CAP FILTER
- F AND Q PROGRAMMING
- MODE SWITCHING



Switched-capacitor integrators easily fit into the state-variable architecture. The switched-cap networks in the diagram (labeled "SCN") indicate where the closed loop gains of internal stages are controlled to program the cutoff/center frequency and filter Q. Filter operating modes 1 through 4 are controlled via switches S1, S2, and S3 and are programmed via the mode inputs, M0 and M1.

Two of these switched-cap state variable sections are provided in each MAX26X series filter. One chip makes a fourth-order filter with no external components. The center frequency and Q programming inputs are shown as well as the internal switches which route feedback signals for various operating modes.

MAX26X Filter Mode Features

Mode	1	2	3	3A	4
M1, M0 Input	0,0	0,1	1,0	1,0	1,1
Outputs	Lowpass Bandpass Notch	Lowpass Bandpass Notch	Lowpass Bandpass Highpass	Lowpass Bandpass Highpass Notch	Lowpass Bandpass Allpass
Notch Freq	f_0	$f_0\sqrt{2}$		$f_0\sqrt{R_H/R_L}$	
Lowpass Gain	-1	-0.5	-1	-1	-2
Bandpass Gain	-Q	$-Q\sqrt{2}$	-Q	-Q	-2Q
Notch Gain:					
as f – 0Hz	-1	-0.5		R_G/R_L	
as f – $f_{CLK}/4$	-1	-1		R_G/R_L	
Highpass Gain			-1	-1	
Typ Clock Limit*	2.7MHz	2.1MHz	1.7MHz	1.7MHz	2.7MHz

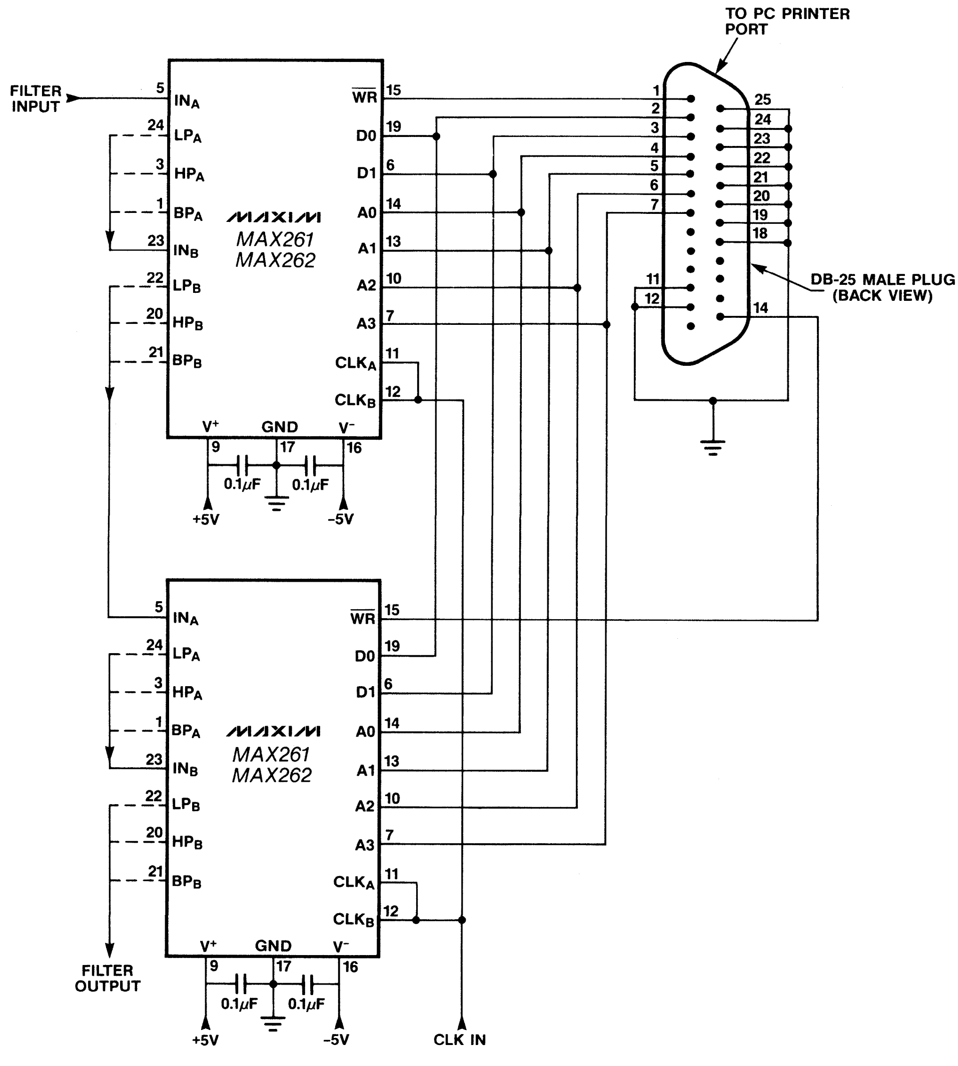
* The MAX260 Clock Limit is lower.

Throughout Maxim's filter data and design software literature, several references are made to "Filter Modes" or "Operating Modes". What are these? In the switched-capacitor state-variable circuit, several feedback connections may be modified with internal switches that are set via the "Mode" inputs. Each mode (The choices are Mode 1, 2, 3, 3A and 4) exhibits somewhat different performance characteristics as outlined in the above table.

MAX26X FAMILY FEATURES

- µP or Pin-Strap Programming
- Independent Frequency and Q Control
- Resistor Programming Also Available — MAX265/266
- 100µV Noise — 88dB Dynamic Range
- Frequency Range to 140kHz
- Highpass, Lowpass, Bandpass, Notch, Allpass

LOADING FILTERS FROM A PERSONAL COMPUTER FOR UP TO 8TH-ORDER FUNCTIONS



NOTE: PIN NUMBERS DIFFER FOR MAX 260. CONSULT DATA SHEET.

The MAX260, 261, and 262 are designed to be microprocessor programmed, however, for prototyping purposes, programming codes for f_{CLK}/f_C ratio, Q, and filter operating mode can be easily loaded into one or two filter ICs from a personal computer. The filter ICs are connected to the computer's parallel printer port as shown and the BASIC program listed on page 25 loads programming data directly to the chips. No interface circuitry is needed. The program asks for the filter programming codes and loads up to four 2nd-order sections in turn. Programming codes for specific filter designs can be generated by Maxim's filter design software or can be obtained from tables in the filter data sheets.

FILTER LOADING BASIC PROGRAM

```
10  PRINT "PRINTER PORT ADDRESS FOR LPT1: "
20  PRINT "          3BC  ( PORT ON DISPLAY ADAPTER )     =0 "
30  PRINT "          378  ( PORT NOT ON DISPLAY ADAPTER ) =1 ";
40  INPUT P : IF P>2 OR P<0 THEN 10
50  IF P=0 THEN PORT=956   ELSE PORT=888 ' SET PORT ADDRESS
60  FOR CHIP=1 TO 2
70  PRINT "CHIP # ";CHIP
80  AB$="FILTER A "
90  GOSUB 180 : REM GET DATA FOR SECTION A
100 ADD=0     : REM FILTER A ADDRESS
110 GOSUB 290 : REM WRITE DATA TO THE PRINTER PORT
120 AB$="FILTER B "
130 GOSUB 180 : REM GET DATA FOR SECTION B
140 ADD=32    : REM FILTER B ADDRESS
150 GOSUB 290 : REM WRITE DATA TO THE PRINTER PORT
160 NEXT
170 GOTO 60
180 PRINT "MODE (1 TO 4, SEE TABLE 5) "; AB$;
190 INPUT M
200 IF M<1 OR M>4 THEN 180
210 PRINT "CLOCK RATIO (0 TO 63, N OF TABLE 2) "; AB$;
220 INPUT F
230 IF F<0 OR F>63 THEN 210
240 PRINT "Q (0 TO 127, N OF TABLE 3) "; AB$;
250 INPUT Q
260 IF Q<0 OR Q>127 THEN 240
270 PRINT
280 RETURN
290 X=(ADD+M-1)
300 GOSUB 450
310 ADD=ADD+4
320 FOR I=1 TO 3
330 X=ADD+(F-4*INT(F/4))
340 GOSUB 450
350 F=INT(F/4)
360 ADD=ADD+4
370 NEXT I
380 FOR I=1 TO 4
390 X=ADD+(Q-4*INT(Q/4))
400 GOSUB 450
410 Q=INT(Q/4)
420 ADD=ADD+4
430 NEXT I
440 RETURN
450 '>>>>>>>  OUTPUT  <<<<<<<<<<<<
460 OUT PORT,X
470 OUT PORT+2,CHIP
480 OUT PORT+2,0
490 RETURN
```

This program asks for programming input instructions for up to 4 filter sections (up to 2 filter ICs). It asks for: MODE (Type in 1, 2, 3, or 4), CLOCK RATIO (f_{CLK}/f_0, Type in 0 to 63), and Q (Type in 0 to 127). In the program prompts there are references to "TABLE 2", "TABLE 3", and "TABLE 5". These can be found in the MAX260/261/262 data sheet. Programming numbers can be taken from these tables or can be generated in Maxim's Filter Design Software.

The filter-loading BASIC program loops through all four filter sections in sequence. A section is loaded when the Q for that section is entered. Other sections remain unchanged when any one is reloaded. Also, the program does not "end" but merely waits for reprogramming data for the next section in the sequence. If no new instructions are entered, the filter's programmed state does not change.

FILTER DESIGN SOFTWARE

- Reduce Design Time of High-Order Filters
- Evaluate Filter Designs
- Supplies Programming Data for MAX26X Filters
- Color Graphics For Viewing Filter Response
- Any PC-DOS or MS-DOS Computer

MAXIM

FILTER SOFTWARE – OPENING MENU

```
--------------------------------------------------------------------------
                     MENU OF AVAILABLE OPTIONS
--------------------------------------------------------------------------
(0) EXIT the program
--------------------------------------------------------------------------
(1) PZ      a routine to evaluate poles, Qs, and zeros
--------------------------------------------------------------------------
(2) RP      to evaluate the resistors needed for filter performance
            for general purpose resistor programmable filter circuits
--------------------------------------------------------------------------
(3) RPCHECK       the reverse of RP, given the resistors, clock frequency
            etc. calculate the performance
--------------------------------------------------------------------------
(4) MPP     for microprocessor controlled parts and pin programmable part
            calculate the digital settings etc. to get required performan
--------------------------------------------------------------------------
(5) FR      after designing one or more stages check the performance
            against the requirements
--------------------------------------------------------------------------
(6) BP      to determine resistor values for bandpass filters using the
            special multiple feedback technique
--------------------------------------------------------------------------
(7) FV    Screen graphics    (8) HELP instructions    (9) Browse in files
--------------------------------------------------------------------------
Choose the appropriate number from the above: 1
```

Designing all but the most basic active filters can become a complex process. Maxim's switched-capacitor filters highlight this point because the circuitry needed for most designs is minimal but the filter stages must still be told what to do via programming. The proper programmed frequencies and Qs for a high performance filter may be far from obvious if the designer has no general filter design experience.

Maxim supplies a series of software programs for use on a personal computer which simplify the design procedure for a wide variety of filters. Programs are provided which turn response specifications into filter programming codes. Also, the frequency response of proposed designs can be checked, and response plots can be viewed and printed.

FILTER SOFTWARE – PROGRAM "PZ" INPUT

```
Type in the shape of the filter, using the initial letter
L = lowpass, H = highpass, B = bandpass, N = notch : B
--------------------------------------------------------------------------
Type in the polynomial for the filter, using the initial letter
B = Butterworth, C = Chebyshev, E = Elliptic : C
--------------------------------------------------------------------------
For this filter, which parameters do you know?

   1. Upper and lower frequency limits of the passband and stopband

   2. Center frequency , pass bandwidth and stop bandwidth

   Select 1 or 2 :  2
--------------------------------------------------------------------------
Enter the maximum flatband ripple, Amax, in db : 1
--------------------------------------------------------------------------
Enter the center frequency ( FCNTR ) in Hz : 1000
--------------------------------------------------------------------------
Enter the passband bandwidth ( BW ) in Hz : 600
--------------------------------------------------------------------------
Do you know the order of the filter? (Y/N) : N
--------------------------------------------------------------------------
Enter the minimum stopband attenuation ( Amin ) in db : 20
--------------------------------------------------------------------------
Enter the  stopband bandwidth in Hz : Hz 2000
```

An example of how one part of the software operates is provided on the next several slides. The "PZ" program is usually the first step taken in a design because it begins with questions on the type and performance of the filter to be designed. The requirement in this example is for a bandpass filter at 1kHz which has a 600Hz bandwidth (This is also Example 1 in Maxim's publication UM-2, the "MAX260-268 Filter Design Software User Manual", page 16). The selected stopband is 2kHz, and passband ripple is 1dB.

FILTER SOFTWARE – PROGRAM "PZ" OUTPUT

```
FILTER TYPE: CHEBYSHEV BANDPASS
specified parameters:
Center Frequency is  1000.000     Pass Bandwidth is   600.0000
    Stop Bandwidth is  2000.0000
Amin=20.00 db  Amax= 1.0000 db
Calculated Parameters:
Pass Band Low Edge is   744.0307   Pass Band High Edge is  1344.0307
Stop Band Low Edge is   414.2136   Stop Band High Edge is  2414.2136
Order=   4  N=   2   Actual Attenuation=20.70
Resonant Frequencies in Hertz:
        Pole            Q
      764.3366        3.1469
     1308.3242        3.1469
      . . . . . . Screen stopped for viewing, press return to continue
Gain adjustment factor for bandpass = GAF =    3.930
For bandpass filters, the gain of the system at center frequency is give
in terms of the gains of the individual stages by the expression:
Bandpass gain at center frequency= H0_BP_1*H0_BP_2* . . . *H0_BP_N / GAF
A more detailed summary can be found on file "fout\PZ1.R"

Do you wish to print out a summary of the design? (Y/N) : N
-----------------------------------------------------------------------
Do you want to check frequency response ? (Y/N) : Y
```

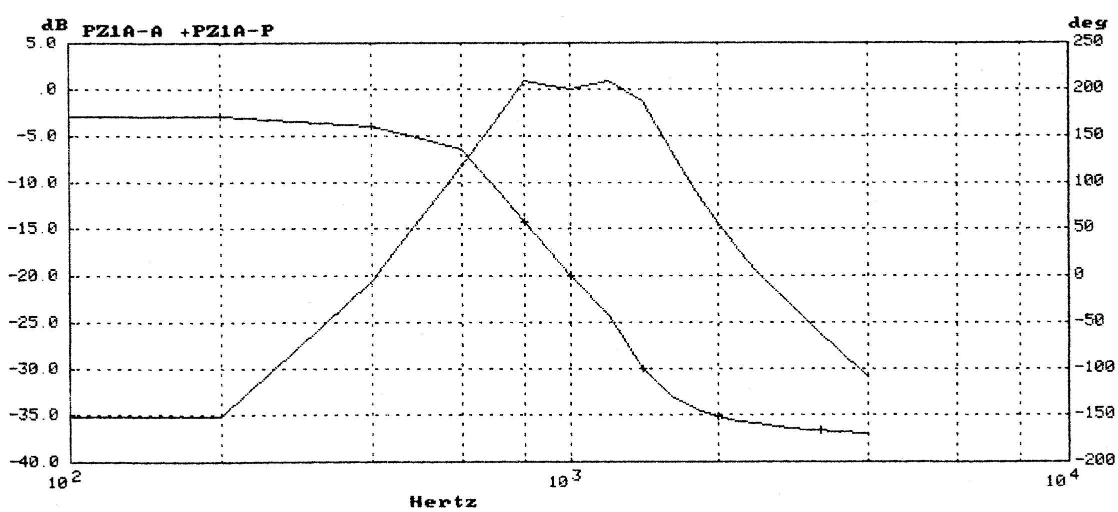
The PZ program responds by telling us:

1) Passband and stopband upper and lower edges
2) The order of the filter
3) The actual attenuation (-20.7dB)
4) Pole locations and Q values
5) Gain adjustment factor (GAF), used to determine the filter gain. The overall filter gain, $A_F = (H_{0BP1} \times H_{0BP2})/GAF$

Where H_{0BP1} and H_{0BP2} are the gains of each section at their own center frequency. Since the gain of these bandpass sections is equivalent to their Q, the filter gain in this example is:

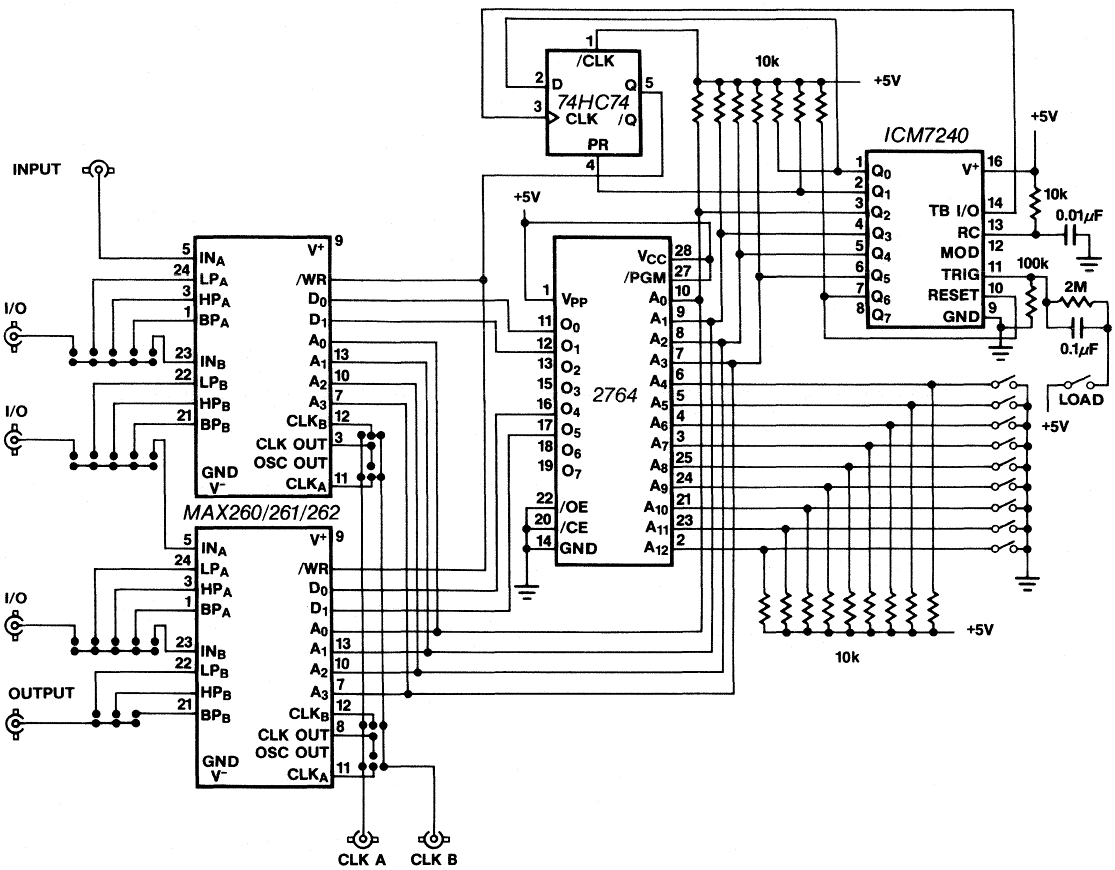
$A_F = (3.1469 \times 3.1469)/3.938 = 2.5147 = +8dB$

FILTER SOFTWARE – OUTPUT RESPONSE PLOT



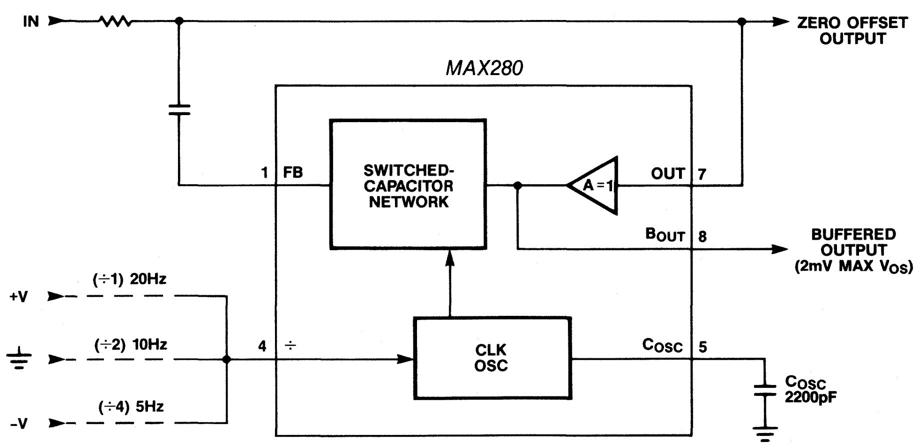
The gain and phase of the filter can be also be plotted by the filter viewing program, FV, included with the design software package. Response curves can be run for different frequency bands if desired.

LOADING MAX260/261/262 WITHOUT A μP



If a processor is not available, a MAX260/261/262 can be loaded from an EEPROM using some discreet logic. Toggle switches select the EEPROM address to be loaded and then the LOAD button transfers the data to the filter. With two filter ICs, any 8th-order filter function can be designed. The EEPROM must be preprogrammed with the proper codes for the desired filter, but up to 256 separate filter designs can be stored in one memory chip. Pin numbers are shown for the MAX261/262. The MAX260 can also be used, but the pin numbers are somewhat different.

MAX280 DC ACCURATE LOWPASS
• TYPICAL APPLICATIONS— 20Hz/10Hz/5Hz ROLLOFF



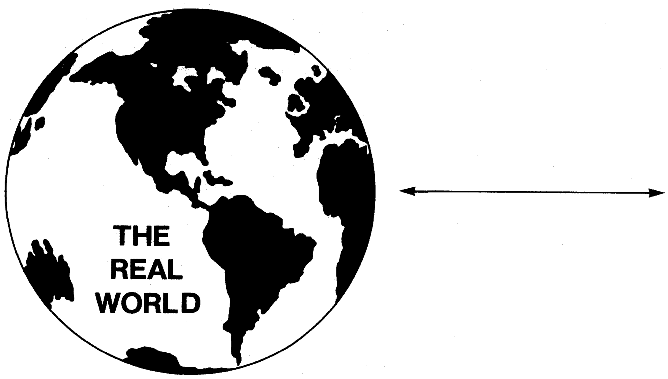
$$C_{OSC} = \left[\frac{140kHz}{f_{osc}} - 1\right] 33pF$$

The MAX280 eliminates DC offsets in lowpass filter applications. This is particularly important in high resolution measurement and instrumentation designs because low noise and low DC errors are often both required. The MAX280 achieves 0V offset by operating its switched capacitor network out of the main signal path. The filter input and output connect to an external resistor, and one external capacitor to form a 5th-order Butterworth lowpass. In addition to the 0V offset output, a buffered output (Pin 8) is also supplied which has a maximum offset of 2mV.

MAX280 FEATURES

- 5th-Order Butterworth Lowpass Filter
- 0mV DC Offset Voltage
- 2mV Max Offset with Internal Buffer
- 0 to 20kHz Input Range
- 10ppm/°C Cutoff Frequency Drift
- Internal Clock
- 8 Pin Package

DATA ACQUISITION



- NEW INTEGRATING A/Ds
- TESTING NOISE AND LINEARITY
- 12-BIT CONVERSIONS AT 500kHz
- CMOS DAC APPLICATIONS

MAXIM

Integrating A/D Converters

Part Number	Resolution	Output Type	Supply Voltage	Supply Current (Typ/Max mA)	Comments	Reference
MAX130	3 1/2 Digit ±2000 Counts	LCD	+4.5V to 14V	0.1/0.25	Replacement for ICL7106	Low T.C. Bandgap
MAX131	3 1/2 Digit ±2000 Counts	LCD	+4.5V to 14V	0.06/0.1	Replacement for ICL7136	Low T.C. Bandgap
MAX133	3 3/4 Digit ±4000 Counts	µP	+9V	0.09/0.2	Digital Multimeters 50ms Conversion	External
MAX134	3 3/4 Digit ±4000 Counts	µP	±5V	0.09/0.2	µP Interface 50ms Conversion	External
MAX136	3 1/2 Digit ±2000 Counts	LCD	+9V	0.06/0.15	Hold Function, Low Power	Bandgap
MAX138	3 1/2 Digit ±2000 Counts	LCD	+2.25V to 7V	0.2/0.8	2 cell Battery Operation ± Signal Input	Low T.C. Bandgap
MAX139	3 1/2 Digit ±2000 Counts	LED	+5V	0.2/0.8	± Inputs (from GND)	Low T.C. Bandgap
MAX140	3 1/2 Digit ±2000 Counts	LED	+5V	0.2/0.8	2-3mA seg I	Low T.C. Bandgap
ICL7106	3 1/2 Digit ±2000 Counts	LCD Drive	+9V	0.6/1.8	Digital Multimeters	Zener
ICL7107	3 1/2 Digit ±2000 Counts	LED Drive	+9V	0.6/1.8	Digital Panel Meters	Zener
ICL7109	12 Bits + Sign ±4096 Counts	8/16 bit µP an UART	±5V	0.7/1.5	3-State Binary Up to 30 Conversion/Sec	Zener
ICL7116	3 1/2 Digit ±2000 Counts	LCD	+9V	0.8/1.8	Same as ICL7106, but Adds Hold Function	Zener
ICL7117	3 1/2 Digit ±2000 Counts	LED	±5V	0.8/1.8	Same as ICL7107, but Adds Hold Function	Zener
ICL7126	3 1/2 Digit ±2000 Counts	LCD	+9V	0.6/0.1	Use ICL7136 for New Designs	Zener
MAX7129 ICL7129A	4 1/2 Digit ±20,000 Counts	Triplexed LCD	+9V	1.0/1.4	DPMs, Instruments Lowest Noise A/D– 3µV (7129A)	External
ICL7135	4 1/2 Digit ±20,000 Counts	Multiplexed BCD	±5V	1.0/2.0	DMM, DPM, Data Loggers	External
ICL7136	3 1/2 Digit ±2000 Counts	LCD	+9V	0.06/0.1	Low Power Version of ICL7106, Very Low Noise	Zener
ICL7137	3 1/2 Digit ±2000 Counts	LED	±5V	0.06/0.2	Low Power when LED Display Turned Off	Zener

SAR and Half-Flash A/D Converters

Part Number	Resolution	Integral Linearity	Conversion Time	Supply Voltage	Input Range	Features	Reference
MAX150	8 bits	1/2 LSB	1.34µs	+5V	+5V	Track/Hold	Internal
MAX154	8 bits/4 ch	1/2 LSB	2µs	+5V	+5V	Track/Hold	Internal
MAX158	8 bits/8 ch	1/2 LSB	2µs	+5V	+5V	Track/Hold	Internal
MAX160	8 bits	1/2 LSB	4µs	+5V	±15V	Fast AD7574	
MAX161	8 bits/8 ch	1/2 LSB	20µs	+5V	±15V	Fast AD7581 Dual port RAM	External
MAX162	12 bits	1/2 LSB	3.25µs	+5V/-12V	+5V	Fast AD7572	Internal
MAX163	12 bits	1/2 LSB	7µs	+5V/-12V	+5V	Sample/Hold	Internal
MAX164	12 bits	1/2 LSB	7µs	+5V/-12V	±5V	Sample/Hold	Internal
MAX167	12 bits	1/2 LSB	7µs	+5V/-12V	±2.5V	Sample/Hold	Internal
MAX172	12 bits	1/2 LSB	10µs	+5V/-12V	+5V	Low cost	Internal
AD578	12 bits	1/2 LSB	3µs	±15V	±10V	Parallel/Serial Output	Onboard
ADC0820	8 bits	1/2 LSB	1.4µs	+5V	+5V	Half-Flash, Internal Clock Track/Hold	
AD7572	12 bits	1/2 LSB	5µs	+5V/-15V	+5V		Internal
AD7574	8 bits	1/2 LSB	15µs	+5V	±15V	Analog V_{IN} > V_{SUPP}	External
AD7581	8 bits	1/2 LSB	66.6µs	+5V	±15V	8 byte RAM	External
AD7820	8 bits	1/2 LSB	1.34µs	+5V	+5V	Half-Flash	External
AD7824	8 bits/4 ch	1/2 LSB	2.0µs	+5V	+5V	4 Channels	External
AD7828	8 bits/8 ch	1/2 LSB	2.0µs	+5V	+5V	8 Channels	External

MAXIM'S A-TO-D CONVERTER FAMILY

	Half Flash	Successive Approximation	Integrating
Speed	1-2μs	3-20μs	>20ms
Resolution (bits)	8	8-12	11-16
Data Acquisition	Yes	Yes	Yes
Display Driver	No	No	Yes
Input Mux	Yes	Yes	Yes

MAXIM

Maxim's A-to-D family includes a variety of converter types and feature combinations. Devices for "data acquisition" have digital outputs suitable for interfacing to microprocessors or other digital subsystems. Those with display drivers can be directly connected to LED or LCD displays, but are generally unsuitable for connection to other digital circuits.

NEW INTEGRATING ADCs THE MAX130/131

- Pin Compatible Upgrade for ICL7106 and ICL7136 At the Same Price
- High Stability Bandgap Reference
- 100μA Maximum Supply Current (MAX131)
- 4.5V to 14V Supply Voltage Range
- Onboard 3 1/2 Digit LCD Driver
- Industrial Temperature Grades and Surface Mount Packages

MAXIM

The MAX130 and MAX131 provide much improved reference stability and lower supply current than the ICL7106 and ICL7136, and for the same cost. The A/Ds' on-chip bandgap reference has a guaranteed tempco of 50ppm/°C for the "A" versions and 100ppm/°C for standard devices. The MAX130 is a drop-in replacement for the ICL7106 but draws only 250μA (compared to 1.8mA for the ICL7106). The MAX131 replaces the ICL7136, but uses only 100μA.

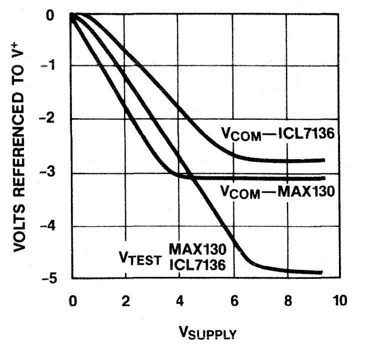
NEW INTEGRATING ADCs THE MAX138/139/140

- Operates with +2.5V to +7.0V Supply
- Measure Positive and Negative Input Voltages
- On-Chip Charge Pump Generates V^- Supply
- Stable Bandgap Reference
- Drive LCDs (MAX138) or LEDs (MAX139/140)
- MAX140 for Low Current LEDs

MAXIM

The MAX138/139/140 contain on-chip charge pumps which create a negative supply using two external capacitors. This allows the A/D to measure negative input voltages, and inputs that swing right to 0V, while being powered from a single positive source. The MAX138 drives LCD displays, the MAX139 drives LED displays, and the MAX140 drives low current LED displays. Pinouts are very close to those of the ICL7136 and ICL7137.

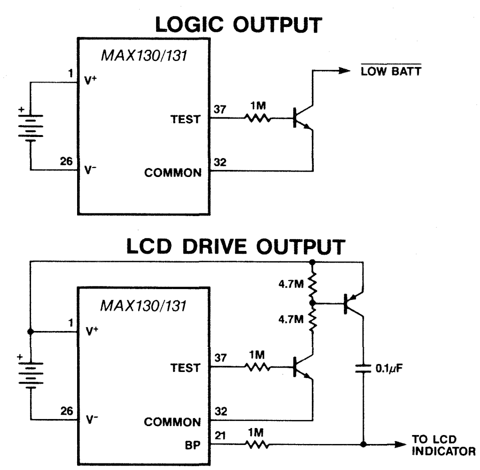
MAX130 AND ICL7136 THE BANDGAP ADVANTAGE



Maxim's bandgap advantage is clearly illustrated by comparing the internal voltage reference (the voltage between V^+ and COMMON) of the MAX130 with that of the ICL7136. The ICL7136's zener reference, besides having higher drift, loses regulation when the supply voltage falls to about 6.5 volts. This limits the ICL7136's useable supply range to above 7 volts.

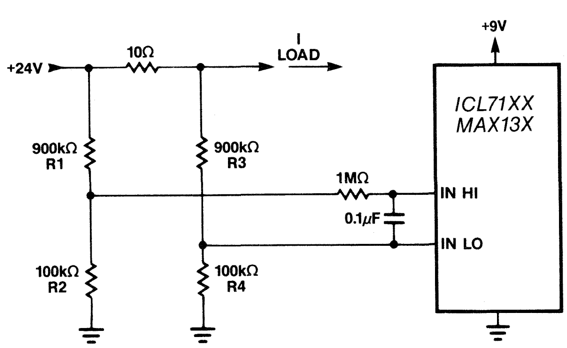
The MAX130/131 Bandgap reference is stable with supply voltages below 4 volts, extending effective battery life, and allowing reliable operation with +5 volt logic supplies. In addition, the MAX130/131 features superior long-term stability and uses less power than the ICL7136.

MAX130/131 LOW BATTERY DETECTORS



A simple low battery indicator can be added to the MAX130 or MAX131 using a single transistor for a logic output, or two transistors to drive an annunciator on an LCD. The voltage cross-over between the COMMON and TEST pins turns the transistor on when the supply voltage drops to approximately 4.0 volts.

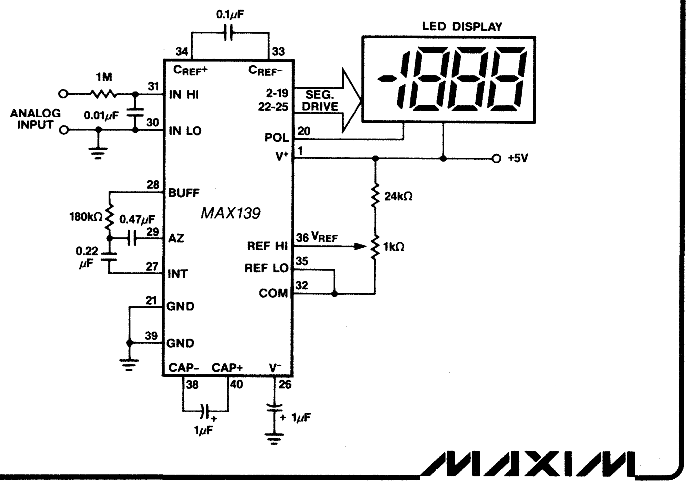
EXTENDING COMMON MODE RANGE



The input range of most integrating A/Ds is limited to approximately 1.5V below V^+ and 1.5V above V^-. However, since the A/D's input impedance is high, differential signals with common mode voltages outside the chip's power supplies can be easily measured. This is done by connecting a resistive voltage divider to each input pin.

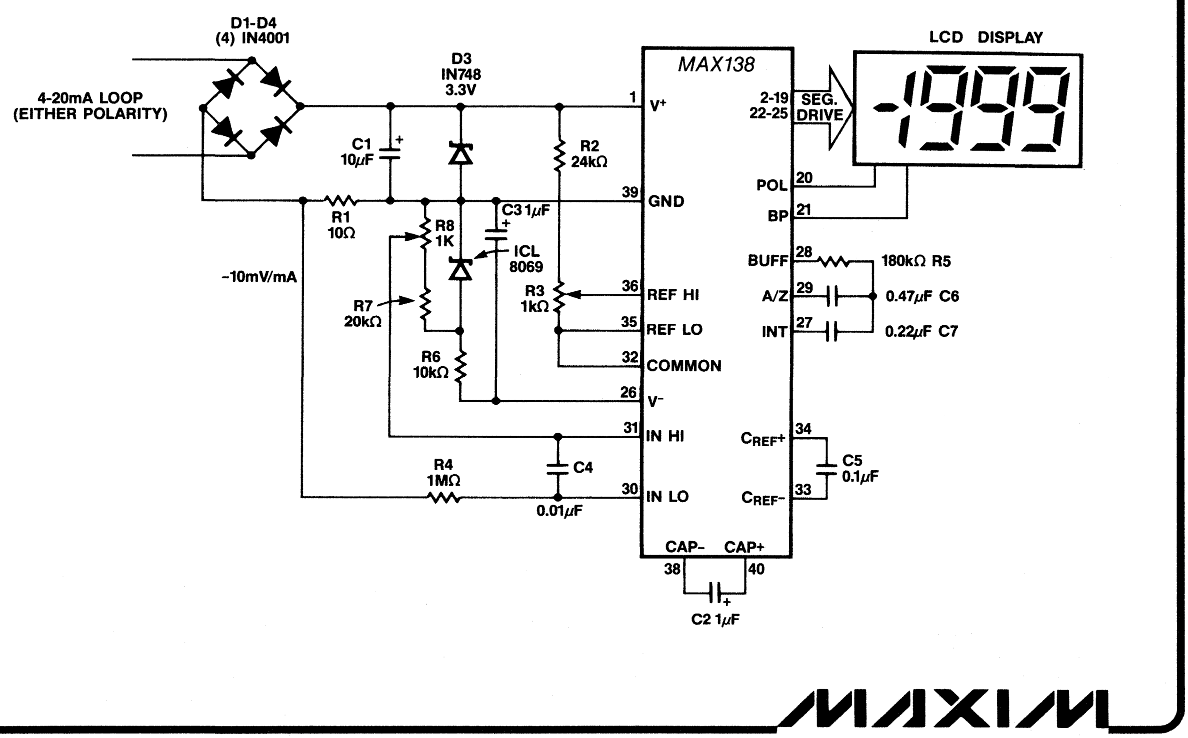
In this example, the current from a +24V supply, flowing to a grounded load, is measured by sensing the differential signal on a current sense resistor at the 24V source. This signal is divided by identical voltage dividers at the A/D's HI and LO inputs. The resulting measurement may be scaled by adjusting either the sense resistor or the A/D's reference to correct for losses from voltage divider attenuation.

+5 VOLT POWERED DVM
±3.5V INPUT RANGE



The MAX138, MAX139 and MAX140 allow DVM functions to be easily included in single +5V powered boards. Either analog input may operate above or below ground, facilitating direct measurements of ground referenced signals (or current shunts) in single supply systems. These parts accomplish this by generating their own V⁻ supply with an internal charge pump. Up to 1mA may be drawn from the chip's V⁻ output (approximately -4.8V) to power other circuitry. Here, a +5 volt logic supply is used to power a MAX139 and LED display. The same approach can be used with LCDs and low current LEDs by using the MAX138 and MAX140, respectively.

4-20mA LOOP POWERED INDICATOR



The unique features of the MAX138 allow it to be used in low voltage applications that were previously not feasible for display driving A/Ds. In this example, a 0-100.0% digital indicator (LCD) measures, and is powered from, an industrial 4-20mA analog current loop. The circuit consumes only 4.5V when inserted in the line. A full-wave bridge at the input allows either input polarity to be measured, but if the input polarity is known, the full-wave bridge can be omitted, reducing the circuit's operating voltage.

A voltage proportional to the loop current is developed across a 10Ω sense resistor, R1. A 3.3V zener clamps the MAX138's supply as the current through the loop changes. IN HI (pin 31) is offset negatively and trimmed by R8 so that the display reads "000.0%" with 4mA loop current. R6 is set for an indication of "100.0 %", with a 20 mA loop current. R8 and R6 can also be adjusted so the display reads directly in milliamps.

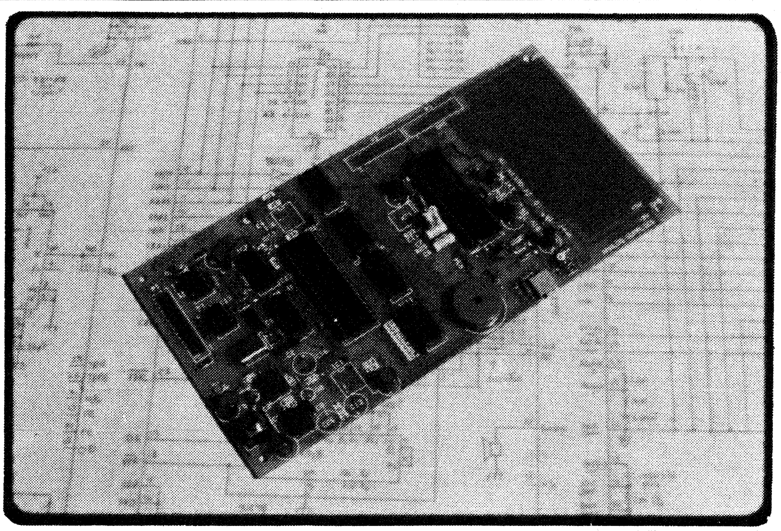
INTEGRATING A/Ds FOR µP-BASED APPLICATIONS

	MAX133/134	ICL7109
Resolution	± 40,000 Counts (BCD)	12 bits + Sign
Conversions/Sec	20	15-30
Input Channels	7 Single Ended	1 Differential
Sensitivity	10µV	100µV
Supply I (Typ)	100µV	700µV
Sleep & Beep	Yes	No

In many measurement applications, the A/D that connects to a microprocessor needs resolution and noise rejection more than speed. In those cases integrating A/Ds are usually preferred....if they can talk to the processor.

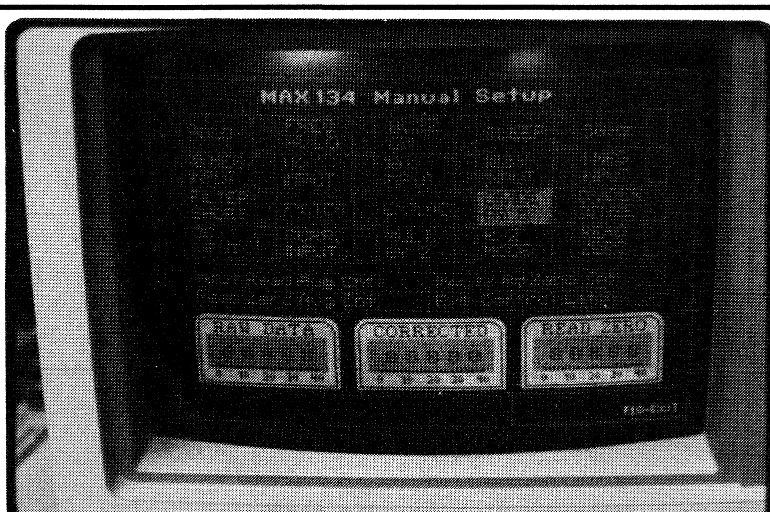
The MAX133 and MAX134 are high resolution integrating A/Ds for data acquisition and instrumentation applications. When controlled by a microprocessor, either A/D can perform auto-ranging measurements from ±400mV to ±4000V full scale. The MAX133 has a 4 bit multiplexed address/data bus. The MAX134 uses 3 address lines and a 4-bit bi-directional data bus.

Maxim's ICL7109 provides low noise 12-bits-plus-sign A/D conversions at up to 30 conversions/second and is designed for easy interface to UARTs and microprocessors.

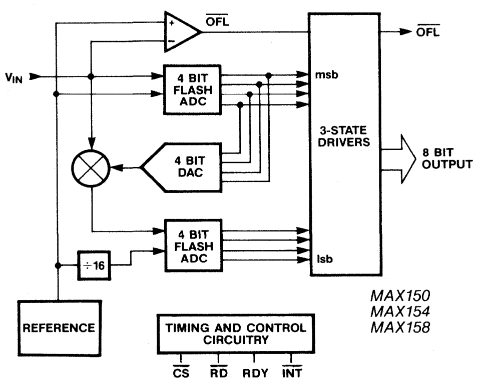


MAX134 DEMO BOARD AND SOFTWARE

A demo circuit board is available for the MAX134 which connects to the RS-232 port of an IBM PC or compatible computer. The interface is opto-isolated so that low level signals can be measured without interference from the computer. Software is also provided to exercise all MAX134 features including input channel selection (or auto-ranging), auto-zeroing, and switched filtering to list a few.



HALF-FLASH A TO D CONVERSION

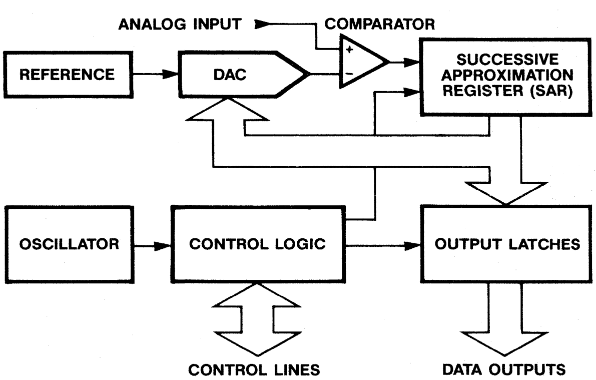


An A/D conversion technique which combines some of the speed advantages of flash conversion with the circuitry savings of successive approximation is termed "half-flash". In an 8-bit half-flash converter, two 4-bit flash A/Ds sections are combined. The upper flash A/D compares the input signal to the reference and generates the upper 4 data bits. This data goes to an internal DAC whose output is subtracted from the analog input so that the difference can be measured by the second flash A/D which provides the lower 4 data bits.

MAXIM'S HALF FLASH ADCs
8 bits; 2.5V Reference On-Board

Part #	Conversion Time	MUX	Replaces
MAX150	1.34µs		AD7820, ADC0820
MAX154	2.0µs	4 ch	AD7824
MAX158	2.0µs	8 ch	AD7828

A-TO-D CONVERSION BY SUCCESSIVE APPROXIMATION



The most frequently encountered type of A-to-D conversion in data acquisition applications is successive approximation (SAR = successive approximation register). SAR converters operate by stepping through a sequence of trial and error comparisons between the input signal and a series of binary weighted reference levels. Each successive comparison provides one output data bit. The result of each comparison also narrows the range in which the input signal is known to reside.

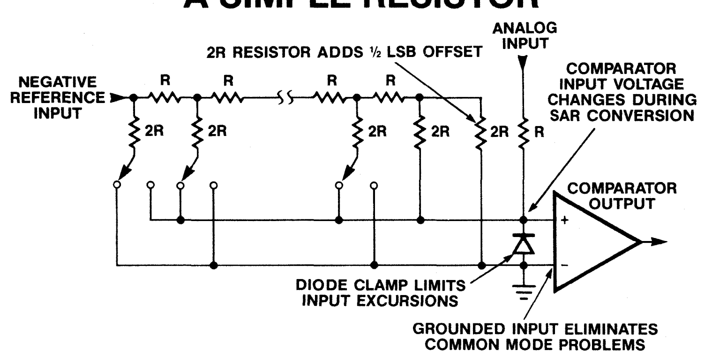
MAXIM'S SUCCESSIVE APPROXIMATION ADCs

Part #	Resolution	Speed
MAX160	8 bits	4μs
MAX161	8 bits	20μs (8 ch)
MAX162	12 bits	3.25μs
MAX172	12 bits	10μs

MAXIM'S SECOND SOURCES ARE FASTER

Orig PN	Speed (μs)	Maxim PN	Speed (μs)	Features
AD7572-5	5	MAX162	3.25	-12V or -15V V^-
AD7572-12	12	MAX172	10	Lower Noise, $10 (1K)
AD7574	15	MAX160	4	Input Range > Supply
AD7581	66.7	MAX161	20	8 Byte RAM

SAR ADC INPUT IS *NOT* A SIMPLE RESISTOR



The analog input to a SAR A-to-D may look like a fixed resistance but actually behaves very differently. The comparator sees the summed currents from the DAC output and the analog input at its "+" input. The voltage at the comparator input varies as the A/D sequences though its SAR routine. This forces the A/D input current to also change as the conversion progresses. The comparator input is diode clamped to minimize these changes and to optimize conversion speed. Nevertheless, the A/D's performance may still be impaired if the analog input signal arrives from too high a source impedance, or from a source whose output does not quickly settle when driving the varying load presented by the A/D input.

FIGURE 1. 1MHz CLOCK

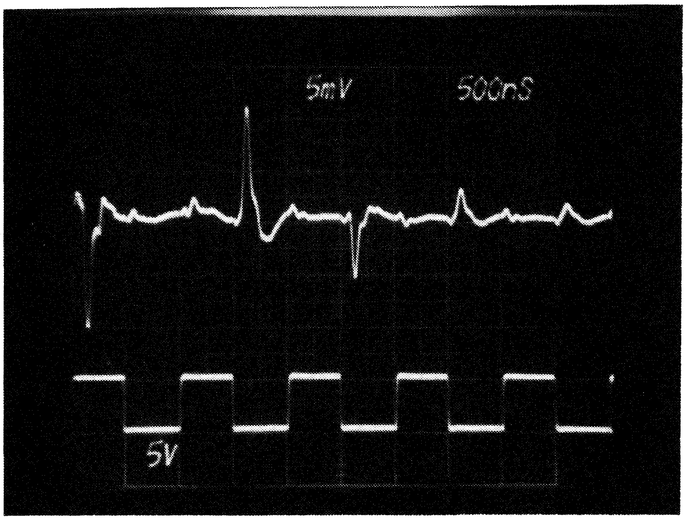


FIGURE 2. 2 MHz CLOCK

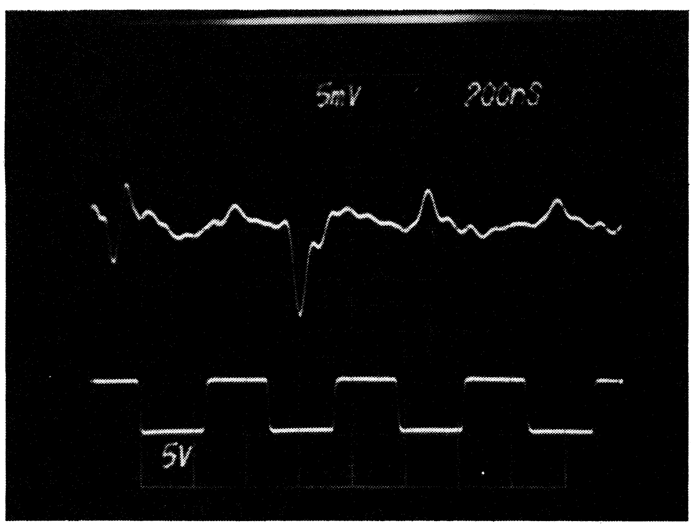
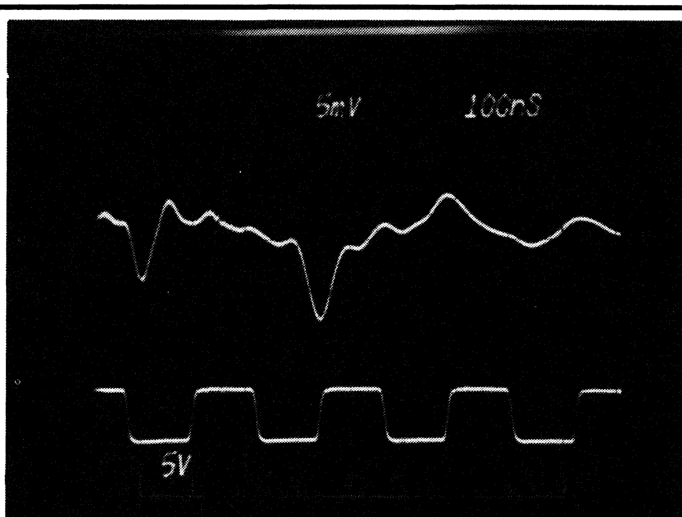


FIGURE 3. 4MHz CLOCK



DRIVING A SAR A/D INPUT

Figure 1 illustrates the disturbance at the input pin of a MAX162 12-bit A/D that is clocked at 2MHz. The input is being driven by an OP-27 amplifier. The amplitude of the pulse disturbances at the input generally decrease as the conversion takes place. This is because the reference DAC's output transitions become smaller as the A/D zeroes in on the answer.

If the driving device (op-amp or sample-hold) does not accommodate the load changes presented by the A/D, the voltage at the analog input terminal deviates from the desired value for a short time. These "glitches" are synchronous with the ADC clock and may be 10mV or more in amplitude. Even though 10mV is well above 1 LSB in amplitude, the pulses do not degrade the A/D's performance **IF** they settle before the comparator's output is sampled at the next falling clock edge.

If the open loop output impedance of the driving amplifier is too high (enabling the glitches to be generated), or if its settling time is too long (enabling the glitches to persist for more than a clock cycle), some errors may be generated even with slow moving input signals. It is interesting to note in the photos that the MAX162 allows 1 1/2 clock cycles for the first glitch to settle, but only 1 clock cycle for each subsequent pulse. This is intentional. Since potentially the largest disturbance occurs at the first bit test in the conversion, the A/D allows more settling time.

Operating symptoms of an inadequate op-amp or sample-hold are missing codes when the analog input signal is DC and free of noise. A good test is to slow down the A/D's clock, giving the driving device more time to settle. Figures 2 and 3 show this test in practice with an OP27 driving the A/D input. With a 4MHz clock (Figure 3, 4µs between conversions) the OP27 does not settle within one clock period, and conversion errors are generated. With a 2MHz clock however (Figure 2), the OP27 settles adequately and 1LSB performance is achieved.

If the desired conversion rate is below the A/D's limit, amplifiers with less than optimum output specifications may be used. The A/D clock of course has to be slowed. The approximate speed at which various op amps can perform adequately with the MAX162 is listed on page 41. There is some variation between different sources of the same device type.

AMPLIFIER SELECTION

- Bandwidth > 2MHz
- Slew Rate > 2V/μs
- Fast Settling to 0.01%
- Check DC Accuracy Specs
- Ensure the Supplies Are Clean
- Test Using 20% Faster Clock

MAXIM

Things to watch for when picking an op-amp to drive the MAX162 or AD7572 are:

- Beware of thermal tails on output settling, a frequent characteristic of "high slew rate" op-amps with undefined settling specs.
- Amps with 0.01% settling specs usually do substantially better than those with only 0.1% numbers.
- Input offset voltage and offset drift.
- High frequency PSRR - This affects settling time but unfortunately is not specified by many manufacturers.
- Some general purpose amps amplifiers will work reasonably well as input buffers at reduced A/D clock rates:

CA3140, OP27 –	+1/2 LSB DNL at 0.6MHz
	1 LSB DNL at 0.7MHz
LT1055, LF411 –	+1/2 LSB DNL at 0.6MHz
	1 LSB DNL at 0.7MHz
	2 LSB DNL at 2MHz

The amplifier-A/D combination should be checked with a faster A/D clock than the design requires.

A SAR A/D's input is typically driven by an op-amp or sample-hold, however, care must be taken with high speed A/D's like the MAX162, because it's tempting to assume that high frequency input signals can be converted accurately with no sample-hold. A 12-bit converter with a 5V full scale range resolves input changes of nearly 1mV, and signals as slow as 1kHz change by this amount in well under the 3.25μs conversion time of the MAX162. In fact for full-scale signals above only 11Hz, some sort of hold function has to be added if the conversion is be accurate to 12 bits.

The acceptance criteria for sample-holds suitable for use with the MAX162 are:

- Aperture time fast enough to capture the highest input frequency - 30ns will preserve the MAX162's performance,
- Acquisition time of about 500ns,
- Hold step error (switching glitch) of less than 1/2LSB (i.e., less than 600μV),
- Hold droop less than 1/2LSB in the cycle time of the ADC (often about 4μs),
- Input offset voltage <600μV over temperature range.

If an off-the-shelf sample-hold is required, the HA-5330 is a good match for the performance of the MAX162. Its offset error, acquisition characteristics, and output stability are all sufficient to avoid degrading the accuracy of the MAX162.

MAX162: THE INDUSTRY LEADER

- 12 bit, SAR A to D Converter
- 3.25µs or 250K Conv/s with 4MHz Clock
- +5, and -12 or -15V Supplies
- On-board Reference & Clock Osc
- Monotonic from -55 to +125°C
- 8 or 12 bit 3-State Interface

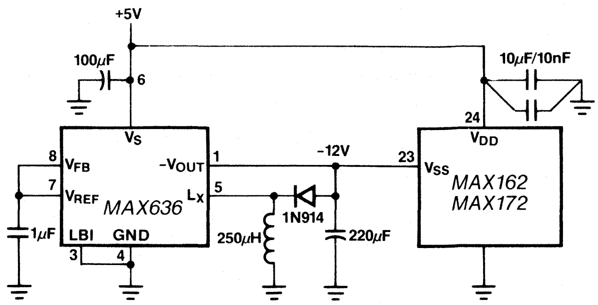
MAXIM

MAX162: THE INDUSTRY LEADER (con't.)

- AD7572 Architecture
- Fastest Second-Sourced 12 bit A/D
- Lower Noise
- Faster Conversion Time
- -12V Operation
- Superior High Frequency PSRR

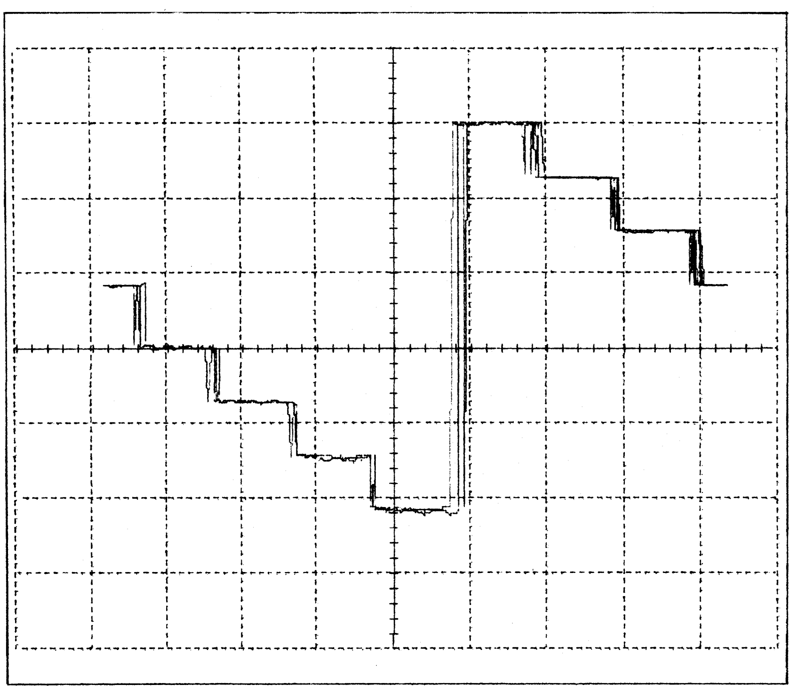
MAXIM

SINGLE +5V OPERATION



ONLY POWER CONNECTIONS SHOWN FOR CLARITY

Because the MAX162 and MAX172 use only 135mW of power, they can easily be powered from a single +5V supply with the help of this low power DC-DC conversion circuit. A MAX636 generates a -12V (or a -15V) supply for the converter's V^- input. A potential problem when using switching power supply circuitry with high speed A/Ds is increased noise but as the accompanying plot shows, the A/Ds's conversion noise is still quite low with this circuit.



500kHz CONVERSION RATE

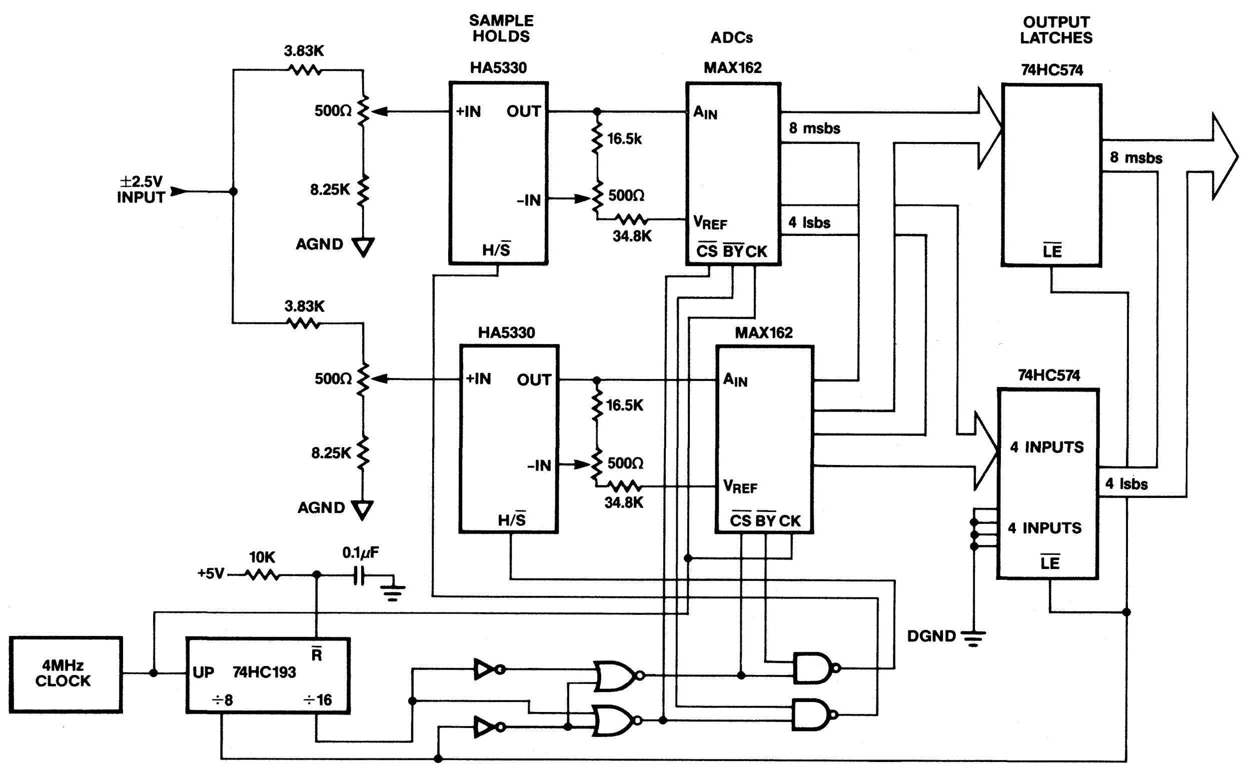
Advantages:

- 12 bits in 2µs
- Pipeline Architecture
- Low Cost Components

Disadvantages

- Needs External Components
- Care Required for Low Noise
- Calibration is Required

500kHz A/D CONVERSION WITH TWO MAX162s

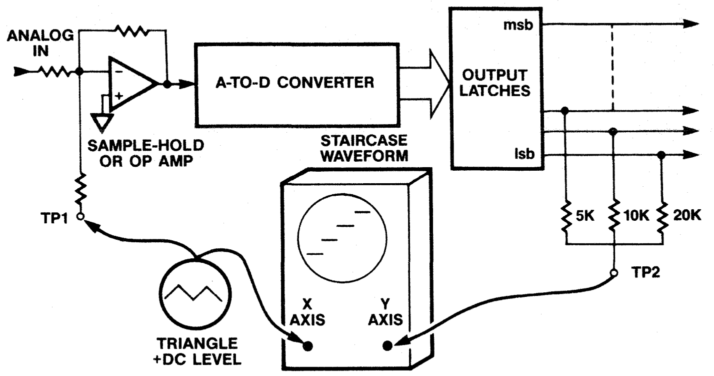


DECOUPLING, POWER, GND AND GROUNDED CONTROL LINES NOT SHOWN

Two MAX162's build a low cost 12-bit A/D that does a conversion every 2µs. The A/Ds connect to one input signal and send their outputs to one 3-state bus. With a 4MHz clock, each MAX162 converts every 4µs. The control logic starts conversions first in one MAX162 and then 2µs later in the other, and synchronizes the sample-hold with the A/D. HOLD mode is started by the control logic, but is sustained by the MAX162's BUSY output. The A/Ds output is then latched into the 74HC574s.

The circuit's throughput rate is 500,000 conversions per second (a new conversion every 2µs), but the output is updated 4µs after a conversion is begun, because of the pipeline architecture. The circuit provides a low cost means of increasing converter speed, without sacrificing other specifications since neither the A/Ds nor the sample-holds need to be capable of 2µs performance. The major part costs are: MAX162ACNG $33.15, MAX162CCNG $25.00, and HA1-5330-5 $15.65.

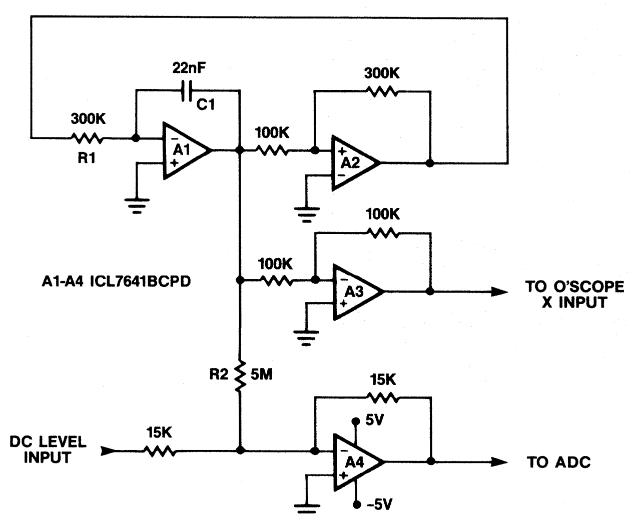
TESTING NOISE AND DNL ERRORS



The noise and linearity of high performance A/Ds can be tested without expensive measurement instruments using the circuit shown. A low amplitude triangle "dither" waveform is summed with a variable DC level and inputted to the A/D. The 3 LSBs of the converter output are summed together through 3 weighted resistors to make a 3-bit "DAC". The input signal and DAC output are connected to an X/Y oscilloscope so that a properly functioning A/D generates a staircase waveform. Eight bits of the A/Ds range can be viewed at one time and any part of the converter's input span can be checked by varying the DC level at the input. Output staircase waveforms from linear, noisy, and nonlinear 12-bit A/Ds a shown on page 46.

Since most of what is used in this test circuit is already included on a typical data acquisition board, a DNL "test point" could be added for in-circuit A/D testing by including a separate input resistor (at TP1) and the 3 output resistors (at TP2) on-board.

TESTING ADCs: TRIANGLE GENERATOR

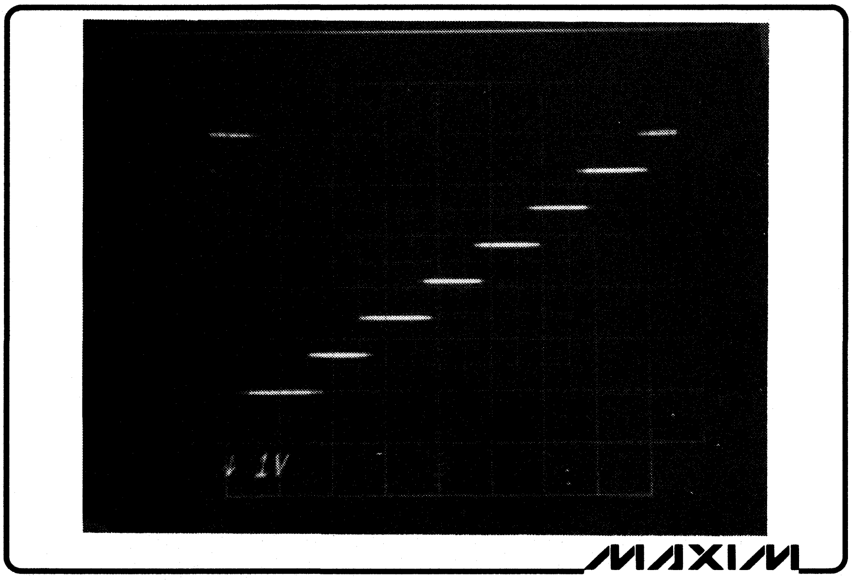


This circuit generates a symmetrical 10mV p-p triangle waveform which is summed with a DC level and connected to the A/D analog input for noise/DNL testing. The DC LEVEL input offsets the triangle waveform over the input range of the ADC. The 10mV amplitude amounts to an 8 LSB span for a 12 bit, 5V full-scale ADC.

The frequency of the sawtooth is critical, and depends on the speed of the converter to be tested. The frequency should be below the point at which the input moves 1/2 LSB during a conversion (if there is no input sample-hold). Generally 10Hz to 1kHz is used. The R1xC1 product determines the sawtooth frequency. The oscilloscope "X" output is the same frequency, but is 3.3V p-p and is not affected by the DC LEVEL input.

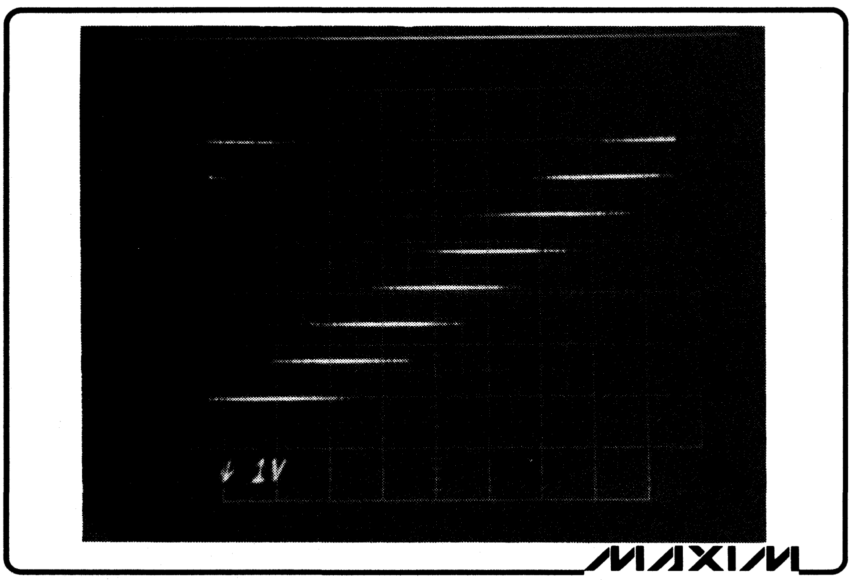
Component values are not critical, and 5% values should work well. The ICL7641BCPD quad op-amp was used in our tests, but most other other quad op-amps will not alter the performance.

Take care to avoid noise on the sawtooth input to the ADC. Use short shielded cables, decouple the op-amp power supplies, and use single point grounding. Such precautions are needed because each 610µV of injected noise represents 1/2 LSB.



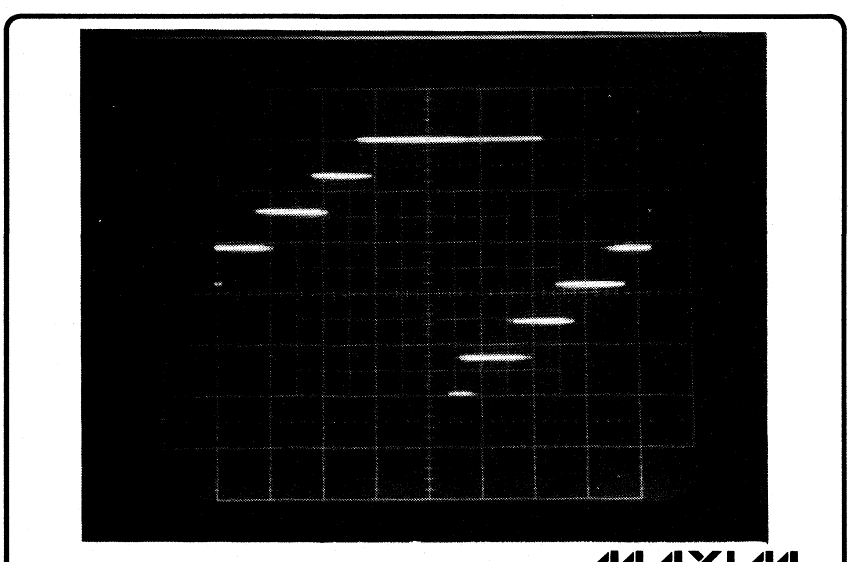
LINEAR AND NOISE FREE A/D

Note the very small amount of overlap from one step to the next indicating less than 1/4 LSB of noise. Also, the uniform step size indicates low DNL error.



NOISY A/D

Here the step sizes are still relatively uniform but there is severe overlap and "bleeding" from one step to the next. With this particular converter, as many as 4 or 5 different conversion results are generated for one DC input.



DIFFERENTIAL LINEARITY ERROR

This A/D exhibits a "wide code" (top of waveform) at mid-scale where the A/D gives one answer for a several LSB span of input signal. This is followed by a very narrow (bottom of waveform) code which is almost missing entirely. The trace also shows that on some conversions the output stays on the same code for as much as 3 LSBs.

THE MAX172: THE LOWEST COST SOLUTION (12 bit – 10μs)

• AD7572JN12	\$35	May 88
• MAX172BCNG	\$12	Now

- 1/3 the Noise
- -12V Operation
- Faster Conversion
- Available in SOIC

MAXIM

COMING SOON

12-BIT ADCs WITH SAMPLE-HOLDS

- 7μs Conversion Time – Including S-H Acquisition
- Pin Compatible with MAX162, MAX172, AD7572
- MAX163 . . . . 0V to +5V Input Range
- MAX164 . . . . -5V to +5V Input Range
- MAX167 . . . . -2.5V to +2.5V Input Range

MAXIM

DACs FROM MAXIM

- Singles, Duals, Quads, Octals
- Voltage or Current Outputs
- 8 bit/10 bit/12 bit Resolution
- Serial and Parallel Interfaces
- Single and Double Buffered Interfaces
- CMOS and Bipolar Technologies

MAXIM

D/A Converters

Part Number	Type	Resolution	Settling Time (μs)	Power (V)
MAX500*	Quad, Voltage Out, Serial In	8-bit	5.0	+12V to +15V
MAX543*	Serial Input, 8-pin pkg	12-bit	1.0	+5V
MAX7624	Improved AD7524	8-bit	0.25	+5V to +15V
MAX7645	Improved AD7545	8-bit	0.25	+5V to +15V
AD565A	Bipolar, Reference	12-bit	0.25	±15V
AD566	Bipolar	12-bit	0.35	±15V
AD7224	Voltage Out	8-bit	7.0	+12V to +15V
AD7225	Quad, Voltage Out	8-bit	5.0	+12V to +15V
AD7226	Quad, Voltage Out	8-bit	5.0	+12V to +15V
AD7228	Octal, Voltage Out	8-bit	5.0	+12V to +15V
AD7520	Multiplying	10-bit	0.5	+5V to +15V
AD7521	Multiplying	12-bit	0.5	+5V to +15V
AD7523	Multiplying, Low Cost	8-bit	0.2	+5V to +16V
AD7524	Multiplying	8-bit	0.25	+5V to +15V
AD7528	Dual, Multiplying	8-bit	0.18	+5V to +15V
AD7530	Multiplying	10-bit	0.5	+5V to +15V
AD7531	Multiplying	12-bit	0.5	+5V to +15V
AD7533	Multiplying, Low Cost	10-bit	0.6	+5V to +16V
AD7534*	Multiplying, μP Compatible	14-bit	1.5	+12V to +15V
AD7535*	Multiplying, μP Compatible	14-bit	1.5	+12V to +15V
AD7536*	4-Quadrant, Multiplying	14-bit	1.5	+12V to +15V
AD7537*	Dual, Multiplying	12-bit	1.5	+11V to +16V
AD7538*	Multiplying, μP Compatible	14-bit	1.5	+12V to +15V
AD7541	Multiplying	12-bit	1.0	+5V to +16V
AD7541A	Multiplying	12-bit	1.0	+5V to +16V
AD7542	Multiplying, 4-Bit Load	12-bit	2.0	+5V
AD7543	Multiplying, Serial Input	12-bit	2.0	+5V
AD7545	Multiplying, Latched Input	12-bit	2.0	+5V to +15V
AD7545A	Multiplying, Latched Input	12-bit	1.0	+5V to +15V
AD7547*	Dual, Parallel Load	12-bit	1.5	+11V to +16V
AD7548*	Multiplying, 8-Bit Bus	12-bit	1.0	+5V to +15V
AD7549*	Dual, Double Buffered	12-bit	1.5	+15V
AD7628	Dual, Double Buffered	8-bit	0.35	+11V to +15V
DAC8212*	Dual, Multiplying	12-bit	1.0	+5V or +15V

* To be introduced.

APPLICATION HINTS

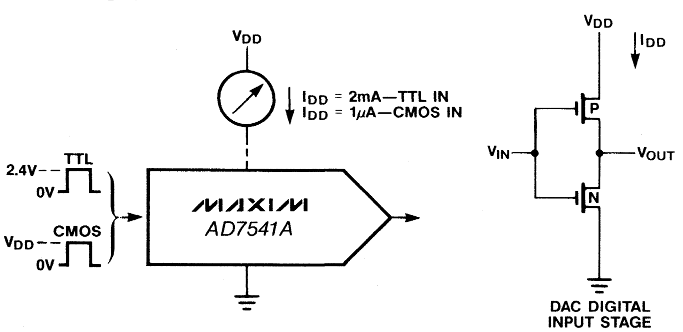
For Minimizing Noise . . .

- Use External Latches
- Use a Low Noise Reference
- Use a Low Noise Amplifier
- Ensure Supplies are Clean
- Pay Attention to Grounding

To minimize noise in CMOS D-to-A converters

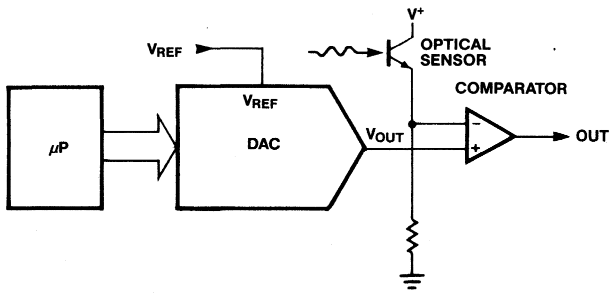
1) External latches eliminate noise feedthrough from the data bus.
2) The reference and the output amplifier are the major sources of wideband DAC noise since most CMOS DACs are essentially passive in the signal path.
3) Power supply rejection ratio (PSRR) decreases at high frequencies. Bypassing power supplies with electrolytic capacitors in parallel with ceramic caps *as close to the DAC as possible*. 10µF and 0.1µF are common values.
4) Pay attention to grounding. Ground current from digital hardware should not be allowed to inject noise into the DAC's ground connection. Use either a low impedance ground path, ground plane, or single point ground.

REDUCING DAC SUPPLY CURRENT



The operating current of most CMOS DACs depends on the logic level of the digital input signals. This is because the input inverter/level shifters in the DAC account for most of the power consumption. If the digital inputs are made to swing from V_{DD} to ground (i.e. CMOS logic levels) rather than to TTL levels, the supply current is typically cut by 100X or more. When the DAC output is changing, the supply must still supply 5 to 10mA for about 20ns, independent of input logic level, but this is usually a small fraction of the total operating time.

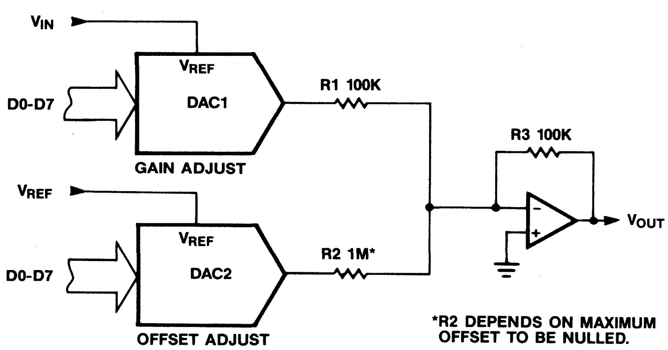
DIGITAL CALIBRATION AND THRESHOLD SELECTION



The circuit is used to discriminate "light" and "dark" conditions in photo-sensing applications where the light levels are better described as "bright" or "dim". Applications include tachometers, motion sensing, automatic readers, and liquid clarity examination.

A calibration "bright" light is applied while the DAC is digitally ramped. The count of the ramp at the trip point is then the high calibration point. Similarly, a "dim" calibration light is applied and a low calibration point is determined. The trip threshold is placed between the two calibration points, and the μP sets this value up on the DAC inputs. Continuously varying light levels are judged "light" or "dark" according to whether or not they exceed this level.

DIGITAL CONTROL OF GAIN AND OFFSET



Two voltage-output DACs (or a dual DAC) digitally control the offset and gain of this circuit. This is especially useful when curve-fitting nonlinear functions for transducer linearization or in analog compression/expansion applications. The input signal enters through the reference input of the Gain Adjust DAC whose output is summed with that of the Offset Adjust DAC. The Offset DAC's reference input is a fixed DC level. The relative weight of each DAC output is adjusted by R1 and R2.

THE MAXIM ADVANTAGE

- AD75XX – Multiplying DACs

  Lower Output Capacitance
  Lower Glitch Energy, Lower Noise

- AD7225/7226 – Quad DACs

  Higher Speeds, Faster Settling,
  Improved Performance with Single Supply,
  Small Outline & PLCC Packages

- All D to A Converters

  Low FIT Rate

MAXIM

NEW MAXIM DAC PRODUCTS

- MAX500 – Quad, 8 bit, Serial Input Voltage Out
- MAX543 – 8 pin Package, 12 bit, Serial Input
- MAX7624 – TTL Compatible, 15V, 8 bit, Latched Input
- MAX7645 – TTL Compatible, 15V, 12 bit, Latched Input

MAXIM

NEW MAXIM SECOND SOURCE DACs

AD565A	AD7534	AD7545A
AD566	AD7535	AD7547
AD7224	AD7536	AD7548
AD7228	AD7537	AD7549

A Major Thrust into D to A Conversion

MAXIM

LOW COST PRECISION REFERENCES MAX672 (10V), MAX673 (5V)

- Improved Pin-For-Pin REF01/REF02 Replacements
- 5ppm/°C Max, 2ppm/°C Typ Tempco
- 15µV p-p Max Noise, Guaranteed
- Pretrimmed to 0.05% Accuracy
- 0.002%/mA Max Load Regulation

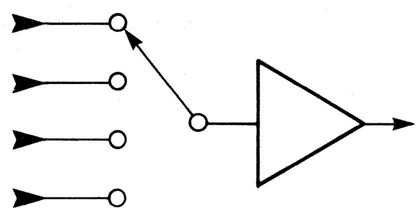
MAXIM

The MAX672 and MAX673 are enhanced pin-for-pin replacements for the popular REF01 and REF02. Improvements include tighter initial accuracy of 10V±5mV (0.005%) and guaranteed load regulation of 10ppm/mA (0.001%/mA). Temperature coefficients are guaranteed at 5ppm/°C. Output noise is also guaranteed at 15µV p-p maximum in a 0.1Hz to 10Hz bandwidth. The references are supplied in 8 pin SOICs (small outline) as well as DIP and TO-99 packages. The MAX672 and 673 are monolithic devices and at $3.50 (1000s) they combine the accuracy expected from hybrid voltage references with monolithic pricing.

Voltage References

Part	Voltage	Tempco (ppm/°C)	Initial Accuracy	Vsupply	Isupply (Typ/Max mA)	Features
MAX670	10V	3	±2.5mV	+15V	9/14	Kelvin sensing
MAX671	10V	1	±1.0mV	+15V	9/14	Kelvin sensing
MAX672	10V	5	±5.0mV	+13V to +40V	1/1.4	Precision/Low Cost
MAX673	5V	5	±2.5mV	+8V to +40V	1/1.4	Precision/Low Cost
AD580	2.5V	10	±10mV	+15V	0.75/1.0	+5V power
AD581	10V	5	±5mV	+15V	0.75/1.0	10mA Output
AD584	2.5V to 10V	5	±5mV	+15V	0.75/1.0	Programmable
AD2700	10V	3	±2.5mV	+15V	9/14	Precision
AD2710	10V	1	±1.0mV	+15V	9/14	Precision
AD2701	-10V	2	±2.5mV	-15V	9/14	Precision
ICL8069	1.23V	10 to 100	±25mV	—	50µA to 5mA	2 terminal bandgap
REF01E	10V	8.5	±30mV	+15V	1/1.4	Trim Input
REF01HP	10V	25	±50mV	+15V	1/1.4	Trim Input
REF02E	5V	8.5	±15mV	+15V	1/1.4	Temp Output
REF02HP	5V	8.5	±25mV	+15V	1/1.4	Temp Output

ANALOG SWITCHES
ANALOG MULTIPLEXERS
VIDEO SIGNAL PRODUCTS
OPERATIONAL AMPLIFIERS



MAXIM

Analog Switches

Part Number	Function	$r_{DS(ON)}$ (Ω max)	$I_{D(OFF)}$ (nA max)	$t_{(ON)}$ (μs max)	$t_{(ON)}$ (ns max)	V_{IL}/V_{IH} (V)	Supply Current (I^+/I^- mA max)	Features
MAX331	4 SPST NC	175	5	600	450	0.8/2.4	0.01/0.01	Improved
MAX332	4 SPST NO	175	5	600	450	0.8/2.4	0.01/0.01	DG201/DG202
MAX333	4 DPDT	175	5	1000	500	0.8/2.4		$\pm$5V to $\pm$18V,
MAX334	4 SPST NC	75	50	120	75	0.8/3.0		+10V to +30V
								$t_{ON} > t_{OFF}$
DG201A	4 SPST NC	175	5	600	450	0.8/2.4	0.1/0.1	Low Power
DG202	4 SPST NO	175	5	600	450	0.8/2.4	0.1/0.1	Normally Open
DG211	4 SPST NC	175	5	1000	500	0.8/2.4	0.1/0.1	No V_{LOGIC} Supply
DG212	4 SPST NO	175	5	1000	500	0.8/2.4	0.1/0.1	Normally Oen
DG300A	2 SPST NO	50	5	300	250	0.8/2.4	0.5/0.1	2.4V_{IH}, Low R_{ON}
DG301A	SPDT	50	5	300	250	0.8/2.4	0.5/0.1	2.4V_{IH}, Low R_{ON}
DG302A	2 SPST NO	50	5	300	250	0.8/2.4	0.5/0.1	2.4V_{IH}, Low R_{ON}
DG303A	2 DPDT	50	5	300	250	0.8/2.4	0.5/0.1	2.4V_{IH}, Low R_{ON}
DG304A	2 SPST NO	50	5	250	150	3.5/11	0.5/0.1	CMOS Logic
DG305A	SPDT	50	5	250	150	3.5/11	0.5/0.1	levels, high
DG306A	2 DPST NO	50	5	250	150	3.5/11	0.5/0.1	speed, Low R_{ON}
DG307A	2 DPDT	50	5	250	150	3.5/11	0.5/0.1	
DG381A	2 SPST NC	50	5	300	250	0.8/4.0	0.5/0.1	Low R_{ON}
DG384A	2 DPDT NO	50	5	300	250	0.8/4.0	0.5/0.1	Low R_{ON}
DG387A	SPDT	50	5	300	250	0.8/4.0	0.5/0.1	Low R_{ON}
DG390A	2 SPDT	50	5	300	250	0.8/4.0	0.5/0.1	Low R_{ON}
IH5040	SPST NO	80	5	400	200	0.8/2.4	0.01/0.01	Very Low Power
IH5041	2 SPST NO	80	5	400	200	0.8/2.4	0.01/0.01	Very Low Power
IH5042	SPDT	80	5	400	200	0.8/2.4	0.01/0.01	Very Low Power
IH5043	2 SPDT	80	5	400	200	0.8/2.4	0.01/0.01	Very Low Power
IH5044	DPST NO	80	5	400	200	0.8/2.4	0.01/0.01	Very Low Power
IH5045	2 DPST NO	80	5	400	200	0.8/2.4	0.01/0.01	Very Low Power
IH5048	SPDT NO	45	5	1000	500	0.8/2.4	0.01/0.01	Low Charge Injection
IH5049	2 SPST NO	45	5	1000	500	0.8/2.4	0.01/0.01	Low Charge Injection
IH5050	SPDT NO	45	5	1000	500	0.8/2.4	0.01/0.01	Low Charge Injection
IH5051	2 SPDT	45	5	1000	500	0.8/2.4	0.01/0.01	Low Charge Injection
IH5140	SPST NO	50	0.1	150	125	0.8/2.4	0.01/0.01	Fast, Low Power
IH5141	2 SPST NO	50	0.1	150	125	0.8/2.4	0.01/0.01	Fast, Low Power
IH5142	SPDT	50	0.1	200	125	0.8/2.4	0.01/0.01	Fast, Low Power
IH5143	2 SPDT	50	0.1	200	125	0.8/2.4	0.01/0.01	Fast, Low Power
IH5144	DPST NO	50	0.1	200	125	0.8/2.4	0.01/0.01	Fast, Low Power
IH5145	2 DPST NO	50	0.1	200	125	0.8/2.4	0.01/0.01	Fast, Low Power
IH5341	2 SPST NO	75	0.5	300	150	0.8/2.4	0.001/0.001	70dB Isolation
IH5352	4 SPST NO	75	0.5	300	150	0.8/2.4	0.001/0.001	at 10MHz

High Voltage Analog Switches

Part Number	Function	$r_{DS(ON)}$ (Ω max)	$I_{D(OFF)}$ (nA max)	$t_{(ON)}$ (μs max)	$t_{(ON)}$ (ns max)	V_{IL}/V_{IH} (V)	Supply Current (I^+/I^- mA max)	Features
MAX341	2 SPST NO	75	60	1000	750	3.5/12	0.3/0.02	
MAX343	2 SPDT	75	60	1000	750	3.5/12	0.3/0.02	High Voltage,
MAX345	2 DPST NO	75	60	1000	750	3.5/12	0.3/0.02	$\pm$60V Operation
MAX348	2 SPST NO	45	60	1000	750	3.5/12	0.3/0.02	

Analog Multiplexers

Part Number	Function	$r_{DS(ON)}$ (Ω max)	$I_{D(OFF)}$ (nA max)	$t_{(ON)}$ (μs max)	V_{IL}/V_{IH} (V)	Analog Signal Voltage Range	Features
MAX310	1 of 8	250	10	1.5µs	0.8/2.4	±15V	70dB Isolation at 10MHz
MAX311	2 of 8	250	10	1.5µs	0.8/2.4	±15V	70dB Isolation at 10MH
MAX358	1 of 8	1500	2	1µs	0.8/2.4	-12.5V to +13.5V	Fault Protected to ±35V
MAX359	2 of 8	1500	2	1µs	0.8/2.4	-12.5V to +13.5V	Fault Protected to ±35V
MAX453	1 of 2	Buffered Output	10	0.12µs	0.8/2.4	±2V	On-Chip Output Amp
MAX454	1 of 4	Buffered Output	10	0.12µs	0.8/2.4	±2V	On-Chip Output Amp
MAX455	1 of 8	Buffered Output	10	0.12µs	0.8/2.4	±2V	On-Chip Output Amp
DG506A	1 of 16	400	5	1µs	0.8/2.4	±15V	Industry Standard
DG507A	2 of 16	400	5	1µs	0.8/2.4	±15V	Industry Standard
DG508A	1 of 8	300	2	1µs	0.8/2.4	±15V	Industry Standard
DG509A	2 of 8	300	2	1µs	0.8/2.4	±15V	Industry Standard
HI508A	1 of 8	1500	2	1µs	0.8/2.4	-12.5V to +13.5V	Fault Protected
HI509A	2 of 8	1500	2	1µs	0.8/2.4	-12.5V to +13.5V	Fault Protected
IH5108A	See MAX358						
IH6108	See DG508A						
IH6116	See DG506A						
IH6208	See DG509A						
IH6216	See DG507A						

Video Amplifiers and Buffers

Part Number	Type	Bandwidth (MHz)	Slew Rate (V/µs)	Output Current (mA)	Features
MAX450	Video Op Amp	10	70	30	Drives 75Ω
MAX451	Video Op Amp	10	70	30	1nA max I_{BIAS}
MAX452	Video Amp	50	150	20	Drives 75Ω
MAX453	Video Mux/Amp	50	150	20	2 Ch Input max
MAX454	Video Mux/Amp	50	150	20	4 Ch Input max
MAX455	Video Mux/Amp	50	150	20	8 Ch Input max
MAX460	Video Buffer	100	1500	100	Low I_B and C_{IN}
AD3554	Video Op Amp	1200	1000	100	High Gain
BB3553	Video Buffer	300	6000	200	Drive 50Ω
BB3554	Video Op Amp	1700	1000	100	High Gain
LH0033	Video Buffer	100	1500	100	Drives 75Ω
LH0063	Video Buffer	300	6000	200	Drives 50Ω
LH0101	Power Op Amp	5	10	2000	0.008% THD

Operational Amplifiers

Bipolar, Chopper Stabilized and Programmable Gain

Part Number	Initial VOS (μV max)	VOS Tempcos (μV/°C max)	I_{BIAS} (pA max)	Supply Voltage	Supply Current (mA max)	Noise DC-1Hz (μV pk-pk typ)	Features
MAX400	10	0.3	2000	±3V to ±18V	4.0	0.15	Non-Chopped
MAX420	10	0.05	30	±2.5V to ±16.5V	2.0	0.3	±15V Operation
MAX421	10	0.05	30	±2.5V to ±16.5V	2.0	0.3	±15V Operation
MAX422	10	0.05	30	±2.5V to ±16.5V	0.5	0.4	Low Current
MAX423	10	0.05	30	±2.5V to ±16.5V	0.5	0.4	Low Current
MAX430	10	0.05	30	±2.5V to ±16.5V	2.0	0.3	No Ext Capacitors
MAX432	10	0.05	30	±2.5V to ±16.5V	0.5	0.4	No Ext Capacitors
ICL7650	5	0.05	10	±2.25V to ±8V	2.0	0.7	Low I_{BIAS}
ICL7650B	10	0.1	20	±2.25V to ±8V	2.0	0.7	Lowest Cost
ICL7652	5	0.05	30	±2.5V to ±8V	2.0	0.2	Lowest Noise
LT1001	15	0.6	4000	±3V to ±18V	2.5	0.15	Non-Chopped
OP07A	25	0.6	2000	±3V to ±18V	4.0	0.15	Non-Chopped
OP07E	75	1.3	4000	±3V to ±18V	4.0	0.15	Non-Chopped
PGA100	0.5	6	1	+5V and ±8V to ±15V		0.2	Programmed Gain and Input Mux

CMOS

Part Number	Description	Compensation	Offset Null	V_{OS} Selection (mV max.)	I_{OS} (pA typ.)	I_B (pA typ.)	I_Q (μA typ.)
ICL7611	Single, Selectable I_Q	Internal	Yes	2, 5, 15	0.5	1	10-1000
ICL7612	Single, Selectable I_Q Extended CMVR	Internal	Yes	2, 5, 15	0.5	1	10-1000
ICL7614	Single, Fixed I_Q	External	Yes	2, 5, 15	0.5	1	100
ICL7616	Single, Selectable I_Q Extended CMVR	Internal	Yes	2, 5, 15	0.5	1	10-1000
ICL7621	Dual, Fixed I_Q	Internal	No	2, 5, 15	0.5	1	100
ICL7622	Dual, Fixed I_Q	Internal	Yes	2, 5, 15	0.5	1	100
ICL7631	Triple, Selectable I_Q	Internal	No	5, 10, 20	0.5	1	10-1000
ICL7632	Triple, Selectable I_Q	Internal	No	5, 10, 20	0.5	1	10-1000
ICL7641	Quad, Fixed I_Q	Internal	No	5, 10, 20	0.5	1	10
ICL7642	Quad, Fixed I_Q	Internal	No	5, 10, 20	0.5	1	1000

MAXIM'S ANALOG SWITCH FAMILIES

Family	Typical Devices	Features
General Purpose CMOS	DG201, IH5040	±15V Operation
High Voltage CMOS	MAX341, MAX343	±50V Operation
Fault Protected Muxes	MAX358, MAX359	±35V Fault Protected
Video Switches	IH5341, IH5352	70dB Isolation at 10MHz
Video Multiplexers	MAX310, MAX311	75dB Isolation at 5MHz, 1 of 8 and 2 of 8 Mux
Video Mux/Amps	MAX453, MAX455	50MHz GBW, Drives 75Ω

MAXIM

FAULT PROTECTED MULTIPLEXERS

- ±35V (70Vp-p) Fault Protection
- Only Nanoamps of Input Current Under Fault Conditions
- All Switches Off with Power Off

MAX358 – 1 of 8 Channels

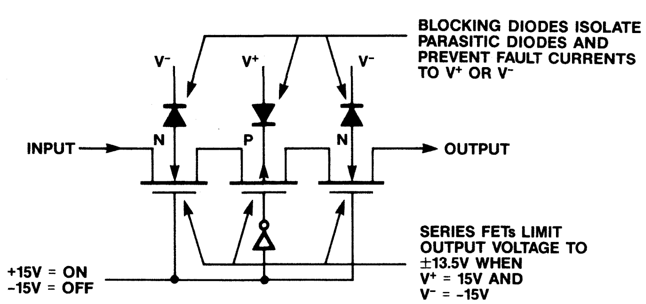
MAX359 – 2 of 8 Differentials

MAX368/369 – With Latches

MAXIM

Maxim's fault protected multiplexers use a series N-Channel, P-Channel, N-Channel structure in the signal path to improve fault over other designs. If the power to the multiplexer is turned off while input signals are applied, ALL channels turn off and only nanoamperes of leakage current will flow. This protects not only the multiplexer but also circuitry connected to its input and output.

FAULT PROTECTED MUX USING JUNCTION ISOLATED CMOS

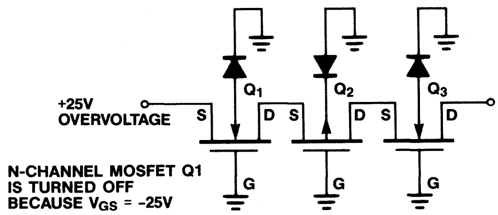


MAXIM

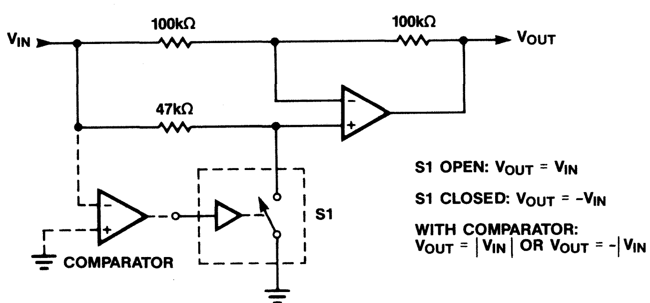
Maxim's protection structure has two significant advantages over the basic resistive current limit scheme used in other fault protected designs. First, fault currents are limited to much lower levels, nanoamps as opposed to 10s of milliamps. Second, Maxim's structure can withstand a continuous ±35V overvoltage which is not limited by power dissipation considerations.

The MAX358 has a "blocking" diode in series with each parasitic body-to-drain/source diode in the three-FET structure. Fault currents are limited to the reverse leakage current current of these diodes (4nA typ). In addition, when the input signal exceeds the power supplies (but not ±35V), the multiplexer output is clamped to 1.5V below the supply rails (i.e., ±13.5 with ±15V supplies). This output limiting works in both directions through the multiplexer.

OVERVOLTAGE WITH MULTIPLEXER POWER OFF



PROGRAMMABLE INVERTER/RECTIFIER

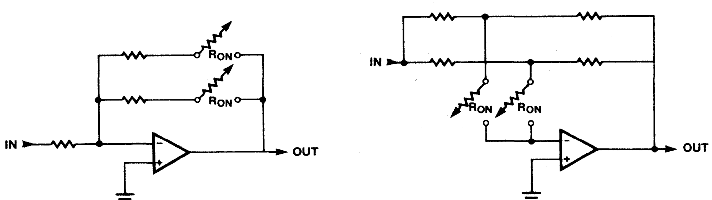


Analog switches can control the functioning of an op-amp. Here the op-amp is alternately an inverter or buffer, under control of the switch polarity. As a buffer, the gain is always 1, but as an inverter, the gain is set by the ratio of the input and feedback resistors. By adding a comparator, the function can be synchronously switched as the input polarity changes, effectively rectifying the output. The output polarity is determined by the switch logic (normally open or normally closed) and the comparator input polarity.

OP AMP GAIN SWITCHING TECHNIQUES

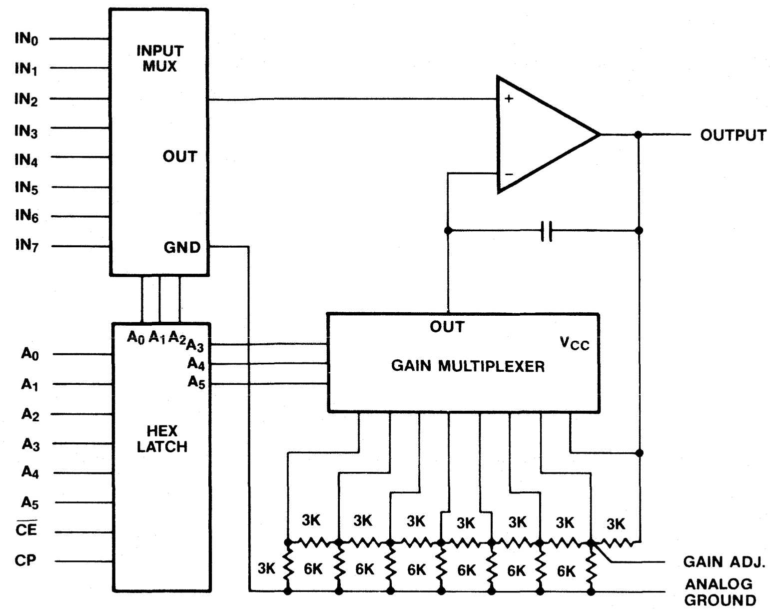
R_{ON} IS PART OF GAIN SETTING NETWORK

R_{ON} IS OUTSIDE GAIN SETTING NETWORK



Analog switches provide a convenient way to control op-amp gain. In the first of two methods shown here, the switch is placed in series with part of the gain determining network, and the switch on resistance (R_{DSon}) becomes part of the overall gain. If the resistors are very large, The switch on resistance (and its temperature coefficient) is insignificant. In the right hand diagram, the switches are in series with the op-amp input, so on resistance has no effect on the gain. In this configuration, however, switch leakage and charge injection may contribute to the op-amp's performance.

PGA100—8 CHANNEL PROGRAMMABLE GAIN AMP



The PGA100 is a high precision, digitally programmable gain amplifier (PGA) combined with an on-board 8 channel multiplexer. Any one of eight analog input channels can be selected for amplification by one of eight gains (1, 2, 4, 8, 16, 32, 64, 128). The gain and selected channel are internally latched for easy interface to microprocessors, and in addition the input multiplexer is fault protected for up to ±35V input signals. A guaranteed settling time of 5µs to 0.01%, means the PGA100 is ideal for high speed, 12 bit, channel scanning applications.

PGA100 FEATURES

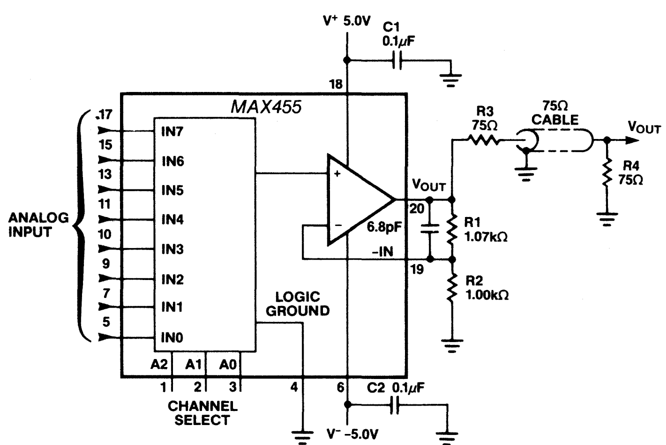
- 8 Inputs Fault Protected to ±35V
- Select Gain From 1 to 128
- ±0.02% Gain Accuracy
- 0.5mV Max Input Offset
- 0.01% Settling in 5µs

MAX452/3/4/5 VIDEO MUX/AMPLIFIERS

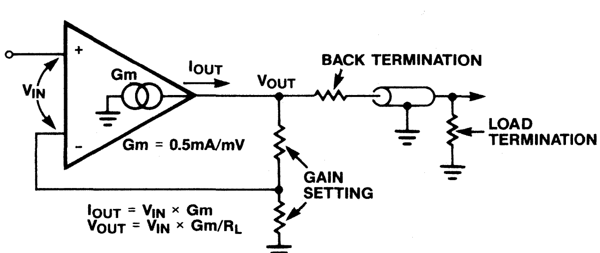
- 50MHz Gain Bandwidth
- Onboard Multiplexer
  8 Channels — MAX455
  4 Channels — MAX454
  2 Channels — MAX453
- High Impedance Inputs
  10pA Typical IBIAS
  9pF Input Capacitance
- 2pF On-State-Off-State Capacitance Change
- Drives 75W Coax ±1V
- Unity Gain Stable
  No External Compensation Needed

The MAX452 50MHz gain bandwidth video amplifier is made in a silicon gate CMOS process. This enabled Maxim to add an onboard multiplexer to make the MAX453, MAX454 and MAX455, which have a 2, 4, and 8 channel input multiplixer, respectively. Since the multiplexer is onboard, the output of the mux need only drive the very small input capacitance of the CMOS amplifier. Because of this, the analog switch impedance can be as high as 2kΩ. This, in turn keeps the input capacitance of each channel down to 7pF in the off state and 9pF when selected.

MAX455 BLOCK DIAGRAM

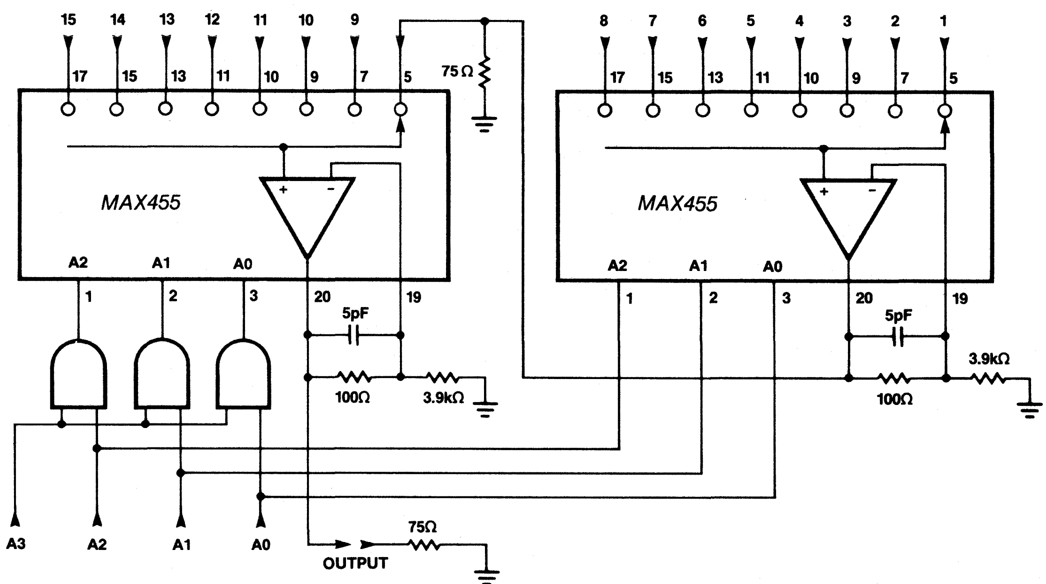


MAX452-455 OUTPUT STAGE



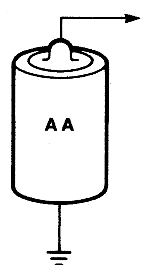
The output stage of the MAX452-455 is a current source. This allows these devices to have very high bandwidth, but it means that the output voltage is proportional to the output terminating resistance. Since the gain setting resistors are in parallel with the output, and are low values (to reduce R-C low pass effects), they must be included in the load calculation. If the load is some distance from the output, back termination is recommended to reduce reflections and match coaxial cables. This requires doubling the voltage gain.

1 OF 15 CASCADED VIDEO MUX



Two MAX455s can be cascaded to form a 1 of 15 video mux by connecting the output of one mux to one of the input channels of a second mux. Although the two devices are usually close to one another, the output of the first mux should be terminated to preserve its bandwidth.

POWER SUPPLY COMPONENTS



- FLYBACK CONVERTER BASICS
- DC-DC TRANSFORMER DESIGN
- 1 CELL TO 5V CONVERTER
- POWER SUPPLY APPLICATIONS

Voltage Regulators & Converters

Linear Voltage Regulators

Part Number	Output Voltage	Input Voltage	Quiescent Current (Typ/Max)	Output Voltage Accuracy	Features
AC-DC Regulators					
MAX600	Fixed 5V or 1.3V to 9V	110/220VAC	70μA/150μA	±4%	Full wave bridge
MAX601	Fixed 5V	110/220VAC	70μA/150μA	±4%	Half wave bridge
MAX602	Fixed 5V or 1.3V to 8V	6 to 9VAC	70μA/150μA	±4%	For use with isolation transformer
MAX610	Fixed 5V or 1.3V to 9V	110/220VAC	70μA/150μA	±4%	Full wave bridge
MAX611	Fixed 5V	110/220VAC	70μA/150μA	±4%	Half wave bridge
MAX612	Fixed 5V or 1.3V to 8V	6 to 9VAC	70μA/150μA	±4%	For use with isolation transformer
DC Linear Regulators					
MAX663	Fixed 5V or 1.3V to 15V	2V to 16.5V	6μA/12μA	±5%	+5V or programmable output
MAX664	Fixed -5V or -1.3V to -15V	-2V to -16.5V	6μA/12μA	±5%	-5V or programmable output
MAX666	Fixed 5V or 1.3V to 15V	2V to 16.5V	6μA/12μA	±5%	Fixed -5V or programmable output voltage Low Battery Detector
MAX667	Fixed 5V or 1.3V to 15V	2V to 16.5V	10μA	±5%	Low Dropout 250mA Output
ICL7663	1.3V to 15V	1.5V to 16V	4μA/10μA	±8%	
ICL7663A	1.3V to 15V	2.0V to 16V	4μA/10μA	±1%	Accurate output voltage
ICL7663B	1.3V to 10V	1.5V to 10V	4μA/10μA	±8%	
ICL7664	-1.3V to -15V	-2V to -16V	3.5μA/10μA	±8%	
ICL7664A	-1.3V to -15V	-2V to -16V	3.5μA/10μA	±1%	Accurate output voltage

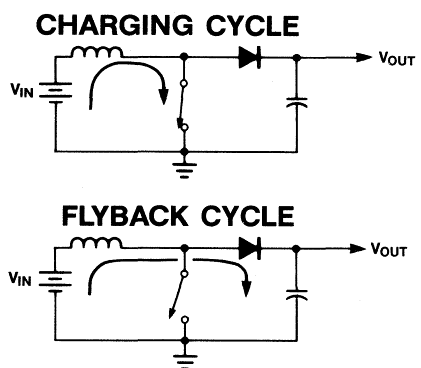
DC-DC Converters

Part Number	Description	Input Voltage	Output Voltage	Comments
Boost Converters				
MAX630/4193	DC-DC Boost Converter	2.0V to 16.5V	$V_{OUT} > V_{IN}$	Improved RC4193 2nd source
MAX631	DC-DC Boost Converter	1.5V to 5.6V	+5V	Only 2 external components
MAX632	DC-DC Boost Converter	1.5V to 12.6V	+12V	Only 2 external components
MAX633	DC-DC Boost Converter	1.5V to 15.6V	+15V	Only 2 external components
Inverting Converters				
MAX634/4391	DC-DC Voltage Inverter	2V to 16.5V	up to -20V	Improved RC4391 2nd source
MAX635	DC-DC Voltage Inverter	2V to 16.5V	-5V	Only 3 external components
MAX636	DC-DC Voltage Inverter	2V to 16.5V	-12V	Only 3 external components
MAX637	DC-DC Voltage Inverter	2V to 16.5V	-15V	Only 3 external components
Step Down Converter				
MAX638	DC-DC Voltage Stepdown	3V to 16.5V	$V_{OUT} > V_{IN}$	Only 3 external components
Boost Converters				
MAX641	High Power Boost Converter	1.5V to 5.6V	+5V	Drives external MOSFET
MAX642	High Power Boost Converter	1.5V to 12.6V	+12V	Drives external MOSFET
MAX643	High Power Boost Converter	1.5V to 15.6V	+15V	Drives external MOSFET
Charge Pump Converters				
MAX660	Negative Charge Pump	2.0V to 6.0V	$-V_{IN}$	100mA Output
MAX680	± Output Charge Pump	2.0V to 6.0V	±10V ($5V_{IN}$)	4 external capacitors
ICL7660	Negative Charge Pump	1.5V to 10V	$-V_{IN}$	Not regulated

Power MOSFET Drivers

Maxim Part No.	Supply Voltage	Peak Output Current (typ)	Max Delay Over Temp	Function
MAX626	4.5V to 18V	2A	60ns	Dual Inverting
MAX627	4.5V to 18V	2A	60ns	Dual Non-Inverting
MAX628	4.5V to 18V	2A	60ns	1 Inverting/1 Non-Inverting
ICL7667	4.5V to 15V		60ns	Dual Inverting
TSC426	4.5V to 18V	1.5A	120ns	Dual Inverting
TSC427	4.5V to 18V	1.5A	120ns	Dual Non-Inverting
TSC428	4.5V to 18V	1.5A	120ns	1 Inverting/1 Non-Inverting

DC-DC BASICS



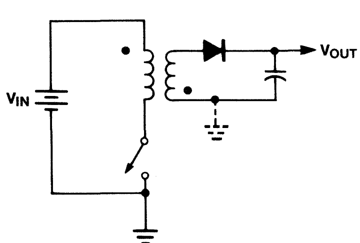
MAXIM

Nearly all of Maxim's DC-DC products employ a DC-DC conversion principle commonly termed "flyback", so named because the coil voltage inverts or "flies back" when power is removed. The best known example of this type of circuit is used to develop spark voltages in automobile ignitions. Flyback circuits are not difficult to understand once we believe the principle that inductors resist changes in their current.

Voltage is first applied to a coil. This induces a current which rises linearly until the voltage is removed. At that point the coil current looks for some other place to flow. A steering or catch diode directs this current to the output. The key to getting a higher voltage out than we put in is that the coil voltage rises to whatever level is needed to allow the current to keep flowing. The circuit output can then be regulated by monitoring the output voltage and controlling the switch which applies power to the coil.

This particular example shows a step-up or "boost" converter, however inverting and step-down circuits are no more complex. They differ only in the positions of the three basic components: coil, switch, and diode.

DC-DC FLYBACK TRANSFORMER



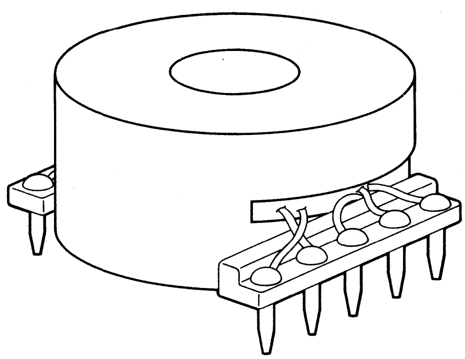
MAXIM

Flyback converters are not limited only to circuits with inductors. Transformers are also frequently used. Flyback transformer operation is very similar to the inductor example above. The only operating difference occurs during the period when the primary circuit is interrupted. Since no current path or steering diode is provided in the primary circuit, the flyback current flows in the secondary winding to the output.

TRANSFORMER ADVANTAGES

- Low Cost NFETs and NPN Transistors
- Electrically Isolated Outputs
- Multiple Outputs
- Simplified Circuitry

WINDING TRANSFORMERS



POT CORE DESIGN

- Start with Inductance and Peak Current
- Calculate LI^2
- Pick Smallest Allowed Core
- Calculate Number of Turns from A_L
- Select Largest Wire Size that Fits
- Check Wire Resistance

Though transformers are a bit more complex than coils in flyback circuits, the increased range of design options can provide cost and performance advantages:

1) Lower cost NFETs and NPN transistors can be used exclusively, even in inverting and step-down circuits. The switch source or emitter can always be grounded so drive circuitry remains simple.

2) With optical feedback or an additional feedback winding, the output can be regulated and electrically isolated from the input power source. This is important in industrial environments where ground integrity cannot be assumed, such as in telecommunications systems.

3) One drive circuit can generate several separate outputs if the transformer has multiple secondaries. One output is actively regulated and the turns ratios of the other secondary windings are sized for the desired output voltages.

4) In high voltage circuits and several "nonstandard" applications, the added cost of a transformer may be more than offset by simplified control circuitry.

POT CORE and TOROID BASICS

Winding inductors and transformers using pot cores and toroids can be very useful when prototyping DC-DC conversion circuits and need not be a painful process. The problem experienced by designers who are unfamiliar with magnetics is that an uncomplicated "working" solution is hard to find in what seems to be an impenetrable mass of specifications in the core manufacturers' catalogs.

Where do you start? Presented here is a short "minimum calculation" procedure for inductor and transformer design based on shortcuts pulled from manufacturers literature. It won't provide the "best" inductor for all applications, but for flyback switching converters which allow some shortcuts, it's a convenient way to designing prototypes and judge the feasibility of a design.

Standard values of inductors are stocked and sampled by several coil manufacturers (See appendix) so once a prototype coil is tried successfully it can often be replaced by a standard product in production. With transformers, winding prototypes is more useful because so few switching transformers are offered as standard products. A homemade prototype can also be a model for the magnetics manufacturer when designing the final production transformer.

"SHORTCUT" INDUCTOR AND TRANSFORMER DESIGN

1) Two basic inductor parameters are needed to start:

a) Inductance - L

b) Peak coil current - I_{pk}

These are determined from the initial DC-DC converter design. How these values are selected in various circuits is covered in Section 4. Most DC-DC converter applications in Maxim's literature will specify the required inductance and peak current. From the above parameters, calculate the LI^2 product in millijoules (mH x Amps^2 = mJ). This is actually two times the maximum amount of energy that the core must store ($E=LI^2/2$) but is the quantity specified in most pot core and toroid literature.

2) Pick the core size from the Pot Core Selection Guide on page 69. Find LI^2 on the X axis and draw a vertical line up from that point to the lowest intersecting diagonal line. Read the inductance factor, A_L, from the Y axis. The diagonal line and the corresponding value from the Y axis represent the smallest core size and the maximum A_L that may be used (A_L is the inductance per turns squared for the core). This means that any core of that size, with an A_L less than the Y axis value, will not saturate. Also, any other core that intersects the LI^2 line is usable if its A_L is less that the Y axis value. Of course if a core with too small an A_L is selected then the needed number of turns may not fit on the core.

In flyback transformers, wire size, resistance, and "fitting" the turns on the core is usually not a problem. This is because the specified inductances are small compared to other types of switching transformers so few turns are needed. This often allows the use of significantly smaller A_L values than provided by the graph on page 69, which reduces core losses. This gain however must be weighed against the increased wire losses.

3) The number of turns of wire, N, to be wound on the core for the desired inductance is:

$N = 31.6 \sqrt{(L/A_L)}$, where L is the inductance in µH.

4) Knowing how many turns are needed, we then want the largest wire that will fit on the core (within reason) to minimize the DC resistance. The available winding area, A_W, can either be directly found in the data for the core bobbin, or by using the Ferrite Pot Core Winding Dimensions table on page 69.

$$T \geq N/A_W$$

T, the turns/cm^2 from the Wire Table on page 70, must be greater than the required number or turns, N, divided by the available winding area, A_W. A wire size with an adequate value for T should be used.

5) Once the wire size is picked, the resistance is checked using data from the Wire Table and the Winding Dimensions table (pages 69 and 70).

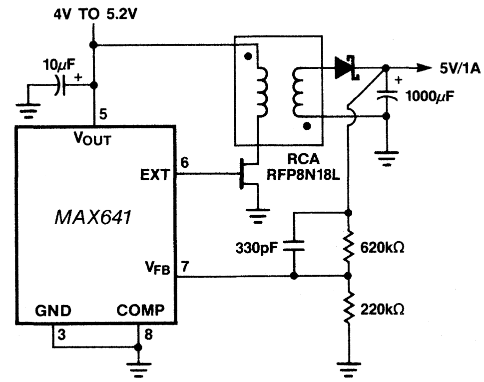
$$R = N l_W r_W$$

N is the number of turns, l_W is the average length of a turn for the selected core, and r_W is the resistance/foot of the selected wire. R should be such that no more than 1 or 2% of the total power is lost in the wire resistance. The power lost due to wire resistance in flyback circuits is roughly:

$$P_R = (I_{pk}^2 \times R)/3$$

If the inductor or transformer is extremely small, then a somewhat larger percentage may be allowed. This is not normally a problem in flyback circuits since the inductance and hence the number of turns required is relatively small compared to other converter types. If the resistance is too high, a larger wire size (with

DC-DC TRANSFORMER EXAMPLE



lower ohms/foot), and possibly a larger core to accept the larger wire diameter, must be used.

DESIGNING TRANSFORMERS

Transformers for flyback converters are no more difficult to wind than inductors. The primary winding is designed in the same manner as above and basic turns ratios are used to design the secondary winding(s). The ratios will usually be proportional to or slightly less than the input/output voltage ratios, i.e., 2/1 for a 10V to 5V circuit or 1/3/3 for 5V to ±15V.

TRANSFORMER EXAMPLE

One application where a transformer is a desirable alternative to an inductor is in this DC-DC converter which generates 5V at 1 Amp from 4 NiCad cells. Since the input voltage varies from 4V to 5.2V neither a step-up nor a step-down converter works over the full input range. Also, the input and output voltages are too low to drive the P-Channel MOSFETs normally used in step-down circuits. A transformer allows us to not only step up or down, but also to use a lower cost N-channel FET grounded source switch. With a transformer we can also cut all power to the load when the converter is shut down, which is something a basic step-up converter is not able to do.

First of all, the required output power, assuming a 0.5V forward diode drop (1N5817), is 5.5W. With a 50kHz clock frequency (MAX641) and a minimum input voltage of 4V, the inductance required, L_P, is:

$$L_P = IN^2/(8f_O I_{OUT}(V_{OUT}+V_D))$$
$$= 4^2/(8 \times 50\text{kHz} \times 1 \times (5 + 0.5) = 7.27\text{uH}$$

The peak current in the transformer, calculated using the highest input voltage (5.2V), is:

$$I_{pk} = V_{IN}/(2f_O L) = 7.15 \text{ Amps}$$

Then LI^2 = 0.372 millijoules (mJ)

Using the Pot Core Selection Guide (page 69), 0.372 millijoules leads us to a 18x11 mm pot core with an A_L value of no more than 330.

We select a standard core with an A_L of 250. The number of turns needed to wind a 7.27µH primary is:

$$N = 31.6\sqrt{(7.27/250)} = 5.4$$

The primary/secondary turns ratio for this application is 1/1 so we also want a secondary winding with the same number of turns. The winding area, A_W, for each section of a two section bobbin, as indicated in the Winding Dimensions table, is 0.084 cm^2.

$$T \geq N/A_W = 5.4/0.084 = 64$$

T is the minimum turns/cm^2 for the wire size that fits on the bobbin. The Magnet Wire Table (Table 3) indicates #18 wire (at 79.1 turns/cm^2) or smaller. The resistance from the Wire table is 6.386Ω/1000 ft. and the average length per turn from the Winding Dimensions table is 0.121 feet, so the winding's resistance is:

$$R = N l_w r_w = 5.4 \times 0.121 \times 0.00638 = 0.0042\Omega$$

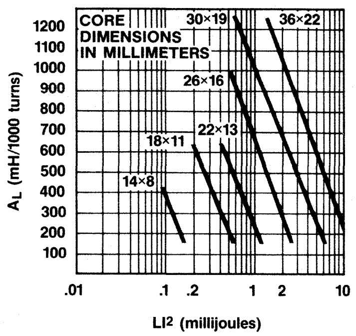
The average wire loss with a peak current of 7.15 Amps is then:

$$(I_{pk}^2 \times R)/3 = (7.15^2 \times 0.0042)/3 = 72\text{mW}$$

This is only 1.4% of the output power and should not be a problem. Our final transformer design is then:

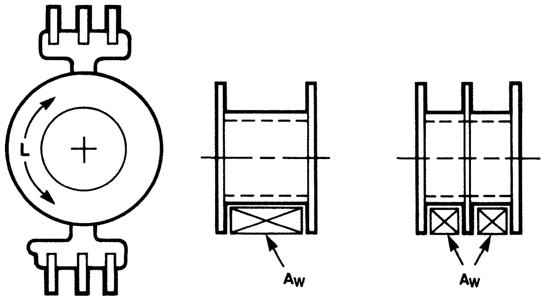
1) 7.27 µH (about 10% tolerance) primary inductance
2) 1/1 primary/secondary turns ratio
3) 18 x 11 mm pot core with an A_L of 250 (example core: Magnetics Inc. #G-41811-25)

POT CORE SELECTION GUIDE



This graph determines the smallest core that will work for a given LI^2 product. See Pot Core and Toroid Basics, page 67, for a detailed explanation.

WINDING AREA AND LENGTH



A_W = WINDING AREA
L = AVERAGE WINDING LENGTH

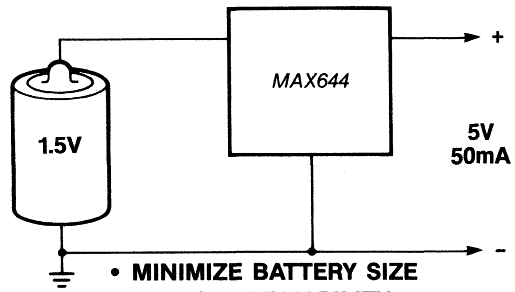
Ferrite Pot Core Winding Dimensions

Core Size (mm X mm)	Winding Area Per Section (cm^2) 1 sect	2 sect	3 sect	Avg Length/Turn (Ft.)
14 X 8	0.098	0.044	N/A	0.0953
18 X 11	0.170	0.084	0.049	0.121
22 X 13	0.292	0.138	0.087	0.145
26 X 16	0.421	.202	0.128	0.173
30 X 19	0.542	0.254	0.159	0.204
36 X 22	0.755	0.357	0.225	0.244

Magnet Wire Table

Wire Size AWG	Wire Area (max) HEAVY	(Circular Mils) QUAD	Turns per in2	Turns per cm2	Resistance Ohms/1000'	Current Capacity (ma) @750 Cir. Mil/amp	Current Capacity (ma) @500 Cir. Mil/amp
10	11,470	12,230	89	13.8	.9987	13,840	20,768
11	9,158	9,821	112	17.4	1.261	10,968	16,452
12	7,310	7,885	140	21.7	1.588	8,705	13,058
13	5,852	6,336	176	27.3	2.001	6,912	10,368
14	4,679	5,112	220	34.1	2.524	5,479	8.220
15	3,758	4,147	260	40.3	3.181	4,347	6,520
16	3,003	3,329	330	51.2	4.020	3,441	5,160
17	2,421	2,704	410	63.6	5.054	2,736	4,100
18	1,936	2,190	510	79.1	6.386	2,165	3,250
19	1,560	1,781	635	98.4	8,046	1,719	2,580
20	1,246	1,436	800	124	10.13	1,365	2,050
21	1,005	1,170	1,000	155	12.77	1,083	1,630
22	807	949	1,200	186	16.20	853	1,280
23	650	778	1,500	232	20.30	681	1,020
24	524	635	1,900	294	25.67	539	808
25	424	520	2,400	372	32.37	427	641
26	342	424	3,000	465	41.0	338	506
27	272	342	3,600	558	51.4	259	403
28	219	276	4,700	728	65.3	212	318
29	180	231	5,600	868	81.2	171	255
30	144	188	7,000	1,085	104	133	200
31	117	154	8,500	1,317	131	106	158
32	96.0	128	10,500	1,628	162	85	128
33	77.4	104	13,000	2,015	206	67	101
34	60.8	82.8	16,000	2,480	261	53	79
35	49.0	67.2	20,000	3,100	331	42	63
36	39.7	54.8	25,000	3,876	415	33	50
37	32.5	44.9	32,000	4,961	512	27	41
38	26.0	36.0	37,000	5,736	648	21	32
39	20.2	28.1	50,000	7,752	847	16	25
40	16.0	22.1	65,000	10,077	1,080	13	19
41	13.0		80,000	12,403	1,320	11	16
42	10.2		100,000	15,504	1,660	8.5	13
43	8.40		125,000	19,380	2,140	6.5	10
44	7.30		150,000	23,256	2,590	5.5	8
45	5.30		185,000	28,682	3,348	4.1	6.2

SINGLE CELL 5V CONVERTER



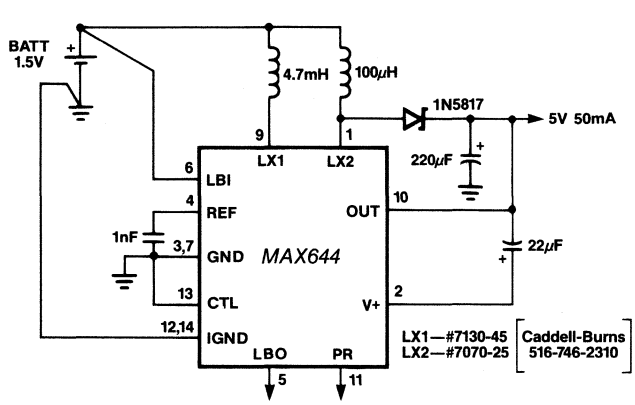
- MINIMIZE BATTERY SIZE
- IMPROVE RELIABILITY
- ANY ONE-CELL BATTERY
  ALKALINE NiCad
  LITHIUM LEAD ACID

The MAX644 efficiently generates a regulated 5 volt supply when only a very low voltage input, such as a single cell 1.3V battery, is available. The MAX645 is optimized for slightly higher inputs, as with two alkaline cells or possibly one lithium cell. 50mA output current is supplied with minimal external components by employing a unique double-conversion technique (patent pending) where one micro-power boost converter generates 12 volts for the internal MOSFET switch of a second internal boost converter. This way the MOSFET receives adequate gate voltage for low on resistance and typically 80% conversion efficiency.

MAX644/645 HIGHLIGHTS

- Low Guaranteed Startup Voltage — 1.15V
- 5V/50mA Output with 1.3V Input
- Minimum Input Voltage — 0.5V to 1.0V
- Standby Current — 80µA (At 1.3V In)
- 80% Conversion Efficiency
- Low Battery Indicator

MAX644—TYPICAL CIRCUIT



Any components that could be integrated on-chip have been included. A total of six external components include:

1) Two small inductors. Surface mount coils may also be used in some cases.

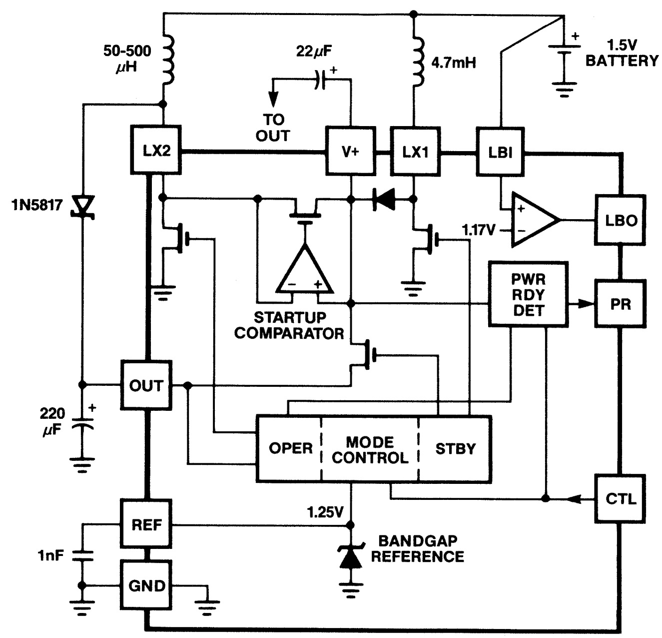
2) Two electrolytic filter capacitors

3) One reference bypass capacitor

4) One diode - A Schottky diode is recommended for best efficiency

The V^+ output is approximately +12V when the chip is operating and drops to 5V when the standby mode is activated. The control input selects between these two modes.

MAX644/645 BLOCK DIAGRAM



A number of features are included in the MAX644 to minimize external components in battery powered applications.

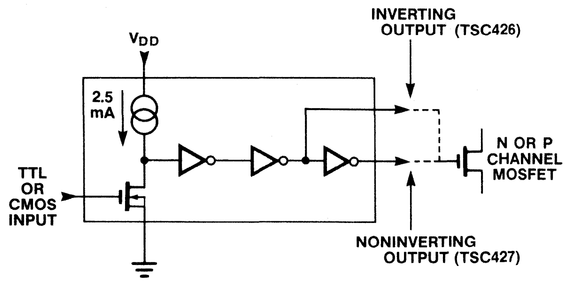
1) A standby mode in which 5V can still be supplied at low current and where quiescent current drops to 80µA.
2) A low battery comparator output which goes low when the input battery voltage drops below 1.15V.
3) An internal 1.25 bandgap reference.
4) A "Control" input which allows the standby or high power mode to be activated by a switch or logic level.
5) A "Power Ready" output which goes high when the 5V output has reached its proper level after power-up or termination of standby.

MAX626/627/628 MOSFET DRIVERS

- 4.5V to 18V Supply Range
- 20ns Typ Rise and Fall Time into 1nF
- TTL/CMOS Compatible Input
- Drives N and P-Channel MOSFETs
- High Performance Replacement for TSC426/427/428 and ICL7667
- 8 Pin DIP and SO Packages

Drivers for power MOSFETs are essentially very low output impedance inverters or buffers, optimized to quickly turn MOSFETs on and off. They are also useful wherever a CMOS buffer or inverter with a 4 ohm output is needed. In addition to power FETs, these devices are often used to drive relays, charge pump circuits, and pulse generators.

MAX626/627/628
DUAL POWER MOSFET DRIVER



COMING SOON
MAX660 – 100mA VOLTAGE INVERTER

- Pin Compatible with ICL7660
- 100mA Output Current
- 10 Ohm Output Impedance
- Charge Pump Uses No Inductors
- Connect Two 47µF Caps

It is important to note that charge pump voltage inverters such as the ICL7660 and MAX660 do *not* have regulated output. Their open circuit output voltage is very nearly -(V_{IN}), but the output voltage is less when a load is applied. The ICL7660's output impedance is typically 55Ω and is guaranteed to be no more than 100Ω. This means that with a +5V input and a 20mA load, the output voltage is typically -4.5V and is -4V worst case.

The new MAX660 has a typical output impedance of only 5Ω (10Ω max). With a 20mA load and a +5V input, the output voltage is -4.8V worst case and only drops to -4.0V (guaranteed) when 100mA is drawn. The MAX660 also has thermal shutdown which turns off the oscillator and output in the event of a short circuit. A control pin is also supplied that increases the switching frequency from 12kHz to 40kHz to reduce output ripple.

COMING SOON
MAX667 – LOW DROPOUT REGULATOR

- 0.12V Dropout Voltage At 100mA
- 250mA Output Current
- Shutdown Control Input
- < 1µA Shutdown Current
- 10µA Quiescent Current
- Dropout Indicator

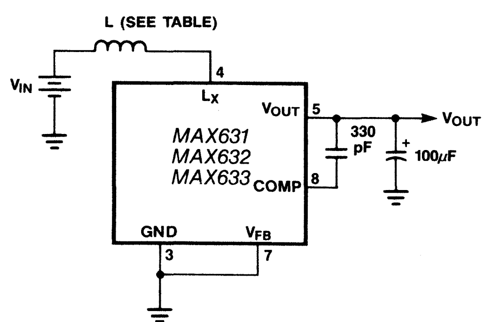
The MAX667 complements Maxim's family of micro-power regulators. Its bipolar PNP output stage supplies 250mA and its minimum input-to-output voltage is no more than the PNPs saturation voltage. This is particularly useful in 5V system powered from 5-cell NiCad or 3-cell lead-acid batteries.

The MAX667 PNP transistor's base current flows to ground and not through the load, therefore the IC includes circuitry which adjusts the base current as a function of the load to minimize quiescent current. The MAX667 also features a shutdown input which turns off the output and reduces the quiescent current to 1µA, and a "Dropout" output which goes high as the output PNP begins to saturate.

POWER SUPPLY APPENDIX

POWER SUPPLY CIRCUIT APPLICATIONS

LOW POWER STEP-UP CONVERTERS

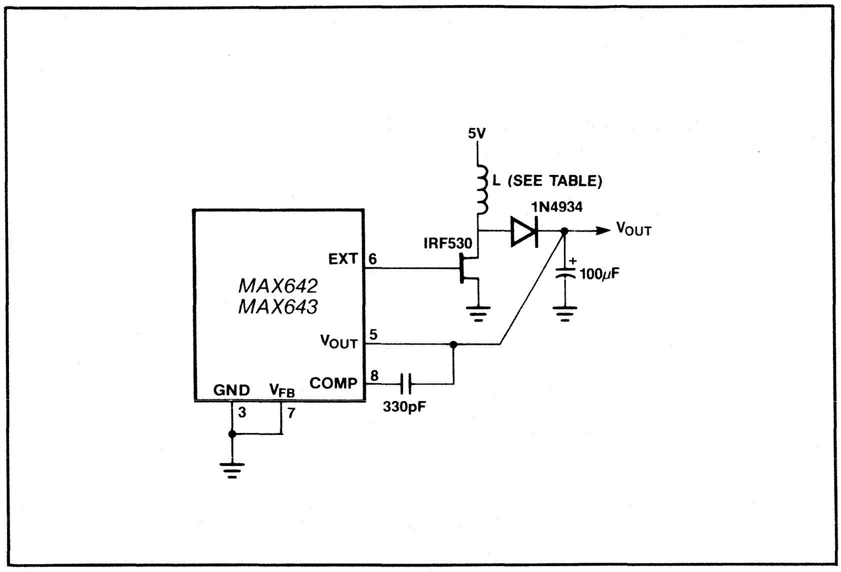


The table lists measured efficiency of circuits using the indicated coils. Efficiencies can be improved slightly by placing a Schottky diode such as the 1N5817 in parallel with the internal MAX631/632/633 diode, from pin 4 to 5. The increase in efficiency is most noticeable for the 5V output circuits.

Maxim Part No.	V_{IN} (V)	V_{OUT} (V)	I_{OUT} (mA)	Typ Eff (%)	Part No.*	Inductor (L) µH	Ω
MAX631	2	5	5	78	6860-21	470	0.44
	2	5	10	74	6860-17	220	0.28
	2	5	15	61	6860-13	100	0.1
	3	5	25	82	6860-21	470	0.44
	3	5	40	75	7070-29	220	0.55
MAX632	3	12	5	79	6860-10	330	0.35
	3	12	10	79	7070-28	180	0.48
	5	12	12	88	6860-21	470	0.44
	5	12	25	87	6860-19	330	0.35
MAX633	3	15	5	73	7070-29	220	0.55
	3	15	8	71	7070-27	150	0.43
	5	15	10	85	6860-21	470	0.44
	5	15	15	85	6860-19	330	0.35
	8	15	35	90	6860-21	470	0.44

* Caddell-Burns, NY, (516) 746-2310

Medium Power Step-Up Converters

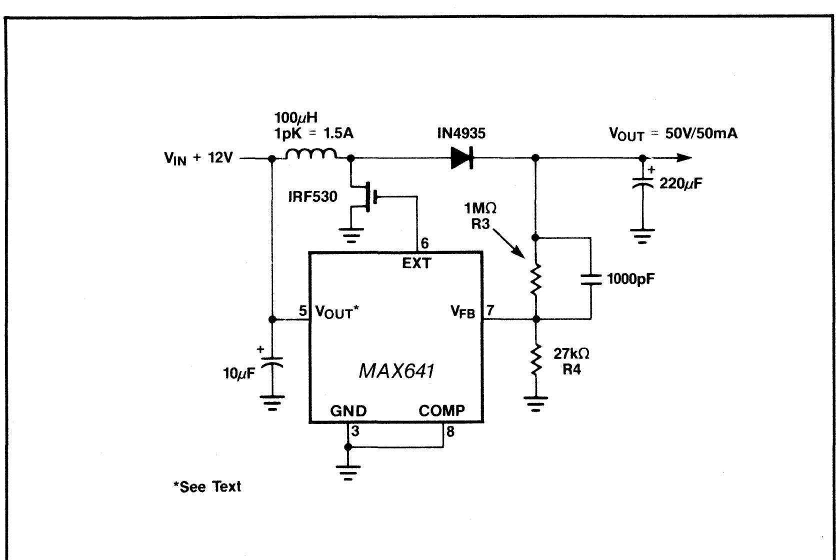


In selecting coils for DC-DC converter circuits that use external MOSFETs it is important to calculate the peak current and to select a coil that will not saturate at that peak current.

Maxim Part No.	V_{IN} (V)	V_{OUT} (V)	I_{OUT} (mA)	Typ Eff (%)	I_{pk} (A)	Inductor (L) Part No.*	µH	Ω
MAX642	5	12	200	91	1.2	6860-08	39	0.05
	5	12	350	89	2	6860-04	18	0.03
	5	12	550	87	3.5	7200-02	12	0.01
MAX643	5	15	100	92	1.2	6860-08	39	0.05
	5	15	150	89	1.5	6860-06	27	0.04
	5	15	225	89	2	6860-04	18	0.03
	5	15	325	85	3.5	7200-02	12	0.01

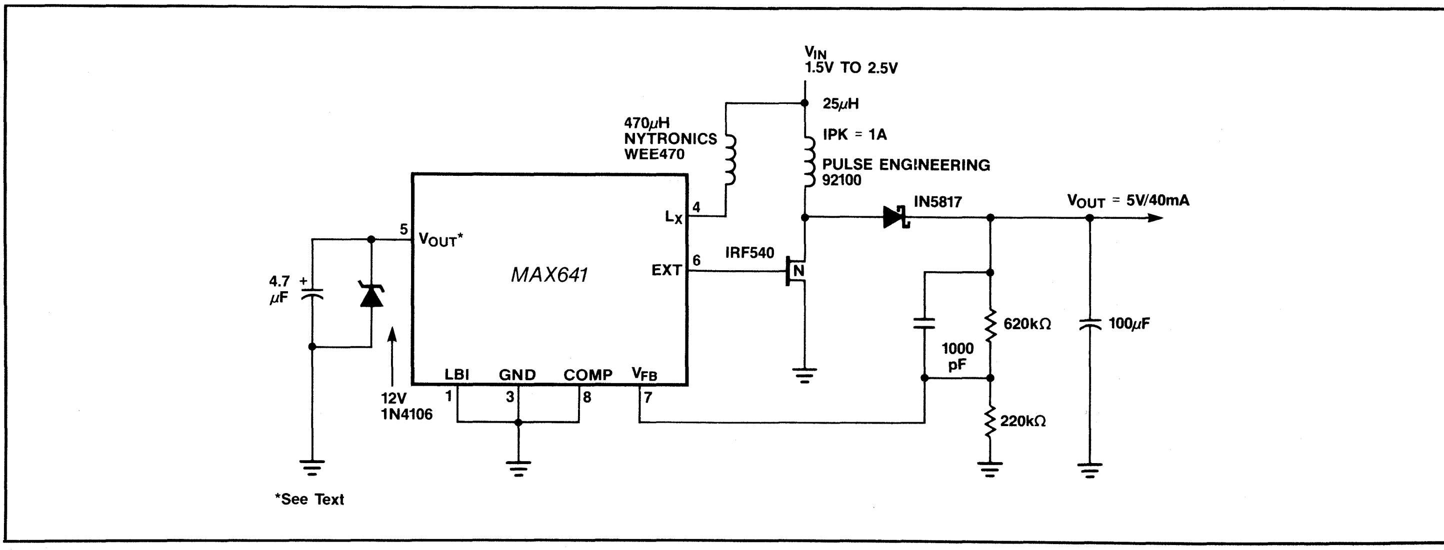
* Caddell-Burns, NY, (516) 746-2310

High Voltage Step-Up Converter



The output voltage limits of the MAX6xx series DC-DC converters can be exceeded once an external FET or transistor with an adequate voltage rating is used as the switch. Here a +12V input is converted to +50V at 50mA by adding an IRF530 N-channel FET which has a voltage rating of 100V. The circuit differs from the basic MAX641 hookup in that an external resistor divider must provide the feedback signal to the V_{FB} input and that chip power comes from the +12V input via the V_{OUT} pin.

Low Voltage Battery to +5V

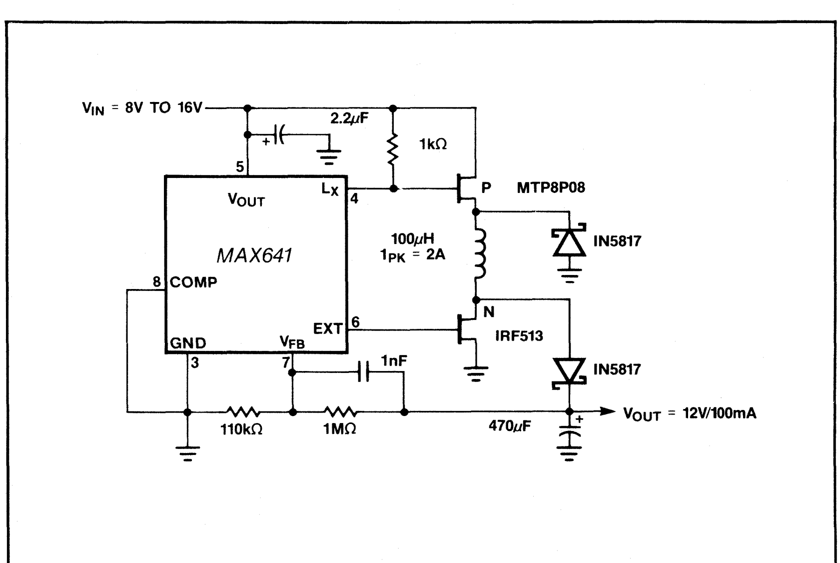


By connecting a second inductor to the Lx output of a MAX641 step-up DC-DC converter, the efficiency and power handling ability of converters with a low voltage input can be dramatically improved. This can supply 5V at 40mA with only 1.5V input. With 2.4V input, it can supply 180mA at 5V.

The 470uH coil connect to the Lx pin forms a second step-up converter whose only function is to generate a local power supply for the MAX641 and the power MOSFET gate. This local supply is clamped to 12V by the zener connected to the MAX641 VOUT pin. The higher gate drive voltage allows the use of the industry standard IRF540 MOSFET and ensures that it will have a low on resistance. Without the second coil (470uH) the IRF540 MOSFET gate drive voltage would be the same as the input battery voltage and the circuit would not turn on.

* VOUT is actually the MAX641's voltage input and not an output per se. The pin is labeled this way because it usually connects to the circuit output to provide power and the feedback signal back to the chip when the MAX641 is used in its standard configuration. When a MAX64x or MAX63x series device is used in other than the basic configurations, such as here, the VOUT pin is frequently NOT the output of the DC- DC circuit.

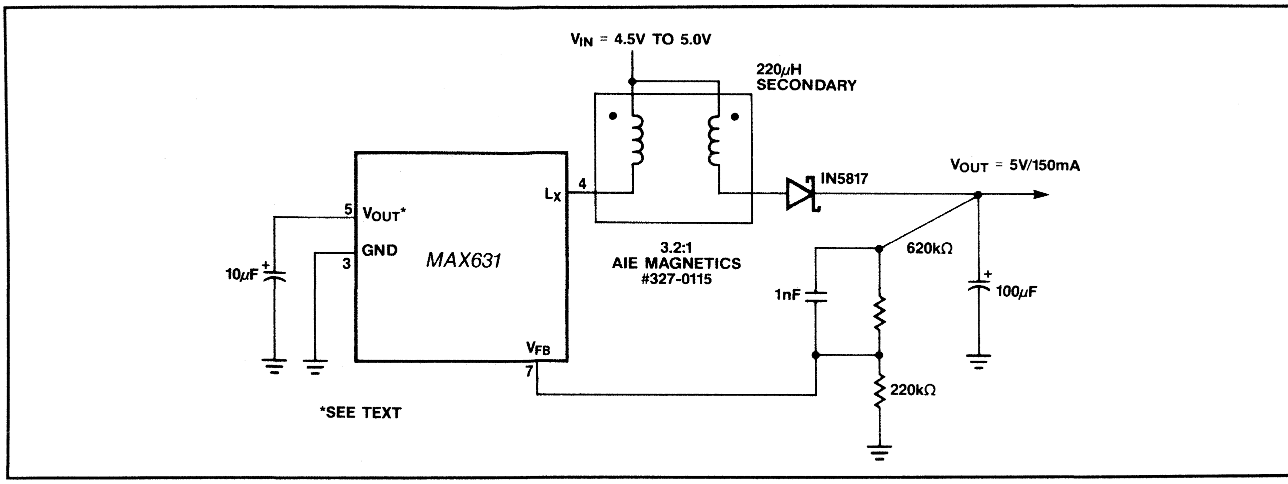
Step Up/Down DC-DC Converter



Positive output step-up and step-down DC-DC converters have a common limitation in that neither can handle input voltages which may be both greater than or less than the output. For example, when converting a 12V sealed lead acid battery to a regulated +12V output, the battery voltage may vary from a high of 15V down to 10V.

By using a MAX641 to drive separate P- and N-channel MOSFETs, both ends of the inductor are switched to allow noninverting buck/boost operation. A second advantage of the circuit over most boost-only designs is that the output goes to 0V when SHUTDOWN is activated. A drawback is that efficiency is not optimum because 2 MOSFETs and 2 diodes increase the losses in the charge and discharge path of the inductor. The circuit delivers +12V at 100mA at 70% efficiency with an 8V input.

Long-Line IR Drop Voltage Recovery

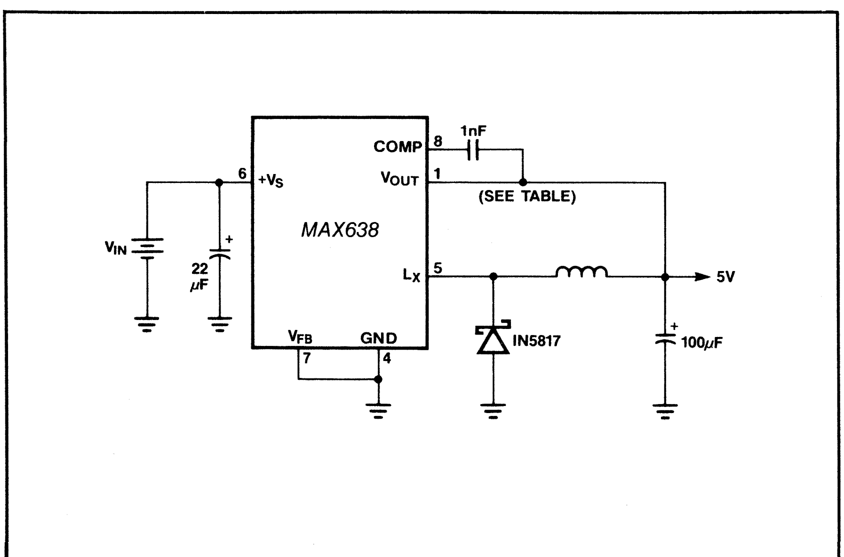


This circuit provides a unique solution to a common system- level power distribution problem: When the supply voltage to a remote board must traverse a long cable, the voltage at end of the line sometimes drops to unacceptable levels. This "+5V to +5V" converter, addresses this by taking the reduced voltage at the end of the supply line and boosting it back to +5V. The can be especially useful in remote display devices such as some point-of-sale (POS) terminals where several meters of cable may separate the terminal from the read-out.

A MAX631 and a small transformer restore the 5V to 4.5V input back to 5V. The 3.2/1 turns ratio of the transformer allows the MAX631 to provide more than its usual output current, without an external MOSFET, at these relatively low operating voltages. Output current is 5V at 150mA with a 4.5V input.

* The MAX631 also makes use of the reflected voltage in the transformer primary to generate a higher supply voltage of about +9V for itself at V_{OUT}. By operating at 9V rather than 5V, the on resistance of L_X is reduced.

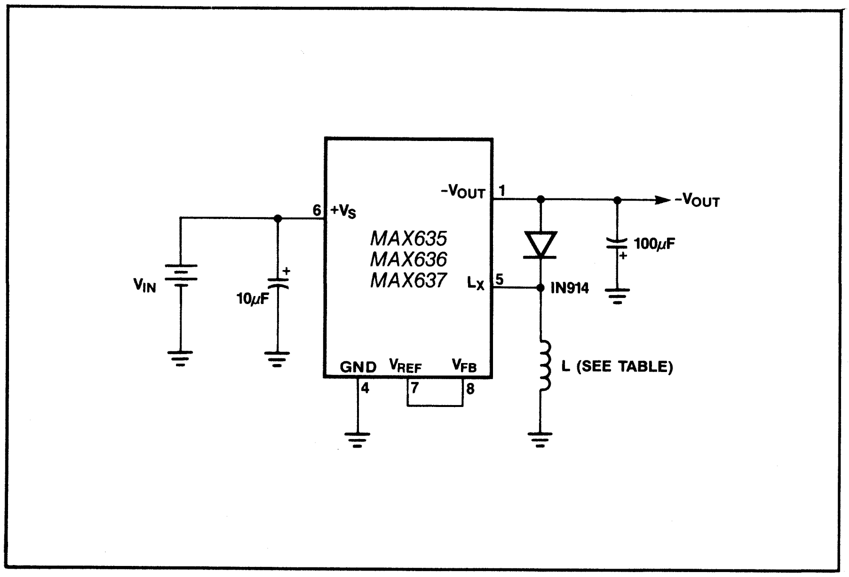
LOW POWER STEP-DOWN CONVERTERS



Maxim Part No.	V_{IN} (V)	V_{OUT} (V)	I_{OUT} (mA)	Typ Eff (%)	I_{pk} (A)	Inductor (L) Part No.*	µH	Ω
MAX638	7-9.5	5	35	92	200	7070-27	150	0.4
	8-9.5	5	55	89	200	7070-27	150	0.4
	10-14	5	50	92	300	7070-30	270	0.6
	12	5	60	92	250	7070-30	270	0.6
	12	5	75	89	300	7070-28	180	0.5

* Caddell-Burns, NY, (516) 746-2310

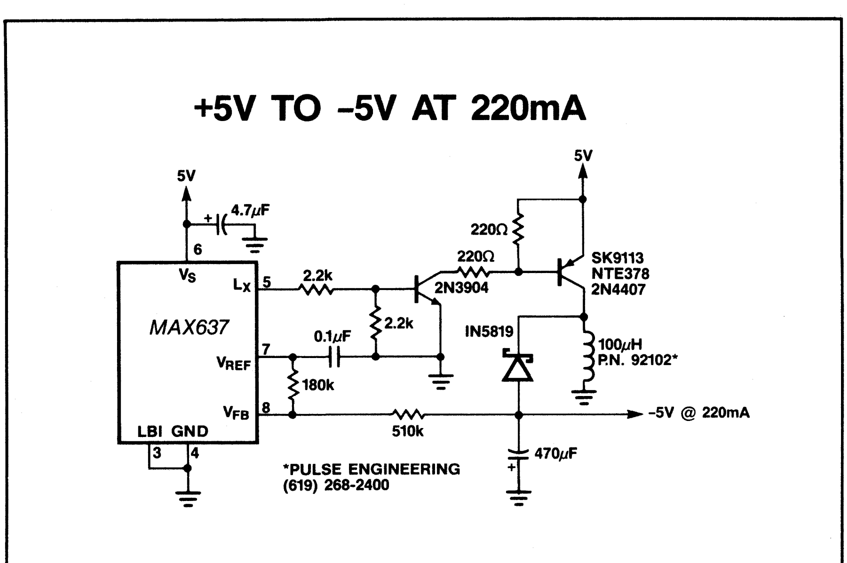
LOW POWER INVERTERS



Maxim Part No.	VIN (V)	VOUT (V)	IOUT (mA)	Typ Eff (%)	Part No.*	Inductor (L) µH	Ω
MAX635	+3	-5	5	60	6860-19	330	0.35
	+5	-5	25	76	6860-19	330	0.35
	+9	-5	40	79	6860-19	330	0.35
	+12	-5	45	85	6860-21	470	0.40
	+15	-5	50	90	6860-23	680	0.55
MAX636	+5	-12	12	74	6860-19	330	0.35
	+9	-12	30	84	6860-19	330	0.35
	+12	-12	40	89	6860-21	470	0.40
MAX637	+3	-15	2	65	6860-21	470	0.40
	+5	-15	8	77	6860-19	330	0.35
	+9	-15	25	85	6860-19	330	0.35

* Caddell-Burns, NY, (516) 746-2310

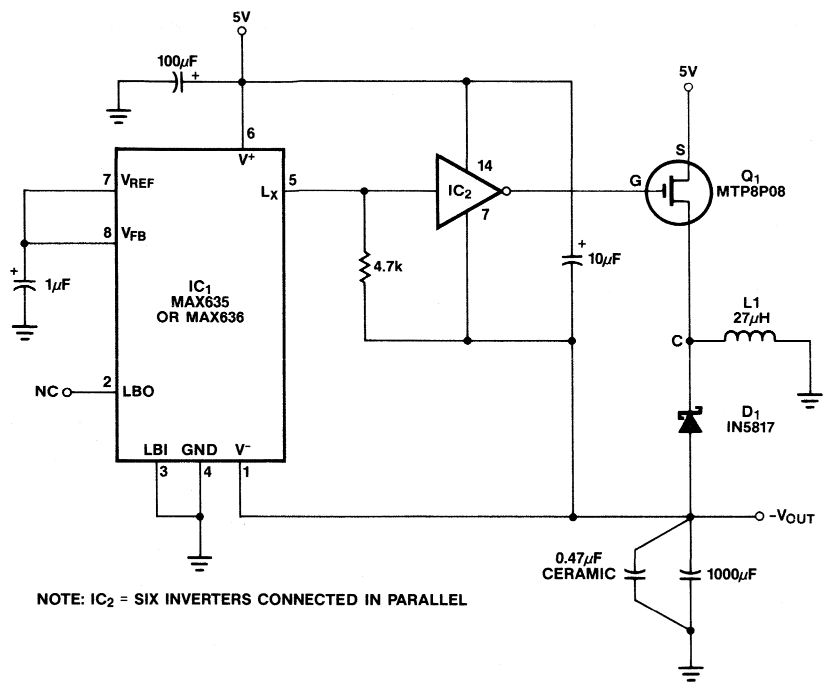
+5V to -5V at 220mA



The absolute maximum peak current rating of the MAX637 Lx pin is 525mA. This does NOT correspond to 525mA of average output current. The MAX637 is suitable only for low power circuits of up to about 20mA output when converting +5V to -5V with no external buffering.

This circuit uses two transistors to buffer the Lx output to achieve 220mA of output current. The 2N3904 is a small signal NPN transistor used to invert the Lx output. The 2N3904 then drives a power PNP transistor. The signal at the collector of the PNP transistor is equivalent to the normal MAX636 Lx output, but it has a much larger peak output current rating. When using external transistor buffers, the output voltage is set by an external feedback resistor network; in this case, the 510kΩ and 180kΩ resistors.

Medium Power Inverters



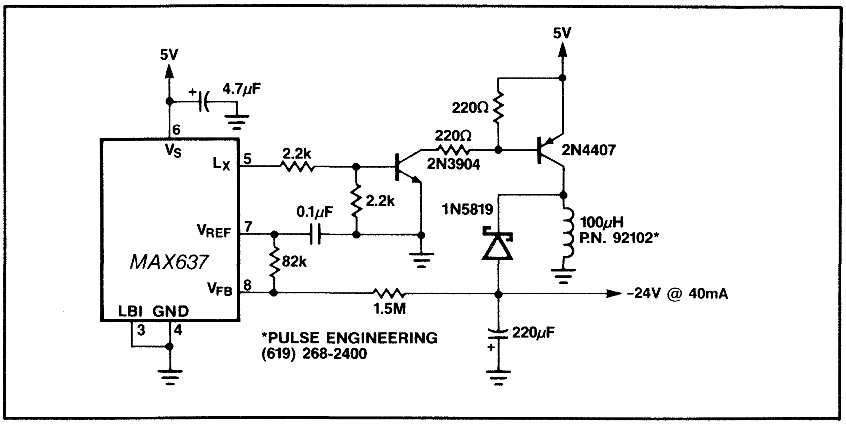
V_{IN}	$-V_{OUT}$	I_{OUT}	Efficiency	IC_1	L_1
5V	-5V	400mA	70%	MAX635	27μH
5V	-5V	500mA	64%	MAX635	18μH
5V	-12V	150mA	75%	MAX636	27μH
5V	-12V	200mA	70%	MAX636	18μH

NOTES:
18μH Coil = Caddell-Burn's (Mineola, NY) Model 6860-04.
27μH Coil = Caddell-Burn's Model 6860-06.

In this circuit a CMOS inverter such as the CD4069 is used to convert the open drain L_X output to a signal suitable for driving the gate of an external P MOSFET. The MTP8P08 has a gate threshold voltage of 2.0V to 4.5V so it will have a relatively high resistance if driven with only 5V of gate drive. To increase the gate drive voltage, and thereby increasing efficiency and power handling capability, the negative supply pin of the CMOS inverter is connected to the negative output rather than to ground. Once the circuit is started, the P MOSFET gate drive swings from +5V to $-V_{OUT}$.

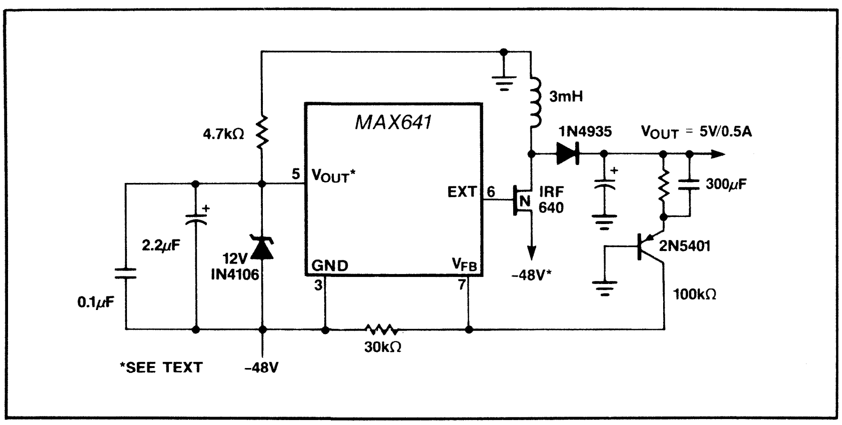
At startup the $-V_{OUT}$ is one Schottky diode drop above ground and the gate drive to the power MOSFET is slightly less than 5V. The output should be only lightly loaded to ensure startup, since the output power capability of the circuit is very low until $-V_{out}$ is a couple of volts negative.

+5V to -24V at 40mA



Similar to the preceding circuit, the 2N4407 PNP transistor is a buffered replica of the MAX637 Lx output. The 2N4407, though, has a high breakdown voltage and can be used to generate a -24V output. The -24V output does not appear directly on any pin of the MAX637 since it is sensed via the 1.5MΩ external feedback resistor.

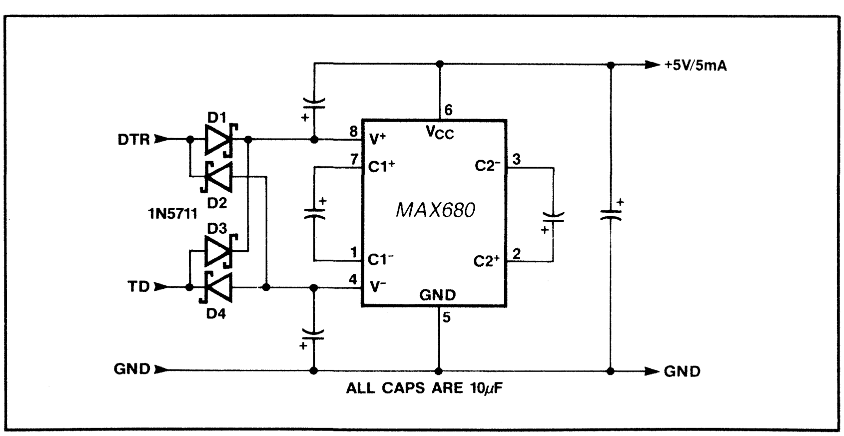
Telecom -48 to 5V at 0.5A



The small current consumption of a MAX641 allows it to be biased at a -50V rail with a shunt zener diode so that it can convert -50V to +5V. This is a common requirement in telecom systems where logic circuitry must be powered from the central office battery voltage.

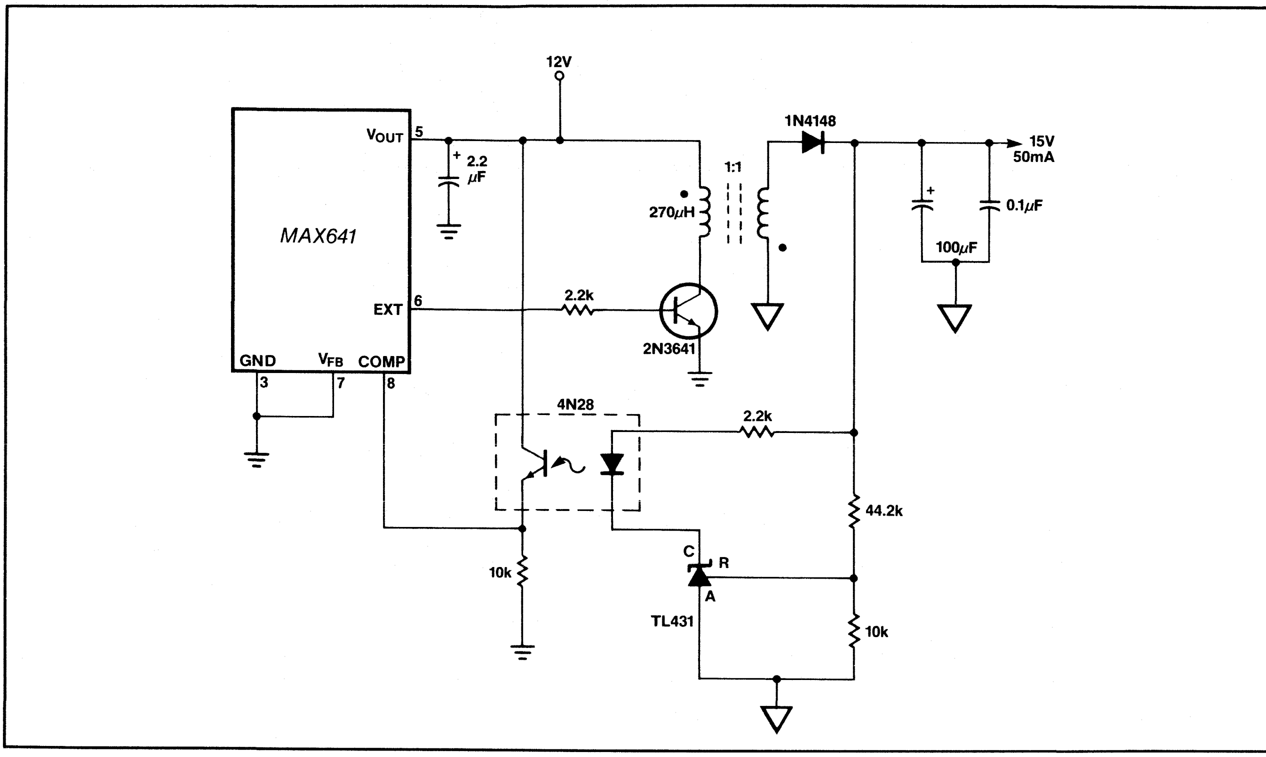
A small high voltage PNP transistor level shifts the a feedback signal from the +5V output down to the MAX641, whose ground pin (pin 3) is tied to the -50V input. The chip is biased this way so that EXT can directly drive an N-channel MOSFET to switch the inductor to -50V. This way the circuit operates much like a step-up DC-DC converter. The 330pF capacitor provides feedforward compensation to stabilized the regulator's control loop.

RS-232 Line to 5V



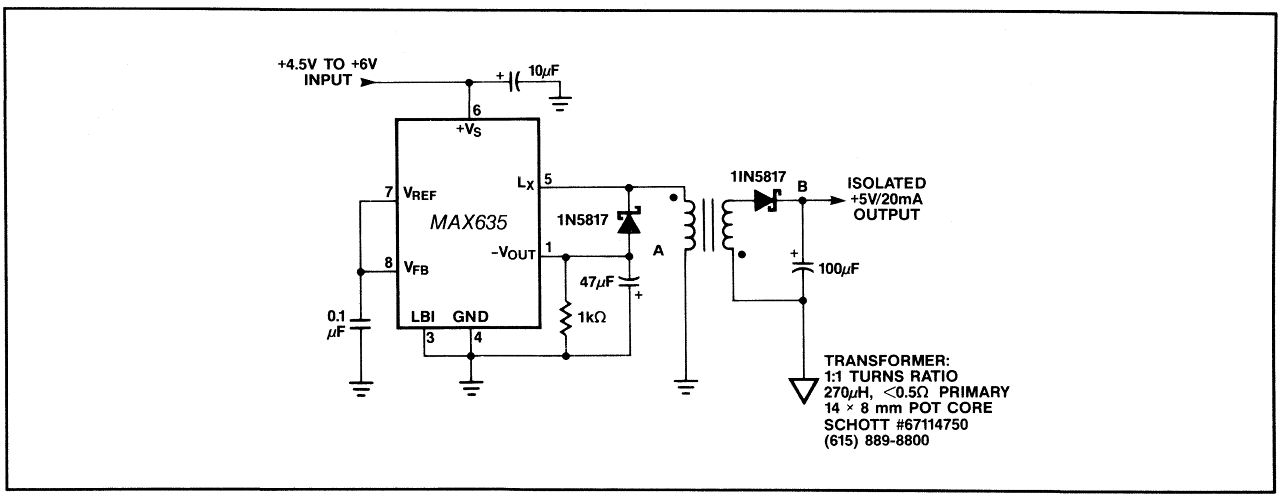
The MAX680 is normally used as a charge pump voltage converter which converts +5V to ±10V. In this circuit it works in reverse, converting RS-232 signal levels to a lower voltage. When the RS-232 inputs are driven by 1488 drivers powered by ±12V the output voltage varies from 5.3V open circuit to 4.5V with a 5mA output load.

Isolated +15V DC-DC Converter



In this circuit a TL431 shunt regulator is used to sense the output voltage. The TL431 drives the LED of a 4N28 opto-coupler which provides feedback to the MAX641 while maintaining isolation between the input +12V and the output +15V. In this circuit the +15V output is fully regulated with respect to both line and load changes.

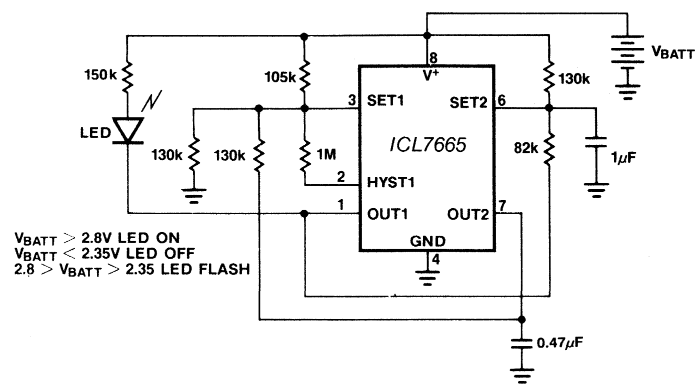
5V to Isolated 5V at 20mA



In this circuit a negative output voltage DC-DC converter generates a -5V output at point A. In order to generate -5V at point A, the primary of the transformer must flyback to a diode drop more negative than -5V. If the transformer has a tightly coupled 1/1 turns ratio, there will be 5V plus a diode drop across the secondary. The 1N5817 rectifies this secondary voltage to generate an isolated 5V output. The isolated output is not fully regulated since only the -5V at point A is sensed by the MAX635.

With careful selection of the transformer the 5V output will be within 10%. Bifilar winding of the transformer provides better load regulation of the isolated 5V output, but this does reduce the isolation by increasing the capacitance between the primary and secondary. The isolation voltage breakdown is determined by the characteristics of the transformer, not the MAX635.

3-STATE BATTERY INDICATOR



WHAT VALUE OF INDUCTOR? A GENERAL DISCUSSION

All the converter types referred to in this note are "flyback" converters. They operate by charging an inductor from a DC input and then discharging the inductor to generate a DC output that is greater than, less than, or of opposite polarity to the input.

The proper inductor value for any DC-DC converter depends on three things: the desired output power, the input voltage (or range of input voltage), and the converter's oscillator frequency and duty cycle. The oscillator timing is important because it determines how long the coil will be charged during each cycle. This, along with the input voltage, determine how much energy will be stored in the coil. The energy at a given instant is a function of the coil's current and inductance:

$E_L = LI_{pk}^2/2$ where

$I_{pk} = V_L t_{ON}/L$

The total power that can be put in (or taken out of) the coil is the energy (E_L) per cycle times the number of cycles per second.

$P_L = E_L f_O$

In the above equations, t_{ON} and f_O are usually interdependent. Most often the clock is a 50% duty cycle square wave so:

$t_{ON} = 1/(2f_O)$

The coil power as a function of input voltage, frequency (duty cycle = 50%), and inductance is:

$P_L = V_L^2/(8f_O L)$, or in terms of t_{ON}:

$P_L = V_L^2 t_{ON}/(4L)$

In step-up and inverting converters, the charging voltage for the coil (V_L) is usually the same as the input voltage (V_{IN}) if switch losses are ignored. In step-down converters, $V_L = V_{IN}-V_{OUT}$ (again ignoring losses) because the coil is connected between the input and output voltage when charging.

The above equations generally describe the design of MAX6XX DC-DC converter circuits. The following sections contain more specific design steps for each converter type.

STEP-UP REGULATOR DESIGN

(MAX630/631/632/633, MAX641/642/643)

First choose V_{OUT}, I_{OUT}, V_{IN}(min), and V_{IN}(max). Remember that in a step-up converter, V_{IN} must be less than V_{OUT}.

V_{IN}(min), and V_{IN}(max) cover the input voltage range, such as beginning and end of life battery voltages. The output power, P_{OUT}, is of course $V_{OUT} \times I_{OUT}$, but the converter also has to make up for losses in the inductor, LX switch, and catch diode. These losses typically add 10 to 25% to the required power.

In a step-up converter, power is supplied both via the inductor and directly from the input voltage. This is because one end of the coil remains connected to V_{IN} (Figure 1) as it supplies current to the output. Therefore:

1. $P_{OUT} = P_L + V_{IN} I_{OUT}$

The power, P_L, that we need from the inductor is then:

2. $P_L = (V_{OUT} - V_{IN} + V_D) I_{OUT}$

$V_D \times I_{OUT}$ accounts for losses in the catch diode. Schottky diodes minimize (1N5817) this loss which can be significant in low voltage circuits. In high voltage circuits (V_{OUT} = 10V and up) or if efficiency is not critical, signal diodes such as IN4148 perform well if their reverse voltage rating is not exceeded.

In order to get P_L out of the inductor, that much power must be put in. In an ideal system, i.e., minimal switch losses, Eq. 3 or 4 provides the power in the coil:

3. $P_L = V_L^2/(8f_O L)$, or in terms of t_{ON}:

4. $P_L = V_L^2 t_{ON}/(4L)$

Solving for the inductor value, L, and substituting Eq. 4 for P_L:

5. $L = V_{IN}^2/(8f_O I_{OUT}(V_{OUT}+V_D-V_{IN}))$

f_O is the DC-DC converter's clock frequency. Equation 5 assumes that f_O is a square wave with a 50% duty cycle. On the MAX630 and MAX4193, the clock frequency can be adjusted. It is preset to a fixed rate (50kHz) on the MAX631/32/33 and cannot be changed. In Equation 5, the *minimum* expected value should be used for V_{IN} to insure that there is adequate output power under all input conditions.

In a non-ideal system, the inductor voltage, V_L, and the input voltage, V_{IN}, are not quite the same, largely because of switch ON resistance (R). The peak inductor current will not be the expected value if this resistance is significant. Instead of $I_{pk} = V_L t_{ON}/L$, the expression for inductor current changes to:

6. $I_{pk} = V(1-e^{-Rton/L})/R$

The expression for output power then changes to:

7. $P_L = (V(1-e^{-Rton/L})/R)^2 f_O/2$

Besides inductance value, the selected coil must also be rated for the current that it must handle. The peak current that the coil sees is:

8. $I_{pk} = V_{IN} t_{ON}/L = V_{IN}/(2f_O L)$

t_{ON} is the coil charging time (for one clock cycle) which is equivalent to one half of one f_O clock period.

When calculating I_{pk} with Eq. 8, the largest expected value of V_{IN} ($V_{IN}(max)$) should be used so that the maximum current under all operating conditions will be considered. I_{pk} is then compared with the inductor current rating and the current rating of the L_X switch. I_{pk} should of course be less than these values.

If I_{pk} exceeds the peak current rating of the internal L_X switch in the MAX630/4193 (550mA) or MAX631/632/633 (475mA) then an external MOSFET or transistor with an adequate current rating must be used. The MAX641/642/643 works best in most such circuits since it is designed to directly drive an external FET.

STEP DOWN REGULATOR DESIGN

(MAX638, and MAX631/2/3 driving P MOSFET)

Choose V_{OUT}, I_{OUT}, $V_{IN}(min)$, and $V_{IN}(max)$. In step-down converters, $V_{IN}(min)$ must be greater than V_{OUT}.

$V_{IN}(min)$ and $V_{IN}(max)$ define the converter's input voltage range, such as the unregulated power supply voltage at high and low power line voltages. Output power is $V_{OUT}I_{OUT}$, but the converter must also be able to supply power to make up for losses in the inductor, Lx switch, and catch diode. If an 80% conversion efficiency is assumed, P_{OUT} is multiplied by 1.25 in the equations below.

In a step-down converter such as the MAX638, the output power is the sum of the power supplied via the coil and the power supplied directly from the input voltage. When the coil charges, it is connected between the input and output so that inductor charging current also flows into the load (Figure 2). When the coil discharges, current flows from ground, through the coil, into the load. Total output power is:

9. $P_{OUT} = P_L + V_{OUT}I_{pk}/4$

P_L is the power supplied by the coil and $V_{OUT}I_{pk}/4$ is the power supplied directly to the load while the coil charges. I_{pk} is the peak charging current of the Inductor. The above equation assumes that the coil charging current rises linearly from 0 to I_{pk} during one half of each oscillator cycle. The *average* coil charging current is then $I_{pk}/4$.

The peak inductor current is a function of the charging voltage (V_{IN}-V_{OUT}), charging time (t_{ON}), and coil inductance (L).

10. $I_{pk} = (V_{IN}-V_{OUT})t_{ON}/L$

11. $I_{pk} = (V_{IN}-V_{OUT})/(2f_oL)$

f_o is the converter's clock frequency. The coil can charge for at most one half of each clock cycle ($1/2f_o$). By substituting Equation 11 into Equation 9, we get:

12. $P_{OUT} = P_L + V_{OUT}(V_{OUT}-V_{IN})/(8f_oL)$

In order to get P_L out of the inductor we must put at least that much in. The power that is put in is:

13. $P_L = (V_{IN}-V_{OUT})^2/(8f_oL)$

By substituting Equation 13 into Equation 12, we get:

14. $P_{OUT} = V_{IN}(V_{IN}-V_{OUT})/(8f_oL)$

In terms of L, and by multiplying P_{OUT} by 1.25 to account for typical losses, we get:

15. $L = V_{IN}(V_{IN}-V_{OUT})/(10f_oP_o)$

The minimum expected value should be used for V_{IN} to ensure that there is adequate power under all conditions.

Besides inductance value, the selected coil must also be rated for the current that it will be handling in the circuit. The peak current that the coil will see is express by Equation 8. When calculating I_{pk}, $V_{IN}(max)$ should be used so that the maximum current under all operating conditions will be considered. I_{pk} is then compared with the current ratings of the inductor and L_X switch, and must be less than these values.

If I_{pk} exceeds the peak current rating of the internal L_X switch in the MAX638 (550mA) an external MOSFET or transistor, that is rated for the current, can be used with a MAX631/2/3 converter. Although they are called "step-up" regulators they can easily be configured for step-down circuits when using an external MOSFET. The MAX638 may also be used with an external transistor, but an inverter is also required.

INVERTING REGULATOR DESIGN

(MAX634/635/636)

Choose V_{OUT}, I_{OUT}, $V_{IN}(min)$, and $V_{IN}(max)$. In an inverting DC-DC converter, V_{IN} may be greater, equal, or less than V_{OUT}.

Output power is $V_{OUT}I_{OUT}$, but the converter has to supply additional power to make up for losses. These typically add 10 to 25% to the required power, depending on external conditons such as component selection and operating voltage. If we assume a conversion efficiency of 80%, the required power is multiplied by 1.25 in the initial design.

In an inverting converter (MAX634/635/636/637), all output power is supplied via the coil. One end of the coil remains grounded when it charges and discharges. The total output power is:

16. $P_{OUT} = P_L$

P_L is the power supplied by the coil. In order to get P_L out of the inductor, P_L must be put in:

17. $P_L = V_{IN}^2/(8f_oL)$

Solving for the inductor value and assuming 80% efficiency:

18. $L = V_{IN}^2/(10 f_O I_{OUT} V_{OUT})$

f_O is the DC-DC converter's clock frequency and assumes that f_O is a square wave with a 50% duty cycle. On the MAX634, the clock can be adjusted while the MAX635/636/637 has a preset oscillator that cannot be changed. In Equation 18, V_{IN}(min) should be used to ensure that there is adequate power under all conditions.

Besides inductance value, the selected coil must also be rated for the current that it must handle. The peak current that the coil sees is:

19. $I_{pk} = V_{IN} t_{ON}/L = V_{IN}/(2 f_O L)$

t_{ON} is the coil charging time (for one clock cycle) which is equivalent to one half of one f_O clock period.

When calculating I_{pk} with Equation 19, the largest expected value of V_{IN} (V_{IN}(max)) should be used so that the maximum current under all operating conditions will be considered. I_{pk} is then compared with the inductor current rating and the current rating of the L_X switch. I_{pk} should of course be less than these values. If I_{pk} exceeds the peak current rating of the internal L_X switch in the MAX634 (550mA) or MAX635/36/37 (475mA), an external MOSFET or transistor with an adequate current rating must be used.

TRANSFORMER DC-DC CONVERTER DESIGN

In designs which employ flyback transformers, step-up, step-down, or inverting converters are all built using the same basic architecture. As with other DC-DC designs, first choose V_{OUT}, I_{OUT}, V_{IN}(min), and V_{IN}(max). Remember that although the power to the load is $V_{OUT} I_{OUT}$, the converter has to supply 10% to 25% more power to make up for diode, switch and transformer losses. In flyback transformer circuits, all output power is supplied via the transformer so:

20. $P_{OUT} = P_T - V_D I_{OUT}$

P_T is the power supplied via the transformer and $V_D I_{OUT}$ is the power lost in the steering diode. In order to get P_T out of the transformer secondary, P_T must be put into the primary.

21. $P_T = V_{IN}^2/(8 f_O L_{PRI})$

Solving for the transformer's primary inductance:

22. $L_{PRI} = V_{IN}^2/(8 f_O I_{OUT}(V_{OUT}+V_D))$

f_O is the DC-DC converter's clock frequency and assumes that f_O is a square wave with a 50% duty cycle. The MAX634 and MAX630 allow the clock frequency to be adjusted while other devices have a preset oscillator (typically 50kHz) that cannot be changed. In Equation 22, V_{IN}(min) should be used for V_{IN} to insure that there is adequate power under all input conditions.

Besides inductance value, the transformer must also be rated for the peak current that it sees. The peak current is:

23. $I_{pk} = V_{IN} t_{ON}/L_{PRI} = V_{IN}/(2 f_O L)$

t_{ON} is the charging time (for one clock cycle) which is equivalent to one half of one f_O clock period.

When calculating I_{pk} with Equation 23, the largest expected value of V_{IN} (V_{IN}(max)) should be used so that the maximum current under all operating conditions will be considered. I_{pk} is then checked against the MAX6XX's L_X switch current rating and also determines the current handling requirements of the transformer. L_{PRI} and I_{pk} are then used for transformer selection or for core selection and the transformer design.

See "Pot Core and Transformer Basics", page 67, for information on winding transformers.

BENCHTOP SHORTCUT

The most direct means of checking I_{pk} is to measure it using an oscilloscope and current probe. If a current probe is not available, a less direct but still effective method is to observe the current by looking differentially across a small sense resistor placed in series with the inductor. 1 Ω usually does well, but a smaller value is necessary if currents over a few hundred milliamps are expected.

The following is a partial listing of sources for inductors, transformers, and cores for DC-DC converter magnetic components. It is by no means intended as a complete list.

INDUCTOR AND TRANSFORMER SUPPLIERS

AIE Magnetics 701 Murfreesboro Rd. Nashville, TN 37210 TEL 615-244-9024	AIE makes a variety of coils and transformers. Their full line catalog lists a variety of inductors. Catalog 5 lists many transformers designed for flyback converters up to 50 watts. Catalog 3 lists many coils types: slugs, toroids, pot cores, etc.
BH Electronics 12219 Wood Lake Dr. Burnsville, MN 55337 TEL 612-894-9590 FAX 612-894-9380	Coils and transformers from miniature surface mount up through 10's of watts.
Caddell-Burns 40 East Second Street Mineola, NY 11501 TEL 516-746-2310	Caddell-Burns has many standard coils in their 6860 and 7070 series which are suitable for use with Maxim's DC-DC converters. 7070 series coils have high current ratings for their size. Efficiencies that match toriods can be acheived.
Dale Electronics East Highway 50 Yankton, SD 57078 TEL 605-665-9301	Standard toroidal inductors in moulded PC mount packaging.
Gowanda Electronics Corp. 1 Industrial Place Gowanda, NY 14070 TEL 716-532-2234	Gowanda makes coils ranging from surface mount devices up through high power toroids, and also has many inductors wound on cylindrical ferrite bobbins. They make transformers in pot cores, E cores, and toroids.
J.W. Miller 19070 Reyes Ave. P.O. Box 5825 Rancho Dominguez, CA 90224 TEL 213-537 5200 FAX 213-631-4217	Miller has standard coils and toriods at low cost, however, avoid the tiny moulded chokes. The series resistance and current ratings of the small moulded parts are generally not adequate for DC-DC applications except in extremely low power applications (mA).
Nytronics Components Group, Inc. Orange Street Darlington, SC 29532 TEL 803-393-5421	Small moulded inductors which generally do not have adequate current capability for DC-DC circuits except in very low current applications (mA).
Pulse Engineering P.O. Box 12235 San Diego, CA 92112 TEL 619-268-2400 FAX 619-268-2515	Pulse Eng. supplies a line of low cost toroids, as well as mounted and moulded toroids. A toroid sampler kit, #845, is useful for prototyping designs. Custom transformers are also available.
Prem Magnetics 3521 North Chapel Hill Rd. McHenry, IL 60050 TEL 815-642-3763 TWX 910-642-3763	Prem supplies low cost, high current toroids.

Schott Corporation
Suite 108
1838 Elm Hill Pike
Nashville, TN 37210
TEL 615-889-8800
FAX 612-885-0834

Schott makes both miniature toroids, pot core inductors and transformers. Custom designs are available.

Torotel Products Inc.
13402 South 71 Highway
Grandview, MO 64030
TEL 816-761-6314
TWX 910-777-7037

Torotel specializes in toroidal inductors. Their general catalog lists hundreds of sizes and inductances. Mil spec versions are available.

POT-CORE AND TOROID-CORE SUPPLIERS

Allen-Bradley Magnetic Products
5900 N. Harrison St.
Shawnee, Oklahoma 74801
TEL 405-275-2100/TLX 796208

Publication MPCC contains information on both soft/linear ferrite and permanent magnetics. The key section is the "W5" section which provides info on their optimum material for DC-DC converters, as well as pot cores and toroids made from this material.

Arnold Engineering Company
300 West Street
Marengo, Illinois 60152
TEL 815-568-2000
TWX 910-642-2790

Arnold makes cores in both iron powder and molypermalloy powder (MPP).

Ferroxcube
5083 Kings Highway
Saugerties, NY 12477
TEL 914-246-2811
TWX 510-247-5410

Ferroxcube's Linear Ferrite Materials and Components catalog lists a wide variety of cores: toroids, pot cores, square cores, EP, EC, etc. The catalog has complete data on the cores and material, but little applications information.

Magnetics
Div. of Spang and Co.
900 E. Butler Rd.
P.O. Box 391
Butler, PA 16003
TEL 412-282-8282
TWX 710-373-3821

Magnetics has a complete line of magnetic materials: The ferrites and MPP cores are the most useful for DC-DC converters. A complete catalog/binder with both inductor design information and complete product data is available. All of the standard core shapes are available: pot cores, EE, EI, bobbin, toroids, etc. The catalog contains useful coil design information.

Micrometals, Inc.
1190 N. Hawk Circle
Anaheim, CA 92807
TEL 714-630-7420
TWX 910-591-1690

Low cost iron powder cores in toroids, E cores, and bars. These are lower cost than MPP or ferrite, but have lower efficiency.

Siemens Components Inc.
Special Products Division
186 Wood Avenue South
Iselin, NJ 08830
TEL 201-321-3400

The "Ferrite Cores and Hardware" short form catalog provides information on pot cores from 3.3x2.6mm to 41x25mm; toroids from 2.5mm to 34mm diameter, and a wide variety of other shapes such as E and RM.

Permag Inc., Regional offices:

Atlanta	404-448-4998
Boston	617-273-2890
Chicago	312-956-1140
Dallas	214-699-1121
Colorado	303-693-6612
Los Angeles	714-952-2091
Minneapolis	612-934-4635
San Francisco	408-738-1080
Toledo	419-385-4621
NY/LI	516-822-3311

Permag is a Distributor of Magnetic Cores for Allen- Bradley, Stackpole, Siemens, and Krystinel.

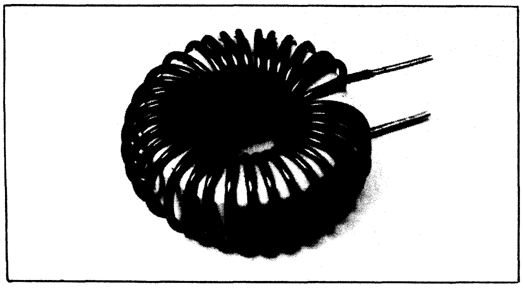
Cores Unlimited, Inc.
8311 Westminster Ave.
Suite 340F
Westminster, CA 92683
TEL 714-894-3062
800-772-CORE
FAX 714-895-4502

Cores Unlimited is an authorized distributor of magnetic cores for Hitatchi and Siemens. They also handle cores from several other manufacturers.

LOWEST COST INDUCTORS

NEW

ELECTRICAL DATA



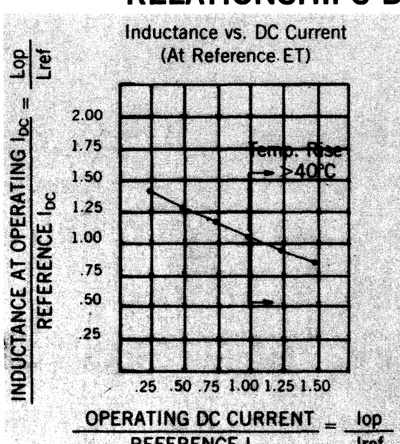
- SMPS AVERAGING FILTER (4)
- CHARACTERIZED FOR GENERAL PURPOSE USE, AND RIPPLE FILTERS
- SINGLE LAYER DESIGNS
- CAN BE USED AS DIFFERENTIAL MODE INDUCTORS IN EMI FILTERS (3)
- MOUNTING PACKAGE AVAILABLE ON REQUEST
- DESIGNER KIT AVAILABLE

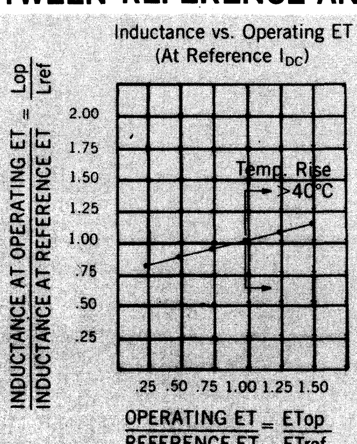
ELECTRICAL CHARACTERISTICS AT 25°C

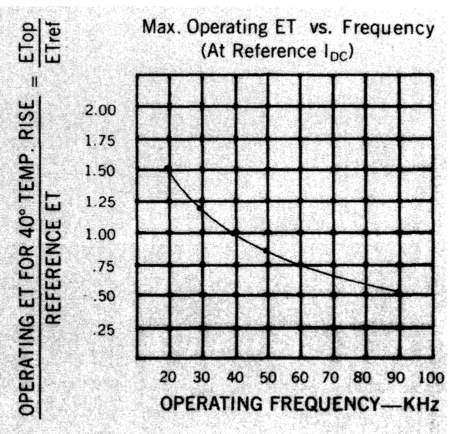
Reference Operating Values					Design Control Values (2)						
Part Number	Klip Mount Option	Inductance Typical (μ Hy)(1)	I_{DC} (AMPS)	ET_{op} (V-μSec)	Inductance No D.C. $^{+20\%}_{-12\%}$ (μ Hy)	1000 Hz Test Volts No D.C.	D C R (OHMS) Max	Coil Size Code	Klip Mount Package	Lead Dia. (In) $\pm.003$	Min. Energy Storage (μJ) (5)
51591		20	2.0	52	32.8	.0034	.06	H		.020	40
92100	K	25	2.5	30	20.7	.0023	.04	A	KM1	.020	75
92101	K	50	2.5	50	45.7	.0047	.07	B	KM2	.020	150
92102	K	100	2.5	90	94.1	.0094	.10	C	KM3	.020	300
92103	K	35	2.5	55	28.4	.0037	.04	B	KM2	.025	110
92104	K	70	3.0	85	61.0	.0076	.05	C	KM3	.025	300
92105	K	145	3.0	140	141.8	.015	.08	D	KM4	.025	650
92106	K	285	3.0	300	264.1	.035	.14	E	KM5	.025	1275
92107		450	3.0	425	436.3	.053	.20	F		.025	2000
92108	K	100	3.5	130	90.7	.012	.04	D	KM4	.032	600
92109	K	165	4.0	240	152.0	.027	.07	E	KM5	.032	1300
92110		270	4.0	350	263.9	.041	.10	F		.032	2150
92111	K	40	4.0	70	37.9	.006	.03	C	KM3	.032	300
51590		12	5.0	44	20.3	.0038	.03	G		.032	150
92112	K	100	5.0	200	90.7	.021	.04	E	KM5	.042	1250
92113		170	5.0	300	159.7	.032	.05	F		.042	2100
92114	K	55	5.0	100	54.9	.009	.02	D	KM4	.042	650
92115		95	7.0	225	96.0	.025	.03	F		.051	2300
92116	K	55	7.0	150	49.1	.015	.02	E	KM5	.051	1300
92117		55	10.0	175	55.9	.019	.02	F		.064	2750

See page 5 for Notes

RELATIONSHIPS BETWEEN REFERENCE AND OPERATING CONDITIONS







4

Pulse Engineering P.O. BOX 12235 SAN DIEGO, CA 92112 619-268-2400

QUALITY COILS SINCE 1946

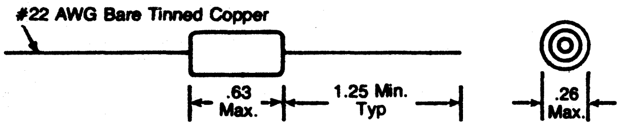
CADDELL-BURNS

258 East Second Street
Mineola, New York 11501-3508
Area Code 516 746-2310

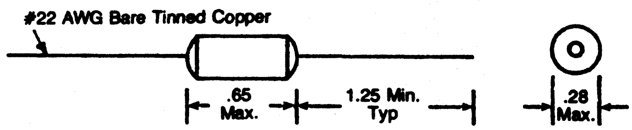
MINIATURE HIGH CURRENT CHOKES

1.0μH - 100mH. 10% Tolerance. Recommended Mounting Pitch — .80"

TYPE 7070



TYPE 7080 (EPOXY ENCAPSULATED VERSION)



NOTES: (FOR BOTH TYPES)

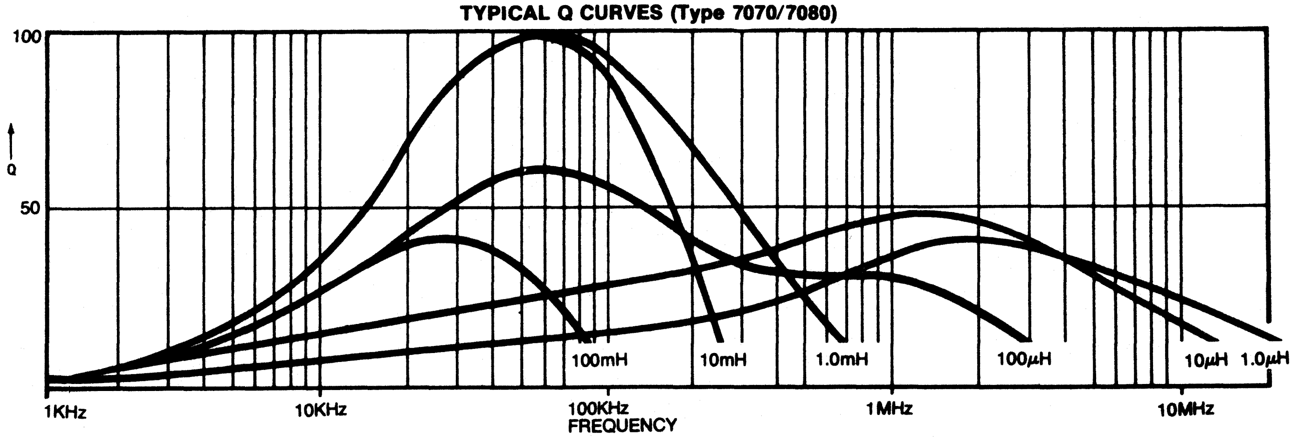
1. INDUCTANCE:
   For 1.0μH thru 8.2μH; effective inductance measured on Boonton 260A Q-meter at 7.9 MHz in accordance with MIL-C-15305.
   For 10μH thru 100mH; inductance (Ls) measured on General Radio 1650B Impedance Bridge at 1 KHz.
2. Current rating (Rated IDC) is based on 0.25 watt power dissipation for approximately 20°C temperature rise. Depending on the application, these units may be operated at up to twice the rated current.
3. Incremental current (INCR I) is the minimum current at which the inductance will be decreased by 5% from its initial (zero-DC) value.
4. Dielectric Withstanding Voltage — 1000 VRMS.
5. Operating temperature range −55° to +105°C.
6. Materials:
   Coil Form: Ferrite
   Cover: TYPE 7070 - PVC shrink tube-flame retardant UL type FR-1 per MIL-I-23053
   TYPE 7080 - Epoxy encapsulated.
   Magnet Wire: Per FED SPEC J-W-001177/9 (MIL-W-583 Type B)

STANDARD VALUES: (Electrical characteristics are identical for both types. Other values are available on special order.)

DASH NO.	NOMINAL INDUCTANCE	MAX.DCR OHMS	MIN.SRF μHz	RATED IDC ma	INCR I ma
-01	1.0μH	.010	155	5,000	6,400
-02	1.2	.011	148	4,800	6,000
-03	1.5	.012	128	4,600	5,000
-04	1.8	.013	120	4,400	4,500
-05	2.2	.014	108	4,200	4,100
-06	2.7	.015	100	4,100	3,600
-07	3.3	.016	96	4,000	3,200
-08	3.9	.017	90	3,800	3,000
-09	4.7	.022	85	3,400	2,700
-10	5.6	.028	76	3,000	2,500
-11	6.8	.031	70	2,800	2,300
-12	8.2	.035	51	2,700	2,000
-13	10	.038	35	2,600	1,900
-14	12	.043	21	2,400	1,700
-15	15	.049	14	2,300	1,500
-16	18	.054	10	2,200	1,400
-17	22	.059	8.0	2,100	1,300
-18	27	.070	6.5	1,900	1,100
-19	33	.077	6.1	1,800	1,000
-20	39	.084	5.7	1,700	940
-21	47	.093	5.1	1,600	870
-22	56	.12	4.3	1,500	790
-23	68	.13	3.6	1,400	710
-24	82	.16	3.2	1,300	650
-25	100	.24	3.0	1,000	590
-26	120	.32	2.7	880	540
-27	150	.43	2.1	760	480
-28	180	.48	1.7	720	440
-29	220	.55	1.6	670	400
-30	270	.62	1.5	640	360

DASH NO.	NOMINAL INDUCTANCE	MAX.DCR OHMS	MIN.SRF μHz	RATED IDC ma	INCR I ma
-31	330μH	.72	1.4	590	320
-32	390	.79	1.3	560	300
-33	470	.88	1.2	530	270
-34	560	1.2	1.1	460	250
-35	680	1.5	1.0	410	230
-36	820	1.7	.96	380	210
-37	1.0mH	1.9	.88	360	190
-38	1.2	2.4	.78	320	170
-39	1.5	2.8	.64	300	150
-40	1.8	3.1	.60	280	140
-41	2.2	4.5	.54	240	120
-42	2.7	5.8	.44	210	110
-43	3.3	8.1	.43	180	100
-44	3.9	8.9	.40	170	95
-45	4.7	10	.38	160	86
-46	5.6	11	.35	150	79
-47	6.8	15	.29	130	72
-48	8.2	17	.26	120	65
-49	10	22	.24	110	59
-50	12	26	.23	100	54
-51	15	34	.19	86	48
-52	18	39	.17	80	44
-53	22	54	.16	68	40
-54	27	62	.15	64	36
-55	33	82	.12	55	32
-56	39	93	.11	52	30
-57	47	120	.096	46	27
-58	56	130	.092	44	25
-59	68	190	.088	36	23
-60	82	210	.080	35	21
-61	100	270	.074	30	19

TYPICAL Q CURVES (Type 7070/7080)



QUALITY COILS SINCE 1946

CADDELL-BURNS

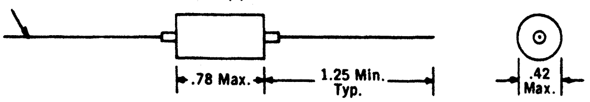
258 East Second Street
Mineola, New York 11501-3508
Area Code 516 746-2310

HIGH CURRENT CHOKES

10μH-1.0H. 10% Tolerance. Recommended Mounting Pitch — 1.25"

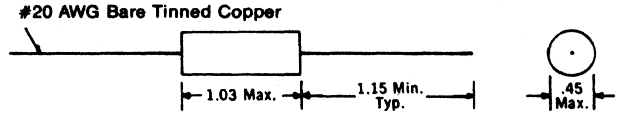
TYPE 6860

#20 AWG Bare Tinned Copper



TYPE 6870 (EPOXY ENCAPSULATED VERSION)

#20 AWG Bare Tinned Copper



NOTES: (FOR BOTH TYPES)

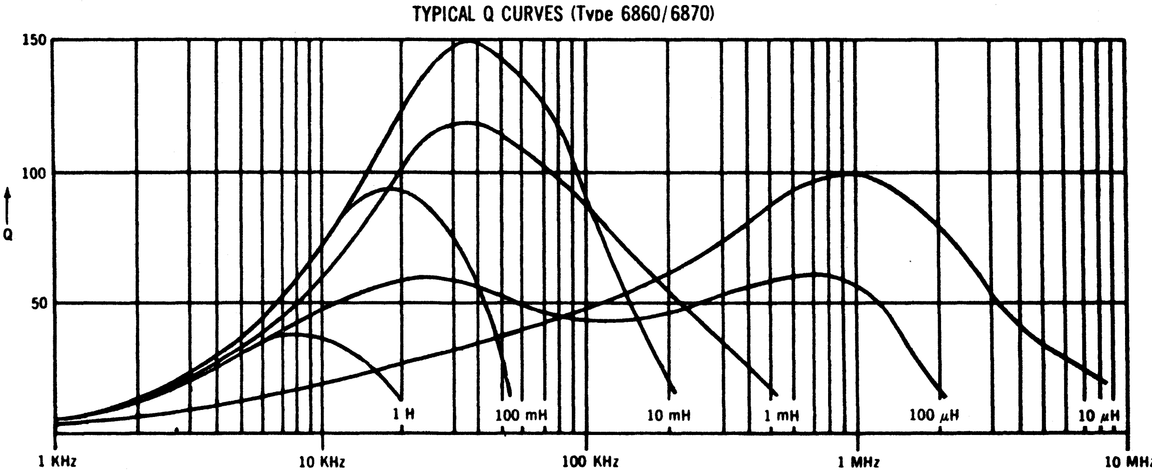
1. Inductance (Ls) measured on General Radio 1650B Impedance Bridge at 1 KHz.
2. Current rating (Rated IDC) is based on 0.5 watt power dissipation for approximately 20°C temperature rise. Depending on the application, these units may be operated at up to twice the rated current.
3. Incremental current (INCR I) is the minimum current at which the inductance will be decreased by 5% from its initial (zero-DC) value.
4. Dielectric Withstanding Voltage — 1000 VRMS.
5. Operating temperature range — 55° to + 105°C.
6. Materials:
   Coil Form: Ferrite.
   Cover: TYPE 6860 - Polyester alpha-cellulose.
   Coating: Polyurethane per MIL-1-46058.
   TYPE 6870- Cover & Coating: Epoxy encapsulated.
   Magnet Wire: Per FED SPEC J-W-001177/9 (MIL-W-583 Type B)

STANDARD VALUES: (Electrical characteristics are identical for both types. Other values are available on special order.)

Dash No.	Nominal Inductance	Max. DCR Ohms	Min. SRF MHz	Rated IDC ma	INCR I ma
—01	10 μH	.023	45	4600	3200
—02	12	.025	40	4500	2900
—03	15	.030	32	4100	2600
—04	18	.032	21	3900	2400
—05	22	.035	12	3700	2200
—06	27	.038	8.5	3600	2000
—07	33	.043	5.8	3400	1800
—08	39	.047	3.5	3200	1700
—09	47	.054	3.2	3000	1500
—10	56	.060	2.9	2900	1400
—11	68	.068	2.7	2700	1200
—12	82	.073	2.5	2600	1100
—13	100	.098	2.3	2300	1000
—14	120	.14	2.1	1900	930
—15	150	.18	1.9	1700	830
—16	180	.20	1.5	1600	760
—17	220	.28	1.3	1400	680
—18	270	.31	1.3	1300	620
—19	330	.35	1.2	1200	560
—20	390	.38	1.1	1100	510
—21	470	.44	1.0	1050	460
—22	560	.48	.90	1000	430
—23	680	.63	.80	890	390
—24	820	.87	.72	760	350
—25	1.0 mH	.96	.65	720	320
—26	1.2	1.3	.62	620	290
—27	1.5	1.4	.58	600	260
—28	1.8	1.7	.53	540	240
—29	2.2	2.3	.44	470	220
—30	2.7	2.6	.39	440	190
—31	3.3	3.5	.36	380	180

Dash No.	Nominal Inductance	Max. DCR Ohms	Min. SRF MHz	Rated IDC ma	INCR I ma
—32	3.9 mH	3.8	.34	360	160
—33	4.7	4.3	.32	340	150
—34	5.6	5.6	.26	300	140
—35	6.8	6.3	.24	280	120
—36	8.2	8.6	.21	240	110
—37	10	9.7	.20	220	100
—38	12	11	.19	210	92
—39	15	15	.17	180	84
—40	18	20	.14	160	75
—41	22	24	.13	140	68
—42	27	26	.12	130	.62
—43	33	35	.10	120	56
—44	39	38	.95	110	51
—45	47	50	.80	100	47
—46	56	55	.72	95	43
—47	68	76	.66	81	39
—48	82	86	.62	76	35
—49	100	99	.57	71	32
—50	120	110.	.52	67	29
—51	150	200	.47	50	26
—52	180	220	.43	48	24
—53	220	300	.38	41	22
—54	270	320	.35	40	20
—55	330	420	.32	35	18
—56	390	480	.29	32	16
—57	470	670	.26	27	15
—58	560	730	.24	26	14
—59	680	870	.19	24	12
—60	820	950	.18	23	11
—61	1.0 H	1100	.17	21	10

TYPICAL Q CURVES (Type 6860/6870)



FREQUENCY

Representative Power Mosfets
(N-Channel)

Part Number	Pkg	R_{ON} (@) at (I_{DS}, V_{GS} = xV)	Volts (max)	Mfg**
BUZ71A	TO-220	0.12 (6A, V_{GS}=10V)	50	MOT/SI/SM
BUZ21	TO-220	0.1 (9A, V_{GS}=5V)*	100	MOT/SI/SM
		0.2 (5A, V_{GS}=5V)*		
IRF513	TO-220	0.8 (2A, V_{GS}=10V)*	100	IR/SI/GE/MOT
		1.2 (1A, V_{GS}=5V)*		
IRF530	TO-220	0.18 (8A, V_{GS}=10V*)	100	IR/SI/GE/MOT
		2.0 (4A, V_{GS}=5V)*		
IRF540	TO-220	0.085 (8A, V_{GS}=10V*)	100	IR/SI/GE/MOT
		0.1 (5A, V_{GS}=5V)*		
IRF620	TO-220	0.8 (2.5A, V_{GS}=10V*)	200	IR/SI/GE/MOT
		1.3 (2.5A, V_{GS}=5V)*		
IRF640	TO-220	0.18 (10A, V_{GS}=10V*)	200	IR/SI/GE/MOT
IRFD121	4p DIP	0.3 (1.3A, V_{GS}=10V*)	60	IR/GE
RFP12NO8L	TO-220	0.2 (1.3A, V_{GS}=5V*)	80	RCA

(P-Channel)

Part Number	Pkg	R_{ON} (@) at (I_{DS}, V_{GS} = xV)	Volts (max)	Mfg**
IRFD9120	4p DIP	0.6 (1.3A, V_{GS}=-10V*)	-100	IR/GE
IRF9520	TO-220	0.6 (3.5A, V_{GS}=-10V*)	-100	IR
IRF9540	TO-220	0.2 (10A, V_{GS}=-10V*)	-100	IR
IRF9543	TO-220	0.3 (10A, V_{GS}=-10V*)	-60	IR
IRF9620	TO-220	1.5 (3.5A, V_{GS}=-10V*)	-200	IR
RFP6P08	TO-220	0.6 (V_{GS}=-10V*)	-100	RCA
MTP2P45	TO-220	6 (1A, V_{GS}=-10V*)	-450	MOT
MTP8P08	TO-220	0.4 (4A, V_{GS}=-10V*)	-80	MOT

* Typical specification, not guaranteed by manufacturer.
** Manufacturer code: IR = International Rectifier, SI = Siliconix, MOT = Motorola, GE = General Electric, SM = Siemens, RCA = RCA

Rectifiers for DC-DC Converter Circuits

Part Number	I_{AVG} (amps)	V_F (volts)	Prv (volts)	Type	Pkg
1N914	0.05	1.0	75	Silicon	Glass
1N4148	0.05	1.0	75	Silicon	Glass
1N4935	1	1.2	200	Silicon	Plastic
1N5817	1	0.45	20	Schottky	Plastic
1N5818	1	0.55	30	Schottky	Plastic
1N5820	3	0.475	20	Schottky	Plastic
1N5821	3	0.5	30	Schottky	Plastic
1N5822	3	0.525	40	Schottky	Plastic
1N5823	5	0.36	20	Schottky	Metal
1N5824	5	0.37	30	Schottky	Metal
1N5825	5	0.38	40	Schottky	Metal
UDS620	6	0.48	20	Schottky	TO-220
UDS640	6	0.48	40	Schottky	TO-220
UES1001	1	0.895	50	Silicon	Bead
UES1002	1	0.895	100	Silicon	Bead
UES1003	1	0.895	150	Silicon	Bead

SUMMARY OF BATTERY CHEMISTRIES

Performance of Primary (Non-Rechargeable) Battery Chemistries

Chemistry Anode/Cathode/Electrolyte	Open Circuit V	Loaded V	Capacity (Ah at 100 hr rate)	Energy (Watts-hr)	Weight (grams)	Energy Density Wh/kg	Energy Density Wh/Liter
Alkaline $Zn/MnO_2/KOH$	1.58	1.5-1.1	14	17	132	125	315
Carbon Zinc (LeClanche) $Zn/MnO_2/NH_4CL$, $ZnCL_2$	1.55	1.5-1.0	4.5	5.4	95	55	100
Mercury $Zn/HgO/KOH$	1.35	1.3	15	18	166	100	450
Lithium Sulfur Dioxide $Li/SO_2/LiBr$	3.0	2.8-2.7	8	22	95	275	440
Lithium Thionyl Chloride $Li/SOCL_2/LiALCL_4$	3.6	3.5-3/4	14	45	113	375	850
*Lithium poly-carbonmonoflouride nLi/(CF) n/α-butyrolactone	3.0	2.6	10	26	94	275	500
*Lithium Manganese Dioxide Li/MnO_2	3.25	3.0-2.0	14	26	130	230	500

* Extrapolated from data for smaller cells.

Secondary or Rechargable Chemistries

Chemistry Anode/Cathode/Electrolyte	Open Circuit V	Loaded V	Capacity (Ah at 100 hr rate)	Energy (Watts-hr)	Weight (grams)	Energy Density Wh/kg	Energy Density Wh/Liter
Nickel Cadmium $Ni/Cd/KOH$	1.35	1.25	4.4	5.3	140	40	120
Lead Acid $PbO_2/Pb/H_2SO_4$	2.1	2.0	2.7	5.4	182	30	100

PRIMARY BATTERIES

Dry Cells

There are three popular cell types based on zinc and carbon: the standard LeClanche dry cell, the "heavy duty" zinc chloride cell, and the alkaline manganese cell. As shown in **Table** 1, there is about a 3 to 1 difference in capacity between the alkaline and LeClanche at the 100 hour discharge rate. The zinc chloride battery falls about midway between the other two. The most significant difference between the characteristics of these three chemistries is their response to high load currents, and their response to continuous versus pulsed applications.

In very low drain applications such as 1mA from a D cell, the LeClanche cell has about 1/2 the capacity of the alkaline (7Ah vs. 16Ah). When the load current is increased to 80mA, the duty cycle begins to play an important role. The alkaline battery loses virtually no capacity when operated continuously at 80mA, but the LeClanche cell capacity drops from 4Ah to 2.3Ah as the duty cycle of the 80mA load is increased from 1h/day to 24h/day. The zinc chloride cell has 6.2Ah and 6.0Ah capacity under 80mA load, 1h/day and 24h/day respectively.

Summary: The alkaline cell has about twice the capacity of a LeClanche under very light loads, and is over 6 times better under continuous duty high current loads.

Mercury

This cell has a very stable output voltage of about 1.35V, except for those cells which have a small amount of manganese dioxide added, which raises the open circuit voltage to 1.4V. It has good storage characteristics, retaining 85% to 90% capacity after two years at 20°C, and about 80% after one year of storage at 45°C. Mercury cells are generally not recommended for use below 0^oC. The "9V" size 6 cell mercury battery has an output voltage of 7.5 to 8.0V for virtually all of its life, and has about 575mAh capacity. The capacity of mercury batteries is nearly independent of duty cycle, except at very high discharge rates. It is generally more expensive than alkaline cells, which have nearly the same performance, except that the output voltage of alkaline cells changes more as the cell is discharged.

Silver Oxide

Primarily used for miniature button and coin cells such as those used in watches and hearing aids, the silver oxide cell has a flat 1.55V discharge. They are best for light loads of C/50 or less. It retains about 70% of its capacity when operated at 0^oC, and about 30% at -20^oC.

Lithium Manganese Dioxide ($LiMnO_2$)

There are two basic types of $LiMnO_2$ cells. A pressed powder cathode is used in low discharge rate cells such as the "coin" cells. Flat and cylindrical cells with higher discharge current ratings use a thin pasted electrode on a supporting grid structure. The open circuit voltage is slightly higher than 3.0V. The output voltage is relatively flat at 2.8V to 3.0V up to the last 20% of the discharge period. Like most lithium chemistries, Li/MnO_2 cells have very good storage characteristics, with about 85% of capacity remaining after 6 years storage at 20^oC.

This cell chemistry is available in sizes from 30mAh to over 1Ah. $LiMnO_2$ cells do not exhibit the voltage delay phenomena sometimes seen in $LiSO_2$ and $LiSOCl_2$ cells.

Lithium Sulfur Dioxide ($LiSO_2$)

This is the most mature of the high energy density, high discharge rate lithium chemistries and has been extensively used in military applications such as sonobouys, radios, and beacons. Early $LiSO_2$ had some safety problems, but manufacturers now extensively test their designs. Indeed, many data sheets have more information about crush tests, nail-through-the-battery tests, and incineration tests than information about discharge characteristics.

Large $LiSO_2$ cells with high discharge rate capability have safety devices such as high temperature fusible links, over-pressure safety vents, and time delay fuses to prevent catastrophic rupturing (battery manufacturers apparently do not like the term explosion).

The capacity of $LiSO_2$ cells is about 25% less than its major competitor, lithium thionyl-chloride ($LiSOCl_2$). $LiSO_2$ cells have excellent storage characteristics, and maintain a high percentage of capacity over a wide temperature range. Typical cells sizes range from 300mAh upwards to many tens of ampere hours. The open circuit voltage is about 3.0V, and the voltage during discharge is relatively flat.

Lithium thionyl-chloride ($LiSOCl_2$)

Presently the highest energy density system available, $LiSOCl_2$ is the choice over $LiSO_2$ in many new applications. In general, $LiSOCl_2$ cells have about 1/3 more capacity for a given volume or weight than do $LiSO_2$ cells, but energy densities vary slightly between manufacturers.

Typical cell capacities range from several hundred mAh through 20Ah, but there are cells up to 8,000 ampere hours available. See "Lithium cells suit high-energy military needs", Don Powers, EDN April 85. The open circuit voltage is 3.9V, and the discharge voltage is a constant 3.5 to 3.9V, depending on the discharge rate. High discharge rate D size cells can be operated up to 10A.

Lithium poly-carbonmonofluoride ($LiCF_n$)

These are available in both coin cells and cylindrical cells up to 1.2Ah. $LiCF_n$ cells are well suited for CMOS RAM backup applications since the self discharge rate is typically only 0.5% per year of storage at room temperature. The cylindrical cells have a high pulse current capability of 1A from a 1.2Ah 2/3A cell (17mm diameter by 33.5mm high, 13.5 grams).

SECONDARY BATTERIES

The two most popular rechargeable chemistries are lead acid and nickel cadmium.

Nickel cadmium

Nickel Cadmium, or NiCad, cells and batteries are available in sizes from about 10mAh to over 10Ah, with a wide variety of packaging styles. 10mAh to 150mAh batteries designed for memory backup are available with pins for direct mounting to printed circuit board, with up to 5 seconds exposure to a solder wave being allowed. Batteries are also available in consumer outlets for the standard AA, C, D, and 9V sizes. The consumer C and D batteries are really repackaged "sub-C" or C_s cells, and both have the same 1.2Ah rating as industrial grade sub-C cells.

Industrial grade C cells have around 2.5Ah capacity, and D cells about 4Ah to 4.8Ah. 9V NiCad batteries come in both 6 cell and 7 cell versions, with nominal terminal voltages of 7.2V and 8.4V respectively. Since the output discharge characteristics is very flat, even a 7.2V output voltage is in most applications an adequate replacement for the 9.5 to 6V output of a 6 cell LeClanche or alkaline battery. NiCad 9V batteries, though, have only 65mAh to 100mAh capacity, much lower than the 500+mAh ratings of alkaline 9V batteries.

Sealed C and D cells are typically recharged at the 0.1C rate for 14 hours. Ni-Cad batteries in memory backup applications are usually continuously current trickle-charged at 0.002C to 0.10C (where C is the cell's rated capacity). Standard cells are normally charged at 0.1C for 14 hours, since this allows a very simple charging circuit --- batteries can be over-charged at 0.1C for for extended periods without drastic reductions in life. Special, fast rate rate batteries can be recharged in 20 minutes with a charge rate of 4C, but high rate chargers must have sophisticated methods of detecting end-of-charge, such as sensing the increase in temperature that occurs when full charge has been reached. Other methods of terminating fast-charge include temperature compensated voltage detection, and detection of the voltage reduction that occurs after full charge has occurred.

Cell capacity is typically reduced to 50% of its initial value after 500 to 1500 deep discharge cycles, or after 5 or more years of continuous trickle charge at 0.005C. NiCd batteries have a self-discharge rate of about 0.5% per day at 23°C when near full charge, and retain 20% to 40% capacity after 5 months. At 30°C, 50% to 80% charge is retained after 30 days.

The open circuit voltage of a fully charged Ni-Cd cell is about 1.3V, and the discharge voltage is 1.24V +/- 100mV for virtually the entire discharge cycle. After the cell voltage has fallen to around 1.1V, the rate a change of the cell voltage increases rapidly. Cells may be completely discharged without damage, but reverse charging at more than -200mV for extended periods will permanently damage cells. This reverse charge condition most often occurs in multi-cell batteries in which the capacity of the various cells is not well matched.

NiCad batteries come in both sealed and vented types. A sealed cell is a closed environment, and allows the escape of gas only under abnormal conditions which cause safety vents to open. The vented cell allows gases to escape from the cell during normal operation. Sealed cells are most common in the sizes from D cell and below, while vented cells are used in very large batteries in such applications as engine starting and mobile X-ray equipment. Vented cells must be periodically inspected and the elecrolyte replenished.

Lead Acid Batteries

Lead acid cells come in many forms, the most common one being the automobile battery, which is typically a vented lead-acid battery. Smaller cells, such as D and X come in fully sealed packages. Typical sealed lead-acid battery capacity is 2.5Ah (10hr rate) for D cells, and 5Ah for X cells. The open circuit voltage of a lead-acid cell can be used to estimate the charge state. For a Cyclon brand cell from Gates, the 25°C open circuit voltage battery is about 2.18V when fully charged, declining linearly to about 1.98V when 10% of capacity is left.

The minimum voltage allowed during discharge is normally 1.6V, and lead-acid batteries should not be allowed to self-discharge below 1.8V. If allowed to self discharged below 1.8V while in storage, the battery will take longer than normal to recharge and the next discharge cycle cannot deliver the rated capacity. Subsequent cycles, however, will result in an increase in capacity to the rated capacity. As with all batteraies, the rate of self discharge is a strong function of temperature. After 5 months of stoarage at 20°C, the typical battery will have 60% capacity remaining. At 40°C, however, the battery would be fully discharged after 5 months.

Lead-acid batteries have extremely low internal impedance, and D cell can deliver up to 100A for short periods of time.

Since the terminal voltage of a lead-acid battery rises sharply as it nears 100% charge, the most common charge circuit is a simple constant voltage supply. A constant 2.35V charger will recharge a battery to 90% within 2 hours, and the battery can be left float charging at 2.35V indefinitely to maintain full charge. If the battery temperature will vary significantly, the float voltage should have a temperature coefficient of -2.5mV/°C/cell. Typical float life is 5 to 8 years at 25°C and 2.35V float voltage, decreasing to 2 years for 2.35V float voltage at 50°C.

For deep discharge service that need a fast cycle time, the charge voltage can be increased to 2.45V to 2.7V for even faster charging, but continuous float charging at more than 2.4V is not recommended. Typical cycle life is 2500 for full discharge, and over 1000 for a 25% depth-of-discharge. It is important to fully recharge lead-acid cells, with a 2.35V optimum for batteries that are cycles once per week, and 2.45V for batteries that are cycled once per day.

ARTICLE REPRINTS - Chronological Order

Power supply chips apply their many talents to running 5-V systems

Three ac-dc converter chips pack enough resources, including on-chip bridge rectifiers, to create accurate 5-V sources that slip into all systems.

Monolithic power converters appeared on the scene only recently, but few are so complete that they carry all related circuitry—including rectifiers and regulators—on chip. Usually, a supply working with an input of 117 V ac demands five or more external components to convert that input into 5 V dc. And if the IC is designed for isolated outputs, it typically needs a regulator, a bridge rectifier, and the ubiquitous transformer. Obviously, the benefits of a chip's size can be tempered quickly by the design considerations and the cost of the external circuitry.

Into the fray come three ac-dc converter chips that simplify and streamline the design of power supplies. They turn 117 or 220 V ac into 5 V dc, with an output current of 0.5 to 50 mA. All three distinguish themselves as the first supply chips to incorporate full- or half-wave rectifiers, circuitry that is complemented by an internal series-pass regulator and built-in overvoltage and undervoltage detectors. Furthermore, each one targets a different application area: The MAX610 sets itself up as a simple line-powered 5-V supply, the MAX611 is designed for systems that control triacs, and the MAX612 uses an 8-V rms transformer to produce an isolated 5-V output. Each one delivers its output to within ±5% over the full temperature range and with load and line variations.

Charles Allen, Maxim Integrated Products Inc.

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Individually, the three chips are impressive (Fig. 1). The MAX610 includes a full-wave bridge rectifier and a 12.4-V zener and needs only two capacitors and a resistor to generate 5 V dc. Though its output voltage is internally preset to 5 V, some simple rewiring of the voltage-setting pin, V_{set}, can be used to select an output of 1.3 to 9 V.

A different look

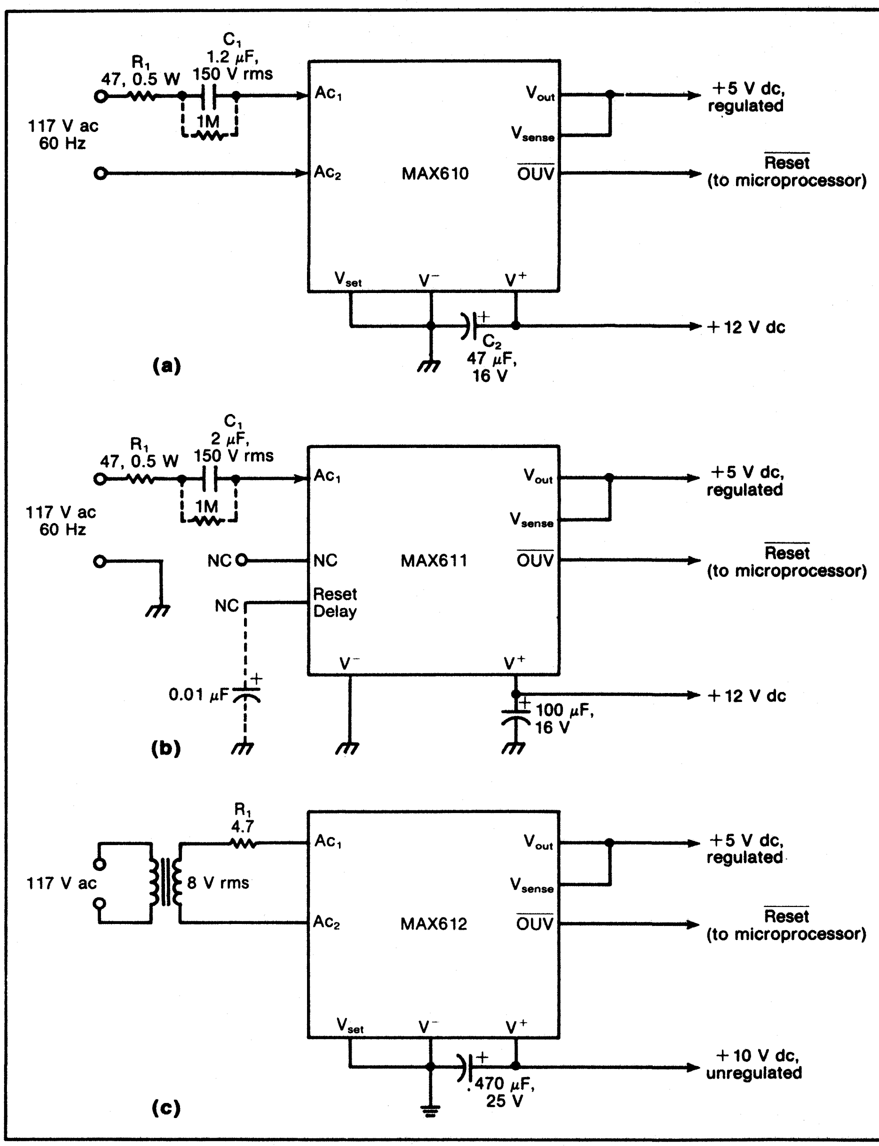
The second chip, the MAX611, replaces V_{set} with the Reset Delay pin, which it uses to set the delay of a monostable timer controlled by the Overvoltage/Undervoltage Detection pin, $\overline{OUV}$. Whenever the chip's output voltage goes out of bounds, $\overline{OUV}$ goes low and stays in that state until the output remains within 7% of the fixed 5-V output for a certain timeout period, which is determined by an external capacitor. Unlike the MAX610, it carries a half-wave rectifier, which enables the output voltage to be referenced directly to one side of the power line, simplifying connections with triacs.

Finally, the MAX612 is intended for isolated power supplies using an 8-V rms transformer. It contains an 18.6-V zener and a full-wave bridge rectifier. Its output voltage can be fixed at 5 V or adjusted for 1.3 to 15 V, the latter through rewiring of the V_{set} pin and with resistive voltage dividers.

The chips' undervoltage/overvoltage detector is set to 5 V ±7%. Whenever the output voltage falls below 4.65 V or climbs above 5.35 V, $\overline{OUV}$ goes low, indicating a fault. Whereas the MAX610 and MAX612 supply ICs use only $\overline{OUV}$, the MAX611 chip adds the reset delay section, which stretches the $\overline{OUV}$ output and in essence acts like a retriggerable monostable circuit.

Whenever the output voltage drops below 4.65 V, an external delay capacitor, optionally connected to the chip's Reset Delay pin (RD), discharges rapidly, sending $\overline{OUV}$ low (Fig. 2). When the regulated output returns to 5 V, the delay capacitor recharges slowly; when its voltage finally hits 8 V, $\overline{OUV}$ goes high. The slow recharging extends the chip's $\overline{OUV}$ fault indication by 30 ms for every 0.01 μF of capacitance connected to RD. When $\overline{OUV}$ is connected to a microprocessor's $\overline{Reset}$ input, that delay provides a sufficiently wide pulse for resetting at power-up and during brownouts and short power interruptions.

The proprietary voltage-setting circuitry bears close examination. When the V_{set} pin is grounded, it selects the preset output of 5 V; on the other hand, connecting an external resistive voltage divider be-



1. The MAX610 power supply, the simplest of a series, aims directly at line-powered 5-V supplies (a). The MAX611 works with triac control systems; an optional capacitor enables the chip to stretch a reset pulse to a microprocessor (b). The MAX612 uses a transformer to produce an isolated output (c).

tween the pin and the output enables the output to be adjusted between 1.3 and 9 V for the MAX610 and between 1.3 and 15 V for the MAX612.

The voltage at the V_{set} input selects the feedback source for the loop error amplifier (Fig. 3). When V_{set} is grounded and the input voltage is less than 100 mV, a comparator connected directly to the pin selects an internal divider string—a fusible link, trimmed for an output of 5 V ±4%—as the feedback source for the error amplifier. When the input voltage rises above 100 mV, the comparator selects the external divider as the input to the error amplifier. All comparisons are based on an internal 1.3-V bandgap reference. The output voltage is figured as 1.3 V × $(1+R_1/R_2)$, where R_1 and R_2 make up the resistive divider between the output and ground.

As opposed to connecting an internal resistor string to V_{set}, this analog feedback arrangement keeps the input current to 10 nA, meaning that full accuracy can be obtained with high-value resistor feedback strings.

Taking that first step

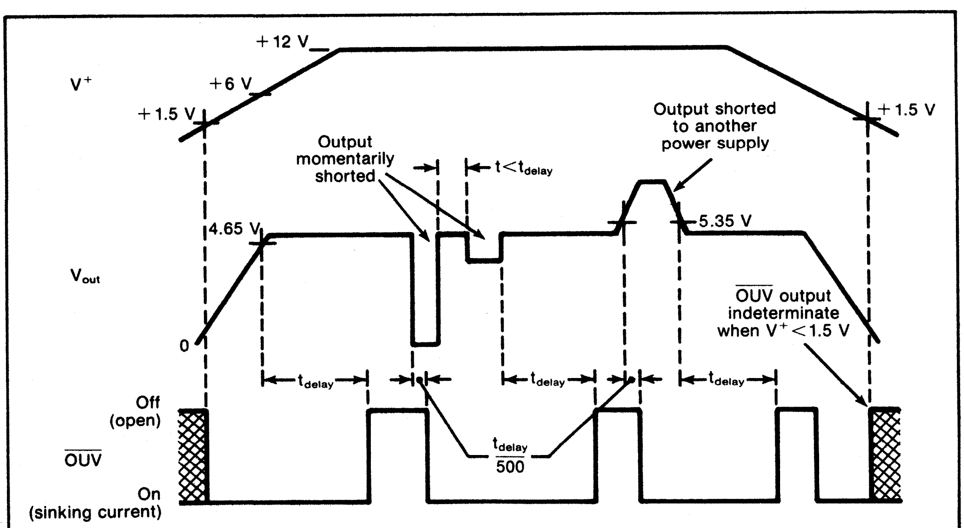
The three chips do their jobs in a relatively straightforward manner. First, they pass the ac line voltage—117 or 220 V—through a capacitor to limit its current. Second, they rectify the voltage, with the built-in zener shunt-regulating it to 12 V dc. Finally, the internal series-pass regulator lowers that voltage to 5 V dc. The current limit of the regulator can be programmed by an optional current-sensing resistor, placed on the voltage output line.

As the current-limiting capacitor, C_1 is critical to the performance of all three chips. Its value must be large enough to supply both the load current and the operating current of the chip at hand. Moreover, because nearly all of the ac line voltage appears across it, it cannot be an electrolytic, and it must be rated for at least 150 V rms or 210 V pk-pk. Polyester and polypropylene film capacitors, rated for continuous operation at power line voltages, can fill the bill.

The limiting resistor, R_1, is far less critical. Usually rated at 47 Ω and 0.5 W, it limits the in-rush current to 3.5 A peak. That current appears when the circuit connects to the 117-V ac line just as the latter's voltage approaches the maximum of 165 V. In fact, the bridge rectifier and zener diode within the three chips can withstand a nonrepetitive peak in-rush current of 5 A for 250 μs.

If the input voltage is not wired permanently to the circuit, placing an optional resistor in parallel with the 1.2-μF input capacitor bleeds off any charge that remains when power is disconnected. Without R_{bleed}, that capacitor could remain fully charged to the line's peak voltage of 165 V, posing a shock hazard through the input terminals.

Naturally, the first step in the ac-dc conversion relies on C_1's ability to limit incoming current. During each half cycle of the 117-V ac input, the voltage across it changes from −152 V to +152 V. (The peak



2. Within the chips, the Overvoltage/Undervoltage Detection pin, $\overline{OUV}$, goes low if the output voltage falls below 4.65 V or rises above 5.35 V. In the MAX611, the line delays going high until the output has stabilized at 5 V for a certain period.

voltage across C_1 is the peak line voltage minus the voltage drop across the power supply chip.)

Recall that the capacitance charge equals the value of the capacitor times the voltage across it ($Q = CV$). Thus each half cycle transfers 1.5 μF times the peak-to-peak voltage of 304 V, for a total transfer of 0.365 mC. Since there are 120 half cycles every second, the current can be figured as 0.365 mC $\times$ 120 = 43.8 mC/s, for an ac current of 43.8 mA.

For other line voltages and frequencies, the current into the full-wave chips can be calculated as:

$$I_{in} = C_1 \times 2 \times f_{line} \times 2[(V\ pk) - (13.5\ V)]$$

or if the input is a sine wave, as:

$$I_{in} = C_1 \times 2 \times f_{line} \times 2[(V\ rms)(\sqrt{2}) - (13.5\ V)]$$

In shorthand, a 1-μF capacitor limits the current to 37 mA when driven by 117 V ac at 60 Hz.

Because the MAX611 has a half-wave rectifier, only half of each cycle is delivered to the raw dc filter capacitor; the other half recharges the current-limiting capacitor. Thus the available output current is half what it would be for the corresponding circuit built with the MAX610. Consequently, its current can be calculated as:

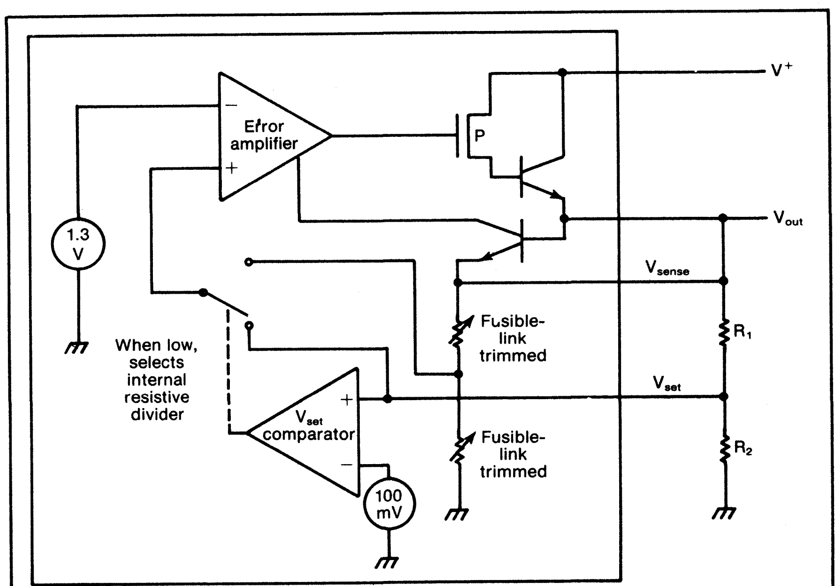
$$I_{in} = C_1 \times f_{line} \times 2[(V\ pk) - (6.5\ V)]$$

where 6.5 V is the average voltage drop across the chip. Handily, the current is about 19 mA for every microfarad of C_1 capacitance, again based on 117 V ac at 60 Hz.

Rectifying the situation

Once the ac input voltage has been current-limited, it is applied to the Ac_1 and Ac_2 input terminals to the on-chip rectifier. That circuit itself is composed of regular pn diodes and synchronous rectifiers (see "A View from the Bridge," opposite).

The bridge rectifier's dc output appears at the V^- and V^+ terminals and is clamped to 12.4-V by the built-in zener diode. Clamping limits the voltage at Ac_1 and Ac_2 to about 14 V (12.4 V plus two diode



3. The V_{set} pin is the center of an innovative voltage-setting circuit within the MAX610 and MAX612. Grounding it selects the preset 5-V output; connecting it to an external resistive voltage divider programs other output voltages. The pin doubles as an analog feedback input.

drops). An external capacitor, connected between V^+ and V^-, filters the rectifier's output. Since current-driven system exhibit a high conduction angle, the value of the filter capacitor need not be as high as in voltage-fed systems. In the MAX610, that filter capacitor can be set at 47 or 100 μF and 16 V.

In the last step, the series-pass regulator produces either a fixed or an adjustable output voltage, thanks primarily to the chips' unusual voltage-setting circuitry. The regulator's programmable current limit is activated whenever the voltage across the optional sensing resistor exceeds 0.6 V. As an interesting note, the series-pass element on the output is a bipolar npn transistor, even though the chip is built with metal-gate CMOS. The transistor base is driven by a large p-channel FET, so that the minimum differential between the values at V^+ and V_{out} is only 1.1 V with a 50-mA current at the output.

Designing with the simplest of the converter chips, the MAX610, is as simple as choosing the optimum component values. Assume, for example, that the chip must generate 5 V dc at 35 mA and that its 60-Hz line ranges from 105 to 130 V rms. Choosing a value of 1.2 μF ±10% for C_1 is sufficient to supply both the load current and the chip's operating current, 70 μA.

The power dissipated in the chip under no-load conditions amounts to the current flowing through the capacitor times 13.5 V (representing the 12.4-V zener and a 1.1-V drop across the diode bridge). Any power delivered to the load reduces the chip's dissipation. Consequently, worst-case dissipation occurs when the line voltage is at its highest value and no load current is being drawn from the 5-V output.

With an input of 130 V rms and a value of 1.32 μF

A view from the bridge

In the evolution of power control ICs, putting a bridge rectifier onto a series-pass regulator would seem to be a natural step. Using a metal-gate CMOS process—rather than a bipolar one—the family of MAX610 power converters successfully puts full-wave and half-wave bridge rectifiers on chip.

Two of the four diodes are pn junctions, formed by the n substrate and p^+ material that normally create the source and drain of p-channel FETs. In an ordinary CMOS IC, forward-biasing the p^+ material would cause latch-up. But the MAX610 contains serial collectors that trap and remove minority carriers, which are injected into the n^- substrate when the p^+n^- junction is forward-biased.

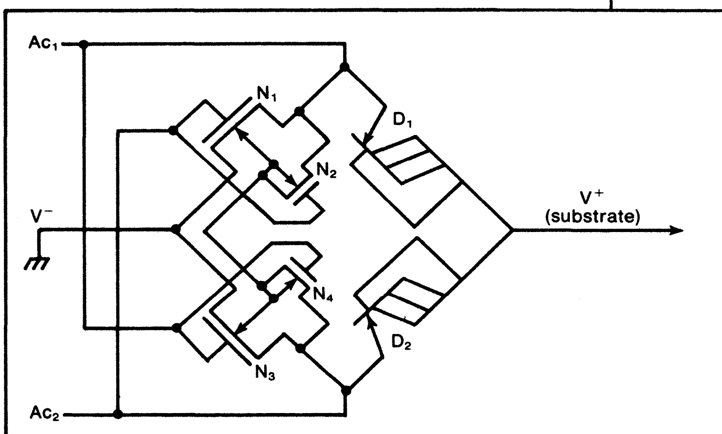
The other two diodes are really n-channel FETs, operating as synchronous rectifiers; therefore their voltage drop is proportional to current. Each one's gates are driven high only when that particular diode is conducting. A slight delay in switching on the FETs does not cause any problem at power line frequencies of 50 to 400 Hz but does degrade efficiency at 1 kHz and above.

Crucial to integrating the bridge rectifier is the substrate of the n-channel FETs. Overall, the substrate must be tied to the source, but since the polarity of the input voltage changes, the FET terminal that acts as a drain for one half of the line cycle is the source for the other half.

When rectifier input Ac_1 is the most positive input, n-channel FETs N_3 and N_4 are turned on (see the figure). N_3 has an on-resistance of about 10 Ω and connects the Ac_2 input to V^-. On the other hand, N_4 connects the four FET substrates to the most negative input potential, Ac_2, and pn junction diode D_1 is forward-biased, connecting Ac_1 to V^+.

When input Ac_2 is the most positive, the operation reverses. D_2 is forward-biased, connecting Ac_2 to V^+. N_1 connects Ac_1 to V^-, and N_2 connects the FET substrates to Ac_1, which is now the most negative potential. The terminal of N_1 that is connected to V^- is now the drain of N_1, and the terminal of N_3 that is connected to Ac_2 is the drain of N_3. During the other half of the cycle, the substrates are connected to Ac_2 and the terminals are the sources of N_1 and N_3.

A standard bridge rectifier with four pn junction diodes has a forward voltage drop of about 1.4 V. The MAX610 bridge rectifier uses only one pn diode at a time. It has a forward voltage drop of about 0.6 V at low currents and a resistance of about 10 Ω. The half-wave rectifier is similar, except that the devices are paralleled to maximize current handling.



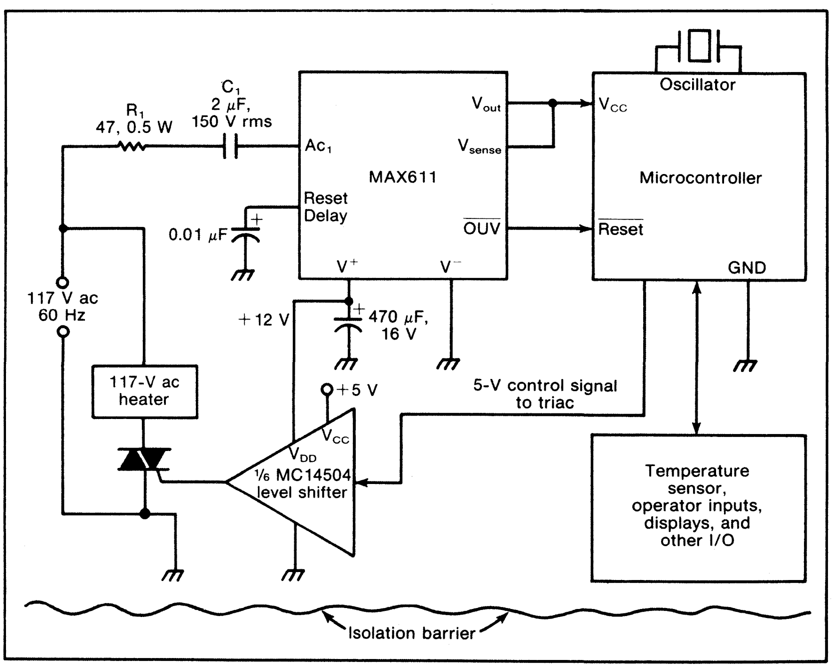
for C_1 (that is, 1.2 μF +10%), the chip's input current is 54 mA, which, when multiplied by 13.5 V, yields a worst-case dissipation of 730 mW—less than the maximum rating allowed for the chip's copper lead-frame 8-pin DIP. Under nominal conditions—C_1 at 1.2 μF, an input of 117 V ac, and the load at 30 mA—the chip dissipates 440 mW. The junction temperature can be lowered, enhancing system reliability, by slipping on an inexpensive heat sink.

Airing the differences

Because the MAX610 represents the simplest version of the converter chip, the other chips give a better idea of the family's scope. Consider an environmental control system that is run by a microcontroller fitted with triac outputs. For this case, the MAX611 is appropriate, but since the chip's 5-V output is not isolated from the power line, all circuitry connected to it, including the temperature sensor, must be surrounded by an isolation barrier that prevents human contact. Here, monitoring the recirculating air from a location that is inaccessible to the user provides enough isolation.

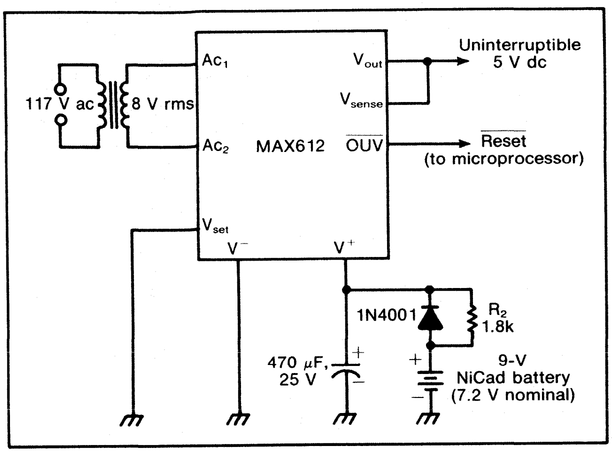
The microcontroller reads the air temperature, compares it with the programmed temperature, and controls the heater through its triac outputs (Fig. 4). The shunt-regulated 12 V dc is used to drive the gate of triacs. For its part, the triac output circuit is designed to be gated off when the converter drives the microprocessor's $\overline{\text{Reset}}$ line low. Thus, the heater can be turned off whenever the supply voltage is outside the microcontroller's operating range and the state of the controller's outputs is unknown.

The MAX612, unlike the other two chips, targets systems that must be isolated from the power line by a transformer. With an operating current of 70 μA, the chip is well suited to uninterruptible power supplies (Fig. 5). If the ac input is disconnected or if it fails, a 9-V nickel-cadmium battery (7.2 V nominal) steps in, supplying the power without interrupting



4. When the MAX611 powers a microcontroller that is monitoring an environmental chamber, it must be treated as if it is connected directly to the 117-V ac input—hence the need for an isolation barrier. The chip's shunt-regulated 12 V dc drives the gate of triacs.

Power supply chips



5. The MAX612 can easily serve as an uninterruptible power source when it hooks up with a 9-V battery. In the event of an extended power outage, $\overline{OUV}$ can be used to turn off most of the system, save its 70 μA of operating current.

the 5-V output. If the power is out for a long period, $\overline{OUV}$ will go low when the battery voltage falls to 5.5 V and the output goes below 4.65 V. If desired, $\overline{OUV}$ can be used to turn off most of the system, leaving only 70 μA of quiescent current and some CMOS memory to save crucial data. When the 117-V ac input voltage is present, the battery recharges itself through the 1.8-kΩ resistor.

The transformer voltage must not exceed the 18.6-V zener voltage pin plus two diode drops. An 8-V nominal transformer comes in handy, since the peak output voltage is still less than the zener voltage for line voltages of 140 V rms or more. Moreover, only 90 V rms is required to recharge the battery and supply the 5-V load current.

In contrast to the situation in a current-driven system, the diode bridge rectifier in this circuit only conducts near the peak voltage of each half cycle, recharging the raw dc filter capacitor, which then supplies the current for the rest of the cycle. With a value of 470 μF, the filter capacitor can minimize the ripple at the input to the series-pass regulator. Switching between power sources on the raw dc side of the chip's regulator eliminates variations in output voltage that occur when systems alternate between a battery and 5 V on the load side.□

How will your A/D really perform?

Simple test circuits can help you get the most out of integrating A/D converters.

Charlie Allen
Maxim Integrated Products
Sunnyvale, California

Integrating analog-to-digital (A/D) converters offer high accuracy, nearly ideal differential nonlinearity, high power supply rejection, and low cost. As such, they're often used in digital multimeters, temperature meters, process control gear, and data acquisition systems.

In many of these diverse applications, integrating A/Ds are often used under conditions other than those for which the converter is specified. There are, however, some simple—yet effective—ways to determine the performance of integrating A/Ds under the actual conditions of your application, as well as methods to optimize this performance.

While the specific converter discussed here is the Intersil ICL-7129A, the test circuits and procedures are applicable to all popular integrating A/Ds. The ICL-7129A is a 4½-digit (±20,000 count resolution) analog-to-digital converter with an onboard triplexed LCD. It has full-scale ranges of both 2 volts and 200

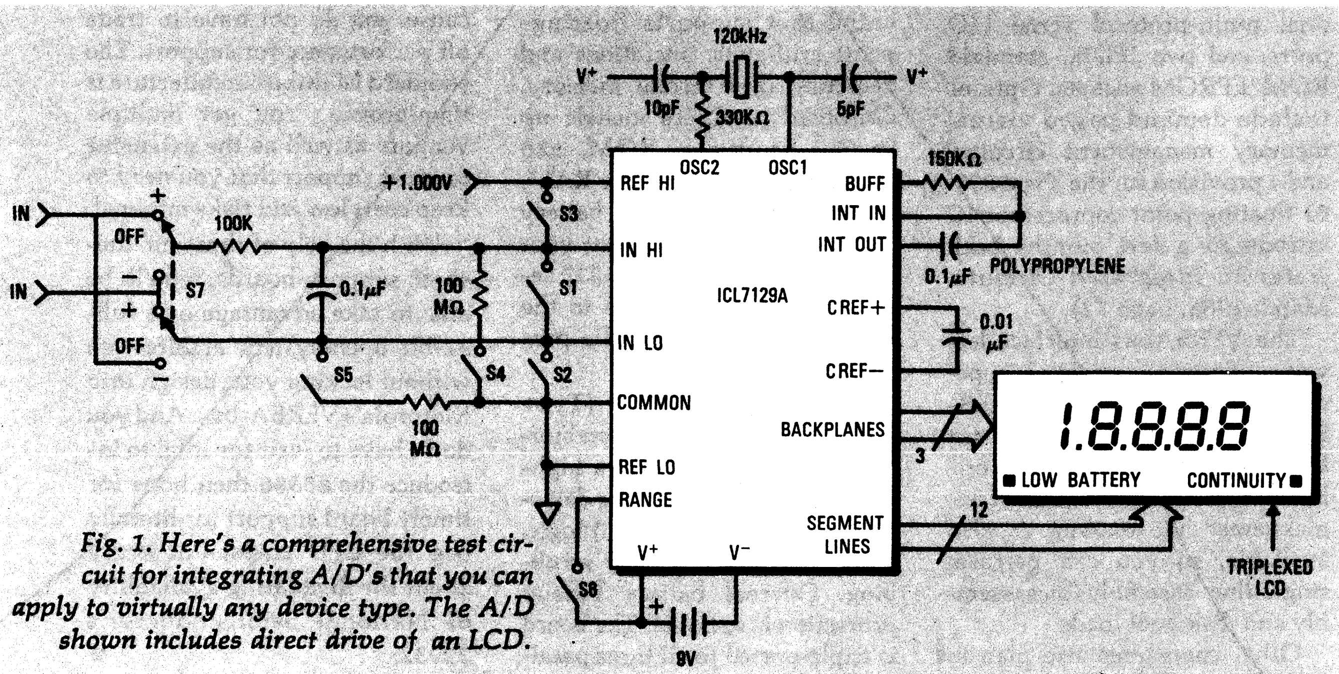


Fig. 1. Here's a comprehensive test circuit for integrating A/D's that you can apply to virtually any device type. The A/D shown includes direct drive of an LCD.

millivolts, with 10 microvolt resolution on the low scale.

The Test Circuit

The test circuit shown in Figure 1 allows you to directly test many of the key specifications of the ICL7129A and other integrating A/Ds. Details are shown in the accompanying table.

A high-resolution input-voltage source is required to measure A/D noise. This voltage source must also have good short term stability but need not be a high-accuracy source. A 6½-digit voltage calibrator, if available, makes an ideal source. However, you can also make an acceptable input voltage source by summing a coarsely-adjusted voltage source with an attenuated "dither" voltage (Figure 2).

Similar to the "dither DAC" arrangement, often used to test successive-approximation A/Ds, the dither voltage is attenuated by a factor of 100 and added to the coarse voltage source before being applied to the A/D.

Measure the dither voltage with a 4½-digit multimeter before attenuation. This provides an effective resolution of 6½-digits referred to the input of the A/D under test. (The resolution of the input voltage source is 1/100 count of the 4½-digit A/D under test.)

Noise Management

Noise is one of the most difficult parameters to measure quantitatively. Yet, it is often the primary factor in selecting an A/D converter IC. Most integrating A/Ds specify a "peak-to-peak noise not exceeded 95 percent of the time." This value is 3.3 times the RMS noise, assuming Gaussian distribution.

The observed effect of noise is jitter, or flicker of the least significant digit (Figure 3). If the peak-to-peak noise is greater than one count, the A/D will display three different results for some input voltages.

When the noise is exactly one count, the display will jump back and forth between two adjacent digits. At ½ count, the display will jump back and forth between adjacent digits when the input voltage is within ±¼ of one count of the transition point between displayed counts, and will show a single reading when the noise is not close to a transition point.

To measure noise using the transition detection method, first adjust the input voltage until the A/D displays one reading almost all of the time, with only an occasional display of the next higher reading. The higher reading should occur approximately once every 20 conversions. Record the input voltage.

Next, adjust the input voltage upward until the higher reading is displayed for almost every conversion, with only an occasional display of the lower reading. Record this second input voltage. The peak-to-peak noise is simply the difference between these two readings.

Use Care

The primary problem with this method is that it tends to be sub-

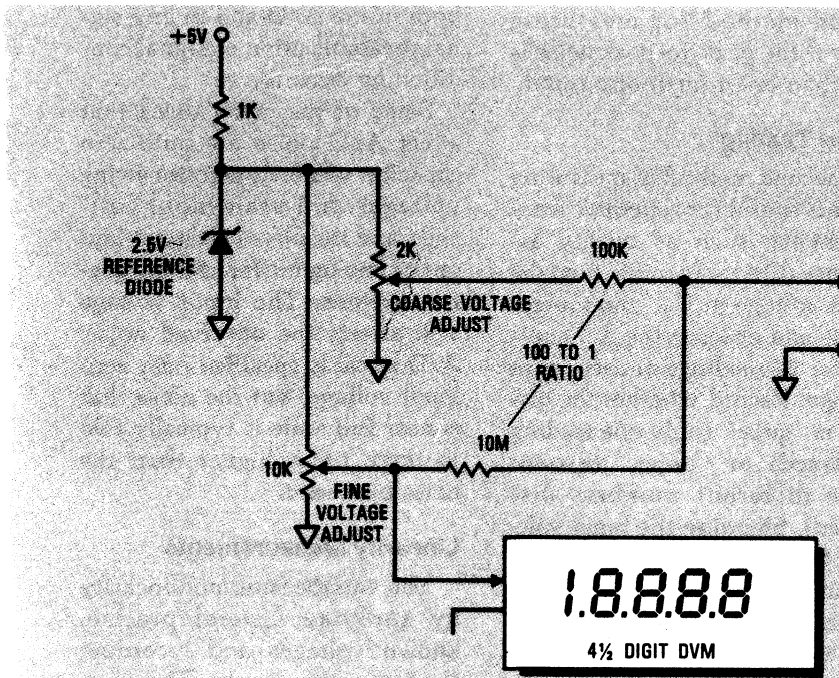


Fig. 2. This high-resolution voltage source for testing A/D converters sums a coarse voltage with an attenuated "dither" source.

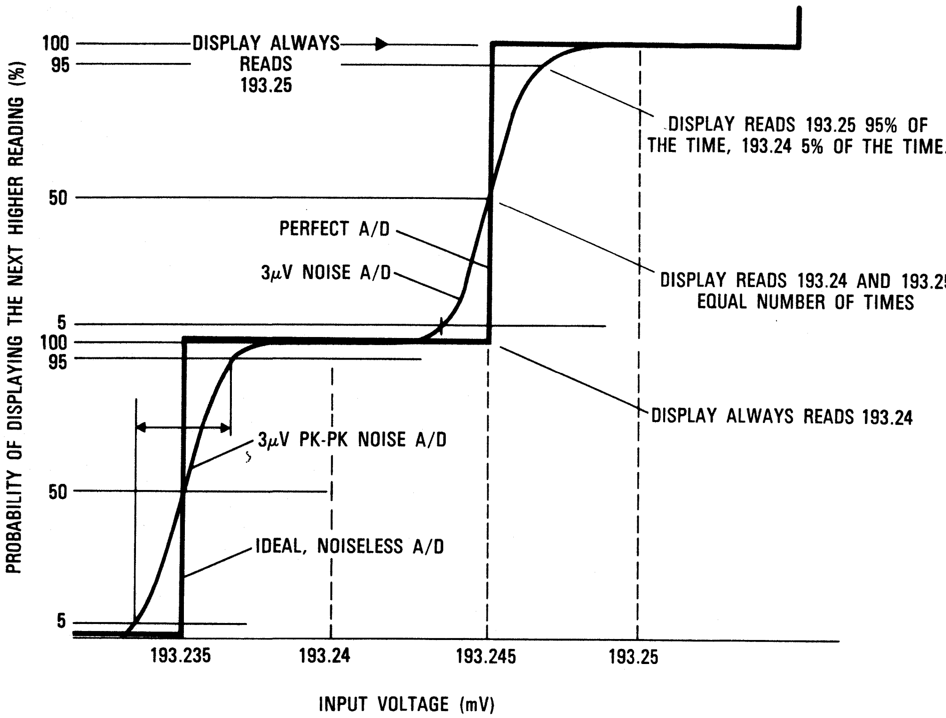


Fig. 3. Noise causes A/Ds to have a transition region, where more than one reading is displayed for a fixed input voltage. Noise is defined as the width of this transition band, measured from the 5 to 95 percent points.

jective; it is very difficult to determine the exact voltage where the adjacent reading is displayed once per 20 conversions. A second problem is that measurements take several minutes per device. Nonetheless, this method gives accurate results when carefully performed. It's the preferred method for measuring noise if the peak-to-peak noise is less than one-fourth of a count.

Noise Testing

A second method of measuring noise is useful for repetitive measurements, such as quality assurance (QA) screening. Step the input voltage in $\frac{1}{10}$ count increments and observe the A/D output for 20 readings at each input voltage. Record whether the display is "quiet" (only one reading displayed) or "noisy" (two or more different numbers displayed). Stepping the input voltage ten times puts the chip through one entire count. The number of "noisy" input voltages is proportional to the peak-to-peak noise.

For example, if the A/D has noise of ½ count, the display will typically have jitter on five of the ten input voltages applied. If the peak-to-peak noise of the A/D is only 0.2 count, the display will typically be stable for eight of the ten input voltages and alternate between adjacent counts with two of the input voltages.

Due to the random nature of noise, repeated measurements of the same device may result in different readings. The average result of several measurements will be the true noise of the device under test.

This method requires little judgement on the operator's part, and about 15 units per hour can be sample tested. Since the noise of an integrating A/D is relatively constant for all devices in a given lot, a sample size of five units is sufficient to characterize each lot.

In addition to evaluating the noise level of A/Ds from various manufacturers, these noise test methods let you determine the effect of circuit design changes, both in the A/D and in any signal preamplication or signal conditioning circuitry.

Some of the many items that affect A/D noise are auto-zero capacitor values, integrator swing voltages, full-scale input voltages, and the physical circuit layout of analog buffer and integrator sections. The input voltage also affects the observed noise. A/D noise is specified near zero input voltage, but the noise that is near full scale is typically two to three times higher than the noise near zero.

Linearity Measurements

You can measure nonlinearity by applying several precise, known voltages and recording the conversion results. The maximum deviation from a "best fit" straight line through the measured points is the sum of the nonlinearity error plus quantizing errors of up to one-half count. The quantizing error term can be eliminated by finding the transition points where two adjacent codes are each displayed 50 percent of the time.

For example, if the display is flickering back and forth between 0009 and 0010, with each reading being displayed about 50 percent of the time, the digital output code is considered to be 0009.5. Accurate nonlinearity measurement requires the use of either a 6½-digit voltage calibrator or a calibrated 6½-digit meter to measure the voltage applied to the A/D converter.

Zero Code Width Measurement

Integrating A/Ds with sign magnitude output format have two zeros: +0000 and −0000. To have a linear transfer function between input voltage and displayed readings, these two zero codes should each be one-half the width of normal codes (Figure 4). Using the standard circuits shown in data sheets, the delay of the A/D zero-crossing comparator is about half a clock cycle. This narrows the width of the zero codes to the desired half-count width. Keep in mind that operating the A/D at a different clock rate or with significantly different component values will affect the zero code width and thus should be evaluated.

You can determine the width of the plus and minus zero codes by applying a small voltage to the input and adjusting this voltage until equal numbers of zeros and ones are displayed.

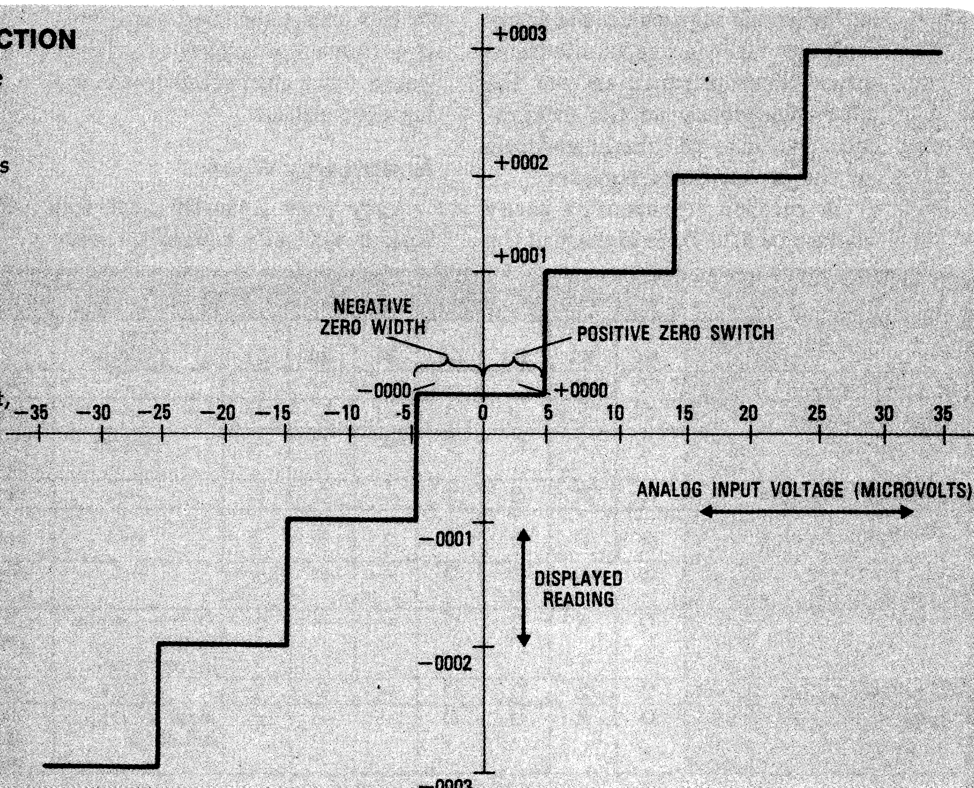
A better way to test for zero offset is to find the transitions between +0009 and +0010. The zero offset is the difference in magnitude between the measured transition points. To determine the zero code width, compare the measured transition points with the desired transition point.

For 200 millivolts full scale, the ideal transition point between 0009 and 0010 is 95 microvolts. Measuring a transition point five to twenty counts away from zero is the preferred method of measuring zero-offset and zero-width. The first few counts around zero may have additional nonlinearities caused by op-amp settling. This may happen when transients occur as a result of the A/D switching from the signal-integrate phase to the reference-deintegrate phase.

Placing a low-value resistor in series with the integrating capacitor can help you adjust the zero-code width of many analog-to-digital converters. The voltage across this resister causes the input voltage to the comparator to cross earlier than the true zero-crossing. It thereby com-

THE A/D TRANSFER FUNCTION

This example shows the discrete steps in the A/D output. The transfer function of an ideal integrating A/D around zero has two zeros, +0000 and −0000, each of which is ½ the width of the other count widths. The delay of the A/D zero crossing comparator shortens the zero code width to ½ count, putting the zero-to-one code transition at the desired 5-microvolt point, rather than at 10-microvolts. Since the comparator delay is a fixed time period, it becomes a large fraction of the clock period as the clock frequency is increased in order to perform more conversions per second. At very high conversion rates, the zero code will disappear. Under standard data sheet conditions, this A/D is guaranteed to display +0000 or −0000 with a zero volt input.



pensates for the comparator delay.

There are several different causes of excessive noise. Look at the integrator output voltage during a conversion. A very low integrator voltage swing reduces the signal-to-noise ratio at the comparator input, increasing the A/D converter noise. The integrator output should swing to within one to two volts of the power supply voltage when you apply a full-scale input voltage.

Next, make sure that the auto-zero capacitor is at least twice the value of the integrator capacitor. This ensures that the noise of the auto-zero loop will be negligible compared to that of the buffer and integrator. (This precaution does not apply to the ICL7129A, which reduces overall system noise by digitally compensating for the A/D offset.)

A third item to check is normal-mode rejection. Sometimes, what at first appears to be noise is caused by 50 or 60 Hz differential mode signals on the input voltage. Either increase the input filter time-constant, or set the clock frequency so the integration period is an integer multiple of the power line frequency.

A related problem is stray pickup of 50/60 Hz signals in the analog section of the A/D's external circuitry. When the observed noise varies in a cyclic fashion as the input voltage is raised, consider it a symptom of stray pickup in the analog integrator section.

This occurs because the input voltage determines the duration of the deintegrate mode, and therefore also determines the amount of 50/60 Hz normal mode rejection. The A/D noise will be lowest when the deintegrate cycle is an integer number of line cycles. Conversely, the noise will be highest when the deintegrate cycle is an odd number of half cycles of the line frequency.

The most sensitive node is the junction of the integrator resistor and the integrator capacitor. The auto-zero capacitor is a potential point for stray coupling in many analog-to-digital converters. Stray pickup by these sensitive nodes can be best be avoided by keeping connections in this area short and surrounding these components with a guard trace connected to the integrator output.

Nonlinearity Woes

Very poor linearity near full scale is typically caused by saturation of the integrator output stage. Increase the integrator resistor value if the integrator output voltage swings within one volt of either supply rail with positive and negative full-scale inputs. Another potential cause of gross nonlinearity is excessive current draw from the buffer or integrator.

Nonlinearity errors of a couple of counts are usually the result of dielectric absorption in the integrator capacitor. The best test for dielectric absorption is to try a Teflon™ capacitor and see if this cures the problem. Also, do not exceed the input common-mode voltage range of the A/D. For best nonlinearity, keep the input voltage at least 1.5 volts above V− and at least one volt below V+.

Rollover Error

Rollover error is the difference in A/D readings for equal positive and negative voltages near full scale (Figure 5). An example is the difference in the readings on a multimeter when you interchange the input leads. Rollover error can be caused by either a difference in offset for positive and negative voltages or by a difference in gain for these voltages.

The gain error is very com-

TABLE OF SWITCH SETTINGS FOR A/D TEST CIRCUIT

	S1	S2	S3	S4	S5	S6	S7	Input Voltage	Measure
Zero Reading	X	X	O	—	—	—	O	—	Display should read + 0000 or −0000
Ratiometer Error	O	X	X	O	—	X	O	—	Display reads 9998 to 10,000
Rollover Error	O	X	O	O	—	O	±	≈199mV	Rollover error is difference in reading as S7 reverses the input voltage polarity
In Hi Input Leakage	O	X	O	X	—	O	O	—	Leakage in pA = reading ÷ 10
In Lo Input Leakage	O	O	O	O	X	O	O	—	Leakage in pA = $\frac{\text{reading}}{10}$
Common Mode Rejection Ratio	X	O	O	O	—	O	±	V_{CM}	Display should read ±0000 for V_{CM} <3V
Linearity	O	X	O	O	—	—	±	O to full scale in 0.1FS increments	Measure input V w/ 5½ digit meter compare w/ displayed reading
Zero Code width	O	X	O	O	—	O	±	≈95uV	Find ±0009 to ±0010 transition
Noise	O	X	O	O	—	—	±	Near 0V and full scale	Measure width of transition band where display shows two adjacent readings

X = Closed O = Open — = Don't care

mon, with stray capacitance on the reference capacitor terminals being the most common cause of poor rollover performance. Test for stray capacitance by increasing the reference capacitor value by a factor of 10. If rollover is signiflcantly reduced, then stray capacitance between the reference capacitor and ground is the culprit. Common-mode errors, which offset the entire transfer curve in one direction, will also cause a rollover error.

Input bias current can cause rollover problems when the input source impedance is very high. The ICL7129A uses an input chopping technique to reduce the noise level. This chopping, however, increases the input bias current to about 8 picoamps. If the input source impedance is one Megohm, this small bias current will cause an offset of 8 microvolts, or 0.8 count. If you can't lower the source impedance or add a buffer stage, you may have to use the non-chopped ICL7129, which reduces bias current to 0.1 picoamps while increasing noise level slightly.

Other possible rollover error sources are reference capacitor leakage and low integrator gain in the A/D. You can prevent stray leakage into the reference capacitor by using wide spacings between the reference capacitor terminals and any other circuitry. Guarding these terminals, however, is not recommended, since this will increase the stray capacitance and aggravate rollover problems. At temperatures above 70 degrees C, increase the value of the reference capacitor to minimize the effect of the internal leakage at the A/D reference capacitor input terminals.

A non-zero reading with zero input is rarely seen in the ICL-7129A but is sometimes observed in earlier generations of A/Ds. You can often trace this problem back to a common-mode voltage between analog ground (or common) and the low (inverting) input terminal of the converter. Connect these two points together directly whenever possible. In all cases, keep the input-low terminal within 2 volts of analog ground and maintain at least a 2-volt margin from both power supply rails.

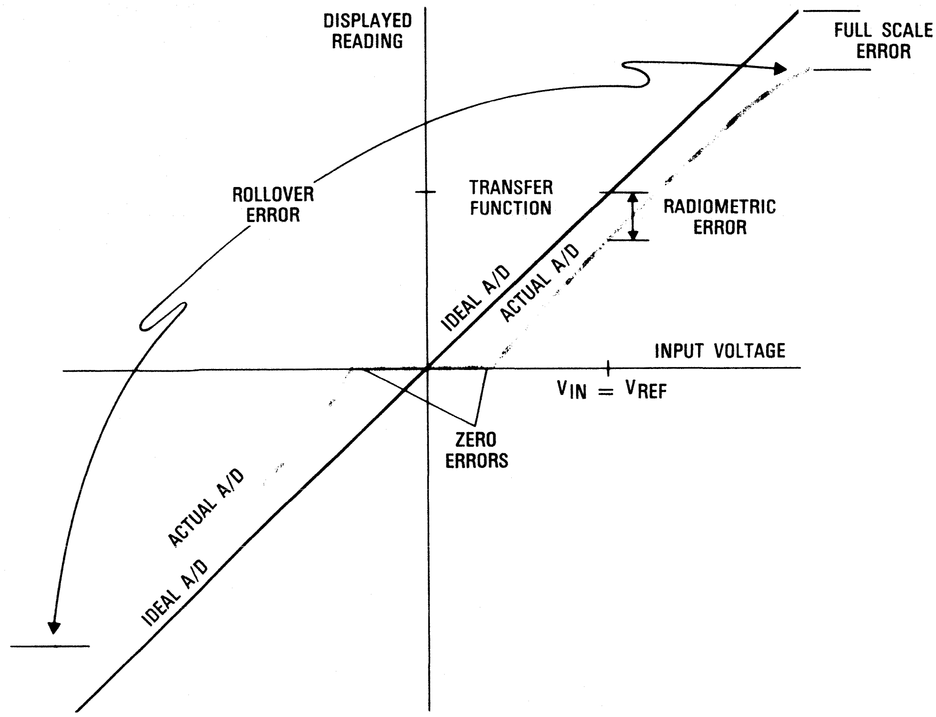


Fig. 5. Error terms of the Intersil ICL7129A are shown here greatly exaggerated. The deviation from a best-fit straight line is the non-linearity error. The difference in displayed readings for equal positive and negative inputs is the rollover error. The deviation from the ideal reading of 10,000 when the reference voltage is applied to the input is the ratiometric error.

Single-supply Operation

Integrating A/Ds can be operated from a single 5-volt supply if you observe several precautions. Keep the input-common and input-low terminals centered between V+ and ground. It may also be necessary to increase the value of the integrating capacitor, so that the integrator swings to about 1 volt from V+ and ground when full-scale inputs are applied. In addition, you should limit the full-scale input to 1 volt and increase the value of the auto-zero capacitor by a factor of four, to reduce noise.

The circuit shown in Figure 1 is suitable for virtually all integrating A/Ds. Most of these converters use an autozero capacitor in series with the integrator input, but this does not change the test procedures. The input voltage, however, must correspond with the full-scale voltage actually being used. The only other significant difference among the commonly available integrating A/Ds is whether they are intended for single-supply or dual-supply operation. For dual-supply converters, separate grounds for the digital section and the analog section are essential. ■

CMOS curbs the appetite of power-hungry dc-dc converter chips

Working with CMOS rather than the standard bipolar process cuts the typical operating current of a dc-dc converter more than a hundredfold.

Designing dc-dc converters involves two basic steps: Selecting a converter control chip and adding the required external components, such as coils, resistors, and capacitors. The efficiency of the control-circuit IC determines how well the converter does its job. For a battery-powered system, high efficiency means that battery life is extended because voltage is converted to the desired value with minimum losses. In the case of ac-powered systems, efficiency translates into cooler adjacent components.

Bipolar control chips for dc-dc converters usually pull about 12 mA, with a minimum required operating voltage of 3.5 V. Quiescent current is on the order of several hundred microamperes. In contrast, the first CMOS converter control IC drastically changes such customary values, setting new standards for these devices. The MAX630's typical operating current comes in at 70 μA, its minimum operating voltage is merely 1.8 V, and the chip's quiescent current is a miserly 0.5 μA.

In addition, the 8-pin device needs no base current for its n-channel MOSFET output transistor, which supplies up to 300 mW when converting 5 V to 15 V. And output power is virtually unlimited when the chip is used with one external power-boosting transistor or power MOSFET.

The chip can best be described while in action, say in a voltage-regulating loop (Fig. 1). When 5 V is first applied, current flows through inductor L and diode D, supplying the chip with 4.4 V for startup. The desired output voltage is selected by changing the value of feedback resistors R_1 and R_2.

Comparator 1 determines whether a specific fraction of the output voltage is equal to the 1.3-V internal reference. When the converter's output voltage is too low, the comparator's output is high and the 40-kHz oscillator pulses are allowed through the NOR gate latch, turning on output MOSFET Q_1. As long as the output voltage is less than the desired voltage, Q_1 enables the inductor to be driven from the 5-V supply with pulses at the oscillator frequency.

Charlie Allen
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Delivering stored energy

Each time Q_1 is turned on, the current increases through L, storing energy. When Q_1 turns off, the polarity of the voltage reverses across the inductor (a result of L di/dt). The voltage at L_X then increases until D is forward-biased and current is delivered to the output.

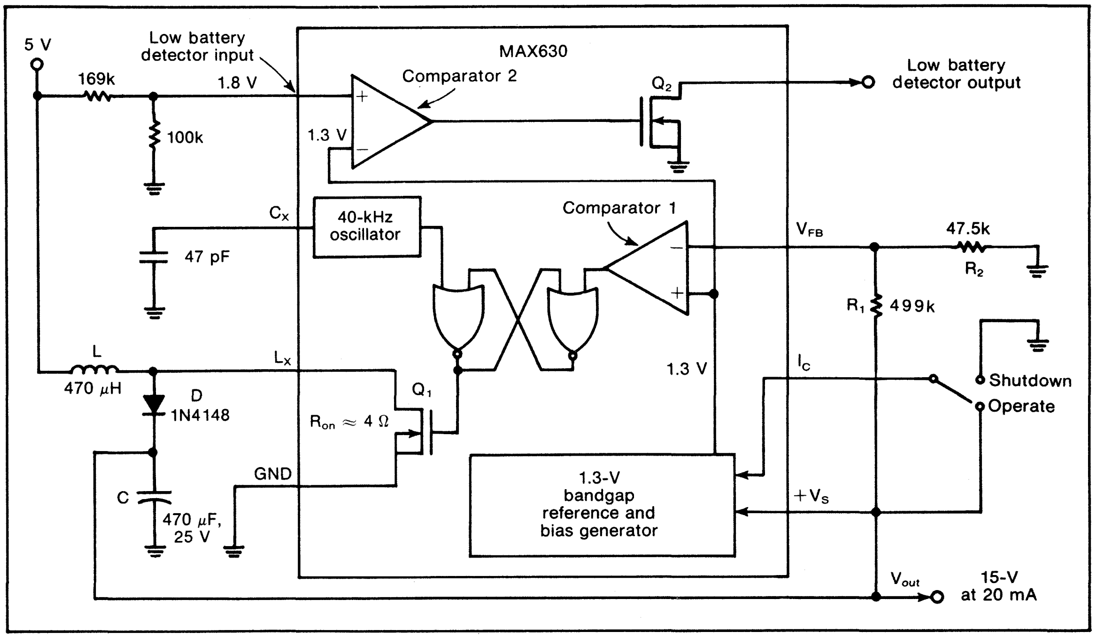
When Q_1 is on, the current rises linearly because $\frac{di}{dt} = \frac{V}{L}$. At the end of the on-time (14 μs for a 40 kHz, 55% duty-cycle oscillator), $I = \frac{VT_{on}}{L} = \frac{5\text{ V} \times 14\,\mu\text{s}}{470\,\mu\text{H}} =$ 150 mA. The energy in the inductor is $P = \frac{1}{2}LI^2 =$ 5.25 μW-s.

Q_1 is turned on and off 40,000 times every second, and at maximum load the power transferred through the coil is 40,000 × 5.25, or 210 mW. Since the inductor supplies only the voltage above the input voltage, at 15 V the dc-dc converter can supply 210 mW/(15 V − 5 V) or 21 mA. The inductor furnishes 210 mW, the battery directly supplies another 105 mW (that is, 21 mA × 5 V) for a total output power of 315 mW. If the load draws less than 21 mA, the IC turns on its n-channel MOSFET output control only often enough to keep the voltage at a constant 15 V.

Thus, contrary to what might be expected, reducing the inductor value increases the available output current. In other words, lowering L increases the peak current, thereby increasing the available power.

When the output voltage reaches the desired value, $1.31\left(1 + \frac{R_1}{R_2}\right)$, the comparator's output goes low and the oscillator pulses no longer turn on Q_1. The output current is then supplied by filter capacitor C, which limits ripple to about 50 mV. As the output voltage drops below the comparator threshold, Q_1 is again switched on, repeating the cycle. On average, the duty cycle at the output is directly proportional to the output current.

An n-channel MOSFET with an on-resistance of about 4 Ω and a maximum current rating of 250 mA (peak) controls the chip's output. MOSFETs have two distinct advantages over bipolar transistors. Their higher speed, which reduces switching losses and allows for smaller, lighter, and less costly magnetic components is the benefit generally mentioned with reference to high-power switchers. It must be kept in mind, though, that if a low-cost molded inductor is employed, circuit efficiency comes in at roughly 75%. An inductor with a lower series resistance, however, boosts efficiency to around 90%. In highly efficient low-power dc-dc converters, the advantage is that MOSFETs require no base current. Alternatively, their bipolar kin must use a portion of the input power for that purpose, reducing circuit efficiency. The accompanying power loss is most evi-



1. Using the MAX630 in a voltage-regulating loop delivers 15 V from a 5-V source. The divided output voltage is applied to comparator 1, where it is checked against the 1.3-V reference voltage. The comparator output enables the oscillator pulses, turning on Q_1 and allowing current to store energy in inductor L. When Q_1 is off, that energy is applied to the output.

dent in voltage-boosting systems, such as up-converters. In fact, the chip's low operating current and MOSFET output control features superior efficiency in such applications. A chip carrying a bipolar transistor, though, could use up to 10% of the total input power.

The oscillator frequency is determined by a single external component, a low-cost ceramic capacitor. A 47-pF capacitor sets the frequency to 40 kHz, a reasonable compromise between the smaller switching losses associated with lower frequencies, and the smaller inductor found at higher frequencies.

On-chip convenience

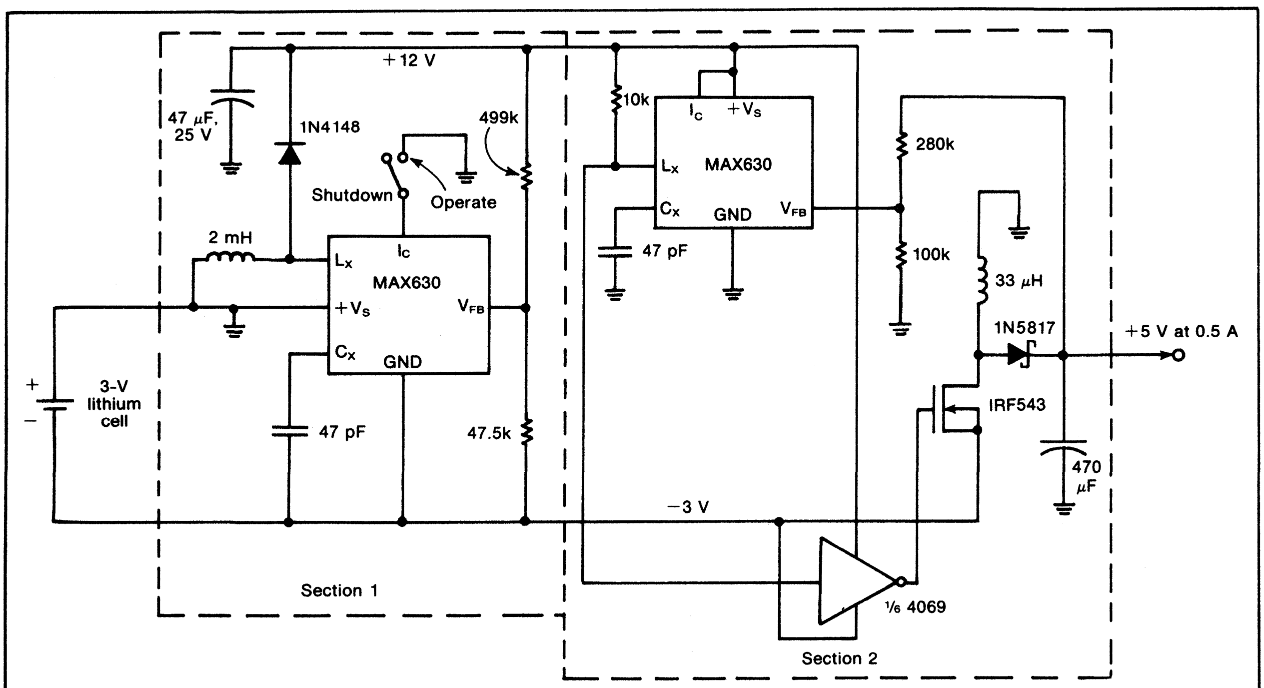
Two of the converter's features simplify power supply design—a low-battery detector and a shutdown mode. Specifically, the first compares the voltage on the Low Battery input with the internal 1.3-V reference. The detector's output is an open-drain n-channel MOSFET. The low-battery detector is also called into play in other voltage-monitoring operations such as determining ac line-power failure in battery-backed systems.

The second feature employs the I_c pin to shut down the device. When in that state, the chip turns off the analog bias generator and draws less than 1 μA of quiescent current. The shutdown condition is entered whenever the I_c pin is left floating or is driven below 1 V.

Some battery-powered systems, however, call for short bursts of high current. The extra circuitry designed to meet that need (Fig. 2) converts −3 V to +5 V in the buck-boost or flyback mode (see "Dc-dc Converter Configurations," p. 178).

The left half of the circuit is similar to that of the 5-to-15-V converter, and supplies 15 V for the gate drive of the external power MOSFET. This gate drive ensures that the external power MOSFET is fully turned on and presents a low resistance.

The right half of the circuit is a −3 V to +5 V buck-boost converter. An advantage of this circuit is that when the chip is turned off the output falls to



2. High current for short periods is obtained by adding a second section to the basic dc-dc converter. The first section furnishes a 15-V output from a −3-V lithium cell and controls an external power MOSFET. The second is a buck-boost converter, which supplies +5 V from −3 V.

0 V. On the other hand, in the standard boost circuit (Fig. 1 again), the output is $V_{BATT} - 0.6$ V with the chip shut down. Also, when the buck-boost converter is shut off, it draws less than 10 μA.

The inductor and output filter capacitor values have been changed to accommodate the increased power levels. At the indicated values, the circuit supplies up to 500 mA at 5 V, with an efficiency of 85%.

An alternative approach to high power conversion (-3 V to $+5$ V) uses a single chip and an inductor with two windings. One winding supplies 10 to 15 V that drives the power MOSFET gate; the second winding delivers power for the 5-V output. Since the MOSFET gate drive is only the battery voltage minus one diode drop during startup, the MOSFET must be a low threshold device. On the other hand, the circuit that pairs up the converter chips will start and operate with power MOSFETs that have high threshold voltages.

The converter IC ensures a continuous supply of regulated $+5$ V, and automatically switches over between line power and battery backup (Fig. 3). When

Dc-dc converter configurations

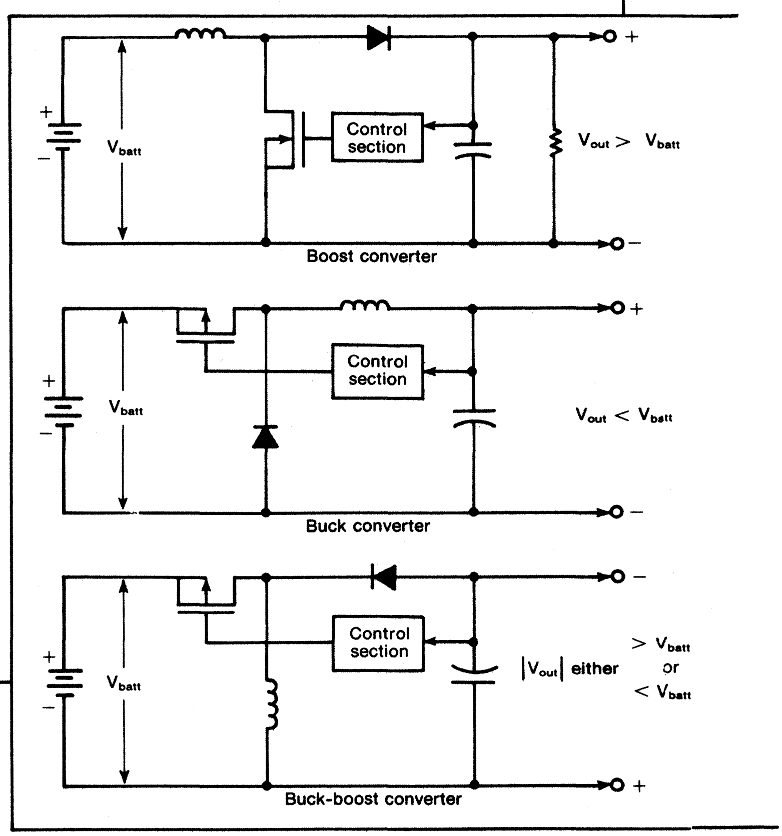
Dc-dc converters come in three basic topologies: buck, boost, and buck-boost—each of which meets a particular need. When the output voltage is greater than the input, the converter is usually operated in the positive voltage boost circuit (see the figure). The buck circuit is employed when the input voltage is always greater than the desired output voltage. Finally, the buck-boost circuit inverts the input voltage, and can be used with an input voltage that is either greater or less than the desired output.

Dc-dc converters also can be classified according to the method by which they control their output voltages. The two most common approaches are pulse width modulation (PWM) and pulse frequency modulation (PFM). Pulse-width-modulated switchers (current-mode control being one variant) are well established in high-power switching supplies that work off the ac line.

Both PWM and PFM circuits control the output voltage by varying the duty cycle. In the former, the frequency is held constant and the width of each pulse is varied. In the latter, the pulse width is held constant and the duty cycle is controlled by changing the pulse repetition rate.

The MAX630 refines the basic PFM approach by supplying a constant frequency oscillator. The chip's output MOSFET is switched on when the oscillator output is high and the converter's output voltage is lower than desired. If the output voltage is higher than wanted, the MOSFET output control is disabled for that oscillator cycle. This "pulse skipping" varies the average duty cycle and thus controls the output voltage.

Unlike the PWM ICs that use an op amp as the control element, the converter chip uses a comparator to check the output voltage against an internal reference. That arrangement reduces the die size, the number of external components, and the operating current.



the line-powered input voltage is 5 V, it furnishes 4.4 V to the chip and trickle charges the battery. The device runs continuously, boosting the 4.4 V to 5 V. When that line-powered 5-V input falls below the 3.6-V battery voltage, the battery supplies the power to the chip. The converter boosts the battery voltage to 5 V, delivering a continuous dc supply to the uninterruptible 5-V bus. Since the 5 V output is always supplied through the chip, there are no power spikes or glitches during power transfers.

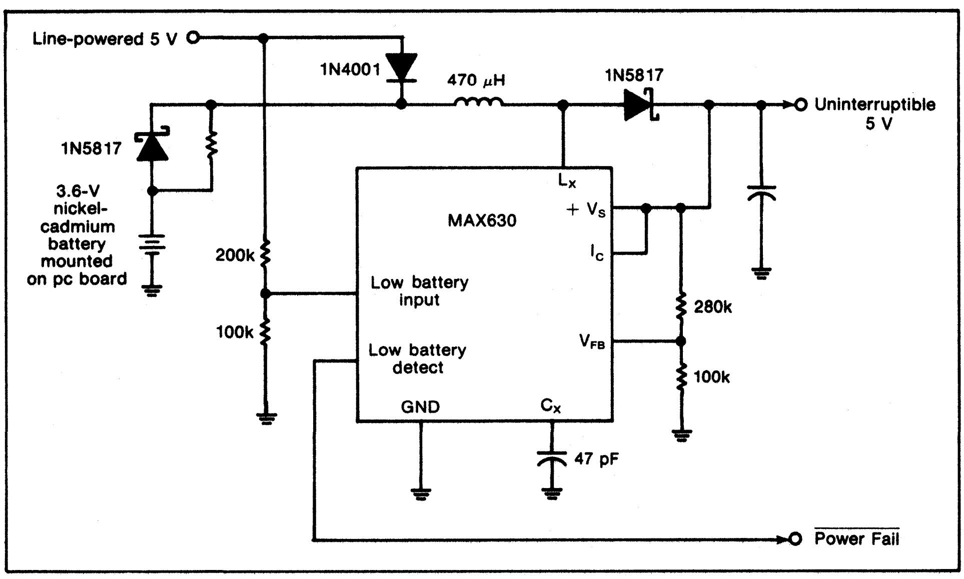
Keeping an eye on the battery

The device's low-battery detector monitors the line-powered 5-V line, and the low-battery detector output can shut down unnecessary sections of the system during power failures. Alternatively, the low-battery detector could monitor the Nicad battery voltage and warn of loss of power when the battery is nearly discharged.

Unlike battery backup systems that use 9-V batteries, this circuit does not need 12 or 15 V to recharge the battery. Therefore, it can be used to supply a 5-V backup for modules or circuit cards on which only a 5-V supply is available.

A common problem in large electronic systems is that, although multiple power supply voltages are used, they are not always available where they are needed. Often only +5 V is distributed, but one section or printed circuit card needs both +5 V and +15 V. The ready availability of power converters enables the designer to have the needed voltage available at his fingertips.□

Charlie Allen has served as the applications engineering manager at Maxim for the past year and a half. Before that, he held the equivalent position at Intersil. He received a BSEE from Michigan State University and has published several technical articles in the past few years.



3. The MAX630 also acts as an uninterruptible 5-V power supply. It converts either the line-powered 5 V or the 3.6-V Nicad battery input to 5 V. The Power Fail output goes low when the line-powered 5-V input fails.

High-precision CMOS op amps accommodate ±15V supplies

Monolithic, dc-stabilized CMOS amplifiers are no longer limited to a ±8V power-supply range: One family of CMOS op amps can operate from ±15V supplies. The op amps' wider supply range makes them suitable for use in a variety of analog systems that require precision dc amplification.

Leonard Sherman,
Maxim Integrated Products

By employing a family of chopper-stabilized CMOS op amps (the MAX420 Series) that can operate from ±2.5 to ±16.5V supplies, you can obtain precision dc amplification in industrial-control, data-acquisition, servo, and other applications that were beyond the capabilities of earlier CMOS chopper-stabilized op amps.

These precision amplifiers can provide good signal conditioning for thermocouples, for example. Despite their advantages—low cost, high reliability, and the ability to measure wide temperature ranges—thermocouples are somewhat difficult to deal with electrically, because they have low output signal levels and require an ice-point reference. Typical thermocouple output signal levels (on the order of tens of microvolts per degree C) dictate that the signal-conditioning amplifiers used with thermocouples must have input offset and drift specifications well below those levels. You need such specs especially if you want to realize resolution and repeatability to a fraction of a degree.

The circuit in **Fig 1** readily amplifies thermocouples' low-level output signals. The MAX422 has a maximum input drift spec of 50 nV/°C, so the amplifier contributes only 0.001° of output error for each degree of shift in ambient temperature. Because the op amp can use ±15V power supplies, the design is simple: A basic inverting summing network combines the thermocouple output, cold-junction compensation, and cold-junction offset.

The trim procedure is also very simple, because gain and cold-junction adjustments don't interact. This lack of interaction is a significant advantage in multichannel setups, which are fairly common in thermocouple measurement systems.

The small-signal npn transistor provides ice-point compensation for the thermocouple. It generates a

Signal-conditioning amplifiers used with thermocouples must have input offset and drift specifications well below typical thermocouple output signal levels.

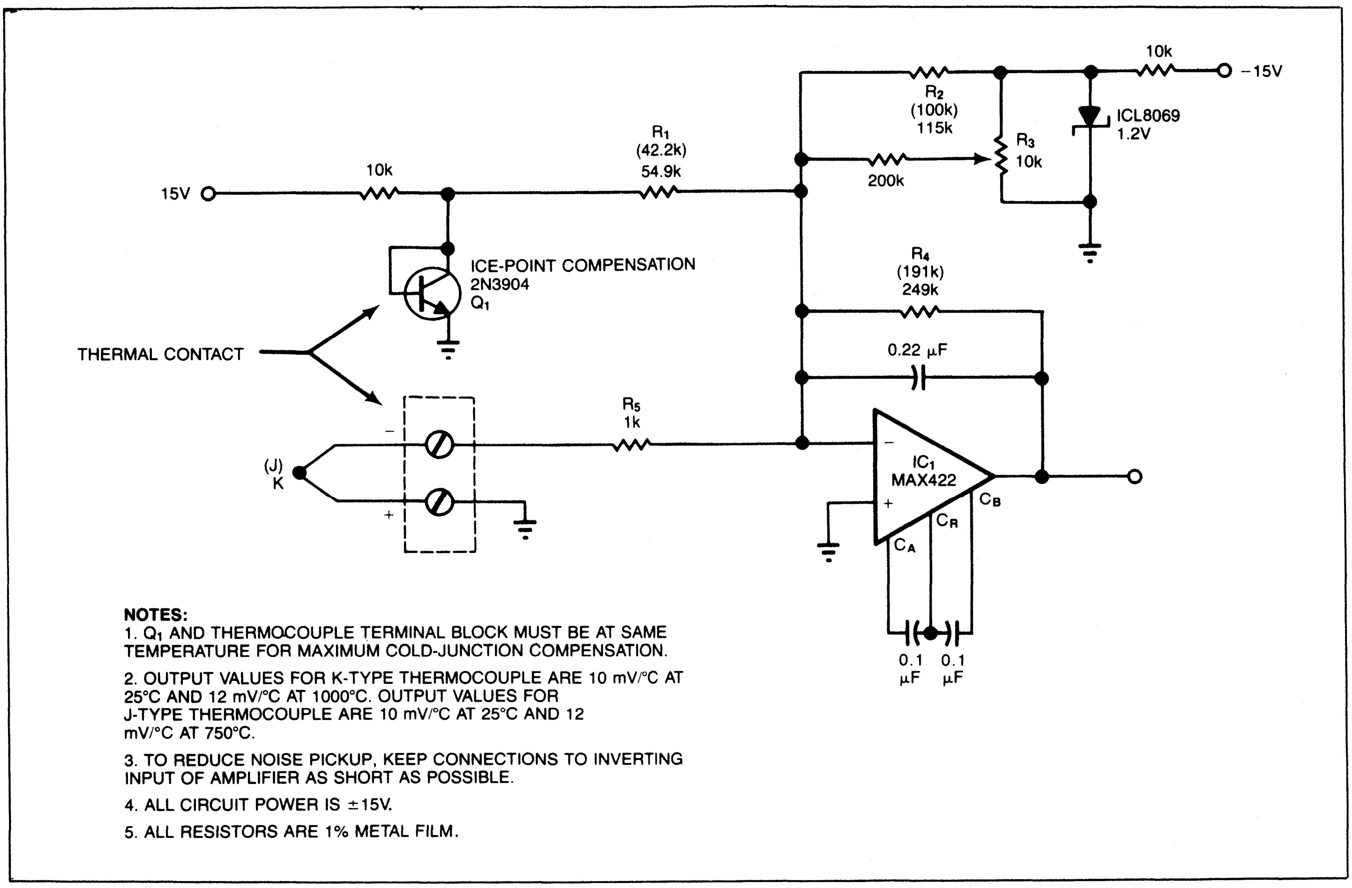


Fig 1—This simple signal-conditioning circuit *uses an inverting summer to combine thermocouple output, cold-junction compensation, and cold-junction offset signals. The circuit is easy to trim, because the gain and cold-junction offset adjustments do not interact.*

−2.2 mV/°C signal, which cancels the thermoelectric error signals generated at the thermocouple's input terminal strip. You should place the transistor as close as possible to the terminal block; to achieve optimum performance, you need to effect thermal contact between the devices.

For temperatures below 200°C (with a common type K thermocouple), the circuit in **Fig 1** provides a 10 mV/°C output. The circuit includes no linearization, so this output factor will change at higher temperatures—12 mV/°C (for type K thermocouples) at 1000°C. Although the circuit component values shown are for J- and K-type thermocouples only, the circuit can accommodate other types of thermocouples. To provide gain and cold-junction compensation, you simply have to calculate new values for R_1, R_2, and R_4.

Platinum resistance thermometers (PRTs) present another classical transducer-conditioning problem. By using platinum wire as the sensing element in a PRT, you can achieve very high accuracy and repeatability over a wide operating-temperature range. As a result, PRTs are well suited for high-accuracy thermometry applications. Like thermocouples, PRTs have a major drawback, however: They provide a low-level output signal. Although the change in resistance over temperature of platinum wire is very predictable, it's also very small (0.3815%/°C), so you'll require large precise gain and long-term stability in order to develop a useful output.

The PRT amplifier circuit in **Fig 2** uses a 3-terminal sensing scheme to eliminate errors from lead resistance, so you can remotely locate the sensor. In addition, the circuit design refers the sensor output to ground, thus minimizing noise-pickup problems. With a 30V supply, the circuit provides a 4- to 20-mA current output with compliance from 3 to 28V.

The REF-01 10V reference combines with IC_1 to generate a precise constant current that biases the PRT and also creates fixed voltages for sensor-offset correction and wire-resistance error cancellation. Lead-resistance effects (RW_1, RW_2, and RW_3) are subtracted from the real temperature signal at the input of the PRT amplifier, IC_2. The circuit in **Fig 2** works on the principle that the resistance of all three leads will be identical. The lead-resistance effects related to RW_2 are insignificant, however, because only op-amp bias current passes through RW_2.

The temperature and correction signals sum at IC_2 as follows: The voltages on the PRT and its ground wire (RW_3) are amplified by 1, the voltage drop on the PRT's positive lead (RW_1) is amplified by $-½$, and the drop across R_4 (an offset resistor) and RW_1 is amplified by $-½$. As a result, the RW terms cancel, and the net output appearing across R_3 is the PRT voltage minus ½

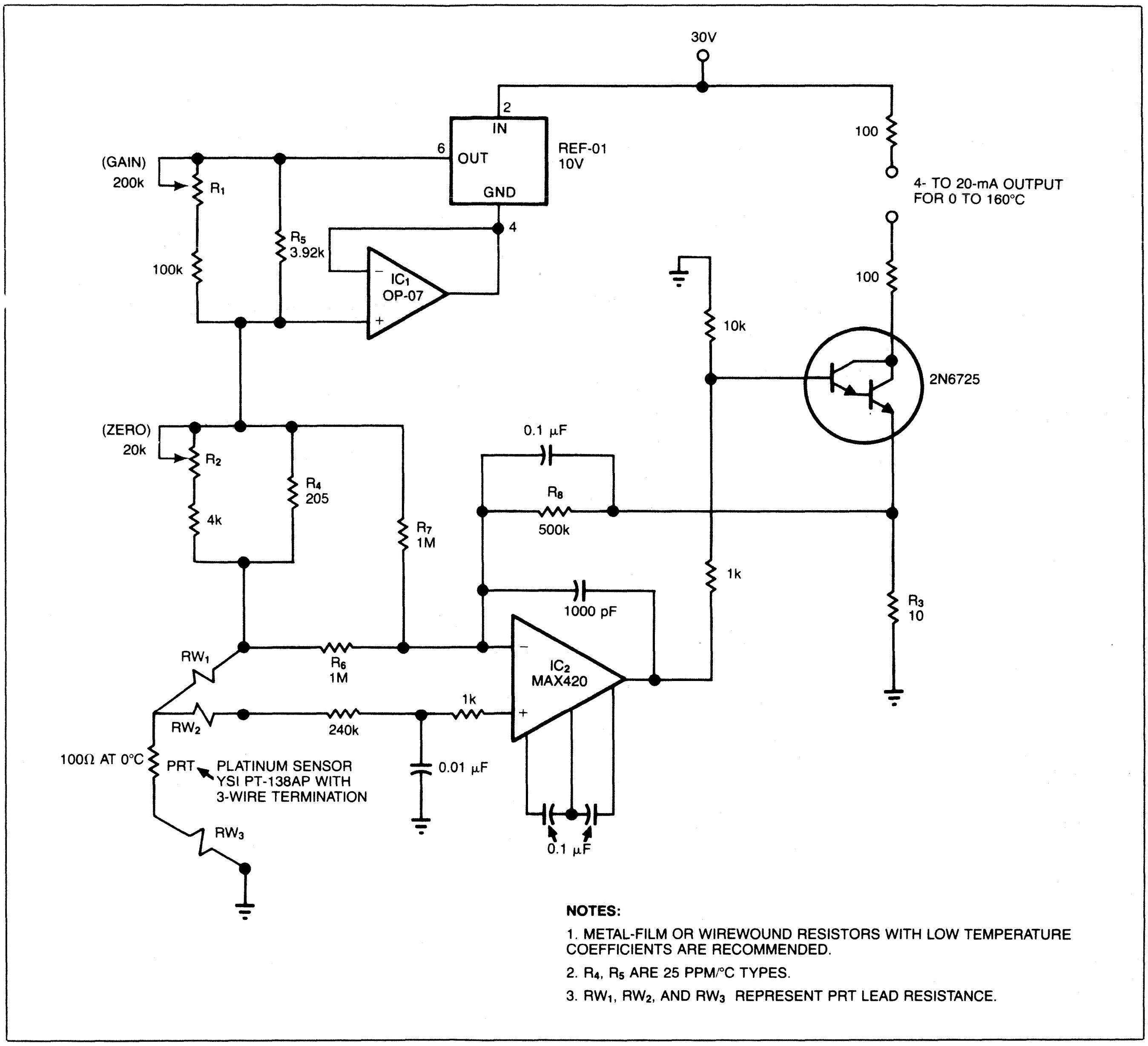


*Fig 2—**By referring the sensor to circuit ground,** this platinum-resistance-thermometer amplifier circuit minimizes noise-pickup problems. Because the circuit uses a 3-terminal sensing scheme to eliminate errors from lead resistance, you can remotely locate the sensor.*

Because they require high measurement accuracy and low signal levels, bridge-type transducers usually need precise amplification.

the offset on R_4. IC_2 drives a Darlington transistor (Q_1); R_3 senses Q_1's output current to provide a feedback signal for IC_2.

To calibrate the circuit, you must adjust R_2 and R_1. R_2 trims the sensor's offset, and R_1 handles circuit gain adjustments. If you adjust the offset first, the gain trim will not interact, so you can probably make each adjustment in only one pass.

Better precision for instrumentation amps

Precision amplifiers are usually a necessity in applications involving bridge measurements (strain gauges, load cells, and some types of pressure transducers), because these applications require high accuracy and low signal levels. In most cases, instrumentation amplifiers can easily handle the 30-mV differential output signals from these bridge-type devices. These instrumentation amplifiers have finite, controllable, differential gain that's fixed or that can be set with one or two resistors.

The high-performance instrumentation-amplifier circuit in **Fig 3** amplifies a small differential signal from a strain-gauge bridge into a large ground-referenced signal. Such a configuration is typical of off-the-shelf instrumentation amplifiers; however, when you use MAX421 amplifiers in the front end, the offset and drift performance you obtain is better by an order of magnitude than that available in off-the-shelf amplifiers.

As with all instrumentation amplifiers employing this 3-stage configuration, the front-end and output amplifiers will both affect overall drift performance. The output amplifier's effect on drift (referred to the input) is divided by the front-end gain, which is approximately 30 (the overall gain is 300). The MAXOP07's V_{OS} specifications (75 μV and 1.3 μV/°C) are divided and added to two times the MAX421 error, yielding a maximum input-referred error of 12.5 μV and 0.15 μV/°C. Even when front-end gain is set at 30, the MAXOP07 contributes more to offset error than do the MAX421s.

The chopping circuits of the front-end op amps are locked together via their clock-control pins. IC_1's INT/EXT clock pin is unconnected, so it operates from its internal 400-Hz clock in normal fashion. The CLK IN pin of IC_1 functions as an output. Since IC_2's INT/EXT

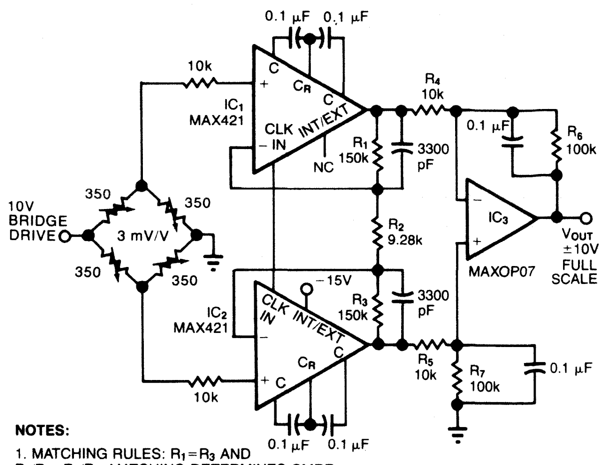
Addressing thermal problems

In an instrumentation amplifier circuit—as in any design dealing with low-level signals—the quest for microvolt-offset and nanovolt-drift performance involves more than just selecting a high-precision amplifier. When you're trying to amplify low-level signals, any number of outside error sources can complicate your task. These errors are troublesome because it's very hard to distinguish them from real signals or amplifier error.

Thermoelectric voltages provide a perfect example of such error sources. The same phenomenon responsible for thermocouple operation can generate significant errors at pin-to-socket, socket-to-board, and board-to-edge-connector interfaces, and even at soldered connections. The level of the voltage generated in such situations can range from one-tenth to tens of microvolts per degree C. In general, designers deal with this problem by minimizing the use of sockets and connectors in low-level circuitry, or by using components designed for low thermal EMF.

Although temperature obviously contributes to thermoelectric errors, thermal gradients in low-level circuitry cause more problems than does the mere presence of heat. Gradients can, for example, cause the normally balanced input connections of a sensitive amplifier to be at different temperatures. These connections then generate different values of thermoelectric voltages that the amplifier's inputs can no longer completely cancel; the final output is an offset error. The most effective way to combat thermal gradients is to keep power dissipation low and minimize air currents in and around low-level circuitry and connections.

You can also solve thermoelectric voltage problems by designing thermal symmetry into the circuit layout. This solution can involve adding dummy resistors and connections so that the thermal mass—as well as the number of thermoelectric error sources in an input pair—will cancel. In addition, you might have to run input traces close to each other and keep their dimensions identical. You might also find it helpful to develop some thermal filtering by minimizing enclosure size, or by insulating sensitive areas.



NOTES:

1. MATCHING RULES: $R_1=R_3$ AND $R_6/R_7=R_4/R_5$. MATCHING DETERMINES CMRR.
2. METAL-FILM OR WIREWOUND RESISTORS ARE RECOMMENDED.
3. IC_1'S INTERNAL CLOCK IS SLAVED TO IC_2 VIA CLK IN PINS.
4. ALL AMPLIFIERS OPERATE FROM ±15V SUPPLIES.
5. R_2 IS A SELECTED VALUE.

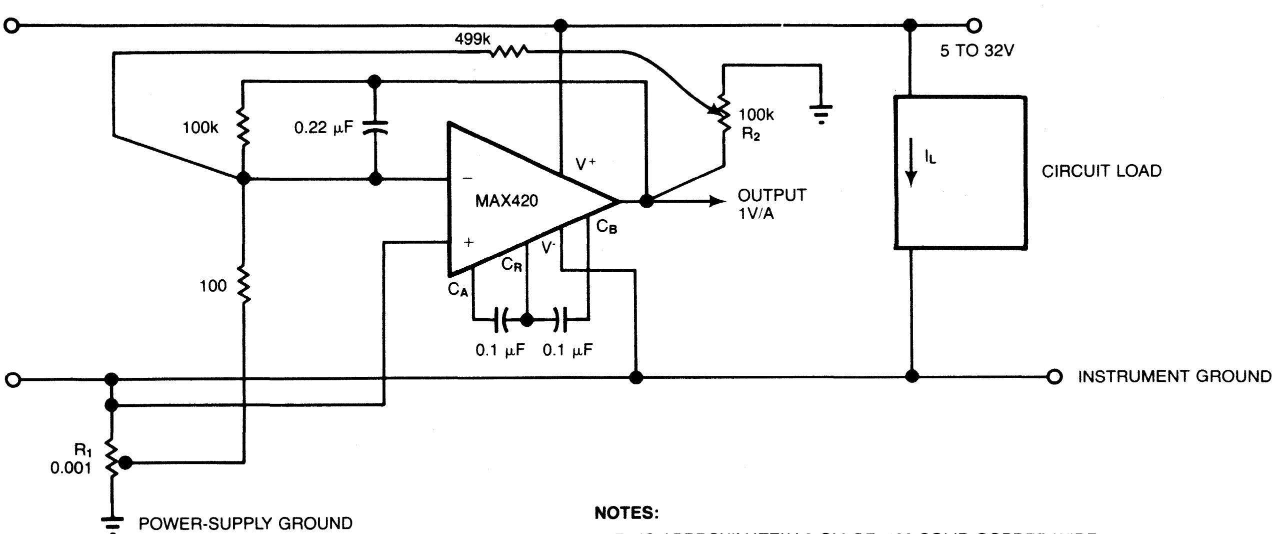
Fig 3—To eliminate the low-level errors caused by clock interaction, *this instrumentation-amplifier circuit locks the chopping circuits of the two front-end op amps together.*

CLK pin is connected to the negative supply, its clock is disabled. IC_1's CLK IN terminal drives IC_2. By synchronizing the input amplifiers in this manner, you eliminate low-level errors caused by clock interaction.

Gain-setting resistor matching limits the amount of common-mode rejection that you can realize. For prototype or test purposes, you can generally achieve 0.1 to 0.01% matching when you use selected 1% resistors. For more precise matching, you'll have to use resistor arrays or resistors that have low temperature coefficients and tighter tolerances. Once you've satisfied your rejection needs, you can adjust R_2 to change overall circuit gain without affecting common-mode-rejection performance.

Amplifying high-level signals

If you're designing circuitry to handle high-current measurements, you'll typically need to use sense resistors, so you'll have to contend with increased power-supply source impedance, and possibly even high power dissipation. Because it uses a low-offset amplifier in a high-current application, the circuit in **Fig 4** eliminates these concerns.



NOTES:

1. R_1 IS APPROXIMATELY 3 CM OF #20 SOLID COPPER WIRE. YOU DETERMINE SHUNT CONTACT POINT AT CALIBRATION TIME. TO AVOID TEMPERATURE-RELATED RESISTANCE CHANGES, DO NOT SOLDER SHUNT CONTACT CONNECTION.
2. USE METAL-FILM DEVICES FOR ALL RESISTORS EXCEPT R_1.

Fig 4—You can make sense measurements *without disturbing the operating load circuitry in this current-sense amplifier circuit because of the low offset specifications of the MAX420 op amp.*

Although temperature obviously contributes to thermoelectric errors, thermal gradients can cause more problems than can the mere presence of heat.

The MAX420 is suitable for this circuit for two reasons: First, because of the MAX420's low offset-voltage specification, you can use a low-value sense resistor (0.001Ω in **Fig 4**). The current measurement will, therefore, have no adverse effects on the operating load circuitry. Second, the MAX420 has a common-mode input range that includes the negative supply, which is circuit ground in this case. When the input range includes ground, the op amp can read the current-sense voltage across R_1 without the need for level shifting. Further, the MAX420 allows the circuit to operate from supply voltages of 5 to 33V.

In **Fig 4,** the sense resistor is actually a short piece of solid copper wire with a movable tap. To calibrate the circuit, you apply a known full-scale current and use the tap as a coarse trim to establish the proper output voltage. You then adjust R_2 to set amplifier gain and develop the precise output-voltage level. If you solder the tap on R_1, be sure to wait until the connection cools completely before you make the fine trim; this way, you'll avoid introducing errors caused by any change in wire resistance at high temperatures.

The circuit in **Fig 4** requires no offset adjustments. With the values shown, the circuit delivers 1V per amp of supply current. You can, however, change the sense-resistor value or the amplifier gain to develop other output ranges. With a 10A supply current, the load will experience a shift in ground potential of only 10 mV, and the sense resistor will dissipate only 0.1W.

You don't need to limit monolithic chopper amplifiers to applications involving low-level signal amplification. Because devices like the MAX420 perform precise amplification, you can use them not only in the primary signal path, but also to stabilize other circuitry. In effect, this approach lets you use CMOS op amps to improve the dc performance of wide-bandwidth or high-power circuitry.

The high-speed 12-bit power D/A converter shown in **Fig 5a** provides a good example of such an application. This circuit uses an LH0101 power op amp (300-kHz power bandwidth and 2A drive capability) as an output stage. The LH0101 has a 15-mV offset-voltage specification, so it can't accommodate 12-bit converter resolution by itself. The MAX421 overcomes the problem by monitoring the LH0101's inverting input and driving the noninverting input so that the summing junction is at 0V (to within 5 μV, the 421's error spec). The result is a high-speed, no-drift DAC circuit.

The MAX421 can use the same power supplies as do the LH0101 and AD565. The voltage divider at the MAX421's output attenuates the LH0101's correction signal to avoid any overdrive problems. Addition of the offset correction has no noticeable effect on the circuit's dynamic performance. **Fig 5b** shows step responses obtained with and without the correction circuitry in operation. The stabilized and unstabilized waveforms exhibit no perceptible slewing or settling-time differences as a result.

MAX420 amplifiers

MAX420 Series amplifiers offer very low zero-offset and zero-offset-drift specifications (5 μV and 0.05 μV/°C, respectively). The parts can operate over a 5V (or ±2.5V) to 33V (or ±16.5V) supply-voltage range. In addition, input bias current for the op amps specs at only 30 pA.

The amplifiers provide low-power operation at any supply voltage, and they offer FET-type bias currents (30 pA) for high-impedance measurements. The series also includes two devices (MAX422 and MAX423) that draw only 25% of the supply current that the standard parts draw, but that don't sacrifice dc performance. The two parts do, however, exhibit some decrease in bandwidth and output drive capability.

The 8-pin versions of the family are compatible with standard op-amp footprints with respect to inputs, outputs, and supply lines. To enable the amplifiers, you simply connect two external capacitors to the pins that conventional amplifiers normally use for offset trimming.

Maximize voltage calibrator performance

To design a voltage calibrator that can generate a 0 to 10.0000V output with 100-μV resolution **(Fig 6),** you must use an op amp with very good offset, drift, and common-mode rejection specs. Although the circuit is designed for reliable absolute-reference stability, it has a ratiometric capability: Both fixed and variable references, based on the same source, are available simultaneously. Such a feature is especially useful in applications involving linearity checks of digital voltmeters or A/D converters. In such cases, relative rather than absolute results have the most significance. Very few off-the-shelf calibrators, if any, provide this dual-reference feature, so if you want such a feature you must build your own calibrator.

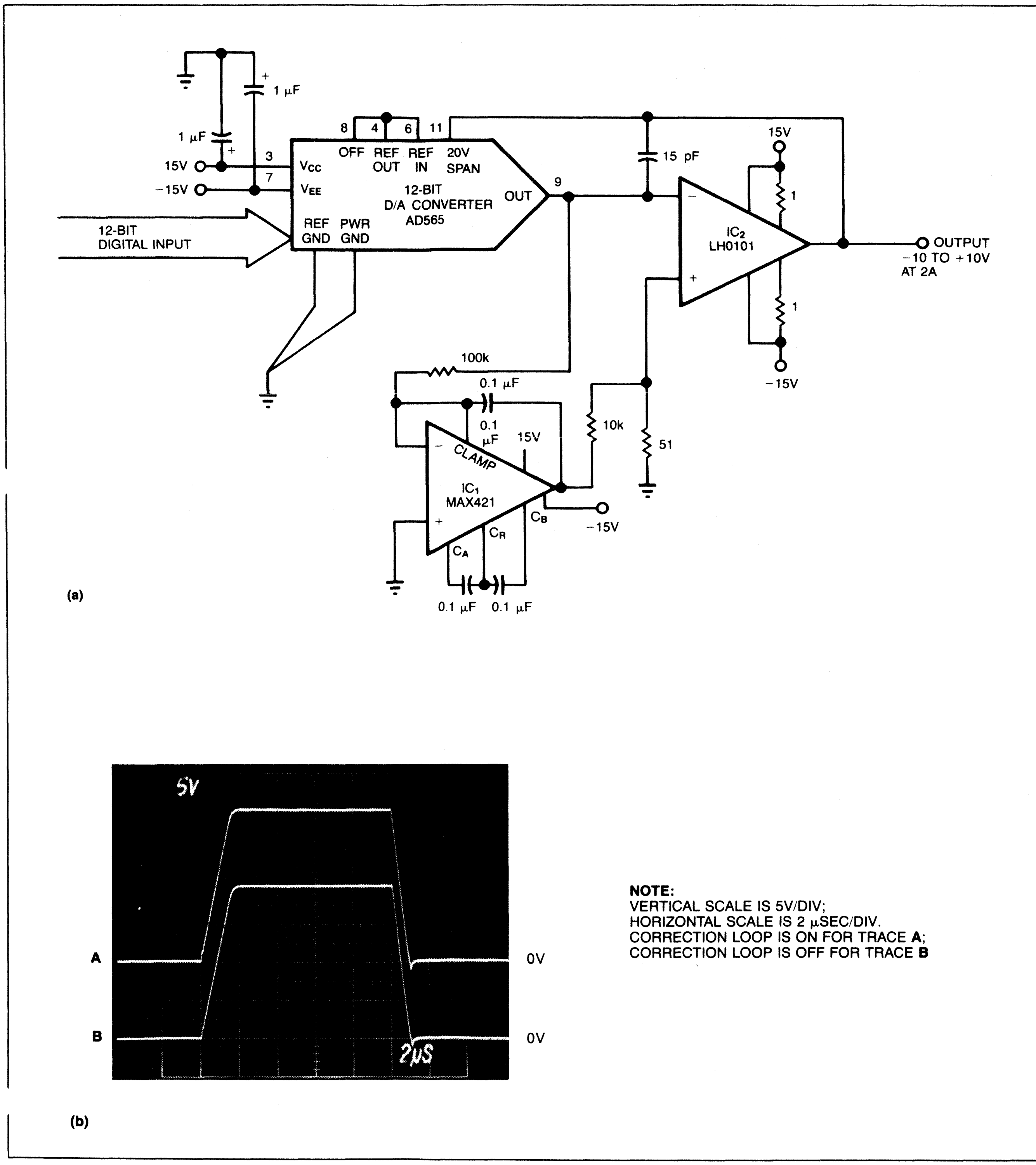
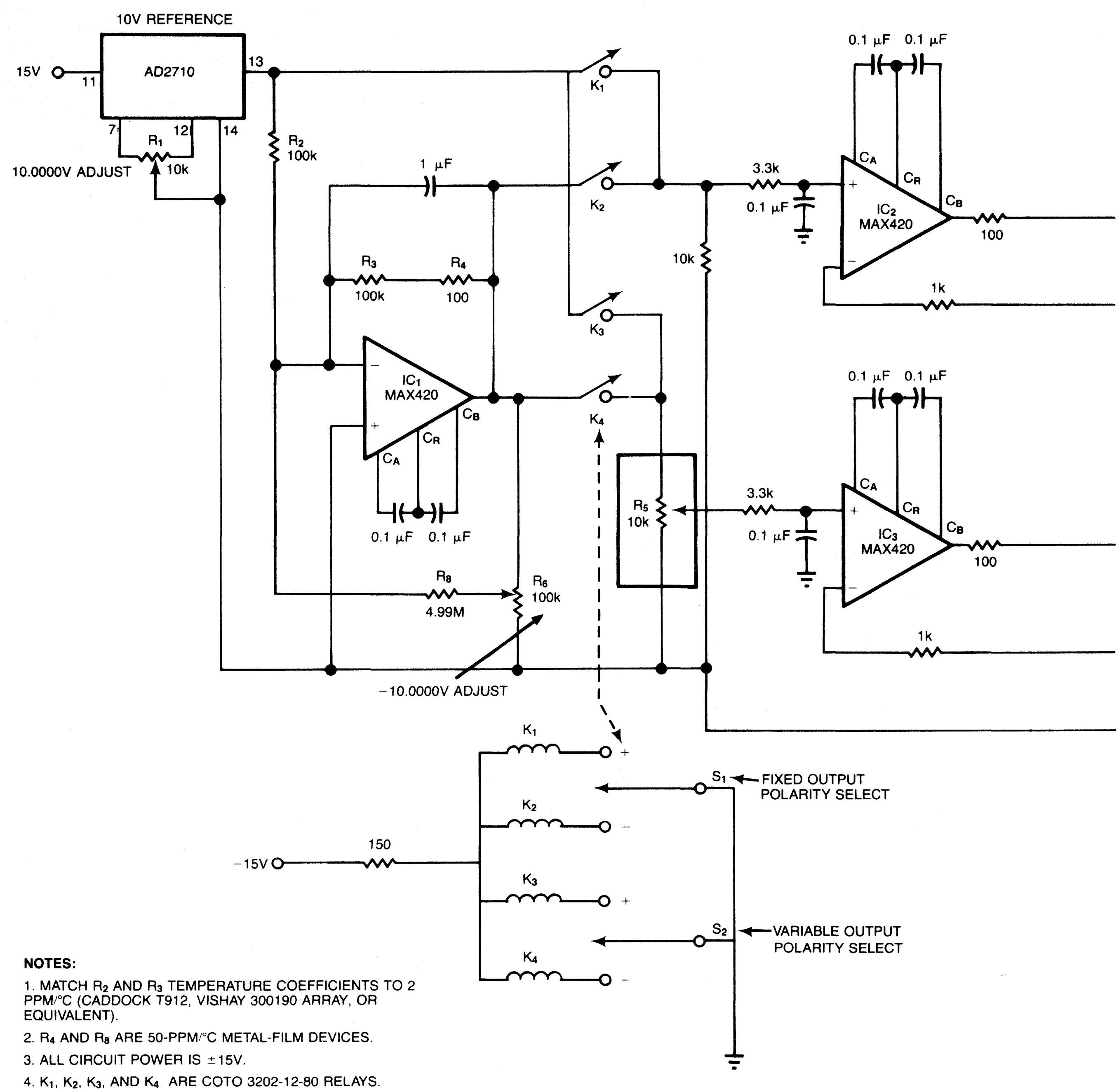


Fig 5—Able to satisfy wide-bandwidth or high-power requirements, *this D/A converter circuit* ***(a)*** *employs the MAX421 for offset correction. As the scope photo shows* ***(b)****, the correction scheme has no noticeable effect on the circuit's dynamic performance.*

Monolithic choppers aren't limited to low-level signal-amplification tasks; they can also provide dc stabilization for other circuitry.



NOTES:

1. MATCH R_2 AND R_3 TEMPERATURE COEFFICIENTS TO 2 PPM/°C (CADDOCK T912, VISHAY 300190 ARRAY, OR EQUIVALENT).
2. R_4 AND R_8 ARE 50-PPM/°C METAL-FILM DEVICES.
3. ALL CIRCUIT POWER IS ±15V.
4. K_1, K_2, K_3, AND K_4 ARE COTO 3202-12-80 RELAYS.
5. R_5 IS AN ESI PRECISION VOLTAGE DIVIDER (DP1311).

***Fig 6—Because it provides a ratiometric measurement capability,** this voltage-calibrator circuit is useful for checking the linearity of digital voltmeters or A/D converters, two tasks for which relative results are more significant than absolute results.*

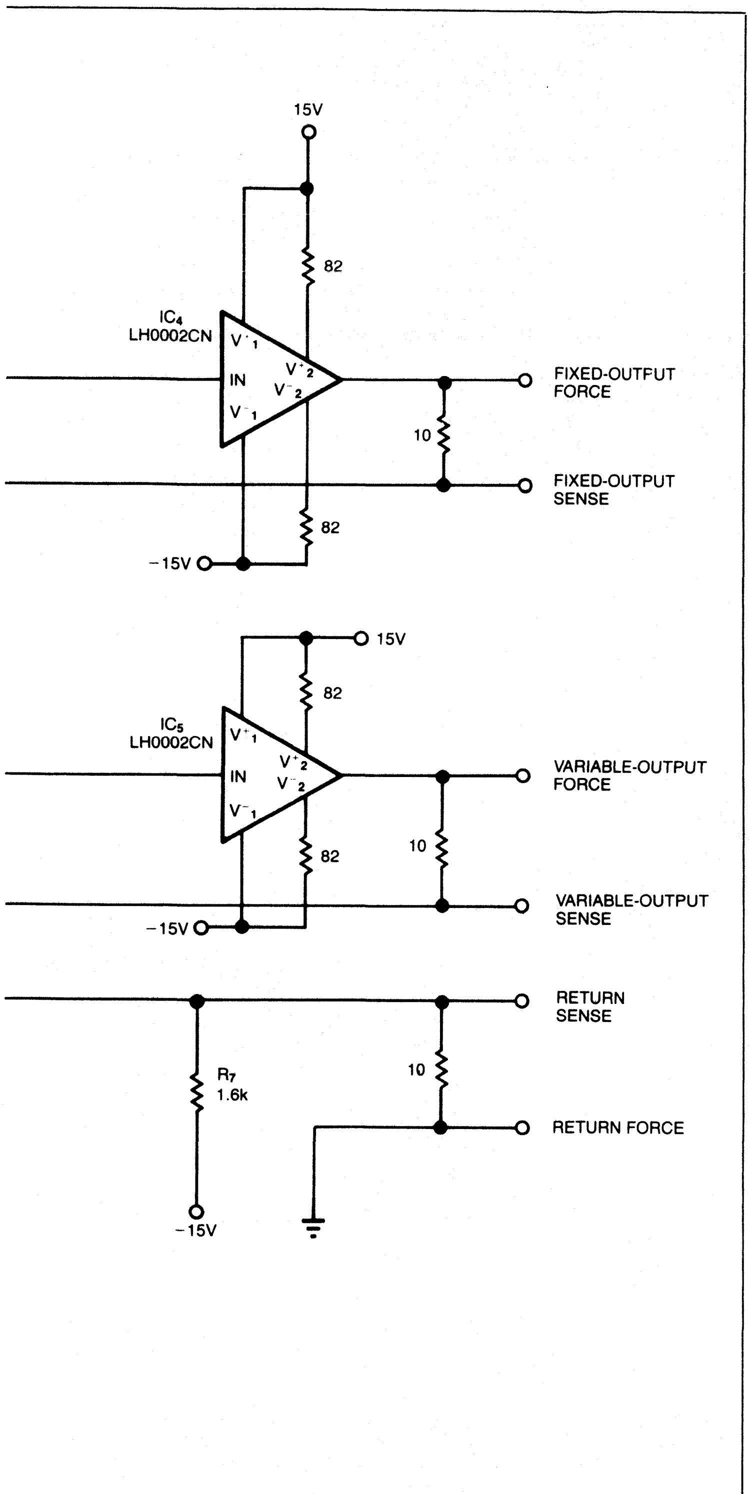


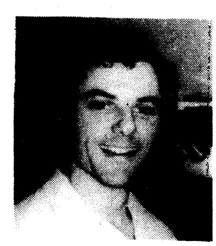
Fig 6's circuit uses an AD2710 voltage reference, which the manufacturer can trim to within 1 mV. If you're going to use the circuit for purely ratiometric measurement, a 1-mV (or even larger) error, and some degree of drift, will cause few problems. For absolute calibration applications, however, you'll require an accurate, low-drift reference. You can adjust the 2710 for a tighter tolerance by using a 10-kΩ multiturn trimmer, as shown in **Fig 6.**

IC_1, an MAX420 connected as a precision unity-gain inverter, provides a negative version of the reference voltage that's adjusted with trimmer R_6. Special reed relays, designed for minimal thermal EMF errors, select either the positive or the negative reference to drive the fixed and variable outputs.

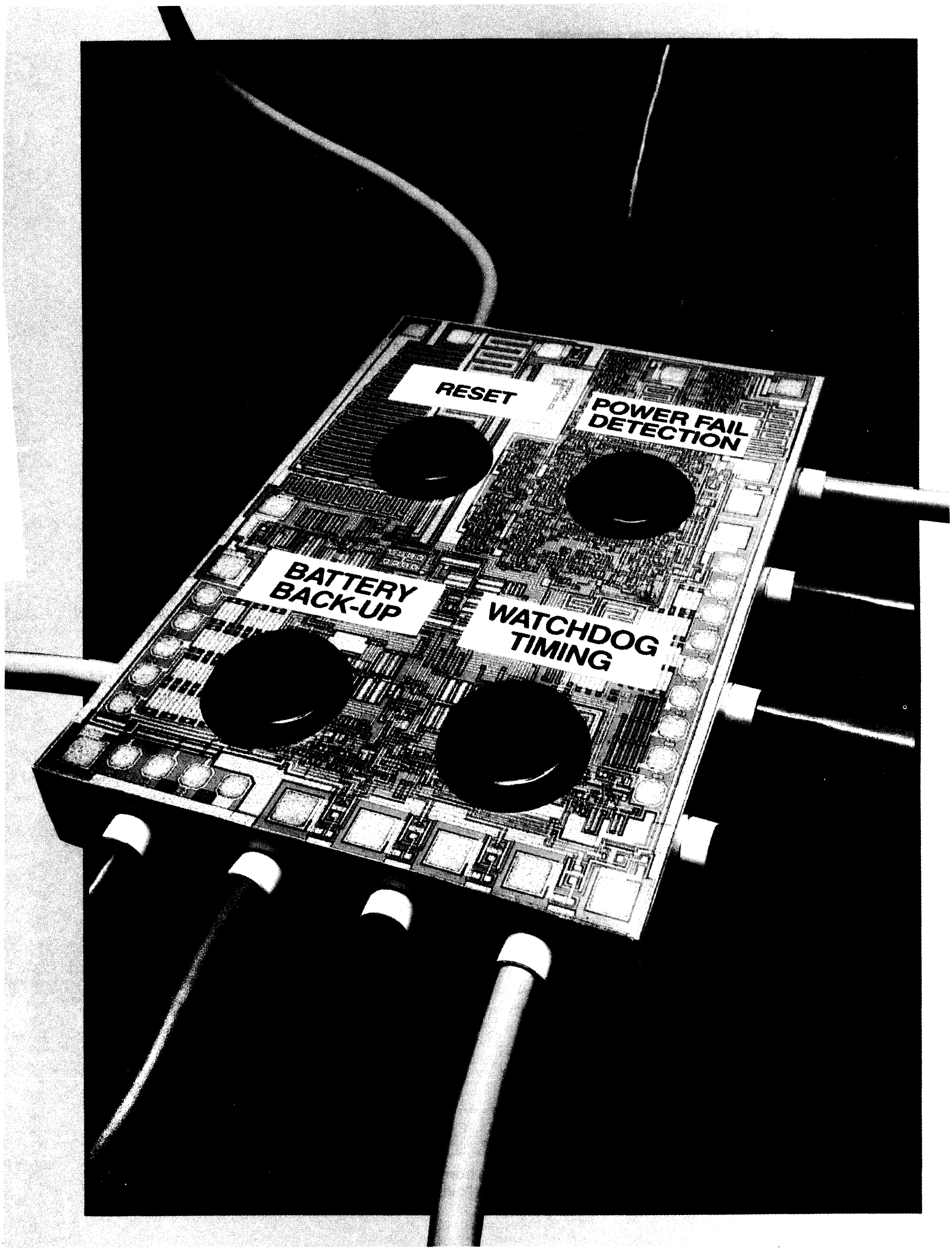
It's not a good idea to use a conventional switch to perform the polarity-select function, because conventional switch contacts generate small thermoelectric voltages. These voltages can introduce significant errors when the output resolution is as low as 100 μV/step, as it is in **Fig 6.** By using four relays, you can select the polarity of the fixed and variable outputs independently.

A Kelvin-Varley voltage divider (R_5), with five decades of adjustment range, divides the fixed reference, developing output increments as small as 100 μV with 20-ppm linearity. The fixed and variable output buffers (IC_2 through IC_5) are composites of MAX420 op amps and LH0101 buffers. This combination provides reasonably good output current drive, and it has less than 10 μV of untrimmed error from offset, common-mode rejection, and other sources. **EDN**

Author's biography



Leonard Sherman is a senior member of the technical applications staff at Maxim Integrated Products (Sunnyvale, CA). In this position, he gets involved in product planning, generates applications literature, and provides customer support. Len has a BSEE degree from Massachusetts Institute of Technology and has been granted one patent. In his spare time, he collects old hi-fi equipment, rates pickles, and watches other people repair automobiles.


RESET
POWER FAIL DETECTION
BATTERY BACK-UP
WATCHDOG TIMING

Analog supervisor chip keeps microprocessor out of trouble

Charlie Allen
Maxim Integrated Products, 510 N. Pastoria Ave., Sunnyvale, CA 94086; (408) 737-7600.

Goal: to sharply reduce the complexity, component count, and space of standard microprocessor supervisory circuits while improving the accuracy and reliability of reset and battery-switchover circuits. Add to that a watchdog timer and write protection. The result: the MAX691 and MAX690 microprocessor supervisory ICs, which pack many common housekeeping functions into 16-pin and 8-pin packages.

To protect processors, a mixed bag of blocks, including threshold detectors, a power-switchover circuit, and a watchdog timer, unite on one IC.

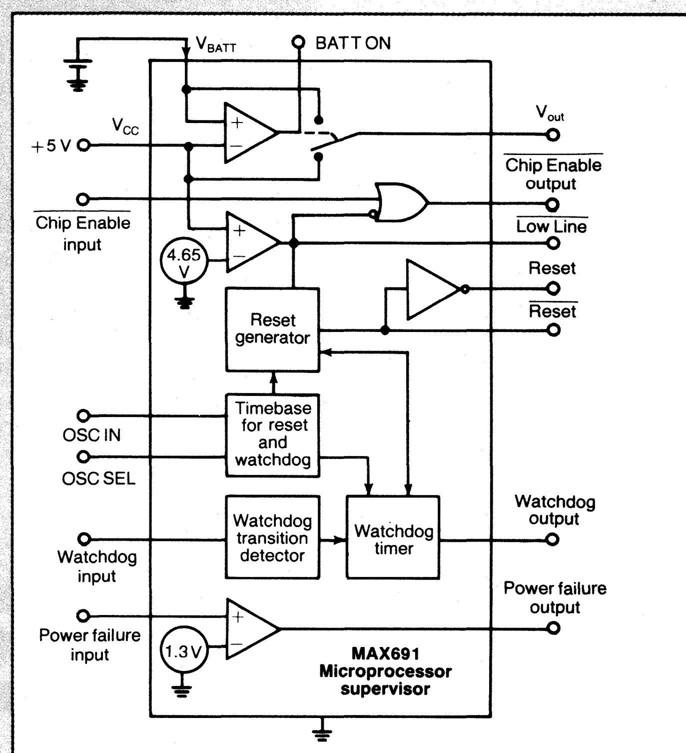
Because the needs of microprocessor systems defined the ICs, the supervisory circuits contain several seemingly distinct functions that microprocessors commonly need (Fig. 1). For one, the chips have a precise 4.65-V threshold detector and a 50-ms timer that generate an accurate reset signal for any power-up, power-down, brownout, or momentary-interrupt condition. For another, power-switchover circuitry offers a battery backup for CMOS RAMs or real-time clocks. On top of that, an uncommitted 1.3-V threshold comparator can serve as a power-failure warning indicator or monitor for the backup battery.

To keep the microprocessor from writing incorrect data into a CMOS RAM or EEPROM, during power-up, power-down, or a brownout, the MAX691 has chip-enabling circuitry. The circuit forces a Chip Select output enable high whenever the +5-V supply falls below 4.65 V. Moreover, the chip's watchdog timer monitors software execution, resetting the microprocessor if execution is disrupted for any reason.

The reset and watchdog time-out periods have three pin-selectable timing options. The first is based on an on-chip oscillator and needs no external components. The second uses an external clock. Finally, the designer can choose the minimum reset pulse width and watchdog time-out period by tuning the oscillator with one external capacitor.

The eight-pin MAX690 has fewer timing options for the watchdog timer and reset-pulse width and no chip-enable gating circuitry.

The reset function relies on a reference-voltage technique with much more reliability than simple RC circuits, which work well if a change occurs abruptly between completely off and +5 V. But if the power is slowly applied or lost, or if a brownout or momentary loss occurs, the RC circuit does not properly reset the system. The reset section contains a bandgap reference, a voltage divider that establishes the 4.65-V reset detection threshold, and a



1. The basic functions provided by Maxim's MAX691 microprocessor supervisory IC are battery back-up for V_{CC}, reset pulses, and watchdog timing to ensure proper operation.

50-ms retriggerable monostable oscillator.

The voltage detector forces the Reset output low whenever the +5-V V_{CC} input drops below 4.5 V. The circuit holds Reset low until the supply stays above 4.75 V for 50 ms, ensuring that the microprocessor receives the minimum reset pulse width specified by the manufacturer. The designer can extend this pulse width by adding a capacitor to the oscillator input pin or shorten it by overdriving the oscillator input. The voltage detector also serves a second purpose: forcing the Chip Enable output high whenever the +5-V input is incorrect.

Although the 4.65-V threshold is a nominal value, the 4.5-V minimum and 4.75-V maximum are guaranteed because the thresholds are trimmed at the wafer level. Trimming involves fusible metal links like those on bipolar PROMs. A Low Line output offers an instantaneous view of the voltage-detector status. The Reset output, however, goes high only after the input voltage has been at a valid level for 50 ms.

The Reset pin offers a "weak output," which works as an active CMOS output driving high or low or as a wired-OR output when connected to open-collector outputs. The pin can also supply the pullup current needed by the open-collector outputs connected to the bus. This dual capability is possible because the pin is a CMOS output but with only a limited source-current capacity.

When connected to a high-impedance input, the weak output drives with full CMOS output swings of ground to V_{CC}. But when connected to an open-collector or wired-OR bus, the limited source-current output lets the weak output be pulled down by any device with a sink current of only 10 μA. Another benefit is the output's guaranteed minimum 1-μA pullup current, which eliminates the need for an external pullup resistor on a wired-OR bus.

The battery-switchover circuitry, although shown in the block diagram as merely an SPDT switch, is more complicated than that. In practice, a voltage comparator checks V_{CC} against the battery voltage. As long as the supply is higher than the battery, V_{CC} connects to the output terminal, V_{out}. The connection, however, is through a pnp transistor whose base-drive modulation circuitry saturates the transistor while minimizing base current.

SHORT-CIRCUIT PROTECTION TOO

When V_{CC} falls to within 100 mV of the battery voltage, the circuit turns off the base drive, and a 400-Ω MOS switch connects the battery to V_{out}. With the MOS switch the voltage drop is about 4 mV, compared to about 500 mV across the diode switch used in a discrete circuit typically found in battery-backup circuits. The base-drive circuitry includes a thermal-shutdown circuit that reduces base drive if the junction temperature is too high, thereby lending short-circuit protection to V_{out}.

With V_{CC} present, battery current is a maximum of ± 1 μA and typically is a 10-nA charging current. This level extends battery life while remaining within the allowable charge current of even the smallest lithium batteries.

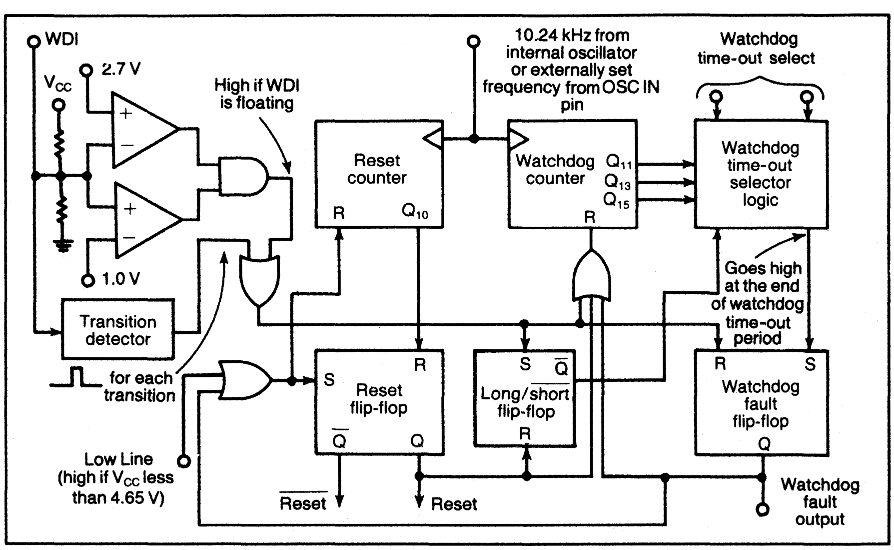
When V_{CC} falls to about 700 mV below battery voltage, a second comparator shuts down all circuitry not needed in the battery-backup mode. In this low-quiescent-current state, the typical draw on the battery is about 600 nA, ensuring a long battery life.

The chip's Battery On pin, which has a sink current of 5 mA, indicates whether V_{CC} or the battery is powering V_{out}. It can directly drive the base of an external pnp transistor if the battery-backed power bus needs more than 50 mA of output current during normal operation.

An internal pnp transistor, meanwhile, is guaranteed to have a maximum voltage drop of 300 mV at 50 mA, a level that will run several CMOS RAMs if their Chip Enable inputs function to reduce the average current. A 0.1-μF bypass capacitor at the output supplies the high-current spikes the RAMs draw for a few nanoseconds during each access cycle.

Besides working with batteries, the chip's power-switchover circuitry supplies short-term power backup with standard capacitors or the new farad-size units.

The third major section of the chip is the watchdog timer, a function often desired but seldom found (see



2. The chip's watchdog circuitry looks for periodic strobes from the program. If no strobe is received in the allotted time, the Watchdog Fault Output, WDO, goes high, and Reset issues a 50-ms reset pulse.

"Watchdog On Guard," p. 108). The circuit monitors system execution, detecting hardware, and software faults. But because not all microprocessor systems can strobe the watchdog timer, the chip automatically disables the timer if the Watchdog Input, WDI, is left floating. When that input floats, two internal resistors bias it at an invalid logic level detected by internal voltage comparators (Fig. 2).

On the job, the microprocessor drives the WDI pin with an I/O line that is periodically strobed. At each strobe, the supervisory chip resets the timer. If the microprocessor fails to strobe WDI for the programmed time period, a watchdog fault occurs. Typical reasons for a failure to strobe include hardware problems, a temporary disruption caused by voltage transients, and software errors that put the microprocessor in an endless loop.

ALARM ACTIVATED

If a watchdog fault occurs, the chip sends a 50-ms reset pulse and the Watchdog Fault Output, WFO, goes low. The output remains low, and the chip periodically pulses the Reset line until WDI is again strobed. For the greatest reliability, the Watchdog Fault output can activate an alarm and trigger hardware that places the equipment controlled by the microprocessor into a fail-safe mode.

The designer can set the time-out period in several ways. One of two internally preset time delays can be selected with a logic level on the Oscillator Input pin while the Oscillator Select pin is high. Slow-response systems, such as instruments that update displays, usually can tolerate errors for several seconds. In this type of system, the nominal 1.6-s watchdog time-out suffices.

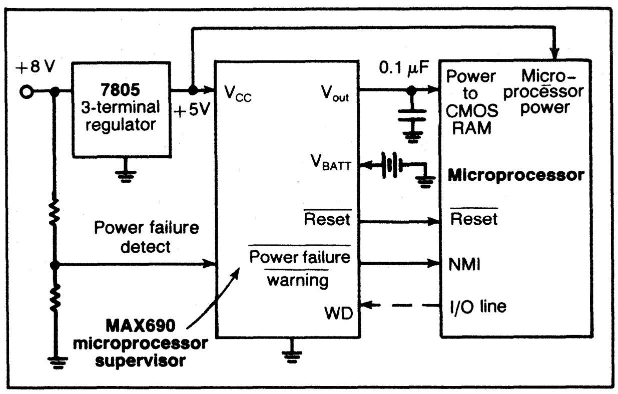
On the other hand, an airplane autopilot or a programmable controller must quickly reset after a malfunction. Such systems usually need a 100-ms or faster time-out period. Since most systems have special power-up routines that run after a reset, the chip waits up to 1.6 s for them, even if the 100-ms time-out period is chosen.

A designer has two ways to externally set the exact time-out period. One way is to overdrive the Oscillator Input pin with an external clock. Because the watchdog timer is an oscillator and a counter rather than the RC monostable circuit, a 200-ns pulse can reset the timer. The designer can then use a decoded Write signal to drive WDI.

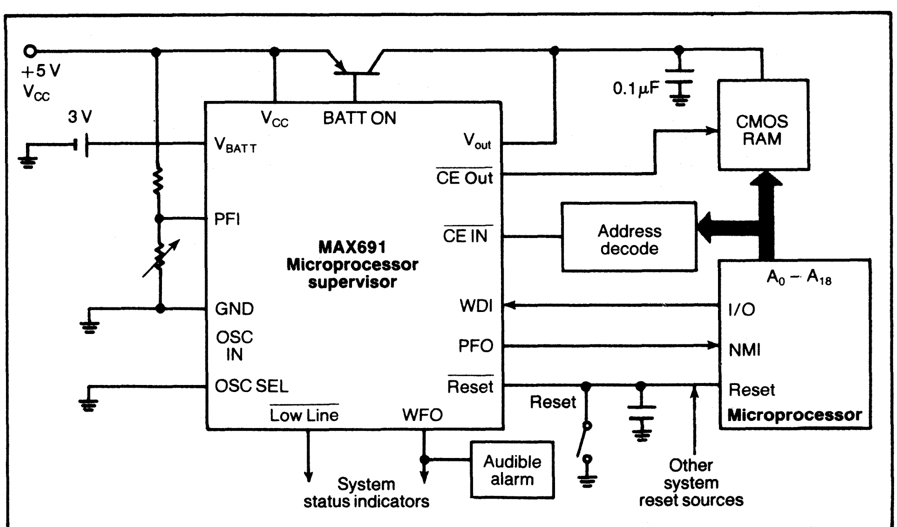
The alternative is to select the external clock mode and connect a capacitor to the Oscillator Input. The internal oscillator operates at 10.24 kHz when Oscillator Select is high and at about 20 kHz when it is low and the Oscillator Input is floating. The latter situation creates a watchdog time-out of 50 ms. Adding capacitance to the Oscillator Input reduces its frequency and stretches the time-out period. For example, a 2-nF capacitance lengthens the time-out to 10 s.

Typically, the MAX690 monitors V_{CC} for a one-board controller system and generates a 50-ms reset pulse when needed (Fig. 3). An +8-V dc source fed to the 7805 three-terminal regulator is scaled by two external resistors and applied to the chip's Power Fail Warning detector, which has a 1.3-V threshold.

As a result, when the raw dc falls below 8 V, the supervisory chip's Power Fail Output goes low, interrupting the microprocessor through its nonmaskable interrupt



3. In a circuit with a three-terminal voltage regulator, the MAX690—a simplified version of the 691—sends a power-failure warning to the microprocessor as soon as the raw 8-V dc begins to drop.



4. With its Oscillator Select pin tied low and Oscillator Input pin left floating the MAX691 has a 25-ms reset pulse, a 50-ms watchdog period during normal operation, and a 200-ms watchdog period after any reset.

(NMI). If the voltage continues to fall, the regulator's 5-V output will begin to drop. When it reaches 4.65 V, the supervisory chip asserts Reset, halting the microprocessor and preventing further access to the RAM. As V_{CC} approaches the battery voltage, the chip replaces V_{CC} with the battery at V_{out}, ensuring continued power to the CMOS RAM. When V_{CC} falls to 700 mV below the battery, the chip shuts down all of its own unneeded circuitry, reducing its current drain to 600 nA.

RAM POWER RESTORED

When the dc voltage begins to build up again, the sequence reverses, except that Reset stays low for 50 ms after V_{CC} rises above 4.65 V. When 5 V is again present, V_{out} restores power to the RAM, guaranteeing a maximum voltage drop of 300 mV between V_{CC} and V_{out} at an average 50-mA current.

A more sophisticated system based on the MAX691 has the Battery On output driving an external transistor to increase power on the battery-backed bus to 250 mA (Fig. 4). Also, the Watchdog Fault Output sounds an external alarm if recovery from a malfunction is impossible. In this system, if V_{CC} is at an invalid level, the Chip Enable gating circuit blocks write cycles to the CMOS RAM or real-time clock, preventing the microprocessor from corrupting the data in the RAM during power-up, power-down, or a brownout.

Tying Oscillator Select low and leaving Oscillator Input floating reduces the reset time-out to 25 ms. This configuration also sets the watchdog time-out to 50 ms during normal operations and 200 ms immediately after a reset.

PRICE AND AVAILABILITY

The commercial version of the MAX690, in a plastic package, costs $3.30 each, and the similarly packaged MAX691 costs $3.80 each, both in quantities of 100. Production quantities will be available in June.

CIRCLE 503

The reset bus contains a manual switch and a 0.1-μF capacitor that make possible manual system resets. This circuit contains a simple method of generating a power-failure warning, needing only the +5-V input and two resistors. The resistors set up a warning threshold of 4.8 V. The processor has only the time it takes V_{CC} to drop from 4.8 to 4.65 V to save any data into RAM.

Because all the sections of the MAX690 and 691 are independent, any unnecessary functions can be ignored. For example, a designer who does not want battery backup and power-failure warning can eliminate the battery and the two resistors and connect the battery and Power Failure Inputs to ground. If the watchdog timer is not desired, the Watchdog Input is left floating.□

Charlie Allen has been applications engineering manager at Maxim for three years. Before joining Maxim, he did similar work at Intersil. Allen has a BSEE from Michigan State University.

Watchdog on guard

Nearly every designer has operated equipment that can lock up, needing a reset to restore normal operation. The watchdog circuit's job is to detect such malfunctions and to automatically issue reset commands.

Performance of this task can take many forms. The most sophisticated is a separately executing microprocessor that independently calculates results and compares them with the master computer's actions. At the lower end are schemes that demand, within a specified time, a response to an interrupt issued by the watchdog circuit. In between are circuits that issue a pseudorandom number or bit stream that the microprocessor must match to avoid a fault indication.

The most common watchdog circuit, though, checks the microcomputer's operation by monitoring an I/O line that is toggled under program control. This scheme assumes that if the processor periodically toggles the I/O line properly, then it is correctly executing the program. Such a circuit can tell if the microprocessor is stuck in a loop so long as the loop does not toggle the I/O line.

A watchdog circuit needs software, the simplest being a few lines of code that toggle the I/O line. The code must be in a section of software that executes frequently enough that the time between toggles is less than the watchdog time-out period. Typically, the watchdog software is in the system or supervisory software, either in a section that responds to a periodic interrupt or in the section that executes when the system is idle.

Another common software technique controls the I/O line from two sections of the program. The software might set the I/O line high while operating in the foreground mode and set it low while in the background or an interrupt mode. Then both modes must execute correctly or else the watchdog timer issues a reset pulse.

Chip-level converters administer local dc dosages

Whether operated from a battery or an off-line power supply, dc/dc converter ICs are an economical alternative to multiple-voltage power supplies

David Bingham, Staff Scientist
Maxim Integrated Products

The availability of standalone power-supply chips will revolutionize the conventional method of structuring power-supply systems. Increasingly, the ubiquitous +5 V will solo as the only bussed voltage while the less common and lower secondary dc voltages are generated where they are needed by dc/dc converter ICs. For the moment, these chips and their menage of external parts will remain board-level circuitry, but the industry is gravitating toward generating the secondary voltages right on the chips that need them.

Putting an end to overkill

Almost without exception, multiple dc power-supply voltages are required in electronic equipment, whether computers, household appliances, industrial data acquisition systems, or military mobile radios. There is typically one dominant dc voltage for the digital electronics and one or more secondary dc voltages for the analog circuitry. An obvious example of such a system is a computer in which the processor, memory, and other logic circuitry are operated from a +5-V supply and some of the interface circuitry (both analog and digital) is run off various other dc voltages such as +10 V, +12 V, ±15 V, and so on.

In the past, the most cost-effective way to obtain the multiple dc voltages has been to generate them in the system's main power supply at power levels that exceed what is actually needed for many applications. Today's more cost-effective strategy involves conscripting dc/dc converter ICs to convert the main (bussed) dc power-supply voltage into the required secondary voltage(s). Typically, the converter ICs are mounted on the same board with the circuitry that requires the nonstandard power-supply voltages. It's a practical maneuver thanks to a

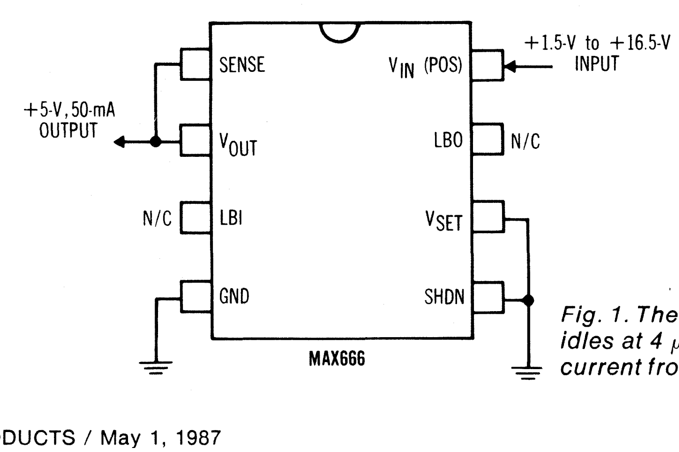


Fig. 1. The MAX666 is a CMOS linear regulator that idles at 4 μA yet can deliver an output current from 5 μA to 50 mA.

new series of CMOS dc/dc converter ICs that Maxim recently introduced. Intended for low- to medium-power applications, each member of this new chip battalion integrates most of the components or glue that was needed by earlier-generation converters. This results in a low component-count design, high power-conversion efficiency, simplicity of use, and low cost. Because of their high efficiencies in particular, these novel dc/dc power-supply circuits are a wise choice for battery-operated military systems—or for any other battery-operated system for that matter.

Linear, switcher, or charge pump

Maxim's family includes the three basic types of dc/dc converter circuits: the linear regulator, switcher, and charge pump. Before choosing between them, designers should be aware of the advantages and disadvantages of each, as outlined in *Table 1.* Because none can fulfill the needs of every application, each has found its niche—a situation unlikely to change for some time to come.

The linear regulator in particular was the first to show up extensively in electronic equipment. It has a fine pedigree, initially being integrated onto silicon in the mid 1960s by Fairchild Semiconductor in the now famous μA723. Subsequent generations of linear regulator ICs have extended the level of available output power and, because of this higher integration, have culminated in three-terminal regulator ICs that require few, if any, external components.

More recently, CMOS technology has brought about a reduction in the quiescent (idling) current of linear regulators. For example, Maxim's MAX663 and MAX666 ICs have slashed this current to 4 μA typ so that the only real power loss occurs in output pass transistors. In addition, the MAX666 has an output current that can range from 5 μA to 50 mA, and includes alluring goodies like shutdown, low-voltage detection, and output current limiting (see *Fig. 1*). A new feature, dual mode operation, proffers selection of either a fixed 5-V output or—with the addition of two external resistors—any output voltage from 2 to 15 V.

In general, linear regulator ICs are most suitable for the battery operation of equipment, such as running a 5-V system from a 9-V battery. *Figure 2* shows a battery-backup power supply using a MAX666 and a 9-V transistor radio battery. Resistor R_S senses the unregulated dc from the line supply and inhibits the output of the MAX-666. A detector informs the system if the backup battery is malfunctioning. The quiescent current drawn from the battery is typically 3 to 4 μA. If a nickel cadmium stack is used, then the optional charging circuit of R_1 and diode D_2 may be added.

Linear regulators are also prudent where noise suppression is important. Other uses include the generation of say +10 V or +12 V from an existing +15-V supply at the board level rather than at the main power supply.

The two drawbacks of the linear regulator are its low power-conversion efficiency and the fact that its output voltage is always less than its input voltage. If the ratio approaches unity, then power conversion efficiencies will be proportionately high. For an efficient regulator such as the MAX663 the power conversion efficiency, assuming equal input and output currents, is:

$$\text{Efficiency} = \frac{V_{OUT}}{V_{IN}} \times 100\%$$

This works out to an efficiency of approximately 50%, assuming, say, a 10-V input and a 5-V output.

Switching regulators are more efficient than linears. If a switching regulator is used to convert, say, +15 V into +5 V, then the conversion efficiency will be about 80%. The figure for a linear regulator is 33% using the formula shown (5/15 V) x 100%. Hence, switchers are called upon extensively for high-efficiency dc/dc power conversion,

Table 1. Comparison of Dc/dc Converters

Feature	Linear regulation	Charge pump	Switcher
Output voltage regulation	Excellent	Poor	Excellent
Voltage step-up	No	Yes, but only in discrete multiples of input voltages	Yes
Voltage step-down	Yes	Yes, but only in discrete fractions of input voltage	Yes
Current step-up	No	Yes, but only in discrete fractions of input current	Yes
Power conversion efficiency	Poor	Excellent - can approach 100%	Good 80%
High power control	Yes	Not suitable	Excellent
Inductor required	No	No	Yes
Output noise	Low	Includes clock frequency plus harmonics	Includes clock frequency plus harmonics, hf hash
Minimum of external components	0	2 capacitors	1 inductor 1 capacitor

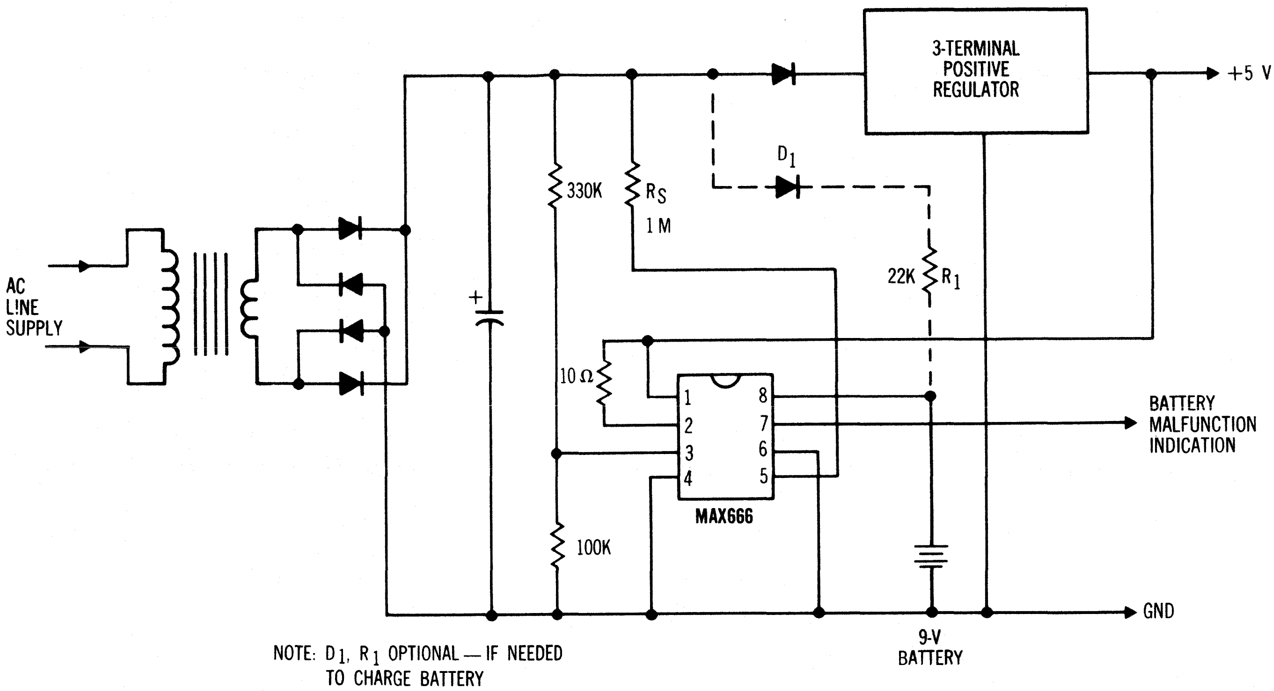


Fig. 2. This uninterruptible +5-V power supply is constructed around a MAX666 and a 9-V battery. If ac power is lost, then the MAX666 will convert the battery voltage into +5 V. The chip's low-battery detector circuitry informs the system if the backup battery is malfunctioning.

and they have the added assets of being smaller and cheaper. (Note that off-line switching power supplies are basically dc/dc converters since the ac line voltage, after being rectified and smoothed somewhat, is connected to the input of a dc/dc converter from which the desired output voltage is obtained). Also, switchers are much more versatile than both linear regulators and charge pumps and are particularly fine where stepup and voltage inversion with reasonable precision is required. (They do, however, cost more than linear regulator ICs.)

A switching power converter, unlike the linear regulator, operates in a discontinuous mode. In its simplest configuration—a two-phase mode—energy is taken from an input power source and stored in the magnetic field of an inductor in one phase. At the end of this phase a current, I, will be flowing in the inductor, L, which will have a stored energy of ½ LI^2. At the beginning of the second phase, the inductor is reconnected through a switch to the output reservoir capacitor. The current will have a value of I at the beginning of the second phase and will either approach or decrease to zero at the end of it. Energy will thus be transferred from the magnetic field of the inductor to the reservoir capacitor. The function of this capacitor is to store the packets of energy received and produce a constant voltage during each complete period of operation. Thus the switcher requires at least two external storage elements—an inductor and a capacitor—because large amounts of energy still cannot be stored on a monolithic IC.

Although the internal operation of the switcher is discontinuous, as viewed from the output load it appears to be continuous. Furthermore, the switcher is extremely flexible in function because of the diversity of ways—called topologies—that the inductor(s) can be connected to the input supply and to the output.

Among switching regulator ICs, the MAX631 is typical of what can now be achieved using conventional CMOS (see *Fig. 3*). Its utility lies

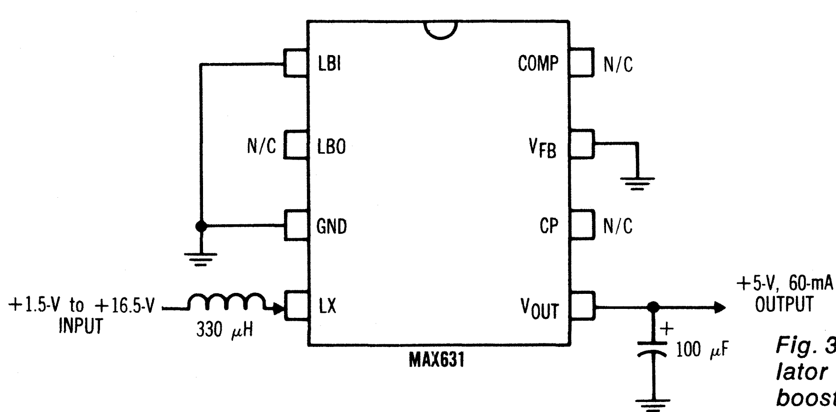
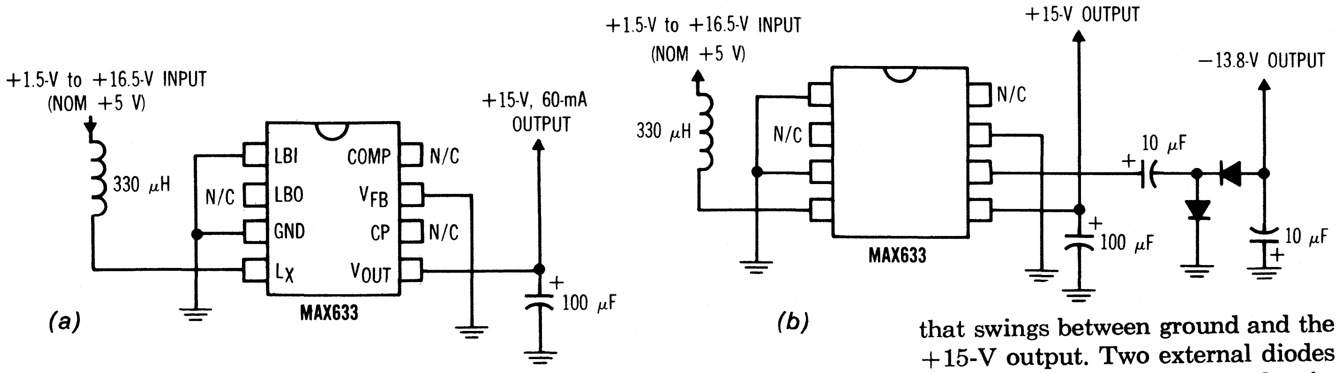


Fig. 3. The MAX631 switching regulator serves well as a voltage booster. It can supply a +5-V output voltage from an input voltage as low as +1.5 V.



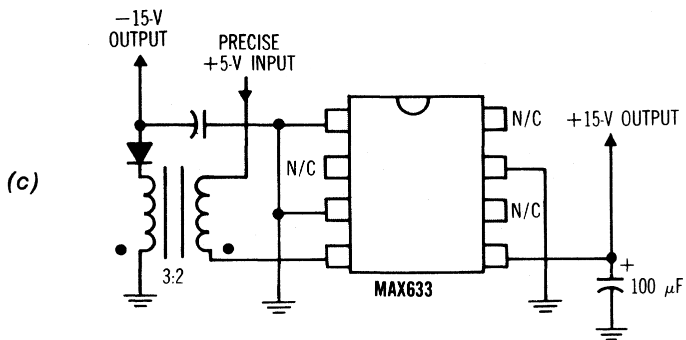


Fig. 4. Boosting a nominal input of +5V to +15 V is an easy job for the MAX633 switcher (a). If a second—but less accurate—higher voltage is needed, then the circuit shown in (b) can be used to derive both +15 and −13.8 V from a nominal +5-V input. Dual ±15-V supplies can be had by using a more precise +5-V input (c).

in its ability to boost a voltage as low as 1.5 V up to +5 V. All of the necessary circuit components except those storing energy (the inductor and reservoir capacitor) are contained on chip. From an applications point of view the MAX631 switcher duplicates the niceties of the MAX666 linear converter. Both have low-voltage detectors, operate with minimum quiescent currents, and feature dual mode operation.

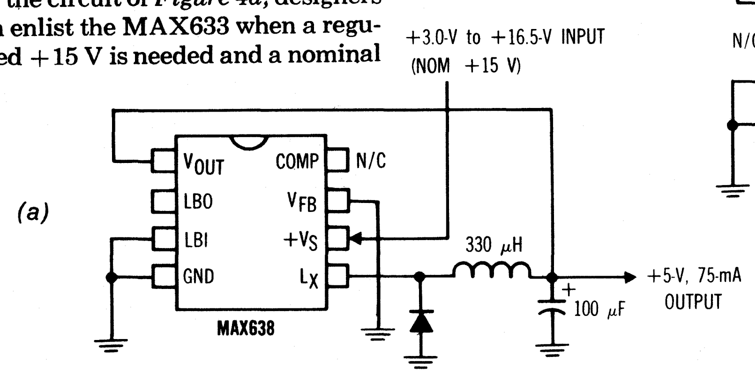
The MAX632 and MAX633 are sister chips to the MAX631 and can be drafted to boost a 1.5-V input to +12 and +15 V, respectively. Using the circuit of *Figure 4a,* designers can enlist the MAX633 when a regulated +15 V is needed and a nominal +5 V is available. The only additional components are an inductor and a capacitor. This same circuit could also operate from a two-cell battery stack to provide +5 V using the MAX631.

Generating a negative voltage

To generate a crude −15 V in addition to +15 V, the circuit of *Figure 4b* can be put into service. A charge pump inverts the +15-V output (which also powers the MAX633 after the circuit has started up). It consists of an on-chip power inverter that swings between ground and the +15-V output. Two external diodes and two capacitors complete the circuit. The output voltage will be less than −15 V, however, because of the forward-voltage drop across each of the diodes. Instead, the negative output voltage will have an open-circuit value of approximately −13.8 V, and it will, of course, be unregulated.

Figure 4c shows how to generate both a +15-V and a −15-V supply by replacing the inductor with a transformer and adding another capacitor and a diode. The accuracy of the −15-V supply will not be as good as the +15-V supply but should be adequate for most applications. Note that this circuit requires a precise +5-V input since the negative output has a power-supply rejection ratio of less than unity.

Other CMOS dc/dc converters from Maxim provide stepdown and inverting outputs, together with 12- or 15-V dual-mode output voltages of either polarity. *Figure 5a* shows how to derive +5 V from an input voltage range of +5.5 to +16.5 V using the MAX638 stepdown or "buck" switcher. If −5 V is needed,



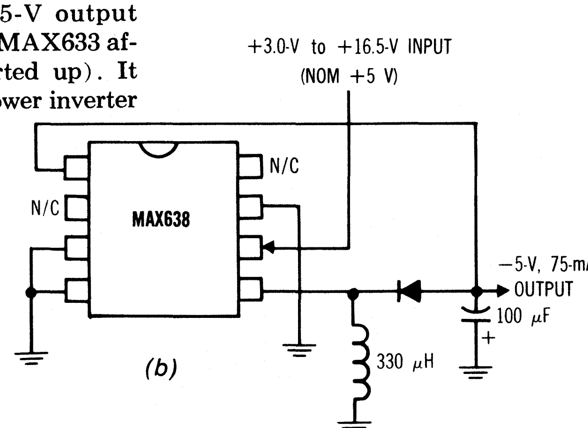


Fig. 5. The MAX638 stepdown switching converter can be conscripted when +5 V is needed and +15 V is available (a). It can also invert +5 V to −5 V (b).

Table 2. Comparison of Inverting Converters

Feature	ICL7660 charge pump	MAX638 inverting switcher
Output voltage accuracy—no load	2% typ	2% typ
Output impedance	60 Ω	1 Ω
Supply rejection	0 dB	70 dB
Useful output current	20 mA	75 mA
Maximum power-conversion efficiency	95% at 2 mA	85% at 35 mA

then the circuit arrangement of *Figure 5b* can be used instead.

One of the drawbacks of using a switcher as a low-power dc/dc converter is that an inductor is mandatory. Because inductors are the least common components in electronic equipment, designers have been reluctant to employ them. However, a wide range of inexpensive inductors from potted miniatures to small toroids are available. The potted variety resemble ¼-W resistors and can handle milliwatts of power, while the toroids are usable at levels of several watts.

The second switcher bugbear is the fact that it produces both electrical (conducted) noise and electromagnetic (radiated) noise. The noise at the output of a switcher and the noise fed back into the input supply will consist of its clock frequency plus harmonics and sometimes subharmonics. In some systems additional filtering may be needed to reduce this noise, thus upping the cost.

The third type of dc/dc converter, the charge pump, is really a modified voltage doubler—or multiplier—which has been around for 80 years. Like the switching regulator, the charge pump operates in a discontinuous mode that has two phases. In one, a flying capacitor is connected to the input supply. In the second, this capacitor is reconnected through switches to an output reservoir capacitor.

This output device serves the same function as the reservoir capacitor in the switcher—namely, to provide a continuous or dc output during both phases. The output voltage of a simple charge pump can therefore be either an inversion of the input voltage or a doubling or halving of the input, depending on how the interconnecting switches are configured. Recruiting more capacitors and switches yields any multiple or positive fraction of the input voltage. If both stepup and stepdown techniques are used together, then virtually any output voltage can be obtained.

The plus side of charge pumps includes high voltage and power conversion efficiencies, which can approach 100% at low power levels, and the need to add only two inexpensive capacitors. Demerits are a lack of output regulation, input voltage tracking, and the fact that they produce switching noise. In general, however, their noise problems are much less severe than the switchers', and in most applications it is not even a consideration.

Probably the most common home of charge pumps has been aboard n-channel RAMs. This circuit arrangement generates on chip a voltage of −3 to −4 V from the +5-V power-supply rail. The negative voltage can then back-bias the substrate, which effectively reduces junction capacitances to improve switching speeds.

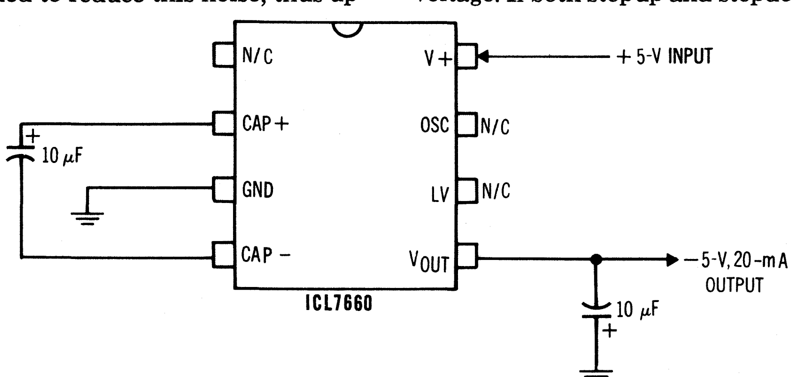


Fig. 6. The ICL7660 charge-pump converter, together with two external capacitors, can be employed to derive −5 V from a +5-V supply. However, it is a tracking converter and has an output that is poorly regulated.

Standalone converters

More recently, CMOS charge pumps have become available as standalone ICs, the trailblazer being the ICL7660. Since its introduction in the late '70s, the ICL7660 has been used by the millions for the local generation of a negative voltage from a system's main (positive) power source. It is most frequently called upon to provide −5 V from the main +5-V supply (see ***Fig. 6***).

Not surprisingly, power-supply designs employing the ICL7660 have been the lowest-cost solutions for interface drivers such as EIA RS-232-C and various analog circuits, including operational amplifiers, analog-to-digital converters, and so on. But the ICL7660 does have its faults, too. For one thing, its output is not well regulated. Secondly, it is a tracking converter: its output voltage changes proportionately with its input voltage. Also, it cannot supply more than about 20 mA of output current. Most of these limitations can be overcome by the use of a switching regulator such as the MAX638. ***Table 2*** compares the features of the MAX638 with those of the ICL7660.

New CMOS charge-pump ICs have been slow in arriving because of their susceptibility to latchup—a nuisance that occurs in a CMOS circuit when its inputs or outputs are forward biased, tripping the device into a high-conductance state similar to a conducting SCR. This condition can only be remedied by removing the input power supply—not a very practical solution. But latchup mechanisms are now much better understood by IC makers, who can take steps to guard against it. New charge-pump ICs such as the MAX-680 tap both positive and inverting pumps to supply ±10 V from a +5-V input. Looking ahead, the improved CMOS charge-pump circuitry will have a dramatic effect in another arena, as well.

Included on chip

Certain standard IC functions require an unusual—but not necessarily highly accurate—dc power, supply voltage(s). Take, for example, RS-232-C line drivers whose outputs must drive a load to approximately ±10 V. These voltages are not critical but must be within a range of ±5 to ±15 V. Increasingly, where the market for such an IC is sufficiently large and the economics amenable, chip makers are going to include a dc/dc power converter right on the same chip. Charge pumps are a practical choice for such converters owing to their high power-conversion efficiency, low power levels, and low cost.

The MAX232 pioneers a new series of ICs that include their own power supplies. The chip's two RS-232-C transceivers have been integrated along with a double charge pump to enable operation from a single +5-V supply. Four external capacitors—two for each charge pump—must be connected to the MAX232.

But external capacitors are not needed by the MAX233, which is a hybrid packed into a 20-pin DIP along with a modified MAX232 that uses a high on-chip chopper frequency plus a proprietary lead frame. □

Multiplexer-amp combo tames losses in wideband circuits

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Teleconferencing cameras, flash converters, video crosspoint switches, and attenuators can make good use of a monolithic video multiplexer chip with zero insertion loss, low input capacity, and 75-Ω line-driving capability. But such a chip—a multiplexer driving an op amp near unity closed-loop gain—did not exist until quite recently.

The problem is that putting the multiplexer and op amp on the same chip presented some difficult trade-offs: a CMOS multiplexer is best for switching with minimal power consumption, and a bipolar op amp is best for maximum gain-bandwidth product. So an all-bipolar chip with diode bridge multiplexer switches, for example, would need a lot of power, but an all-CMOS chip of conventional design would be too slow.

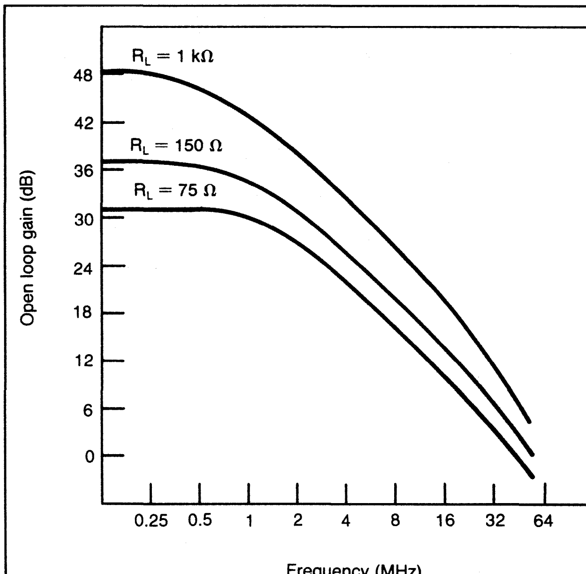
By making a slow CMOS op amp simulate a much faster bipolar one, a multiplexer-amplifier chip drives TV-camera switching chains.

The conventional solution—relying on two separate chips, say a 200-MHz CMOS multiplexer and a 100 to 300 MHz bipolar op amp—is both difficult and not very cost-effective. It is difficult to stabilize a very wideband op amp near unity closed-loop gain, and it is wasteful of bandpass.

Similarly, wideband multiplexers are all designed with low resistance (R), typically 50 Ω, to directly drive low impedance (75-Ω) cables without an amplifier. This low R means high channel capacitance (C), as much as 50 pF, which typically might limit the bandpass to 50 to 100 MHz because of input RC rolloff. Opting for a multiplexer with higher R and lower C—assuming one was available—would not be of much help. With the multiplexer and op amp only 0.25 in. apart, 2 to 3 pF of stray wiring capacitance would still limit the bandpass.

A monolithic solution to this designer's dilemma is Maxim's MAX455, a 50-MHz, 250 mW, ±5V, eight-channel multiplexer-amplifier. Other versions have four and two channels as well as the bare video amplifier, the MAX454, 453, and 452, respectively. While there are some TV applications in which a separate bipolar op amp and a special multiplexer may be necessary, the all-CMOS device supplies low input loading, easy stabilization against oscillation, and general cost-effectiveness.

The MAX455 design is far from conventional, since there is no way to get enough op amp bandwidth from CMOS MOSFETs with a 5-μm line width. The trick was to make a relatively low open-loop gain amplifier look like a multistage op amp when driving cable impedance and yet look like a single stage gain to prevent excessive phase shift and oscillation. This trick eliminates the need for phase compensation to prevent oscillation at low closed-loop gain. Without the need for compensation, the bandpass of the device is improved by a factor of 2 to 10 or more. However, voltage gain will



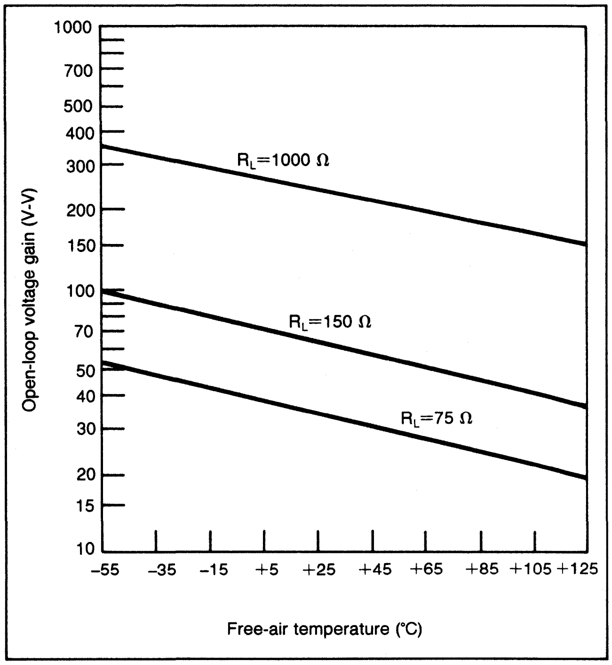
1. Both the open loop gain and the gain-bandpass increase with load, which becomes apparent when the MAX455, configured as an instrumentation amplifier with a 20-dB closed loop gain, drives a 1- kΩ load. The gain-bandpass product is 150 MHz.

be proportional to load resistance.

So instead of three amplifier stages (typical of an op amp) there are just two: a 38 V/V differential input stage driving a push-pull transconductance output stage with the help of current mirrors for biasing. The output stage has gain proportional to output R and equal to unity with a 75-Ω load. The low gain is not a drawback, since the amplifier usually runs at a closed-loop gain of 0 or 6 db; its dc transfer function is quite linear, and it is always driving a well-defined resistive load. The amplifier also behaves like a true op amp in terms of low drift, high common mode rejection and high power supply rejection. Typical MAX455 specs are offset voltage of 1 or 2 mV; offset input drift of 15 to 20 μV/°C; CMRR of 80 dB; power supply rejection of 66 dB at low frequencies and 20 dB at 4 MHz.

The amplifier is unity-gain stable when driving a 75-Ω load (±1 V maximum). With a gain of +6 dB, it is stable with 150 Ω (±2 V max), and at a gain of +20 dB, it is stable at 1-kΩ load. The gain-bandwidth increases with load resistance (Fig. 1). The gain also drops with increasing temperature (Fig. 2).

Unlike many fast video amplifiers, it is simple to get unity-gain stability even on a breadboard without using a ground plane. Ground planes are still recommended, since anyone who has ever worked with high-speed amplifiers knows the frustration of trying to stabilize them. Just bypass the MAX455 chip's supplies with 0.1 μF ceramic capacitors to ground and keep all leads short.



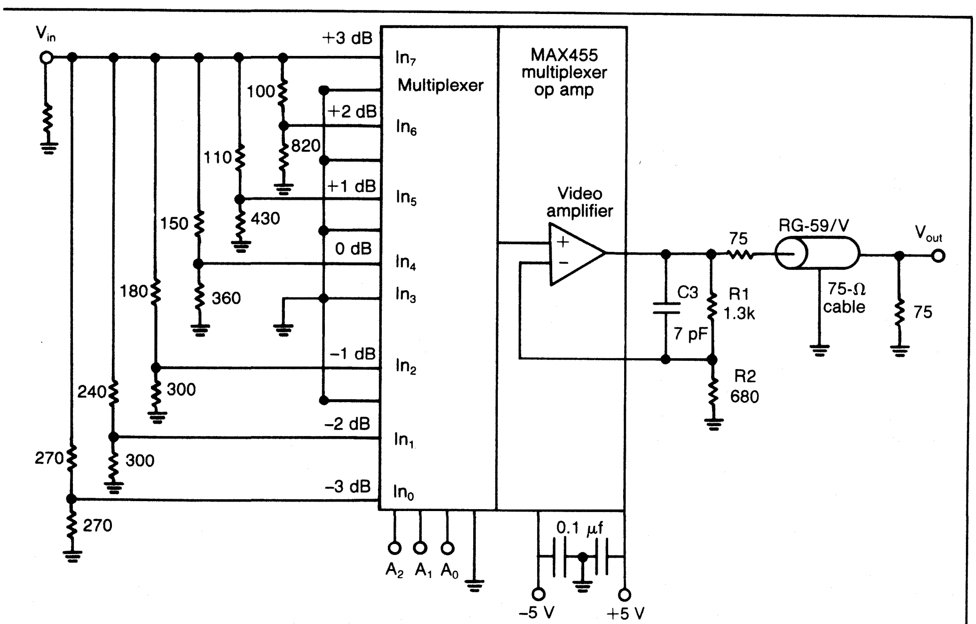
2. The gain of MOSFET transistors, however, drops with increasing temperature. A MAX455 chip with a closed-loop gain of 6 dB and driving a 150-Ω load would lose 1% of its gain when operating at +85°C instead of +25°C.

OPTIMIZING THE MULTIPLEXER

The chip is also much more stable than a bipolar amplifier when driving a high-capacity load typical of many flash comverters. With loads greater than 100 pF, however, a series R should be inserted serially at the output such that the RC product is 10 ns or more at unity gain.

The MAX455 loads the video cable far less than would a separate multiplexer-op amp combination since the multiplexer-out/amplifier is an internal node deliberately not brought out to a bonding path to keep parasitics to a minimum. This allows smaller switches, compared with those in a separate multiplexer, which keeps the channel input capacitance down to about 7 pF on and 3.5 pF off. The multiplexer switch areas were also held to a near-circular shape to minimize the ratio of on-resistance to capacitance.

The multiplexer-out/amplifier-in RC rolloff was held to 80 MHz—about equal to the uni-



3. This video amplifier-attenuator has 25-MHz typical bandpass and 70 dB off-channel isolation in the TV band. It provides seven steps from -3 to +3 dB in 1-dB increments, plus a no-output position. The resistors R1 and R2 set a gain of 9 dB, 6 of which cancel out the 6-dB loss in driving the back-terminated cable. C3 reduces peaking caused by the amplifier input capacity.

ty-gain amplifier bandpass—so the combination gives a typical 50-Mhz bandpass for unity gain driving 75 Ω. Input voltage protection was not sacrificed to get the low input capacitance. The inputs have electrostatic discharge protection to greater than ±2000 V, and a fault current of 100 mA or more is needed to cause SCR latch-up.

The MAX455 multiplexer section uses "T" switches to attain high off isolation and interchannel isolation, typically 70 dB at 4 MHz. This high isolation can only be preserved with careful external wiring. A groundplane should be used, and all unused pins, which provide internal interchannel shielding, should be connected to it. Only channels two and three are physically close and a ground trace should therefore be run between them to cut external crosstalk.

TV SPECS

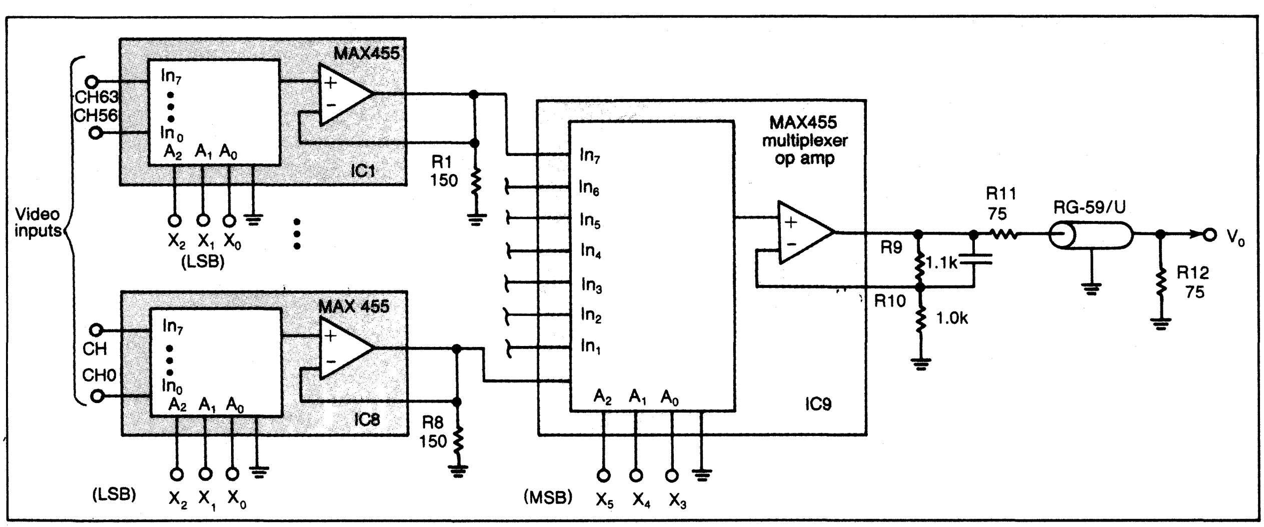
The multiplexer-amplfier can be used for both visual (TV) and instrument (flash converter) switching. For picture work, like switching teleconferencing or broadcast studio cameras, what's critical is differential gain and phase shift. This is measured as change in a small (0.28 V, 3.58 MHz) signal on a large, slowly varying, 0.7-V ramp. Change in gain means change in contrast over the picture, and change in phase corresponds to a change in color. Ideal specifications would be a phase shift of 0.1° and a gain change of 0.1%. No color or contrast changes would then be noticeable even in a long chain of multiplexers for switching cameras, attenuators, mixers, and so on.

The MAX452 video amplifier (the MAX455 without multiplexer) has a phase error of twice this, or 0.2°, while the MUX455 with eight-channel multiplexer has a phase error of about 1.2°. (The present design optimized bandpass somewhat at the expense of phase linearity.) The gain error is about 0.5% with or without the multiplexer. Both errors are tolerable though when only a few multiplexer/amplifiers are used in series, such as for small studios, teleconferencing, and remote surveillance. To do better even with a bipolar op amp of several hundred megahertz gain-bandpass may require special multiplexers such as reed relays, which are being phased out, or p-i-n diodes, which take a lot of power.

A typical link in the picture chain might be an amplifier-attenuator for setting the video level for the best signal-to-noise ratio. This could be done in inconvenient binary increments with a digital-to-analog converter, but it is easily accomplished in one decibel step changes with a single MAX455 multiplexer op amp (Fig. 3).

If more than eight multiplexer channels are needed, MAX455s can easily be cascaded with no special precautions other than adding 150 Ω to ground at the first stage output to insure stability (Fig. 4). There is a roll off in the second stage. R9 is 1.1 kΩ to compensate for the 1.5% gain loss of IC1 through IC8.□

Greg Schaffer is a senior member of the technical staff at MAXIM. He has a BSEE from the Massachusetts Institute of Technology, an MSEE from the University of California, Berkeley, and an MS in computer science from the University of Arizona. One of his two patents, a switching amplifier for noise reduction, resulted in a CMOS a-d converter with a resolution of 1 μV per count.



4. A 64-channel multiplexer can be built by using eight MAX455s to select 8 of the 64 channels and a final MAX455 chip to select each multiplexer. The first stage is operated at unity gain and 150-Ω load, which supplies some 40 MHz peaking to compensate for the high-frequency rolloff of the second stage. The -3 dB frequency is typically 35 MHz.

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Electronics Test

INSTRUMENTS, SYSTEMS AND TECHNOLOGY FOR TEST

New Method Improves Speed and Accuracy of Switched-Capacitor Filter Testing

by Tom Boydston, Teradyne, Inc., Sunnyvale, CA and Tai-Tsong Wang, Maxim Integrated Products, Sunnyvale, CA

New Method Improves Speed and Accuracy of Switched-Capacitor Filter Testing

An enhanced digital-signal processing scheme for testing switched capacitor filters can decrease test times and improve measurement accuracy over conventional analog and standard DSP-based testing methods.

by Tom Boydston, Teradyne, Inc., Sunnyvale, CA, and Tai-Tsong Wang, Maxim Integrated Products, Sunnyvale, CA

The cost and performance advantages offered by monolithic switched-capacitor techniques have made these filters a popular alternative to conventional analog filters—their bulkier and nonprogrammable counterparts.

Unfortunately for the production test engineer, the unique benefits of switched-capacitor filters can be offset by the serious testing difficulties they engender. A new technique, however, promises to speed device testing while improving measurement accuracy over conventional analog and standard digital signal processing (DSP)-based testing methods. The scheme, which is called the exact DSP test solution, uses phase information to solve parameters such as Q and F_o quickly and accurately.

This article will describe the implementation of three different methods—analog, DSP approximation, and the exact DSP test solution—in switched-capacitor filter (SCF) testing. The measure of success of each test will be the speed and accuracy achieved.

The popular MF10 switched-capacitor filter will be the example device under test (DUT) to illustrate the three test techniques. Several reasons led to the selection of this particular programmable SCF. To begin with, the device, introduced by National Semiconductor Corp. (Santa Clara, CA), has been available for many years and is available from several firms, including Maxim Integrated Products, Texas Instruments (Nouston, TX), Exar (Milpitas, CA), EG&G Reticon (Sunnyvale, CA), Linear Technology (Milpitas, CA) and others. What's more, the MF10 is a good example of a device with an exceptional set of testing problems.

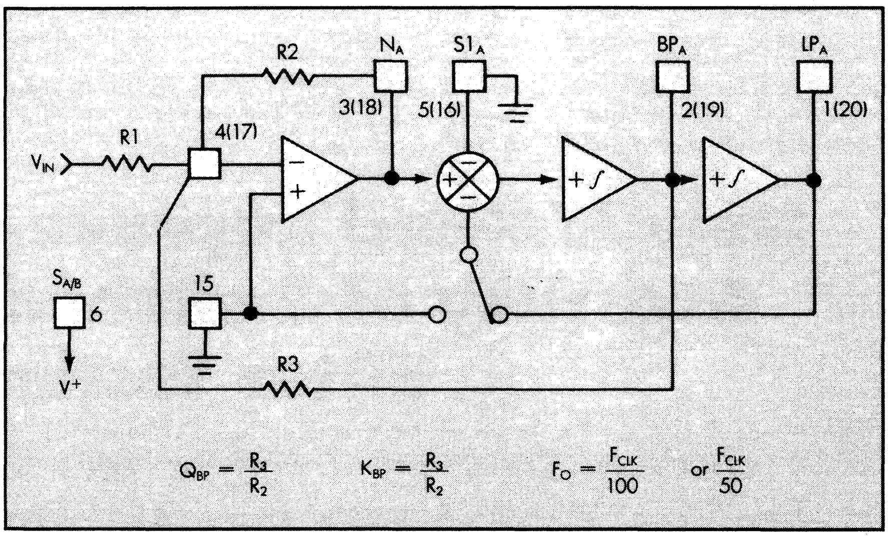

Fig 1. The model MF10 switched-capacitor filter is composed of two identical, second-order, independent active filter sections. For testing, it is configured as a bandpass filter.

For example, test times for a complete run test on the MF10 can exceed several seconds per device, and measurement inaccuracies for Q can be as high as 2 percent for bench-level testing (where the procedures can be fine-tuned) and even worse when using automatic test equipment (ATE). The exact DSP test solution, however, can reduce test times to less than one second and increase tenfold the accuracy of the critical F_o and Q measurements using high-throughput ATE.

Two identical, second-order, independent active filter sections comprise the MF10 (Fig 1). The center frequency can be digitally programmed to be 1/100 or 1/50 of the input device clock frequency (F_{clk}). An external resistor ratio controls the gain quality factor—the ratio of center frequency to the bandwidth. Even though SCFs are sampled-data systems, it's important to note that their transfer functions can be described effectively using continuous frequency domain functions; however, this is true only if the sampling rate is sufficiently greater than the center frequency.

Techniques for testing programmable SCFs

Of the three techniques to test switched-capacitor filters, the first approach, a straight analog method, can be implemented with either bench test equipment or production ATE. The second method, DSP approximation, utilizes some of the unique advantages of digital signal processing. The exact DSP test solution allows higher accuracy and faster test times; it can be easily implemented on mixed-signal (analog and digital) production ATE.

When testing programmable SCFs such as the MF10, necessary decisions to be made include selection of clock frequency (F_{clk}) and choice of the filter configuration (lowpass, highpass, bandpass, etc.) to test Q and F_o. In order to limit the complexity of the testing problem and keep the cost of test minimal, device manufacturers generally select as the test condition for second-order filters, such as the MF10, the bandpass filter mode at a F_o that is typically no higher than 5 kHz and a Q of 10.

In order to understand the differences between the three test techniques, a review of the continuous frequency domain transfer function of a second-order bandpass filter should prove useful. The transfer function is:

$$H_S = \frac{-KS}{S^2 + \frac{W_O}{Q} S + W_{O^2}}$$

Substituting:

$$S = j2\pi F \text{ and } W_O = 2\pi F_O$$

the transfer function of the filter can be described by either amplitude vs. frequency or phase vs. frequency. The amplitude vs. frequency equation then becomes:

$$H(F) = \frac{KF}{2\pi[(F_O{}^2 - F^2 + (\frac{F_O}{Q})^2 F^2]^{1/2}}$$

The phase vs. frequency equation has the following form:

$$\Theta(F) = \tan^{-1}[\frac{Q(F_O{}^2F^2)}{F_OF}] - 180°$$

where F_o is the center frequency of the bandpass filter, f is the input frequency, and Q is defined as $Q = F_o/(F_h - F_l)$. F_h and F_l are −3 dB from F_o in the amplitude equation H(f), or −225 degrees and −135 degree phase shift in the phase equation θ(f) (Fig 2). Due to the nature of sampled data SCFs, F_h and F_l measured using amplitude measurements will not coincide exactly with those measured using phase shift measurements.

A simple analog test method

By using the above set of equations, an analog method for testing Q and F_o can be devised based on the phase information contained in the difference between two sine waves. A dual-input timer-counter measures the time difference—the phase shift—between the two sine waves; the phase shift determines the Q and F_o. Furthermore, by using a reference with the output of the DUT, the absolute phase shift through the filter is subsequently measured. The source sine wave frequency is then changed until the measured phase shift is −225°, −180°, and −135°, corresponding to F_h, F_o and F_l.

Although this method provides a simple and straightforward approach to test for Q and F_o, its liabilities—long test times, poor repeatability, and low accuracy—outweigh the benefits. For example, one source of inaccuracy is in the input section of the timer-counter comparators or zero-crossing detectors. Indeed, with a 2-V p-p, 5-kHz sine wave as the input to the zero-crossing detectors, and assuming a ±50-mV uncertainty band around 0 V, the result will have a maximum accuracy of only ±2.87 degrees. This error translates directly into a variation in F_o of ±4.8 percent.

Taking multiple samples for each measurement minimizes the effect of phase jitter by the zero-crossing detectors. But because only one sample is taken per cycle of the input signal and measurements must be taken for F_l, F_h, and F_o, the test time tends to get unacceptably long.

Another source of error is due to the nature of an SCF sampled-data system, where the output of the filter is a staircase signal. For an input signal at exactly F_o of the SCF, the output can have either 100 or 50 quantized levels. Error can be introduced into the analog method depending upon which quantized level the zero-crossing detection occurs. For example, by missing the detection of one adjacent level, an error can be as much as 3.6°, or 7.2° in the phase measurement.

The DSP approximation test method

To reduce the long test times associated with the analog test method, a simple multitone DSP approximation method is used with a mixed-signal ATE. The test time is reduced because several tones can be supplied to the device simultaneously and the output of the device is digitized. Thus, all the data is captured within one acquisition of the digitizer. This simultaneous sourcing and measurement of test signals has traditionally been used to reduce test time.

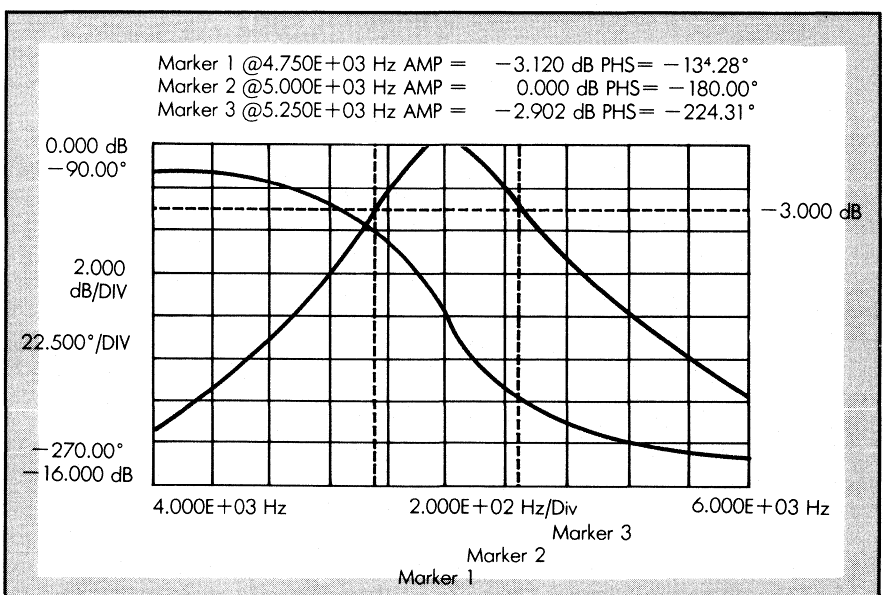

Fig 2. Two frequency domain transfer functions describe a second-order bandpass filter—amplitude vs. frequency and phase vs. frequency. F_h and F_l are −3 dB down from F_o in the amplitude equation H(f), or −225 and −135° phase shift in the phase equation θ(f)

Digitizing the bandpass-filtered output of the DUT and performing a Fast Fourier Transform (FFT) on the time domain samples results in data displayed in the frequency domain. The output of the FFT will produce a magnitude at each of the discrete frequencies. By interpolating between the two magnitudes nearest the −3 dB bandwidth frequency values, and assuming a straight line between these two points, the frequencies at the −3 dB points can be approximated (Fig 3 and Fig 4). The same approach can be used for F_o by searching for the two maximum amplitudes nearest F_o and interpolating F_o. Finding the bandwidth and F_o allows the value of Q to be calculated.

One major disadvantage with the DSP-based amplitude approximation is that the slope of a bandpass curve at F_o is almost equal to zero. For example, a 5-Hz variation from an F_o of 5 kHz produces only 0.02 percent variations in the amplitude. Thus, a 2-V p-p input tone will only produce a 400 μV difference at the output of the bandpass filter. This presents a difficult measurement for the test system when searching for the peak F_o amplitude value. System resolution and noise will adversely affect F_o repeatability and, consequently, will also affect Q.

This SCF testing method does, however, have a very important advantage. Although the accuracy of the DSP-amplitude approximation is, at best, equivalent to the straight analog approach, the test time is reduced by up to 50 percent because the DSP permits the sourcing and measurement of multiple waveforms simultaneously. For greater test time reductions and markedly improved test repeatability the exact DSP technique is an effective, new alternative.

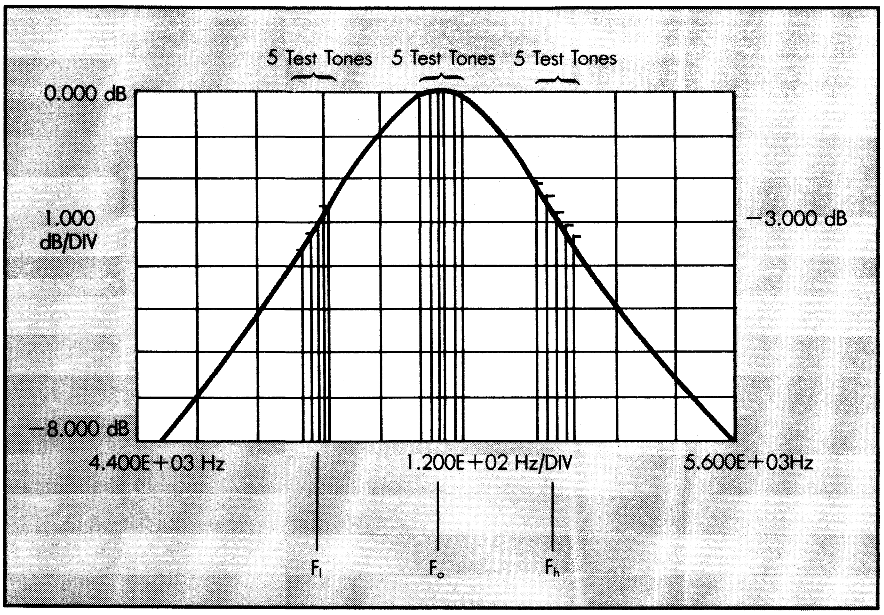

Fig 3. DSP-based approximation method for determining F_l, F_o, and F_h.

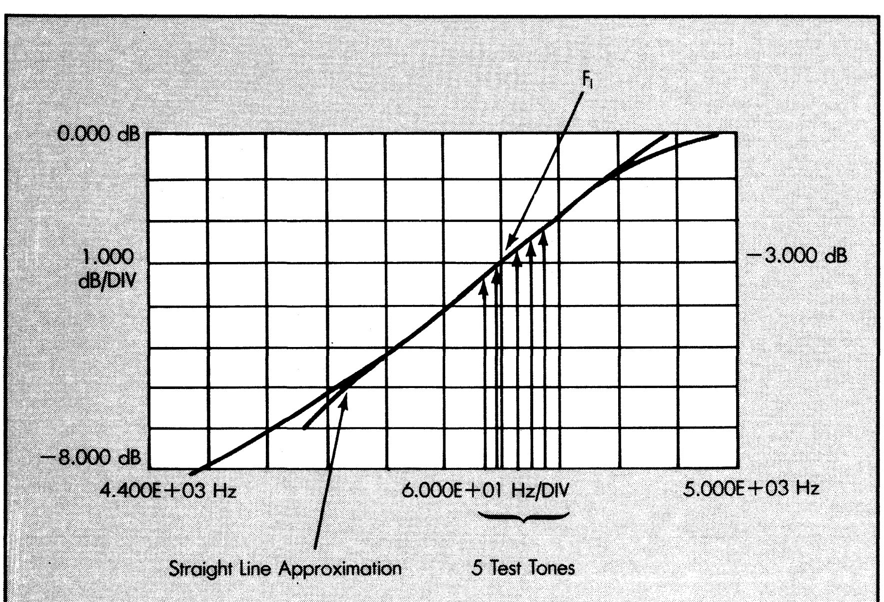

Fig 4. DSP-based approximation method for determining F_l using a straight-line approximation, interpolating the −3-dB point, and calculating F_l using $y = m x + b$.

The exact DSP test solution

Two frequency domain transfer functions describe a second-order bandpass filter—amplitude vs. frequency and phase vs. frequency (Fig 2). Either transfer function can be used to obtain Q and F_o, as both equally describe the characteristics of the device.

One form of the exact DSP test solution uses this phase information to solve for Q and F_o. When time-sampled data are transformed to frequency data via the FFT, the resultant frequency data are complex vector pairs describing the magnitude and phase of each tone. Sampling and performing an FFT on tones that bypass the DUT, and then comparing them to the phase-shifted tones that pass through the device, determines the absolute phase shift at the test frequencies.

The exact DSP test solution that uses phase information is best described from a graphical point of view (Fig 5). Because an ideal second-order bandpass filter will exhibit the same characteristic reversed "s" curve (which tends to flatten out for lower Qs and become steeper for high Qs), only two points in frequency and phase need to be identified in order to draw the characteristic curve that describes the Q and F_o.

For example, two test tones, F_{test1} and F_{test2}, which, for the sake of convenience, are placed approximately at F-low (F_l) and F-high (F_h), have been input to the DUT, the resultant output tones have been digitized, and an FFT has been performed on the time samples of the two-tone waveform (Fig 6). A phase shift through the DUT can be determined and two frequency-phase pairs are obtained. The phases, θ_{test1} and θ_{test2} represent the phase shift through the DUT at F_{test1} and F_{test2}. These two frequency-phase pairs are all that is needed to "draw" the singular and unique phase curve that describes the frequency transfer function of the bandpass filter. F_o and Q can now be extracted from the curve.

The problem encountered with the analog method's zero-crossing detector and phase jitter is not a problem when digitizing time domain data from the output of the DUT. A mixed-signal test system that provides exact internal synchronization guarantees that the samples will be taken at the middle of quantized output levels. Integrated test system synchronization and the nature of time sampled data will reduce any phase jitter, or aperture jitter of the digitizer, by averaging the samples taken. Furthermore, because the samples are taken at a high digitizing rate, multiples of samples can be taken within one cycle of the test frequency, thus reducing the effect of system noise and increasing repeatability.

With the exact DSP test solution, test time is dramatically reduced over either of the previous methods. When compared to the analog method, the exact DSP test solution reduces test time by simultaneous sourcing and measurement of multitone signals. The approximate DSP method, in contrast, requires longer sampling acquisition time than the exact DSP test solution due to the required placement of numerous tones with as small a frequency resolution as system limitations allow. Here, sampled data acquisition time is inversely proportional to the frequency resolution demand of the method. The exact DSP solution (using phase) takes samples faster because it allows the test engineer to place a minimum of just two tones with a much larger frequency resolution at approximately F_l and F_h (the −3 dB points above and below F_o).

A similar exact DSP solution using the amplitude vs. frequency equation alternatively can be used to measure Q and F_o. Rather than use the phase information from an FFT of the sampled output data, the magnitude data can be substituted. When using this approach, three test frequencies are required rather than the two required for the exact DSP solution using phase information. The extra tone is required because the equation has a third term, k (the gain factor).

Two test system factors determine the accuracy and repeatability of the exact DSP solution: frequency accuracy and test system synchronization. The latter includes such factors as synchronization between the multitone source, the digitizer, and the device clock.

System synchronization

To measure the phase delay through the SCF accurately, a test system should provide strict control of the frequency and phase synchronization between (1) the multitone waveform source, (2) digitization of the DUT output waveform and (3) the device F_{clk} (which determines the center frequency, F_o, of the filter). A test system architecture that incorporates multiple internal low phase jitter digital synchronization paths not only increases the overall accuracy and repeatability of the system, but, in addition, it gives the test engineer added flexibility.

A bus architecture that integrates analog, digital, and DSP signals be-

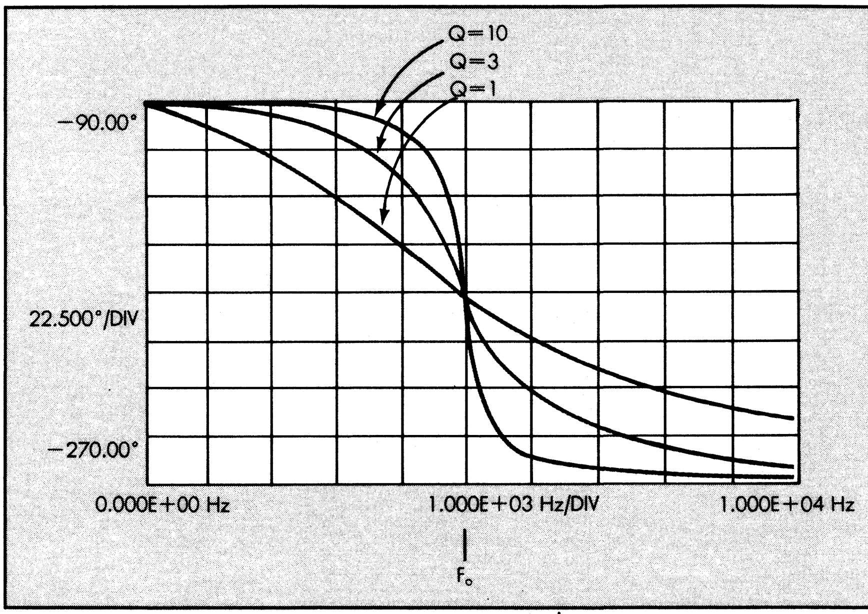

Fig 5. Phase and frequency for a bandpass filter with the same center frequency F_o, but 3 different Q values.

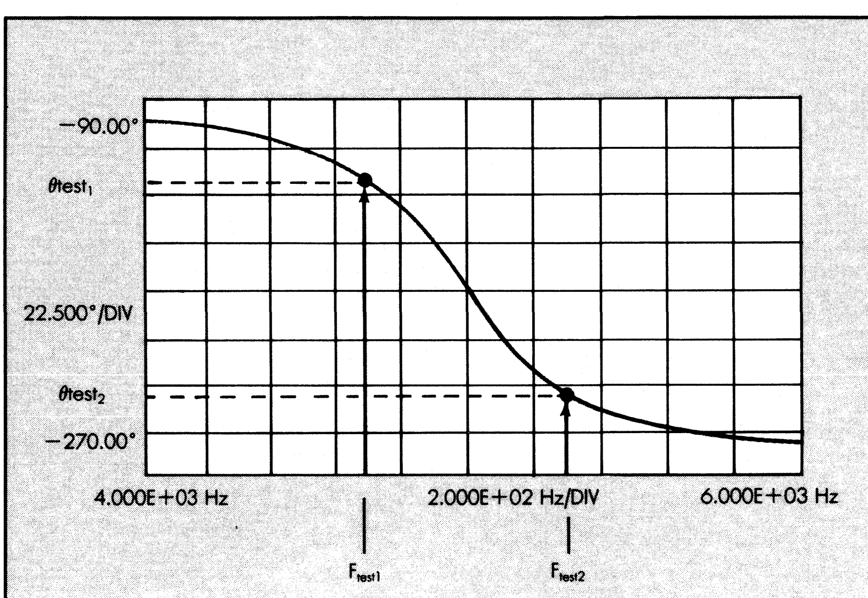

Fig 6. The Exact DSP solution uses only two test tones to describe the phase curve through a bandpass filter. Once the curve is "fixed," Q and F_o can be extracted.

fore they go to the DUT solves complex programming problems of digital-to-analog clock synchronization by giving the test engineer a straightforward path for synchronization. Additionally, complex application hardware is not required for synchronization.

The Teradyne A370 Test System is used to illustrate the measurement technique for an exact DSP test solution using a test system with a multiple internal synchronized bus structure.

To supply a 2-MHz master clock (Fig 7), Teradyne's M621 High Speed Digital Subsystem, a digital pattern generator and processor, is used. A 500-kHz clock derived from this master clock is then sent to the SCF as the device's F_{clk}.

Through a Vector Bus synchronization path this 2-MHz clock is also sent to the ac subsystem and to an arbitrary waveform generator, such as Teradyne's IM101 Programmable Waveform Generator. By internally dividing the 2-MHz clock by 20, a 100-kHz clock is derived and used to source the multitone waveform to the DUT. A 1-bit programmable sync signal imbedded at a specific time sample of the multitone waveform is used to trigger the start of digitization of the SCF output. The waveform digitizer samples the output of the DUT at a 20 kHz rate, a clock rate that is also derived from the 2-MHz master clock.

After the waveform is digitized, a high-speed array processor module, Teradyne's M634, performs an FFT to get the amplitude and phase information for all tones. To obtain the phase delay through the SCF, a local relay is all that is required to provide a calibration path to measure the phase directly from the output of the local buffer, U1. The phase measured from the SCF output is subtracted from the constant system delay and is called the SCF phase delay. Alternatively, the master synchronization clock can be sourced from the ac subsystem to synchronize all signals of the user's choice under high-level software control.

Breakthroughs in cost and accuracy

Switched-capacitor filtering techniques have experienced widespread use in the recent past and offer many cost and performance advantages over analog filters. To test filters economically and with high precision, however, is not a trivial task. The exact DSP test solution offers a breakthrough in both high accuracy and low test cost. Test times have been shown to be reduced by as much as a factor of five to below one second for a complete production test program. Measurement repeatability for the critical tests of Q and F_o can be demonstrated to be better than 0.2 percent due to system synchronization. Higher repeatability allows a relaxation in test limit guardbands that results in higher yields. The exact DSP test solution can be performed using either the phase equation or the amplitude equation. Finally, an element that is critical to the success of this test method is system synchronization such as the Vector Bus architecture will provide. ET

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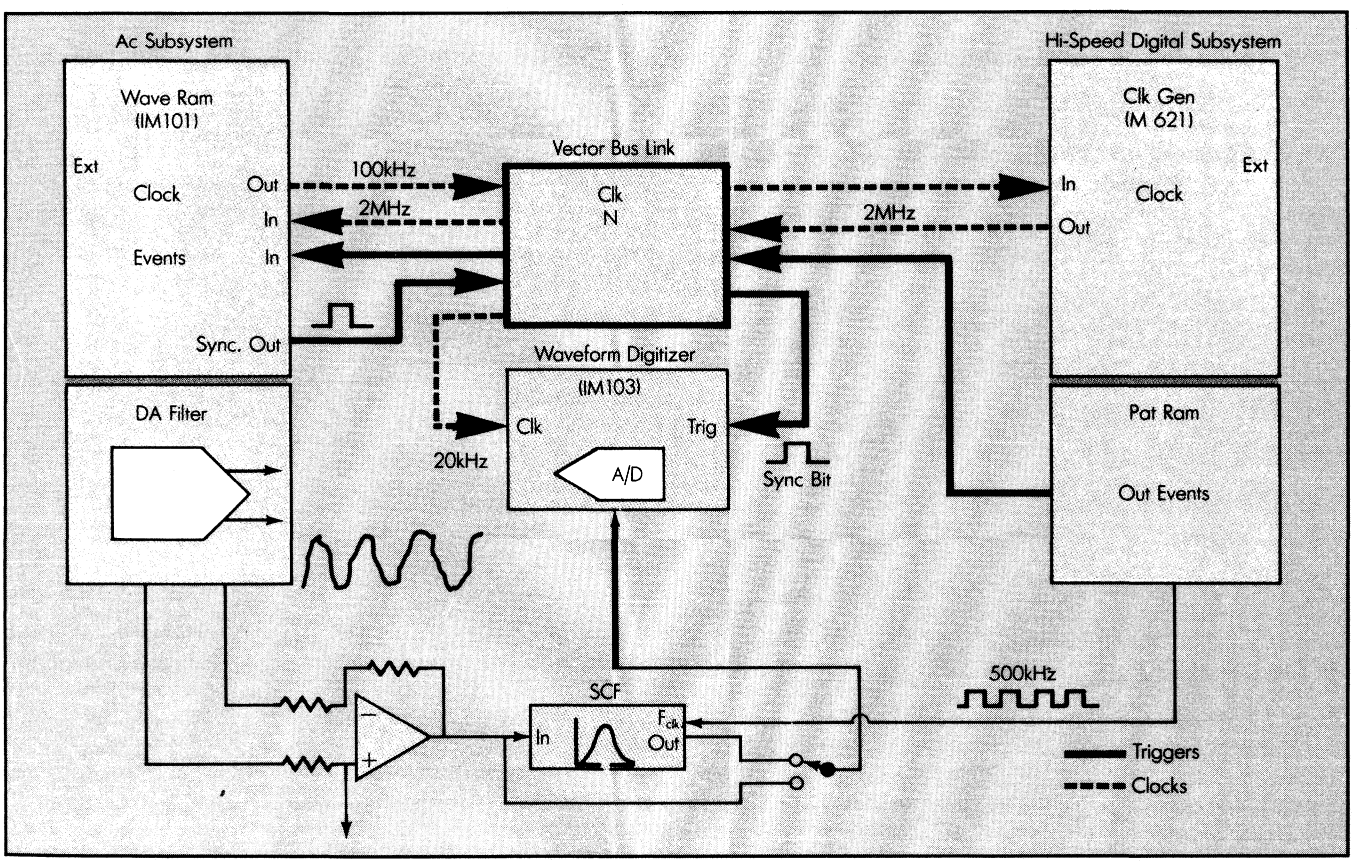


Fig 7. Switched-capacitor filter test setup to supply a 2-MHz master clock.

Step-up converter produces 5V from 1.5V

Gerald Grady
Maxim Integrated Products, Sunnyvale, CA

You can produce a regulated 5V output from a 1.5V battery cell by using the step-up dc/dc-converter circuit shown in **Fig 1.** The circuit can operate with V_S as low as 1V, but it requires at least 1.5V to start. The output can deliver 100 mA when V_S is 1.5V or 1.7A at 70% minimum efficiency when V_S is 3V.

Supply voltage for the switching regulator IC_1 appears first at pin 4 (start-up mode) and then at its own V_{OUT} terminal, pin 5. The chip includes an oscillator, bandgap reference, three voltage comparators, a catch diode, and associated control circuitry. An internal MOSFET lets you implement low-power applications; higher power calls for an external device: Q_1 in **Fig 1.** The on-chip oscillator provides a 55-kHz square-wave drive to both the internal and external MOSFET.

When Q_1 turns off, current through inductor L_2 drives the drain-node voltage higher. Similarly, current through L_1 drives the voltage at pin 5 higher when the internal MOSFET turns off. This action generates two independent voltages, each higher than V_S. Q_1 and L_2 generate sufficient overhead voltage to enable the regulator chip to produce a regulated 5V output, and the internally generated voltage ensures adequate gate drive to Q_1. The internal voltage, clamped by the 10V zener diode D_3, ranges from 10V (turn-on) to 15V (normal operation). Q_1's resulting $R_{DS(ON)}$ is only 0.085Ω.

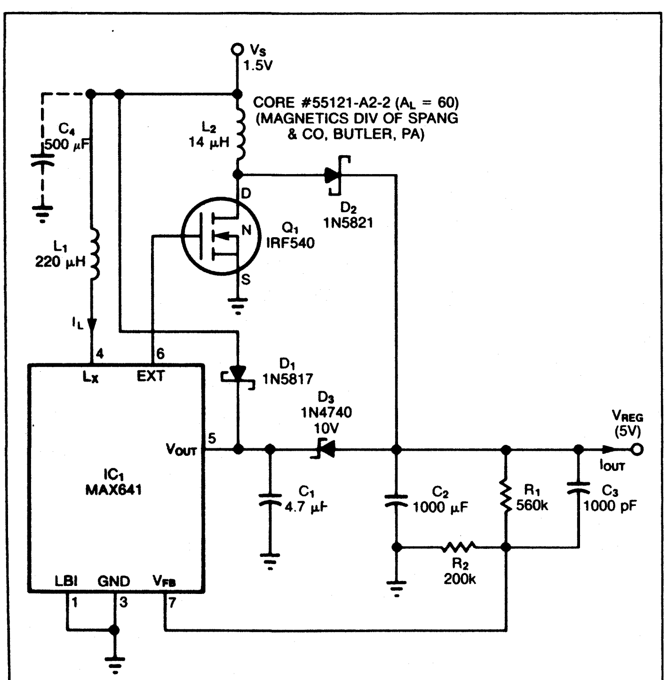


***Fig 1—This switching-regulator circuit** converts a dc input (as low as 1.5V) to a higher regulated value—to 5V, for example.*

TABLE 1—OBTAINABLE VALUES OF I_{OUT} MAX

	L_2		
V_S	14 µH	27 µH	50 µH
1.5V	105 mA	48 mA	30 mA
2.0V	220 mA	95 mA	60 mA
2.5V	390 mA	160 mA	100 mA
3.0V	1.75A	1.5A	1.25A

Resistors R_1 and R_2 determine the regulated output level. For V_{REG} outputs other than 5V, set

$$R_1=R_2\left(\frac{V_{OUT}}{1.31}-1\right).$$

You choose an R_2 value in the range from 10 kΩ to 10 MΩ.

For low values of V_S, losses in the internal and external diodes and the Q_1 inductor sharply limit the maximum output current. The following design equations let you determine component values while calculating this current. First, Q_1 must be able to handle the peak current I_{PK} of inductor L_2:

$$I_{PK}=V_S t_{ON}/L_2,$$

where $t_{ON}=0.55/f_{OSC}$. For this circuit, then, $I_{PK}=1.07A$.

The circuit loss V_{LOSS} is

$$V_{LOSS}=I_{PK}(R_{DS(ON)}+2R_{L_2})+V_{D_2},$$

where R_{L2} is the dc resistance of L_2 and V_{D2} is the forward voltage drop of D_2. For this circuit, $V_{LOSS}=0.56V$. The output current is

$$I_{OUT}=\frac{0.5L_2 I^2_{PK} f_{OSC}}{V_{REG}-(V_S-V_{LOSS})-I_{L_1}}.$$

Therefore, $I_{OUT}=103$ mA for the circuit shown. **Table 1** shows the output currents you can obtain for various values of V_S and L_2. The corresponding conversion efficiencies range from 70 to 85%. **EDN**

DC/DC converters adapt to the needs of low-power circuits

High cost, quiescent current, and circuit complexity have often restricted switching power supplies to high-power applications, for which the switchers' high efficiency, wide input range, and reduced size and weight offset their drawbacks. Now, however, you can employ switchers in low- and medium-power applications as well.

Len Sherman, *Maxim Integrated Products*

Designers of dc/dc-conversion products are now addressing the special requirements of low- and medium-power applications. As a result, you can apply switching techniques' advantages in battery-powered portable equipment, telemetry devices, and consumer products.

A key requirement for designers of battery-powered products is that they minimize the number of cells used in the product. Substituting, for example, two large cells for a stack of six or seven smaller ones yields not only reductions in size and weight but also increased reliability and energy density. An efficient, low-power step-up voltage converter used in conjunction with a few high-capacity, low-voltage cells makes such a trade feasible, especially in an application where a stack of expensive rechargeable batteries would be the alternative.

The circuits shown in **Figs 1** through **7** are all

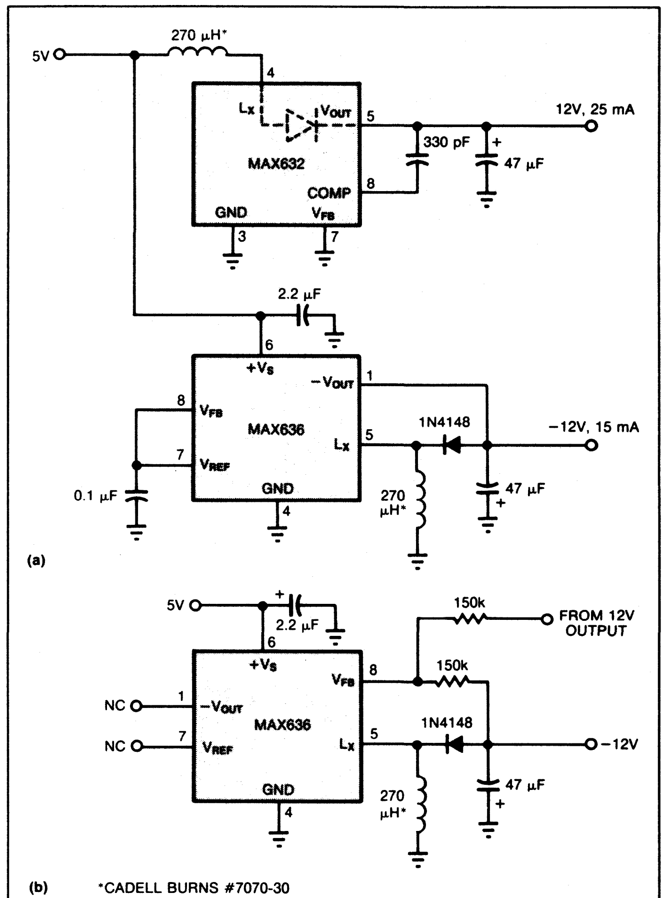


Fig 1—You can tailor this ±12V supply *to provide either independently regulated outputs* ***(a)*** *or a tracking negative output* ***(b)****. The inductors don't exact too great a size penalty: Each measures only 0.6 in. long by 0.26 in. in diameter.*

The flyback configuration keeps circuitry compact, and it adapts not only to voltage boosting but to buck and buck/boost configurations as well.

flyback-type switching dc/dc converters (the same type that generates 10- to 20-kV supplies for television, video display terminals, and oscilloscopes) that operate at 50 kHz (see **box,** "Flyback converters' internal operation"). The flyback configuration keeps the circuitry compact, and its versatility allows it to accomplish more than simple voltage boosting.

Derive ±12V from digital system's supply

Often, a digital system powered by a 5V supply includes a few analog functions that require ±12V. The circuit shown in **Fig 1** uses two dedicated 8-pin converters—the MAX632 and MAX636—to derive 25 mA at 12V and 15 mA at −12V from a 5V logic supply. You can configure the circuit for independently regulated outputs **(Fig 1a)** or for tracking regulation **(b).**

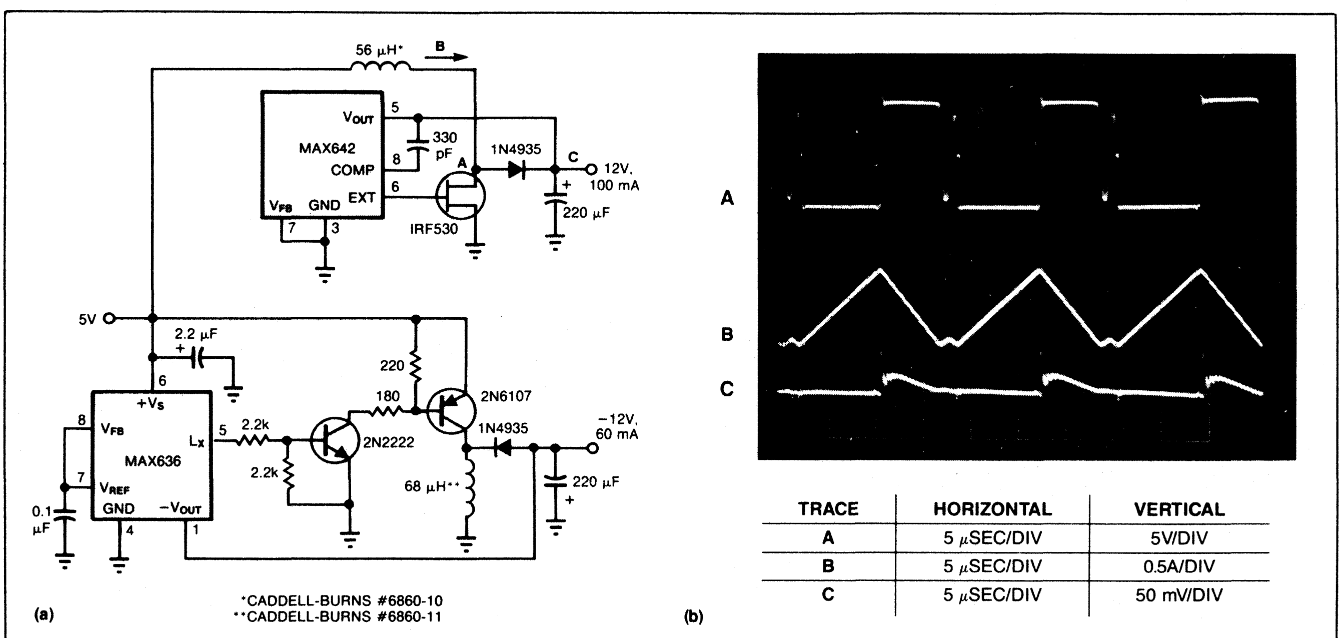
The positive converter's efficiency is 85%; the inverter's is 75%. You can improve these efficiency figures slightly by using Schottky diodes rather than the MAX632's internal diode and the 1N4148 signal diode connected to pin 5 of the MAX636. If you opt to use a Schottky diode with the MAX632, connect it in parallel with the chip's internal diode (that is, between pins 4 and 5).

With several popular types of high-current rectifier diodes, such as ones in the 1N4000 Series, efficiency and overall performance are poor for high-frequency (greater than 10 kHz) dc/dc conversion. Many of these diodes were designed to pass high current only at 120 Hz; therefore, they waste energy at 50-kHz operating frequencies. In addition, these slow rectifiers might also allow the inductor's discharge voltage to reach excessive levels before the rectifier turns on and directs current to the load.

Small-signal diodes, such as the 1N4148, are fast enough and work well in applications that require less than 50 mA. High-speed rectifiers, such as the 1N4935, are suitable in applications that require as much as 1A. Schottky diodes provide the best performance with respect to speed and forward voltage drop, and they can significantly improve efficiency in low-voltage, high-current applications. However, you'll have to decide on the basis of your individual application whether their higher cost and relatively low reverse breakdown voltage eliminate the Schottky diodes from consideration.

External MOSFET increases power

If your application requires higher power than **Fig 1**'s circuit provides (if, for instance, you need the power for a data-acquisition board or a high-level industrial controller), then you can modify the circuit by adding an



TRACE	HORIZONTAL	VERTICAL
A	5 µSEC/DIV	5V/DIV
B	5 µSEC/DIV	0.5A/DIV
C	5 µSEC/DIV	50 mV/DIV

***Fig 2—With the addition of a few** external components **(a)**, the circuit of **Fig 1** can supply currents of 100 mA at 12V and 60 mA at −12V. Traces A, B, and C **(b)** represent the switch voltage, inductor current, and output ripple for the 12V supply.*

external power MOSFET, as shown in **Fig 2a,** and obtain 100 mA at 12V and 60 mA at −12V. The power MOSFET drops the 12V converter's efficiency to 80%, but driving the power MOSFET doesn't require any additional parts.

The scope photo **(Fig 2b)** shows some of the key waveforms in the step-up circuit. Trace A is the voltage waveform at the drain of the IRF530 MOSFET (under full load), trace B is the inductor current, and trace C is the ripple voltage at the 12V output. The ringing found on trace A near the end of each discharge cycle is normal and is due to the inductor's interaction with stray capacitance when the inductor current decays to nearly zero. As you can see from trace C, this ringing has no effect on the output waveform.

Compensate for IR drops

Not only might you need to derive ±12V from a 5V supply, you might also need to derive a regulated 5V level from a nominal 5V supply that suffers from an unacceptable voltage drop because of IR effects in long power-distribution cables. You can efficiently boost the voltage back to a regulated 5V by using the circuit shown in **Fig 3.**

That circuit operates at input voltages as low as 4.5V. The transformer's 3.2:1 turns ratio allows the circuit to supply more than the MAX631's usual output current without requiring external power transistors. This circuit provides as much as 150 mA of output current at 5V. You can wind the transformer on a 14×8-mm pot core, or you can obtain the transformer by ordering the standard part number listed in the schematic.

When the MAX631's L_X switch turns off at each half cycle of its 50-kHz clock, the reflected voltage in the transformer's primary generates a 9V supply voltage for the MAX631 at the V_{OUT} pin. Operating the MAX631 at 9V rather than at the 4.5V provided at the input increases the gate-source voltage of the internal MOSFET, consequently reducing the MOSFET's on-resistance. This circuit requires the external feedback resistors at V_{FB} because, unlike the previous circuits, this circuit doesn't allow you to use V_{OUT} as the feedback input for the regulator.

Derive 12V from 8 to 15V input

The simple boost converters of the previous examples are inadequate for some battery-powered applications. For example, the unregulated output of a 12V sealed lead-acid battery varies from worst-case peaks of 15V down to as little as 8V when it is deeply discharged. Therefore, you can't derive a regulated 12V output from a 12V lead-acid battery by using a simple boost converter, such as one of those illustrated in **Figs 1** and **2,** because a boost converter can't accept an input voltage that is greater than its output voltage. Conversely, a buck converter can't accept an input voltage that's less than its output; therefore, a simple buck converter won't work either. A buck/boost converter, as the name implies, is a combination of buck and boost circuitry that successfully addresses the challenge of

Text continued on pg 150

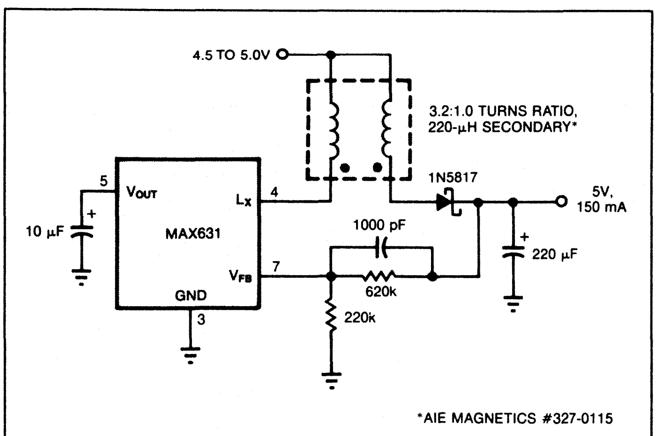


Fig 3—This simple circuit boosts *a supply voltage that might have sagged substantially because of IR drops in long cables.*

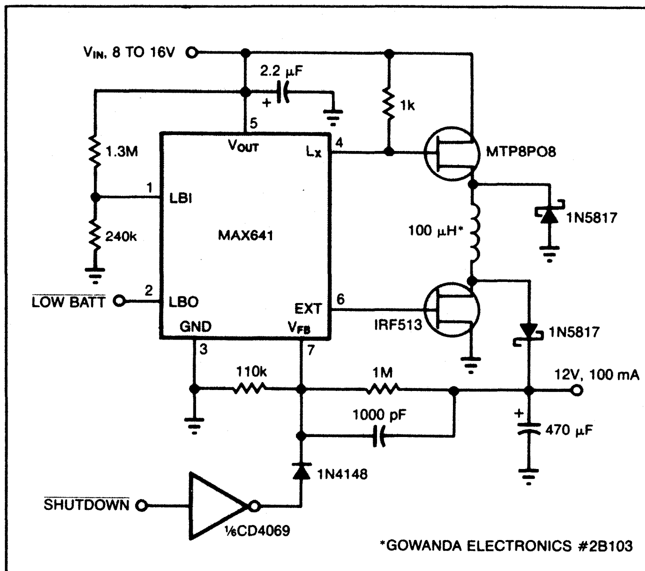


Fig 4—A buck/boost converter *can accommodate wide input-voltage swings, such as the 8 to 15V swing typical of a 12V sealed lead-acid battery. The LOW BATT output indicates when input voltage drops below 8V. Pulling SHUTDOWN low turns off the circuit.*

Flyback converters' internal operation

In a flyback converter, voltage applied to an inductor or transformer primary via a switch causes inductor current to rise for a fixed period of time. When the voltage is switched off, the magnetic field stored in the transformer collapses, causing the secondary to supply current to the load. With the MAX640 and MAX630 Series devices, this switching occurs at 50 kHz. You can use these devices to step up the voltage, step it down, or invert it just by changing the configuration of the switch (transistor), coil, and steering diode.

Fig A illustrates the MAX641's internal operation. When the output voltage drops below the preset (or externally set) value, the error comparator switches high and connects the internal oscillator to the L_X and EXT outputs. EXT is typically connected to the gate of an external n-channel power MOSFET (although the external MOSFET isn't necessary for most of the low-power circuits discussed in the accompanying article). When EXT is activated, the MOSFET turns on and off at the oscillator frequency.

When EXT is high, the MOSFET switches on, and the inductor current increases linearly, storing energy in the coil. When EXT switches the MOSFET off, the coil's magnetic field collapses, and the voltage across the inductor changes polarity. The voltage at the catch diode's anode then rises until the diode is forward-biased, delivering power to the output. As the output voltage reaches the desired level, the error comparator inhibits EXT until the load discharges the output capacitor to a point at which the error comparator connects the oscillator to the L_X, and EXT generates output once again.

The MAX641 doesn't have a V_{IN} pin. Input power to start the dc/dc converter is supplied via the external inductor (and external diode, if used), to the V_{OUT} pin. If you use an external catch diode, connect its cathode to V_{OUT}. Once the converter is started, it's powered from its own output voltage. This bootstrap design ensures that the external MOSFET has the maximum gate drive and, consequently, the minimum R_{ON}.

One external component that you must select is the inductor. Although the inductance of many types of coils, such as RF chokes and air-core inductors, frequently falls in the appropriate range for dc/dc converters (50 to 500 μH), these inductors typically saturate at only a few milliamps and therefore are not a good choice for your dc/dc-converter design.

A saturated inductor ceases to behave as an inductor. It can no longer store energy in its magnetic field, so the mechanism that normally limits the inductor current no longer operates; all that limits the current is the series resistance. This resistance is quite low; consequently, the current can rise to an excessive, and possibly destructive, level.

The scope photo in **Fig B** shows the switch voltage (trace A) and inductor-current waveforms (trace B) for an inductor that's well on its way to saturation. Compare these waveforms with the normal performance illustrated in **Fig 2b** on pg 146. The A and B waveforms in both photos are of the same A and B nodes of the 12V boost circuit in **Fig 2a. Fig B** reflects the effects of using an inductor with an inadequate current rating in **Fig 2a**'s circuit.

When you look at **Fig B,** you'll see that, in the middle of the

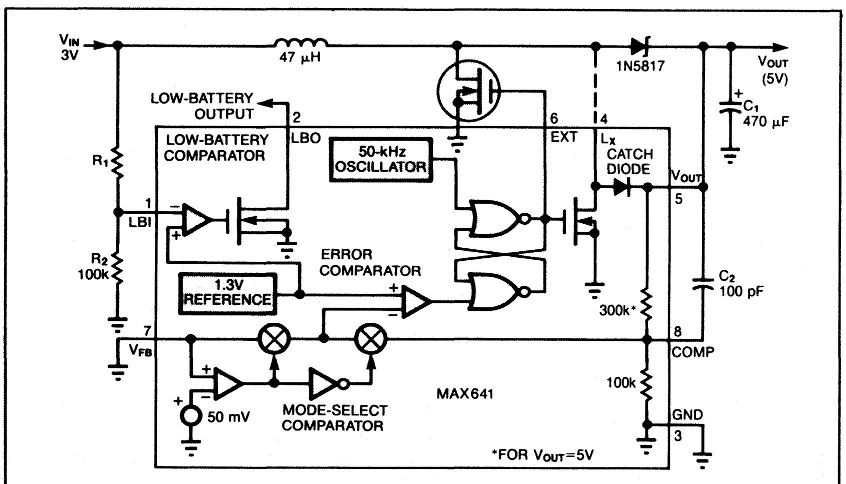


Fig A—This block diagram illustrates the MAX641's operation. For many low-power applications, the external MOSFET and Schottky diode are unnecessary.

charge cycle, above the 0.5A level, the current waveform's slope increases markedly, indicating the onset of saturation. At this point, the effective inductance of the coil decreases because the current through the inductor has risen to the saturation level. The rising edge of the switch-voltage waveform is much slower in **Fig B** than in **Fig 2b** because the inadequately rated inductor takes several microseconds to come out of saturation.

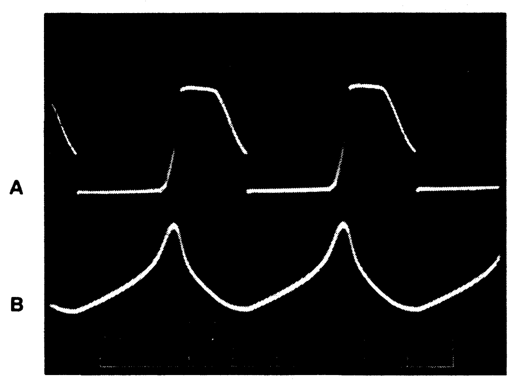
An inductor doesn't saturate as long as its operating current is less than its rated maximum current. At first glance, it would seem easy enough to specify the maximum current rating for your inductor, but what you have to watch out for in your dc/dc designs is that the peak inductor current is often four to six times the converter's average current output. In the case of flyback converters, this peak current flows not just under peak load conditions, but each time the current switch turns on. For this reason, you must give careful consideration to the current rating of your converter circuit's inductor.

Besides the care required in the selection of inductors, another often-overlooked area of concern in dc/dc-converter design is that encompassed by grounding, shielding, and bypassing. The quality of ground connections is key to the performance of dc/dc converters. Because the peak current in an inductor or switch (transistor) can reach several amps, you must provide these points with very-low-impedance paths to the supply common. For example, in the inverting circuit of **Fig 2a**, the coil current typically exceeds 1A. For best results, use separate paths to ground for the high-current paths so that they are separated from the chip's power and feedback connections. If you don't have the option of separate traces, then use as heavy a single trace as you possibly can to carry the high current back to the supply.

Loop instabilities, caused by interactive ground connections or stray capacitive pickup, can also severely limit the performance of an otherwise sound dc/dc-converter design. Some of the symptoms of these problems are high ripple voltages at the output, efficiency that's lower than expected, and "motorboating," or low-frequency oscillation.

Motorboating occurs when the control loop of the dc/dc converter produces pulses in periodic clusters of 10 to 20 pulses rather than at more or less random intervals. Motorboating can be caused by one or more of the following phenomena: stray pickup at the feedback node, unwanted feedback to the reference, and feedback via the ground or power-input pin.

If the cause is stray pickup at the feedback node, add a lead compensation capacitor (100 to 1000 pF) from the feedback terminal or COMP pin to the circuit output or reduce the size of your connections at the feedback input in order to reduce stray capacitance to ground. If unwanted feedback to the reference is the culprit, bypass the reference and power-input pins to ground (using 0.1 to 1.0 μF). If your circuit is suffering from feedback via the ground or power-input pin, bypass the power-supply input (1.0 to 10.0 μF). You should also separate high-ground-current connections from the reference, feedback, chip-ground, and chip-power connections.



TRACE	HORIZONTAL	VERTICAL
A	5 μSEC/DIV	5V/DIV
B	5 μSEC/DIV	0.5A/DIV

Fig B—The marked increase *in the current waveform's slope (trace B) illustrates the onset of saturation for an inductor with an inadequate current rating. Trace A represents switch voltage.*

You must sometimes develop 5V from a nominal 5V input that has sagged because of IR drops in long power-distribution lines.

the wide input-voltage swing associated with the sealed lead-acid battery.

The circuit of **Fig 4** is a buck/boost converter that provides 100 mA at 12V and accepts 8 to 16V inputs. Both ends of the circuit's inductor are switched by separate power MOSFETs, which the MAX641 drives directly via its L_X and and EXT outputs. These outputs operate out of phase, so the p- and n-channel FETs turn on at the same time. When both the n- and p-channel FETs turn off, the two Schottky diodes steer the coil's discharge current to the 12V output. A slight drawback of this circuit is that the converter's efficiency is less than that of a pure buck or boost converter, because the two MOSFETs and two diodes increase losses in the charge and discharge current paths. Nevertheless, the circuit still delivers 100 mA at a respectable 70% efficiency figure.

An additional benefit of this type of circuit is that you can control its operation with a TTL-level signal. Overriding the V_{FB} input with a high-level TTL signal (such as the diode-coupled inverter output in **Fig 4**) fools the MAX641's internal feedback circuitry into thinking that the output is too high, so the chip turns off both MOSFETs. The circuit's idle current is around 400 μA.

Obtain 50V from a 12V supply

If you need to generate voltages higher than the 5 and 12V levels of the circuits shown in **Figs 1 through 4**, consider a configuration such as the one shown in **Fig 5**. It provides a 50V output from a 12V input and is simpler than **Fig 4**'s circuit: Because the output is higher than the input, a simple boost configuration suffices.

The circuit uses an IRF530 n-channel MOSFET in conjunction with a MAX641 dc/dc controller. In this circuit, the 50V output is not connected directly back to the V_{OUT} pin because that pin has a maximum voltage rating of 18V. The circuit uses an external resistive divider network to provide feedback to the V_{FB} input. The V_{OUT} pin obtains power for the MAX641 directly from the 12V supply. The only components that must withstand high voltages are the MOSFET, the steering diode, and the output filter capacitor: They're rated at 100V, 200V, and 100V, respectively.

A different twist to high-voltage dc/dc conversion is the requirement to power low-voltage logic circuitry from a high-voltage source—for instance, the telephone system's −48V battery voltage. The circuit of **Fig 6** uses a basic boost configuration to convert −48V to 5V. A small-signal, high-voltage pnp transistor shifts the feedback signal from the 5V output to the MAX641, whose ground terminal (pin 3) is tied to the −48V input. The output, at 5V with respect to ground, forces about 43 μA through the 100-kΩ sense resistor and the emitter of the 2N5401. This current is sent through the 30-kΩ input resistor at V_{FB}, placing this pin 1.3V above the ground pin (or at −46.7V). Because the internal reference of the MAX641 is a 1.3V bandgap reference, the 1.3V bias level at the feedback input closes the feedback loop.

This biasing scheme allows the EXT output to directly drive the n-channel MOSFET, switching the inductor to the −48V input without level shifting of the MOSFET's drive signal. The 330-pF capacitor provides feedforward compensation, which stabilizes the regula-

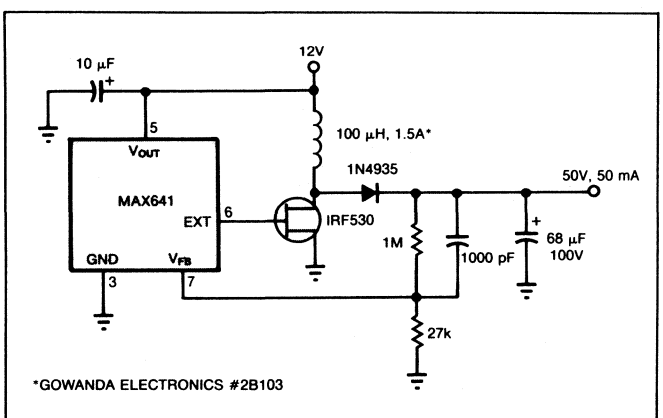


Fig 5—Only the power MOSFET, *catch diode, and output-filter capacitor need to withstand high voltages in this 50V supply circuit.*

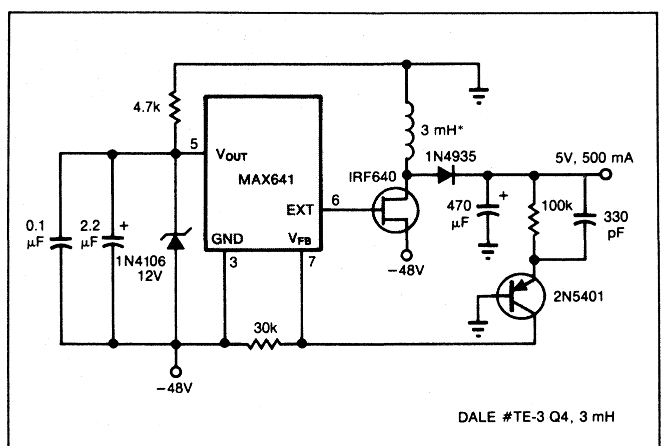
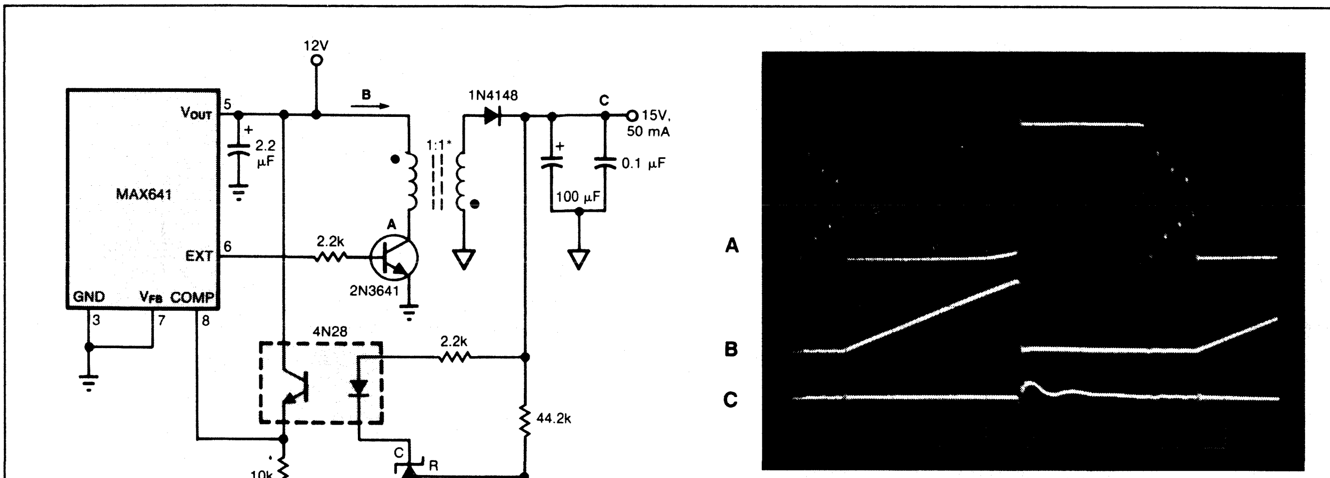


Fig 6—Telecomm applications *often require you to develop your logic-level supply from −48V. Suitable for such applications, this circuit delivers 5V at 500 mA.*

A buck/boost converter can deal with the wide input-voltage swings associated with sealed lead-acid batteries.



TRACE	HORIZONTAL	VERTICAL
A	5 µSEC/DIV	10V/DIV
B	5 µSEC/DIV	200 mA/DIV
C	5 µSEC/DIV	100 mV/DIV

***Fig 7—This circuit (a) provides 50 mA at 15V** with an isolation rating of 500V—a function of the transformer and opto-isolator. In the scope photo **(b)**, traces A, B, and C represent the switch voltage, primary current, and output-voltage ripple.*

tor's control loop and improves the regulator's transient-load response.

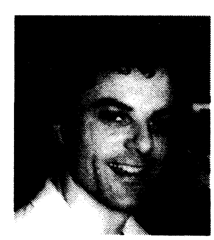
Generating an isolated supply

In large analog systems and in industrial-control systems, you must often provide power that is electrically isolated from the main system's power source. This isolation is necessary to prevent ground loops, to protect measurement hardware from dangerous voltages, and to reject common-mode signals. The circuit in **Fig 7a** generates a regulated 15V, 50-mA output that is fully isolated from the 12V input supply. The circuit's output power is supplied by a 14×8-mm pot-core transformer, and the feedback signal returns to the unisolated side of the circuit via an opto-isolator.

Although the peak primary current of the transformer is within the ratings of the MAX641 converter IC's internal switch, you must use an external transistor to drive the transformer. The reason you need this external transistor is that when the transistor turns off, the 15V secondary voltage is reflected to the primary, placing 30V across the transistor. This 30V exceeds the MAX641's 18V rating. The transformer primary's voltage, current, and ripple voltage are illustrated in traces A, B, and C, respectively, of the **Fig 7b** scope photo.

To transmit the feedback signal across the isolation barrier, the 15V output is divided and compared with the 2.75V reference of a TL431 shunt regulator. When the voltage at the TL431's reference input exceeds 2.75V, the TL431 draws current through the opto-isolator's photodiode. The opto-isolator's transistor then pulls the COMP input of the MAX641 high, turning off the EXT output. The COMP input connects to the MAX641's internal voltage divider, and thus the opto-isolator's transistor can control the MAX641. The components specified in **Fig 7a** provide an isolation rating of 500V. **EDN**

Author's biography



Leonard H Sherman is a senior member of the technical staff at Maxim Integrated Products in Sunnyvale, CA. Leonard received his BSEE from MIT, and he has one patent to his credit. Leonard enjoys playing volleyball and collecting old hi-fi equipment in his spare time.

High-speed buffers help solve problems in circuit applications

Although high-speed, unity-gain buffer amplifiers have been available for several years, recent versions serve a wider variety of applications. The high speed of today's devices makes them attractive for use in S/H circuits, active filters, and video switches.

Bob Underwood, *Maxim Integrated Products*

Modern high-speed buffer amplifiers solve a variety of circuit problems, but the design tradeoffs that increase their speed can also degrade their dc performance. Fortunately, these effects are predictable and, in most applications, correctable.

Although most of the circuits that follow will operate properly if you substitute an equivalent device, you must first check each device's specifications—particularly its input resistance, output-drive capability, and supply-voltage requirement. In some cases, the choice of a particular device can affect your circuit's performance. In other cases, you may need to adjust circuit values to optimize the buffer's performance. When your designs call for buffer amplifiers, consider one of the more popular devices, such as the LH0033, LH0063, and BB3553 (see **Table 1**). Several pin-compatible and improved versions of the devices are now also available.

Because a buffer amplifier's input provides a high-impedance load, designers often use such amplifiers in transducer or low-signal-level applications. However, the buffer amplifier isn't a lightweight contender. Its output can drive a moderate to heavy load. If a buffer amplifier's input is dc coupled to a transducer or other signal source, then the buffer's input impedance is simply its resistance and capacitance as specified in its data sheet. Note, though, that the relationship between input bias current and input voltage is nonlinear in many buffers. So, you must make certain that the input-resistance values for a particular buffer amplifier are specified over the input-voltage ranges you'll use for it.

Several common situations require ac coupling at the

TABLE 1—REPRESENTATIVE BUFFER AMPLIFIERS

TYPE	MANUFACTURERS
BB3553	BURR-BROWN, MAXIM
EL2003	ELANTEC
HA5002	HARRIS
HA5033	HARRIS
HOS100	ANALOG DEVICES
LH0002	ELANTEC, NATIONAL
LH0033	ANALOG DEVICES, ELANTEC, MAXIM, NATIONAL
LH0063	MAXIM, NATIONAL
LM110	NATIONAL
LT1010	LINEAR TECHNOLOGY CORP
MAX460	MAXIM
OPA633	BURR-BROWN

Buffer amplifiers provide high input impedance and drive a moderate to heavy output load.

buffer's input: Such coupling is necessary when you operate the buffer from a single power supply, when you remove a dc level from the signal, and when you use transducers that don't furnish a dc signal. In these applications, you must connect a resistor between a dc supply and the buffer's input to supply the buffer's bias current. The resistor's value must be low enough to supply the buffer's input current without causing too much voltage drop. However, the resistor's value must also be high enough so that it doesn't load the transducer excessively. In either event, the dc-bias resistor usually dominates the buffer circuit's input resistance. Remember that when a transducer supplies a capacitive output, the buffer amplifier's input resistance limits the low-frequency response of the circuit.

A typical bootstrap circuit (**Fig 1**) provides an input impedance that exceeds the impedance of any of the circuit components. Although the MAX460 buffer operates from a single supply in this circuit, you can reconfigure the circuit to operate with conventional split power supplies. During operation, an ac signal passes through the input capacitor, appears at the input to the buffer, and is available at the buffer's output at nearly unity gain. The output signal is capacitively coupled back to the low end of the input resistor network, which results in an effective multiplication of the resistive value. The circuit's total input capacitance arises from several sources; the intrinsic buffer-input capacitance, stray capacitances within the circuit layout, and the dc bias resistor's capacitance. You can reduce all of these capacitances by judiciously using shields and guards.

The circuit of **Fig 1** was tested while operating from a source resistance of 1 MΩ in series with a 10-pF capacitor. Under these conditions, the measured low-frequency −3-dB point was 3.3 Hz, and the low-frequency input circuit's time constant was 48 msec. These values are equivalent to those you would measure for a 4800-MΩ shunt-input resistor and a 10-pF series capacitor. The high-frequency input time constant was 0.7 μsec and had a −3-dB point of 227 kHz, which equates to an input capacitance of 0.7 pF when you use the 1-MΩ series resistor. An input capacitance this small might be difficult to reproduce because it depends greatly on the configuration of the driven guards connected to the output. Also, the buffer amplifier's case was connected to the output, and the test circuit had no special mechanical support for the input node. However, it should be possible to obtain a 1- or 2-pF input capacitance by connecting guards to the output and by supplying Teflon standoffs for mechanical support.

Buffer amplifiers can also serve in sample-and-hold (S/H) amplifier applications. In most such circuits, you need a buffer amplifier so that the output load does not discharge the hold capacitor (**Fig 2**). You can consider almost any of the available buffer amplifiers for this application, but in terms of low input current, some are better than others. For example, the MAX460 was designed for this type of application; its input current is low and is virtually independent of the input voltage.

In an S/H application, the switch's characteristics are critical. Ideally, the switch should have no offset voltage, low or zero on-resistance, and low or zero leakage, both across its contacts and from its output contact to its control terminal. Any capacitance between the switch's control terminal and its output couples a charge from the control input to the charge-storage capacitor. As a result, switching from the sample mode to the hold mode adds an error voltage to the analog signal being held by the circuit. It's sometimes possible to calibrate an S/H circuit to account for such a constant

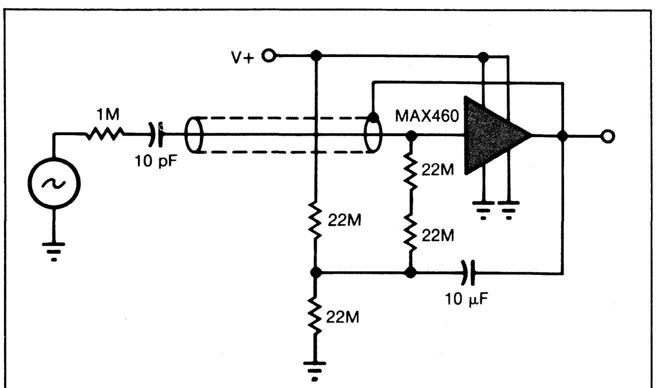


***Fig 1—This basic buffer circuit employs bootstrapping techniques** that provide an ultrahigh-impedance input for a transducer signal. The MAX460 buffer amplifier operates from a single power supply, but you can reconfigure the circuit to operate from a dual supply.*

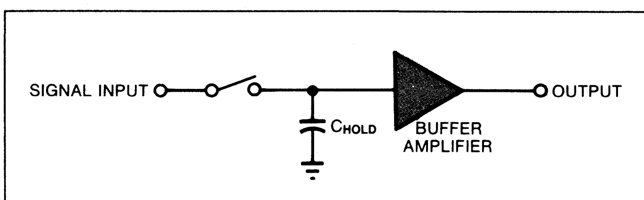


***Fig 2—This basic sample-and-hold circuit** charges a capacitor by passing an analog signal through a switch. Opening the switch holds the charge on the capacitor. The buffer amplifier's high-impedance input prevents the output load from discharging the capacitor.*

The bootstrap technique creates an input impedance value that exceeds the impedance of the individual circuit components.

Because the level-translator circuit is inherently nonsaturating, the transistor's storage time is not a critical consideration. In fact, you can use high F_T RF transistors for all four devices. The collector loads of Q_3 and Q_4 are returned to the analog input voltage (buffered by IC_1), which keeps the analog and digital circuits completely balanced. Note that the circuit also provides a set of dummy switches. The circuit's balanced nature is the secret of its excellent hold-step performance. The measured charge injection of the circuit in breadboard fashion—and without any circuit adjustments—was 100 mV into a 33-pF hold capacitor, which represents a 0.165 pJ charge injection. A stray capacitance of 1 pF between the FET switch's gate and the hold capacitor would just about account for such a small hold step.

The breadboarded circuit was adjusted, with a short piece of stiff wire, to add about 1 pF of capacitance between the compensating gate and the hold capacitor. By moving the wire, you can adjust the circuit to put out 0V for a 0V analog input. When properly adjusted, the circuit gave an output error of only a few millivolts over the S/H circuit's entire ±10V analog-input range. Also, replacing any of the semiconductor components had no effect on the hold step. In short, you don't need closely matched components for this type of S/H circuit.

Construct active filters, too

You can also use buffer amplifiers to build filters of various configurations. Although lowpass filters and other filter types do have their uses, notch and highpass filters have the most practical applications. For example, you can use a 2-pole highpass filter to remove low-frequency signal components with little effect on signals that occur above the filter's cutoff frequency. As **Fig 4** shows, the basic filter circuit exhibits a damping factor of 1 and a cutoff frequency of 1 kHz. The damping factor is controlled by the ratio of R_1 to R_2. You can scale the filter's frequency by changing the capacitor values, but both values must be equal.

Often referred to as a bridged-tee notch filter, the circuit shown in **Fig 5** consists of a series resistance and a shunt capacitance, bridged by a series-capacitance and a shunt-resistance section. In this filter circuit, dc levels pass through the series resistors, while at frequencies well above the notch the resistors have no effect. Instead, the high-frequency signals pass through the series capacitors and go on through the buffer to the output. Only the accuracy of the components, the loading by the output buffer amplifier, and any stray capacitance incurred in the filter's construction limit the maximum signal rejection at the notch frequency.

The readily available component values shown in **Fig 5** create a notch filter centered at 60 Hz. If you use 1%-tolerance resistors and 2.5%-tolerance capacitors, adjust the notch by trimming R_3 for a null at 60 Hz. When properly adjusted, the notch is deeper than −60 dB at 60 Hz, and the notch width is about 2.5 Hz at −40 dB.

Filters aren't the only signal-processing application for buffer amplifiers. You can use buffer amplifiers in video-signal switches, too. For example, an IH5352 chip lets you construct a moderate-performance 2-input, 2-output video crosspoint switch. But because the IH5352 has high shunt capacitance and high series resistance, problems can arise when you use the chip directly in high-performance video circuits. Luckily, you can alleviate the problems by buffering the video switch's input and output signals.

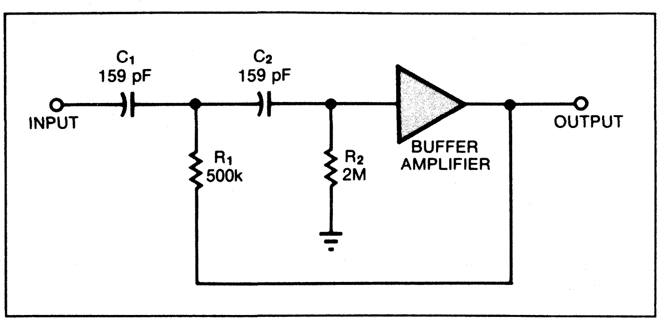


***Fig 4—This basic 2-pole, highpass filter** requires only a buffer amplifier and a few passive components. This circuit has a cutoff frequency of 1 kHz and a damping factor of one. The damping factor is set by the ratio of R_1 to R_2.*

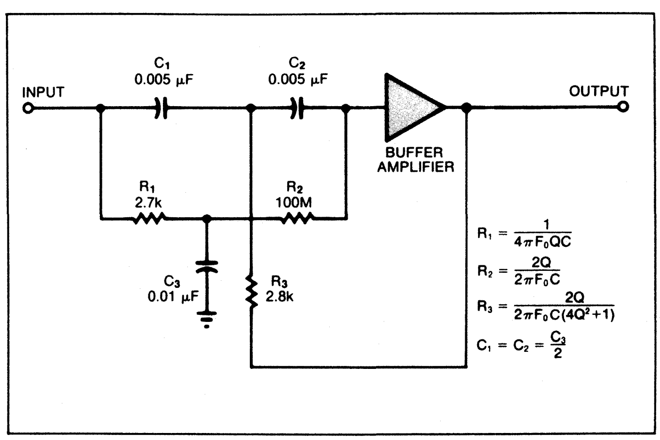


***Fig 5—This bridged-tee notch filter** produces a −60 dB notch at 60 Hz. The buffer amplifier provides the necessary high input impedance.*

error, but often the charge injection is not constant; it varies with input voltage, temperature, or time. Thus, it's best to minimize any charge-injection errors by minimizing the switch's parasitic capacitances.

The SD5000 DMOS switch IC is a good choice for S/H applications: When the chip's switch is closed, it has no offset voltage. The switch's capacitances are reasonably low, and it offers tolerably low on-resistance and leakage current. Because the SD5000 device provides four independent switches, you can connect two switches in parallel to further lower the circuit's on-resistance. You can also use the two remaining switches as a dummy capacitor, which lets you balance any charge that might be injected by the active switch section. Such a technique can result in nearly a tenfold reduction in the injected charge, without any need for manual circuit adjustments. However, if you require even better performance, you can add manually adjustable components to the circuit.

The hold capacitor is the heart of all S/H circuits. As such, it must have a high breakdown voltage and low leakage current, and it should be made of a material that has a low dielectric absorption. Typically, polycarbonate, polystyrene, and polypropylene exhibit good dielectric characteristics, but glass, mica, and most ceramics do not.

Some S/H circuits—particularly those that employ low-charge-injection techniques—may require only a switch, a hold capacitor, and an output buffer. More often, however, the circuit demands an input buffer amplifier, too. Without an input amplifier, the signal source must supply the hold capacitor's charging current.

Build high-speed S/H circuits

A realistic S/H circuit **(Fig 3)** contains an input buffer amplifier (IC_1), an output buffer amplifier (IC_2), and a DMOS FET switch. A voltage-level translator circuit made up of transistors Q_1 through Q_4 converts the ECL-compatible input signals to a voltage (referenced to the analog signal voltage) that drives the DMOS FET switches. The circuit employs Q_1 and Q_2 to form a differential amplifier that accommodates the balanced incoming ECL signals (Q and $\overline{Q}$). The four transistors are high-speed types that have a gain-bandwidth product (F_T) in the gigahertz range. Transistors Q_1 and Q_2 must have a breakdown-voltage rating that at least equals the circuit's positive supply voltage plus the ECL common-mode voltage of −2V. In this circuit, a breakdown-voltage rating of 17V is adequate. Likewise, transistors Q_3 and Q_4 require breakdown voltages of at least 25V to allow for a −10V analog input.

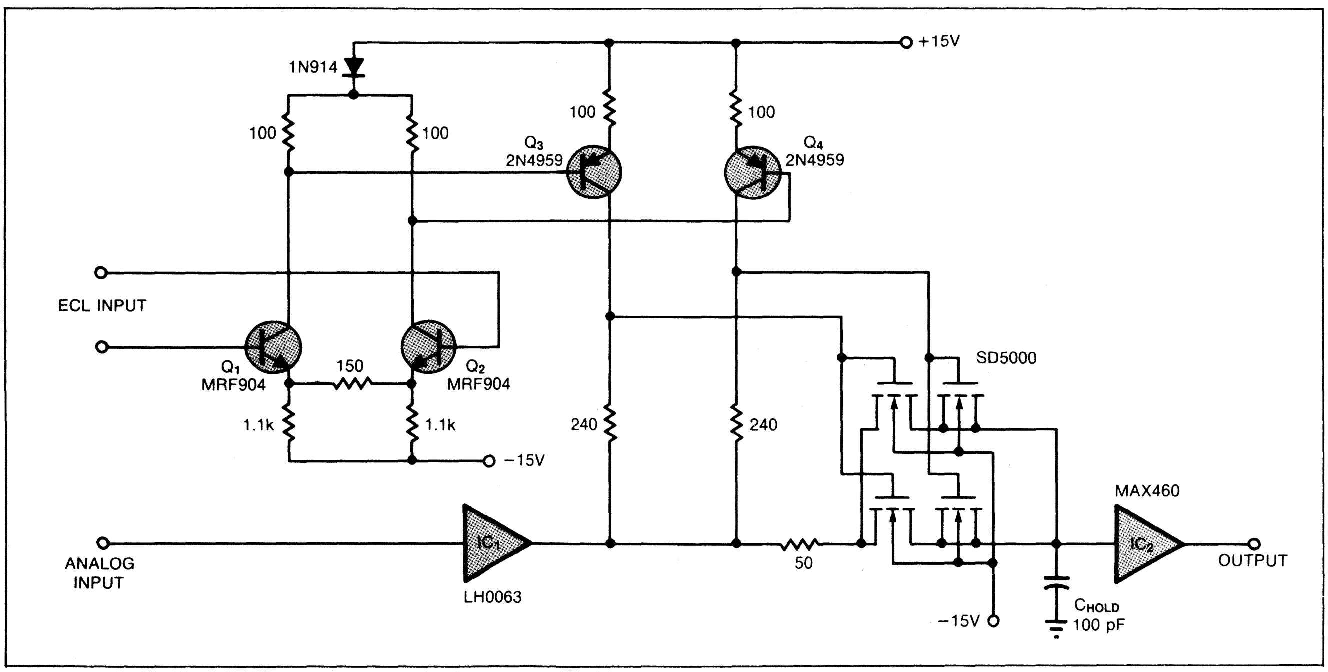


Fig 3—This high-speed S/H circuit *includes an input buffer, a DMOS switch, and an output buffer. High-frequency transistors Q_1, Q_2, Q_3, and Q_4 form a level translator that converts the ECL input signal into a voltage that can drive the switch's gates.*

In the circuit shown in **Fig 6**, the two video inputs feed through input buffer amplifiers (IC_1 and IC_2) to generate low-impedance signals that drive the analog-switch IC. Depending upon the input signal's source, each buffer-amplifier input may require a termination resistor. The termination resistors keep the amplifiers from saturating if an input is unconnected or is fed with an ac-coupled signal. A disconnected channel, for example, could turn on the switches and feed a high dc voltage to the output buffers. Because of the buffer amplifiers' input capacitances, a slight impedance mismatch may exist at the input terminals, but it usually doesn't limit the circuit's performance.

If you can accept a 6-dB loss through the switches, you can eliminate the output buffer amplifiers (IC_3 and IC_4). Such a loss might be tolerable, but switch-resistance values can vary considerably, which leads to an uncertainty about the actual signal loss. The output buffer amplifiers solve the loss-uncertainty problem and at the same time provide a low-impedance output that drives transmission lines either directly or through an accurate termination resistor. When operating from a ±15V power supply, the video circuit **(Fig 6)** handles a ±10V signal.

Restore dc levels

You can also use buffer amplifiers to restore the dc portion of a video signal. For example, because of the inherent ac coupling used in baseband video circuits, you frequently encounter video signals that have indeterminate dc sync levels. Because composite-video signals include dc sync pulses and ac picture information, the brightness of the scene influences the voltage level of the sync pulses as the complete video signal passes through an ac-coupled circuit. You can use several methods to re-reference the signal to ground.

The classic dc-restorer circuit **(Fig 7a)** passes the ac input signal through a capacitor (C_1) to a diode-resistor shunt (D_1-R_1). The time constant for R_1 and C_1 must be long enough to pass the lowest frequency component of interest in the input waveform, yet short enough to recover quickly from a fast change in the input waveform. However, the diode's forward-voltage drop produces a signal that is dc-restored to −0.7V, not to ground. By adding a resistor and an additional diode **(Fig 7b)** you can restore the video signal's sync pulse to

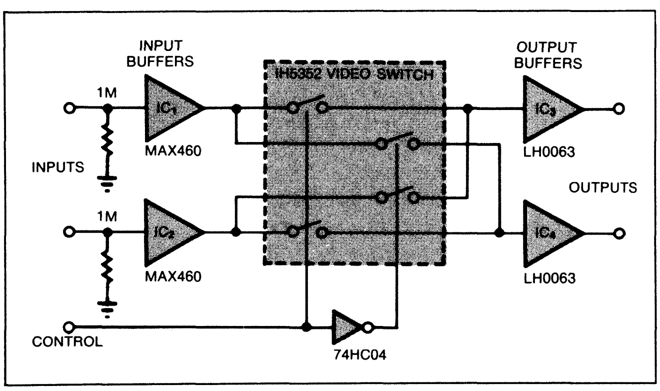


*Fig 6—**Four buffer amplifiers help overcome** the inherent limitations of a video-switch IC. The input buffers produce low-impedance signals that drive the analog switches, while the output buffers eliminate gain variations that result from unmatched switch resistances.*

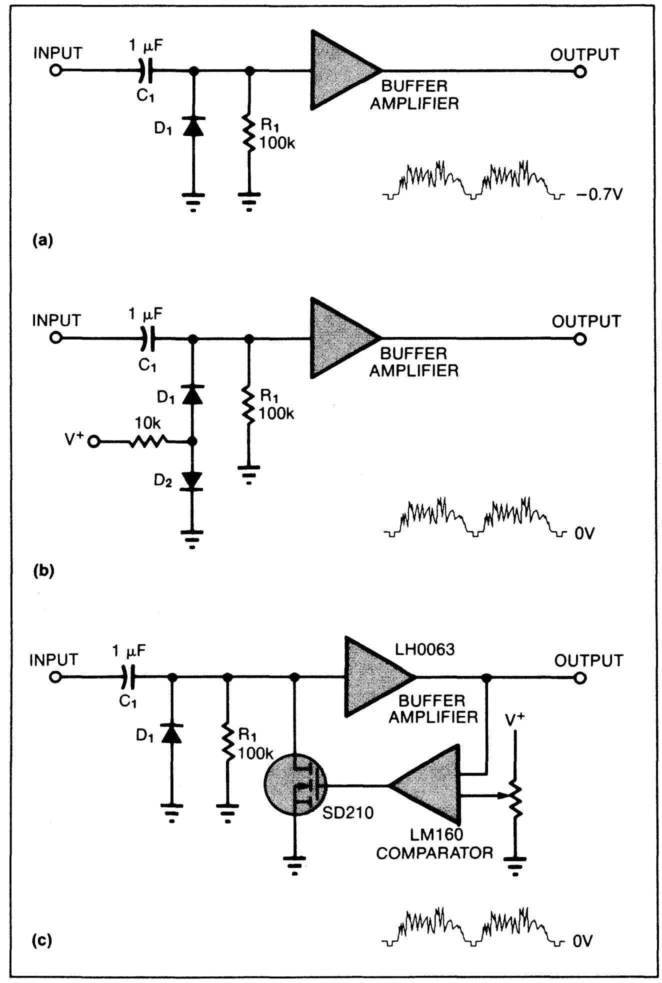
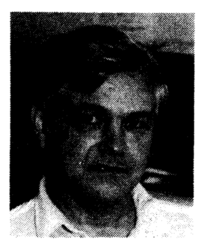


*Fig 7—**Baseband video systems frequently require restoration.** The basic circuit **(a)** couples an input signal through a capacitor to a diode-resistor combination that restores the dc reference to −0.7V. Adding a resistor and a diode **(b)** lets the circuit restore the sync pulse to 0V. You can further refine the circuit by adding an analog switch and a comparator **(c)**.*

0V. The added diode (D_2) produces a 0.7V output that should exactly cancel the forward voltage of D_1. The added diode also cancels D_1's -2 mV/°C temperature coefficient. In practice, the current in D_1 depends on the input signal's characteristics, so an exact cancellation of forward-voltage drops and temperature coefficients is difficult to achieve. You won't be able to match the diodes' voltages to better than several tens of millivolts. The circuits shown are suitable for positive signals, but they can process negative-going signals if you reverse the polarity of the diodes.

A more sophisticated approach **(Fig 7c)** requires you to add an analog switch and a comparator to the basic dc-restoration circuit. In the new circuit, you bias the comparator a few hundred millivolts above ground so that the comparator recovers the composite sync information, rejects the video part of the signal, and turns on the SD210 analog switch during sync time. The comparator's action puts a voltage on C_1 that equals the input signal's voltage during the previous sync pulse. In effect, the circuit subtracts the stored charge from the incoming waveform, which yields a 0V signal during the sync interval. Diode D_1 speeds the recovery from any fast change in the input signal's dc level. Without the diode, the comparator might force the analog switch to conduct during the complete video cycle or at least until R_1 could act to bring the buffer amplifier's input voltage back into range. **EDN**

Author's biography



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Selection criteria assist in choice of optimum reference

It's not always easy to select the most suitable precision voltage reference for your application. These devices often require parametric and economic tradeoffs. Further, parameters that are crucial in some systems are missing from or presented unclearly in many data sheets. An overview of selection criteria can help you make the choice.

Ron Knapp, *Maxim Integrated Products*

In choosing a precision voltage reference, you should look beyond initial accuracy, temperature coefficient (TC), and cost. Other factors that determine the suitability of a reference for your application are the device's power dissipation, noise, long-term stability, package size, ease of use, TC linearity, and the manufacturer's definition of TC. Familiarity with these selection criteria will help you avoid unpleasant surprises when you characterize your prototype system.

Before going into the details of the various selection factors, it's useful to briefly review the different types of references available and to explain the principles of operation of each type. The overview will give you some insight concerning the performance you can expect from the various references. Reference circuits comprise three categories: bandgap cells, zener-diode-based references, and heated-substrate types. Most voltage references fall into the first two categories and derive their fixed output from a bandgap cell or a zener diode. The third type of reference obtains additional stability by mounting the bandgap or zener circuit on a heated substrate.

Bandgap references depend on the behavior of diodes (or the equivalent base-emitter junctions of transistors). The following equation predicts the operation of such junctions with a high degree of precision.

$$V_{BE} = V_{G0}\left(1 - \frac{T}{T_0}\right) + V_{BE0}\left(\frac{T}{T_0}\right) + \frac{nkT}{q}\ln\left(\frac{T_0}{T}\right) + \frac{kT}{q}\ln\left(\frac{I_C}{I_{C0}}\right),$$

where

- V_{G0}=the extrapolated bandgap voltage (about 1.2V) at 0°K
- n=process-dependent constant; value 1.5 to 3
- q=charge of an electron
- k=Boltzmann's constant
- T=temperature in °K
- I_C=collector current
- T_0=reference temperature for V_{BEO} and I_{CO}
- I_{CO}=reverse saturation current at T_0
- V_{BEO}=V_{BE} value for the conditions T_0 and I_{CO}.

The diode's temperature coefficient is large but predictable and repeatable (−2 mV/°C or −3100 ppm/°C). Thus, you can achieve stability by balancing the diode's

The V_{BE} equation's third and fourth non-linear terms limit the performance of bandgap references by making a flat voltage/temperature response impossible.

TC with a TC of equal magnitude and opposite sign. Such a TC exists for the *difference* between the forward voltages of two diode junctions operating at different current densities. Because the ratio of mismatch governs the TC's value, the bandgap circuit is compatible with good IC design—parameter values should depend on accurate ratios based on layout geometry, rather than on absolute quantities that are difficult to control.

You can calculate the desired difference voltage (ΔV_{BE}) with high predictability, directly from the diode equation

$$\Delta V_{BE} = \frac{kT}{q} \ln\left(\frac{J_1}{J_2}\right),$$

where J_1/J_2 is the ratio of current densities. To obtain zero TC, you add the expression for V_{BE} to the one for ΔV_{BE}, differentiate the sum with respect to temperature (T), and set this quantity equal to zero. The result is

$$V_{G0} = V_{BE0} + \frac{kT}{q} \ln\left(\frac{J_1}{J_2}\right).$$

Solving this equation for the J_1/J_2 ratio tells you that an approximate 8:1 ratio gives the best result (a near zero TC). Scaling the transistor areas gives an IC designer accurate control of this ratio.

In a basic bandgap circuit **(Fig 1)**, V_{BE} is the base-emitter voltage of Q_1, and ΔV_{BE} appears across R_2. The ratio of R_1 and R_2 scale ΔV_{BE} to a voltage (V_1) whose TC cancels the TC of V_{BE}. The amplifier then raises the 1.2V sum of V_1 and V_{BE} (the bandgap-cell voltage) to a higher level at V_{OUT}: usually 2.5 to 10V. Unfortunately, the amplifier multiplies noise as well. A 10V scaled output, for example, increases the bandgap cell's noise voltage by an approximate factor of 8 ($10 \div 1.2$).

Commonly available bandgap-reference voltages are 10, 5, 2.5V, and the bandgap-cell voltage itself, 1.23V. Typical TCs range from 5 to 50 ppm/°C. The V_{BE} equation's higher-order, logarithmic third and fourth terms limit the performance of these references by making a flat voltage-temperature response impossible. What's more, some of the equation's coefficients are process-dependent—particularly n, which is related to the carrier mobility of dopant in the silicon. The quantity n poses a problem because you cannot easily determine its value by making electrical measurements during production.

Because most bandgap references are constructed in silicon monolithic form, they are relatively inexpensive ($3 to $20). Many designs employ curvature correction to compensate for the logarithmic nonlinearity in the TC, but none offer an exact cancellation.

Zeners have rock-bottom TCs

The second type of voltage reference—based on a zener diode—achieves TCs as low as ±1 ppm/°C. Zener diodes have a positive or negative TC, depending pri-

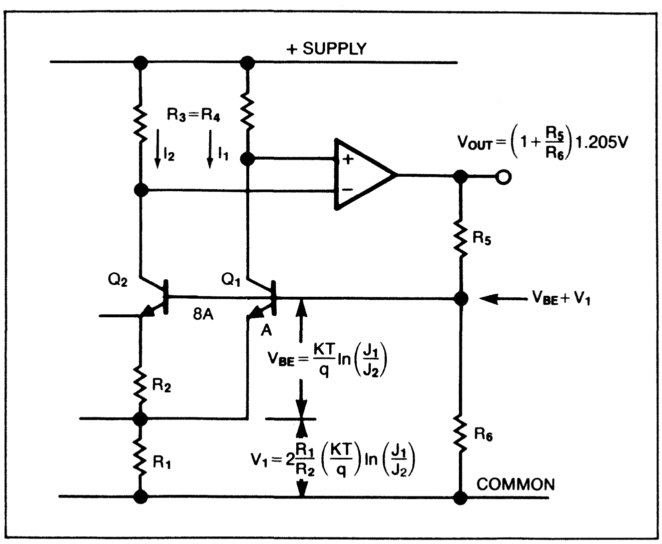


***Fig 1—A bandgap voltage reference** generates the sum ($V_{BE}+V_1$), in which the two voltages have equal and opposite temperature coefficients. The amplifier then raises the sum to a more convenient voltage level.*

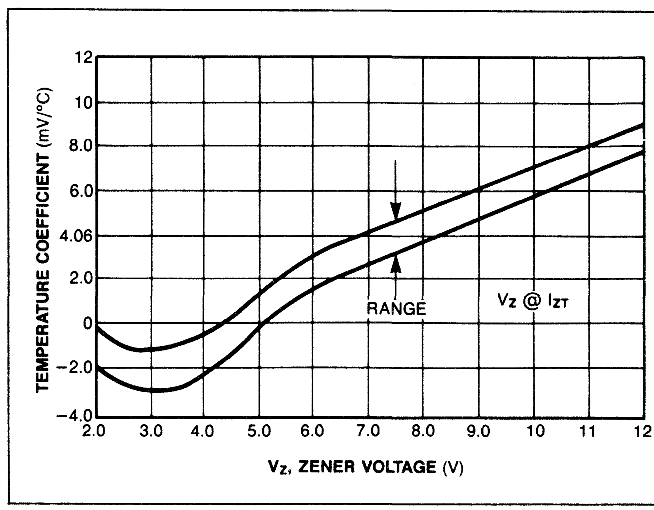


***Fig 2—Zener diodes produce a zero-TC voltage** near 5V—the level for which the mechanisms of negative-TC field-emission breakdown and positive-TC avalanche breakdown are in balance. However, the zero-TC ideal is difficult to achieve on a production basis.*

marily on the breakdown-voltage value and to a lesser degree on the operating current. The zener breakdown involves two mechanisms: field-emission breakdown, which dominates below 5V and produces a negative TC, and avalanche breakdown, which occurs above 5V and yields a positive TC. Although complex and difficult to quantify, these breakdown mechanisms should be in balance at approximately 5V, yielding a near-zero TC. Tests corroborate this contention **(Fig 2)**.

Unfortunately, 5V zener diodes exhibiting the utopian zero TC are difficult to produce. The problem is that the negative TC breakdown mechanism is flukey and difficult to reproduce consistently in production. The positive TC breakdown, on the other hand, is predictable and eminently repeatable for devices using routine semiconductor-production processes. Another characteristic that's predictable and repeatable is the negative-slope temperature characteristic of a forward-biased diode.

Because of the difficulty of producing a zero-TC zener diode that depends purely on breakdown mechanisms, it's evident that the TC of a zener-diode reference should not depend solely on the absolute zener breakdown voltage. A class of zener diodes, called TC zeners, takes a compensatory approach by balancing the negative TC of a forward-biased diode (−2 mV/°C) with the equal and opposite TC of a 5.6V zener diode. The output voltage is therefore 6.3V (0.7V+5.6V). These references offer 5- to 100-ppm/°C TCs and require operating currents from 0.5 to 7.5 mA. You must maintain the specified operating current to obtain the guaranteed TC.

Precision references need laser trimming

To achieve accuracies as tight as ±0.01% in precision references, manufacturers use laser-trimmed thin-film resistors. Diffused resistors embedded within silicon exhibit not only hysteresis, but also high TC, poor TC matching, large voltage coefficients, and poor stability. Thin-film resistors, deposited on the chip's surface, are found in such voltage references as the REF01, AD581, AD2700, and the MAX670.

The secret to the precision references' accuracy is to trim the thin-film resistors by laser before attaching a lid to the package. This critical operation determines a reference's initial accuracy and its long-term stability. Fuse-link blowing and resistor-link trimming are alternative schemes for trimming the absolute voltage, but the chip area required with these methods makes them prohibitively expensive for tight-tolerance adjustments.

Thick-film resistors have insufficient stability for use in precision references; therefore, hybrid products such as the MAX670, AD2700, and AD2710 include TaN (tantalum nitride) or NiCr (nichrome) thin-film resistors, sputtered on a ceramic substrate of 99.6% alumina (Al_2O_3). Before trimming each lot of references, the manufacturer determines the optimum settings for laser power and focus by executing a test matrix of experimental laser cuts.

For each power setting, the system makes a staircase trim pattern in which each right-angle turn marks an additional increment of focus **(Fig A)**. After completion of the focusing and system-calibration steps, quality-control personnel inspect the trim process every 30 minutes to ensure uniform cuts throughout the manufacturing lot. The system achieves extremely clean trims in this way **(Fig B)**. To prove its stability, each device must maintain initial accuracy after trim during a 48-hour, 150°C burn-in operation.

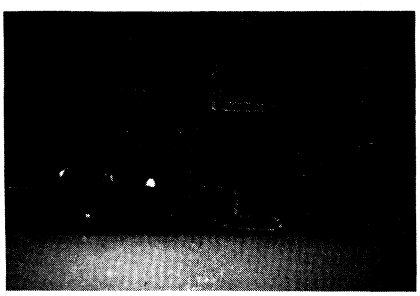


Fig A—A staircase test matrix helps to optimize focus and power levels *in a laser system used for trimming precision thin-film resistors.*

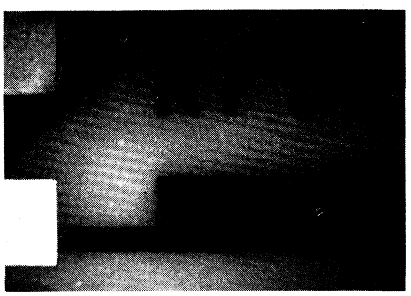


Fig B—After calibration, a laser-trim system cuts cleanly *through a thin-film resistor. The calibration depends on the staircase setup technique of* ***Fig A.***

You can easily achieve a 1-ppm/°C TC by mounting a zener-reference circuit of reasonably low TC on a heated substrate.

The AD2700 and MAX670 series of hybrid references, for example, use a 1N827 zener diode—chosen for low noise, low dynamic impedance (10Ω max), and good TC linearity. (Why use a hybrid? Fabrication of these TC zener references involves a specialized process, involving extra steps not always available in a standard bipolar process.) The products' initial 10-ppm/°C TC is that of the zener diode. Active laser trimming then lowers the TC by adjusting the zener-diode current, thereby creating additional 3- and 1-ppm/°C product grades (see **box**, "Precision references need laser trimming").

The manufacturer calculates the required zener current using actual TC values, obtained through oven tests on unsealed devices. Note that the amplifier in **Fig 3** supplies current to the zener, which in turn supplies an input voltage to the amplifier. To ensure circuit startup, R_4 supplies current to the zener and the amplifier uses ground as its negative supply, thereby eliminating $V_{OUT}=0V$ as an unwanted stable state. Note that the amplifier in a zener reference contributes less output noise than does the amplifier in a bandgap reference, because the zener voltage requires less amplification.

Heater trades P_D for stability

The third type of reference, based on either a bandgap or zener voltage, uses a local heater to maintain the substrate at a constant temperature, usually 10 or 15 degrees above the upper limit of the operating range. If the circuit's TC is reasonably low (20 to 30 ppm/°C), such a reference can easily achieve a TC of 1 ppm/°C. The disadvantage is power dissipation—an LM199 at −55°C, for example, requires as much as 28 mA at 15V for the heater alone.

Also, the LM199's output voltage stabilizes at 1 ppm/°C but the initial accuracy is only ±5%. To meet the ±0.1% or ±0.01% tolerances required in data-converter applications, therefore, you must add a precision op amp and scaling resistors and then cope with these components' additional cost and error contributions. The proper evaluation of a reference application involves these issues as well as many of the following ones, which are not always covered explicitly in the data sheet.

Confusion surrounding the specification of temperature coefficient, for instance, is partly a matter of definition. Two definitions are popular. In the "box" method, V_{OUT} for an in-spec device must remain within a rectangle formed by T_{MIN}, T_{MAX}, and the maximum specified ΔV_{OUT} (**Fig 4**). ΔV_{OUT} is the product of the nominal output voltage (V_{NOM}), the specified TC, and the operating-temperature range. For the AD2700L,

$$\begin{aligned}\Delta V_{OUT} &= V_{NOM}\,(TC)\,(T_{MAX} - T_{MIN})\\ &= 10V(3\ ppm/°C)\,[85-(-25)°C]\\ &= 3.3\ mV.\end{aligned}$$

In other words, V_{OUT} will change no more than ±3.3 mV between any two temperatures in the operating range. This maximum change, added to the ±2.5-mV initial-accuracy spec, produces a total error band of 5.8 mV above and below the nominal V_{OUT} (10V).

The "butterfly" method, on the other hand, refers everything to 25°C and allows the manufacturer to use

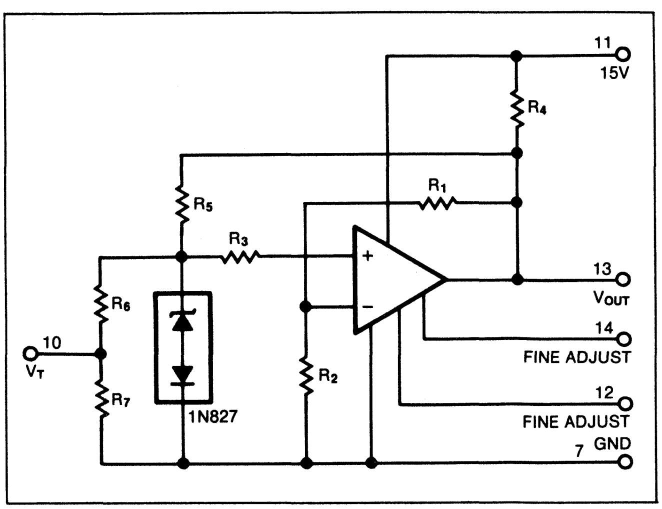


Fig 3—The amplifier in this zener-diode reference bootstraps the zener voltage by delivering current to the zener while the zener delivers voltage to the amplifier. R_4 provides start-up current to the zener.

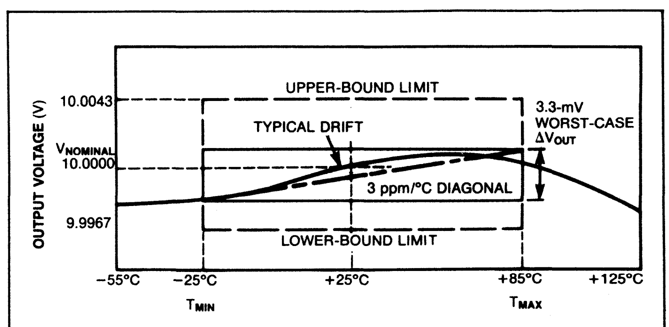


Fig 4—In the "box method" of specifying TC, the operating-temperature range and the maximum allowed change in V_{OUT} form the sides of a rectangle, and the slope of the rectangle's diagonal becomes the TC.

different TCs in determining the error bands at temperatures above and below 25°C (**Fig 5**). The AD2710K, for example, specs a change of ±0.9 mV over the 25 to 70°C range (10V×2 ppm/°C×(70−25)°C). You must add to this the initial tolerance of ±1 mV at 25°C, resulting in a maximum possible error of ±1.9 mV at T_{MAX} (70°C).

Such systems as DVMs and data-acquisition instrumentation often use the box method for specifying total error, because users aren't likely to calculate accuracy using the TC specs. This approach has a disadvantage —the whole 3.3-mV error change in the example of **Fig 4** could occur between, say, 25 and 70°C, yielding an effective TC of 7.33 ppm/°C, which exceeds the maximum specified TC (3 ppm/°C). A worst-case analysis over temperature, however, must allow for this much change anyway, regardless of where it occurs in the operating-temperature range.

Because temperature testing plus the reading and recording of data are costly, manufacturers usually base TC specs on only a few data points. These should include at least 25°C and the endpoints (T_{MIN} and T_{MAX}). Using the endpoints alone, for example, can make the reference appear better than it actually is if the TC curve is symmetrical and parabolic.

You should avoid using "typical" specs for TC and absolute accuracy; only tested and guaranteed limits for minimum and maximum have meaning. A data sheet should also identify the temperatures used in the calculation of the device's TC. The AD2700L data sheet, for instance, lists 25°C plus the endpoints (−25 and 85°C). The AD2700U data sheet lists these three as well as the extended endpoints of −55 and 125°C.

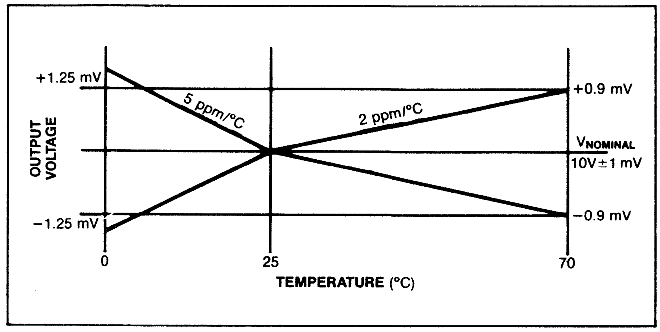


***Fig 5—The "butterfly" method of TC specification** normalizes the variation of V_{OUT} with respect to 25°C. You then extend wing-shaped error bands to the operating-temperature extremes.*

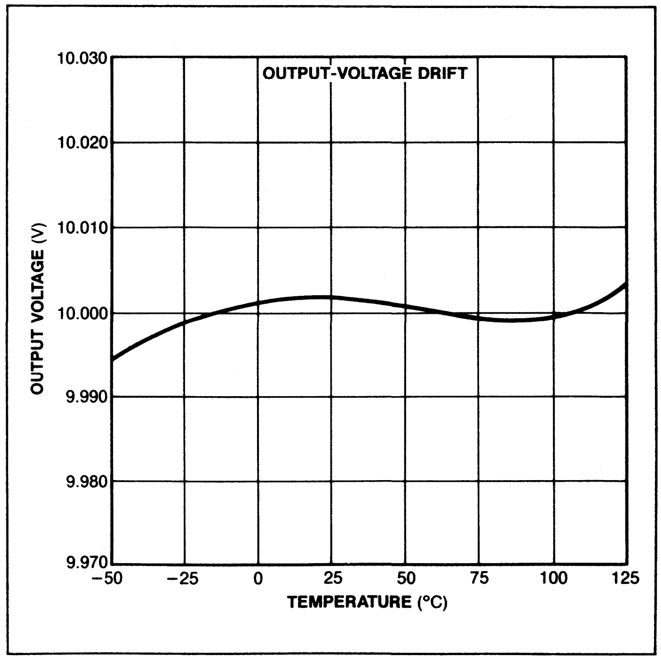


***Fig 6—The AD581's V_{OUT}-vs-temperature characteristic has an S-shaped curve.** This characteristic is typical for bandgap references that include correction circuits for TC linearity.*

Correction yields S curve

Although voltage-reference data sheets seldom specify TC linearity, the characteristic curves for V_{OUT} over temperature contain the most useful TC-linearity information that a manufacturer can provide. For bandgap references these curves are parabolic or S-shaped (**Fig 6**), depending, among other factors, on whether the device includes a linearity-correction circuit. The TC linearity of zener-based references depends mainly on the zener diode, and the reference will include one of two diode types, depending on the intended temperature range and the desired linearity (see **box**, "Zener diodes determine TC linearity").

Another important specification is noise, which appears on most data sheets as a typical value but seldom has a guaranteed limit. Because noise testing is difficult, manufacturers usually guarantee maximum values by performing sample testing only, if that. What's more, because a designer can easily filter or band-limit the higher frequencies by adding capacitors, noise specs cover the 0.1- to 10-Hz range in nearly all cases. (The suppression of low-frequency 1/f noise, however, requires impractically large capacitor values.)

Data sheets usually specify noise in terms of $nV/\sqrt{Hz}$, an expression that allows you to calculate output noise for the bandwidth of interest. At the same time, you usually convert this quantity to the more useful μV p-p, especially for converter applications:

Bandgap references usually have a parabolic TC characteristic that assumes an S shape if the device includes circuitry to effect linearity correction.

First, multiply nV/√Hz by the square root of the system bandwidth to obtain the noise magnitude in nV rms. Then (assuming the noise has a Gaussian distribution), multiplication by 6 will give you the approximate peak-to-peak noise you can expect for that bandwidth.

Noise measurement is difficult

Lack of equipment is part of the difficulty manufacturers face in measuring noise. For example, Quantec makes a noise tester commonly used for testing op amps and transistors, but that instrument requires a nominal 0V bias for the circuit node under test. Spectrum analyzers make good noise testers, but not many have the dynamic range and the low noise floor necessary to measure, say, 10-μV signals riding on 10V dc. Frequency range is another complication. Spectrum analyzers come in high- or low-frequency models (above or below 100 kHz), so one model doesn't cover the measurement range needed for many applications—0.1 Hz to several megahertz.

You can measure noise directly using a Tektronix storage oscilloscope with a 7A22 plug-in amplifier, which has 10-μV/div sensitivity and selectable lowpass and highpass filters that cover 0.1 Hz to 1 MHz. The

Zener diodes determine TC linearity

The TC linearity for a zener-based voltage reference depends on the type of zener diode in the device. Most hybrid references include one of two types of TC zener (in die form), and only a few zener manufacturers can guarantee 5- to 10-ppm/°C performance for these products. One zener type has an alloy-diffused junction in a vertical configuration **(Fig A)**, wherein the anode serves as a bond pad on top of the die and the cathode as the substrate (backside) of the chip.

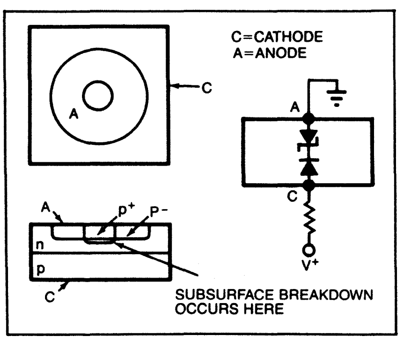


***Fig A—Alloy-diffused zener diodes** feature a vertical configuration in which a top-surface bond pad forms the anode connection and the die substrate forms the cathode connection.*

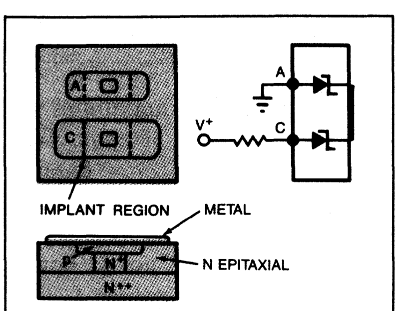


***Fig B—The lateral geometry of ion-implanted zener diodes** places both diode connections on the top surface of the chip.*

The other type of zener features an ion implant and lateral geometry **(Fig B)**, and has both connections on top of the chip. For this type, the substrate must float unconnected, because the substrate is the junction of two zener diodes—one operating as a zener in the breakdown mode, and the other operating as an ordinary forward-biased diode. The zener voltage is 5.6V,

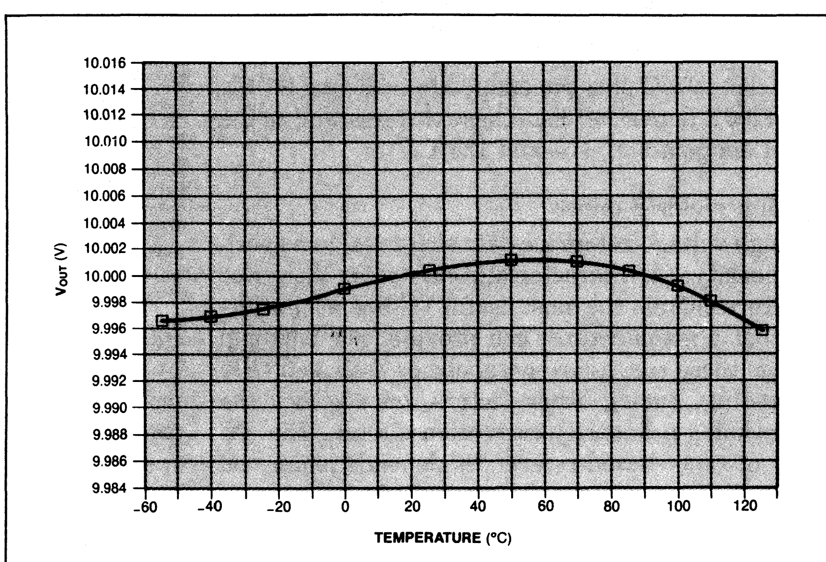


***Fig C—An ion-implanted-zener reference** such as the AD2700 exhibits a concave-down TC characteristic and better overall linearity than does a diffused-zener type for the range −55 to +125°C.*

lowpass settings don't include 10 Hz, however, and the amplifier's input-voltage limitation may require that you ac-couple the signal. The coupling capacitor then forms a highpass filter of a few hertz that precludes the use of the 0.1-Hz highpass setting.

For a more convenient method of noise testing with a storage oscilloscope, you use a low-noise op amp configured for a gain of 100, a 0.1-Hz highpass input filter, and a 10-Hz lowpass output filter **(Fig 7).** The gain boosts 10-μV signals to 1 mV—within the range of most oscilloscopes—and allows use of an OP07A (whose 0.6-μV p-p max noise contributes less than 60 μV p-p noise at the output).

To measure noise, set the scope amplifier's vertical-input coupling to dc. Allow the filter to settle and the reference to warm up (about 30 sec in most cases). Clear the screen in storage mode and set the time base for single-trigger mode at 1 sec/div. Set the scope to save mode or maximum screen persistence and measure the peak-to-peak noise for 10 sec. (Observation for 10 seconds is the accepted method, even though the time constant for 0.1 Hz is only 1.6 sec.) A scope photo based on this technique **(Fig 8)** shows about 20-μV p-p noise for the AD581—typical for most bandgap references—

and when operated at the proper current, it produces a TC of 2 mV/°C—a TC equal to and opposite that of the forward-biased diode. For this reason, nearly all temperature-compensated zener diodes have a total voltage of 6.3V (5.6+0.7V). You can create a higher output voltage by connecting multiple forward-biased diodes in series with a higher-voltage zener diode.

Both TC-zener types specify V_{OUT} as 6.3V±5%, but the actual tolerance for ion-implanted types is tighter (typically ±40 mV, or ±0.6%), vs ±300 mV (±4.7%) for alloy-diffused types. The tighter tolerance of ion-implanted zener diodes allows the reference manufacturer to target gain-resistor values more closely, do less laser trimming, and thereby provide better V_{OUT} stability.

TC linearity is the most noticeable difference between the two zener types. The implanted zener's concave-down curve exhibits better overall linearity from −55 to 125°C **(Fig C),** but the diffused zener has better TC linearity from 0 to 70°C **(Fig D).** Both the forward-biased diode and the zener diode contribute to the nonlinearity, and these effects increase at low current.

Accordingly, most TC zeners have operating currents in the 0.5- to 7.5-mA range, which is an order of magnitude higher than that of zeners normally found in an IC. High current (sufficiently beyond the value at the zener's breakdown voltage) also ensures low noise.

Though it's a tedious procedure, you can always characterize the reference over temperature and then compensate for the TC nonlinearity by using a temperature sensor, A/D converter, and software lookup table. The well-controlled ion-implant process offers a compromise solution, however—the use of zener diodes in which the TC curves and 25°C voltages are repeatable from lot to lot.

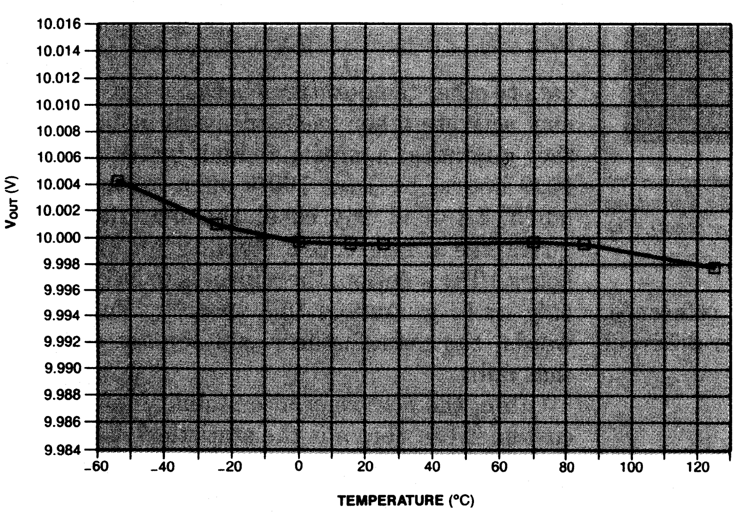


Fig D—The output of a diffused-zener reference *such as the AD2710 provides excellent TC linearity from 0 to 70°C, but suffers in linearity outside that range.*

Often, the statistical data taken by the manufacturer on life-test samples is the best stability information you can obtain about a reference.

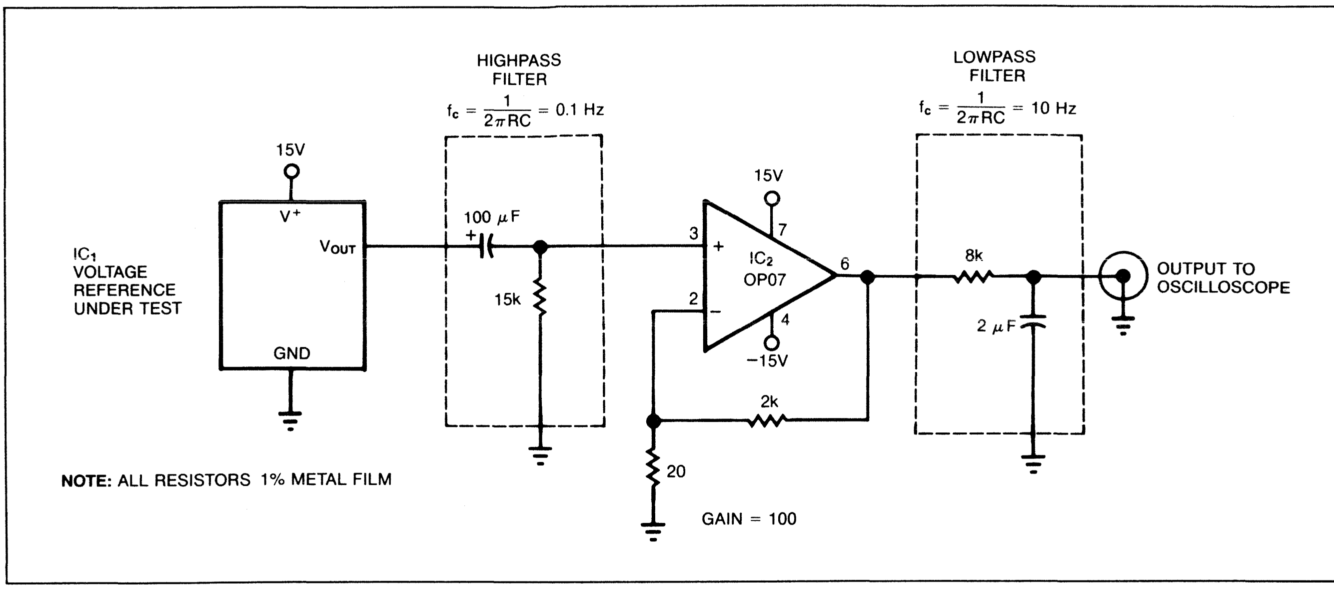


***Fig 7—Introducing highpass and lowpass filters** and a low-noise op amp lets you measure voltage-reference noise using a storage oscilloscope.*

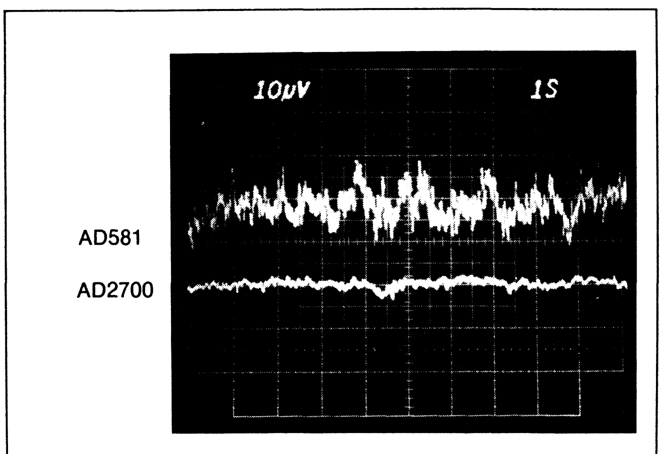


***Fig 8—This scope photo shows the noise levels** typical for a bandgap reference (upper trace) and a zener-based reference (lower trace). The scale is 10 µV/vertical div; 1 sec/horizontal div.*

TABLE 1—REFERENCE-NOISE COMPARISON

	NOISE (µV p-p)	
BANDWIDTH	AD581 (BANDGAP)	AD2700 (ZENER)
0.1 TO 10 Hz	20	5
1 TO 100 Hz	50	8
1 Hz TO 3 kHz	220	30
1 Hz TO 300 kHz	600	200

and about 5 µV p-p for the AD2700 zener reference. **Table 1** compares noise for these devices over different bandwidths.

Long-term stability

Long-term stability can be the most important spec in a reference application, but—as in the case of noise—this parameter seldom receives a thorough characterization in the data sheet. Most manufacturers specify stability as 25 to 100 ppm (typ) per thousand hours at 125°C. They cannot accurately extrapolate this stability data to other temperatures because those temperatures may activate other mechanisms of instability. Nor can they guarantee a maximum limit by testing all parts for 1000 hours, because 100% burn-in testing costs too much. (And in any case, the manufacturer cannot guarantee a reference's stability for the second 1000 hours.) The solution, therefore, is to either test samples only or to guarantee this spec "by design" (in other words, the manufacturer will replace customer parts that fail).

Often, the best reference-stability information that a customer can obtain is the statistical data taken by the manufacturer on life-test samples. Maxim, for example, records long-term stability for a set of sample devices operating continuously for several thousand hours at 55°C (a realistic operating temperature that is higher than the room ambient temperature but lower than T_{MAX}). Such data **(Fig 9)** for the AD2700, for instance, *Output-current specs are misleading unless they specify V_{OUT} limits such as those in the spec for load regulation.*

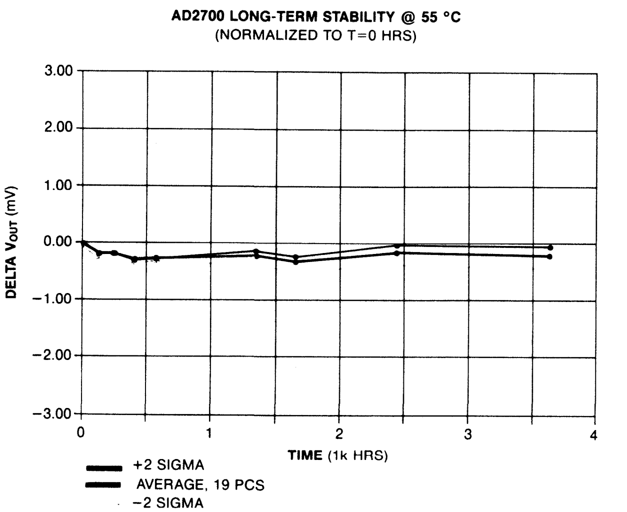


***Fig 9—The average stability of AD2700 voltage references over 3600 hours** at 55°C appears in the center curve. The upper and lower curves denote 2-sigma boundaries that encompass 90% of the 19 units tested.*

shows that V_{OUT} drifts about 250 μV negative and then remains within ±50 μV of that level. The center curve represents typical performance; the upper and lower "2-sigma" curves encompass 90% of the devices, based on the standard deviation of measured values.

I_{OUT} specs can be misleading

Output-current specs are misleading unless they specify V_{OUT} limits such as those included in the spec for load regulation. Note how this parameter reveals important differences in several reference devices. The AD2700, for example, has a 741-type output circuit that can sink and source current equally well within a range of ±10 mA. V_{OUT} changes 0.5 mV max for a 0- to 10-mA change in output current, resulting in a load regulation of 50 μV/mA max.

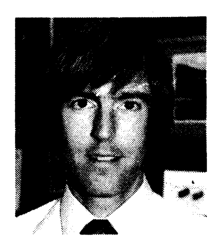
The MAX671 has Kelvin outputs that provide load regulation of 10 μV/mA max. The 10V REF01 monolithic reference, on the other hand, has a simple emitter-follower output that can only source current (to ground); its load regulation is 1 mV/mA max over 0 to 10 mA. For the AD580, this same 1-mV/mA limit represents lower performance because V_{OUT} is only 2.5V. The 10V references AD581 and AD584 can source as much as 10 mA at 25°C but specify the load regulation (500 μV/mA max) to only 5 mA. These two devices have limited current-sinking capability over the MIL temperature range. They guarantee 5-mA source current over the full operating-temperature range.

Measure V_{OUT} vs V_{SUPPLY}

Line regulation and power-supply rejection ratio (PSRR) are two other important parameters for voltage references. They represent the change in V_{OUT} that results from fluctuations in supply voltage. Line regulation is a dc test whose results are usually expressed in μV/V or mV/V. PSRR can be a dc test, but usually the test conditions for this parameter include a range of frequencies or a specific frequency. The line-regulation spec has the advantage that self-heating effects are included in the output-voltage change. PSRR, on the other hand, has more realistic test conditions. At 60 Hz in particular, self-heating effects average out but the power supply may offer poor regulation, degrading the stability of V_{OUT}.

Finally, consider the implications of temperature hysteresis in your application. A reference output V_{OUT1} at temperature T_1 should return to V_{OUT1} after you cycle the device to T_2 and back to T_1. If not, the output exhibits hysteresis. The cause is thermal stress within the IC, which in turn causes expansion of the silicon with temperature—and this effect is aggravated by the contact of dissimilar packaging materials that have different coefficients of expansion. With the exception of that for the LT1021 (Linear Technology Corp, Milpitas, CA), voltage-reference data sheets rarely mention hysteresis. **EDN**

Author's biography



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Back-to-basics approach yields stable references

Achieving the accuracy and stability that IC voltage references promise isn't necessarily a "piece of cake," but if you return to your EE roots and do a little old-fashioned analog-circuit analysis, you can obtain impressive results.

Ron Knapp, *Maxim Integrated Products*

Analog-IC manufacturers make it *look* simple to achieve the voltage-reference accuracy and stability that used to present major challenges for circuit designers. Today, obtaining stability of a few parts per million per degree should be a routine task. Nonetheless, ignoring facts of life such as noise and I·R drops can transform a seemingly simple job into one as complex as that faced by reference designers 20 years ago. Attention to circuit basics can make obtaining precise rock-solid reference voltages in the late 1980s as uncomplicated as the vendors of the ICs intend it to be.

Selecting a low-temperature-coefficient, precision voltage reference starts with careful consideration of your noise requirements. If the reference is too noisy, the highest dc accuracy and the cheapest price won't mean anything. Determine the signal-to-noise ratio your application requires. If you intend to use the reference with an A/D or D/A converter, the reference noise should be less than 1/10 the resolution. For example, a 12-bit ADC with a 0 to 10V input has a 1-LSB resolution of 2.44 mV. The maximum noise from the reference should be no more than 240 μV p-p. In this case, a bandgap reference such as an REF01 or AD581 will suffice. For a 14-bit converter with an LSB size of 610 μV, a noise limitation of 60 μV will require an AD2700, MAX670, or equivalent.

Often, you can lower the wideband noise with a large capacitor placed on the reference-device output. A 10-μF capacitor is large enough to prevent oscillation problems and will typically decrease the high-frequency noise (above 1 kHz) by a factor of 3 or 4. Some references, like the AD584, have noise-reduction pins that allow you to add an external capacitor. A smaller capacitor (0.01 μF) placed in parallel with the feedback resistor (if the inverting input of the reference amplifier is available) will filter the noise from both the reference device and the amplifier, but will also adversely affect the turn-on time and the circuit's response to load changes. Nevertheless, there is little you can do about low-frequency noise, and therefore most reference data sheets place great importance on noise in the 0.1- to 10-Hz band.

Thermal effects take second place

Taking into account noise considerations, the second most important spec of a voltage reference is the temperature coefficient, or TC. Don't ignore initial

Ignoring facts of life like noise and I·R drops can turn what appears to be a simple job into a complex one.

accuracy, though, or take it too lightly, thinking that you'll be able to adjust it. Remember that any components you add can jeopardize the TC, long-term stability, and reliability—all it takes is one component that drifts out of calibration. For example, using a reference's trim-adjust feature or scaling its output with an external gain stage will probably affect the TC.

To set the gain of an op amp, you should use TC-matched thin-film resistor networks. When you use separate RN55D metal-film resistors with TC specs of ±50 ppm/°C, you introduce a TC error of 100 ppm/°C if one resistor TC is +50 ppm/°C and the other is −50 ppm/°C. The fine-trim adjustment of the AD2700 is somewhat interactive with the TC; 1 mV of adjustment changes the TC by 4 μV/°C or 0.64 ppm/°C referred to the output. The best advice to follow about reference trim adjustment is "don't do it." It is much better to calibrate the gain elsewhere—at the D/A converter, for example. Better yet, make gain adjustments in software. If there is a temperature-independent gain trim elsewhere in the system, you can use the AD2700's fine-trim output-voltage adjustment to change the device's TC (and, incidentally, affect its output voltage).

High-resolution converters need TLC

Most high-resolution converters, such as 16-bit DACs and ADCs, guarantee linearity consistent with resolution, but rarely do they guarantee absolute accuracy, which includes gain error, to that level. Because the gain accuracy depends primarily on the accuracy of the voltage reference (whether internal or external), the temperature coefficient of the reference determines the useful temperature range for converter accuracy **(Fig 1)**. Applications where absolute accuracy is critical include weighing scales, data-acquisition measurement systems, automatic test equipment (ATE), and laboratory instruments (such as DVMs and programmable voltage standards).

First, consider a 12-bit-system example. A 12-bit A/D converter with its linearity specified to ±½ LSB requires a reference with a TC of no more than 2.67 ppm/°C from 25 to 70°C to maintain a gain accuracy within ½ LSB, or 1.2 mV out of 10V FS. A suitable reference would be the AD2710KD, which is specified to 2 ppm/°C max from 0 to 70°C.

For a 16-bit system, you can use a reference guaranteed to 1 ppm/°C, like a MAX671 or an AD2710LD, to obtain true 16-bit absolute accuracy over the 7.5°C range from 25 to 32.5°C. Between these temperatures, the output of the reference changes no more than 75 μV—an amount equivalent to less than ½ LSB on a 16-bit, 10V FS converter.

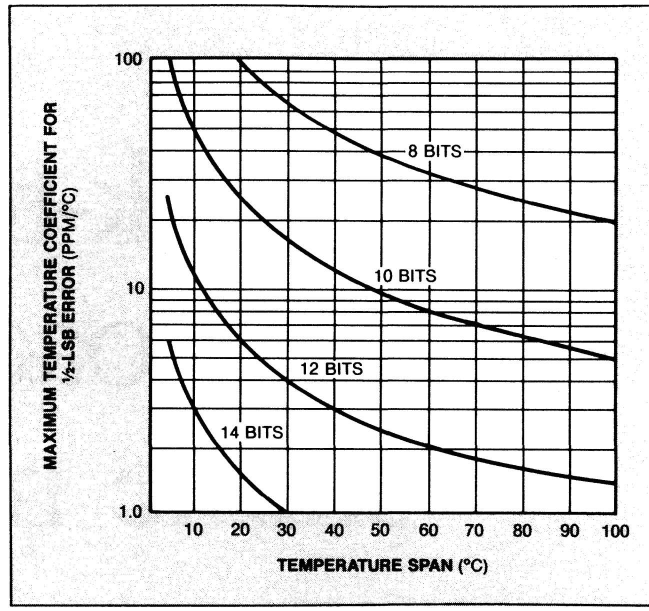


Fig 1—Increasing the resolution *of an A/D or D/A converter decreases the temperature range over which it delivers absolute accuracy comparable to its resolution.*

Most voltage references on the market produce a single-ended output between V_{OUT} and GND, as in the AD2700, REF01, AD580, AD581, and AD584. In these types of devices, I·R voltage drops can cause errors that can spoil the accuracy of the output voltage. The reason is that the load current must pass through the V_{OUT} pin and the quiescent supply current must pass through the GND pin **(Fig 2)**. If the output happens to be sinking current—for example, if the load is connected between V_{OUT} and V^+ as in **Fig 3**—then the load current also returns through the GND pin. In most cases, however, the load is connected between V_{OUT} and GND; the load current flows into the reference from V^+ and out to the load through the V_{OUT} pin. In these cases, the I·R drop on GND won't disturb the output voltage, but the I·R drop on the V_{OUT} pin will always remain.

You can minimize single-ended output errors by using a device whose package incorporates internal Kelvin connections (that is, separate force and sense lines) to connect the die to the package pins **(Fig 2)**. Running both force and sense bond wires internally from the chip to the output pin places the "force" bonding wire's resistance inside the output amplifier's feedback loop. This technique in effect eliminates connection resistance except for that of the pin itself and that of the wiring or metal trace between the output and the load. With many references operating at full output current, even if you connect the load directly to the pin, the voltage drop in the pin resistance itself is large enough to equal the initial accuracy spec.

Small errors add up

For example, the AD2710LD has an initial accuracy of 10.000V ±1 mV max. The device is enclosed in a 14-lead ceramic sidebrazed DIP that can have a pin resistance of 0.05Ω (**Fig 2**). If your circuit draws the full output current of 10 mA, the resulting voltage drop in series with the load will be 0.5 mV—half the initial accuracy spec. This drop has the effect of lowering the output voltage to 9.9995V (assuming that the factory set it to exactly 10.0000V with no load).

If you connect the load to V^+ as in **Fig 3**, the error will be twice as great, because the GND I·R drop is additive. The result is a further decrease in load voltage, to 9.9990V. In fact, the output-lead resistance is the dominant contribution to the load-regulation spec of 50 μV/mA, which is equivalent to 0.05Ω. The I·R-induced errors can expand into several millivolts if there is any length of wiring or pc-board trace that measures a few tenths of an ohm.

For constant-current loads, it's possible to simply adjust out the errors caused by I·R drops. However, the TC of the reference may be unacceptable because the TC of the pin resistance is too high. Gold has a TC of 4000 ppm/°C, or 0.4%/°C. With this TC, the 0.05Ω resistance in the above example would increase by 40% between 25 and 100°C. At 100°C, the resistance is 0.07Ω and produces a 0.7-mV error at a load current of 10 mA.

Sense at the load

A voltage reference can easily eliminate all the problems associated with I·R drops if it uses Kelvin outputs with separate force and sense pins joined at the load. The sense pin carries only a small, constant current, such as that which flows in the gain-determining feedback resistors. The force pin carries all of the variable load-dependent current. You close the feedback loop at the load by connecting the force and sense pins there; that is, you place the wiring and pin resistances inside the feedback loop.

Because of limitations on the number of package pins, some references provide only one GND pin and offer Kelvin connections only on the output. This setup is adequate if the output sources current but does not sink it, as is true in the majority of applications. The MAX670 and MAX671 contain full Kelvin connections

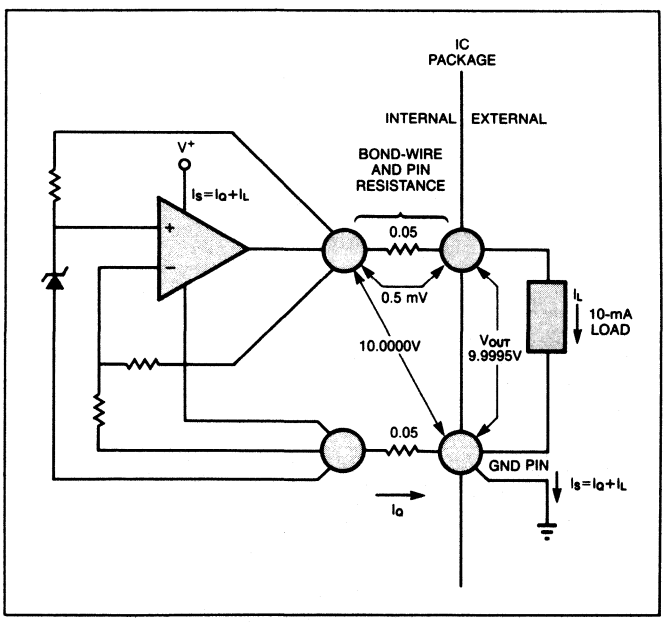


*Fig 2—**When this reference supplies current** to a load, voltage drops inside the package are inside the feedback loop and have little effect on accuracy. The voltage drop across the package's output pin can be significant, however.*

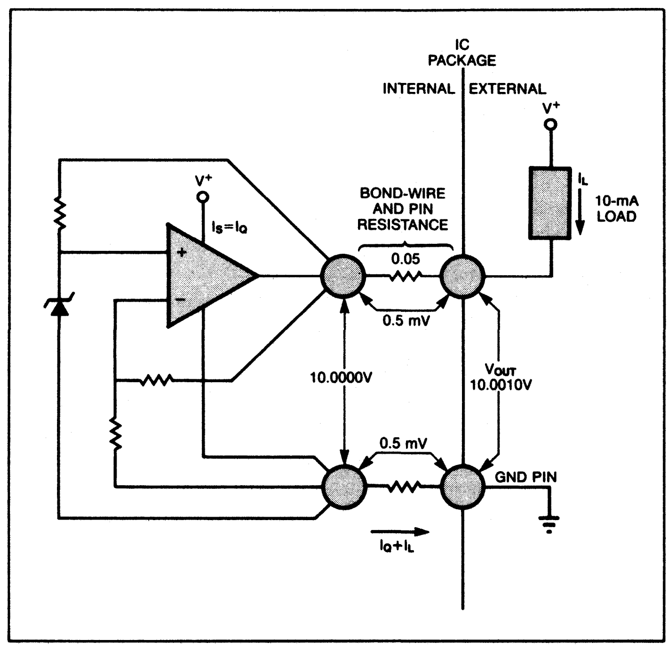


*Fig 3—**If a reference sinks current** into its output from the positive supply, both the voltage drop across the package's ground pin and the drop across its output pin will affect accuracy.*

Don't take the initial accuracy too lightly, thinking that you'll be able to adjust it.

on both output and GND (**Fig 4a**). These devices can source or sink 10 mA.

Why are Kelvin output connections so important? First of all, they make the voltage reference easier to use: You can preserve accuracy without providing extra-wide printed-circuit traces or limiting the load current or the length of the wire that carries it. (You do, however, need to consider the possibility of loop instability caused by the inductance of the load-current-carrying wires and capacitive loading of the feedback network.)

Fifty milliohms *can* be a big deal

D/A and A/D converters with 12- to 16-bit resolution often require separate voltage references. In the case of a 16-bit converter, if the full-scale input equals 5V, 1 LSB has a value of 76.25 μV. As good as the AD2700 is, with a load-regulation spec of 50 μV/mA, it takes only a 1.5-mA change in the output current to cause a 1-LSB error. Such a 1.5-mA change means that, if you were to use a reference such as the REF01, with a load-regulation spec of only 1 mV/mA, the result would be a 20-LSB error. In a 12-bit, 10V-FS converter, the same 1.5-mA reference-current change causes the REF01's output voltage to change by more than ½ LSB. Fortunately, except for transients that occur when the converter's code changes, the reference input current of an ADC or DAC is normally constant.

Sometimes, though, you must switch the A/D converter between a unipolar 0 to 10V range and a bipolar −10 to +10V range. You can do so by using relays to switch the converter's bipolar-offset-resistor input to GND or to the voltage-reference output. This arrangement causes the load on the reference to vary by 1 mA. The MAX670's Kelvin outputs alleviate concern over output-voltage changes caused by such output-current changes.

Buffering—analgesic for pain of high current

The MAX670 and the MAX671 are unusual in the way that their Kelvin sense lines are further divided between two separate pins (**Fig 4a**). This arrangement allows you to add an output buffer transistor or amplifier for higher current. It also allows you to place the added components within the reference feedback loop and thus maintain the specified performance at the load (**Fig 4b**). For example, an LH0101, together with a MAX670, can supply 10V at 2A with a 3-ppm/°C TC. In this way, the MAX670 can serve as an ultrastable, low-noise power-supply regulator with an output current ranging from hundreds of milliamps to a few amps, depending on the external buffer components.

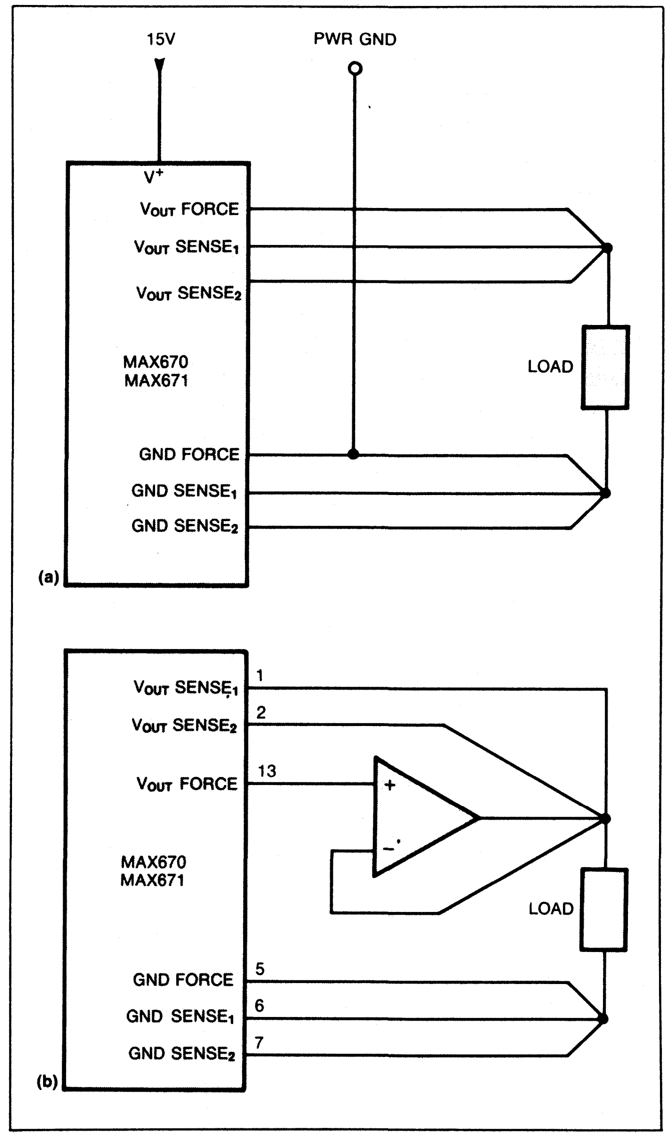


***Fig 4—By providing a pair of sense terminals** for both the output and ground signals (**a**), this reference can compensate for voltage drops outside as well as inside the package (**b**). The configuration also allows you to enclose a high-current output buffer within the feedback loop.*

You can use an amplifier to buffer references without Kelvin connections, but the voltage at the load is subject to added errors such as offset, drift over temperature, output-impedance-induced voltage drops, and voltage variations caused by line regulation. If you know the load current to within ~20%, you can supply high current regardless of the type of reference, even if

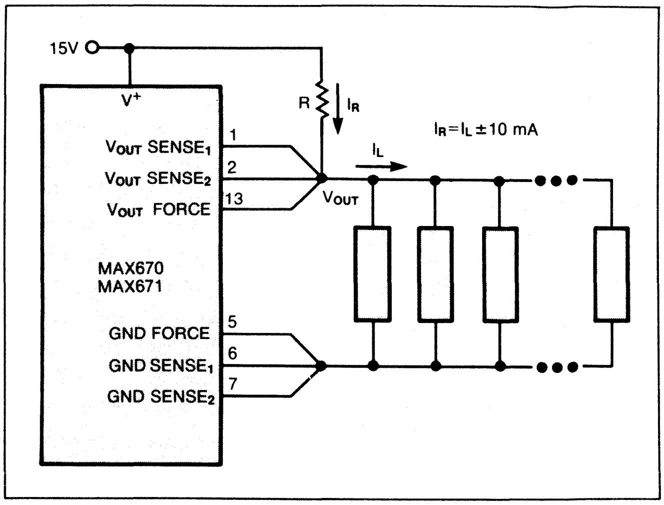


***Fig 5—A pullup resistor** acts as a poor man's output buffer by delivering most of the load current. Even though it delivers a small fraction of the load current, the reference still controls the output voltage.*

it is one with a single-ended output (**Fig 5**). In such a special case, you can use a pullup resistor to supply the nominal load current from V^+ to V_{OUT}. The reference output then only needs to sink or source the error current—the difference between the actual load current and that supplied by the resistor. Most IC-op-amp outputs supply at least ±10 mA. Ideally, if the pullup current exactly equals the load current to ground, the output current from the voltage reference will be zero. When using references like the REF01, which can source current but cannot sink it, you must guarantee that the current in the pullup resistor is less than the load current, so that the reference always supplies some current. The REF01 supplies 10 mA to ground, so you should select the pullup resistor to supply 5 mA less than the load current. That way, the REF01 will nominally supply 5 mA, a value in the middle of its range.

This technique is prevalent in ATE, where one reference supplies the reference input to perhaps dozens of D/A converters, which set the voltage or current of the pin drivers that supply signals to the device under test. A similar situation arises in drift testing large numbers of D/A converters in a temperature chamber; a reference outside the chamber drives the reference input of all of the converters.

All in all, there are three advantages to using a pullup resistor to boost a reference's output-drive capability: Adding a single passive component is simple and cheap; you preserve the accuracy and TC performance of the reference without resorting to Kelvin connections; and you don't need extra supply current (as you would if you used a buffer).

Why not design your own?

Voltage references seem like simple circuits, so you might be tempted to design your own with discrete components, but you should consider the tradeoffs carefully. To make a bandgap reference like the REF01, for instance, you need two transistors carrying equal currents with an 8:1 current-density ratio. In other words, one transistor must have 8× the emitter area of the other. Matched pairs that have this area ratio are not commonly available, but you could use a pair of identical devices and set the current ratio with resistors, except that the TC of the resistors must match as do the TCs of R_1 and R_2 in **Fig 6**. In addition, you still have to amplify the 1.2V bandgap voltage, something that requires an op amp with matching gain resistors (R_5 and R_6 in **Fig 6**).

If you want to construct a reference similar to the AD2700 (**Fig 2**), you can do so with a 1N829 5-ppm/°C zener diode and an op amp, but again don't forget the task's nontrivial nature. First of all, the diode's several-dollar price tag is a significant expense. And, in discrete form, the best temperature-compensated diodes have TC specs higher than the AD2700 spec. Assuming you can accept 5 ppm/°C, however, you'll need a low-

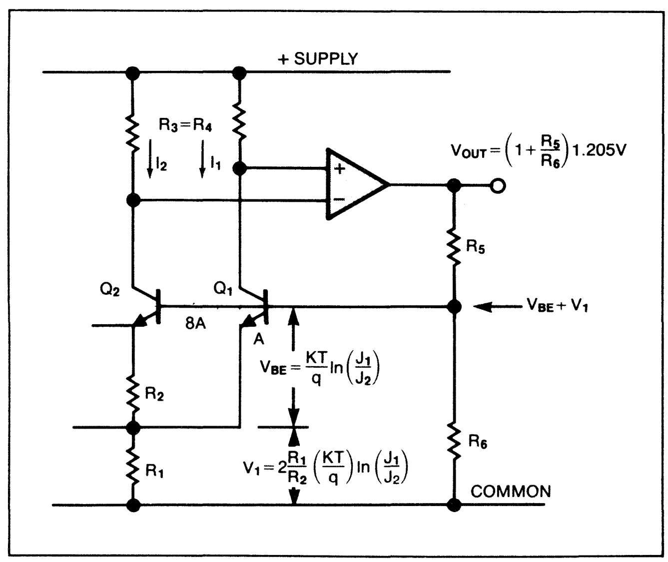
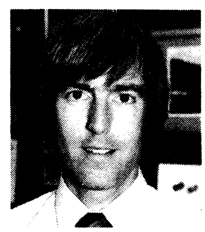


***Fig 6—A bandgap voltage reference** generates the sum, $V_{BE}+V_1$, in which the two voltages have equal and opposite temperature coefficients. The amplifier then raises the sum to a more convenient voltage level.*

drift op amp. The spec of 5 ppm/°C translates into 50 μV/°C, and therefore you'll need an op amp like the OP07, with an offset voltage drift significantly smaller than 50 μV/°C. The OP07's 2.5-μV/°C drift, when multiplied by the required gain of 1.59, contributes 0.4 ppm/°C (10V/6.3V·0.25 ppm/°C=0.40 ppm/°C). You can reduce this drift further by substituting a MAX400 op amp: It has a 0.3-μV/°C maximum offset-voltage drift spec, which translates to only 0.05 ppm/°C. Assuming that you use a thin-film resistor network, you should allow a 0.5-ppm/°C tracking TCR. You should also be aware of thermocouple effects of as much as several μV/°C, a result of interconnections between different metals. The thermocouples' sensing and reference junctions are at slightly different temperatures because of gradients across the board. Finally, if you add up the cost of the components and the time to build, test, and calibrate the circuit, you can easily appreciate the value of purchasing a complete, tested, and guaranteed precision voltage reference. **EDN**

Author's biography



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Five volts powers negative 2W regulator

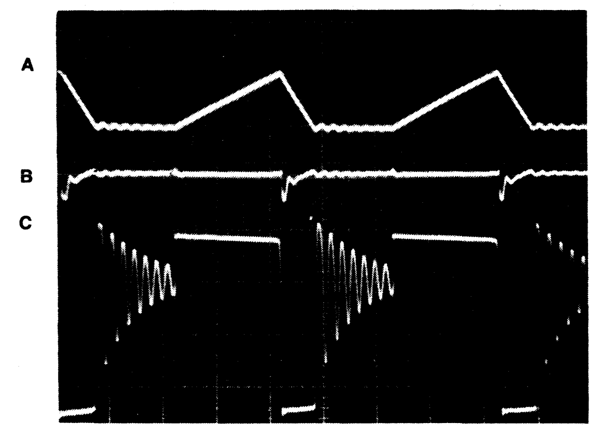
Leonard H Sherman
Maxim Integrated Products, Sunnyvale, CA

Using a 5V supply, the simple circuit of **Fig 1** generates a regulated, negative-voltage power source capable of delivering several hundred milliamperes. The circuit produces a −5 or −12V regulated output depending on whether IC_1 is a MAX635 or a MAX636. The circuit's output power is approximately 2W.

IC_1's L_X output (pin 5) drives the gate of Q_1 via the inverter, IC_2. (Form the inverter by connecting six inverters in parallel. Using the CD4069C, for example, connect the 1, 3, 5, 9, 11, and 13 inputs together, and the 2, 4, 6, 8, 10, and 12 outputs together.) Note that connecting the negative supply terminal of the inverter (pin 7) to the negative output voltage supplies maximum gate drive to Q_1, which improves efficiency by minimizing the MOSFET's on-resistance.

During operation, the inductor current (trace A in **Fig 2**) rises to a peak of 2A, which is much higher than the output current and typical of flyback-converter circuits. The inductor must handle these current peaks without saturating. Output ripple and noise are about 100 mV p-p (trace B). Trace C (node C in **Fig 1**) reveals that the inductor voltage begins near 5V when Q_1 is on and "flies back" to $-V_{OUT}$ when Q_1 turns off. Stray capacitance causes the inductor waveform to ring when Q_1 turns off, forcing the inductor current to zero. Only about 10 mV of this oscillation appears at the output.

Table 1 lists circuit performance for several operating conditions. Lowering the inductor value from 27 to 18 μH, for example, boosts the output-current capability by 20 to 40%, but sacrifices some conversion efficiency. Reducing the inductance creates higher peak currents that increase output power, but increase energy losses as well. Q_1 needs a heat sink when operating with the 18-μH inductor. **EDN**



TRACE	HORIZONTAL	VERTICAL
A	5 μSEC/DIV	2A/DIV
B	5 μSEC/DIV	200 mV/DIV
C	5 μSEC/DIV	5V/DIV

Fig 2—These Fig 1 waveforms show *inductor current (trace A), output voltage (trace B), and flyback voltage across the inductor (trace C).*

TABLE 1—PERFORMANCE CHARACTERISTICS

V_{IN}	$-V_{OUT}$	I_{OUT}	EFFICIENCY	IC_1	L_1
5V	−5V	400 mA	70%	MAX635	27 μH
5V	−5V	500 mA	64%	MAX635	18 μH
5V	−12V	150 mA	75%	MAX636	27 μH
5V	−12V	200 mA	70%	MAX636	18 μH

NOTES:
18-μH COIL=CADDELL-BURNS'S (MINEOLA, NY) MODEL 6860-04.
27-μH COIL=CADDELL-BURNS'S MODEL 6860-06.

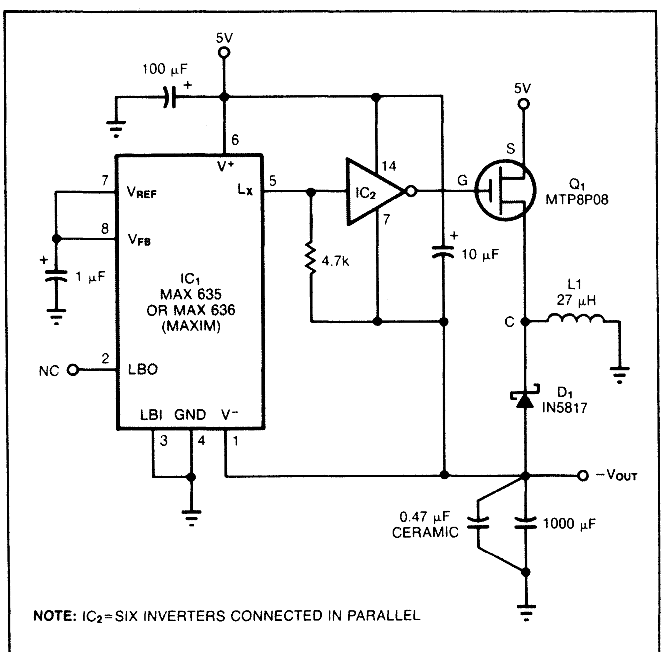


Fig 1—This negative regulator *supplies −5V when using a MAX635 or −12V when using a MAX636.*

Transistor powers low-dropout regulator

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The monolithic regulator chip in **Fig 1,** combined with an external pnp transistor, forms a very-low-dropout regulator. The composite regulator can supply several hundred milliamps at 5V from an input as low as 5.3V. Such low-dropout performance suits battery-powered applications, because it extends the useful life of batteries having sloping discharge curves, such as sealed lead-acid and lithium batteries.

The monolithic regulator derives its supply current from the base circuit of the external pnp transistor. The feedback-resistor ratio sets the output voltage:

$$V_{OUT}=1.3V\times(R_1+R_2)/R_1.$$

If the output-voltage feedback to the chip's V_{SET} input is below the bandgap-reference voltage (1.3V), the supply current into V_{IN} (the pnp transistor's base current) increases. The transistor multiplies this base current by β and delivers it to the load. The circuit's quiescent current is a function of the transistor's β and load current.

When there's no load, the quiescent current is typically 10 μA. For larger load currents, the quiescent current is simply the load current divided by the transistor's β. The regulator chip can sink 40 mA max. When you enable the chip's shut-down input, the circuit consumes 6 μA typ. R_4 supplies current to the chip under no-load conditions.

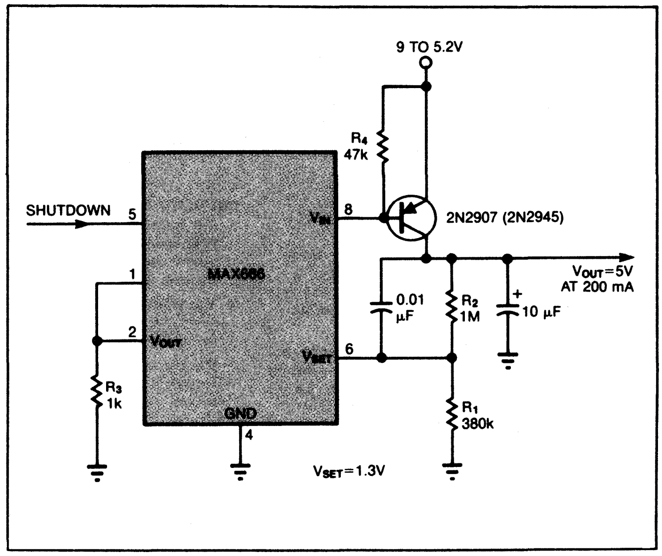

***Fig 1—A monolithic regulator chip driving a dummy load** sets the base current of an external, series-pass pnp transistor; the result is a very-low-dropout regulator for batteries whose output voltage droops under load.*

R_3 can limit the transistor's base current. The chip's V_{OUT} pin will try to raise its voltage level to that of the V_{IN} pin when the output voltage of the chip is low. Reducing R_3 has the effect of supplying larger base currents to the external transistor.

You can substitute a 2N2945 for the 2N2907 shown in **Fig 1.** With this substitution, the circuit will supply a 5V, 100-mA max output from a 5.1V input. **EDN**

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