analog
dialogue

A forum for the exchange of circuit technology: Analog and Digital, Monolithic and Discrete

1V   200µS

1.003 ON FLUKE 931
1.000 ON 536
0.972 ON D.V.M.

1V   200µS

1.003 ON FLUKE 931
1.000 ON 536
0.820 ON D.V.M.

1V   200µS

1.003 ON FLUKE 931
1.000 ON 536
0.986 ON D.V.M.

MONOLITHIC TRUE-RMS-TO-DC CONVERTER (Page 3)

1V   200µS

1.001 ON FLUKE 931
1.000 ON 536
0.972 ON D.V.M.

1V   2mS

1.008 ON FLUKE 931
1.000 ON 536
0.968 ON D.V.M.

500mV   500µS

0.573 ON FLUKE 931
0.570 ON 536
0.598 ON D.V.M.

1V   500µS
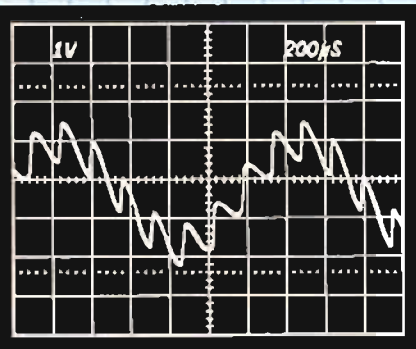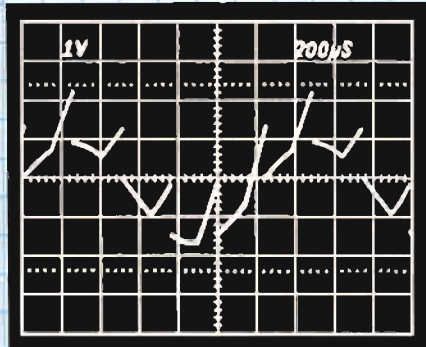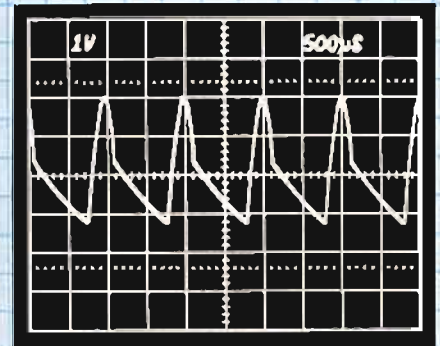
1.004 ON FLUKE 931
1.000 ON 536
0.918 ON D.V.M.

1V   2mS
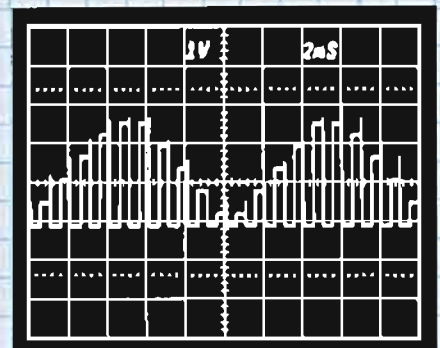
1.005 ON FLUKE 931
1.000 ON 536
0.957 ON D.V.M.

**ANALOG
DEVICES**

Editor's Notes

The youngest among our readers were born during the era of the transistor, and their consiousness probably embraces the existence of the integrated circuit. They grew up with color television and the miniature transistor portable radio, entered their teens in the era of the handheld calculator and the digital watch, and they probably view the burgeoning development of sophisticated computing tools and electronic games as a not-very-remarkable incremental evolution.

It would be hard to imagine a time when such everyday sounds as the roar of jet engines and the whirring of helicopters, such sights as radomes and offshore drilling rigs could have existed only in the imagination, if at all.

And yet, such a time did exist, as somewhat older readers can attest. But most of those older readers grew up in an era when radio and airplanes were everyday phenomena, television an intriguing new development, the vacuum tube a miracle of sorts. It was hard to imagine a time when mass spectrometers and gas chromatographs would be production tools, scanning electron microscopes an everyday aid to Q.C., and computers simply components to be engineered into instruments and other little black boxes.

No doubt, those readers, in turn, were told by their parents that a time existed when automobiles had to give the right-of-way to horses, when airplanes were a rarity, when transmission of voices, or pictures, or data (what's that?) through the ether (if it exists) using electronics (what's an electron?) were in the realm of the supernatural, if anywhere.

Their grandparents, in turn, could tell of times when the only means of rapid communication over distance was through the use of carrier pigeons; *their* grandparents could tell of times when a trip from Boston to San Francisco (if it existed) was a long, hazardous journey, with the probability of completion during one's lifetime distinctly less than 1.0.

Having straddled a good part of this recent era of technological progress, we have been witness to phenomena that, in earlier times, would have been considered miraculous. Yet, we are, in our workaday life, near the leading edge of the growth of a technology driven by human curiosity and inventiveness in a social environment that encourages technological progress by providing tangible rewards.

A perhaps relevant question is, how long can the technological progress (in which we have an increasingly important stake) continue? The best answer is that it *must* continue, whether — if you favor it — as a natural extension of human capability, or if only — in the pessimistic view of some — that technology is increasingly necessary to get us out of the fixes that technology gets us into, in a series of escalating crises.

And indeed, humankind's ability — manifested in part through technology — to cope with disease, starvation, adverse climate, and wild beasts (if not with our own warlike nature) has helped to bring some 4×10^9 living humans into the world. Some further unpredictable technological "miracles" would appear to be required to keep these billions alive, improve the quality of their life, and perhaps help to find a solution to the delicate problem of keeping the world's population in balance by other than catastrophic or morally repugnant means.

We are thus convinced of the continued need for technological progress. As noted earlier, though, to flourish, technology requires a climate that provides an inclination towards research and invention, freedom in its exercise, and tangible rewards for success. Such a climate still exists in the West, but are there clouds on the distant horizon?

Dan Sheingold

THE AUTHORS

Our authors in this issue are all "repeaters;" their photos and brief biographies are to be found in recent issues. Here are a few words about each, just in case you can't lay your hands on a copy of one of those back numbers.

A. Paul Brokaw (page 10) is Director of Product Planning at Analog Devices Semiconductor. A hardened circuit designer and inventor, with few remaining illusions, he is well qualified to write about the realities of active-circuit interfacing (see also DIALOGUE 9-1).

Lew Counts (page 3) is Staff Engineer at Analog Devices Semiconductor. Among his earlier successful products are the 433 multifunction multiplier-divider and the prize-winning 440 rms module. He contributed mightily to our *Nonlinear Circuits Handbook* (see also DIALOGUE 9-3).

Bob Craven (page 6) is Staff Design Engineer at our Hybrid Circuits facility. Bob has a number of winning designs to his credit, including the "QZ" converter series and the AD562 12-bit DAC. The AD572 looks like another sure winner (see also DIALOGUE 8-2).

Barrie Gilbert (page 3) is a consultant to our Semiconductor division. His numerous designs and publications include the basic translinear circuit concept, and its application to such IC's as the AD534 multiplier-divider and the AD536. He also fathered the AD537 monolithic V/f converter (see also 9-1).

Dave Kress (page 3) is Product Marketing Specialist at Analog Devices Semiconductor. An experienced design and applications engineer, he has been responsible for the integrity and authority of many of our IC data sheets and is well-known on the seminar circuit (see also DIALOGUE 10-1).

MONOLITHIC IC RMS-TO-DC CONVERTER
AD536K Laser-Trimmed to 0.2% Max Error; Needs No User Adjustment
Operates on Single or Dual Supply; dB and Linear Outputs Provided

by Lew Counts, Barrie Gilbert, and Dave Kress

The AD536* is the industry's first monolithic rms-to-dc converter. It permits the true-rms value of complex wideband ac and dc signals to be measured with errors less than 0.2% of reading ±2mV without external trim circuits. Its bandwidth, for signals greater than 0.1V rms, is 20kHz, for the above accuracy, and 100kHz for –3dB.

Computing the true rms value of a waveform gives an exact measure of the power in the signal, regardless of the wave shape. (Rectified average, which has been used extensively in multimeter instruments, is an accurate measure only for sine waves, but shows errors of 10% or greater for many other waveforms.) The photograph shows three types of waveform for

readily handle signals with crest factors as great as 10; at crest factor of 6 (pulse duty-cycle of 1:36), error is increased by only 1% of reading.

In addition to the linear output, a dB output, having a 60dB dynamic range, is available; it is especially useful in audio applications, and in acoustic noise measurements. Using an externally supplied reference current, the user can set the 0dB level at any convenient value from 0.1V to 2.0V.

The 536 is a low-power device, drawing only 1mA at any supply voltage in its rated span of 5V to 36V (±3V to ±18V, with bipolar supplies); this is a useful feature for battery powered applications, and – with the small size – for portable instruments.

The AD536 is packaged in a 14-pin hermetically sealed dual in-line ceramic DIP. It is available in two 0° to 70° versions, the AD536J, with maximum pretrimmed error of 0.5% ±5mV, and the AD536K, with maximum error of 0.2% ±2mV. Prices are $15 and $26.50 (1-24), and $9.95 and $18.50 in 100's.

The accuracy and low price of the AD536 result directly from a unique, automatic laser-trimming process in which all AD536
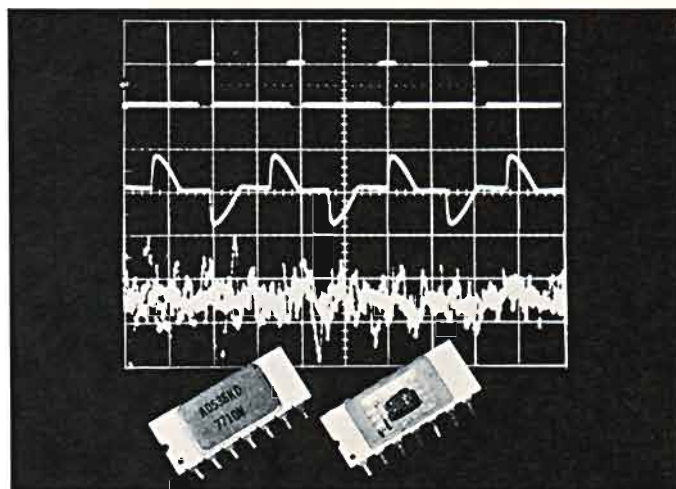
chips are trimmed, *at the wafer level,*[1] for input-and-output offsets, positive-and-negative waveform symmetry, and full-scale accuracy. As a result, no external trimming is required (unless even-better performance — by up to a factor of two — is desired).

As Figure 1 shows, the only external component needed for rms measurement is an averaging capacitor, to set the averaging period, and thus to establish the low-frequency accuracy and ripple level, as well as the response speed and settling time. Use of the internal buffer amplifier and external passive elements to add one or two poles of active filtering can improve settling times at least tenfold without increased ripple. Information for filter design can be found on the data sheet and in *Dialogue* 9-3 (page 21).
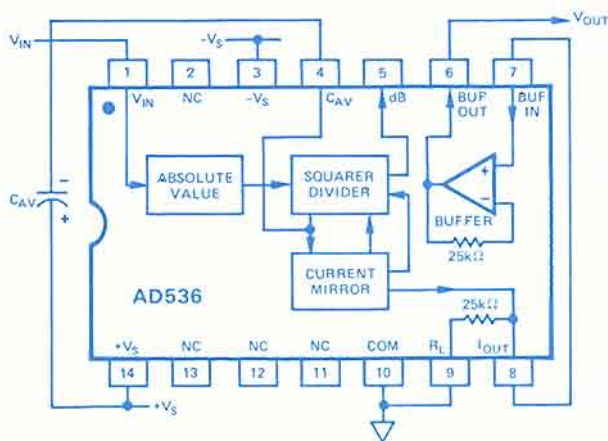
Figure 1. Standard connection for rms-to-dc conversion.

*For complete information on the AD536, use the reply card.
[1] See "Laser-Trimming IC's on the Wafer", DIALOGUE 9-3, 1975.

HOW IT WORKS[2]

The AD536 computes the root-mean-square relationship:

$$V_{rms} = \sqrt{\overline{V_{IN}^2}} \qquad (1)$$

where V_{rms} is a voltage, and $\overline{V_{IN}^2}$ is the average of the squared instantaneous values of the input voltage over a nominal period.

It does so by performing the implicit computation,

$$V_{rms} = \overline{\frac{V_{IN}^2}{V_{rms}}} \cong \left[\frac{\overline{V_{IN}^2}}{V_{rms}}\right] \qquad (2)$$

The approximation indicated in (2) can be considered exact if the averaging time is sufficiently long for V_{rms} to be essentially constant for the range of periods that is of interest.

Figure 2 is a simplified schematic of the AD536. In the AD536, the computations indicated in (2) are performed with currents. The four basic sections of the device are an input absolute-value voltage-to-current converter, a one-quadrant squarer-divider core, a current-mirror, and buffer-follower.
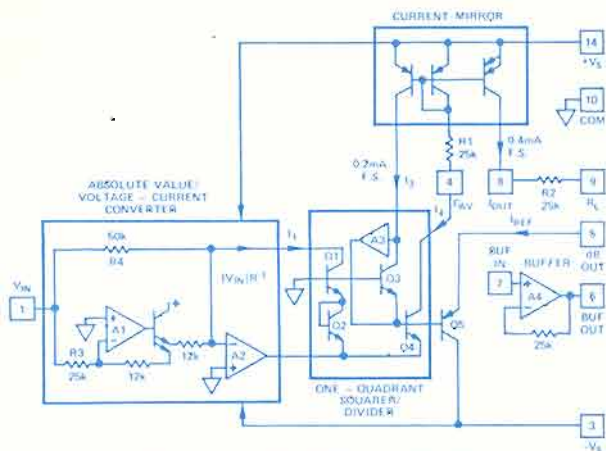

Figure 2. Simplified schematic.

The input stage takes the absolute value of the input signal (i.e., *rectifies* it) and converts it to a current, I_1, which is applied to the squaring input of the squarer-divider. The output current, I_4, drives the current mirror through a low-pass filter, formed by R1 and C_{AV}. If $R_1 C_{AV}$ is a long time, compared to the longest period of the input signal, the current entering the current-mirror is $\overline{I_4}$. An identical current, I_3, is returned to the division input, to compute the quantity, $I_4 = I_1^2/I_3$. Since I_3 is equal to the average value of I_4, it is also equal to the desired value of output current.

The current mirror produces another identical current, I_{OUT}, which is thus a current corresponding to the rms value of input voltage. It can be either used as a current or converted to a voltage by R2, which can then be unloaded by the buffer follower.

The dB output is derived from the emitter of Q3, since the voltage at this point is proportional to $-\log V_{IN}$. Emitter follower, Q5, buffers and level-shifts this voltage, so that the dB output voltage is zero when the external current, I_{REF}, applied to the emitter of Q5, is approximately equal to I_3.

[2] The reader who is unfamiliar with rms circuits will find much useful information about rms measurements (and other nonlinear devices, such as multipliers, dividers, logs, etc.) in the 536-page *Nonlinear Circuits Handbook*, available from Analog Devices, Inc., P.O. Box 796, Norwood MA 02062, for only $5.95 (Check or Master Charge).

APPLICATIONS

Besides its obvious applicability to a wide variety of physical measurement, such as noise and vibration, the low cost and small space required by the AD536 (due to both its small size and the fact that it's so complete that it needs very little auxiliary circuitry) permit OEM designers of instruments, apparatus, and systems to add measurement options at lower cost than ever before.

Since many audio signal-handling circuits utilize single supplies and ac coupling, the single-supply connection is one which will prove popular. Figure 3 shows a way of accomplishing operation with a single 5V supply — with the AD2026* low-cost digital panel meter — to build what must needs be the world's lowest-cost high-accuracy rms digital meter.
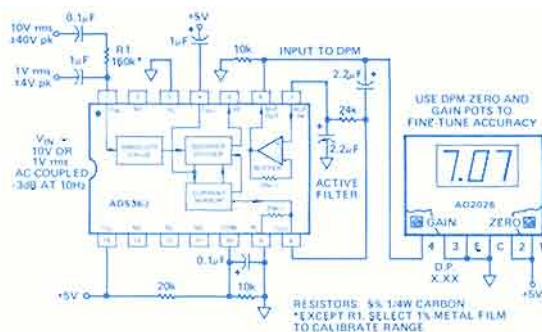
Figure 3. Low-cost true-rms DPM, using AD536 and AD2026 Panel Meter, and single 5V supply, for true rms ±1 digit ±0.5% of reading.

The input common, at pin 10, is biased off ground; it is critical that no extraneous signals be coupled into this point. The input is coupled via a capacitor; and the dc return is via the internal amplifier circuits. The input high-pass time constant is determined by the coupling capacitance and the 16.7kΩ input resistance; the values shown in the diagram are for 10Hz cutoff. Signal range is 0 to ±4V peak, 0.999V rms. The external 10kΩ load provides a path for sink current.

Decibel measurements will also be popular. Figure 4 shows a scheme for building an analog dB-meter with linear dB readout, using a single supply, such as a 9V battery. The AD580* precision 2.5V reference provides a regulated voltage to the zero-dB set potentiometer, R1, and also a current for biasing pin 10 off ground via the 2.2kΩ resistor. ▶▶▶

Figure 4. Single-supply rms-dB meter using the AD536.

*Use the reply card for information on the AD2026 and the AD580.

6-CHANNEL SCANNING DIGITAL THERMOMETER
AD2036 Linearizes J, K, or T Thermocouples, Reads °F or °C
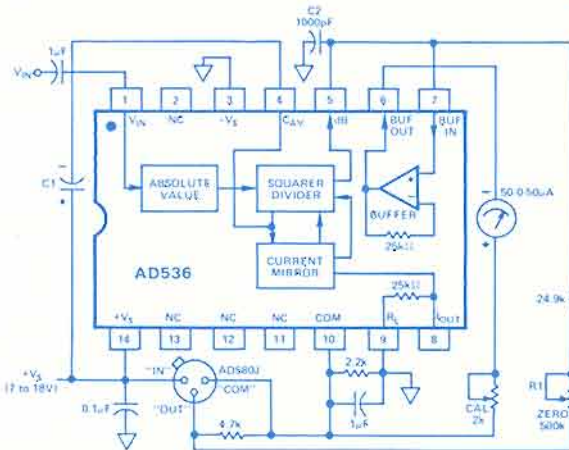## Flexible Scan Can Be Manual, Automatic, or Remote-Controlled

The AD2036* is a 3½-digit, line-powered digital panel meter, with an opto-isolated front end. It accepts inputs from up to six thermocouples (furnished by the user) and provides switching, cold-junction compensation, linearizing, conversion, 0.5" (13mm) LED display, and parallel BCD output — including channel identification. Because all these functions are combined in one versatile instrument, the cost is less than half what it would be for the user to integrate a temperature meter, having proper linearization, with a multiplexer or scanner assembly providing appropriate functions and interfacing for thermocouples. Price is $349, 1-9, shrinking quickly to $319, 10-24, and $289, 25-49.

The choices between Celsius and Fahrenheit display, and J, K, and T thermocouple compatibility, permit you to choose among 6 optional instruments, having temperature-measurement ranges of

J(Fe-Cn): –60°C to +760°C, or –76°F to +1400°F
K(Cr-Al): –60°C to +1350°C, or –76°F to +2000°F†
T(Cu-Cn): –100°C to +400°C, or –148°F to 752°F

with 1° resolution (0.1° available for limited ranges on special order) and typical accuracies to within 2°C, 1.8°C, and 1.5°C, averaged over the respective ranges.

The AD2036, the first known low-cost self-contained digital thermometer that automatically scans a number of thermocouples, is especially useful for multi-point temperature measurements for remote data-acquisition and data-logging, and for temperature monitoring in design, laboratory instrumentation, manufacturing, and quality control.

SCANNING
A particularly useful feature in data-logging applications: the AD2036 can cycle through its six channels continuously and automatically, dwelling on each channel for 3.2s (1.6s and 0.8s can be made available on special request). It can also be stimulated to perform single cycles of scanning, stopping at channel 0. Individual channels can be addressed by means of a 3-bit-word control input, for remote computer or microprocessor control. In addition, a rotary switch, shown

```
c 0 - 0 0 1 0   °C
c 1 + 0 0 2 5   °C
c 2 + 0 0 7 2   °C
c 3 + 0 0 7 5   °C
c 4 + 0 0 5 1   °C
c 5 - 0 0 4 5   °C
```

Figure 1. Brief tape segment from printer, showing channel number and temperature.

*Use the reply card for complete information on the AD2036.
†3½-digit limit
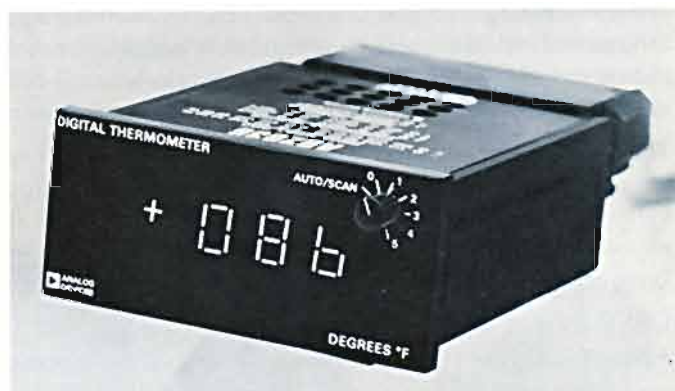
in the photograph, permits manual interruption of the scan and addressing of individual channels.

Whatever the means of selection, a digital output is available for identifying the channel that has been selected, in addition to its BCD data output. This is especially useful in applications with printers and processors. Figure 1 shows a segment of printer tape, with temperature readouts and channel identification. The application of the AD2036 with a printer will be discussed in relation to Figure 3.

HOW IT WORKS
Figure 2 is a block diagram of the AD2036. The analog front end is above the dashed isolation barrier, the digital logic below; they operate from separate windings of the power transformer. The common-mode rating (analog input to digital ground) is ±350V peak max., and the common-mode rejection is 120dB at 250V rms, for 250Ω unbalance, common-mode voltage $dV/dt < 10^6$ V/s.

The thermocouple voltages are applied via a barrier strip, differentially multiplexed, and the selected one is applied to the amplifier in series with cold-junction-compensation voltage. The signal is filtered, linearized, and converted to digital at a 5/s rate, by circuitry essentially the same as that of the AD2022
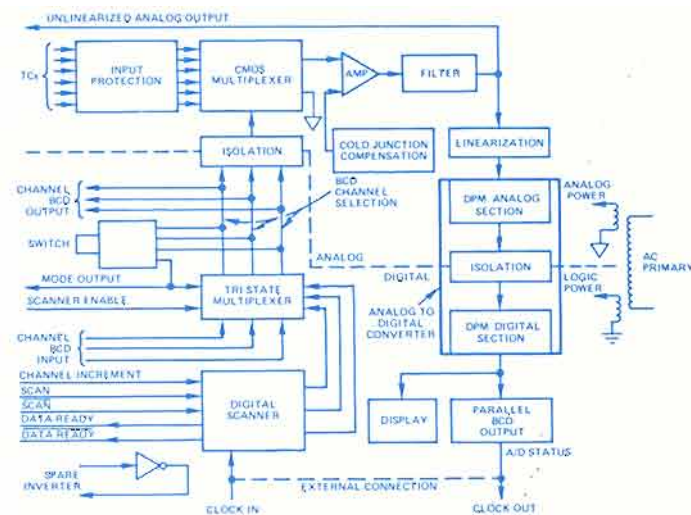

Figure 2. Block diagram of the AD2036.

Functional DPM.* (In fact, the AD2036 uses the standard Analog Devices ac-line-powered-DPM case, with the industry-standard panel cutout.) The converter drives the display and the parallel BCD output circuitry.

The tri-state multiplexer drives the analog multiplexer switches via opto-isolators, to choose the appropriate channel, except when overridden by the manual switch; the same signal that drives the analog multiplexer also provides the channel output information. When the scanner is enabled, it initiates switching between channels. When it is not enabled, the channel BCD input can choose the channel.

Though 5 conversions take place per second, 3.2 seconds are normally allowed to elapse before the Data-Ready output indicates that the data is valid. A minimum of 2 seconds are required for worst-case settling (of the filter) for a full-scale input change, as might occur when switching channels.

OPERATING WITH A PRINTER

Figure 3 shows the input and output connections for operating with a printer. A single scan may be initiated via a momentary logic "0", applied to $\overline{SCAN}$, via a grounded pushbutton or pulse source. When DATA READY goes high, the printer status, $\overline{BUSY}$, goes low. This *holds* the BCD Data output and the Channel Number until the data has been printed and $\overline{BUSY}$ goes high. When $\overline{BUSY}$ goes high, the hold is released and the channel counter is incremented. After 3.2s (in the standard unit), DATA READY again goes high; the cycle continues until data have been printed for all six channels. Each automatic or manual initiation of the scan causes the sequence to repeat. To continuously scan all six channels, $\overline{SCAN}$ is held low.
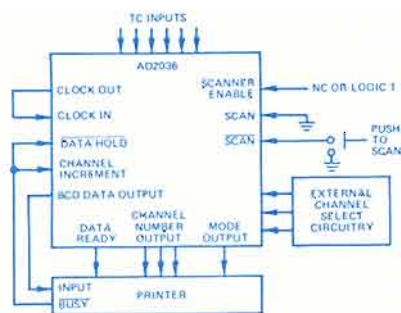

Figure 3. AD2036 with printer.

Continuous scan of a single channel is effected by selecting the channel with the switch, or by an external 3-bit word, with $\overline{SCAN}$ at logic "0".

ISOLATION AND RELIABILITY

Optoelectronic isolation is a standard feature. It prevents external digital load currents from entering the analog circuitry via common ground paths and also enables safe temperature-monitoring of equipment that cannot be isolated from the ac line. Since reliable operation is important, especially in on-line applications, great pains have been taken to design and produce a reliable instrument — starting with the use of LSI chips to minimize parts count, continuing with continual quality-assurance monitoring during assembly and test — including the use of automatic test equipment to test each DPM, both the board and the final assembly — and concluding with one week of failure-free +50°C cycled power burn-in.

*For information, use reply card. Or see *Analog Dialogue* 10-2, page 10.

The AD572* is a complete 12-bit successive-approximation analog-to-digital converter. In addition to the basic analog switch array and precision-resistance ladder network, it includes an internal clock, voltage reference, comparator, and analog buffer. Its hybrid design utilizes MSI digital and linear monolithic chips, and active laser-trimming of high-stability thin-film resistors, to combine module-like performance, flexibility, and ease of use, with IC size, price and reliability.

Performance characteristics of the AD572 include: maximum linearity error of ±0.012% of full-scale range (FSR) at 25°C, gain tempco (including the reference) below 15ppm/°C, power dissipation of 900mW, and conversion time less than 25μs. The AD572 is available in three performance grades; the "AD" and "BD" -suffixed versions are specified from –25°C to +85°C, and the "SD" from –55°C to +125°C; MIL-STD-883 processing is available.

Internal scaling resistors permit jumper-programmed analog signal-input ranges of ±2.5V, ±5V, ±10V, 0 to +5V, and 0 to +10V. The internal precision +10V reference has external drive capability. All digital I/O signals are TTL-compatible; the output is positive-true and available in both serial and parallel form.

PACKAGE AND RELIABILITY

The AD572 is packaged in a hermetically-sealed, welded, 32-pin metal DIP (1.74"x1.14"x0.235"H — 44.2x29x5.9mm), having pins on 0.9" (22.9mm) row centers. The package is sealed in a dry N₂ atmosphere, by a one-shot resistance-welding method, which generates essentially no heat rise within the package. Consequently, the accuracy of the trimmed resistors is preserved throughout the sealing operation. The metal package provides excellent electrostatic shielding, which permits error-free device operation in noisy environments typical of computer-interface applications.

For assurance of a level of reliability consistent with its outstanding performance, each device receives a stringent precap visual inspection, high-temperature preseal and postseal stabilization burn-ins, temperature cycling, acceleration, and fine- and gross-leak testing. In addition /883B-suffix devices receive further processing per MIL-STD-883, Method 5004, Class B.

CIRCUIT DETAILS

The various unipolar and bipolar analog input ranges are pin-programmed externally. If the unity-gain buffer is not required for impedance transformation, the analog voltage input can be applied directly to the appropriate span resistor.

A proprietary 12-bit analog current-switch chip and Si-Cr thin-film resistance-ladder network (packaged and available separately as the AD562 12-bit d/a converter) constitute the feedback weighing DAC. The network is actively laser-trimmed to calibrate all the bit-ratio scale factors to within ±0.0005% of

*For complete information on the AD572, use the reply card.

12-BIT HYBRID IC A/D CONVERTER
AD572S: Successive Approximations, No Missing Codes, -55° to 125°C
Complete Device Includes Laser-Trimmed Reference

by Bob Craven

FSR; this permits a guarantee of no missing codes over the respective temperature ranges specified for the various options.

TTL storage registers output both parallel and serial data in positive-true form. Parallel-data output coding is binary for unipolar ranges, and either two's-complement or offset-binary for bipolar ranges, depending on whether $\overline{BIT\,1}$ (pin 13) or its logical inverse, BIT 1 (pin 12), is used as the MSB. Parallel data becomes valid approximately 200ns before the STATUS flag returns to logic 0, permitting parallel-data transfer to be clocked directly on the 1-to-0 transition of the STATUS flag.

Serial-data coding is binary for unipolar, and offset-binary for bipolar input ranges. Bit-serial output — MSB first and LSB last — is in NRZ (non-return-to-zero) format. Serial and parallel data outputs change state on positive-going clock edges. Serial data is guaranteed valid on all negative-going clock edges, permitting serial data to be clocked directly into a receiving register on those edges.
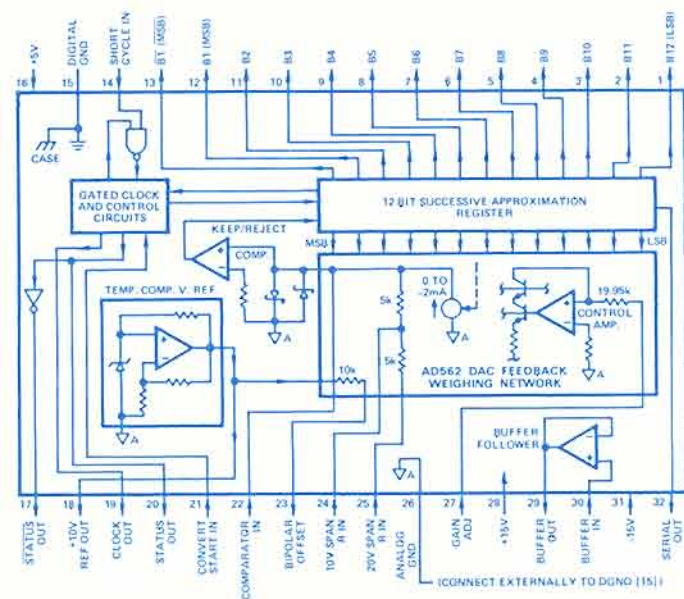

Figure 1. Functional diagram and pinout of the AD572.

A SHORT CYCLE input permits the conversion to be truncated for a tradeoff between speed and resolution. For 8, 10, 12 bits, the respective maximum conversion times are 17, 21, 25μs.

CONSTRUCTION

The AD572 substrate assembly is shown in Figure 2, with the critical analog components — the reference subassembly and the AD562 feedback element — identified. The main thick-film substrate contains a multi-layer interconnect system and thick-film resistors (used in the non-critical digital and analog biasing circuitry). The precision voltage reference comprises a separate thin-film substrate subassembly, bonded to the main

substrate. This construction technique combines the low-cost design flexibility of thick film with the high performance of thin-film and monolithic technologies to provide outstanding value at low cost ($95-"AD", $125-"BD", $250-"SD", 1–24).
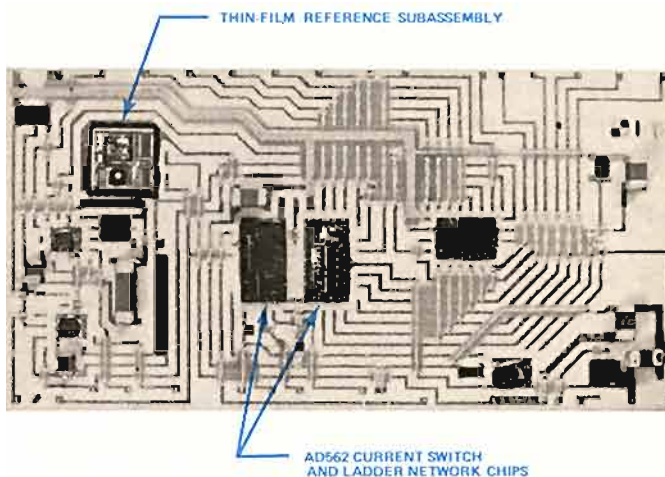

Figure 2. Photomicrograph of the AD572, showing run patterns and mounted components.

PERFORMANCE CHARACTERISTICS — AD572SD
Typical at +25°C, ±15V, +5V, unless otherwise noted

ANALOG INPUTS		
Voltage Ranges:	Unipolar	+5V, +10V
	Bipolar	±2.5V, ±5V, ±10V
Input Impedance (Direct Input)		
Ranges:	+5V, ±2.5V	2.5kΩ
	+10V, ±5V	5kΩ
	±10V	10kΩ
BUFFER AMPLIFIER (Unity Gain)		
Impedance		100MΩ, min.
Settling time to 0.01% FSR		2μs (for 20V step)
RESOLUTION		12 bits binary (0.024% FSR)
RELATIVE ACCURACY (error)		±0.012% FSR
LINEARITY TEMPCO - No Missing Codes		Guaranteed, -55°C to +125°C
GAIN TEMPCO (Internal Reference)		±15ppm/°C, -25°C to +85°C
		±25ppm/°C, -55°C to +125°C
INTERNAL REFERENCE		10V, ±5mV
Tempco		±10ppm/°C, max
CONVERSION TIME		25μs, max
CONVERSION COMMAND		Pulsed logic "1", 200ns, min Leading edge (0-to-1 transition) resets registers. Trailing edge (1-to-0) initiates conversion.
DIGITAL OUTPUTS		All data positive-true, TTL-compatible
Parallel data (13 lines)		Bit 1 (MSB) to Bit 12 (LSB), & Bit 1 (MSB)
Serial data (NRZ format)		1 line, single bit
Status		"1" during conversion, "0" otherwise
Status		"0" during conversion, "1" otherwise
Internal Clock		500kHz; used to clock serial data
POWER REQUIREMENTS		+15V ±5%dc, +25mA
		-15V ±5%dc, -20mA
		+5V ±5%dc, +50mA
OPERATING TEMPERATURE		-55°C to +125°C

THREE NEW ANALOG I/O BOARDS FOR µC'S
RTI-1201: 4-Channel Output Board for Intel SBC 80 Users
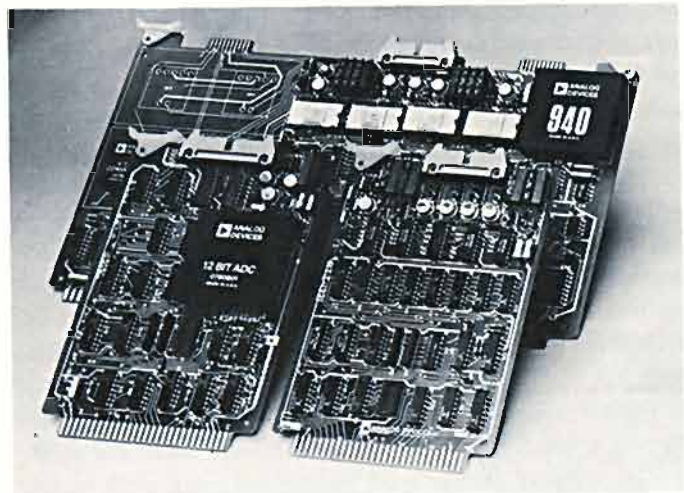## RTI-1220 & RTI-1221: Input and Output Boards for PRO-LOG µC'S

The cover of the last issue of this Journal (11-1) was graced by a photo and block diagram of the RTI-1200*, the progenitor of an extended family of microcomputer-compatible input/output subsystems. The RTI-1200 "real-time interface" is functionally, electrically, and mechanically compatible with the Intel SBC 80 microcomputer; in software, its functions are mapped on up to 16 selectable 1k blocks of computer memory-space.

The RTI-1200 contains a large number of features which make it exceptionally useful in interfacing real-world analog phenomena to a microcomputer. Such features include

- a complete up-to-32-channel data-acquisition subsystem, including multiplexer with protected inputs, (software-) programmable-gain sample-hold amplifier, 12-bit a/d converter
- two optional 12-bit d/a converters with voltage or 4—20mA current-output options, and two logic-driver outputs
- a real time pacer-clock system, a fully-active socket for a 1k PROM, memory-inhibit and 8-level interrupt capability, and many other features — integrated into the computer software system — that make efficient use of program and computing time
- "card select", to permit up to 16 RTI-1200's to share the same block of memory addresses.

POPULATION EXPLOSION
So well-received was it (and imitated), that its success has led us to develop a line of analog I/O products, designed to have a similar degree of sophistication compatibility with a range of popular microcomputers. Here, we report on a multi-channel *output* board for the SBC 80, with many of the properties of the RTI-1200, and on input and output cards for Pro-Log microcomputers. In future issues, you can expect to read about a great many more compatible subsystems, including analog I/O subsystems for Motorola and Texas Instruments microcomputers, and additional cards for use with Intel, Pro-Log, and Control Logic products.

RTI-1201 OUTPUT BOARD FOR SBC 80
The RTI-1201* is a 4-channel analog output card that is electrically and mechanically compatible with the Intel MDS and SBC 80 microcomputers, as well as the Analog Devices RTI-1200 analog I/O card. Also available are four high-power (30V, 300mA) open-collector, software-controlled logic drivers, for system control functions. Designed with the end-user in mind, the RTI-1201 is versatile, efficient, and easy to use. Figure 1 is a block diagram of the RTI-1201.

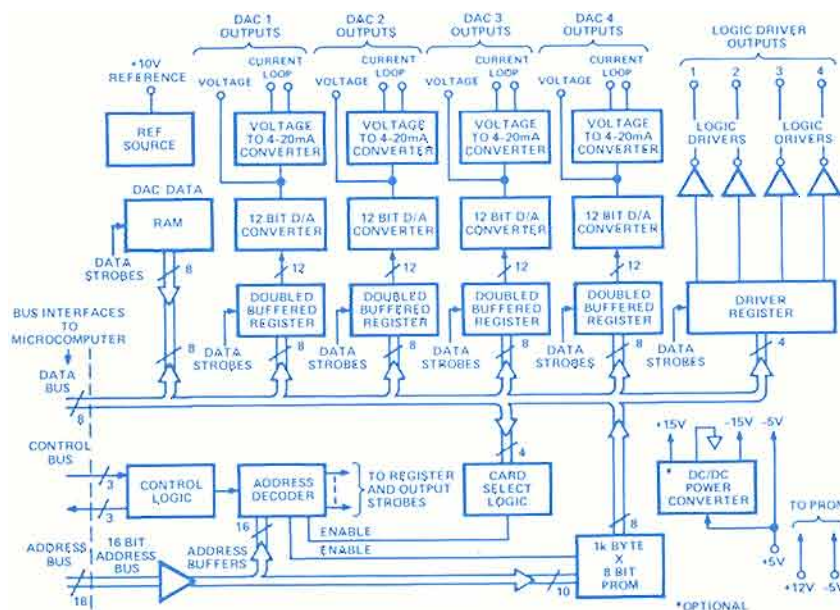It is configured for four 12-bit DAC's, software-driven via

Figure 1. RTI-1201 functional block diagram

*For technical information, use the reply card.

double-buffered registers. The user can configure the DAC's individually for a variety of output spans (±2.5V, ±5V, +5V, ±10V, +10V) and codes (binary, 2's complement, offset binary). Analog output data is loaded into each converter in either two-byte (4 MSB, 8 LSB) format for 12-bit conversion, or single-byte (8 MSB) format for 8-bit conversion, *with 12-bit accuracy.*In two-byte form, the data for a single DAC can be loaded with a single SHLD instruction. Double buffering permits simultaneous 2-byte loading of the DAC, allowing its output to move directly from one value to another, without suffering an intermediate value.

DAC-input data to the RTI-1201 can be read back, sparing scratchpad memory and software overhead. Memory mapping gives the user the full benefit of the 8080's memory-reference instructions. A flash *reset* feature allows a single-byte instruction to reset any or all of the DAC's, and jumpers permit them (and four logic drivers) to be reset on a *system reset* command.

Any or all of the four channels can be equipped with a 4–20mA current-loop converter for process-control applications. The voltage outputs have remote-sensing capability, to avoid degradation of accuracy at distant destinations by IR-drops. All channels track very well, because they share a single reference. The on-board precision +10V reference voltage is also available externally; conversely, the common reference input can be driven from an external source.

Power required is ±15V, when available; an optional dc-dc converter permits the RTI-1201 to be powered by the computer's +5V supply.

Like the RTI-1200, the RTI-1201 interfaces to the SBC 80 microcomputers as a 1k block (1024 bytes) of memory; it can occupy any one of 16 selected blocks, so that it need not interfere with already-committed address space. A memory "overlay" feature is included, so that the RTI-1201 can be used with an MDS system having a full complement (64k) of memory. The RTI-1201's addresses are offset by 1k with respect to the RTI-1200's block, allowing a pair to occupy 2k of memory space. The unique card-select feature allows up to 16 RTI-1201's to share the same block of memory, interchangeably. As with the RTI-1200, the active I/O and control functions occupy 16 bytes of the 1024; the remainder are available to a socket wired for a 1k PROM, either for functions associated with the RTI, or simply for augmentation of system memory.

Price of the RTI-1201 depends on the chosen options. Prices range from $298 to $654 (1–9).

RTI-1220 AND RTI-1221 FOR PRO-LOG μC's

The RTI-1220* and RTI-1221* are small, low-cost analog I/O interface subsystems. The RTI-1220 is a complete, 16-channel, data-acquisition subsystem on a single board. The RTI-1221 is a complete, 4-channel, analog-output subsystem on a board. Both are directly compatible with Pro-Log microcomputers, and both are easy to interface to other 8080- and 6800-based μC's.

These low-cost analog I/O boards were designed for anyone considering a small microcomputer system. Typical users might include designers of instrumentation, industrial equipment, simulators and trainers, lab equipment, data loggers, automatic test equipment, etc.

As the RTI-1220 block diagram shows (Figure 2), it is a complete data-acquisition subsystem, including protected multiplexer, adjustable-gain amplifier, sample-hold, and 8- or 12-bit successive-approximations a/d converter. Like other members of the RTI family, it uses memory-mapped interfacing and has card-select logic to permit up to 8 RTI-1220's to reside in one memory location.

It occupies 8 consecutive memory addresses (FXF8 to FXFF); its "X" location is user-selectable on the board. Multiplexer channel addresses may be incremented using a memory increment instruction; the active channel number may be read back. Data can be read in 8 (MSB's) and 12 (4 MSB + 8 LSB)-bit formats.
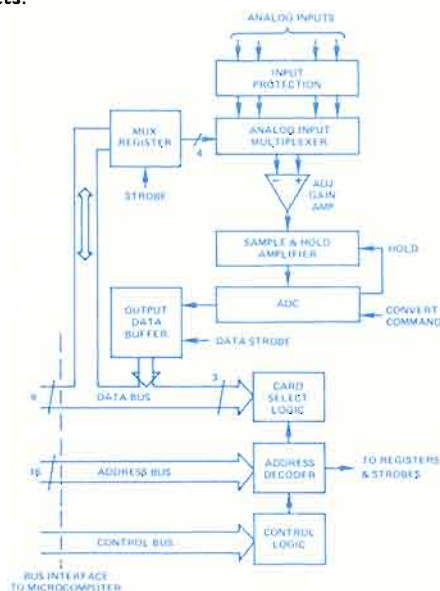

Figure 2. RTI-1220 block diagram.

The RTI-1221 (Figure 3) has four 4-quadrant multiplying DAC's and a memory map somewhat like that of the RTI-1201, occupying 16 memory addresses (but without the PROM socket), from FXF0 to FXFF. A DAC *reset* capability is included, and card-select logic permits 8 units to share a memory block.

Both units are functionally, mechanically, and electrically compatible with Pro-Log μC's. All address, control, and data-bus connections are made by simply plugging the cards into a user-wired Pro-Log card cage. Analog connections are separated "across the board" from digital. Cost for RTI-1220/RTI-1221: 8-bits, $269/$235; 12/10 bits, $330/$275 (1–9). ▶▶▶
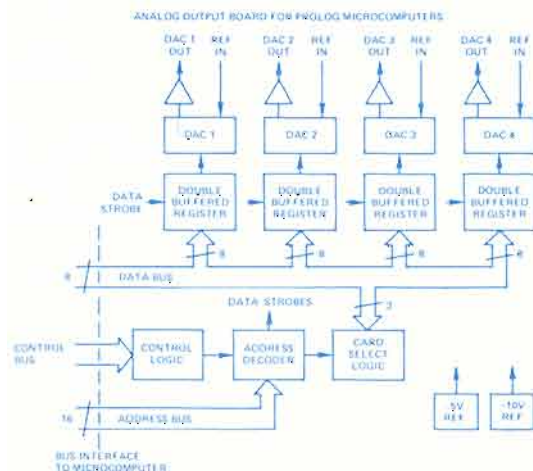

Figure 3. RTI-1221 block diagram.

ANALOG SIGNAL-HANDLING FOR HIGH SPEED AND ACCURACY
You're Paying for Accuracy in IC Converters;
Don't "Blow It" in the Analog Circuit Wiring

by A. Paul Brokaw

You've bought an IC a/d or d/a converter that's specified for 10-bit-and-better resolution and accuracy. Or, you've bought a current-ouput DAC with submicrosecond settling to 1/2 LSB. Much design effort, technological development, and process competence have been expended to solve the hardest part of your interface problem. But . . . you aren't out of the woods yet! Here are some of the issues that you will have to come to grips with to preserve speed, resolution, and accuracy:

1. If your DAC is a current-output type and you want voltage, the use of an op amp requires that you deal with the dynamic and steady-state signal-interfacing problems.

2. You will have to minimize interference introduced via common power-supply connections.

3. You will have to decide where "ground" should be and how to keep it there.

4. If "ground" is remote, you will have to couple to it without reduced accuracy or succumbing to interference.

5. If your analog signal is being converted by a successive-approximations converter, you may have to buffer the source from fast transients incidental to conversion.

To become aware of these potential problems is to have taken the first step towards solving them. Since all circuits and systems differ in important little ways, there are no "cook-book" solutions that can be blithely employed for satisfactory results in all cases. However, a little thought will go a long way towards solving them. The purpose of this Brief is to remind you of some of the things you should be thinking about.

DAC's AND OP AMPS — DYNAMIC PROBLEMS

A current-output DAC is usually connected to the summing point of an inverting op amp, and then the feedback loop is closed via the internal "span" resistor, R_F, as Figure 1 shows. The output impedance of the DAC can generally be treated as a parallel combination of resistance and capacitance. The shunt capacitance, C_O, combines with R_F to add a pole to the open-loop response, which may result in poor closed-loop response.
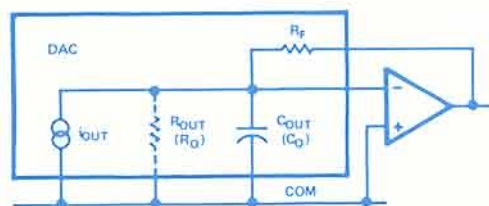
Figure 1. Equivalent circuit of current-output DAC.

Figure 2 shows how the open-loop amplitude and phase response might appear if the spurious pole due to C_O is below the undisturbed system-crossover frequency. Not only will the closed-loop bandwidth be reduced, but —more seriously— excess phase shift will be introduced. The extra phase shift reduces the system frequency stability margins and may cause ringing (and perhaps even oscillation).
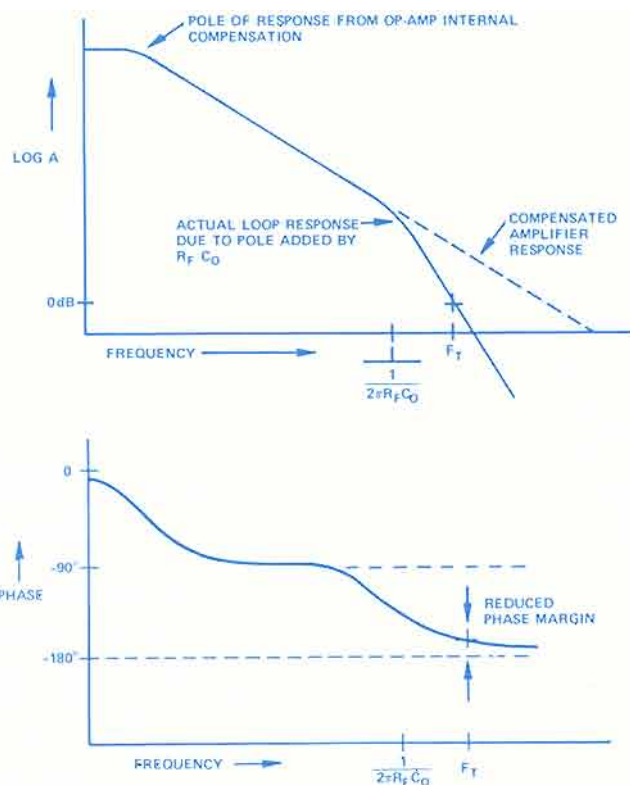
Figure 2. Amplitude and phase response of the circuit of Figure 1. The additional pole increases settling time by reducing bandwidth and increasing both overshoot and ringing.

As Figure 3a shows, the loop-stability margins can be restored by connecting a feedback capacitor, C_F, in parallel with the feedback resistor. This capacitance creates a zero in the open-loop transfer function, which can be adjusted to correct the phase margin. However, if R_{OUT} is very large (as is often the case with current-output DAC's), the large pole-zero mismatch remaining (Figure 3b) may result in slow settling.

Even with finite values of R_{OUT}, a small residual pole-zero mismatch (Figure 3c) may result in long-settling "tails"; the DAC output voltage may appear to settle quickly, but then it slowly changes —by a significant amount— to its final value, over the course of tens of microseconds, or even milliseconds.[1]

The residual mismatch will be eliminated when the DAC-output circuit and the feedback network form a frequency-compensated voltage divider, i.e., when $R_O C_O = R_F C_F$. This condition can usually be satisfied, but sometimes it requires large values of C_F. Unfortunately, C_F—which introduces an open-loop zero—also produces a closed-loop pole, which reduces the overall bandwidth and results in increased settling time.

R_F is generally fixed by the desired DAC gain; the minimum value of C_O is a property of the converter not under the system-

[1] This process is discussed in some detail, with waveforms, in the Appendix to an article, "Settling Time of Operational Amplifiers," by Bob Demrow, appearing in ANALOG DIALOGUE 4-1 (1970).
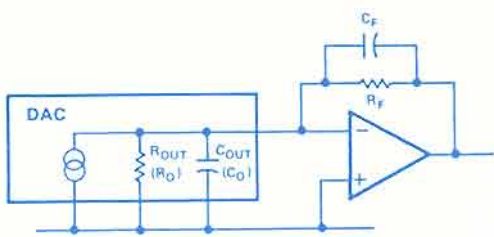
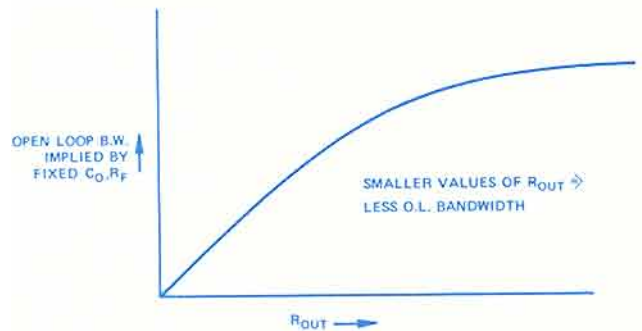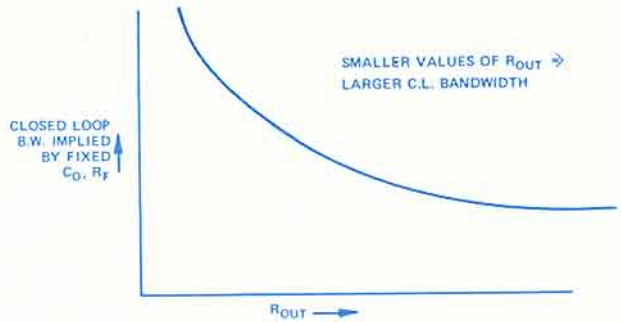Figure 3a. Improving loop stability by the use of feedback capacitance, C_F.
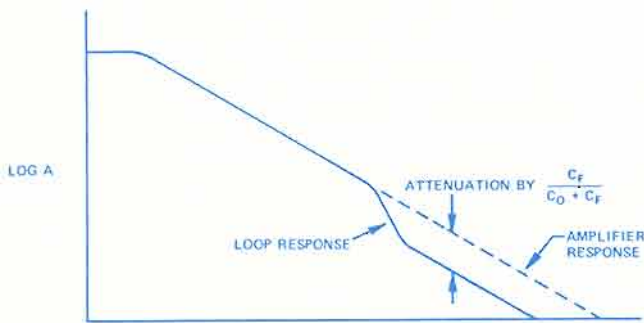

Figure 3b. Response of circuit 3a, neglecting R_{OUT}. Pole- Zero mismatch may yield poor transient response.
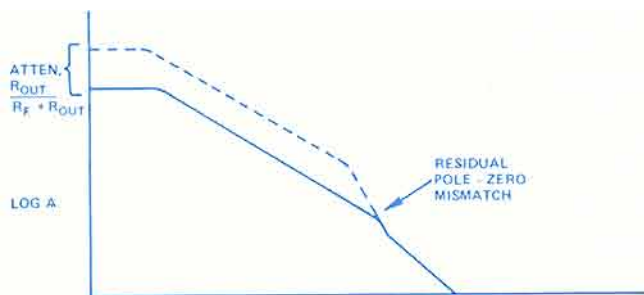

Figure 3c. Response of circuit 3a with finite R_{OUT}.

designer's control. Therefore, C_F and R_O are the only two parameters that can be manipulated (reduced). As R_O' (the effective value of R_O) is reduced by shunting the DAC output with a resistor, the required value of C_F is reduced, and the closed-loop bandwidth is increased (Figure 4). The unity-gain bandwidth of the op amp, b, limits the open-loop system bandwidth, which, in turn, limits the realization of closed-loop bandwidth. As R_O' is reduced, the *open-loop* bandwidth obtainable for a fixed op-amp bandwidth, b, is also reduced.

A compromise can be reached by adjusting R_O' to provide the same open- and closed-loop bandwidth. For a fixed C_O and R_F, the values of R_O' and C_F can be determined from:

$$R_O' C_O = R_F C_F = \frac{1 + \sqrt{1 + 8b\pi R_F C_O}}{4b\pi} \qquad (1)$$

The resistive component at the DAC output also influences the effect of the amplifier offset (V_{OS}) and noise on the overall output voltage. Both are magnified by $(1 + R_F/R_O)$.

Figure 4. Effect of varying R_{OUT} (R_O') on open-loop and closed-loop bandwidth.

DAC's AND OP AMPS - NULLING PROBLEMS

Perhaps the best way to control V_{OS} in an op amp used with a DAC is at the source—to choose an op amp with sufficiently low offset over the temperature range (such as the AD510). The next-best way is to null the op-amp's offset by the standard V_{OS} trim, taking pains to connect the pot wiper to the appropriate supply terminal *at the device*.[2] The amplifier's offset-trim adjustment should be used *only* for V_{OS} nulling; if it is used to compensate for offsets caused by the flow of bias current through the feedback resistor, as well as for offsets occurring in external circuitry, the amplifier input stage will have to be unbalanced, which will cause its V_{OS} tempco to be degraded.

If the amplifier lacks offset-adjust terminals, or if it is necessary to compensate for the additional sources of offset mentioned above in one convenient place, there are two commonly used ways of providing the trim; they are shown in Figure 5. The more-desirable approach is shown in 5a; the correction is applied to the amplifier's positive input terminal, as a voltage. Since it is effectively in series with V_{OS}, the V_{OS} correction is unaffected by changes of R_O'.

The less-effective way is to introduce a current at the summing point, as shown in 5b. If the resistances in the circuit (including R_O') are constant, there is no problem. However, if R_O' can vary, the output offset will change. If the change of R_O' is a function of the applied *digital code*, the result can be increased differential nonlinearity.

[2]The reasons for this are well-documented in the Application Note, "An IC-Amplifier User's Guide to Decoupling, Grounding, and Making Things Go Right for a Change," by the author, available from Analog Devices. A heavily edited version appeared in EDN Magazine, October 5, 1975.

$$V_O = i_O R_F + \left[(V_A - V_{OS})\left(1 + \frac{R_F}{R_O}\right)\right]$$

$$\left[R_O \| R_F = 100\Omega\right]$$
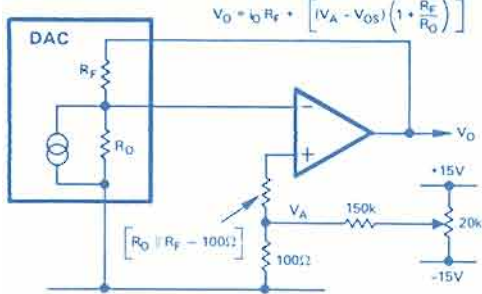
(a) Nulling offset with voltage applied to op-amp reference input.

$$V_O = i_O R_F + \left[V_{OS}\left(1 + \frac{R_F}{R_O}\right) - i_1 R_F\right]$$
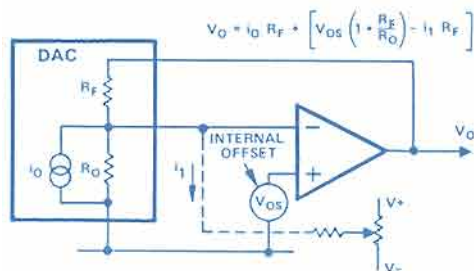
(b) Nulling offset with current added at op-amp summing point.

Figure 5. External offset-null methods.

For example, if the DAC is an inverted R-2R-ladder type, as shown in Figure 6, the output resistance, R_O, approaches R for codes containing many 1's, 3R for codes containing a single 1, and ∞ for all-0's. If R = 10kΩ, the resistance looking back into the network is about 10kΩ for more than four 1's and 30kΩ for a single 1. Thus, for the one-bit transition from 0011111111 to 0100000000, the error voltage, $V_{OS}(1 + R_F/R_O)$, changes from 2 V_{OS} to $(4/3)V_{OS}$. If the offset had been nulled at all-0's $(1 + R_F/R_O = 1$, since $R_O \to \infty)$, the offset error will be $+V_{OS}$ at the first code and $(+1/3)V_{OS}$ at the second code; the incremental change of error will be $(-2/3)V_{OS}$. If V_{OS} is not much smaller than the voltage equivalent of the least-significant bit, a tangible error will result. It will be especially pernicious in the case of a multiplying-DAC application with small analog inputs. The solution is simple: use Figure 5a instead of 5b.
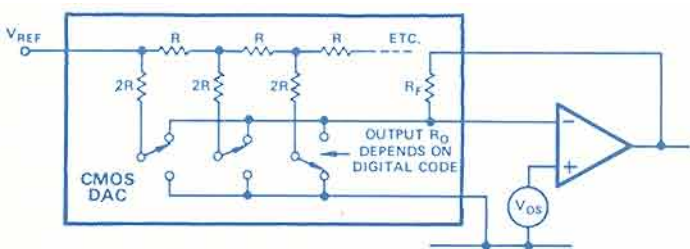

Figure 6. Variable output resistance of inverted R-2R ladder in CMOS and voltage-switching DAC's.

"Foreign" currents in common ground and power lines can introduce offset, noise, and other errors that will be amplified in the same way as V_{OS} errors. It is important to refer the amplifier circuit (and its external V_{OS} trim), the load across which the output voltage is developed, and the DAC's reference input — all of these — to the DAC terminals, in the manner shown in Figure 7.
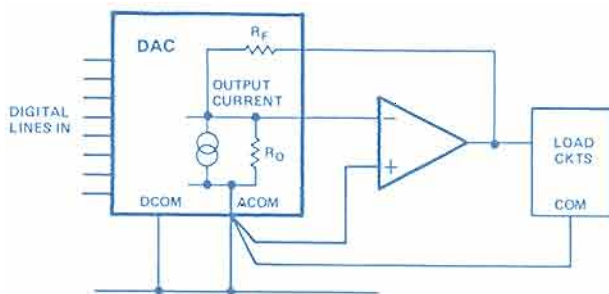

Figure 7. Referring buffer amplifier and load circuits to analog common.

BYPASSING AND DECOUPLING

In "virtual-ground" systems, such as an op amp, driven by a current-output DAC, the DAC output current doesn't actually return to ground, but to one of the power supplies, by way of the op amp's output stage (Figure 8). To reduce the impedance in the high-frequency current path, the bypass capacitor should be connected so as to return the currents from one (or both) power terminals to ground *at the DAC*. If the DAC output is active, it may require bypassing of its own supplies for the same reason.

WARNING: You and your drafting department may have conflicting objectives. Your objective is to design circuits that work and to communicate the important details to whoever assembles them. Your drafting department (or so it may seem) has the objective of drawing nice, neat, squared-off diagrams, in which the lines representing conductors are nicely equipotential. You may have noticed that, in Figures 7 and 8, these niceties have been avoided. The lines are configured to resemble closely the job that the wires perform, converging at the common analog connection. Again, the bypass-capacitor lead, in Figure 8, wends its way purposefully around the op amp's acute angle to its power-supply terminal, rather than shooting straight up to meet the power-supply line (a sure recipe for costly debugging). If you think your drafting depart-
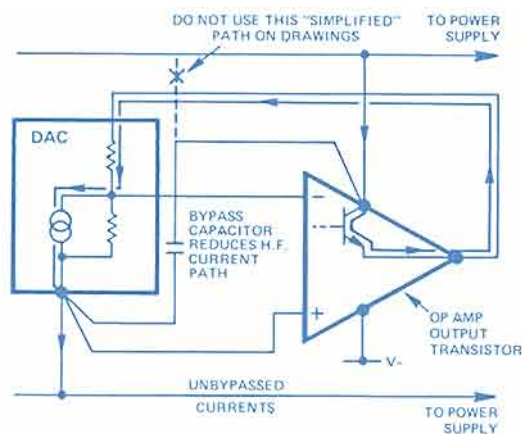

Figure 8. Bypassing power supplies for virtual-ground applications. Arrows show unbypassed current flow.

ment may have a mind of its own, you may want to include a special message for the person who builds the circuit to be sure that it gets built the way you want it built.

Figure 9a shows an example of ineffective decoupling. Here, the op amp drives a load, which connects to a long ground line (returning to the power-supply terminal), and the supply-decoupling for the amplifier returns to the power supply through another long line. The return path for the load current is as long as, or longer than, the supply lines powering the op amp. The "local" decoupling is not only ineffective; it may actually contribute to noise on the power-ground bus.
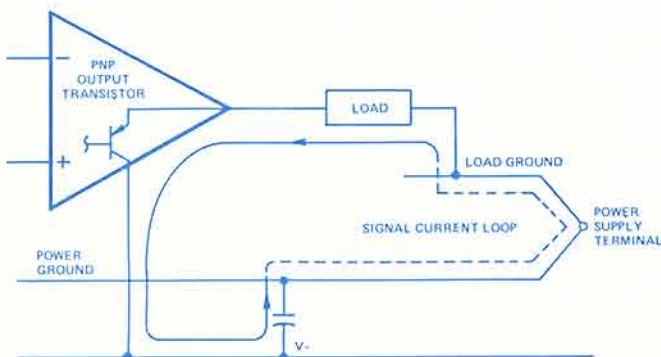
(a) Decoupling for negative supply ineffective.
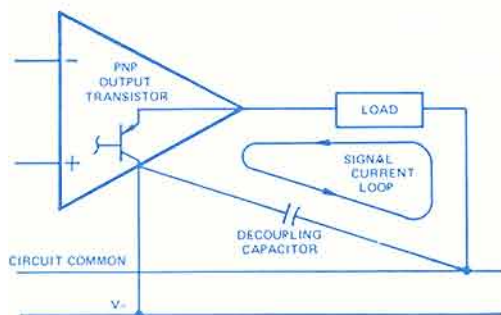
(b) Decoupling negative supply optimized for "grounded" load

Figure 9. Effective and ineffective decoupling.

The cardinal rule of decoupling is: *Make it easy for the current to get back by the shortest path.* Figure 9b shows a more-effective scheme, in which the decoupling capacitor connects by the shortest path between the load return and the load-voltage control element. Here, an op amp, swinging a resistive load-circuit negative, drives the load from an internal PNP transistor, connected to V−. Decoupling the V− pin of the op amp to the low side of the load provides the most direct return path for high-frequency currents, and bypasses them around ground and power buses.
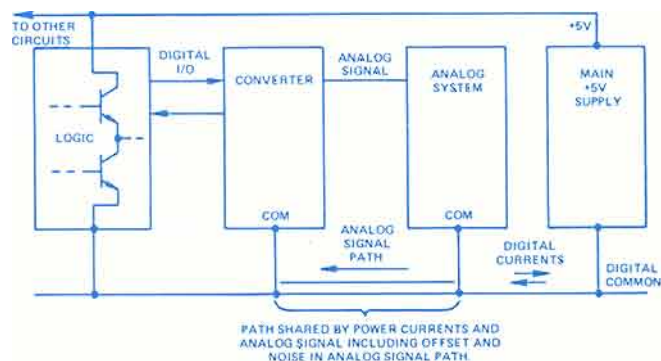
GROUNDING

Great amounts of effort, and many decoupling components, are spent in the attempt to correct problems created by poor ground-current management. In large systems, and in systems which deal with both high-level and low-level signals, "ground" (or common bus) management becomes an important aspect of design. The worst sin− allowing low-level analog signals to share conductors with logic returns or power connections− is an invitation to trouble.

Figure 10a shows an example of a path, shared by digital and analog signals, between the common connections of a converter and an analog system. If the least-significant bit is worth 2.5mV, and digital on-off current fluctuations of the order of 100mA are flowing an a lead with 0.1Ω resistance, the resulting 4-LSB uncertainty (not including spikes and glitches) suggests that it would be folly to waste money on a 12-bit converter, when 10-bit resolution is the very best that can be hoped for, because of wiring limitations.

As Figure 10b shows, in concept, an analog subsystem can be locally interconnected, with a single-wire connection to the digital common. This signal connection carries only the digital currents required for the *converter's* digital interface. Moreover, analog signals are not forced to share a conductor, even with those currents. The analog subsystem should be powered by a supply with a local common return, which may be connected to the digital common but does not share any current-carrying conductors. Ideally, there are no "foreign currents" flowing between the analog system and the digital system, except for those within the converter. If the two systems are joined only at the converter, the foreign currents share the shortest path, and their effect is minimized.

(a) Shared path produces interference and errors.
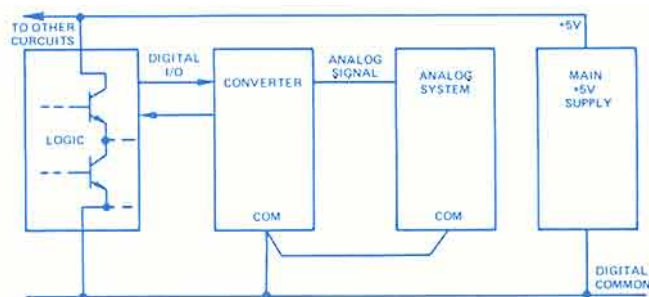
(b) This connection minimizes common impedance between analog and digital (including converter digital currents).

Figure 10. Proper and improper grounding.

In practical systems, it is often impossible to avoid multiple foreign-current paths. In systems which include several d/a and a/d converters, for example, each converter is a path for digital currents, yet it must have access to the analog

signal common. Frequently, the ground problem in such systems can be treated by using an analog common which handles *only* analog signal returns - and a separate system of returns for all digital or high-level signals (Figure 11). Occasionally, a third system of analog power commons may be used to advantage. Since the analog common must be connected to the digital common at no more than one point, safety diodes should be added to any modularized system. These diodes prevent large voltages from developing between ground systems if the key grounding unit, or "Mecca", should be removed from the system.
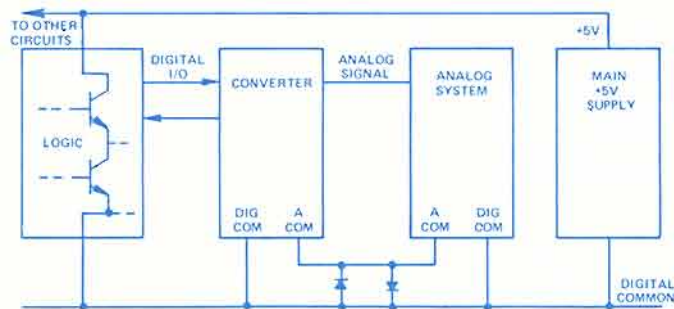

Figure 11. Improved ground current management (analog and digital common must be joined in either converter or analog system. Diodes are fault protection if this connection is broken.)

WHEN COMMON GROUND IS IMPRACTICAL

In large systems, it is often impractical to rely on a single common point for all analog signals. In these cases, some form of differential (or even *isolation*) amplifier is required to translate signals between ground systems. For the inveterate op-amp user, a simple subtractor, or "dynamic bridge" circuit may come to mind. These circuits translate a signal which is referred to one ground system into a similar or amplified signal, referred to a different ground system (Figure 12). The common-mode rejection of the amplifier and a resistance-ratio match are used to eliminate the effects of voltage differences between the two grounds, or common points.
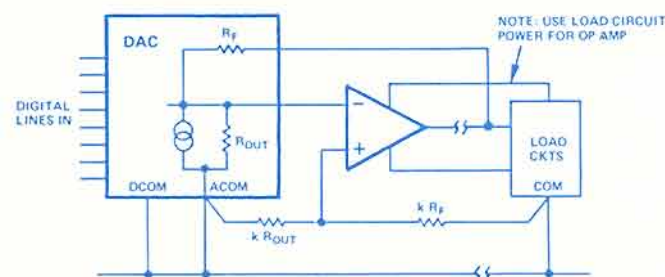

Figure 12. Use of differential amplifier to eliminate the effects of common-mode voltage.

It is generally wise to power the op amp from the power available at the *load* side of the circuit, and/or to decouple it with respect to the *load* common. The reason for this can be deduced from the circuit architecture of the most-common types of op amps (Figure 13).
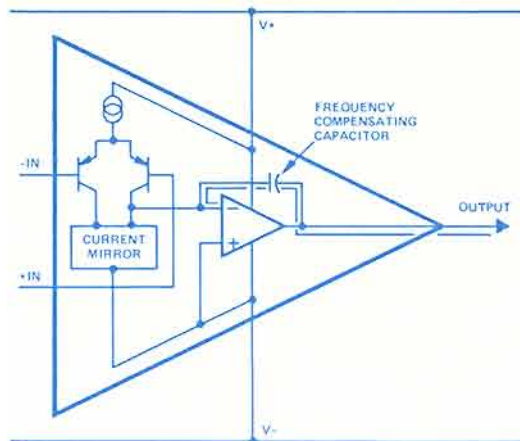

Figure 13. Typical op-amp circuit architecture. Reference for output integrator is V−.

An op amp converts a differential input signal to a single-ended output signal. In many popular op amps, the differential-to-single-ended conversion is done with respect to V− (some use V+), and the resulting signal drives an integrator.[3] The integrator characteristic is used to frequency-compensate the amplifier, and the integrator input is referred to the single-ended output, at V−. The integrator acts as a unity-gain follower for fast signals applied to its non-inverting (or reference) input. As a result, signals applied to the V− terminal have their high-frequency components conveyed directly to the output. Signals having frequency components above the amplifier *closed-loop* bandwidth will be transmitted from V− to the output with little or no attenuation.

As Figure 14a shows, if the op amp used as a subtractor amplifier is powered from or bypassed to the same common line as the input signal, any high-frequency signals associated with that common will appear as part of the output signal. If the ground-noise includes appreciable high-frequency noise (such as logic currents produce), the common-mode rejection will be defeated.

If, on the other hand (14b), the op-amp supply terminals are referred to the *output* signal common, no extraneous signals are coupled into the integrator. Any ground noise appears as a common-mode input signal and is reduced by the common-mode rejection of the amplifier (which is typically very much better than the negative-supply-voltage rejection at high frequencies).

Since noise-rejection performance of the subtractor depends on carefully matched source and feedback resistance ratios, it cannot be used in all situations. Whenever the source impedance cannot be controlled, or is exceptionally high, the subtractor (or dynamic bridge) becomes impractical. In this situation, ground noise and other remote-grounding difficulties can often be avoided by the use of an *instrumentation amplifier*.

IC instrumentation amplifiers, such as the AD521, accept differential input signals at high impedance, provide a fixed gain (which can be selected without introducing overall feedback that joins the input and output circuitry), and

[3] The reference mentioned in footnote 2 provides considerable detail regarding the integrator-reference and compensation schemes of some 32 device families.
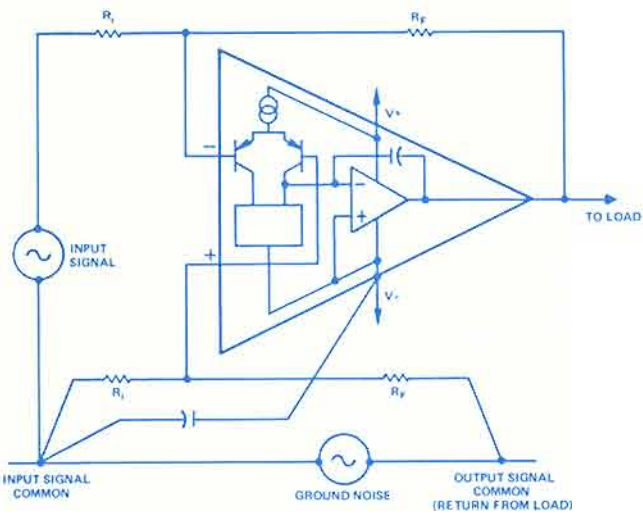
(a) Decoupling to *input* common includes ground noise in the path from the load to the integrator driving the output.
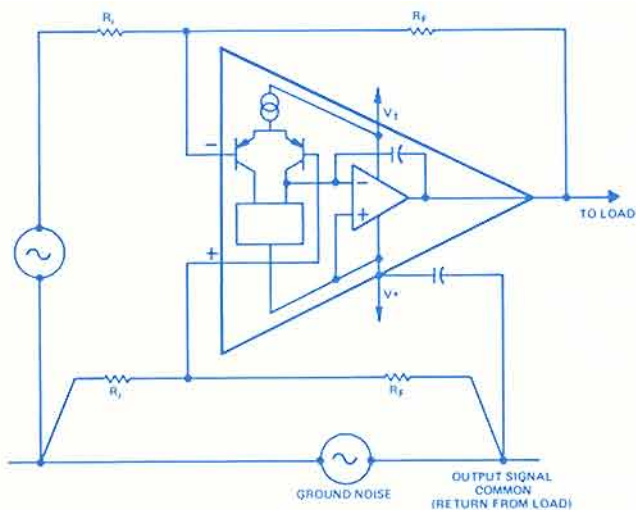
(b) Decoupling to *output* common eliminates ground noise from integrator reference path. Ground noise is minimized in output signal.

Figure 14. Proper and improper decoupling of subtractors using op-amp with integrator referred to V−.

develop the output voltage with respect to a reference terminal, which may be connected to the input common of a remote load-circuit (Figure 15).

Some instrumentation amplifiers are quite versatile and can provide additional functions, while isolating the common
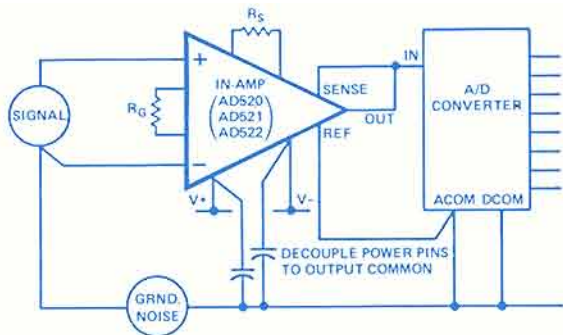

Figure 15. Use of instrumentation amplifier to interface separate ground systems.

returns. For example, the output-reference terminal can be used to add fixed or variable bias voltages to the output.

If the common-mode voltages are very large, or if galvanic isolation is essential for safety, isolation amplifiers, such as the 286 (described elsewhere in these pages), or amplifiers powered by dc-to-dc converters may be highly desirable.

A/D CONVERTERS

The input impedance of many analog-to-digital converters changes during the conversion process and can affect the performance of an amplifier furnishing the input signal.

For example, in successive-approximation converters, the input current is compared to a trial current (Figure 16). The comparison point is diode-clamped, but it may swing plus-and-minus several hundred millivolts. This gives rise to a modulation of the input current. The output impedance of a feedback amplifier is made artifically low by the loop gain. At high frequencies, where the gain is low, the amplifier output impedance rises to its open-loop value. In the case of most IC amplifiers, the open-loop output impedance is a minimum of 25 ohms, due to current-limiting resistors (but is more typically 100-200 ohms).
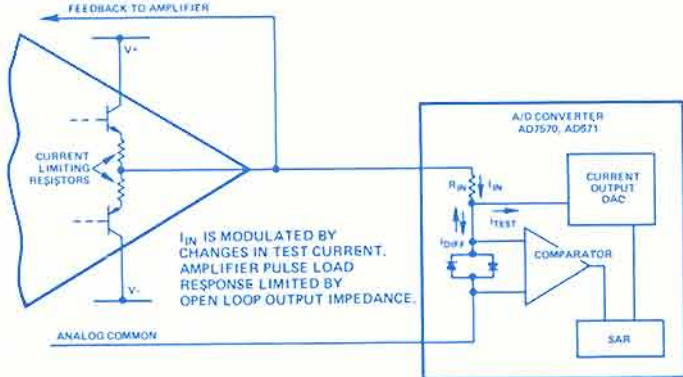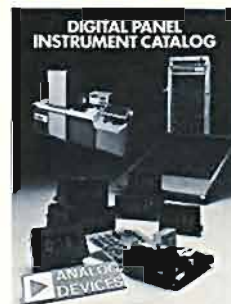

Figure 16. Relationship between successive-approximations A/D converter and op amp that is the source of the input signal.

Even a few-hundred microamperes, reflected from the change in converter loading, can introduce errors in instantaneous input voltage. If the conversion speed and the bandwidth of the amplifier are compatibly fast, the output may return to the nominal voltage before the converter makes its comparison, so that little or no error is introduced. However, many precision amplifiers have relatively narrow bandwidth. This means that they recover very slowly from output transients. Naturally, precision amplifiers are more likely to be used in high-resolution systems, where small errors are less tolerable. As a result, fast, high-resolution systems may suffer from amplifier output-transient errors.

There are a number of ways of solving the problem. Perhaps the easiest is to use an a/d converter that has its own on-board buffer, e.g., the AD572, or most modular types. In other cases, a sample-hold, with low output impedance, may serve as a buffer, as well as to provide the sampling function. Another solution is to use (carefully) a wideband op amp, such as the AD509, which does not include output current-limiting resistors. Finally, it is not difficult to construct an inside-the-loop buffer that can stiffen the output of a slow, accurate amplifier.

Figure 17 shows a simple unity-gain buffer, constructed from an NPN and a PNP transistor in a compound connection. The output impedance of this buffer remains low at high frequencies. A good rule-of-thumb for selecting transistor types to use in complementary-compound is that the input device — in this case, the NPN— should be a high-frequency transistor, and the output device— the PNP— should be a relatively slower transistor.
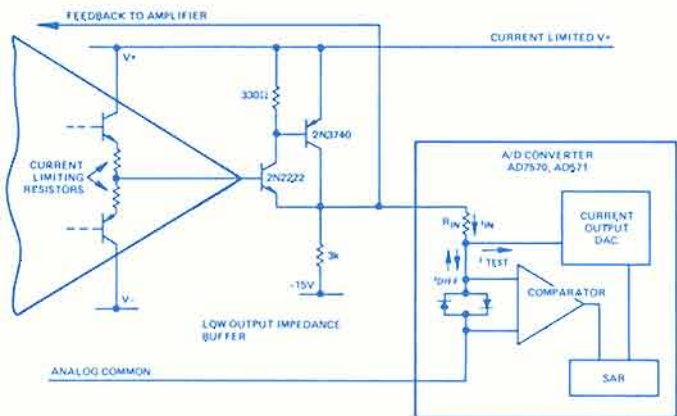

Figure 17. Inside-the-loop buffer provides stiff drive for unipolar ADC.

Since the buffer is not current-limited, a small power-device, capable of pulling down a 200-300mA current-limited supply without damage, has been suggested as the PNP. If the system is definitely safe from overloads, a smaller PNP can be used. This buffer is intended for positive unipolar signals; the 3kΩ resistor provides ample bias to keep the output impedance low over the active range.

A more-complex, protected buffer, for better performance with bipolar input signals is shown in Figure 18. An AD580 voltage reference can be used as a constant-current load to keep the buffer active over the bipolar range. This buffer also includes a bypassed resistor to limit the available output current without pulling down the power bus.
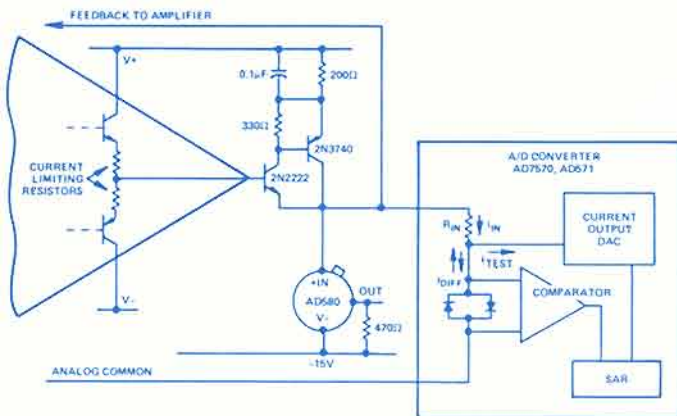

Figure 18. Protected buffer capable of driving bipolar signals into ADC.

TO CONCLUDE

As we told you, our objective was to make you aware of some of the analog problems of implementing interface circuitry, to start you thinking about how to solve them, and to give you some concrete ideas (but not "cookbook remedies"). We hope that they will help make you next system startup somewhat less painful.
▶▶▶

Worth Reading

FREE DIGITAL PANEL-METER CATALOG

This 84-page data book* will tell you much of what you'd like to know about DPM's, in general, and just about everything you need to know about Analog Devices DPM's, in particular.

An aid to the engineer who designs with, specifies, and uses digital panel meters, the book comprises four major sections: "Orientation," "First-Choice Selection Guide," and two data sections, "First-Choice Product Descriptions" and "Secondary Product Descriptions."

"Orientation" is a brief guide to understanding and selecting DPM's; it includes definitions of terms and specs, application notes, and a review of what it takes to design and produce reliable, low-cost DPM's. "First-Choice Selection Guide" is a compact tabular presentation of comparable salient data on all DPM's recommended for new designs, to lead the designer to the one-or-two models most appropriate for the application.

"First-Choice Product Descriptions" are essentially *complete data sheets* on all products in the "First-Choice" category. "Secondary Product Descriptions" include abbreviated data sheets on older products that are still available.

ANALOG-DIGITAL CONVERSION NOTES

by the Engineering Staff of Analog Devices, edited by Dan Sheingold, 1977, $5.95.†

Most readers of this Journal own or know about our little *Analog-Digital Conversion Handbook*, of 1972 vintage. Still an excellent compilation of tutorial and applications material, its product references (and by association, the rest of its contents) may seem antiquated, because it was published long before the ready availability of low-cost IC converters and microprocessors.

To provide a version reflecting the current state of the mart to accompany our recent series of converter seminars, we have modernized Parts I and II ("Converters at Work" and "Converters"), integrated two extensive, wholly new, chapters, "Subsystems and Data Communications" and "Converter Microcircuits," and published the expanded Parts I and II in a pleasing, large (8½" x 11"), 246-page format, with a new title.

The chapter titles may provide a brief flavor of the volume: Data-System Components, Data Acquisition, Data Distribution, Subsystems and Data Communication, Analog Functions with Digital Components, Applications of Converter Systems; and, in Part II, Understanding Converters, Converter Microcircuits, Converter Design Insights, Testing Converters, Specifying Converters, Applying Converters Successfully.

*For your copy of the new DPM catalog, use the reply card.
†The book may be purchased by sending checks or Master-Charge information to P.O. Box 796, Norwood MA 02062.

LOW-COST ISOLATION AMPLIFIER
At $37 (100'S), 286J Makes Multi-Channel Isolation Easy
Well-Received Features of the 284J Improved and Extended

Not too long ago, we introduced in these pages the Model 284J*, a superb, compact, low-cost isolator. Naturally, its economy and performance caused a number of our customers to develop a voracious appetite for them. A few users, who were involved with low-level signals in multi-channel applications, wondered if we could provide an additional version, modified slightly to utilize synchronized carriers, to avoid the possibility of small errors due to beat frequencies developed by small amounts of crosstalk.

And while we were at it, could we do something to increase the isolated front-end power-supply capacity? It was a useful innovation, but it seems that there was just never quite enough voltage or current . . . and could the programmable-gain range be upped to 1-100, instead of 1-10? Could we do all this without increasing the package size or the price, and without decreasing the common-mode voltage and rejection ratings?

Neat trick, but we could and did! And with a price *reduction*.

Enter the 286J.* It provides isolation at continuous dc common-mode voltages up to 2500V and has a minimum common-mode rejection of 110dB at 60Hz with 5kΩ source unbalance. Its leakage current spec of 1.5μA rms *max* at 115V ac, 60Hz, is designed to improve on the 10μA *max* spec of UL544. In addition, it is designed to meet IEEE standards for transient voltage protection (SWC) and MIL-STD 202E environmental testing.

Its nonlinearity is low (±0.05% for peak-to-peak outputs to 10V), it has high gain stability (0.01%/°C and 0.001%/1000hr) and wide signal range (20V p-p at output and input for unity-gain connection).

To answer needs mentioned earlier, its isolated dual output power is ±15V dc at up to ±15mA, its gain range is 1-100V/V, and upwards of 1000 channels could be used with synchronized oscillators. The package size — same as for the 284J — is 1.5"x1.5"x0.62" (38x38x15.8mm). It

*For technical data, use reply card.

4 CHANNEL, ISOLATED DATA ACQUISITION SYSTEM
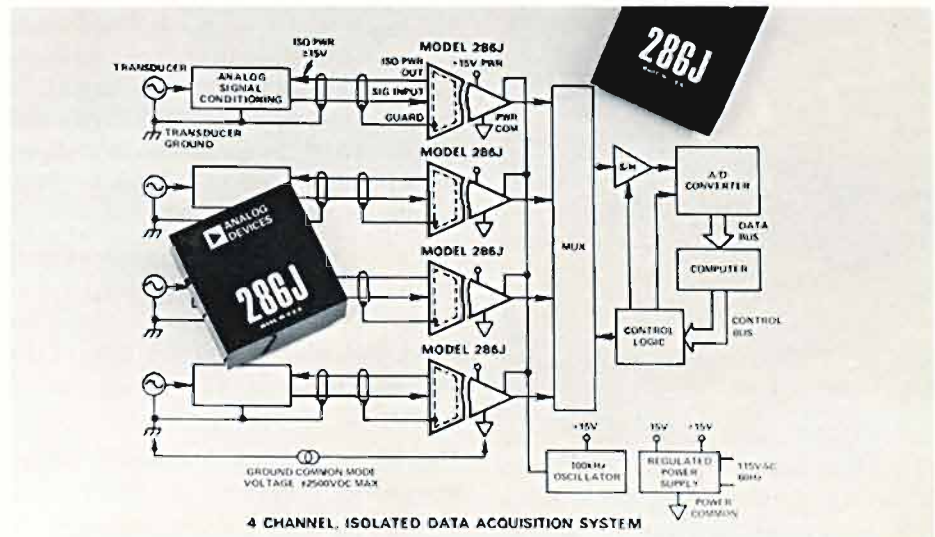
operates from a single dc supply, 8V to 16V, at 13mA quiescent current.

Typical applications are ground-loop elimination in industrial and process control, high-voltage protection and ground-loop elimination in data-acquisition systems, isolation in biomedical and patient-monitoring instrumentation, and general off-ground signal measurement.

Figure 1 is a block diagram of the 286J, showing the transformer coupling used internally for power and signals, the isolated front end, with signal inputs and auxiliary isolated power outputs, and the bidirectionally-protected adjustable-gain input amplifier.
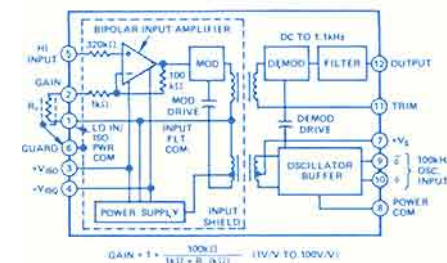

Figure 1. Block diagram - model 286J.

The key to its usefulness in multi-channel systems is the realization that the output of a low-power oscillator can be applied to a number of power amplifiers to furnish the internal drive requirements for each amplifier independently. Thus,

the output of a low-cost 100kHz oscillator, Model 281, can be applied to the ϕ and $\bar{\phi}$ terminals of as many as 16 286J's, to provide excitation for their internal ac power generation (Figure 2). Addition-
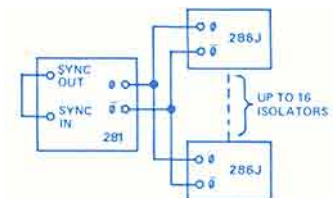

Figure 2. 281/286 connection for driving from 1 to 16 isolators.

al oscillators can be synchronized for additional blocks of 16 channels. The 281's design is very simple, the user can easily make-instead-of-buy — and is encouraged to do so (Figure 3). Price of 286J is $59(1–9); Model 281, $18(1–24). ▶▶▶
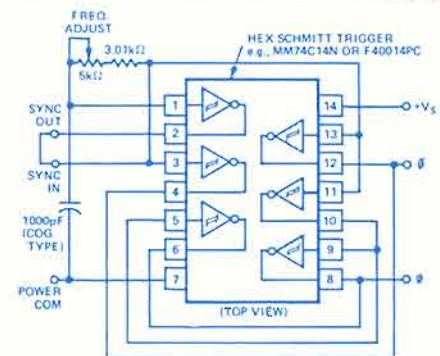

Figure 3. Model 281 100kHz oscillator—logic and interconnection diagram.

DISPLAY-LESS DPM
AD2023 DPM Module Ready-to-Go With Display of Your Choice

The AD2026 digital panel meter* (*Dialogue* 10-2) uses an I^2L chip to eliminate all but 13 electrical parts for a 3-digit DPM. Its low cost, compactness, and implied high reliability made it an instant success. Everyone loved the cost, but not everyone could use it as is — do-it-yourselfers wanted chips, others wanted different packages or displays.

Our response to the clamor is the AD2023, a 2"x2"x0.4" (51x51x10.2mm) ready-to-go DPM module, without case, display, or gain pot. The standard AD2023 byte-serial output has 3 7-segment-decoded digits; the AD2023B has 3 BCD digits, for flexible display choice.

The module can be used wherever a digital display of analog voltage is desired — in instrument housings, on cards (flat or on edge), and wherever space is tight or the readout is unusual (e.g., vertically arrayed numerals).

The high-impedance limited-differential input range is +999mV to -99mV, with automatic polarity and overrange. The 4/s conversion rate can be speeded up to 48-168/s. Controls include "lamp test" and "hold." 1–9/100+ prices are $39/$28 and $37/$27(B). ▶▶▶
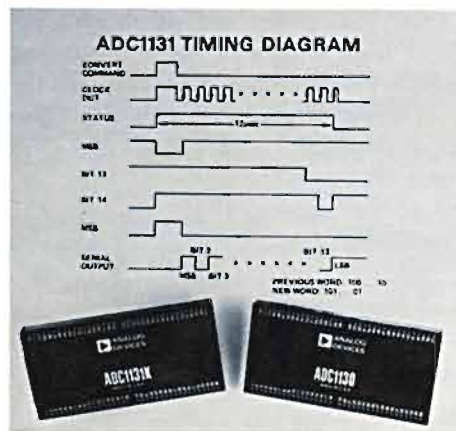
HIGH-PERFORMANCE 14-BIT ADC'S
ADC1130: 14-Bit Conversion in 25µs [$239, 1-9]
ADC1131K: 14-Bits in 12µs, ½LSB Max Nonlinearity

The ADC1130 and ADC1131* are high-speed 14-bit a/d converters, packaged in compact 2"x4"x0.4" (51x102x10.2mm) modules. Using successive approximations, they convert analog voltages to binary, offset binary, or 2's-complement-coded outputs. Data outputs are available in both parallel and serial (NRZ—non-return-to zero) form.

Four analog input ranges are available: 0 to +20V, 0 to +10V, ±10V, and ±5V, chosen by appropriate connections to the device terminals. A short-cycle feature is also included, to permit reduced conversion time, at the expense of lower resolution. Conversion time is proportional to the number of bits, as pin-programmed.

The combination of high resolution—with no missing codes—and short conversion time is useful in a number of applications, including wideband data-digitizing, multichannel computer interfacing, high-accu-

racy data acquisition, X-Ray tomography, and nuclear-accelerator instrumentation—to name a few.

Two versions of the ADC1131 are available, differing in differential nonlinearity specs; ADC1131J — 1LSB max, ADC1131K — ½LSB max. Prices for ADC1131J/K are $279/$369(1–9). ▶▶▶

10V IC REFERENCE
AD581L Laser-Trimmed to ±5mV ±5ppm/°C,max
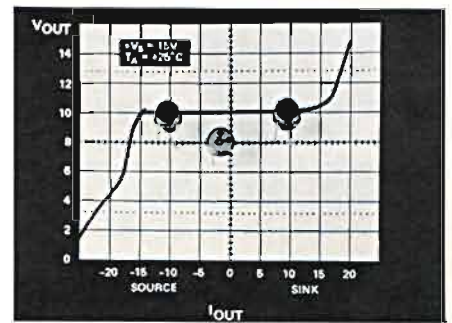
The AD581* is a 3-terminal temperature-compensated monolithic bandgap voltage-reference with a precise 10.00V output level that is independent of dc input voltage, from 12 to 40V. Both the initial voltage and the temperature coefficient are laser-trimmed at the wafer stage (LWT) for high performance at low cost.

The AD581 is recommended as a fixed reference (or reference source) for 8-, 10- or 12-bit a/d and d/a converters and other analog circuitry requiring external precision reference voltage.

Besides its use, with no external components, as a 3-terminal reference, capable of sustaining either positive or negative current flow, it can also be used as a 2-terminal "zener" reference of ±10V (either positive or negative polarity), with an external series resistor.

TWO-TERMINAL -10 VOLT REFERENCE
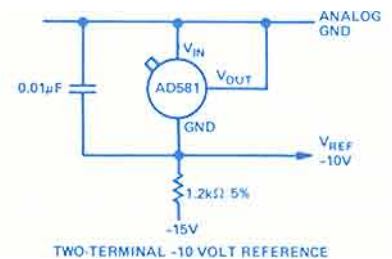
Six accuracy grades (J/K/L and S/T/U) cover both the 0° to 70°C range and the -55° to +125°C range; The S/T/U are also available with full processing to MIL-STD-883, Class B requirements. Initial accuracy for both triads is to ±30/±10/±5mV *max*, and max tempco is 30/15/5ppm/°C (10ppm/°C for AD581U). Prices (100's) are $2.85/$5.95/$11.95 and $9.95/$15.95/$25.75. ▶▶▶

*Use the reply card for technical data.

Potpourri

An Eclectic Collection of Miscellaneous Items of Timely and
Topical Interest. Further Information on Products
Mentioned Here May Be Obtained Via the Reply Card.

NEW PRICES . . . The already low-cost AD2026 3.1-digit DPM, originally introduced at $39
in 100's, has been reduced in price to $29 in 1000's, $34 in 100's, and $49 by 1's.
Improved chip yields and a fast one-way ride down the learning curve are to thank. . .
That learning curve has helped with the '2nd Generation" types generally, and we're
sharing the reduced costs with you. For example, AD2025 and AD2028 (4+3/4 digits) have
been reduced from $259 to $206 and from $249 to $202 in small quantity; the basic
AD2016 was $138, is now $124; the basic AD2021 was $128, is now $115; and so it goes
. . . Watch this space . . .

NEW PRODUCTS . . . The 16-bit DAC1136 is a new low-cost DAC, pin-compatible with our
other "QM" products. Price is only $260 for the DAC1136K (±½LSB nonlinearity) . . .
The SDC1700 basic 12-bit design has spawned 14-bit and 10-bit variations. As you may re-
call, these are low-profile (0.4"H) modular synchro-to-digital converters with self-
contained microtransformers. The 14-bit SDC1704 family start at $440 (1's), and the
10-bit SDC1702's start at $325 . . . The AD2011 was a pioneering rms DPM; but now that we
have the AD2033, with better specs and lower cost, plus log-ratio capability (see
Dialogue 11-1), we are honorably retiring the AD2011.

NEW LITERATURE . . . Use the reply card to get our free new 8-page Power-Supply Catalog.
It covers our whole line of ac/dc power supplies and dc-dc converters, including three
new high-powered chassis-mount models: the 5V/2000mA Model 956 ($89, 1-9), the ±15V/350mA
dual Model 973 ($85), and the ±15V/500mA dual Model 975 ($99) . . . The new power-supply
catalog also shows price reductions on the Model 955 (5V/1A), to $72(1-9) and Model 970
(±15V/200mA) to $68 . . . Now available in hardcover is the MICROPROCESSOR SYSTEMS
HANDBOOK, by Dr. D. P. Burton and Dr. A. L. Dexter - 200 solid pages of information on
µP's and their application with converters. For your copy, send a check for $9.50 (or
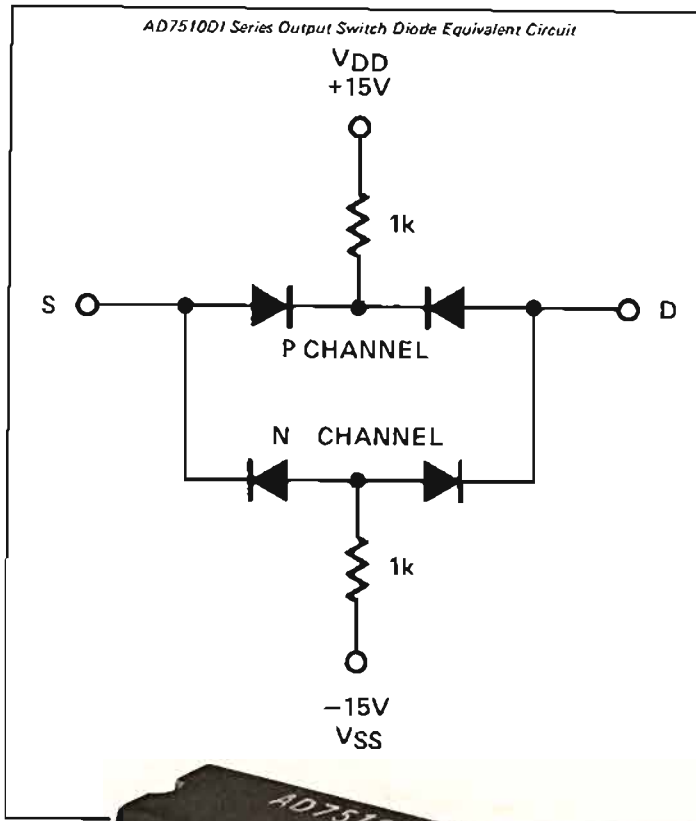complete Master Charge information) to Analog Devices, Box 796, Norwood MA 02062.

NEW INFO ON PRODUCTS . . . LED-type DPM's have been getting brighter displays . . .
Model 51A/B op amp, in inverting applications (such as with DAC's), can settle to 0.1% in
140ns and drive 100Ω loads (100mA at 10V), and -get this- operate over the full-MIL
temperature range, -55°C to +125°C. The data-sheet max temperature spec of 100°C is for
differential operation.

GSA . . . Government agencies and contractors are welcome to write on letterhead for
copies of our GSA Amplifier Price List for GSA Contract GS-00S-04972. Included are
current-, logarithmic-, operational-, and isolation amplifiers . . . 3 other GSA price
lists are currently available: Converters & DPM's, Power Supplies & Synchro Converters,
and Analog Multipliers.

ERRATA . . . DIALOGUE 11-1: Page 14, Figs. 3 & 4 - functions of pins 25 & 24 are inter-
changed; HBS is pin 25, LBS is pin 24 . . . Model 757 Log-Ratio Module data sheet: Fig. 8,
photodiodes are connected backwards . . . AD534 data sheet: The pinout in the Top View is
correct. (Z_1 and Z_2 are interchanged in the chip diagram.) . . . In the SHORT-FORM GUIDE,
the proper units for Ib of the AD515 electrometer IC op amp are pA, not nA; that is,
0.075pA for AD515L . . . AD514 data sheet: offset trim pot 10kΩ, not 10 ohms.

IN THE LAST ISSUE (Volume 11, Number 1, 1977): Versatile Analog-to-Microcomputer Interface
(the RTI-1200 — for Intel SBC 80 and others) . . . Accurate, Low-Cost, Easy-to-Use IC
Multiplier (the monolithic AD534) . . . Highest-Performance Low-Cost 10-Bit Monolithic IC
D/A Converter (the AD561, with on-chip buried Zener reference) . . . CMOS Converters as
I/O Devices (by Ivar Wold) . . . SONICGUIDE(tm) Helps Blind Ambulate Freely (and uses
AD531 Multiplier-Divider) . . . 284J Isolator in Isolated Current Loops . . . Other New
Products: Stable, Linear V/f Converters (458 & 460), True-RMS Digital Panel Meter(AD2033),
18-Bit DAC (DAC1138) . . . Corrected Bound Circuit . . . plus editor's notes, authors, etc.

Overprotection can affect a CMOS switch for life.

AD7510DI Series Output Switch Diode Equivalent Circuit

VDD +15V

1k

S — P CHANNEL — D

N CHANNEL

1k

−15V VSS

But not Analog Devices' AD7510 family of DI CMOS analog switches. They belong to a whole new generation. With positive overvoltage protection, but without any inhibition on performance.

We accomplished it through a unique design, utilizing "on-chip" resistors in series with the power supply. It provides as much as ±25V overvoltage protection. But the resistors only switch in when an overvoltage condition occurs. So normal performance never suffers. And you get both the main assets of an analog switch: a low "ON" resistance of 75Ω and a low leakage current of 400pA.

The equivalent circuit of the output switch element shows that, indeed, the 1 kΩ limiting resistors are in series with the back-gates of the P- and N-channel output devices — not in series with the signal path between the S and D terminals.
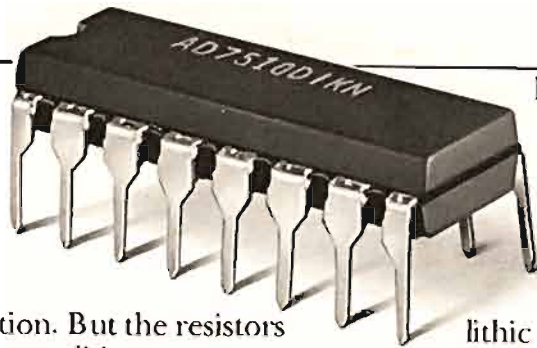
This design, combined with our dielectrically-isolated CMOS fabrication process, prevents latch-up. And allows TTL/CMOS direct interfacing. We also included two other measures of security. Silicon nitride passivation to ensure long term stability and monolithic construction for reliability. Now when it comes to protecting CMOS switches so they can survive in the real world, Analog Devices knows best. Write for our 8-page technical bulletin on the entire family of DI CMOS protected analog switches, to Analog Devices, the real company in precision measurement and control.

◢ ANALOG DEVICES
The real CMOS switch company

For information on CMOS switches and multiplexers, use the reply card