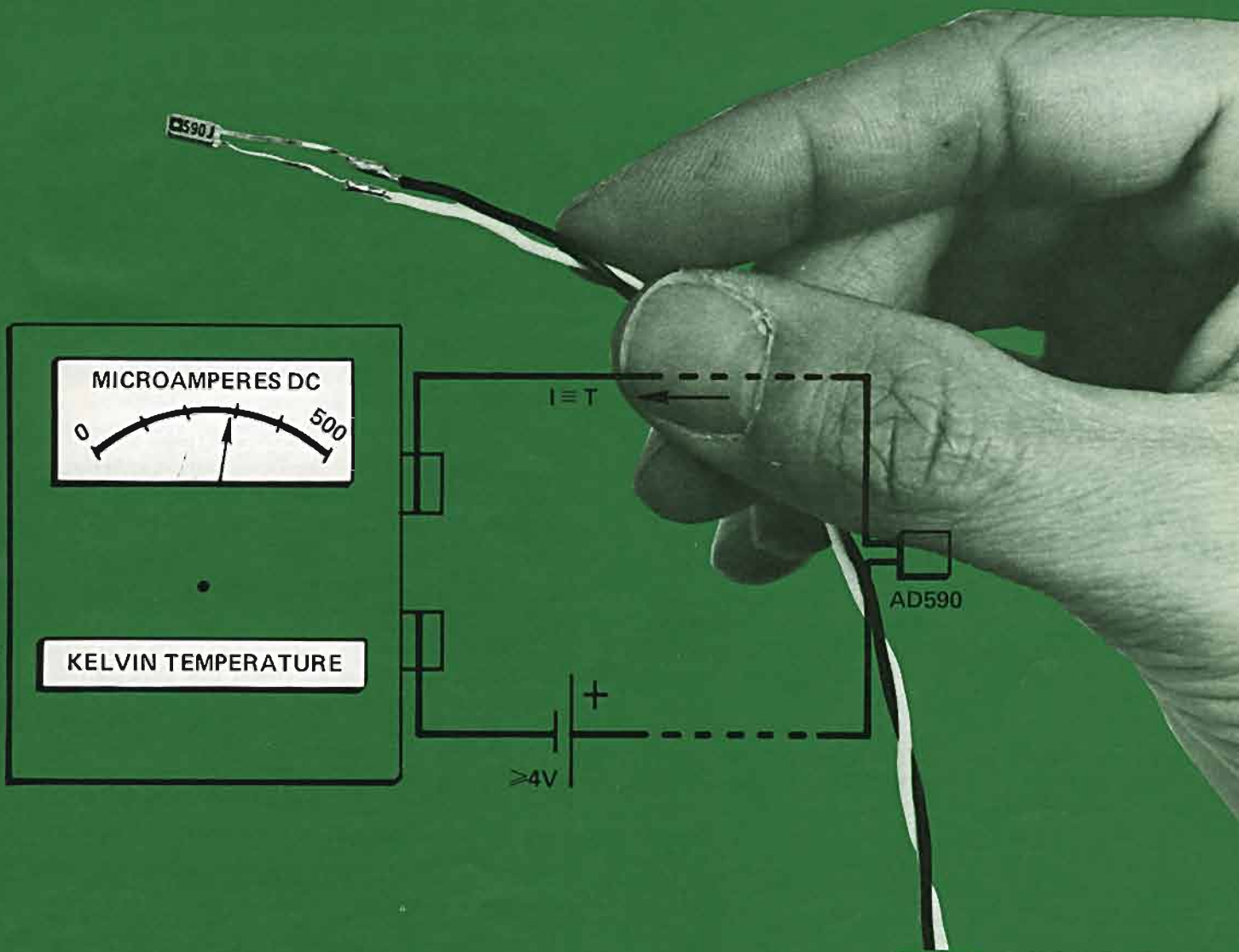


# analog dialogue

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

MONOLITHIC  $1\mu\text{A}/\text{K}$  TEMPERATURE-TO-CURRENT TRANSDUCER (Page 3)

Also in this issue:  
Monolithic Complete 10-Bit A/D Converter  
see page 6  
Complete Contents on Page 3





# Editor's Notes

## TESTIMONIAL



If you are familiar with how things tend to be in a multi-divisional company where one division makes products that another division can use, you probably know that the engineers in the *user* division tend to be a hardbitten bunch of skeptics, who have learned from experience what the magnitude of the optimism-discounting coefficient must be. We think it's less in our Company than most, but it's a necessary adjunct to survival and prosperity and provides the constructive feedback one would hope for from a user.

Consider then our pleased surprise to encounter in an internal memorandum the following comment about the AD571, the monolithic *complete* 10-bit a/d converter: "I have tested...the AD571 samples...Kudos to Paul Brokaw and his crew for the 571 is in order; its performance is superb — as good as most modular 10-bit converters, and better than anyone expected."

### COVER OF DIALOGUE 11-2

The cover of our last issue, which introduced to the World our AD536 monolithic rms-to-dc converter, received a number of favorable comments and one embarrassing query. You may recall that it showed a variety of waveforms; associated with each were three readings, one using the rapidly becoming-obsolete mean-absolute-deviation sine-wave-calibrated "rms," one using the AD536, and the third using a Fluke 931, which happened to be the nearest thing to a standard that we had on hand.

If you are familiar with the problem, you will know that you would expect the AD536 and the 931 to be in reasonable agreement for all readings, and the sine-wave-calibrated full-wave rectified meter to differ by as much as 10% or more. And so it was: the AD536 and the 931 were within 0.5% of one another in all cases.

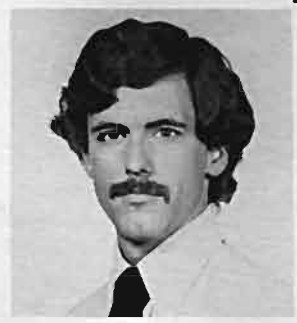
Interestingly enough, the AD536 errors tended to put its readings on the low side, which meant that a typical comparison would be between 1.000V for the 931, 0.997V for the 536, and 0.82V for the DVM. Now, most people would rather compare 1.000 with 1.003 rather than 0.997 — it's a much easier mental subtraction and makes the numbers appear a lot closer (merchants aren't very helpful in that regard when it comes to pricing — isn't \$10.02 an easier price to visualize accurately in relation to \$10.00 than \$9.98?)

In order to do this honestly, and not falsify the data, we decided to normalize the readings to the lower of the two, the 536, and to remove the units, so that we were dealing with dimensionless ratios, rather than volts. It hardly crossed our minds that this could have had the psychological effect of making the 931 look like a less-accurate meter; we found it hard to believe that *anyone* who knew Fluke could fail to recognize which device was the reference. Well, we can once again attest to the validity of Murphy's Law. Our friends at Fluke called us rather worriedly, so we decided to set the record straight. We're sorry if we've caused them any anxiety — the AD536, for all its excellence, will never be a 931!

Dan Sheingold

## THE AUTHORS

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(more authors on page 18)

# analog dialogue

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# 1μA/K IC TEMPERATURE-TO-CURRENT TRANSDUCER

At \$1.95 in 100's, 2-Terminal AD590J is Easy to Choose & Use

Temperature -55°C to +150°C, Linearity error to 0.5°C

by Mike Timko and Goodloe Suttler

The AD590\*, a two-terminal monolithic-IC temperature transducer, produces an output current that is proportional to absolute temperature ("PTAT"). The device acts as a high-impedance constant-current regulator, passing 1μA/K; performance is essentially independent of terminal voltage, from +4V to +30V. State-of-the-art laser trimming, in conjunction with extensive final testing, insures that AD590 units may be used interchangeably.

The device is calibrated to provide an output current of 298.2μA (AD590L) at 298.2K (+25°C). The AD590L is tested over its operating temperature range, -55°C to +150°C (243.2K to 448.2K), to be within 1°C of the ideal PTAT response. Maximum guaranteed nonlinearity over the rated temperature range is ±0.5°C. AD590's are hermetically sealed in the TO-52 package; they are also available in chip form.† MIL-STD-883 processing to Level B is also available.

The most-outstanding feature of the AD590 is the ease with which it can be applied to a wide variety of demanding temperature-sensing problems. Unlike conventional thermoelectric sensors (e.g., thermocouples, thermistors, RTD's), no linearity correction is needed for display of standard temperature scales. In remote applications, thermocouple wires or Kelvin resistance connections are unnecessary, since the high output impedance of the AD590 allows the use of any well-insulated pair of wires. Besides making it possible to ignore voltage effects, such as lead drops and thermocouple effects, the AD590's current output makes the device especially easy to multiplex, as the Applications section will show. The AD590 is thus an attractive way to measure temperatures between -55°C and +150°C, especially in view of its low cost (\$1.95 in 100's, for AD590J).

## HOW THE AD590 WORKS<sup>1</sup>

The circuit of the AD590 exploits a fundamental property of transistors to produce a voltage that is PTAT. That voltage is then converted to a current that is PTAT by means of low-temperature-coefficient thin-film resistors on the chip. Finally, the entire supply current of the device is made proportional to that PTAT current.

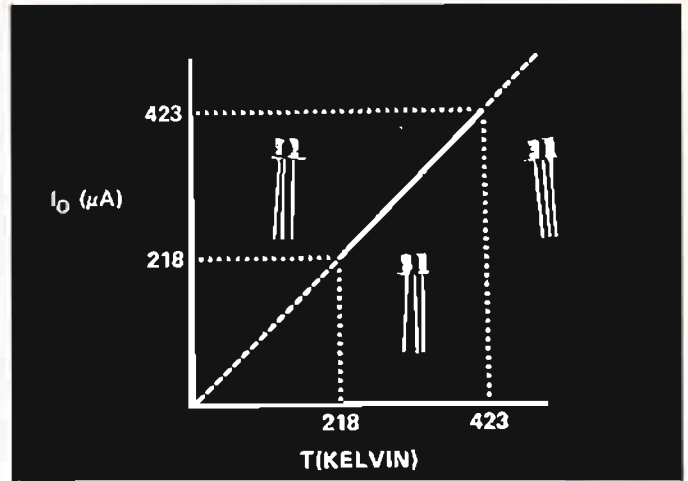
The relationship between base-emitter voltage ( $V_{be}$ ) and collector current of a bipolar transistor is the key device property for this purpose. The difference between the  $V_{be}$ 's of two identical transistors, operating at a constant collector-current ratio,  $r$ , is PTAT, since

$$\begin{aligned} \Delta V_{be} &= \frac{kT}{q} \ln \left( \frac{rI}{I_{cco}} \right) - \frac{kT}{q} \ln \left( \frac{I}{I_{cco}} \right) \\ &= \frac{kT}{q} \ln r \end{aligned} \quad (1)$$

\*Use the reply card to obtain complete information on the AD590.

†A miniature ceramic package will soon be available; see cover photo.

<sup>1</sup>For a detailed description of the circuit, see M. P. Timko, "A Two-Terminal IC Temperature Transducer," *IEEE Journal of Solid-State Circuits*, Vol.SC-11, pp. 784-788, December, 1976.



where  $k/q$  (the ratio of Boltzmann's constant to the charge of an electron) is about  $86.2\mu\text{V}/\text{K}$ ,  $I_{cco}$  is the same for both transistors, and  $T$  is the temperature of both transistors, kelvin.

Figure 1 shows a simple circuit that can employ this principle to produce a current that is PTAT. If one assumes that the output impedance and  $\beta$  of all transistors are infinite, the current mirror, Q3 and Q4, forces currents  $I_{c2}$  and  $I_{c1}$  to be equal, since the base-emitter voltages are equal.

Q2 is made up of eight transistors that are identical to Q1, connected in parallel. Therefore, the current density in Q1 is 8X that of Q2. According to (1), the difference of the  $V_{be}$ 's

(continued on the next page)

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of Q1 and Q2, appearing across resistor, R, is

$$V_T = \frac{kT}{q} \ln 8 = 86.2 \times 10^{-6} \text{ V/K} \times 2.08 \times T$$

$$V_T/T = 179 \mu\text{V} \quad (2)$$

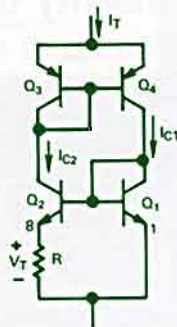


Figure 1. Simplified schematic.

Note that the voltage is PTAT and independent of the actual magnitudes of the collector currents. The collector currents are determined by  $V_T/R$ . The total current flowing through the circuit is twice the current through R, since the currents in the two branches are equal. Therefore,

$$I_T = 2 \frac{V_T}{R} = \left[ 2 \frac{k}{qR} \ln 8 \right] T \quad (3)$$

If  $R = 358 \Omega$ ,  $I_T/T = 1 \mu\text{A/K}$

This basic circuit would, of course, work only for perfect transistors. The complete circuit of the AD590, shown in Figure 2, realizes the desired properties with practical transistors. The four transistors of Figure 1 can be seen in Figure 2 as Q11, Q9, Q1-Q2, and Q3-Q4. The rest of the circuitry has been added to make the two sides of the circuit operate symmetrically. The key to the success of this configuration is Q10, the diode-connected NPN; it forces all the additional circuitry to run at the same current as Q11, thus preserving the temperature-proportionality of the total current.

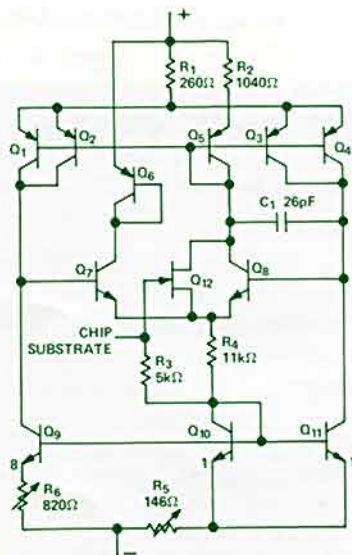


Figure 2. Complete AD590 schematic diagram.

The chip substrate, which is at  $V_-$  on most IC's, is connected through R3 to the collector of Q10 in the AD590. At temperatures above  $125^\circ\text{C}$ , the leakage currents of the isolation pockets can become significant; this substrate connection ensures that these leakage currents are accounted for as part of the PTAT current, rather than as an additional error term.

R5 and R6 are laser-trimmed at the wafer stage. They permit either positive or negative incremental trimming. C1, R3, and R4 provide frequency stability for the AD590's closed-loop circuitry. Q6 insures that the AD590 can be connected backwards without damaging either the transducer or the measuring equipment connected to it. Figure 3, the I-V characteristic of the AD590 at three different temperatures, shows that, like a true current-source, the AD590 is independent of applied voltage greater than about 3V. A  $10\text{k}\Omega$  ( $100 \mu\text{A/V}$ ) load line indicates the impedance scale.

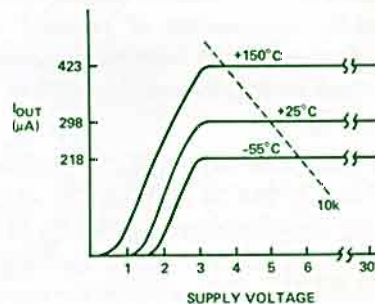


Figure 3. Current vs. voltage.

### AD590 ACCURACY AND EXTERNAL TRIMMING

The accuracy with which temperature may be measured using the AD590 depends on how it is applied and the degree of external adjustment used. The breakdown rating of  $\pm 200\text{V}$ , case to either terminal, permits the AD590 to be placed in intimate contact to measure the temperature of conducting devices operating at common-mode voltages substantially higher than the AD590's supply. The data sheet discusses the AD590's thermal properties in various media.

With no adjustment by the user, the AD590L's indicated temperature is guaranteed to be within  $\pm 1^\circ\text{C}$  at  $25^\circ\text{C}$ , and within  $\pm 2.4^\circ\text{C}$  over the entire temperature range ( $-55^\circ\text{C}$  to  $+150^\circ\text{C}$ ). Device errors consist of a *calibration error*, proportional to absolute temperature, and a deviation from proportionality (*nonlinearity*). The change of maximum calibration error from its  $25^\circ\text{C}$  value is a maximum of  $1.4^\circ\text{C}$  at  $+150^\circ\text{C}$  and  $0.7^\circ\text{C}$  at  $-55^\circ\text{C}$ . The maximum linearity error is  $\pm 0.5^\circ\text{C}$ . With a 5V supply, a microammeter will read temperature (kelvin) to the specified device accuracy (Figure 4).



Figure 4. Direct-reading absolute-temperature analog thermometer.

The scale factor can be trimmed by adjusting a voltage readout resistance (Figure 5) to correct the reading at a single midrange temperature, usually  $25^\circ\text{C}$ . Since the calibration error is PTAT, the remaining error is principally nonlinearity. For increased accuracy over limited temperature ranges, a two-point trim may be used, as shown in Figure 6. In this configuration, R1 provides an adjustable offset for one end of the range, e.g., 0V output at  $0^\circ\text{C}$ , and R2 provides an adjustment for the other end of the range, e.g., 10V output at  $100^\circ\text{C}$ , for a  $10^\circ\text{C/V}$  calibration.



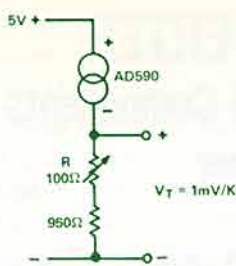


Figure 5. One-point calibration.

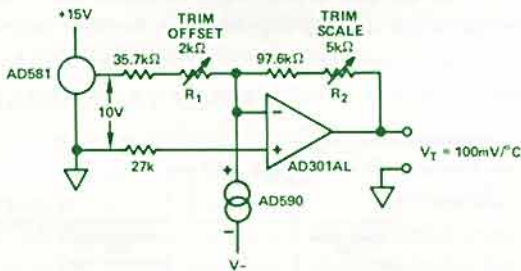


Figure 6. Two-point calibration for 0°C to 100°C range.

## APPLICATIONS

Because the AD590 is a two-terminal current generator, it can be instrumented in many useful circuit configurations. For example, a number of temperatures can be *summed* (hence *averaged*) by connecting the devices in *parallel*. The *lowest* of several temperatures can be read by connecting AD590's in *series*.

Figure 7 shows a simple scheme for measuring temperature difference. As shown, zero output at identical temperatures can be trimmed at one point with  $R_T$ , and gain can be set to any desired sensitivity by the choice of  $R_3$ .

If  $V_{S+}$  and  $V_{S-}$  are caused to differ greatly, the difference of the dissipations will cause a differential temperature rise, which may be useful for measuring ambient thermal resistance in applications such as fluid-level detection and anemometry.

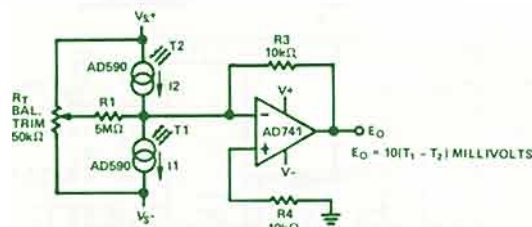


Figure 7. Temperature difference measurement with AD590's.

Figure 8 shows the use of a standard low-cost DPM for readout of temperature on any of the standard scales. For temperatures K or °R only, the AD580 offsetting circuit can be omitted.

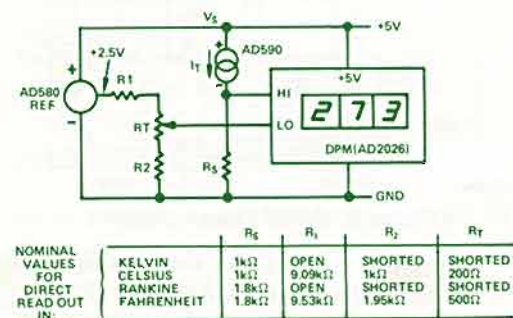


Figure 8. Reading temperature directly with a DPM.

Figure 9 is a multiplexing scheme unique to the AD590. Here, the PTAT current through any one of 80 remote sensors can be read, independently of CMOS switch resistance, via only 18 wires, as addressed by a 7-bit word. The eighth bit can be an *inhibit* line that turns all sensors off for minimum dissipation while idling.

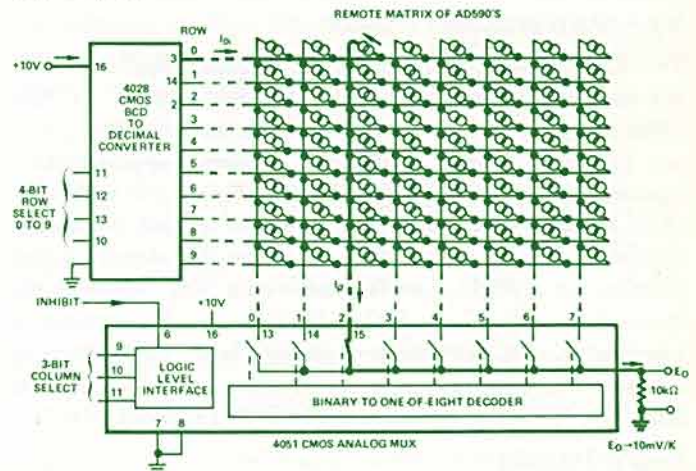


Figure 9. Matrix multiplexing scheme for temperature readout. Heavy line shows path for  $I_{O2}$ .

## AD590 SPECIFICATIONS

(typical @ +25°C and  $V_S = +5V$  unless otherwise noted)

MODEL	AD590J/K/L
<b>ABSOLUTE MAXIMUM RATINGS</b>	
Forward Voltage (E+ to E-)	+44V
Reverse Voltage (E+ to E-)	-20V
Breakdown Voltage (Case to E+ or E-)	±200V
Rated-Performance Temperature Range	-55°C to +150°C
Storage Temperature Range	-65°C to +175°C
Lead Temperature (Soldering, 10s)	+300°C
<b>POWER SUPPLY</b>	
Operating Voltage Range	+4V to +30V
<b>OUTPUT</b>	
Nominal Current Output @ +25°C (298.2K)	298.2μA
Nominal Temperature Coefficient	1μA/°C
Calibration Error <sup>1</sup> @ +25°C, max	±5.0°C/±2.0°C/±1.0°C
Absolute Error <sup>1</sup> (-55°C to +150°C)	
Without External Calibration Adjustment, max	±9.0°C/±3.8°C/±2.4°C
With +25°C Calibration Error Set to Zero, max	±2.0°C/±1.0°C/±1.0°C
Nonlinearity <sup>1</sup> (-55°C to +150°C), max	±2.0°C/±0.5°C/±0.5°C
Current Noise	40pA/√Hz
Power Supply Rejection	
+4V < $V_S$ < +5V	0.5μA/V
+5V < $V_S$ < +15V	0.2μA/V
+15V < $V_S$ < +30V	0.1μA/V
Case Isolation to Either Lead	10 <sup>10</sup> Ω
Effective Shunt Capacitance	100pF
Electrical Turn-On Time <sup>2</sup>	20μs
Reverse Bias Leakage Current <sup>3</sup> (Reverse Voltage = 10V)	10pA
<b>PRICE</b>	
(1-24)	\$3.00/\$6.00/\$12.00
(25-99)	\$2.40/\$5.00/\$10.00
(100-999)	\$1.95/\$3.95/\$7.95
<b>PRICE -/883 Versions<sup>4</sup></b>	
(1-24)	\$7.50/\$10.95/\$18.00
(25-99)	\$6.40/\$8.50/\$15.00
(100-999)	\$4.95/\$6.90/\$11.50

### NOTES:

<sup>1</sup> See data sheet for complete explanation of errors.

<sup>2</sup> Does not include self-heating effects; see data sheet, page 5, for explanation of these effects.

<sup>3</sup> Leakage current doubles every 10°C.

<sup>4</sup> The AD590J, K and L are available with full processing of MIL-STD-883A Class B. A complete list of procedures is available on request.



# COMPLETE MONOLITHIC 10-BIT A/D CONVERTER

## Successive-Approximation AD571 Contains All Required Components

### Technologies Include I<sup>2</sup>L, Laser Wafer-Trim, Buried Zener

by A. Paul Brokaw

The AD571\* is a complete 10-bit successive-approximation a/d converter. It includes the components required for 25 $\mu$ s (30 $\mu$ s max) conversion to within  $\pm 1/2$ LSB, including DAC, voltage reference, clock, comparator, successive-approximation register, and three-state output buffer, all on a 120 x 151mil<sup>2</sup> (3 x 3.8mm<sup>2</sup>) chip. The first converter of its kind, it employs bipolar technology to combine linear and high-density digital circuitry on a single manufacturable chip. Key factors in the successful production of the AD571 are I<sup>2</sup>L (Integrated Injection Logic), automatic dynamic laser wafer-trimming of stable thin-film on-chip resistors, and a stable buried-Zener reference. It is a low-cost device, \$24 in 100's for the AD571J.

### EASY TO USE

Figure 1 is an application diagram for the AD571, showing the signal source, plus the connections for the two power-supply voltages and the CONVERT signal. That's all that's needed for operation. Note that there are no "clock" or "reference" terminals. The AD571 has its own internal reference and its own internal clock. It is a successive-approximation converter, so no external integrating capacitors or resistors are called for. And choice of the bipolar input range requires only simple pin-strapping of a control input.

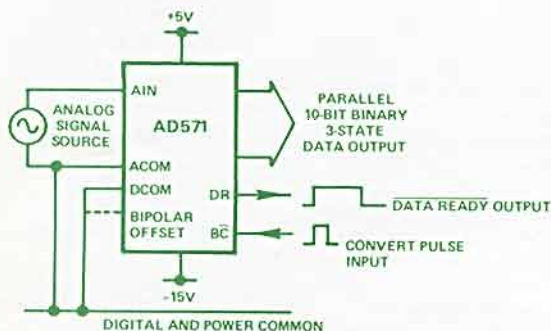


Figure 1. Typical AD571 application, showing power, signal, and control connections.

The self-contained buried Zener reference not only simplifies the use of the AD571; it also ensures excellent long-term stability. The subsurface breakdown of the buried Zener means that the junction is free from hot-carrier trapping and the effects of charge in the passivating oxide. Successive approximations means that the AD571 can be fast (25 $\mu$ s), and the internal clock frees the AD571 from problems with clock-line noise immunity or requirements to synchronize clock and CONVERT signals.

The control input, labeled  $\overline{BC}$  ("BLANK AND  $\overline{CONVERT}$ "), when high, "blanks" the 3-state output lines and inhibits conversion; it also keeps the DATA-READY ( $\overline{DR}$ ), or status, line high. When  $\overline{BC}$  goes low, a conversion is initiated, and the  $\overline{DR}$  line remains high; when the conversion is complete, data appears on the output lines, and  $\overline{DR}$  goes low. Figure 2 shows two modes of operation. In the "normal", or convert-pulse

mode, a short positive pulse clears the registers of the AD571 and starts a new conversion on the trailing edge. In the multiplex mode, the  $\overline{BC}$  line is held high, to blank the 3-state outputs, except when data is requested to be read out; then, a "0" on  $\overline{BC}$  will cause the AD571 to convert and to output 10-bit data at the end of conversion.

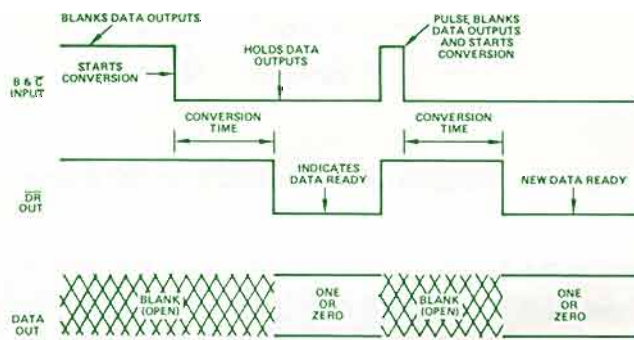


Figure 2. AD571 control and output format.

### WHAT'S INSIDE?

Figure 3 shows the block diagram of the complete AD571. It is organized along conventional lines, with a current-output DAC (digital-to-analog converter) developing a voltage across the input resistor. The analog comparator detects whether this voltage is greater or less than the applied analog input signal and controls the response of the successive-approximation register (SAR). The register operates the DAC to test the cumulative effect of the binary-weighted currents, one at a time.

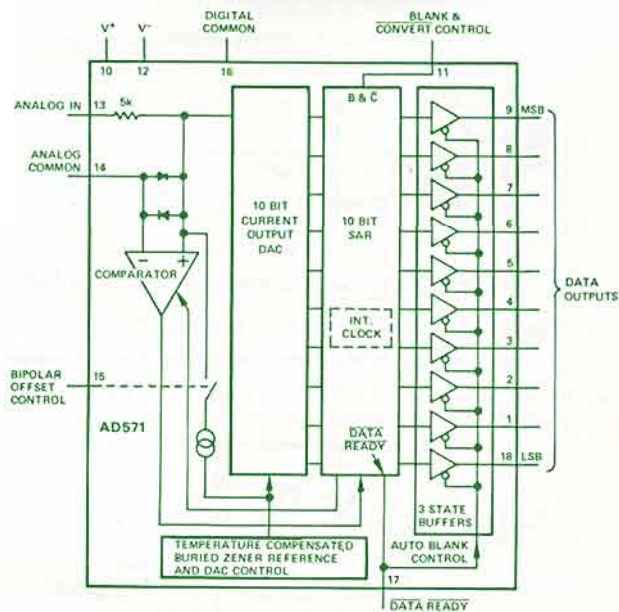


Figure 3. AD571 block diagram.

As each bit is tested, starting with the most-significant bit, its weight is retained as part of a linear sum of currents if the comparator indicates that this sum is less than the total weight of the analog input signal. Bit-weights which cause the current

\*Use the reply card for complete data on the AD571.

sum to exceed a value of current, determined by the ratio of the input voltage to the input resistance, are rejected and removed from the current sum.

After all 10 bits have been tested, the sum will be within one least-significant bit (1 LSB) of the analog input signal weight. The positions of those bits which form the sum indicate the binary code, or digital number, corresponding to the analog input. The one-zero pattern of these bits drives the three-state output buffers, and the DATA READY output signals that the conversion cycle has been completed. This signal also activates the 3-state buffers, so that the result of the conversion appears on the 10 parallel-binary output lines.

### WHY (AND WHAT IS) I<sup>2</sup>L?

Figure 4 shows a generalized flow chart of a successive-approximations algorithm. Even this simplified flow chart shows the necessity to set up tests, make decisions, and keep track of the location of the current operation in an n-step sequence of events. It is this logical complexity that has, in the past, made difficult the development of a complete, monolithic ADC.

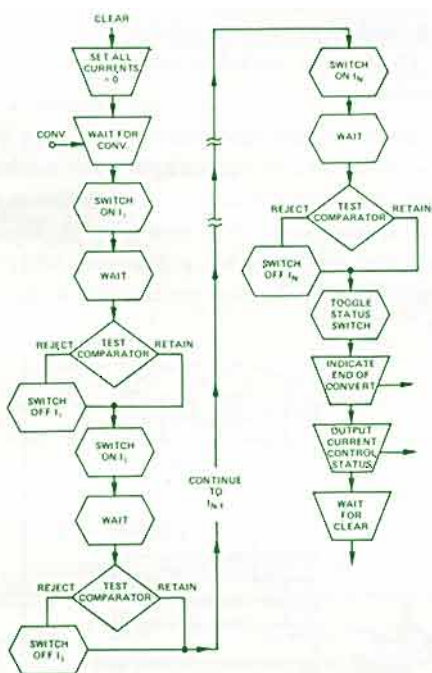


Figure 4. Successive-approximation flow chart.

The process used for the best (stable, high-resolution) linear circuits produces junction-isolated bipolar transistors that are large compared to bipolar logic transistors or MOS devices. Besides being large, they are poorly suited to conventional logic circuits. For these reasons, it is impractical to design a SAR that uses linear-bipolar-processed transistors. Most of the available monolithic ADC's (e.g., the AD7570 and AD7550) rely on MOS (metal-oxide semiconductor) processing, which makes possible efficient logic designs for the SAR or counters. Unfortunately, most MOS processes are not well-suited to the analog requirements of the complete a/d converter-on-a-chip. The key to success is high-density integrated injection logic, which can be produced by a bipolar process similar to that used for linear circuitry and implemented on the same chip.

Figure 5 shows a section through a conventional junction-isolated NPN transistor. The active region of the transistor is in the small area directly under the emitter diffusion. The sub-

stantial remainder of the area of the transistor is associated with the isolation and spacing required to prevent voltage breakdown. The vertical NPN structure can be operated in the inverted mode, with the emitter and the collector interchanged. This mode does not result in high-performance linear operation, but it can be made quite adequate for saturated-logic operation. Furthermore, multiple N+ diffusions into a single base-region allow the single emitter at the bottom to serve several inverted-mode transistors, under the control of a single base.

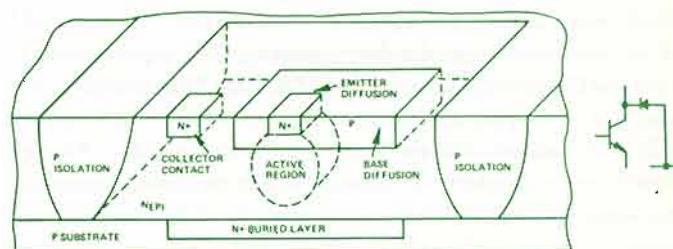


Figure 5. Junction-isolated NPN transistor.

Figure 6 shows several multiple-collector inverted-mode transistors. These transistors share the common N emitter region and need not be individually isolated. They are spaced only as far apart as is needed to permit separate base control. A P-type diffusion running between the rows of NPN transistors can be biased so that it injects minority carriers into the N epitaxial region. These carriers are collected by the NPN base diffusions, which act as the multiple collectors of a PNP transistor, as the drawing of the equivalent circuit shows in schematic form.

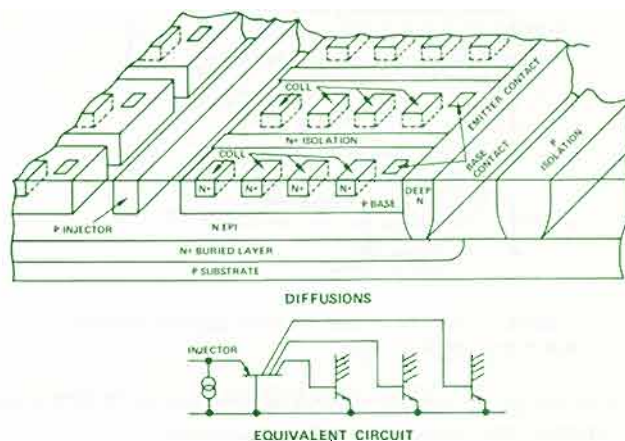


Figure 6. Junction-isolated I<sup>2</sup>L gates and injectors.

When a current is driven into the injector, it activates all the NPN transistors in the array. This dense array of injected transistors can be interconnected to perform logic functions; hence the term, *integrated injection logic*. An important feature of the I<sup>2</sup>L made in this way is that we can use the same diffusions to make conventional diffusion-isolated transistors, optimized for linear performance, on the same chip.

I<sup>2</sup>L has an important influence on cost. A commercial SAR chip area is 120 x 151 mil<sup>2</sup>, including *everything*: the 10-bit attempted with linear transistors, it would be 3 or 4 times bigger, yields would be correspondingly 10 to 20 times lower, and costs similarly higher. In the case of the AD571, the whole chip area is 120 x 151 mil<sup>2</sup>, including *everything*: the 10-bit DAC, including its laser wafer-trimmed thin-film resistors, the LWT voltage-reference circuitry, and the high-level logic inter-



faces are all made with  $I^2L$  — all on the same chip. It is interesting to note that the AD571 is less than 1/3 again as large as the AD561 10-bit DAC (which is similar to the AD571's DAC).

## PUTTING THE AD571 TO WORK

As noted earlier, the AD571 is simple to use. Just apply power and the analog input, and strobe  $\overline{B\overline{C}}$  for 10-bit binary data. Nominal supply voltages are +5V and -15V. The basic converter is affected but little by changes in the positive supply, which may be +5V, +10V, +12V, or even +15V. The threshold of  $\overline{B\overline{C}}$  adapts to applied voltage, meeting TTL requirements at +5V and CMOS requirements at +15V. The digital outputs also adapt to the supply level, maintaining a "1" level at about 2V below the supply voltage, and "0" at about 0.4V. Though slightly more sensitive to changes in the negative supply, the converter operates reliably from as little as -12V.

Analog and digital common connections are brought out separately. While they may be joined right at the AD571, it is possible to join them remotely, for convenience in system grounding. Operation is normal with as much as  $\pm 200\text{mV}$  between ACOM and DCOM, and no damage will occur below  $\pm 1\text{V}$ . In systems involving remote grounds, where it is possible that opening a connector or removing a card elsewhere will float the commons or cause them to be forced apart, we suggest that safety diodes be installed locally between the commons and/or power common (Figure 7).

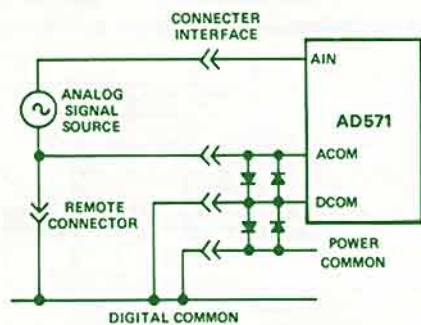


Figure 7. Safety diodes protect against remote open-ground circuits.

Most of the power-supply current is returned to DCOM inside the AD571. The current in ACOM (see Figure 3) comprises the comparator input current and the residual current at the summing node during the conversion process. This current, conveyed by the clamp diodes, may switch in the range 0 to 2mA during conversion. In addition, in the unipolar mode (BIPOLAR OFFSET pin grounded), ACOM will sink about 2mA continuously, returned to the  $V^-$  terminal via the DAC. This current may be used to offset the analog range, as the data sheet explains more fully. To permit transient currents to flow freely in the clamp diodes, the offset network should not be capacitively bypassed.

Figure 8 illustrates one strategy for grounding and bypassing the AD571, where necessary. C15 bypasses the -15V supply locally. A small series resistor may be used to break up supply-line resonances in large systems. The -15V return should ideally connect directly to digital common; however, it is often associated with the analog signal source and returned to the AD571 by way of the analog-digital link, as shown. The positive supply

can be locally decoupled, to minimize noise on a + supply at the converter, an important consideration in bipolar applications, since the BIP OFF current is derived from the positive supply.

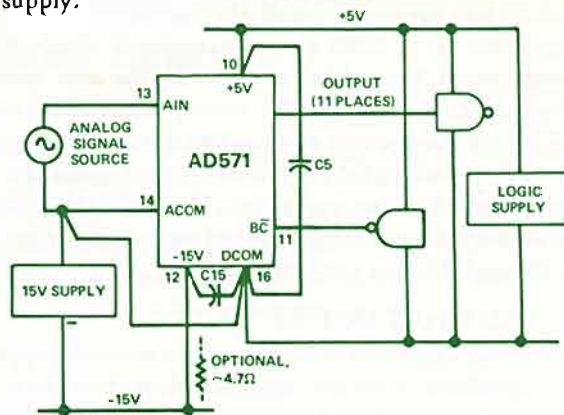


Figure 8. Decoupling the AD571.

The length of the analog signal-and-return path should be minimized, and the analog signal source should be at low impedance at high frequencies. ANALOG DIALOGUE 11-2 (1977) pp. 10-16 is a useful reference to grounding and decoupling techniques.

The AD571 can be used in most circuits requiring 10-bit-accurate a/d conversion and can replace hybrids or modules in such designs. It also has several features which allow it to be used in innovative applications. For example, in Figure 9, two analog signals are measured by a pair of AD571's, and the results are digitally time-division multiplexed under the control of a flip-flop.

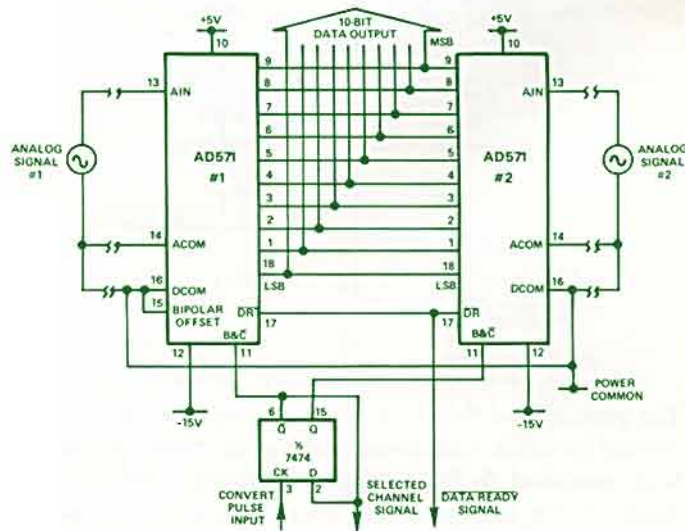


Figure 9. Multiplexed pair of AD571's.

Each CONVERT pulse alternates the state of the flip-flop. For example, Q may go high to blank the output of channel 2, while  $\overline{Q}$  goes low to initiate a conversion on channel 1. At the end of conversion, DR goes low to indicate that data is ready; and  $\overline{Q}$  low indicates that channel 1 has been selected. The output bus is driven by the output of channel 1. On the next CONVERT pulse, the flip-flop output changes state, channel 1 is blanked,  $\overline{DR}$  goes high, and a conversion is initiated on channel 2.  $\overline{Q}$  high indicates that channel 2 has been selected, and, at the end of conversion,  $\overline{DR}$  goes low and the bus is driven by the output of channel 2. ▶▶▶



# 12-BIT CMOS MULTIPLYING D/A CONVERTER

## AD7541 Is Capable of Low-Feedthrough 4-Quadrant Operation

### Monolithic IC is Laser Wafer-Trimmed for 12-Bit Linearity

by John Newell

The AD7541\* is a low-cost 12-bit CMOS multiplying digital-to-analog converter housed in a standard 18-pin dual in-line package. Pin-compatible with the earlier AD7521<sup>1</sup>, the AD7541 is the first such device to be laser-trimmed at the wafer stage (LWT) for less than 0.01% conversion nonlinearity (the AD7541K/B/T); its thin-film resistors impart comparable linearity in response to positive or negative analog ("reference") signals. Gain error is trimmed to within  $\pm 0.3\%$ . Price of the AD7541KN is only \$30, 1-24, and \$12 in 1000's.

The uses of the AD7541 are many, since it can be considered as either a wide-range variable-reference DAC or, alternatively, as a digitally controlled attenuator with a nominal full-scale gain of 0.9998, and settling time less than  $1\mu\text{s}$  following a gain change. The AD7541 is always used with an external operational amplifier, which may be chosen to emphasize the salient aspects of desired overall performance (e.g., cost, speed, stability, etc.) Figure 1 shows the simplest mode of connection.

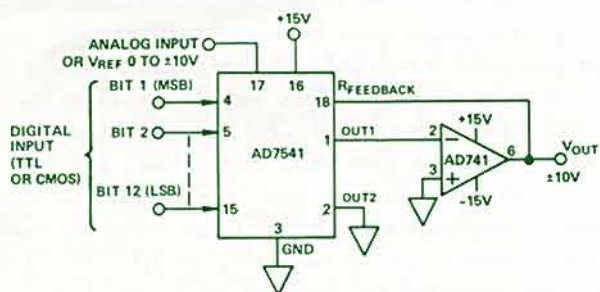


Figure 1. Simple application of the AD7541.

#### INSIDE THE AD7541

Like its predecessors, the AD7541 circuit consists of a thin-film R-2R ladder network and a set of CMOS switches (Figure 2). In response to TTL or CMOS logic signals, the switches steer the fixed, binary-weighted currents, developed in the R-2R ladder, either to the OUT1 line (digital input *high*), where they are summed to obtain the feedback current of the op amp, or to the OUT2 line (digital input *low*), which is grounded in unipolar (digital) applications or summed in an inverting amplifier in bipolar ("4-quadrant multiplying") applications. The on-chip feedback resistor,  $R_F$ , transduces the  $I_{OUT1}$  current to a voltage, which appears at the output with inverted polarity. For  $V_{REF}$  positive, negative, or zero, the output will be negative, positive, or zero.

Though the converter is simple in concept and easy to use, careful attention has been paid to a number of factors to

\*Use the reply card to obtain technical data on devices mentioned here.  
<sup>1</sup> Converters of this type are well-documented in the Analog Devices literature. Besides the data sheets, a few useful references to it and related types are:

- ANALOG-DIGITAL CONVERSION NOTES (246-page book, \$5.95)
- ANALOG DIALOGUE: 8-1 (the AD7520); 8-2 (AD7520 applications); 9-1 (AD7521); 9-2 (AD7570 successive-approximation ADC using AD7520 DAC circuit); 9-3 ( $\mu\text{P}$ -compatible 10-bit double-buffered version, AD7522); 10-1 (audio applications for CMOS DAC's); 11-2 (good practice for using CMOS DAC's with op amps).

obtain 12-bit useful resolution and comparable accuracy, over a wide range of temperature.

The key to success is to graduate the *on* resistance of the switches (in either position) in inverse proportion to the currents flowing through them, so as to make the switch drops equal. For example, in the MSB, Bit 2, and Bit 3, the switch resistances are typically  $15\Omega$ ,  $30\Omega$ ,  $60\Omega$ , and for bit-currents of  $0.5\text{mA}$ ,  $0.25\text{mA}$ , and  $0.125\text{mA}$ , the corresponding voltage drops are all equal to  $7.5\text{mV}$ . Thus, the switch error effectively appears in series with the reference; this causes a gain shift, which can be corrected elsewhere, but does not degrade the linearity. In the AD7541, weighted-resistance switches are used for bits 1 through 9, additional series resistance is used for bits 10-12.

Since the ratio error in thin-film networks can seldom be maintained at better than 0.05% using standard photo-lithographic techniques, the exactness of the ratio in the R-2R ladder network is obtained by laser-trimming at the wafer-probe stage, an automated technique pioneered by Analog Devices and used successfully on a number of our products (some in this issue).

Once the ratio is set, the resistors track to within  $1\text{ppm}/^\circ\text{C}$ . The switches track to within  $100\text{ppm}/^\circ\text{C}$ ; however, since the ratio of the effective resistance in series with each switch to its  $R_{ON}$  is approximately 1000:1, the switch mismatching contributes only  $0.1\text{ppm}/^\circ\text{C}$  to the linearity temperature-coefficient. This accounts for the excellent nonlinearity tempco, which is tested on all extended-temperature-range units.

Because the switches are in series with the ladder, the gain tempco of the tracking of  $R_F$  and the network resistance (about  $1\text{ppm}/^\circ\text{C}$ ) is augmented by the  $0.5\%/^\circ\text{C}$  switch-resistance tempco, multiplied by  $R_{ON}/2R$ , or about  $4\text{ppm}/^\circ\text{C}$ .

Other error-generating factors that had to be dealt with in the design included resistance of metallization, parasitic diode leakages, and the tendency for large voltage outputs to increase the dissipation of  $R_F$  (which would tend to produce a thermally generated integral linearity error, if the resistance of  $R_F$  were to change substantially; in the AD7541, this error-contribution is less than  $1/4\text{LSB}$  at  $10\text{V}$  output). ▶▶▶

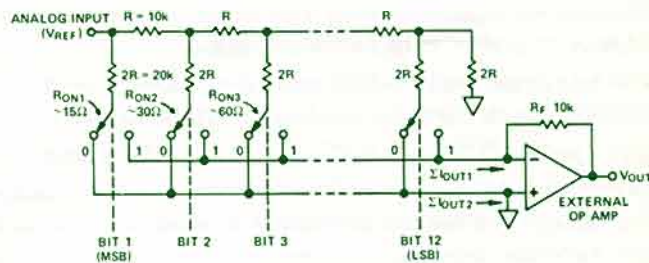


Figure 2. Functional schematic of the AD7541.

# NEW REAL-TIME INTERFACE BOARDS FOR $\mu$ C's

## A First: Six I/O Boards for the 16-Bit TM-990/100M. Plus, A Cost-Effective Input Board for 8080-Based Microcomputers

by Russ VerNooy

Analog Devices RTI "Real-Time Interfaces" are complete, cost-effective analog input-output systems that are physically, electrically, functionally, and software-compatible with the most-popular microcomputers.

If you were a reader of *Analog Dialogue* 11-1, you encountered the first of the series, the RTI-1200,\* an input/output board designed for direct interfacing with the SBC 80/10 and similar systems. In the most-recent issue, 11-2, you became acquainted with the RTI-1201\* 4-channel output board, designed for SBC 80 compatibility, and a pair of Pro-Log-compatible input and output boards, the RTI-1220\* and the RTI-1221.\*

In the present pages, your acquaintance is broadened to include the first analog I/O boards specifically designed for a 16-bit microcomputer, the Texas Instruments TM-990/100M, plus a new analog input board for 8080-based microcomputers.

### INTERFACING 16-BIT COMPUTERS

The 8080 instruction set, RTI features, and the hand-in-glove memory-mapped interface, make it relatively easy to interface an 8-bit microcomputer to analog data having useful resolutions of 10-, 12-bits, and more. Unfortunately, that's only part of the problem. Whenever the data must be manipulated or involved in calculations, time-gobbling double-precision operation are necessary, greatly reducing the overall system throughput rate. It makes sense, in this day and age, to use a 16-bit computer to deal with 12-bit data — if one is available. The TM990/100M could very well be the answer to such needs; not only does it handle 16-bit data directly, it is also less costly than some of the 8-bit competition.

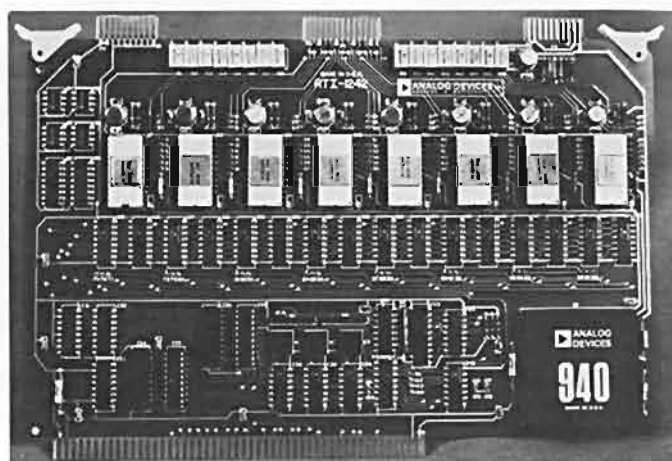
These were all quite cogent reasons for Analog Devices to decide to offer a set of input/output "RTI" boards to match the TM990/100M. The present products include:

- The RTI-1240\* 12-bit Analog-Input Subsystem, with its 16 input channels, expandable on-board to 32, and off-board (using on-board control logic) to 256 channels. Available in two forms, "R" for resistor-programmed gain (1 to 1000) in the instrumentation amplifier, and "S" for software-programmed gain (1,2,4,8) to obtain up to 15 bits of dynamic range, on a channel-by-channel basis.
- The RTI-1241\* Analog-I/O Subsystem, having the same features as the RTI-1240\*, plus two twelve-bit DAC's.
- The RTI-1242\* and RTI-1243\* Analog Output Subsystems, having 4- and 8-channel 12-bit DAC outputs, with some unusual properties, to be discussed below.

The boards are available from many distributors in the U.S.A., who also handle the Texas Instruments microcomputers.

**RTI-1240 Analog-Input Subsystem** Besides the flexibility of channel number and gain choice, and the memory-managed interfacing (with flexible addressing) to allow effective use of the TMS9900 instruction set for high throughput rates and

\*For data sheets on Analog Devices products mentioned here, use the reply card.



versatility, there are a number of software and hardware features that deserve specific comment. For example, the RTI-1240 allows the user to choose between simple software and high throughput rates. In the CPU HOLD mode, only a single instruction is required to initiate a conversion and cause the data to be read about 25 $\mu$ s later. On the other hand, in the polled-status mode, the CPU can be programmed to initiate a conversion, then perform other tasks, and return to poll the converter's status or read the a/d data. Such flexible software-controlled features impart a high degree of versatility to microcomputer systems employing the RTI-1240. Analog inputs are protected for faults of up to 20V beyond the supplies (as in all RTI's).

External *convert* commands may be applied, and the *convert* command may increment the channel (multiplexer) automatically, if desired. Status words provide information on conver-

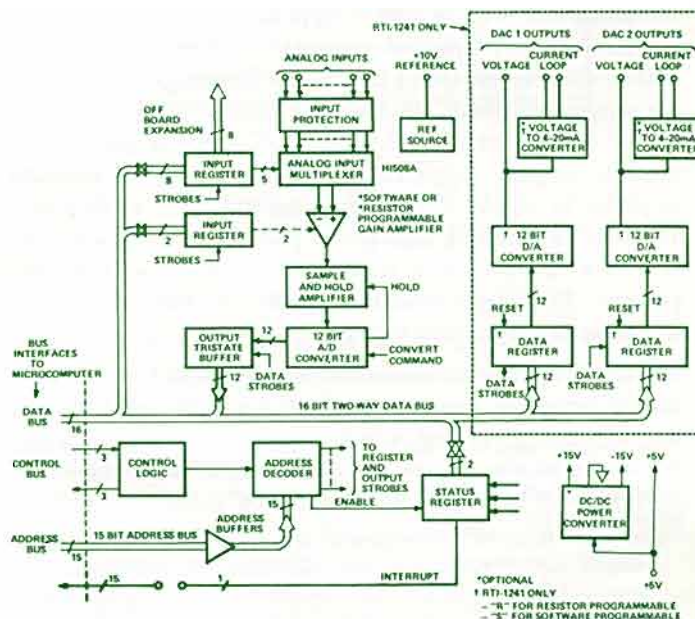


Figure 1. RTI-1240 & RTI-1241 functional block diagram.



sion status, underrange (in the programmable-gain amplifier in "S" versions), and channel- and amplifier-gain. The analog input may be either current (4-20mA current loop) or voltage, and either single-ended or differential. A +5V power option, for operating the board from computer power, is available at small additional cost. Figure 1 shows the block diagram of the RTI-1240 and the RTI-1241. Price of the RTI-1240-R is \$445 (1-9); the RTI-1240-S is \$495.

**RTI-1241 Analog Input/Output Subsystem** has all the features of the RTI-1240 (R or S), plus two 12-bit d/a converters. Optional 4-20mA current-loop outputs are available, and the DAC ranges and coding are independently settable. Though an on-board reference source is provided, an off-board common reference system may be used. The DAC's can be reset to facilitate initialization or for quick shutdown. Price of the basic RTI-1241-R is \$560; the RTI-1241S is \$610.

**RTI-1242 and RTI-1243 Output Subsystems** There's a reason we have not called this an "analog output" subsystem: in addition to the 4 channels (or 8 channels - RTI-1243) of 12-bit analog DAC output, there are also 8 independent channels of software-controlled 1-bit digital output, capable of sinking 300mA and withstanding +30V of supply voltage, for use in on-off control functions. These open-collector logic drivers can be used for relays, solenoids, or valve controllers, or as a high-current 8-bit digital-output port.

The DAC's may utilize the on-board reference or external references. Each DAC has independent range and code selection, with remote load-sense capability to make the output accuracy independent of line drops. The RTI-1242 and RTI-1243 can be equipped with data readback by the user so that it is unnecessary to use the computer to retrieve the data that was latched into a given output channel. Also, the DAC's and the logic drivers may be reset, as in the case of the RTI-1241. Figure 2 shows a block diagram of the RTI-1242 and RTI-1243. Prices are \$395 and \$675 for the basic 4- and 8-channel models.

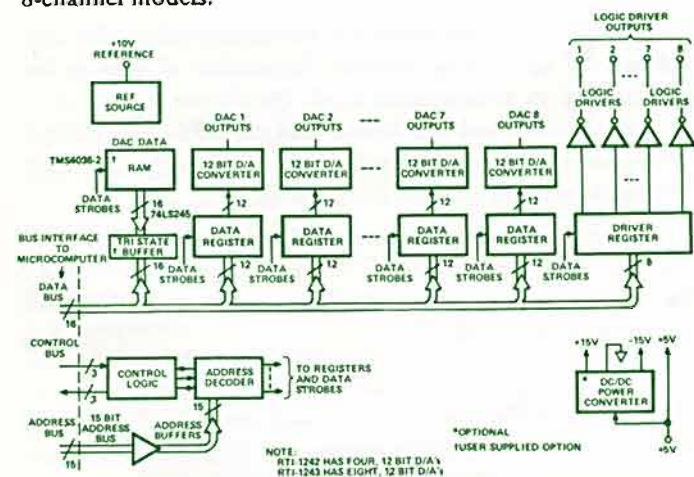


Figure 2. RTI-1242 & RTI-1243 functional block diagram.

## A NEW BOARD FOR 8080-BASED COMPUTERS

In addition to providing easy-to-use I/O boards for promising new microcomputers, we continually add to our line of useful interface products for existing popular  $\mu C$ 's. The RTI-1202\*, a complete analog input subsystem, is electrically and mechanically compatible with the Intel SBC 80 series of single-board computers. It is also compatible with our RTI-1200 and -1201

interfaces, the MDS microcomputer development system, the National Semiconductor BLC 80 board-level computer, and other MULTIBUS-compatible computers.

The RTI-1202's basic function is data acquisition for a  $\mu C$  system (Figure 3). This is accomplished with protected-input multiplexers, a selectable-gain instrumentation amplifier, a sample-hold amplifier, and either an 8-bit or a 12-bit ADC.

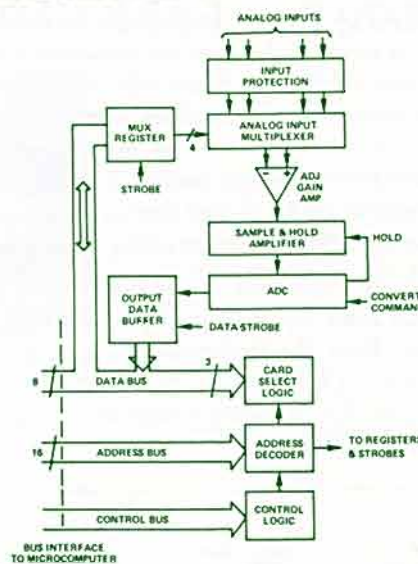


Figure 3. RTI-1202 functional block diagram.

The basic RTI-1202 will accept up to 16 single-ended or pseudo-differential inputs, or 8 differential inputs, fully protected to  $\pm 20V$  beyond the supply rails. A multiplexer expansion option doubles the number of channels on-board, and on-board registers exist for off-board expansion to 256 channels.

The instrumentation amplifier can be resistor-programmed for any gain from 1 to 1000, to provide preamplification or normalization for input signals. The sample-hold amplifier goes into hold when a convert command (software or external) is received. At the end of conversion (EOC), the S/H returns to sample (track). A wirewrap jumper may be used to allow convert commands to increment the multiplexer register to the next channel for autoscan operation.

In normal memory-mapped operation, the CPU continues with operation and data-handling during each conversion. The EOC signal sets the EOC bit in the status word and can also be jumpered to cause one of eight levels of interrupt.

Four methods are available on the RTI-1202 to ensure that valid ADC data will be read: CPU hold (jumper option), EOC interrupt, looping to check the EOC bit status, and simply providing sufficient time in the program for conversion to be complete before reading the converter output.

The 12-bit-ADC version has two simultaneously available output formats, 12 bits in two bytes (4 MSB, 8 LSB, right-justified) and 8-most-significant-bits in a single byte. The ADC can be jumpered for binary, 2's-complement, or offset-binary coding for unipolar or bipolar analog input signals.

The RTI-1202 has an on-board socket for user selection of the 3 most-significant hex digits of the address, allowing flexible location of the RTI's 8 words throughout the entire 65K memory map of the 8080-based system. Price of the basic RTI for 12-bit conversion is \$399. ▶▶▶



# LOW-DRIFT SUPERBETA OP AMP

## The AD517L Has 25 $\mu$ V Offset and 0.5 $\mu$ V/ $^{\circ}$ C Drift

### Max Bias Current is Less than 1nA Initial, 1.5nA Over Temperature

by Doug Grant

The AD517\* is a high-accuracy internally compensated monolithic operational amplifier designed for applications calling for low drift, low bias and offset currents, high open-loop gain and common-mode rejection, and simplicity of use. Typical applications include high-precision voltage references and wide-range voltage-measuring circuits, output circuits for high-resolution current-output DAC's, bridge-instrumentation circuitry, voltage followers, and high-accuracy analog computing circuits.

The AD517L is laser-trimmed at the wafer level for offset voltage less than 25 $\mu$ V and drift less than 0.5 $\mu$ V/ $^{\circ}$ C, with no external nulling. Even the lowest-cost version, the AD517J, has a maximum input offset of less than  $\pm 30$ ppm of full-scale output range over the operating temperature range without external trimming (Figure 1).

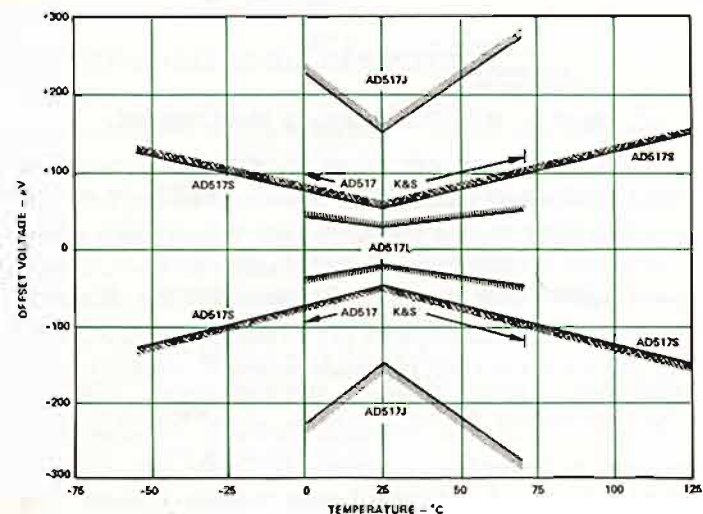
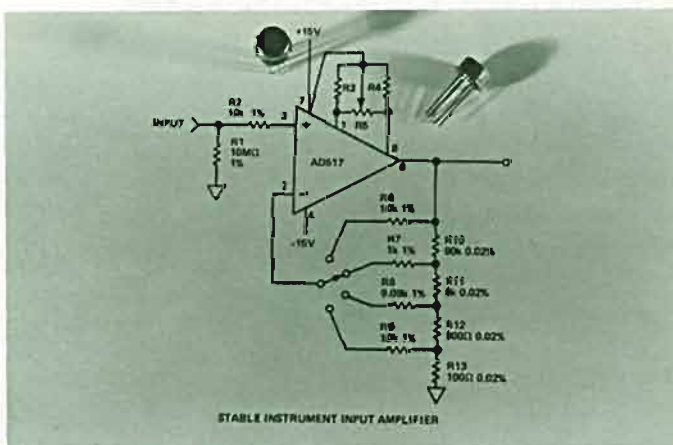


Figure 1. Limits of offset for the AD517.

Super- $\beta$  input transistors provide bias currents as low as 1nA (unipolar) maximum and offset currents as low as  $\pm 0.25$ nA max. While the figures for initial bias current are greater than those for popular FET-input amplifiers at room temperature, the input currents of AD517 decrease with increasing temperature, while FET bias currents double approximately every 11 $^{\circ}$ C. In Figure 2, which compares the AD517 with a typical FET-input amplifier having 50pA bias current, note that FET bias current increases sharply at the higher temperatures, yet the sensitivities of both types to temperature at 25 $^{\circ}$ C are comparable.

Unlike some other precision monolithic amplifiers currently on the market, the low value of initial bias current in AD517 is achieved without internal current nulling. This has two advantages: first, the bias current is always of the same polarity and may be nulled out by a simple unipolar adjustable current "leak"; second, by matching of the parallel source resistance connected to the two inputs, the net current-related error will

\*For complete data on the AD517, use the reply card.



be that due to the considerably smaller offset current, an improvement by a factor of four, tracking over the temperature range.

Gain of the AD517 is  $10^6$  minimum at rated load (5mA), which helps the overall linearity at high closed-loop gain. Common-mode rejection of the AD517(K/L/S) is 110dB minimum ( $V_{CM} = \pm 10$ V), a useful factor in voltage-follower and differential circuit applications. The thermally balanced chip ensures high open-loop gain that is independent of thermal gradients induced by output loading, offset nulling, and power-supply variations.

Although voltage noise (typically 2 $\mu$ V p-p, 0.1Hz to 10Hz) is somewhat higher than for the AD510 and the AD504, current noise is less, and the voltage uncertainty (drift and noise) is comparable.

The input stage is fully protected, permitting differential input voltages of up to  $\pm V_S$ , without degradation of gain or bias current due to reverse breakdown. The output stage is short-circuit protected and can drive a load capacitance of 1000pF. Since internal compensation is provided, the AD517 is basically stable without external components.

Housed in the hermetically sealed TO-99 metal can, the J/K/L versions are provided for 0 $^{\circ}$  to 70 $^{\circ}$ C operation, and the S version is for -55 $^{\circ}$ C to +125 $^{\circ}$ C. Prices are \$4.50/\$7.50/\$12.50/\$18.00 in 100's. ▶▶▶

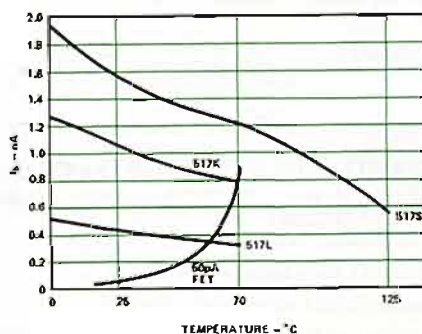


Figure 2. Typical input bias current vs. temperature.



## FAST 6-DECADE LOG-ANTILOG AMPLIFIERS

Models 759N/P Compute  $\text{Exp}_{10}(V)$  or  $\text{Log}(I \text{ or } V)$ , Either Polarity

\$22 (100's), Device is Complete, Small (1.1" x 0.4"), Fast (200kHz@1μA)

Models 759N and 759P\* are the industry's smallest, lowest-cost complete dc log-arithmic amplifiers with pretrimmed internal reference, scale factor, and temperature compensation. They will convert linear current or voltage to logarithmic output voltage, or linear input voltage to exponential output voltage (antilog). Currents can be "logged" over a 6-decade range (1nA to 1mA), voltages are "logged" (input) or "anti-logged" (output) over a 4-decade range, 1mV to 10V.

Log conformity is to within 2% over five decades of current (10nA to 1mA) and all four decades of voltage, and to within 1% over four decades of current (20nA to 200μA) and 3 1/3 decades of voltage (1mV to 2V), as Figure 1 shows (for current).

Three output scale factors,  $K = 2, 1, 2/3$  V/decade, are available by simple pin selection, providing the relationships shown (for current) in Figure 2. The equation describing ideal log performance, for current inputs, is:<sup>1</sup>

$$E_o = -K \log_{10} \left[ \frac{I_{SIG}}{I_{REF}} \right]$$

For voltage inputs, the ideal logarithmic relationship is:

$$E_o = -K \log_{10} \left[ \frac{E_{SIG}}{E_{REF}} \right]$$

and for antilogarithms (exponentials), the ideal relationship is:

$$E_o = E_{REF} \exp_{10} (-E_{SIG}/K)$$

### APPLICATIONS

Log amplifiers in log mode are useful for applications requiring compression of wide-range analog input data, linearization of transducers having exponential outputs, and analog computing, ranging



from simple translation of natural relationships in log form (e.g., computing absorbance as the log-ratio of input currents) to the use of logarithms in facilitating analog computation of terms involving arbitrary exponents (powers and roots) and multi-term products and ratios.

Antilog amplifiers are useful for applications requiring expansion of compressed data, linearization of transducers having logarithmic outputs, analog function fitting or function generation, obtaining relationships or generating curves having voltage-programmable rates of growth or decay, and in analog computing.

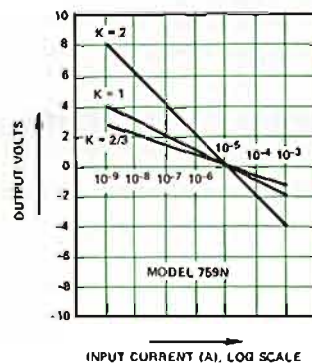


Figure 2. Output as a function of input current, log connection.

### EASE OF CONNECTION

Figure 3 is a block diagram of the 759N. A

log/antilog amplifier consists of an operational amplifier and an element with anti-logarithmic transconductance (i.e., the voltage into the element produces a current that is an exponential function of the voltage). As the figure shows, for logarithmic operations, the input current is applied at op-amp summing point, and the feedback circuit causes the amplifier output to produce whatever voltage is required to provide a feedback current that will exactly balance the input current.

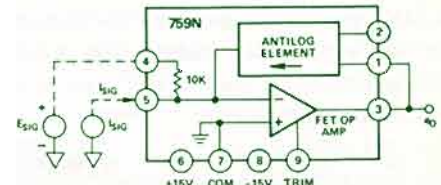


Figure 3. Functional block diagram of model 759N.

In antilog operation, the input voltage is applied directly to the  $K = 1$  or  $K = 2$  input of the antilog element (or both, for  $K = 2/3$ ), producing an exponential input current to the op-amp circuit. The feedback resistance transduces the current to an output voltage.

The 759 is available as a pretrimmed "P" or "N" option, depending on the polarity of the input signal to be "logged". Logarithms may be computed only for positive arguments, therefore the internal  $I_{REF}$  must be of appropriate polarity to make the ratio positive. "N" indicates that the input electron flow is negative, hence current or voltage is positive; "P" indicates that only negative voltage or current may be applied in the log mode. Either version will take the antilog of input signals of either polarity, but the N version provides positive output voltage, and the P negative output.

Price of these compact modules is only \$34 (1-24) and \$22 (100+).



<sup>1</sup>A useful guide to log/antilog specifications and applications (and many other uses of non-linearity) is the NONLINEAR CIRCUITS HANDBOOK, available from Analog Devices, Inc., Box 796, Norwood, MA 02062. @ \$5.95.

\*For technical data on Model 759, use the reply card.

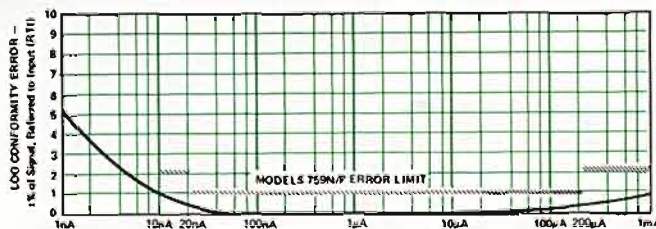
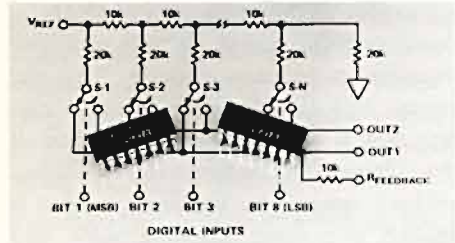


Figure 1. Log conformity error for models 759N and 759P.

## 8-BIT CMOS DAC

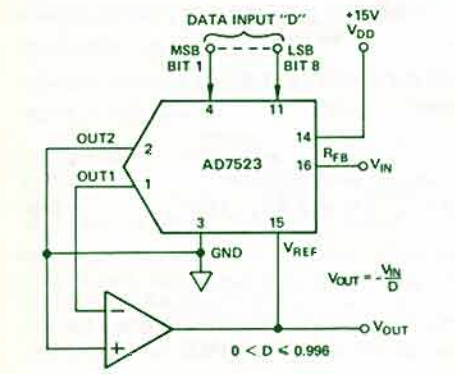
4-Quadrant Multiplying  
Lowest Cost (\$2 in 1000's)



The AD7523\* is an 8-bit CMOS multiplying digital-to-analog converter packaged in a 16-pin plastic DIP. It is compatible physically and in function with the AD7520 but is designed for lowest cost in applications calling for 8-bit resolution, CMOS logic levels, and low power consumption.

The AD7523 is guaranteed monotonic over its 0° to 70°C temperature range and is available in three versions (J/K/L), which provide 8-bit, 9-bit, and 10-bit linearity ( $\pm 1/2$ ,  $\pm 1/4$ ,  $\pm 1/8$  LSB) over the temperature range. The AD7523LN, for example, provides 256 monotonic output levels, each having a relative accuracy to within 0.1% of full-scale gain.

A two- or four-quadrant multiplying DAC, the AD7523 performs as a digitally controlled attenuator, and can provide either gain or attenuation, depending on its mode of connection to the external output op amp. Gain-settling time is 200ns max, and ac sine-wave feedthrough is  $\pm 1$  LSB max at 200kHz. The illustration shows its application in the feedback path for division. Small-quantity prices are \$4/\$9/\$15, dropping to \$2.50/\$5.60/\$9.35 in 100's.



\*For technical data, use the reply card.

## 3-DIGIT MONOLITHIC ADC CHIP

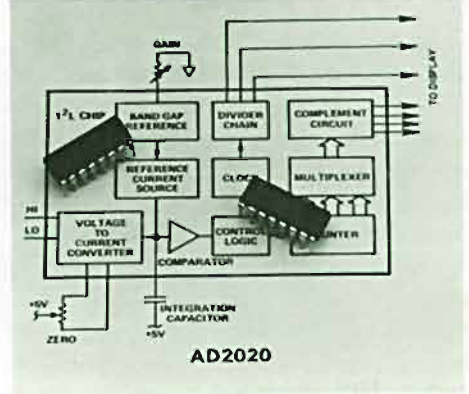
AD2020: Basic I<sup>2</sup>L Converter in AD2026 DPM  
Use it for Custom DPM's, APM Replacement

The AD2020\* is a low-cost 3-digit analog-to-digital converter that needs only 10 additional support components for a complete 3-digit meter. Its economy, simplicity, and reliability (proven by use in tens of thousands of AD2026 digital panel meters), and the variety of displays that it can drive, make it a "natural" as a basic building block of instrumentation modules, equipment, and systems, and for analog panel-meter replacement.

The AD2020, like the AD571 (page 6), utilizes integrated injection logic (I<sup>2</sup>L), to combine linear circuitry and high-density logic on a single bipolar chip. The input amplifier, band-gap reference, comparator, counters, clock, control logic, multiplexer, and drivers implement a complete dual-slope conversion on the chip. Only the integrating capacitor and any desired gain and zero adjustments are connected externally. In this flexible design, the decoder-drivers, displays, and switching transistors are furnished by the user to fit the display application.

Assembled in a 16-pin plastic DIP, the AD2020 draws only 50mW of power from a single +5V supply. Normally specified for 0° to 75°C, it is also available for extended temperature-range operation.

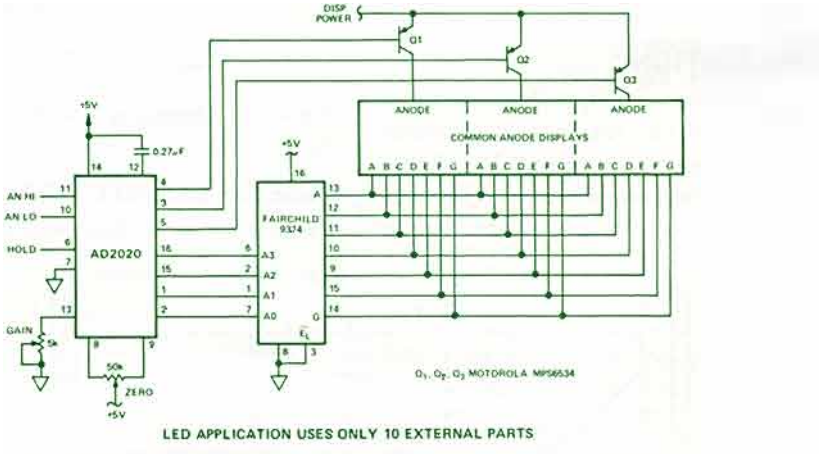
The AD2020 has differential inputs and a unique "sesqui-polar" input range:



+999mV to -99mV. This range is especially useful for analog panel-meter replacement, in temperature measurement, and in any unipolar application where a meaningful device- or system zero-adjustment range is essential.

Its small size permits it to be mounted in standard analog panel-meter cases, with display, for retrofitting APM's in instruments and power supplies.

Almost any kind of display can be used, including light-emitting diodes (LED's) — see Figure— liquid-crystal displays (LCD's), and the recently announced Hewlett-Packard ambient-viewing wide-temperature-range displays (HDSP-353-, 4030, etc.) The AD2020 is priced at \$13.50 in small quantity and \$9.00 in 100's.



LED APPLICATION USES ONLY 10 EXTERNAL PARTS

\*For data on the AD2020; the complete display-less module, the AD2023; or the complete DPM, AD2026, use the reply card.



## AD1408/AD1508 288 SYNCHRONIZED ISOLATION AMPLIFIER

### 8-Bit D/A Converters, \$28 (100's); Small (1"² X 0.56"); Linear (0.05% max NL)

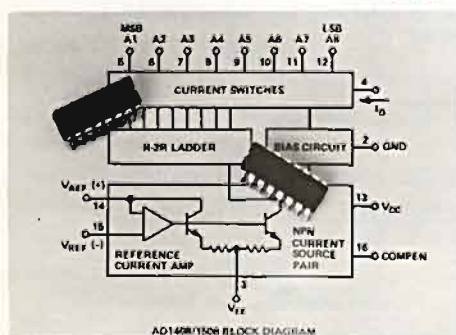
### Direct "MC" Replacements 947/948 Drivers Service up to 8 288's

The AD1408 and AD1508\* are low-cost monolithic integrated-circuit 8-bit digital-to-analog converters in the hermetically sealed ceramic 16-lead dual in-line package. Furnished as high-performance direct replacements for popular types bearing the same numerical designation, each device contains a set of matched bipolar switches, a precision resistor network, and a control amplifier.

The operating temperature range for the 1408 is 0° to 75°C, and the 1508 operates over the -55°C to +125°C range. There are three grades of the 1408, distinguished by relative accuracy (error relative to full-scale output current); the 1408-7/8/9 have maximum error of ±0.39%/±0.19%/±0.1% over the temperature range, and the 1508-8/9 have maximum error of ±0.19%/±0.1% over their rated temperature range. Prices in 100's range from \$3.10 for the AD1408-7D to \$15.95 for the AD1508-9D/883B.

In addition to the availability of improved accuracy, the AD1408/1508 have faster settling times, 250ns; this permits the performance of existing circuits employing the 1408/1508 to be upgraded. Lower power consumption of the Analog Devices types (typically 157mW) means improved stability and faster warmup.

The AD1408 and AD1508 are recommended for all low-cost 8-bit requirements; they are also suitable for upgrading overall performance where earlier, less-accurate, and slower 1408/1508 devices have been designed in. Chips are available for hybrid applications. ▶▶▶

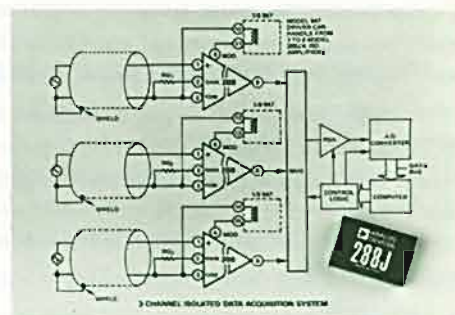


\*Use the reply card for technical data.

Model 288\* is a low-cost, transformer-isolated dc amplifier, with a synchronized design optimized for multi-channel industrial instrumentation and control applications. The 288 has single-resistor-adjustable gain of 1V/V to 1000V/V and maximum nonlinearity of 0.05%(K). Input drift is less than 5μV/°C. The protected front end has a maximum common-mode voltage rating of 850V dc, continuous, and high common-mode rejection (92dB minimum at 60Hz, with 1kΩ source unbalance).

The 288 will interrupt ground loops and leakage paths, and will protect circuitry against high-voltage transients. In low-level multi-channel data-acquisition systems, process-monitoring and computer-interface systems, the 288 offers complete galvanic isolation and protection against damage from fault voltages, within its ratings.

Two compatible drivers are available to



provide the required external modulator/demodulator drive signals for up to eight 288's, with provision for further expansion. The 947 has eight synchronized, isolated modulator drive outputs, providing both channel-to-channel isolation and input-to-output (and dc supply) isolation. The 948 affords the same degree of input-to-output isolation where the input channels share a common guard reference. Price of the 288J/K is \$39/\$49 (1-24), and the 8-channel 947/948 drivers are \$39/\$29. ▶▶▶

## SYNCHRO-TO-BCD CONVERTERS

### For Synchro/Resolver Data to 14 Bits

### Transformer-Coupled 60-2600Hz, Tracking to 75rps

The SBCD1752/1753/1756/1757 series\* of Synchro-to-BCD Converters translate synchro or resolver analog data into 14-bit (0° to 359.9°) or 13-bit-plus-sign (±180°) BCD data with 0.1° resolution. Applications of these devices include:

- Monitoring rotary shaft-angle position with decimal-display readout.
- Furnishing shaft-position inputs for equipment configured to receive BCD data.
- Sensing shaft position when inputs are digital thumbwheel switches or decade boxes.

These converters use a Type 2 tracking servo loop, and can follow a rotating input without error. Reference frequencies of 50Hz through 2.6kHz are available in the standard models and provide for tracking at rates up to 75rps with no velocity lag.

All models are transformer-isolated, in-

cluding the 60Hz units, for high common-mode rejection, 500V isolation, and no errors due to solid-state Scott-T's.

To reduce the number of different devices, units operate over a wide range of frequencies, and the signal and reference input voltage ranges are programmable. For example, 60Hz units can operate at frequencies up to 400Hz. 400Hz units can operate at frequencies up to 2.6kHz; by changing external resistors in the signal and reference input circuits, the user can program the devices to accommodate higher-than-normal voltages. Two of the types operate from ±15V, +5V supplies, the others from +15V & +5V.

Unlike earlier designs, the conversion is provided with a single module, rather than two, and at less cost. Prices start at \$420 in small quantity. ▶▶▶

## APPLICATION IDEAS FOR MULTIPLYING DAC'S

### Digitally Controlled Gain is Useful in Filter Circuits

by John Newell

The uses of CMOS multiplying DAC's are becoming more widely known and more fully developed as more types with a wider range of resolutions and accuracies have become available and —most important— as prices have rapidly decreased\*.

This family of digital-to-analog converters, sired by the AD7520 in 1973, has an important feature not possessed by many DAC types: they can work with positive or negative reference voltage (for example, an ac analog input) without switching or circuit modification, and with negligible ac feedthrough at frequencies well beyond the audio range. The analog channel has wide dynamic range.

In such applications, it is useful to think of these devices as digitally programmed analog gain (or attenuation) channels, —digitally controlled potentiometers— rather than as d/a converters with wide-range reference capability.

#### ATTENUATION AND GAIN

The simplest application for a multiplying DAC is in the digitally controlled attenuator shown in Figure 1. Here, if the value of the digital word, relative to full scale (fractional binary), is  $D$ , ( $D = A_1 2^{-1} + A_2 2^{-2} + \dots + A_n 2^{-n}$ ,  $A_i = 1$  or  $0$ ) then the output is

$$V_o = -D V_{IN} \quad (1)$$

The number of values of gain (including zero) is  $2^n$ , for a binary converter, 4096 in the case of the case of the AD7541.\* The 8-bit AD7523\* has 256 steps of gain, and the "L" version permits each to be set to within 0.1% of full scale.

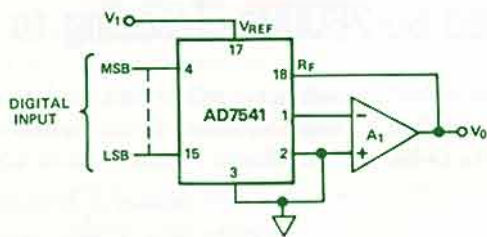


Figure 1. Digitally controlled attenuation.

In this circuit, the analog input is applied to the "reference" terminal of the ladder network, and the feedback from the output of the op amp is applied to the on-chip "feedback" application resistor, which tracks the ladder resistance.

The ladder transduces the input voltage to a current proportional to the digital word, and the feedback resistor develops the output voltage. Distortion and bandwidth are to a great extent determined by the choice of amplifier; such circuitry is well-suited to audio applications.<sup>1</sup>

\*See pages 9 and 14.

<sup>1</sup>See "Audio Application Ideas for CMOS DAC's" by Walter Jung, ANALOG DIALOGUE 10-1, 1976, pp. 16-17

Gains greater than unity can be achieved by connecting resistance in series with  $R_F$ , but a wide-range trim is necessary and the tracking advantage is likely to be lost.

Another way of obtaining gains greater than unity is to apply the DAC for the inverse function. If the signal is connected to  $R_F$ , and the output is fed back to the analog (reference) input of the DAC, then the nominal transfer function is:

$$V_o = -\frac{V_{IN}}{D} \quad D > 0 \quad (2)$$

Gains from 1 to 4096 are attainable with the AD7541. However, like divider-connected analog multipliers, the multiplying DAC circuit has errors that are inversely proportional to the magnitude of  $D$ —including gain, offset, and response time. When the DAC is used with a high open-loop-gain, low-drift op amp, such as the AD517, a linear programmable-gain amplifier, capable of handling reasonably low-level, low-frequency data, can be built.

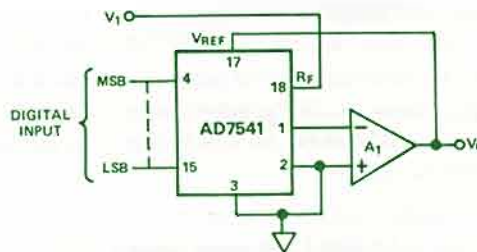


Figure 2. Digitally controlled gain.

In a data-acquisition system, simple logic (hard-wired or processor) can be used to determine the number of leading zeros in a trial conversion at unity gain and to set the gain of the amplifier for the maximum resolution of which the circuit is capable for a second conversion. Naturally, the resolution exceeds the calibration accuracy, but this may be quite in order for wide-range signals (Figure 3).

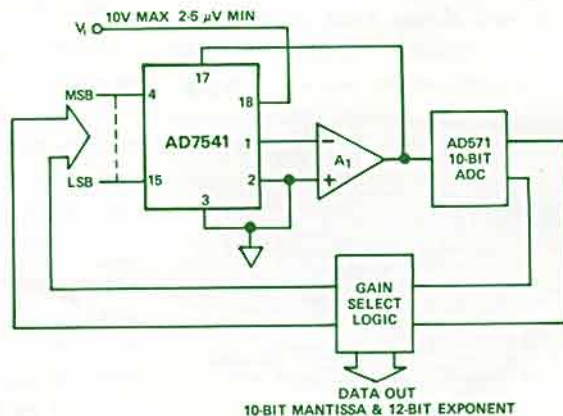


Figure 3. Ranging A-D converter.



## FILTERS

It is well-known that analog multipliers can be used to program the gain of filters in response to analog voltage.<sup>2</sup> If the multiplier (which serves as a voltage-controlled attenuator) is replaced by a multiplying DAC (which serves as a digitally controlled attenuator), the cutoff frequency is under digital control. Figure 4 shows a first-order filter with cutoff frequency proportional to the digital number (D); both high-pass and low-pass responses are available. If  $R_1 = R_2 = R_3 = R_4 = R$

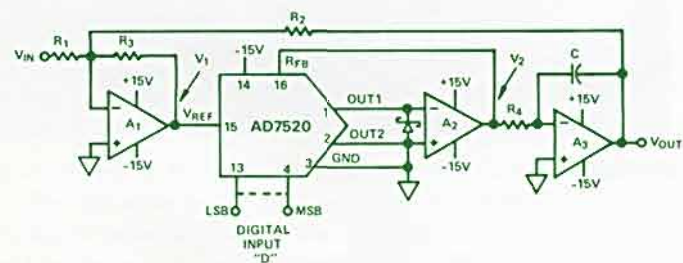


Figure 4. Digitally programmable low-pass filter.

$$V_{OUT} = \frac{D}{RC} \int V_1 dt \equiv \frac{D}{RCp} V_1 \quad (3)$$

and

$$V_1 = -(V_{IN} + V_{OUT}) \quad (4)$$

hence,

$$V_{OUT} = -\frac{D}{RCp} V_{IN} - \frac{D}{RCp} V_{OUT} \quad (5)$$

therefore,

$$V_{OUT} = -\frac{V_{IN}}{1 + \frac{RCp}{D}} \quad (\text{low pass}) \quad (6)$$

and

$$V_1 = -\frac{\frac{RCp}{D}}{1 + \frac{RCp}{D}} \quad (\text{high pass}) \quad (7)$$

The cutoff frequency,  $f_c = \frac{D}{2\pi RC}$ , is proportional to the digital input. In this example, a simplified case was chosen, in which all the external resistors are equal. In practice, however, insertion gain (or loss) may be necessary, or resistor-switched frequency ranging might be desirable. For the more-general case, the low-pass filter gain is  $R_2/R_1$ , and the cutoff frequency is

$$f_c = \frac{D}{2\pi R_4 C} \frac{R_3}{R_2} \quad (8)$$

Note that, except for the normal minor gain trim, the DAC network resistance is not involved in the filter time constant, since the converter is connected as an attenuator, with the feedback resistance tracking the network resistance. It is possible to show a simpler-looking configuration, which saves one operational amplifier (Figure 5). However, the cost is a wide-range trim, to compensate for variation in the initial ladder resistance, and poor temperature tracking, unless the capacitor's tempco is equal and of opposite polarity to the network tempco.

<sup>2</sup>NONLINEAR CIRCUITS HANDBOOK, D. H. Sheingold, Ed., pp. 138-141, Analog Devices, Inc., P.O. Box 796, Norwood MA 02062, \$5.95.

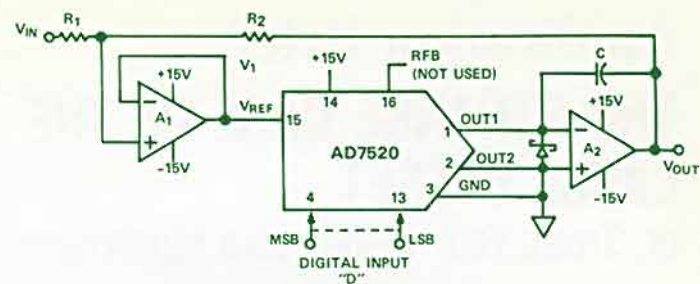


Figure 5. Programmable low-pass filter, simpler circuit.

## PROGRAMMABLE TIME-DOMAIN FILTERING

It is sometimes useful to have a filter with an indicial (unit-pulse) response that can be arbitrarily adjusted in the time domain. This can be done digitally, but the cost may be excessive if the analog signals to be processed are at high frequency, and especially if the output is desired in analog form.

Figure 6 shows a way of accomplishing such filtering. A tapped analog bucket-brigade delay line (charge-coupled device) delays the signal by a length of time proportional to the clock period. The outputs of the taps are delayed versions of the input signal. If the input signal is a pulse of unit amplitude, it will be propagated from one tap to the next with a delay depending on the clock frequency and the number of elements in each segment of the line.

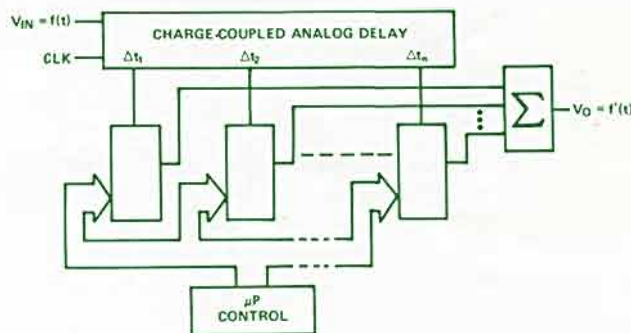


Figure 6. Programmable digital filter.

Each output is unloaded by a follower, if necessary, and applied to the analog input of a multiplying DAC. The digital input adjusts the gain of that individual tap, and the outputs of the DAC circuits are summed to provide the total response of the filter. If the input is a unit pulse, the response will simply be the responses of the taps, in order. For any other input waveform, the response of the filter will be appropriately related to the indicial response (since the system is linear), as long as the clock rate is substantially greater than twice the highest frequency present.

With this system, the filter characteristics can be changed by software, but the bandwidth for filtering can be several hundred kHz, and the process can be implemented with a low-cost microprocessor.<sup>3</sup>



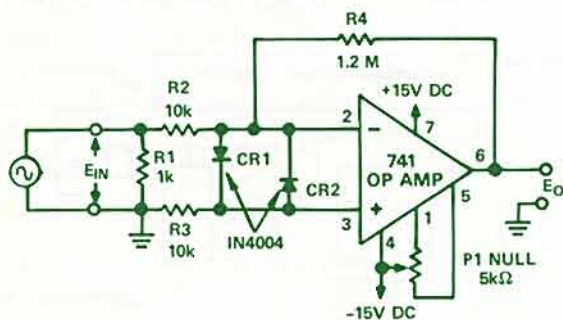
<sup>3</sup>In fact, such a circuit has been successfully implemented, with a SC/MP microprocessor and AD7530's, by Con Cremin, of Telectron, Dublin, Ireland.

## THE STRANGE CASE OF THE LARGE OFFSET or, Treat Your Diodes Like Mushrooms

by Richard P. Johnson\*

The use of back-to-back protective diodes, a time-worn, tried-and-true arrangement used to prevent damage to op-amp input stages, in a garden-variety application with a 741-type amplifier, led to the manifestation of an interesting phenomenon. If you have had similar symptoms, without being aware of the cause, the following words may be helpful.

While checking out the circuit shown, on the test bench, an attempt was made to adjust the offset-null pot for zero dc output under no-signal conditions. Every time the operator's hand was brought near the circuit board, the dc output level shifted randomly over a +1V to +5V range. There was no evidence of instability, oscillation, or stray signal pickup, yet the movement of an opaque (aha!) object over the circuit board caused a definite level shift.



The op-amp output level shift was found to bear a direct relationship to the amount of light incident upon the circuit board. For example, when the small 50-watt bench-type reflector floodlamp was turned off, the level-shift disappeared almost entirely and null action was controllable and consistent. The overhead flush-mounted fluorescent lamps had a less-important effect but introduced an ac signal superimposed on the dc level shift. When the ambient light was removed, the op-amp output level was entirely normal.

The cause was (you guessed it!) photovoltaic action of the 1N4004 silicon glass-encapsulated diodes, which produced a net photocurrent of up to several microamperes through the feedback resistor, causing the output level-shift. This was checked out by a measurement involving only the diodes, a dummy load, and a digital voltmeter. The diode having the larger value of photocurrent determined the polarity of the voltage.

Conclusion: Use diodes having opaque encapsulation, or keep glass-encapsulated diodes in the dark during test and use. (Ed. note: Or use amplifiers with built-in protection, e.g., AD517, if the application isn't quite "garden variety".) ▶▶▶

\*Richard P. Johnson is a Research Engineer at Russelectric Inc., South Shore Industrial Park, Hingham MA 02043.

## FOUR NEW PUBLICATIONS FROM ANALOG DEVICES



*Microprocessor Systems Handbook*, by Dr. D. P. Burton and Dr. A. L. Dexter, 200pp., an introductory book on microprocessors that integrates information on real-world (analog) interfacing into a basic text on digital systems. Send check for \$9.50 or Master Charge information to P.O. Box 796, Norwood MA 02062.

*Analog Devices Short-Form Catalog* This free new 32-page catalog of Analog Devices products "in a nutshell" provides salient comparative data on data-acquisition and signal-conditioning products currently being offered. Check the card or ask your sales engineer for a copy. Domestic prices included.

*Multiplier Application Guide* 50 pages of good, free, information on applications of Analog Devices analog multipliers. Intended to help answer the question "What are analog multipliers (and dividers, squarers, square-rooters, and multifunction components) used for in the real world?" Use the reply card.

*A User's Guide to Instrumentation Amplifiers*, by Jeff Riskin, a free 12-page guide to instrumentation-amplifier considerations, specs, and applications, by our Manager of Microcircuit Applications Engineering (and designer of the high-performance DIP-packaged AD522). Use the reply card.

## MORE AUTHORS



Russ VerNooy (page 10) is Marketing Specialist in microcomputer peripherals for our Instruments and Systems Group. Russ has a BSEE from Southern Illinois University, where he was a teaching assistant and a member of TBΠ. He is widely travelled and has performed classified electronics work in the U.S. Air Force.

Doug Grant (page 12), Microcircuits Applications Engineer at Analog Devices Semiconductor, has a BSEE from Lowell Technological Institute and is currently working on an MSEE. He has worked in converter development at Hybrid System, and  $\mu$ P-based data-acquisition systems and analytical instrument interfaces at Adams-Smith, Inc; he joined Analog Devices in 1976 as a Design Engineer. Doug is active in Amateur Radio.





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

IN THE LAST ISSUE (Volume 11, Number 2, 1977) . . . Monolithic IC RMS-to-DC Converter (AD536) . . . 6-Channel Scanning Digital Thermometer (AD2036) . . . 12-Bit Hybrid IC A/D Converter (AD572) . . . New Analog Interfaces for Microprocessors . . . Analog Signal-Handling for High Speed and Accuracy . . . Plus these New-Product Briefs: Low-Cost Multi-Channel Isolation Amplifier (286J) . . . Add-a-Display DPM Module (AD2023) . . . High-Performance 14-Bit ADC's (ADC1130 and ADC1131) . . . 10V Monolithic Pretrimmed Precision Reference (AD581).

SHOWS . . . In 1978, Analog Devices will be exhibiting in these electronics shows (U.S.) ELECTRO (May 23-25 in Boston) . . . WESCON (September 12-14 in Los Angeles) . . . ISA (October 16-20 in Philadelphia) . . . MIDCON (December 12-14 in Dallas) . . . If you're in the neighborhood, come in and see us.

PATENTS . . . We've been informed that the following patents have recently been issued to Analog Devices Engineers: . . . U.S. 4,029,974, to A. Paul Brokaw, "Apparatus for Generating a Current Varying With Temperature." Should you copy our AD559, this will be one of the patents you'll be infringing . . . U.S. 4,034,366, to John Memishian, Jr., "Analog-to-Digital Converter With Controlled Ladder Network." This is one of the patents you'll be infringing if you copy our ADC1102, ADC1130, or ADC1131.

ERRATA . . . In the ANALOG-DIGITAL CONVERSION HANDBOOK, page I-24, and the ANALOG-DIGITAL CONVERSION NOTES, page 13, there is an error in the non-inverting case in Figure 5a. For the circuit shown, the equation is:  $V_{OUT} = 2.22V_{IN} - 2.5V$ . If you need the equation,  $V_{OUT} = 2.22(V_{IN} - 2.5V)$ , which offsets the input rather than the output, change the 90k resistance to 40.5k, and the 14.21k resistance to 17.6k . . . In the AD2036 Data Sheet, Figure 4, page 4, use the complements, DATA HOLD and DATA READY, rather than the direct terminals indicated. In Figure 3, use DATA HOLD . . . In ANALOG DIALOGUE 8-1 (1974), Figure 1 on page 4, there should be an interconnection dot at the intersection of the IOUT1 line and the bus connecting the left-hand contacts of the switches . . . In the AD510 data sheet, the next-to-last paragraph on page 3 should read: "For gains of 10, the frequency response is down 3dB at 30kHz; for gains of 100, 3kHz. Full output of  $\pm 10V$  can be attained up to 1500Hz." Disregard the last paragraph . . . In the data sheet for the Model 752 Log Transconductor, Figure 2 on page 3, the ground return of A2 should be external via pin C.

PRODUCT NOTE . . . The Model 434 Multifunction Component data sheet indicates an "optional trim" circuit on page 2 but does not provide a procedure for using it to best advantage. Such a procedure might read as follows: "Significant improvement in performance can be achieved with external trimming. Input and output offset can be reduced to zero, improving the accuracy to typically better than 0.15% on Model 434B. Use the optional trim circuit shown on page 2:

1. Allow the unit to warm up for 5 minutes. Set  $V_x$  to +10V and  $V_z$  to zero (connect it to common). Adjust the output trim potentiometer for 0mV out.
2. Set  $V_x = V_z = +10V$ . Note the output reading.
3. Set  $V_x = V_z = +100mV$ . Adjust the  $X_o$  pot for the same output reading as Step 2.
4. Set  $V_x = +100mV$  and  $V_z = 10mV$ . Adjust the  $Z_o$  pot for +1.000V. Iterate Steps 3 and 4 until correct readings are obtained for both steps.
5. Repeat Steps 1 through 4 if further refinement in accuracy is desired. This procedure is intended to yield best linearity. Scale factor must be adjusted externally.

ANALOG DEVICES IN THE TRADE PRESS . . . "What to Look for in Analog Input/Output Boards", by Norm Bernstein, ELECTRONICS, January 19, 1978, page 113 . . . "I/O Boards - Your Microcomputer's Link to the Analog World", by Eric Janson, ELECTRONIC PRODUCTS, February, 1978, page 43 . . . "How to Select and Specify Board-Level Data-Acquisition Systems", by Norm Bernstein, INSTRUMENTS AND CONTROL SYSTEMS, March, 1978, page 55 . . . "I<sup>2</sup>L Puts it all Together for 10-Bit A-to-D Converter Chip", by A. Paul Brokaw, ELECTRONICS, April 13, 1978.



# The first thing we did was make it best. The best thing we did was make it first.

The first. The only. The award-winning, complete, self-contained digital to analog converter that offers true 18-bit resolution. That's 4ppm! The Analog Devices DAC1138.

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## The 18-bit DAC.



(actual size)