analog dialogue

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

DIFFERENTIAL INSTRUMENTATION AMPLIFIERS

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Modules — see page 14

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"High-Rel" A/D & D/A Converters

High-Performance Dedicated Divider

Line-Powered 4½-Digit Panel Meter

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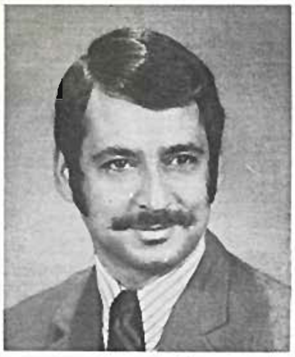
ANALOG DEVICES

Volume 9, No. 1, 1975

Editor's Notes



Ten years ago, Analog Devices first opened its doors. These have been exciting years. They have seen a frail fledgling grow into a major factor in our industry. Since those first few high-performance op amps, our catalog has grown in both numbers and categories. One may now find digital panel meters, A/D and D/A converters, CMOS switches, analog multipliers and dividers, linear IC's, dual transistors, and one of the broadest amplifier lines in the field. Yet we think the next decade will be even more fraught with opportunities, despite possibly disappointing times for the economy looming immediately ahead. If you're interested in our business and its prospects, write for our 1974 annual report.

George A. Philbrick – 1974

A pioneer in electronic analog computing and a devotee of feedback, he is generally acknowledged to have been the originator of the modular operational amplifier, which is today so important to electronic circuit designers. Scholar, gentleman, mathematician, writer, artist, teacher, poet, his fertile mind has produced many innovations in automatic control and electronic circuit design. Moreover, his every accomplishment invoked a feeling of *style*, of unusually high order, that left an indelible impression on all who came in contact with him.

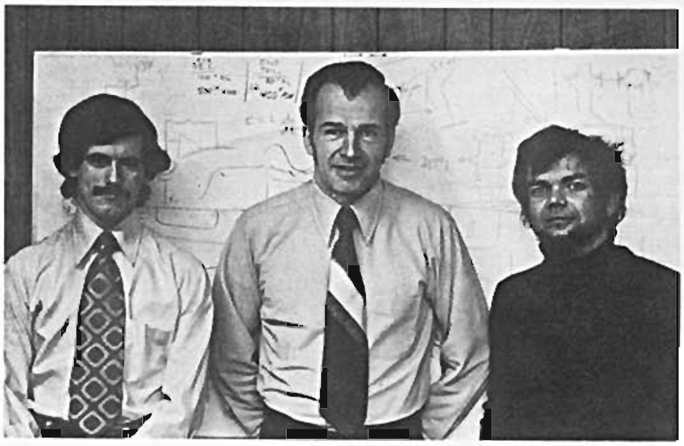
We who have had the privilege of more extended acquaintance with him have known him to be a person of infinite complexity, yet of unexpected simplicity, too. An immensely private person, his thoughts were nevertheless global; no aspect of the world nor of humankind escaped his interest. Impressive in size and appearance, as well as in intellect, he was by no means perfect; yet, those he had befriended and those he had wronged would all agree that he was indeed a special person.

There are many in our field who did not know him, but all should be aware of our generally unrecognized debt to him. Perhaps someone else would have proposed the operational amplifier as a circuit component if he hadn't. But only he could have: originated it (as a component of the modular unit operator, of which he was an early advocate), recognized its implications, communicated them to a technical world with verve and authority, built up a successful amplifier business (virtually unchallenged for 10 years), and nearly made "Philbrick" a synonym for modular operational amplifier (instead of "op amp"). Imagine today's world without the operational amplifier – if you can!

His presence has enriched all of technology; his departure is a matter of regret for all of us, his beneficiaries.

Dan Sheingold

THE AUTHORS



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MONOLITHIC DIFFERENTIAL INSTRUMENTATION AMP

2ND-GENERATION DESIGN HAS IMPROVED PERFORMANCE, RELIABILITY, CONVENIENCE

by Mike Timko, Bill Maxwell, and A. Paul Brokaw

The AD521* is a monolithic differential instrumentation amplifier with high input impedance, supplied in a 14-pin ceramic dual in-line package. Unlike an op amp, the AD521 is a true committed differential amplifier; it multiplies an input signal by a programmable fixed gain determined by the ratio of two externally-connected resistors.

Designed as a successor to the monolithic AD520 – a pioneering model introduced in these pages in 1972 (Vol. 6, No. 1) and now highly popular – the AD521 is a completely-new second-generation IC gain block with lower cost, simpler (but non-interchangeable) connections, and improved performance and reliability.

Salient features include internal compensation, protected inputs (for levels as great as 15V *beyond the supply voltage*), fast response (40MHz gain-bandwidth, 100kHz full-peak bandwidth, settling time of 5μs to within 0.1% of a 10V step), and low noise. You would expect common-mode rejection to be high, and it is: 110dB (G = 1000). AD521 performance is completely specified for gains from 1 to 1000V/V (the usual gain range for amplifiers of this type), but the circuit is actually capable of useful operation, with good performance, connected for gains ranging from 0.1V/V (important for normalizing large input signals) to well in excess of 10,000. Performance characteristics of the AD521 are tabulated in some detail on page 5.

The integrated-circuit AD521 is not an op amp. Designed and specified for operation with fixed closed-loop gains, it is a close relative of both the old-style rack-mounted instrumentation amplifiers and the newer modular instrumentation amplifiers (such as the 606 – see page 14), except for its very small size and cost, and differences in the details of performance characteristics. As such, its applications in front-end preamplification, ground-loop elimination, and line receiving are well known.† More important, the low cost, small size, and ease of use of the AD521 make it inviting to consider its use in large numbers (i.e., "amplifier per channel") to avoid the complexities and frustrations of low-level multiplexing.

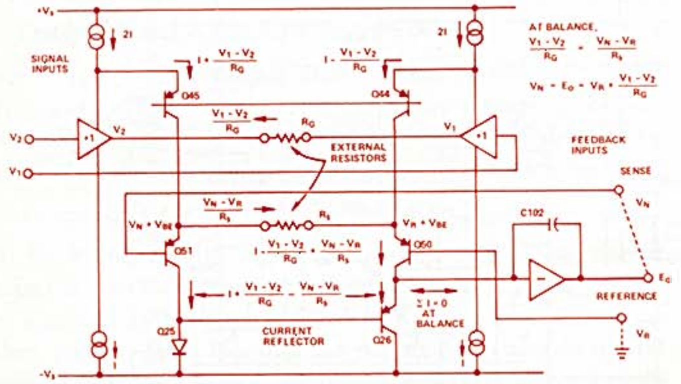
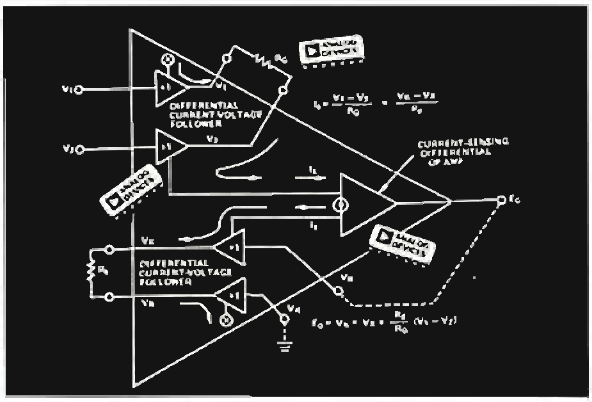


Figure 1. Simplified schematic diagram of the AD521 instrumentation amplifier.



DESIGN PRINCIPLES

Figure 1 is a simplified schematic diagram of the AD521. Like the AD520, it has a set of high-impedance input terminals (V1 & V2) and high-impedance feedback terminals ("Sense" and "Reference").

The input voltages are applied to a pair of unity-gain voltage followers, whose outputs appear across an externally-connected resistor, R_G. The current through R_G is $(V_1\text{-}V_2)/R_G$.

(continued on page 4)

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*For complete information on the AD521, request **N1**.

†See, for example, Chapter III-2 in the *Analog-Digital Conversion Handbook* (Analog Devices, Inc., 1972, $3.95, P.O. Box 796 Norwood, MA 02062).

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The followers are both excited by separate constant-current supplies, a negative supply, I, and a positive supply, 2I. Neglecting all other currents to the followers, the net currents flowing in the emitter circuits of Q45 and Q44 depend only on the two current sources associated with each follower and the current flowing through R_G. The currents through the emitters of high-β transistors Q45 and Q44 are:

$$I_A = 2I - I + \frac{V_1 - V_2}{R_G} \tag{1}$$

$$I_B = 2I - I - \frac{V_1 - V_2}{R_G} \tag{2}$$

Emitter-followers Q51 and Q50 are fed by the "sense" and "reference" input voltages. The current through the resistor R_S, which is connected externally between the emitter terminals, is therefore $(V_N-V_R)/R_S$. Since the emitters of Q51 and Q50 are connected to the collectors of Q45 and Q44, the sums of the currents at those nodes flow through the collectors of Q51 and Q50. The current through Q51 and Q25 is reflected at the emitter of Q26 and compared with the collector current of Q50. The net excess current flows through the summing point of the output integrator and charges capacitor C102.

In the usual connection, the output voltage E_o is fed back as the "sense" voltage V_N, and the "ref" terminal is grounded. The resulting negative feedback causes the fast integrator to seek a "steady state" in which its net input current is zero. In that condition,

$$I - \frac{V_1 - V_2}{R_G} + \frac{V_N - V_R}{R_S} = I + \frac{V_1 - V_2}{R_G} - \frac{V_N - V_R}{R_S} \tag{3}$$

$$V_N - V_R = \frac{R_S}{R_G}(V_1 - V_2) \tag{4}$$

$$E_o = V_N = \frac{R_S}{R_G}(V_1 - V_2) + V_R \tag{5}$$

Thus, the output difference (V_N-V_R) is constrained to develop a current through R_S that is equal to the current developed by (V_1-V_2) in R_G; the ratio of (V_N-V_R) to (V_1-V_2) at balance (i.e., the gain) is determined by the resistor ratio, R_S/R_G. In principle, gains from zero to more than 10,000 can be determined by the value of this ratio.

Output-stage bandwidth is determined by R_S and C102, the internal compensating capacitor. External capacitance, not necessary for compensation, can be connected essentially in parallel with C102 to reduce the overall bandwidth, if desired.

Unlike the AD520, the AD521's input section is independent of the output feedback loop. This design feature made frequency compensation easier and results in less interaction between the gain and zero-offset adjustments. It will be seen that gain can also be controlled by attenuating the feedback from the output to (V_N-V_R).

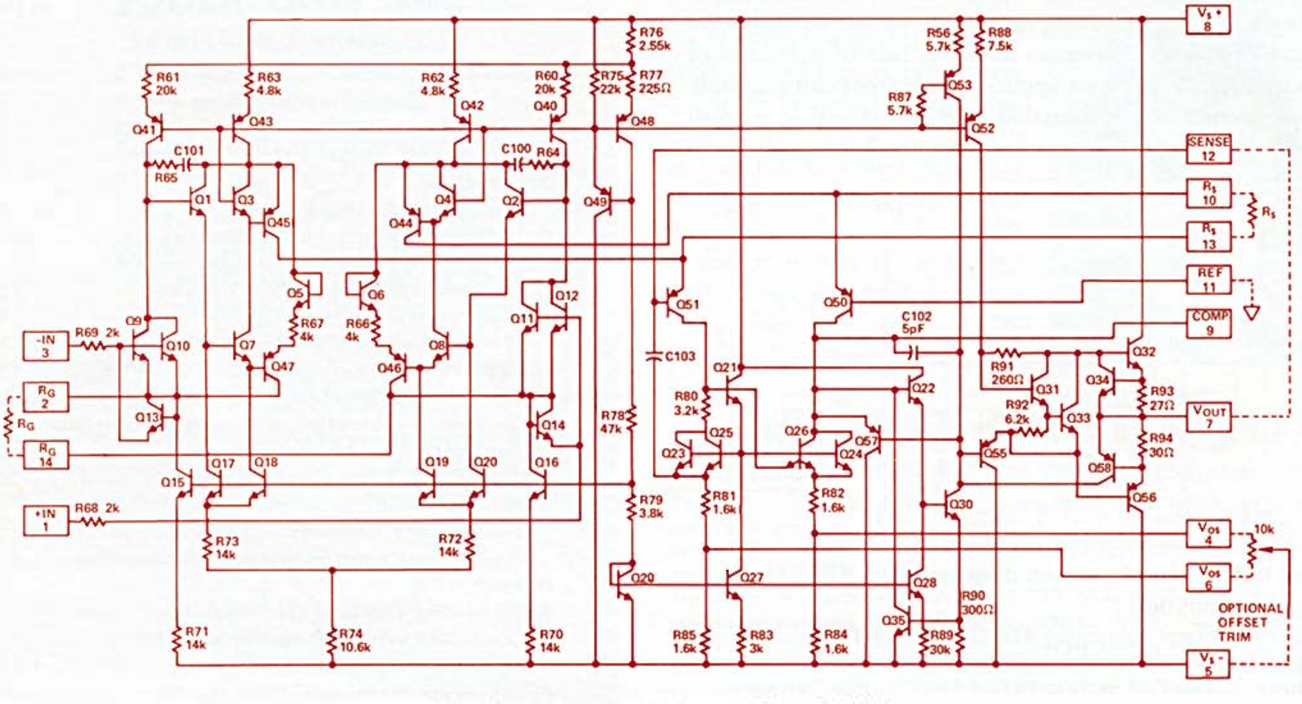
The complete circuit of the AD521 is shown in Figure 2. Its performance is tabulated on the opposite page. Three standard versions are available: the AD521J at lowest cost, the AD521K for highest performance, and the AD521S, specified for operation over the "military" temperature range.

APPLICATIONS

Uses of the Reference Input

The output of the AD521 is quasi-differential. That is, if the Sense feedback terminal is connected to the "Output" terminal, the output voltage that is determined by the ratio of R_S/R_G appears between the "output" terminal and the arbitrary point to which the Reference input is connected. If the Reference is connected to the common "ground" of the measurement system (which the AD521 is to feed), the system will receive the amplified AD521 input as its input; if the reference is at 2 volts above system common, the AD521 output will be translated by +2V.

The Reference input is useful for zeroing measurement offsets, whether they occur in the source, the amplifier, or the system



Complete schematic diagram of the AD521

that follows. It can be controlled manually or digitally (using a D/A converter). It can even be used for *automatic* zero correction, either periodically or on demand. In automatic zero correction, the output is sampled while the input is in a "zero" condition. The sample is stored and fed back in proper polarity to the Reference input, where it provides the very-nearly-exact correction voltage required (until the correction voltage – or the input – drifts off). Typically, zeroing could be programmed to occur just before an output signal reading is taken, to insure up-to-date correction. Figure 3 shows a scheme for accomplishing automatic zeroing.

Uses of the Sense Input

The Sense input is, in effect, the feedback "summing point" of the output section. If, instead of a direct connection from the output terminal to the Sense terminal, there is intervening circuitry, the output voltage will be whatever is required to cause the voltage difference between the Sense terminal and the Reference terminal to conform to the AD521's defining equation. For example, if there is attenuation, the result will be a corresponding augmentation of the device's gain; e.g., if $R_S/R_G = 1000$, and the output is fed back to the Sense input through a 10:1 attenuator, the overall gain will be 10,000. To minimize errors due to bias current, the effective resistances in series with the Sense and Reference terminals should be equal.

Possibilities of Combined Operations

The creative designer may find that the flexibility afforded by two cascaded sources of gain, together with the voltage independence (one might almost say "isolation") of the current-coupled input and output sections, and the immunity to (and fast recovery from) voltage overdrive will suggest a great variety of unconventional combined operations. In the "Application Note" section of future issues of this Journal, more real working circuits will be discussed, including a comparator that amplifies a differential input signal by a predetermined ratio, compares it with a 2.5V reference (furnished by the AD580 – see page 6), and provides a 2-level output (greater-less than the reference) with hysteresis (for noise-immunity), all with a single IC, the AD521.

Understanding Offsets

The output offset of an instrumentation amplifier with zero input is usually a linear function of the gain setting, viz.,

$$\text{Output offset (at gain G)} = V_{OSO} + G\,V_{OSI} \qquad (6)$$

where V_{OSO} is the output offset, measured with G = 0; and V_{OSI} is the "input offset", determined by measuring the output offset at G = 1000, subtracting V_{OSO}, and dividing by 1000. To refer (6) to the input signal, divide both sides by G:

$$\text{Output offset (r.t.i.)} = V_{OS} = V_{OSO}/G + V_{OSI} \qquad (7)$$

If, for example, with R_G open-circuited ($G \to 0$), the output offset is 30mV (without nulling), and if it increases to 630mV at a gain of 1000, V_{OSO} = 30mV and V_{OSI} = 0.6mV. We can then predict, from (6), that the output offset at gain of 100 will be 30 + 60 = 90mV, or, from (7), 0.9mV, referred to the input.

The optional zeroing adjustment allows V_{OSO} to be adjusted to near-zero. At balance, the V_{OSO} sensitivity to power-supply variations will be an order-of-magnitude less than the specified value.

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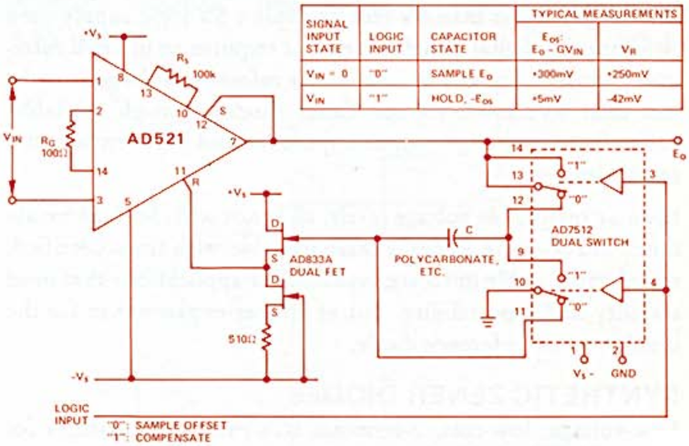
PERFORMANCE SPECIFICATIONS OF THE AD521

(Typical at V_S = ±15V, R_L = 2kΩ, and T_A = 25°C, unless specified otherwise)

GAIN				
Range for specified performance	1 to 1000			
Equation	$G = R_S/R_G$			
Error (% deviation from R_S/R_G)	±0.25 - 0.004G			
Nonlinearity (%)	0.1, max			
Temperature coefficient (ppm/°C)	±3 ±0.05G (J, K), ±15 ±0.4G (S)			
OUTPUT				
Rated	±10V, ±10mA, min			
At maximum operating temperature	±10V, ±5mA, min			
Impedance	0.1Ω			
DYNAMIC RESPONSE	G = 1	G = 10	G = 100	G = 1000
Small-signal bandwidth (+3dB)	>2MHz	300kHz	200kHz	40kHz
Small-signal bandwidth, ±1% flatness	75kHz	26kHz	24kHz	6kHz
Settling time (10V step to within 10mV)	7μs	5μs	10μs	35μs
Slewing rate	10V/μs	10V/μs	10V/μs	10V/μs
Full peak response	100kHz	100kHz	100kHz	100kHz
Overload recovery to within 10mV				
Differential (30V - 10μs pulse, V_{cm} = 15V)				50μs
Common-mode (30V - 10μs pulse)				10μs
VOLTAGE OFFSET (may be nulled)	AD521J	AD521K	AD521S	
Input offset voltage, mV typical (max)	2 (3)	0.5 (1.5)	0.5 (1.5)	
Vs. temperature μV/°C typical (max)	7 (15)	1.5 (5)	1.5 (5)	
Vs. supply, μV/%ΔV_S	3	3	3	
Output offset voltage, mV typical (max)	200 (400)	30 (200)	30 (200)	
Vs. temperature, μV/°C typical (max)	150 (400)	50 (150)	(50 (150)	
Vs. supply (improves if offset is nulled) mV/%ΔV_S	0.5	0.5	0.5	
SIGNAL INPUTS				
Differential input impedance	3×10^9Ω‖1.8pF			
Common-mode input impedance	6×10^{10}Ω‖3pF			
Input range for specified performance	±10V			
Maximum voltage without damage, power ON or OFF				
Maximum differential	30V			
Maximum common-mode	V_S ±15V			
Common-mode rejection, dc to 60kHz, 1kΩ source unbalance, dB typical (min)				
G = 1	74 (70)	80 (74)	80 (74)	
G = 10	94 (90)	100 (94)	100 (94)	
G = 100	104 (100)	114 (104)	114 (104)	
G = 1000	110 (100)	120 (110)	120 (110)	
REFERENCE TERMINALS				
Input resistance	20MΩ			
Bias current	2μA			
Voltage range	±10V			
Gain to output	1.0			
NOISE (voltages referred to output)				
Voltage (μV p-p) 0.1Hz to 10Hz	$150\sqrt{1 + (0.003G)^2}$			
(μV rms) 10Hz to 10kHz	$30\sqrt{1 + (0.04G)^2}$			
Current (pA rms) 10Hz to 10kHz	15			
POWER SUPPLY				
Operating range	±5V to ±18V			
Quiescent current drain	5mA max			
TEMPERATURE RANGE	AD521J	AD521K	AD521S	
Specified operation	0° to 70°C	0° to 70°C	-55° to +125°C	
Operating	-25° to +85°C	-25° to +85°C	-55° to +125°C	
PRICE ($U.S.)				
1-24	$12.75	$18	$30	
100+	8.50	12	20	



SIGNAL INPUT STATE	LOGIC INPUT	CAPACITOR STATE	TYPICAL MEASUREMENTS	
			E_{OS}: $E_O - GV_{IN}$	V_R
V_{IN} = 0	"0"	SAMPLE E_O	+300mV	+250mV
V_{IN}	"1"	HOLD, $-E_{OS}$	+5mV	-42mV

Figure 3. Automatic zeroing configuration employing dual CMOS switch and dual FET with the AD521. With C = 0.1μF settling time is 250μs; drift rate is 0.75mV/s. With C = 0.02μF settling time is 80μs; drift rate is 3mV/s.

MORE ABOUT THE AD580 MONOLITHIC IC VOLTAGE REGULATOR
OR... LOW-COST CONSTANTS FOR ANALOG CIRCUITS & SYSTEMS

by A. Paul Brokaw

Today, voltage-regulator circuits are principally used in power supplies for decreased output impedance and ripple, and improved stability. However, the low cost and excellent performance of the new monolithic regulators, together with their ease of use and small size, bring them face-to-face with the Zener diode for applications that require constant voltage in analog circuits. Here are some of the salient issues and a description of the AD580 precision regulator circuit, introduced in ANALOG DIALOGUE 8-2.*

Block diagrams of analog circuits show 3 classes of voltage – 1) signals: inputs from transducers, results of analog processing, intermediate results at the outputs of op amps, etc., 2) service voltages: excitation power supplies, digital control logic, etc., and 3) constants: converter and system reference voltages, algebraic constants for analog computing, and multiplier-divider scale factors.

The constants used in analog circuitry have diverse accuracy requirements, which are fulfillable by a gamut of devices, ranging from Josephson junctions and electrochemical cells to the highly-popular Zener diodes. Often, one can simply tap off a fraction of the dc supply voltage with a resistive divider and "stiffen" it, if necessary, with the low output impedance of an operational amplifier circuit. This last approach is useful only if the supply has adequate stability and repeatibility – which could be costly if the system requires a high-capacity power supply.

ZENER DIODES AS REFERENCES

Temperature-compensated avalanche breakdown diodes are available to operate from upstream voltages greater than 6V. These diodes have reasonably low dynamic impedance; circuits having quite low temperature sensitivity can be designed using selected devices with constant-current excitation.† If the available supply is less than 6V (for example a 5V logic supply used to operate a digital panel meter that requires an internal reference), or if – for any reason – the reference voltage is to be less than 6V, lower-voltage Zener diodes, though available, suffer from both high output impedance and high temperature sensitivity.

Even at reasonable voltage levels, all is not well. Surface breakdown of avalanche diodes is often unstable with time. Certified, tested, time-stable units are available for applications that need stability and repeatability, but at greater expense than for the usual low-cost reference diode.

SYNTHETIC ZENER DIODES

Low-voltage, low-cost, 2-terminal IC's exist as substitutes for Zener diodes. These "band-gap" devices produce a reference voltage that depends inherently on the properties of transistor junction potentials. Since V_{BE} depends on bulk properties of the semiconductor material, its long-term stability is essentially unaffected by the surface phenomena that cause instability in avalanche diodes.

Like Zener diodes, however, these 2-terminal band-gap reference circuits need an auxiliary current supply, which itself needs regulation for best performance. If the load is variable, the no-load supply current has to be greater than the maximum anticipated load current (even if it is required only briefly); this limits their utility in low-power circuits, e.g., with batteries.

Since synthetic Zeners are no more convenient to use than conventional reference diodes, and tend to cost more for equivalent temperature-stability, they tend to be used only where low-voltage operation or long-term repeatability have been essential requirements. Furthermore, the basic 3-transistor cell operates at 1.2 volts, requiring amplification in reference applications; but 1.2V is insufficient common-mode voltage to operate op amps (such as the 741) when used with a single-ended supply (an economical mode of operation for reference amplifiers).

ENTER THE AD580

The AD580 unites the stable low-voltage operation of the band-gap circuit with the active circuitry needed to make a self-contained Input-Output device on a single monolithic chip in an easy-to-use 3-terminal package. Easily chosen for special-purpose applications where low power drain, excellent long-term stability, and low-voltage operation are essential, its simplicity, low cost, and convenience would also appear to make it a natural choice wherever a stable voltage with substantial low-impedance output capability is desired.

It will operate successfully from voltage as low as 4.5V, and it will easily handle as much as 30V without external current-limiting. Though it supplies an output current of up to 10mA, it idles at about 1mA when lightly loaded, minimizing power dissipation and self-heating. It can operate from batteries or logic supplies, and its 2.5V output is sufficient to bias transistor circuits (including most op amps) into the operating range.

THE BAND-GAP PRINCIPLE APPLIED TO THE AD580 DESIGN

Most designers are familiar with the -2mV/°C temperature coefficient of the base-emitter voltage of a transistor biased into an active region. The exact value depends on the base-emitter voltage, but it is so reproducible for a given device, that diode-connected transistors are often used as thermometers. If the emitter current is made proportional to temperature, the V_{BE} tempco is nearly constant over a wide temperature range.

If V_{BE} for several devices were plotted as a function of temperature and extrapolated towards absolute zero (-273°C), the straight lines would have differing slopes but would intersect at

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*For technical data on the AD580, request **N2**.

†A good example of such a circuit can be found in the AD562 data sheet. Request **N3**.

the same value of voltage: 1.205V, the "bandgap voltage of silicon, extrapolated to zero," as shown in Figure 1. If it were possible to generate a voltage that increased proportionally with temperature at the same rate at which V_{BE} of a given transistor decreased, then the sum of the two voltages would be constant, and equal to the bandgap voltage, a physical constant.

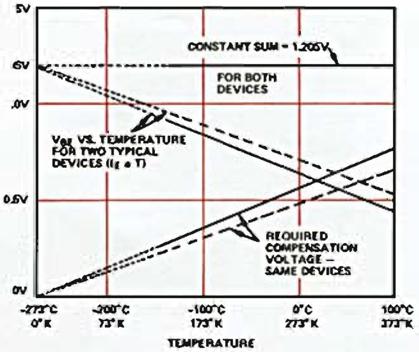

Figure 1. Extrapolated variation of base-emitter voltage with temperature ($I_E \alpha T$), and required compensation, shown for two different devices.

We can obtain such a voltage by amplifying the difference between the V_{BE}'s of similar transistors operating at different values of current density (J_1 and J_2 in A/m²), that is

$$\Delta V_{BE} = \frac{kT}{q} \ln \frac{J_1}{J_2} \qquad (1)$$

where k/q is the ratio of Boltzmann's constant to the unit of electronic charge (86.17μV/°K), T is absolute temperature, °K, and ΔV_{BE} is the difference of the base-emitter voltages. Since ΔV_{BE} is proportional to temperature (if J_1/J_2 is constant), it can be scaled up from its normal value of about 26mV·ln (J_1/J_2) near room temperature to a value that, summed with V_{BE} at the same temperature, gives a total of 1.205V, which will then be (ideally) independent of temperature.

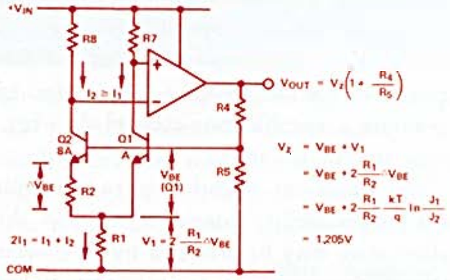

Figure 2. Basic bandgap-reference regulator circuit.

Figure 2 is a simplified version of a scheme that implements this relationship and also provides an amplified output. The two transistors, Q1 & Q2, produce the voltage difference, ΔV_{BE}, amplify it, add it to the V_{BE} of Q1, and scale the output up to a stable 2.5V. Here's how it works:

The amplifier output V_{OUT} adjusts the base voltage of transistors Q1 & Q2, via the voltage divider R4-R5, in order to establish an equilibrium condition in which the collector currents of Q1 & Q2 are equal (since $R_7 = R_8$, and the net input to the amplifier at equilibrium is "zero"). The emitter area of Q2 is 8x that of Q1, so for equal currents at equilibrium, $J_1 = 8J_2$.

The base-emitter voltage-difference, ΔV_{BE} (across R2) is therefore equal to (kT/q) ln 8; the design value of R2 is determined by the desired level of current through Q1 & Q2, which must be equal to (kT/qR_2) ln 8. The sum of the two currents, equal to $(2kT/qR_2)$ ln 8, flows through R_1. Since the base voltage, V_Z, is equal to the sum of the V_{BE} of Q1 and the voltage across R_1,

$$V_Z = V_{BE} + 2\frac{R_1}{R_2}\frac{kT}{q}\ln 8 \qquad (2)$$

The ratio, R_1/R_2, is the value necessary for $V_Z = 1.205V$. Integrated-circuit process control makes V_{BE} predictable, so that R_1/R_2 can be predetermined and implemented with stable, low-tracking-TC, thin-film resistors, deposited on the silicon chip.

To establish a given value of base voltage (e.g., 1.205V), the output voltage must be $(1 + R_4/R_5)\,V_Z$. For the AD580, the nominal output voltage is 2.5V; the appropriate resistance ratio, R_4/R_5, is determined by thin-film resistors deposited on the chip. Figure 3, a complete schematic of the AD580, shows how the op-amp function is implemented. The cell output, at the collectors of Q1 & Q11, drives the output via a level translator and a Darlington current-booster. The amplifier input circuit is bootstrapped to the output to increase the effective open-loop gain and lower the output impedance to less than 1 ohm. R3 compensates for the base-current drop in R4.

Versions of the AD580 are available with output voltage within 2% of nominal and with temperature coefficients as low as 10ppm/°C. The "rough-and-ready" 3% AD580J provides a complete voltage regulator with <0.6% change over a 70°C temperature range and less than 1Ω output resistance, for only $2 in 100's.

By the use of a feedback configuration with an op amp, which provides both the output and the AD580's excitation, the nominally better-than -0.03%/V (7 -30V) line regulation can be greatly improved, the output voltage may be adjusted to any value greater than 4.5V, and the output current may be beefed up.

Besides the many applications implied in the first paragraph (and elsewhere), the AD580 can also be used as a series-connected 2-terminal *current* load. If a fixed resistor is connected between the output terminals, a constant current of approximately 2.5V/R + 1mA will flow between V_{IN} and the AD580's "common", regardless of the applied voltage (from 4.5V to 40V).

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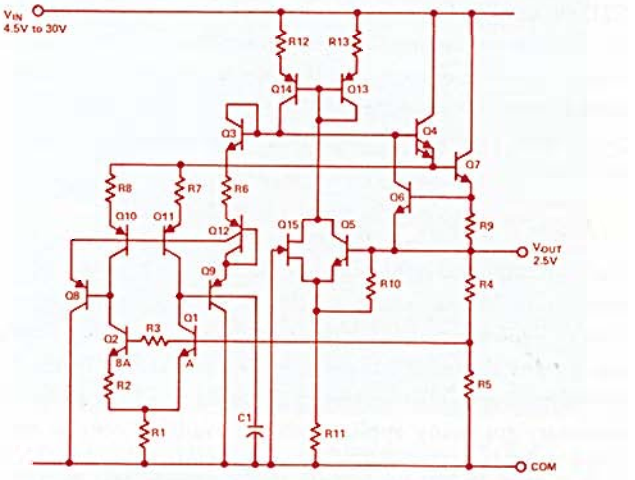


Figure 3. Schematic circuit diagram of the AD580.

IS THERE AN "883" FOR MODULES ?
CONVERTER MODULES FOR 'HIGH-RELIABILITY' APPLICATIONS

by Jim Fishbeck

"High reliability" is somewhat of a misnomer. No manufacturer will long prosper in business unless users find that the product is highly reliable, i.e., it works in the specified manner whenever (and for as long as) needed. The military services, for many years the major user of electronic equipment, faced with the problem of buying very large numbers of items from very large numbers of unknown vendors (all low bidders!) for use under a wide range of environmental conditions, have established a large number of "MIL Standards" to apply to every aspect of the components, assembly, test, and qualification techniques to be used by a system contractor.

Though this does not guarantee high reliability per se, it does ensure a commitment on the part of the supplier to at least document its conformance to a set of minimum standards. If the supplier is already committed to high reliability as a source of customer satisfaction for all of its customers, its credibility may be even higher when it offers a line of products, for application in harsh environments, specifically meant to meet certain military standards (however pacific the application or end use). Analog Devices introduces a line of standard conversion products that embrace both commitments.

Electronic equipment is often used in military, industrial, and scientific applications where it may be exposed to hostile environments, yet must perform properly when called upon. Furthermore, the equipment may be physically remote or otherwide hard to get at, so that component replacement and repair is either difficult or impossible. Clearly, the reliability of equipment used in such applications is of transcendent importance.

Engineers designing apparatus or systems for high reliability usually specify that "high-rel" components be used. Many such components are readily available; they profess to meet certain standards that were originally established by the Military, but are now widely accepted and understood. For example, many microcircuits (such as hybrid and monolithic integrated circuits) are available with qualification to MIL-M-38510, or at least with processing to the applicable requirements of MIL-STD-883. Many discrete semiconductors are available that meet the requirements of MIL-S-19500. Passive components, such as resistors and capacitors, can be specified via reference to applicable MIL documents. Even such components such as jacks, switches, relays, and transformers, can be specified via reference to MIL standards, such as MIL-STD-202.

THE USERS DILEMMA

Unfortunately, as sophisticated circuits, such as A/D and D/A converters, became more widely adopted as "buy" rather than "make" subassemblies, project engineers designing "reliable" equipment containing these devices have had to face a special problem. Complete converters offering the high performance necessary for many applications are available only as encapsulated modules (assemblies of components, tested and guaranteed to perform a functional operation under specified conditions), and none of the commonly-used military specifications and standards apply directly to encapsulated modular circuits. This situation has forced the equipment designer to

1) Obtain waivers to use "commercial-grade" converters; or

2) Generate his own specification drawings, in the hope that some converter-module manufacturer can be persuaded to bid on supplying a custom product that meets the "high-rel" requirements that the system designer (or his customer) deem necessary; or, (worse):

3) Design his own converter(s), thus essentially duplicating an available technology, but at far higher overall cost.

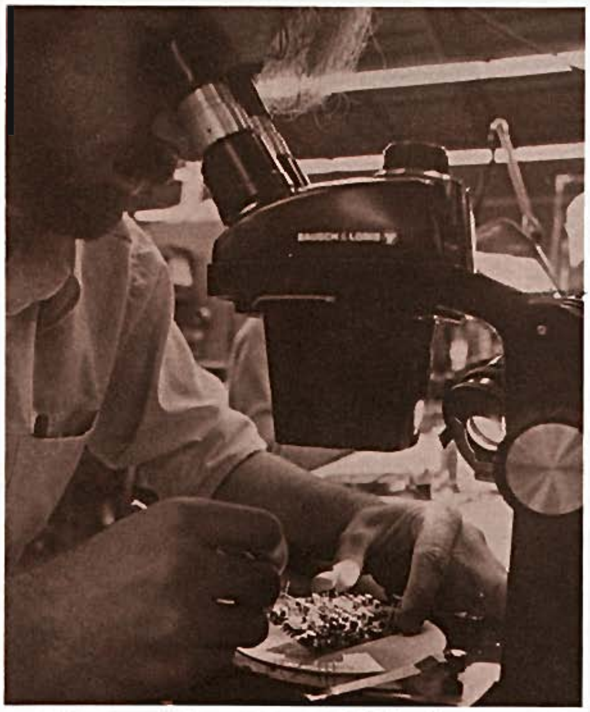
None of these approaches is really satisfactory. If the designer specifies commercial-grade converters and either accepts deviations from the required specs or designs the system to tolerate the deviations (e.g., by building-in a constant-temperature "oven"), he may compromise the performance, reliability, or cost of the overall system. If he attempts to specify a high-reliability converter *a priori*, he must first expend considerable effort to generate a specification-control drawing, then find (a) supplier(s) to bid on it, and finally be prepared to go through a long-drawn-out negotiation to determine exactly what specs a high-reliability converter can and should meet. The third alternative may be wasteful by its diversion of time and resources, and may in any case eventually result in poor performance and reliability.

This dilemma is quite natural, because in equipment design, the converter is a component, while in actual fact (if it isn't an IC) it is an electronic subassembly, albeit wholly self-contained. But most military performance standards and specifications are oriented towards individual components, not towards electronic subassemblies.

One approach to an economic solution of this problem would be for an experienced manufacturer of reliable converters to develop a standard line of documented "high-rel" converters, based on what customers have historically sought. The manufacturer's extensive experience with his own process would provide a good indication as to what could be achieved. Of utmost importance would be a specification-control document originated by the module manufacturer that completely specified the converter. This specification could then be copied directly by the equipment designer with little—if any—modification and would become the source-control drawing for the procurement of that particular high-rel converter.

ANALOG'S ANSWER

Analog Devices embarked on just such a program. Past experience with customers requiring high-rel modular converters indicated that 4 of the company's standard converter types, if available in fully-documented versions, could fulfill the converter requirements for a great many high-reliability system applications. These products were the ADC-12QM 12-bit A/D Converter, the DAC-12QS 12-bit voltage-output D/A Converter, the MDA-12QD 12-bit current-output DAC, and the SHA-2A high-speed Sample-Hold.*



Since a high-rel module must contain high-rel components, the first step was to specify exactly what components were to be used and the specifications that they were to meet. A predicted MTBF ("mean time between failures") was then calculated for each module, in accordance with MIL-HDBK-217A.

Next, we ran a series of electrical tests on sample lots of each module type at -55°C, 25°C, and 125°C to verify the performance capability of the modules at the extremes of temperature.

Then a series of environmental tests was devised that, as a whole, would indicate the modules' ability to operate in harsh environments. The tests were selected from appropriate military test standards, and sample lots of each module type were put through their paces by an independent outside environmental-testing laboratory.

We developed a manufacturing procedure that would ensure that the modules were manufactured in accordance with high-rel requirements; it included a quality-conformance inspection procedure that all units would have to pass. The illustrations show pre-encapsulation visual inspection and preparation for 168-hour high-temperature burn-in.

Finally, and perhaps most important as a time-saving convenience for the user, each type was documented with its own specification-control drawing. These drawings, averaging 17 pages in length, include details of component requirements, environmental specifications, MTBF, electrical performance specifications, and quality-conformance inspection requirements. They also include absolute-maximum ratings, recommended operating conditions, hookup information, block diagram, and physical outline drawing.

The first four modules in the new high-reliability product group are the ADC1111, the DAC1112, SHA1114, and DAC1117.†

THE PRODUCTS

THE ADC1111 is a high-reliability version of the popular ADC-12QM 12-bit A/D Converter, It has a 25μs maximum conversion time and can operate over the -55°C to +125°C temperature range. It is packaged in a 2" x 4" x 0.4" (51 x 102 x 10.2mm) module, priced at $475 (1-9).

THE DAC1112 is a 12-bit voltage-output D/A Converter that is a high-rel version of the DAC-12QS. It operates over the full military temperature range, and any of 5 voltage-output ranges can be programmed by the user. It is a 2" x 2" x 0.4" (51 x 51 x 10.2mm) module, priced at $475 (1-9).

THE SHA1114 is a high-speed, high-reliability sample-hold amplifier. A high-rel version of the SHA-2A, it can accept input signals in the range ±10V and operate over the full military temperature range. It is a 2" x 3" x 0.5" (51 x 76 x 13mm) module priced at $425 (1-9).

THE DAC1117 is a hermetically-sealed 12-bit current-output DAC, packaged in a compact 1" x 1.5" x 0.4" (25.4 x 38 x 10.2mm) module. Capable of operating from -55°C to +125°C, it has a ±1mA or a 0 to -2mA output range. It is the high-rel version of the MDA-12QD, priced at $555 (1-9). ▶▶▶

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*For data on these standard "commercial-grade" types, request **N4**.

†For a brochure describing this product line, request **N5**. For a copy of the complete specification-control drawing for one or more of these products, you may use the reply card to request it specifically by type.

LINE-POWERED 4½-DIGIT PANEL METER—AD2008
HAS OPTO-ISOLATED FRONT END, HIGH CMR & NMR, OPTIONAL RATIOMETRIC INPUT, BCD OR PULSE-TRAIN OUTPUTS

The AD2008* is a 4½-digit line-powered DPM designed for making high-accuracy measurements in adverse electrical environments. The results of the measurement are interfaced to humans via a large, highly-legible gas-discharge display, and to data-acquisition systems via a choice between 2 optional data-output formats (both of which are compatible with the run of popular logic hardware).

The AD2008 measures voltages over a full-scale range of ±1.999V with an error less than ±0.005% (50ppm) of reading ±50μV (25ppm) of full scale ±1 digit. A fully-floating opto-isolated input section allows measurements to be made with common-mode voltages up to 300V rms with better than 100dB of common-mode rejection. The AD2008 achieves normal-mode (line-frequency) rejection better than 60dB by integrating the input voltage for 3 line periods.

The excellent noise rejection and common-mode specifications of the AD2008 make it particularly well-suited for industrial process-control readout, where a noisy electrical environment and high common-mode voltage is the rule, not the exception.

On the other hand, its excellent stability, accuracy, and versatility make it ideal for in-house test and calibration equipment for electrical instrumentation. Because it is ac-powered, it is easy to apply, without requiring an external power supply.

It is ideal for digital weighing systems, and not only because of its full 4½-digit stability and accuracy. It can also be adjusted to center the last digit as required by Federal standards (NBS Handbook 44).

The AD2008 can measure the ratio of two voltages, a useful feature when measuring with a transducer that is sensitive to variations in its excitation voltage. The AD2008/R option will accept external reference voltages in the range of 0.6 to 1.3V, while maintaining its specified accuracy. In a departure from the industry norm (greatly simplifying application), the ratiometric inputs are buffered for high input impedance, protected against overvoltage, and filtered for noise rejection.

OUTPUTS AND DISPLAY

The nucleus of the AD2008 digital circuitry is a MOS/LSI chip that provides 4½ decades of counters and latches. Its use results in reduced circuit complexity, enhanced reliability, very modest power requirements (hence cooler operation) and small size. In its simplest form, the AD2008 provides a display only; the character-serial output of the logic chip is reserved entirely for internal use. ("Character serial" involves a train of pulses that turn on (or off) the 7 segments of each digit, in a multiplexed sequence.)

Two options are available at modest additional cost: The /B version offers full parallel BCD data for 4½ digits, polarity, and overload lines. The /X version has a train of pulses, the number of which (20k *max*) is proportional to the ratio of the integral of the input signal (over 50ms) to the reference. The train of pulses at 200kHz may be counted externally in binary, BCD, complementary, or whatever form is desired, and used as an input to a data-handling system, a printer, SERDEX,† remote displays, etc. To ensure universal applicability, all data outputs are compatible with DTL, TTL, CMOS, and PMOS, and will drive as many as 6 TTL loads. Thus, the user of the AD2008 has considerable versatility, as well as accuracy, at his command, for data acquisition.

The AD2008 displays the digitized data on large 0.55" (14mm) Beckman 7-segment gas-discharge displays that are visible at distances up to 50' (15m) and over angles of 130°. Inputs are available for testing each segment of the display, blanking the display, or blanking the polarity signs. Inputs for 4 decimal-point positions are also provided.

A unique "tri-state" Trigger-Hold Input controls 4 operating modes. With no input, there are 2½ conversions/second. Apply logic "0" to *hold* the latest conversion; pulse it with "1" to initiate conversions, at up to 5/s; hold it at "1" for fast, free-running conversion (up to 10/s).

The AD2008 is packaged in a newly-designed (Analog Devices) case that fits the same 1.682" x 3.930" (42.72 x 99.82mm) panel cutout used by many other DPM manufacturers. In order to make mounting approach the earlier snap-in design in simplicity, a new method has been devised: The user slips the DPM into the panel cutout, snaps in two side rails, and tightens two screws on the rear plate to secure the DPM to the panel. Red or amber snap-on lenses (with or without ADI logo) may be specified; they are coated to prevent distracting reflections from the lens surface.

The overall depth of the AD2008, from lens to edge connector (not including mating female) is 5.83" (148.1mm). All versions of the AD2008, including 5 line-power (V & f) options, and the ratiometric option — but excluding the data-output options — are the same price: $295 (1-9), $195 in 100's. The parallel-BCD "B" version is $60 additional ($40 in 100's), and the pulse-train "X" version is $30 additional ($20 in 100's).

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*For a data sheet on the AD2008, request **N6**.

†For information on the Analog Devices SERDEX SERial Data EXchange modules for interfacing data to teletypewriters or corresponding ports on data-handling equipment in ASCII, request **N7**.

IC DAC's – PURE AND APPLIED

DAC1132: 12 BITS WITH REGISTERS

The DAC1132* is a complete, high-performance 12-bit D/A converter, packaged in a compact 2" x 2" x 0.4" (51 x 51 x 10.2mm) module. Everything is included: registers, reference, output op amp, connections for 5 output-voltage ranges, an AD562 basic IC DAC, and connections for gain and offset trims, if needed.

Performance specifications include ±½LSB linearity error, monotonic performance (0 to +70°C), 2μs settling time to 0.01%, and 8ppm/°C gain-temperature coefficient. Price is $159 (1–9).

A complete, ready-to-go DAC assembled in a compact, potted module, the DAC1132 provides convenient positive-true-input-to-positive-full-scale-output conversion, with a choice of jumper-programmable output ranges (±2.5V, ±5V, ±10V, 0-5V, 0-10V).

The TTL/DTL-compatible latching input register accepts either binary or offset-binary (bipolar) inputs. Improved cost, size, and performance make the DAC1132 a natural successor to earlier devices (e.g., DAC-12QM) for new designs. The only ADI 12-bit DAC with registers having lower cost is the larger, lower-performance DAC1118§, at $112 (1–9).

The key to the DAC1132's high performance/size ratio is our AD562† digital-to-current converter. The 2-chip AD562 contains the precise current switches and laser-trimmed stable thin-film resistor network that are essential for high-performance conversion.

*For data on the DAC1132, request **N8**.
§For information on DAC1118, request **N9**.
†For information on the AD562, request **N3**.

AD7521 MONOLITHIC 12-BIT DAC PROVIDES 4-QUADRANT MULTIPLICATION

The AD7521* is a monolithic thin-film-on CMOS 12-bit multiplying D/A converter, supplied in an 18-pin dual in-line package.

Like the AD7520 10-bit DAC, which was introduced a little more than a year ago and has won heartwarming acceptance, the AD7521 has a number of attractive features:

- 4-quadrant multiplication—positive or negative analog input (V_{REF}) and positive or negative digital input (offset binary)
- low power dissipation (typically 20mW)
- low analog feedthrough error (10mV *max* at 100kHz, 20Vp-p input, all bits off)
- low nonlinearity temperature coefficient (2ppm of full scale/°C *max*)
- only a single +15V supply required
- low cost ($20.75, 1–49; $15, 100+ for AD7521JN).

Six standard versions of the AD7521 are available, in the hermetically-sealed ceramic package, graded for relative accuracy and operating temperature range; 3 low-cost 0 to +70°C versions are available in plastic dual in-line packages.

Like the AD7520, the AD7521 consists of an R–2R ladder attenuator and a set of double-throw CMOS switches, as shown in Figure 1. The binary-weighted output current of each leg is switched between the two output terminals, which are held at ground (or virtual ground) potential. Since the network and the switches will both pass either positive or negative current, the input voltage V_{REF} may be positive or negative. The sum of the currents to i_{OUT_1} flow through the summing point of the output op amp and develop an output voltage (via the on-chip feedback resistor) equal to the product of V_{REF} and the fractional digital number N. The excellent tracking properties of the thin-film resistor network maintain a <10ppm/°C gain-temperature coefficient, despite the moderate variation of absolute resistance with temperature.

In offset-binary handling of bipolar digital signals, the complementary output bus (i_{OUT_2}), normally grounded, is connected to the summing point of an op-amp current-inverter circuit, the output of which is summed with i_{OUT_1} plus a minuscule LSB-compensation current. Since V_{REF} may also be positive or negative, the circuit functions as a 4-quadrant D/A multiplier, with very low distortion of the V_{REF} signal.

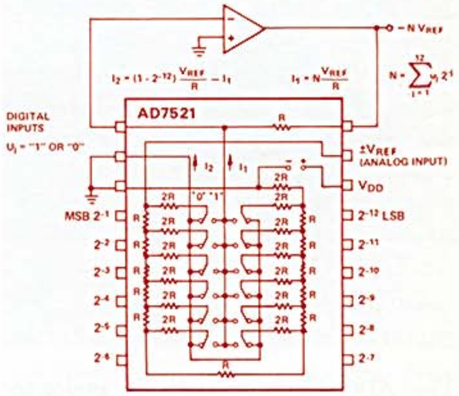

Figure 1. Functional schematic diagram of the AD7521 multiplying DAC.

RESOLUTION VS. ACCURACY

The three grades of conversion accuracy referred to above involve maximum linearity errors of 0.2%, 0.1%, and 0.05%. While ordinarily the accuracy of a converter is expected to match its resolution (an *essential* property for many applications), there exist applications calling for greater or less accuracy. An example is a 4-bit converter controlling a power supply with 16 output levels, each of which must be accurate to within 0.01%, a case of 4-bit *resolution* and 12-bit (or better) *accuracy*.

On the other hand (and of particular relevance to the AD7521), a voltage or a gain setting may be required with an accuracy to within (say) 1% of actual value over a 32:1 range. In that case, if the least *distinguishable* bit is always less than 1/128 (2^{-7}) of the actual signal, a 12-bit converter can handle a signal range of 32:1 (2^5), or about 30dB, with the required resolution and accuracy, even without the traditional ±½LSB accuracy spec.

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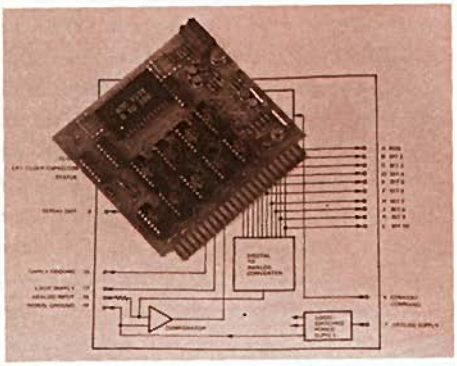
*For a data sheet describing both the AD7520 and the AD7521, request **N10**.

3 NEW A/D CONVERTERS

ADC1123: 10-BIT CMOS LOW-POWER DESIGN, <600μW QUIESCENT, 76mW @ 1kHz



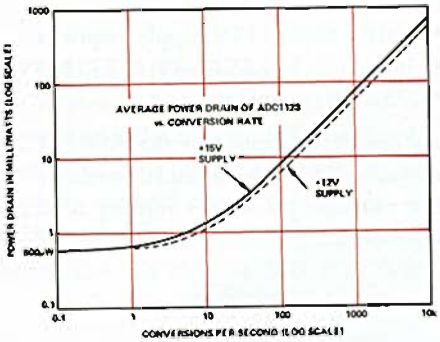
The ADC1123* is a 10-bit analog-to-digital converter designed for low power drain, both quiescently and while converting. The use of CMOS logic and "power shutoff" when a conversion is not in process result in a continuous quiescent power drain of only 600μW plus 75μJ of additional energy for each conversion (with +15V supply). The resulting average power drain is plotted as a function of conversion rate in the adjoining illustration. You can see that a continuous average conversion rate of 1kHz will require less than 76mW (5mA) at +15V, while at 1 conversion per second, the power drawn is only 675μW (45μA).

A further saving of 24% of the power requirement at higher conversion rates, with little sacrifice in top conversion speed, can be obtained by the use of a 12V supply (a "natural" possibility for battery operation).

This low power consumption makes the ADC1123 especially relevant for a range of applications, from portable instruments or remote battery-powered installations to the conservation of both primary and cooling power in large data-acquisition systems.

The ADC1123 may be operated by a single +12V to +15V source, with better than ±½LSB power-supply rejection, thus permitting significant battery terminal-voltage drift with little or no adverse effect. The logic section may be separately powered by a supply in the range +6V to +15V; since the "1" level of the CMOS logic is determined by the supply voltage V_{DD}, the user can readily match the converter's logic levels to those of his system.

The ADC1123 produces both parallel and serial data outputs; it can be ordered with inputs scaled to accept any of 4 ranges. Other performance specifications include 100μs maximum conversion time (at the +12V "low" end of the supply range), ±½LSB relative-accuracy error, and no missing codes, 0° to 70°C. It is furnished as a 3.65" x 4.10" (93 x 104mm) card, at a price of $299 (1-9). ▶▶▶



*For complete information on the ADC1123, request **N11**.

ADC1109: 10 BITS IN 4μs

The ADC1109* demonstrates that a high-speed analog-to-digital converter doesn't have to be expensive. Though it performs complete 10-bit conversions in less than 4μs, this 2" x 3" x 0.4" (51 x 76 x 10mm) module has a low price of only $159 (1-9), $109 (100+).

The ADC1109 accepts 0-10V unipolar or ±5V bipolar inputs. Using the successive-approximations technique, it converts the input voltage to natural binary, offset binary, or 2's complement digital codes. Digital outputs are available in both parallel and serial form.

Performance specifications include a gain-temperature coefficient of ±30ppm/°C and a differential-nonlinearity temperature coefficient of ±7ppm/°C. Relative accuracy error is ±½LSB.

At its low price in quantity, the ADC1109 may well open up new applications in the areas of multiple-channel ("converter per channel") data acquisition and high-speed signal processing. ▶▶▶

*For complete information on the ADC1109, request **N12**.

ADC1102: 12 BITS IN 8μs, NO MISSING CODES 0° TO 70°C

The ADC1102* is a high-performance 12-bit analog-to-digital converter that performs a complete conversion in <8μs. Its specifications include relative-accuracy error less than ±½LSB and low temperature-coefficients of gain (±10ppm/°C) and differential nonlinearity (±3ppm/°C). It is guaranteed to have no missing codes over the entire 0°C-to-70°C operating range.

The input circuit can be jumper-programmed to accept an analog input of ±5V, ±10V, 0-10V, or 0-20V range. The digital output codes, determined by successive-approximations, are natural binary for unipolar inputs and either offset-binary or 2's complement for bipolar inputs. Data outputs are available in both parallel and serial form.

An external control terminal permits faster conversions to be obtained by reducing the number of bits introduced into the successive approximation. It permits the fastest overall throughput rate, yet it also provides accurate 12-bit conversion for those signals that need it.

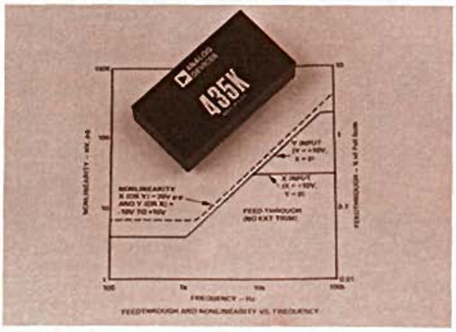
The ADC1102 is 2" x 4" x 0.4" (51 x 102 x 10mm). Its speed/cost compares favorably with those of the 15μs ADC-12QU† and the 3.5μs ADC1103-003†: $350 vs. $331 and $495 (1-9). ▶▶▶

*For information on the ADC1102, request **N13**.

†For data on these converters, request **N14**.

HIGH-PERFORMANCE MULTIPLIER, DEDICATED DIVIDER

435: 0.1% MULTIPLIER



Model 435* is a high-accuracy variable-transconductance multiplier/divider with less than 0.1% error *untrimmed* (K version). With external trims, the error may be further reduced by a factor of 3. Offset drift is less than 0.002%/°C, and total-error drift is less than 0.01%/°C. The "J" version, with initial error of 0.25% has low offset drift, less than 0.003%/°C.

Maximum nonlinearity for Model 435J/K is 0.1%/0.05%; feedthrough is 20/10mV p-p. Noise is correspondingly low, 250µV rms (5Hz to 10kHz), a significant improvement over previous designs, permitting considerably better signal resolution.

Frequency-related parameters are respectable. Bandwidth is 300kHz small signal, 30kHz full-power, with a 2V/µs slewing rate. With a 20V step applied, settling is to within 0.1% in 20µs. At 2kHz, the vector error is about 1%, and the nonlinearity and feedthrough errors are still less than 0.1%, though they increase with frequency to some extent, as shown in the illustration.

Combining high accuracy, low noise, and moderate bandwidth, Model 435 is an ideal choice for a wide gamut of 4-quadrant (±10V inputs and output in proper polarity relationship) multiplier applications, including squaring and remote control of gain. Trimmed, it can divide over a 100:1 range with <1% error.

The 435 scale factor (1/10V) can be trimmed (±1% range) by an externally-applied dc voltage. This permits remote calibration of gain in system applications. Occupying 3.07" x 1.65" x 0.65" (78 x 42 x 16.5mm) 435 is priced at $98/$145 (1-9).

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*For information on the 435, request **N15**.

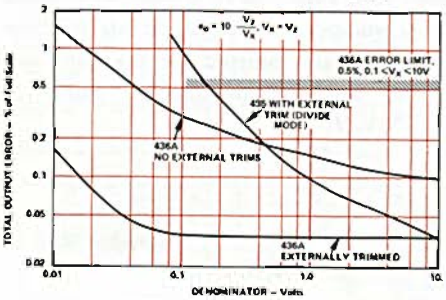
436 DIVIDER: PRETRIMMED TO 0.25%

Model 436* is a 2-quadrant variable-transconductance analog divider designed to provide accurate measurements of the ratio of its two inputs over a wide dynamic range. Available in two versions (436A, 436B), it provides accuracy to within (±0.5%, ±0.25%) over a 100:1 dynamic range (equal inputs), without requiring external trims. It can be externally trimmed to within (±0.3%, ±0.1%) and will then maintain an error less than (±0.5%, ±0.25%) over a 1000:1 (10V to 10mV) dynamic range.

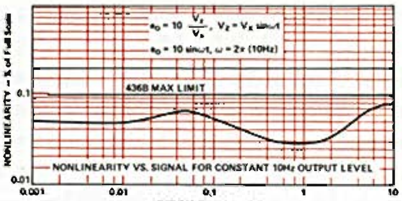
"Two-quadrant division" means that the numerator (V_Z) can be any voltage in the range ±10V, the denominator (V_X) any positive voltage $\leqslant$+10V (but $\geqslant$ the numerator magnitude). The output voltage will be equal to $10(V_Z/V_X)$ with proper polarity. The Model 434†, introduced in these pages in 1974 (8-1), is a 3-variable (YZ/X) multiplier-divider with even better dc performance, but it is a single-quadrant device (+ output and inputs).

The use of feedback enables 4-quadrant multipliers to divide in 2 quadrants, but reduction of feedback at low denominator levels reduces accuracy in proportion, so that even a device with the outstanding characteristics of the 435 (adjacent column) "runs out of gas" over wide dynamic ranges (see below)§.

Besides its excellent low-frequency performance, Model 436 has very respectable bandwidth characteristics (V_X any dc value from 0.1V to 10V): 300kHz small-signal bandwidth (-3dB), 30kHz full-power response, 2V/µs slewing rate, and 10µs settling time to ±0.5% (±10V step). This is considerably better than might be expected from either log-antilog dividers, such as the 434, or inverted-multiplier dividers, which have bandwidth that decreases in proportion to denominator voltage.



The excellent linearity of the 436 design (0.1% *max*, 0.05% *typ* for Model 436B, see illustration) makes it a useful device for controlling gain by a voltage, either remotely or automatically (AGC), with low distortion. In such applications, low noise (200µV rms, 10Hz-10kHz, and 750µV rms, 10Hz-300kHz) is also useful.



The low error of the 436 makes it useful in all kinds of ratiometric measurements—efficiencies, loss or gain ratios, % distortion, impedance magnitudes, elasticity (stress/strain), etc. Ratios may be taken of instantaneous, average, peak, or rms measures (as long as the input constraints are observed). In conjunction with sample-holds (such as SHA-1A)‡ ratios may be taken of quantities measured at different instants of time. In bridge measurements, the 436 may be used to compensate for supply variations, by the use of a denominator proportional to supply voltage.

Model 436 uses hermetically-sealed semiconductors and operates over a wide temperature range (-25°C to +85°C). Maximum error *with no external trim* is ±2% over that range, and ±1% over the restricted (but respectable) 0°C to 70°C range. Dimensions of this compact device are 1.5" x 1.5" x 0.62" (38 x 38 x 15.8mm), and price (A,B) is $75, $95 (1-9).

▶▶▶

*For information on the 436, request **N16**.

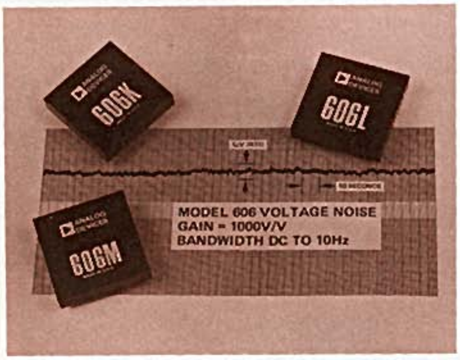
†For information on Model 434, request **N17**.

§We recommend the 540-page Analog Devices *Nonlinear Circuits Handbook* (1974, $5.95) as a useful source of information on design, applications, specification, and testing of devices such as multiplier-dividers, logs, etc.

‡For information on SHA-1A, request **N18**.

LOW NOISE WIDEBAND INSTRUMENTATION AMPLIFIER

MODEL 606M HAS 0.25μV/°C DRIFT, 1μVp-p INPUT NOISE (0.01–10Hz)



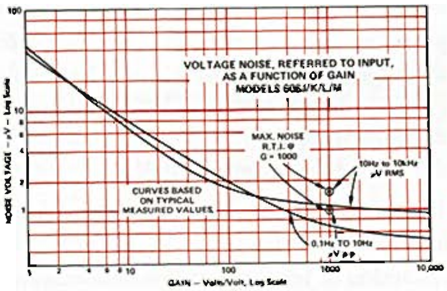
Model 606* is a high-performance differential instrumentation amplifier that combines the best attributes of fast data-amplifier designs with the accuracy normally associated with slower-responding models. This would appear to make it the industry's fastest high-accuracy module. It requires only one external resistor to set its gain, has bandwidth of the order of 100kHz over a 1000:1 gain range, and offset drift as low as ¼μV/°C *max* (referred to input, G = 1000, 606M). Nonlinearity error ≤ 0.002% (*max*, G = 100), input noise is 1μV p-p *max* (0.01-10Hz), and common-mode rejection is 90dB *min*. The unit will work well with gains up to 10,000V/V.

APPLICATIONS VERSATILITY

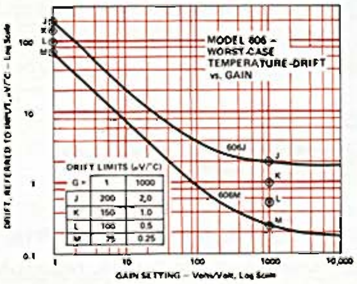
Model 606 is, of course, recommended for handling microvolt-level signals from thermocouples, strain-gage-bridge, and null-detector circuits. High CMR at substantial common-mode voltage levels permits recovery of low-level signals in noisy industrial environments. The high gain-bandwidth and fast settling suggest applications in dynamic instrumentation, such as wideband strain gages, and in data-acquisition systems. In addition to applications requiring a high degree of accuracy and stability, the 606 is of especial value where low power consumption is essential. Its 75mW power drain and ability to operate with supplies from ±9V to ±18V make it especially useful in portable recorders and measuring instruments. Its outstanding gain stability (10ppm/°C, 6ppm/month) is achieved by the use of high-precision components (including the gain-setting resistor supplied externally by the user).

LOW NOISE AND DRIFT

Low voltage noise and drift are essential when performing low-frequency differential measurements. Model 606 is excellent in both departments, as shown in the illustrations: Peak-to-peak noise (0.01-10Hz) is 1μV *max*, and wideband noise (10Hz-10kHz) is only 1.5μV rms *max*, referred to the input (G = 1000). The maximum noise in a 10Hz bandwidth thus contributes an error equal to 0.01% of full-scale output (10^{-7} x 10V).



To provide an economical choice tailored to the needs of specific applications, the Model 606 is provided in four grades, which differ *only* in drift specifications—J, K, L, M: ¼, ½, 1, 2μV/°C *max*, referred to the input, at gain of 1000V/V. The corresponding drift rates at unity gain are 200, 150, 100, 75μV/°C *max*. The curves show worst-case drift rates at intermediate values of gain, plotted for the 606J and 606M, and extrapolated to gains of 10,000V/V.

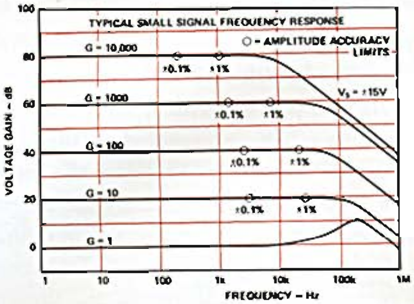


COMMON-MODE REJECTION

The 606 is designed to be applied as a high-impedance ($10^9\Omega$) differential preamplifier. In two common forms of application, it may be used either with balanced sources, such as bridges, with large common-mode voltages, or in "single-ended" unbalanced circuits, as a line-receiver to eliminate ground-loop voltages. It excels in both applications, having 120dB of common-mode rejection with balanced sources, and a guaranteed 90dB with 1kΩ source unbalance (G = 1000, 0-100Hz, V_{CM} = ±10V). CMR is essentially "flat" from dc to 100Hz at all values of gain (1-10,000).

FAST RESPONSE

Bandwidth (-3dB) varies little with gain, as the illustration shows. Approximately 80kHz at G = 1000, it increases to 200kHz at G = 10. Full-power response and slewing rate are 12kHz and 0.8V/μs, independently of gain, The output settles to within 0.01% of the final value of a 10V output step in 50μs, and to within 0.1% in 30μs.



SENSE & REFERENCE, AND OTHER MATTERS

Reference and Sense terminals are provided to allow flexibility of the output function. For example, an offset voltage can be applied at the reference terminal (at low impedance) to shift the output level, irrespective of gain. The ±5mA (*min* @ ±10V) output current can be boosted, through the use of an inside-the-loop follower, by connection of the Sense terminal to the booster output.

Available from stock in a low-profile 2″ x 2″ x 0.4″ (51 x 51 x 10.2mm) module, Model 606 is provided in 4 grades (see "Drift"). Prices are (1-9) – J, $69; K, $85; L, $98; M, $150.

▶▶▶

*For further information on Model 606, request N19.

THIN FILMS: RESISTOR NETWORK, 2-FILM SUBSTRATES

AD1804 ATTENUATOR 0.001–1 IN 9 STEPS

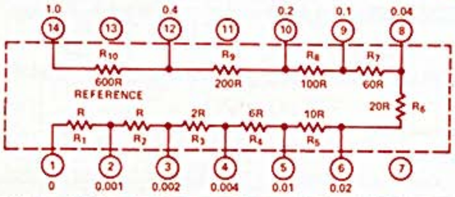
The AD1804* is a resistive divider consisting of 10 accurately-matched-and-tracking nickel-chromium resistors on an alumina substrate, enclosed in a hermetically-sealed dual-in-line or flat-pack package.

The resistance values are chosen to give three decades of partial-BCD attenuation: 1/10, 2/10, 4/10; 1/100, 2/100, 4/100; and 1/1000, 2/1000, 4/1000, with initial ratio errors to 0.01% and temperature tracking to 1ppm.

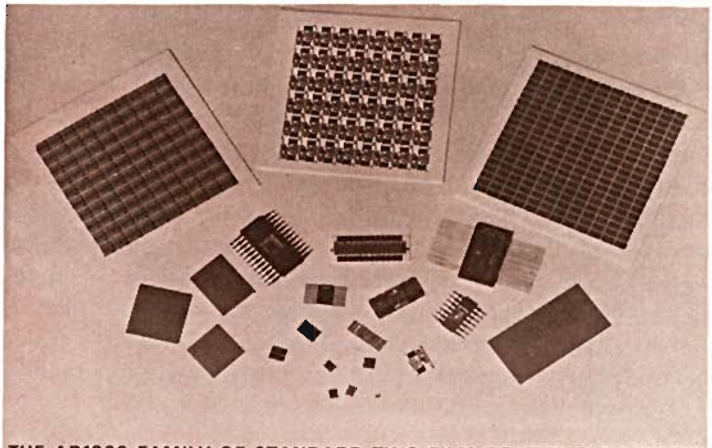
The network is designed to provide stepped attenuation (or gain–used in the feedback path of an op-amp follower) in instrumentation, calibration, and manual- or automatic-ranging circuits. Besides the values given above (obtainable as a simple 3-terminal divider), up to 50 integral multiples of 10^{-3} are available *differentially*, e.g., 1, 2, 3, 4, 6, 8, 9, 10, 16, 18, 19, 20, 30, 36, 38, etc.

CMOS switches or multiplexers, such as the AD7500 series, † may be used for remote and/or digitally-controlled switching.

Standard available options include initial ratio errors from ±1% to ±0.01% and temperature ranges 0° to 70°C and -55° to +125°C. Though the absolute magnitude of resistance is unimportant in most applications, the reference resistor is available with ±1% tolerance. Total resistance of the standard unit is 1MΩ. "High-Rel" versions to MIL-STD-883 are available. Custom designs, with other ratios or resistance values, are available, as are a variety of other standard general-purpose networks§. Prices for the AD1804 are as low as \$6.20 in 100's. ▶▶▶



STANDARD COATED SUBSTRATES USE NiCr 0.005%/YR STABILITY SPEC,-50dB NOISE



THE AD1900 FAMILY OF STANDARD TWO-FILM COATED SUBSTRATES

Analog Devices has, for some time, supplied gold-nickel-"nichrome" 3-layer thin-film substrates to customers who perform their own circuit-etching, dicing, assembly, and test. Although we've publicly discussed the two-film process that has been used[1,2] with excellent results (performance and reliability) for manufacturing our own precircuits, circuits (both chips and packages), and hybrid assemblies, we've been somewhat diffident about offering 2-layer substrates, because of the extra bonding strength afforded by the extra nickel layer to better survive the additional handling they encounter in the harsh, cruel, outside world.

However, the 2-layer process has so well proved itself, we are now offering these substrates for sale to customers who make their own thin-film resistor networks. (Of course, substrates with a third 100-200nm nickel barrier layer are still available.)

The 2-film system consists of a NiCr resistive coating covered by a 3.8μm gold conductive layer. Surface finishes of either 0.08-0.1μm or 0.13-0.25μm are available. A back-side gold conductive coating (for shielding) is optional.

The AD1900† series 2-film coated substrates may be specified with sheet resistivities from 50-500Ω/square, excess noise to -50dB, power densities to 50W/in^2 (7.8W/cm^2), long-term stability to 0.005%/year @ 25°C, and substrate sizes from 0.025" (0.064cm) square to 3.5" (8.9cm) square (of which 3" (7.62cm) square is usable). The substrates come pre-cut or pre-scribed in three material classifications: MIL, commercial, and prototype, for design and cost flexibility.

The number of standard variations amounts to about 1200. For example, the AD1922, available from stock, has 250Ω/square sheet-resistivity, 25 ±2 mil (0.064 ±0.005cm) substrate thickness, 0.13 to 0.25μm of surface finish, and a gold backside coating. It is priced at \$3.60/in^2 (\$0.558/cm^2) in 100in^2 (645cm^2) quantity.

Critically-important film adhesion-strength is a result of a tightly-controlled process, together with a quality-control system (one of the few) that can be certified to MIL-STD-9858A. High-volume batch processing permits rapid deliveries of realistic production quantities.

Typical applications include hybrid assembly (as networks and/or mother-boards) for avionic and commercial equipment. The substrates may be bonded with either conductive or non-conductive epoxy die-bonding, and conventional ultrasonic or thermocompression wire-bonding techniques. ▶▶▶

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*For technical information on the AD1804, request **N20**.

†For a technical brochure on CMOS switches, request **N21**.

§For information on other standard R networks, request **N22**.

†For information on 2-film substrates, request **N23**.

[1,2] See "High-Precision Thin-Film Resistance Networks" in *Dialogue* 8-1, and "Comparing Thin Films for Precision Resistor Networks" in *Dialogue* 8-2.

Application Briefs

THE ADC1105 IN INDUSTRY

CONVERTER'S PULSE-TRAIN OUTPUT PERMITS NUMERIC DISPLAY OF TWO VARIABLES AND THEIR DIFFERENCE IN PHYSICAL UNITS

by Dean Brodie

In industrial processes, the difference between two process variables is often a measure of subsystem performance. Examples: the difference between inlet and output temperature of a heat exchanger measures heat transfer, the difference between the suction and the discharge pressure of a centrifugal pump measures pumping ability, and the pressure differential in a Venturi is a measure of fluid flow. Thus, it is often desirable to measure the outputs of two transducers, display their magnitudes (with appropriate physical units), compute the difference, and display it in appropriate physical units. The new ADC1105* analog-to-digital converter can be readily used as the heart of a subsystem which handily accomplishes all four tasks.

The dual-slope ADC1105 is designed to be used with external counters and registers. Each conversion produces a string of output pulses, of number proportional to the analog input voltage. Figure 1 shows the ADC1105 as a building-block of a parallel-output converter.

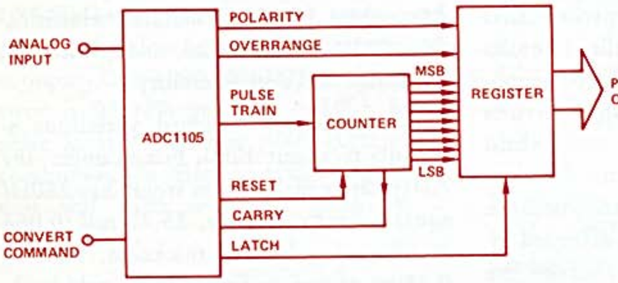

Figure 1. Basic converter block diagram

The conversion cycle begins when a Convert Command, generated either internally or externally, is applied. The counter is reset to zero, input integration begins, and counter pulses are generated. When the counter reaches full-scale and overflows, a *carry* signal, sent back to the ADC1105, stops signal integration and starts reference integration, in the opposite direction. When the integrator voltage reaches zero, the pulse train stops and the accumulated count is latched into the output register. Polarity is determined at the end of signal integration; "overrange" is generated during reference integration, and becomes valid at the end of conversion.

Figure 2 shows configuration, which converts two analog inputs to digital form and computes their difference, using the ADC1105.† The cycle begins with the input switch at channel B. A conversion is performed, and the primary-counter output at the end of conversion is strobed into register B. A second conversion is then performed with the input switched to A. Since A ⩾ B, the reference integration will require at least as many counts for A as for B. At the time, during the reference integration, when the output of the primary counter just equals the output of register B, the digital comparator actuates the difference-count-enable latch, and the difference counter starts its count. At the end of the second conversion, both register A and the difference register are strobed, and the difference count enable latch is reset. The output of the difference counter exactly equals the difference between outputs A and B.

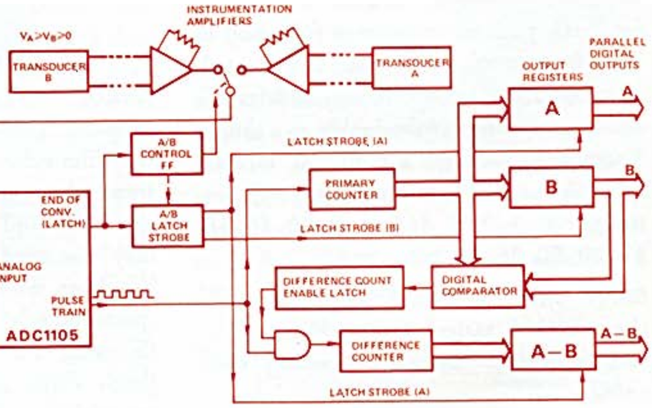

Figure 2. A, B, (A – B) converter configuration

Figure 3 shows a more-detailed version of this scheme, to remotely display the suction, discharge, and differential pressures of a centrifugal pump. We assume that the suction pressure range is 0–99lb/in^2, and the discharge pressure, always ⩾ suction pressure, varies from 0 to 999lb/in^2.

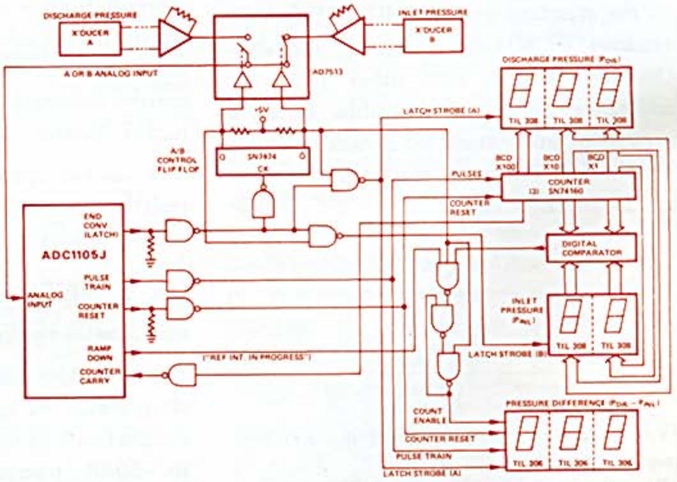

Figure 3. Pump-monitor example showing digital readout.

continued on page 18

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*For complete information on the ADC1105, request **N24**.

†Why not use analog differencing and 3 multiplexed conversions? If the signals are noisy, it is better to compute their difference *after* integration. This scheme also permits a 50% higher throughput rate, by providing all outputs in 2 cycles instead of 3.

CURRENT INVERTER WITH WIDE DYNAMIC RANGE
A USEFUL ADJUNCT TO INTEGRATED-CIRCUIT LOG DEVICES

by Barrie Gilbert

The circuit described here provides an output current of opposite polarity to the input current. The output current can cover a range substantially greater than 4 decades without severe degradation of accuracy. The circuit is best-suited to drive a low-impedance load, such as an op-amp summing point.

The preferred direction of input current flow for logarithmic devices based on NPN "Paterson diode" transistor circuitry[1] is positive (from the signal source to the log-amp summing point). Unfortunately, many transducers (e.g., photomultipliers, ion chambers, chromatographs) provide a *negative* current flow, which calls for an additional inverting amplifier.*

The conventional current-to-voltage inverter (Figure 1) loads the log-amp summing point, increasing the overall "noise gain," and causing the offset voltage of the log amp to limit dynamic range at the low end (from 6+ decades down to about 3 decades). With a slight circuit change, it can be rewired as a current inverter (Figure 2), increasing the output impedance and unloading the log-amp summing point. However, dynamic range is still limited by the linear I-to-V conversion: the inverter's output rating at the high end and its V_{OS} at the low end limit the practical range to about 3 decades of input current.

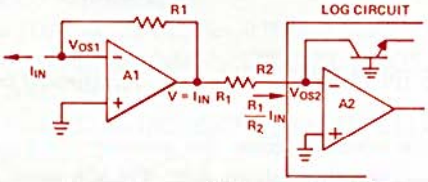

Figure 1. Current-to-voltage sign inversion. Offset error is $(|V_{os1}| + |V_{os2}|)/R_2$, adding to the signal and causing "nonlinear" logarithmic conversion at the low end.

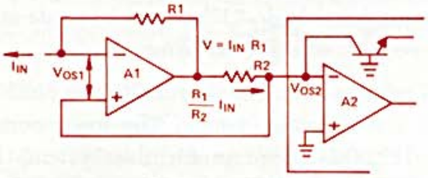

Figure 2. Linear current inverter. Offset error, V_{os1}/R_2, adds to the signal, causes nonlinear conversion for low I_{IN}.

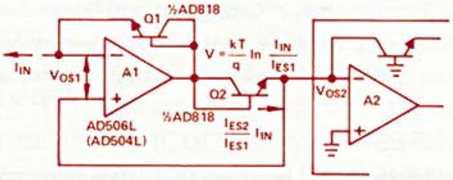

Figure 3. Logarithmic current inverter. Amplifier offset V_{os1} and V_{BE} mismatch errors affect scale factor only, result in an offset error following log conversion, independent of I_{IN} level.

Ideally, one might desire a resistance pair that changed with current, being low at high current and high at low current. Figure 3 shows such a resistance pair. The inverting-amplifier output voltage changes by a fixed amount (about 60mV) for each decade of input current change. An input/output current range of 10^6, for example, ideally requires an intermediate output voltage change of only 0.36V. Diode-connected transistor Q1 converts I_{IN} to a logarithmic output voltage; and Q2, with an (ideally) equal voltage drop, converts the voltage back to a reverse current outflow, equal to I_{IN}.

SOURCES OF ERROR

1. V_{BE} mismatch in the log transistors introduces a scaling factor equal to I_{ES2}/I_{ES1}, or, in terms of the measured ΔV_{BE}, $\exp(\Delta V_{BE_0} q/kT_0)$. Note that this factor is not fundamentally current- or temperature-dependent for well-matched and isothermal transistors, such as those in the monolithic AD818).

2. V_{os1} (the inverting op amp's offset) introduces a scale factor equal to $\exp(V_{os1} q/kT)$. It drifts with the op amp's thermal drift, at about 40ppm/μV in the vicinity of +25°C.

3. I_{os1} (input offset current) of the op amp is translated as an offset error. Like the bias current of A2, it will be the major limiting parameter to operating range at the low end.

4. Because the salient factor in this circuit is *compression of dynamic range* (approximate log behavior would suffice), R_{EB} (emitter bulk resistance of the NPN's) doesn't seriously affect operation, since the current densities—and therefore the excess of emitter voltages (over the log term)—are nearly equal. Scale factor is determined at low I_{IN} by ΔV_{BE} and V_{os}, and at the milliampere level by R_{EB1}/R_{EB2}. This error, and the distortion of log behavior at low I_{IN} due to low β, can both be unimportant if the transistors behave alike and are thermally intimate.

IMPLEMENTATION

Almost any op-amp-dual-transistor pair would work well. A good cost/speed/accuracy compromise for wide-range I_{IN} would involve the AD506L and the AD818.† The AD818 provides good initial match and low R_{EB}. The FET-input AD506L's low bias current permits wide dynamic operating range; its low drift (10μV/°C *max*) provides a reasonably-constant scale factor. With its fast slewing rate (3V/μs *min*), speed is limited principally by small-signal behavior, even for a many-decade step change. For narrower ranges, the AD504L's† low V_{os} drift gives better gain stability vs. temperature. Fig. 3 is still preferred to Fig. 2, because the required slewing rate is at least an order of magnitude smaller for a full-scale step, which befits AD504.

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[1] An extensive treatment of logarithmic circuitry can be found in the Analog Devices' *Nonlinear Circuits Handbook*, available for $5.95, P.O. Box 796, Norwood, MA 02062.

*Most log modules are available in two "sexes," e.g., 752N & 752P, 755N and 755P, 756N & 756P (request **N25** for information), but a user employing this approach will not have to stock both types. More important, the earliest IC log devices have been limited by available technology to the use of NPN log transistors.

†For data on the AD818, AD506, and the AD504, request **N26**.

Potpourri

MORE AUTHORS

Barrie Gilbert, Staff Design Engineer at Analog Devices Semiconductor ("Non-Loading Current Inverter with Wide Input Range", p.17) received the H.N.C. (Higher National Certificate) degree in applied Physics from Bournemouth (England) Municipal College. He has worked on speech coding and synthesis for the U.K. Government, sampling oscillography for Mullard, Ltd., and high-speed oscillography and other advanced measurement systems for Tektronix, where he developed many of the Company's early integrated circuits for oscillographic systems, including the transconductance multiplier (for which he won an ISSCC "best-paper" award). After additional IC design work at Plessey Semiconductors, he joined Analog Devices, where he is doing advanced development work on a variety of linear function circuits and converter products. He is the author of over two dozen technical papers and has been awarded several patents.

Richard Frantz ("Voltage Measurements in Biomedical Research", *Dialogue* 8-2) is a product marketing engineer at Analog Devices Semiconductor. Prior to joining ADS, Rich did marketing work for a small instrument manufacturer. He holds a B.S. in Electrical Engineering from Clarkson College of Technology.

Dean Brodie ("The ADC1105 in Industry", p.16), a marketing specialist with the Modular Converter group at Analog Devices, is a Cornell University E.E. graduate. Prior to joining ADI, he learned a great deal about instrumentation from the user's point of view as a U.S. Navy officer responsible for the operation and maintenance of submarine nuclear-propulsion plants.

ERRATUM – *Analog Dialogue* 8-2

We "blew" the first paragraph, p.19. It should read:

> How it works: Since the voltages appearing at the positive inputs of A1 and A2 must be duplicated at the negative inputs, the current through R0 is determined by the difference voltage, V_{IN}, which is magnified by the sum of the resistance ratios, to appear as the difference voltage $V_A - V_B$. Since the common-mode voltage appears at both input terminals, it cannot cause current to flow through R0; therefore, the output common-mode level is the same as the input common-mode level.

ON MODULAR POWER SUPPLIES

As we noted in a recent issue, "Just about every application for the products described in these pages requires a power supply. Yet the glamour of some of our other devices makes it easy to forget that Analog Devices also provides a line of power supplies . . ." With the publication of this 8-page guide, "Modular Power Supplies",[*] the task of remembering is made easier. Besides listing the product line as of late 1974, the guide also introduces 7 new models with some interesting features:

- Models 952 (±15V @ 100mA), 955 (+5V @ 1A), 970 (±15V @ 200mA), and 971 (±12V @ 240mA) are chassis-mounted types with easy-to-use terminal-strip connections replacing the printed-circuit-board pins.
- Models 922 (+5V @ 2A) and 925 (±15V @ 350mA) are high-power metal-cased types using thermal-barrier isolation between the heat-dissipating elements and the heat-sensitive elements for compactness and long life.
- Model 902-2 is a low-profile 7/8"H (22.2mm) version of the popular 902 (±15V @ 100mA).

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ADC1105 IN INDUSTRY

(continued from page 16)

Three decade counters form the primary counter; a positive edge-triggered flip-flop selects A/B, TIL308 numeric LED displays (with latches) are used for registers A and B, and TIL306 numeric displays (including counters & latches) comprise the difference counter and register. With identical inputs, the digital comparator produces logic "1". It can be made at low cost by *nor*-ing the outputs of eight *exclusive-or* gates.

The 0.01% relative accuracy and 1:20,000 resolution of the ADC1105K are not needed here. The lower-cost ADC1105J (0.1% and 1:2,000, adequate for this system) is used. The AD7513[§] dual SPST analog switch was chosen because of its TTL compatibility and low price. Its 55Ω *on* resistance is in series with the much greater (100kΩ/V) input impedance of ADC1105. Both switch-resistance temperature-drift and differential resistance between switches are negligible. Pages 3–5 & 14 discuss 2 possible choices for the instrumentation amps.

OTHER USES

The flexibility provided by the ADC1105's pulse-count output can be exploited to solve a host of additional instrumentation problems. For example, with appropriate choice of counters, it can be used to produce outputs scaled directly in engineering or physical units; if the count is preset at the beginning of reference integration, digitally-programmed offsets can be applied. Ratiometric capability (for such ratios as efficiencies, attenuations, etc.), and excellent temperature stability contribute to the usefulness of this device.

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[*] For your copy, request **N27**.

[§] For information on CMOS switches, request **N21**.

Worth Reading

CONTENTS OF THE LAST ISSUE, BOOK REVIEW, HANDBOOK ERRATA

LAST ISSUE OF *ANALOG DIALOGUE*
Volume 8 (1974), Number 2

If you haven't seen the last issue of *Dialogue*, you can get a copy by requesting **N28**. Here's what you've missed:

- *12-Bit Integrated-Circuit D/A Converter (AD562)*
- *Specifying and Measuring a Low-Noise FET-Input Op Amp (AD514)*
- *Comparing Thin-Films for Precision Resistor Networks, NiCr vs. Tantalum Nitride*
- High-Resolution FET-Input Op Amp (52)
- Fast FET-Input Op Amp (51)
- Power Supply: ±240mA @ ±12V (Model 921)
- Versatile DPM uses line or logic power (AD2006)
- A/D with Ratiometric Capability Can Read Out in Arbitrary Physical Units (ADC1105, see also p. 16, this issue)
- Fast FET-Input IC Operational Amplifier (AD528)
- Low-Cost Voltage Reference (2.5V) for Analog Circuits (AD580, see pp. 6-7, this issue)
- Seven Thin-Film High-Accuracy Standard Resistor Networks
- Eight New Analog CMOS Switch Arrays (5 Multiple Switches, 3 Multiplexers)
- Fast-settling DAC (0.2% in 30ns, 0.05% in 60ns, DAC1106)
- Multipurpose 12-Bit DAC: TTL/DTL/CMOS, Positive True (DAC1109)
- Low-Cost 12-Bit DAC Includes Register (DAC1118)
- Application Briefs:
  - *Handling Variable Data Rates with SERDEX: Fast to Slow, Slow to Fast, Steady to Burst, Burst to Steady*
  - *Voltage Measurements in Biomedical Research– How FET-Input IC Op Amps Can Help, a Brief Summary*
  - *Keys to Longer Life for CMOS (Protection Against Abuses)*
  - *Using DPM's in Our Own Test Equipment for Fast, Easy, Accurate Testing of Electronic Products*
  - *Design Ideas: Ways of Using the AD7520 10-Bit Monolithic Multiplying DAC*
- Handbook Errata, New (Free) Literature, Contents of 8-1
- Editor's Notes: On CMOS, A Plea for *Dialogue*, Authors
- Advertisement: The Model 440 Low-Cost RMS-to-dc Converter

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BOOK REVIEW

LOW-NOISE ELECTRONIC DESIGN, by C. D. Motchenbacher and F. C. Fitchen, Wiley-Interscience, 1973, 368 pp plus 18-page Preface.

Low-Noise Electronic Design is a practical and complete guide to the analysis and design of quieter circuits for the dc-to-audio-frequency range. The primary emphasis of the book is on noise generated by active and passive components within circuits and sensors, rather than on externally-induced or coupled *interference*, such as pickup noise from fluorescent lights, motors, etc. (See *Grounding and Shielding Techniques in Instrumentation* by Ralph Morrison, Wiley, 1967, for a good discussion of interference-noise avoidance.)

Messrs. Motchenbacher and Fitchen have organized their book in a most useful fashion, starting with a list of symbols and physical constants,* and concluding with noise data on IC's, FET's, and transistors. Each chapter includes a summary and a set of questions oriented toward the practical application of the material covered. The summaries alone provide a succinct guide to good low-noise-design practice.

Throughout the book, the theoretical is related to the practical by the use of design examples based on the parameters of commonly-available devices. An example is the discussion of the degradation of noise performance after breakdown of the emitter-base junction of a transistor, provided in the chapter on "Excess Noise in Transistors." The chapters on noise in passive components (capacitors, resistors, transformers) are particularly provocative for designers who are used to thinking of such components as "noiseless."

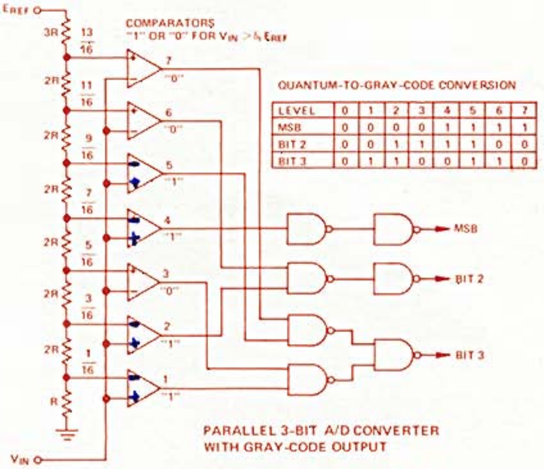
I recommend *Low-Noise Electronic Design* to anyone who uses or designs low-noise, low-frequency amplifiers. ▶▶▶

Lewis Counts

ERRATA–A/D CONVERSION HANDBOOK (1972)

This list, complete as of 12/17/74, doesn't include peccadilloes.

I-25	Figure 5c, equation under block diagram: divisor of 3rd term on r.h. should be 100.
II-26	5th line from the bottom: expression is FS/(2^n – 1).
II-29	Table 12, beside DIGITAL CODE, in both columns, Serial or Parallel
II-49	Line 6: by more bits than would be needed. . . Figure 27: NMR = -20 log (etc.) 10/T marking over heavy ordinate
II-53	Figure 29: Replace by the following:



QUANTUM-TO-GRAY-CODE CONVERSION

LEVEL	0	1	2	3	4	5	6	7
MSB	0	0	0	0	1	1	1	1
BIT 2	0	0	1	1	1	1	0	0
BIT 3	0	1	1	0	0	1	1	0

PARALLEL 3-BIT A/D CONVERTER WITH GRAY-CODE OUTPUT

II-138	9th line from bottom: . . .each step is (1 ± ½)LSB
II-142	2nd par., lines 2 & 3: . . .input signal (A/D) or count (D/A) . . .
xvii	Amplifier, isolation, I-4, 15, 82, III-31, 32, 40-42

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*There is an unfortunate and atypical typographical error in the list of physical constants (page xvii). The value of q, the charge on an electron, is missing a digit. It should be 1.602×10^{-19}C.

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