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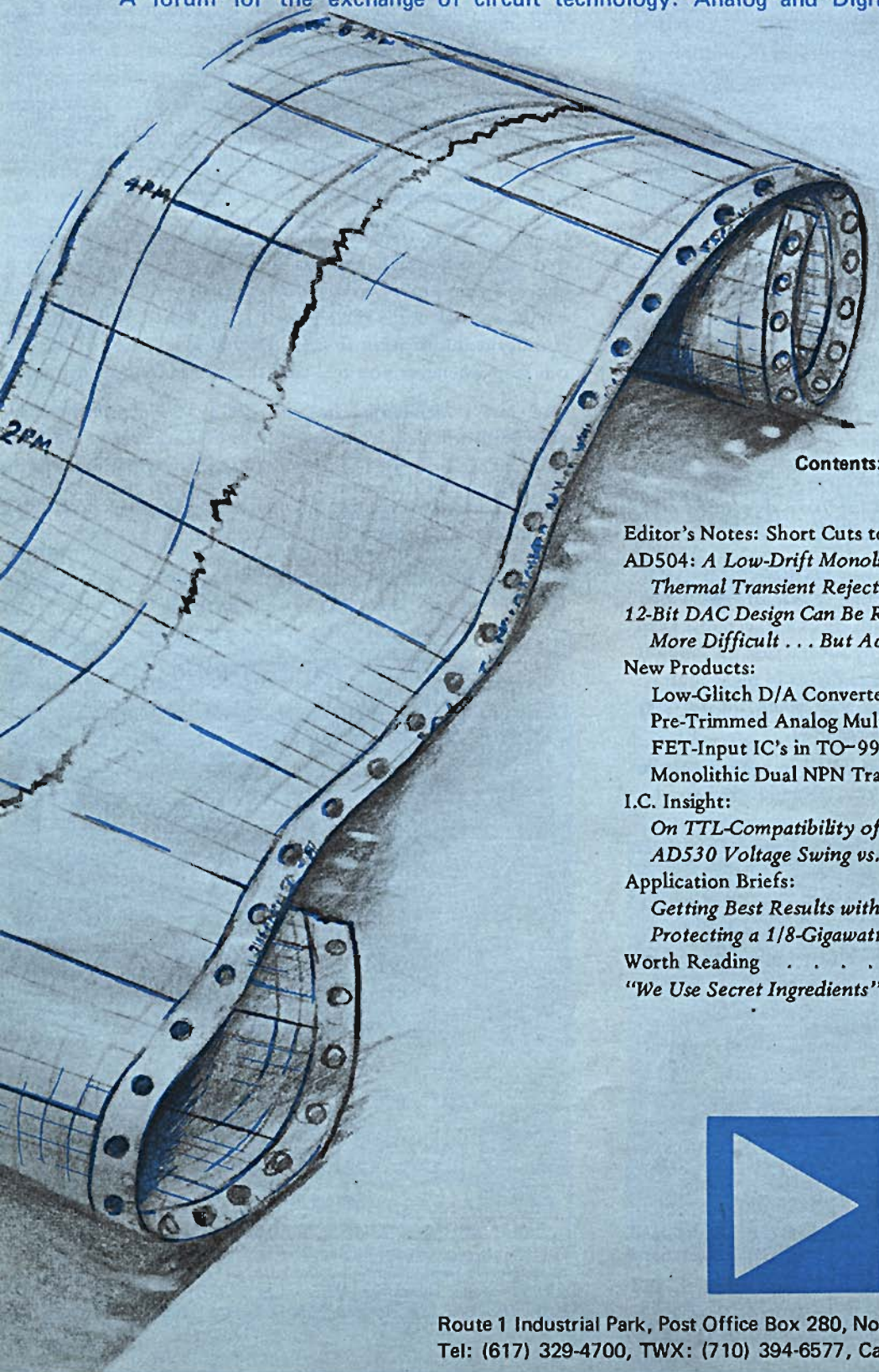

MONOLITHIC OP AMP WITH 1.0μV/°C DRIFT
See Page 3

Contents: Volume 5, Number 3, June, 1971

ANALOG DEVICES
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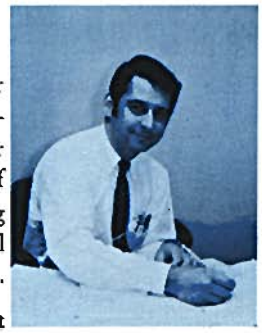
Editor's Notes

THE URGE TO SHARE TECHNIQUES



Despite the availability of computer keyboards and slipsticks, many engineers have the commendable passion for performing as much as possible of computation in their heads, relegating to machines only those computational tasks that constitute sheer drudgery.

Simple examples of problems that "everyone" solves by mental tricks are: instead of multiplying by 5, from right to left, with carries, multiply by 10/2 (i.e., add a zero and divide – from left to right – by 2); instead of multiplying by 99, multiply by 100 – 1 (i.e., add two zeros and subtract the original number); instead of multiplying by 25, multiply by 100 and divide by 4, etc.

Here are a couple of additional ideas that may lighten some of your routine tasks:

1. *Converting dynamic range from binary digits to "dB:"* Just multiply the number of bits by 6. For example, "7 bits" is equivalent to 42dB. Why?

$$\begin{aligned} \text{"dB"} &= 20 \log_{10} (\text{ratio}) = 20 \log_{10} (2^n) \\ &= 20\, n \log_{10} (2) = 20\, n\, (0.301) \\ &\approx 6\, n \end{aligned}$$

Yes, it's the same as calling for n octaves at "6dB/octave.") It quite naturally works in reverse: for 70dB dynamic range, 12 bits are needed. For bipolar ranges, remember that it may be necessary to add an extra bit for the sign (p-p vs. 0-p).

2. *Adding forward and subtracting pairs of multi-digit numbers by adding:* Most people with checking accounts keep continual track of their new balance by subtracting the amount drawn by check from the old balance. Errors in these subtractions lead inevitably to the day of grief when the account must be compared against the bank statement.

Many mathematically-minded people find it quite easy to *add* forward. For example, this sum

$$\begin{array}{r} 278.56 \\ +194.67 \\ \hline \end{array}$$

can be computed by inspection as 473.23, from left to right, just by making mental note of the carries that will be required. But somehow, if the same pair of numbers is involved in a *subtraction*, it is quite difficult to subtract forward without becoming rather confused by the "borrows."

Well, if you can add forward, you can almost as easily subtract forward, by using the same scheme computers use. You can perform the subtraction,

$$\begin{array}{r} \$278.56 \\ -194.67 \\ \hline \end{array}$$

by *adding* ($805.33 – $1,000.00) to $278.56. 805.33 will be recognized as the "10's complement" of 194.67, or the 'change you would get from a $1,000. bill. It is calculated very easily, by inspection: Just subtract each digit from 9, and add 1 to the last integer, viz., 9–1, 9–9, 9–4, 9–6, 10–7. And if you can add forward, the sum

$$\begin{array}{r} 278.56 \\ +805.33\ (-1{,}000.00) \\ \hline \end{array}$$

is, by mental calculation, $83.89. It can be checked instantly by mentally adding 194.67 and 83.89 (=278.56).

With just a small amount of practice (and your checking account will furnish almost daily opportunities), you'll never subtract the old way again, especially if you discover the simplification of subtracting forward until you need to borrow, then applying the complement to the remainder only. E.g.,

$$\begin{array}{r} 632{,}000 \\ -210{,}508 \\ \hline \end{array}$$

By inspection, you can write the first two digits (4,2), see that the next digit will be 1 (because of an anticipated borrow), then add 492 (the 10's complement of 508), to arrive at the answer, 421,492.

If you've never trained yourself to add forward, we suggest that it is well worth your while to do so. Not only does it improve your ability to calculate mentally, but the sum comes out in natural order (i.e., from left to right), with the most significant information first (and if you're just estimating, you can stop whenever you feel that the information is adequate).

And just remembering how to get a 10's complement by inspection will help immeasurably in making sure you get the right change from a $10 bill, or a 10,000-lira note (about $16).

Finally, there is the satisfaction of not being a slave to either a machine or a rote method.

Dan Sheingold

THE AUTHORS (page 6)

Cy Brown, an MIT graduate, with many years of solid experience as designer, salesman, marketer, engineering manager, is now a member of ADI's conversion products marketing staff.

Wayne Marshall has considerable experience in both communications and high-accuracy test equipment. He now designs high-accuracy D/A/D products at ADI.

ANALOG DIALOGUE

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Monolithic Operational Amplifier with 1μV/°C Drift

by M. A. Maidique

The AD504L is the first monolithic operational amplifier available with less than 1μV/°C voltage drift, single-capacitor compensation, fully-protected inputs, and capable of driving as much as 1000pF of load capacitance stably. In addition to having a low offset vs. temperature coefficient, it is also designed to have minimal response to thermal transients generated on the chip, well-behaved response to thermals generated in the outside world, negligible turn-on transients, and fast warmup.

With gain > 10^6, CMR of 120dB, offset less than 1mV (adjustable to zero), it is ideally suited for numerous low-level applications in precision measurement, telemetry, and data acquisition. Typical circuits for which it is an ideal choice include preamplifiers for thermocouple and bridge outputs, stable voltage sources, and buffering of passive circuitry.

An entirely new design, it marries thin-film resistor technology with improved monolithic processing techniques to obtain near-ideal performance from a single chip. It is packaged in a hermetically-scaled TO-99 can, and its connections are compatible with those of AD741 and similar op amp families.

WHY AD504?

The AD504* family of monolithic IC's was developed to meet a growing need for a low-drift IC op amp that is easy to compensate, has reasonable bandwidth and slew rate over the range of likely closed-loop gains, may be used with hefty capacitive loads without oscillating, can be overdriven without fear of destruction, and draws low quiescent supply current.

Their primary advantage over existing low-drift IC's is *simplicity of application:* single-capacitor compensation means that elaborate stabilizing networks, with their many degrees of freedom, are unnecessary. Ability to drive reasonable capacitive loads means that series output resistors, which degrade hard-won dc accuracy (when they are used outside the loop), are unnecessary. Protected inputs means that extra protective resistors and diodes are unnecessary. All of these factors affect the "seated" cost of using the amplifier, at all levels, from breadboarding and prototyping, to parts purchase and inventory, to

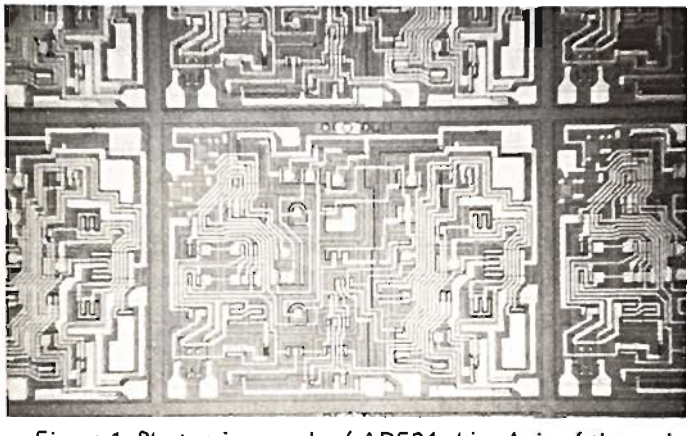
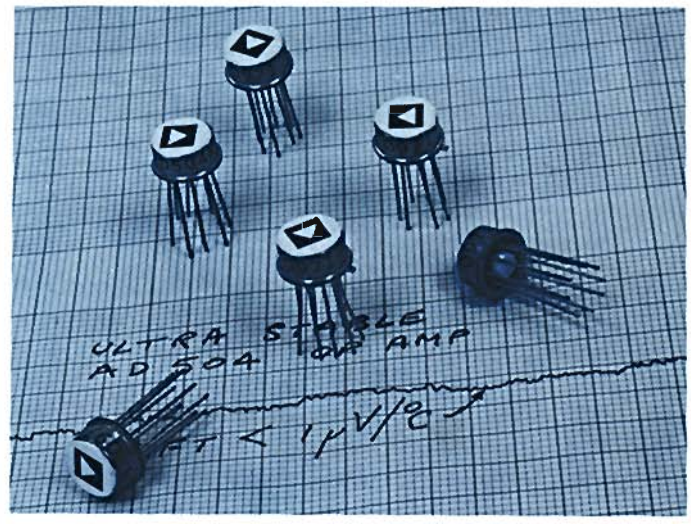


Figure 1. Photomicrograph of AD504 chip. Axis of thermal symmetry runs horizontally. Input transistors form square array at left; output transistors and drivers are grouped about the axis at right.

*For further information on AD504, use reply card. Circle C1



debugging and quality control. One might also mention the hard-won "real estate" that is preserved when those components become unnecessary.

Not only is the AD504 a clear advance over existing "725"-type IC's but it even challenges Analog's "183" discrete op amp family,* on a nearly spec for spec basis. Furthermore, the external compensation (with only one capacitor) is a real advantage. It allows bandwidth to be optimized by the *user*, as a function of closed-loop gain. (At a gain of 200, for example, response of 183 is down 3dB at 3kHz; the AD504 is within 3dB to 110kHz, with a 3.9pF compensating capacitor).

CIRCUIT DESIGN

The basic amplifier circuit (shown simplified in Figure 2) consists of differential input transistors, Q1 and Q2, with current-source loads, Q3 and Q4, followed by a PNP-Darlington common-emitter gain stage, Q5 and Q6, with a current-source load. A follower stage, Q7, Q8, and Q9, then drives the push-push output stage, Q10 and Q11. The complete circuit (Figure 3, next page) includes this basic amplifier scheme and the auxiliary circuitry used for biasing, buffering, input- and output protection.

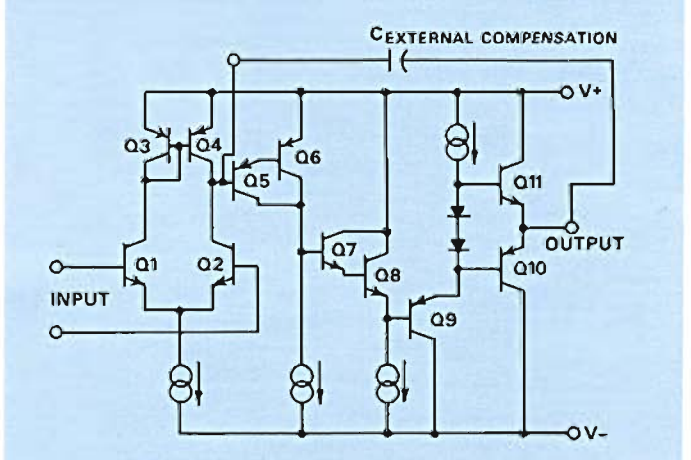


Figure 2. Basic Amplifier Schematic

*For data on the 183, as well as the still-superior 180 and 184, and the 233 and 260 chopper-stabilized types, use the reply card. Circle C2

In order to minimize the effects of temperature gradients developed within the chip, both the input and the output circuits utilize the concept of thermal symmetry. Both sides of the input stage consist of pairs of paralleled transistors (Q1, Q2, and Q3, Q4) diagonally disposed about a point on an axis passing through the "thermal center" of the output and driver stages. This point is the effective "center of thermal mass" of the input stage. As will be shown, all temperature changes resulting from variations in dissipation in the output and driver stages will be communicated along the axis, and will thus affect the input stage symmetrically.

Protection against differential input overload voltages up to ±18V (also the maximum V_s rating) is provided by the current-limiting network R1 and R2 and the diode action of Q11 and Q12. All of the input-protection circuit elements, as well as the collectors of the input transistors, are "boot-strapped" by the Q32–Q36 network, thus reducing leakage at the input, augmenting the common-mode rejection to 120–140dB, and making it feasible to consider the use of "super-β" transistors in variations of the basic design. The input stage develops high gain because of the high impedance presented by the PNP current-source load transistors, Q7 and Q8. The currents are stabilized by the use of thin-film (±10ppm/°C match) emitter resistors. Circuit values are chosen such that the input NPN transistors account for the major portion of the amplifier's temperature drift.

In addition to the split input transistors, the output and driver stages are designed to consist of 6 transistors, Q28, Q29, Q31, Q37, Q38, Q39. One of the drivers and the NPN output stage are placed directly on the axis of symmetry; the other driver and the vertical PNP are split into two sections each,* thus maintaining perfect symmetry of the power stages (where most of the temperature changes arise) about the axis. This technique allows unbalanced thermal feedback to be virtually eliminated, and it also makes possible a dc gain of the order of 10^7.

As is the case with the AD741, the second stage is basically an integrator; therefore, a single 390pF capacitor, connected externally, is sufficient to compensate the amplifier for unity closed-loop-gain applications (Figure 6). The excellent across-the-board performance of this amplifier, contained on a single 67 x 90 mil silicon chip, is discussed in the next section.

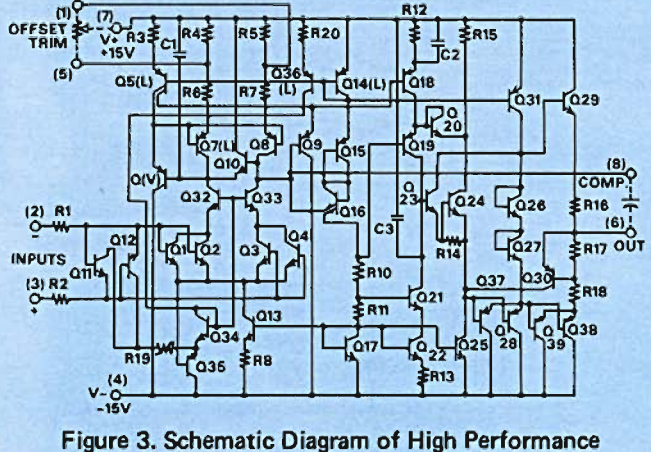


Figure 3. Schematic Diagram of High Performance Operational Amplifier (AD504)

*Patent pending

PERFORMANCE

The AD504 family feature open-loop gain greater than 10^6 at normal supply voltage, with common-mode rejection greater than 120dB. Offset voltage is less than 1mV, trimmable to zero, using a 10kΩ potentiometer. With the offset trimmed to zero, the drift is less than 1μV/°C for the AD504L, 3μV°C for AD504K, and 5μV/°C for AD504J, or 0.1, 0.3, and 0.5ppm/°C of ±10V full-scale output. Alternatively, at the cost of non-zero initial offset (which may be externally trimmed elsewhere in the system), the temperature coefficients may be trimmed to well below the zero-offset values.

At the normal ±15V of supply, the AD504K draws only 2mA of quiescent current. When operated at ±5V, the current drain is only 0.6mA, yet gain of 500,000 and full output of ±2.5V are guaranteeable.

Offset current of AD504K is only 15nA, and drift vs. temperature is less than 50pA/°C; source resistance for balanced circuits is thus 60kΩ at the "knee" of the offset vs. resistance curve. Bias current is 100nA, and its temperature coefficient is 1nA/°C (3kΩ at the "knee" for unbalanced circuits).

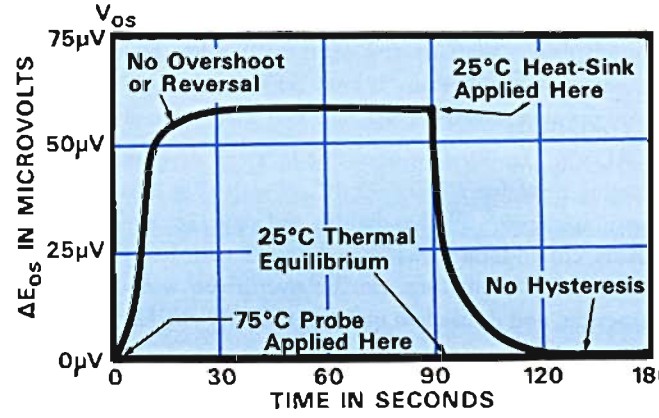


Figure 4. AD504 Response to Thermal Shock

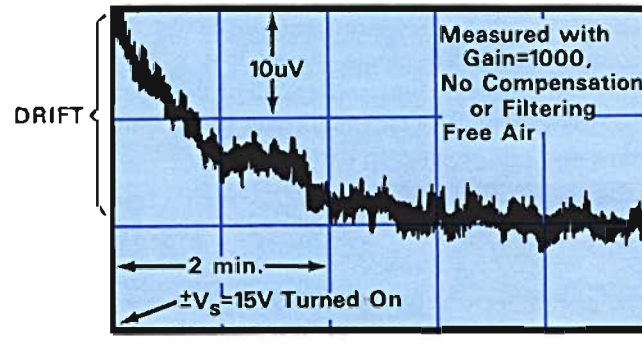


Figure 5. AD504 Turn-On Drift

Thermal transients are small and over-with quickly, as shown in Figures 4 and 5. In Figure 4, a 50°C step change in ambient temperature, applied to the can via a room temperature heat sink, then a 75°C thermal probe, and back to the heat sink, results in settling to the final value within about 30 seconds, for both increases and decreases of temperature. Note that the offset goes directly to the new value, without spikes or reversals. In Figure 5, a turn-on transient of 20μV in free air settles completely within 2 minutes. If a heat sink were used, the initial transient would be smaller, and settling would be complete within 30 seconds.

Figure 6 shows the small-signal frequency response, for both open- and closed-loop gains, for a variety of compensating values. Note that the circuit is completely stable for C_c = 390pF, with −3dB bandwidth at 300kHz; with C_c = 0, the −3dB bandwidth is 50kHz, at a gain of 2,000. More important, at unity gain (390pF), the full-power bandwidth (Figure 7) is 2kHz (0.12V/μs slew rate); at gain of 10 (39pF), it increases to 20kHz (1.2 V/μs), a considerable improvement over "725-type" amplifiers.

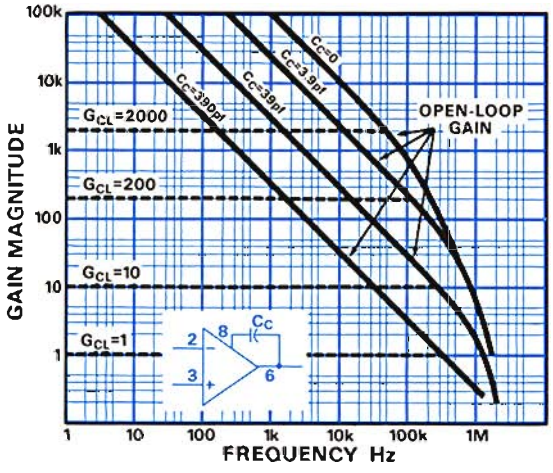


Figure 6. Small-Signal Gain vs. Frequency for AD504 with Several Values of C_c

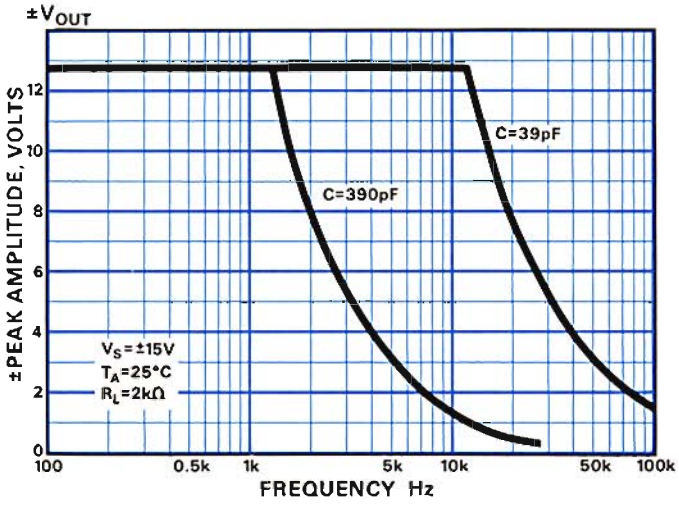


Figure 7. AD504 Output Swing vs. Frequency

APPLICATIONS

The AD504 is specifically designed for high-precision applications calling for error budgets in the parts-per-million category, such as voltage references and accurate analog-computing and data-reduction applications. Because of its low drift and noise, it is also well-suited to pre-amplification of small signals in the dc-to-audio frequency range. Its excellent common-mode rejection and protected inputs qualify it for bridge measurements, follower applications, and use in such instruments as variable-range high-input-impedance voltmeters. Its low power drain (typically 5mW with a ±5V supply) and insensitivity to temperature variations make it ideal for use in battery-powered instruments.

One interesting application is with an expensive, low-drift, temperature-compensated zener diode, to form a high-precision voltage reference (Figure 8). The amplifier maintains the zener at essentially-constant current. If the output is taken from the zener, the loading will affect the zener voltage slightly. If loaded at the emitter, the effect will be negligible, and the amplifier's low drift becomes highly advantageous.

AD504 PERFORMANCE

(Typical at 25°C, ±15V Supply)

Parameter	Typical Value
Low-Frequency Gain, Open-Loop (±10V, $R_L \geq 2k\Omega$)	10,000,000
(±10V, R_L = 1kΩ)	2,500,000
Frequency Response	
Unity Gain, Small Signal	300kHz
Unity-Gain, Full Power	1.5kHz
Unity-Gain, Slew Rate	0.12V/μs
Rise Time	3μs
Input Offset Voltage	
Initial Offset (Adjustable to Zero)	1mV
Average vs. Temperature (V_{OS} Adj. to Zero)	±1μV/°C
vs. Supply Voltage	5μV/V
Input Bias Current	50nA
Input Difference Current	5nA
Average vs. Temperature	50pA/°C
Input Noise Voltage (5Hz to 10kHz, rms)	1.0μV
0.01Hz to 1Hz, peak-to-peak	2μV
Input Voltage Range	
Absolute Maximum, Differential	±15V
Common-Mode Voltage, Maximum	±V_S
Common-Mode Rejection (±10V)	140dB
Supply Voltage, Maximum	±18V
Supply Current, Quiescent	2.5mA
Power Consumption	66mW
Output Current, Short-Circuit	±25mA

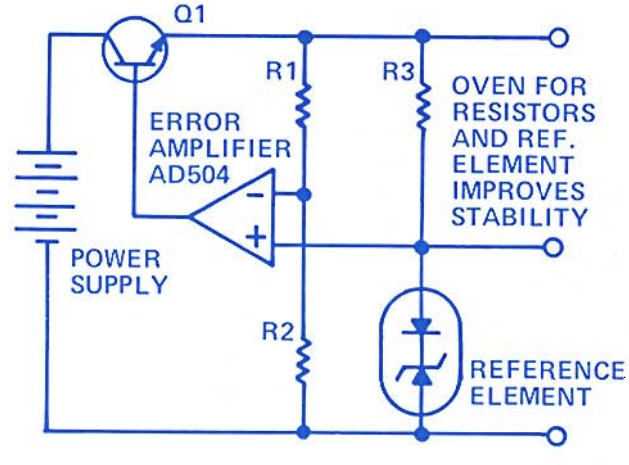


Figure 8. Zener Reference Supply*

THE AUTHOR

Dr. M. A. "Mitch" Maidique is Vice President, Engineering at Nova Devices, Inc. For further information consult Volume 4, No. 2, of *Dialogue*.

*Adapted from a figure appearing in INSTRUMENTS and CONTROL SYSTEMS– March 1971

12-Bit DAC Design Can Be Routine . . .
16 Bits is More Difficult, But Achievable

by Cyril H. Brown and Wayne Marshall

Complete D/A Converters in modular form, characterized by virtually all practicable levels of resolution, are available commercially from Analog Devices* and an increasing number of other sources in the industry. The prices of complete units are highly competitive with the costs of "do-it-yourself," especially for resolutions of 10 bits or less.

For higher resolutions, the design problems (principally obtaining switch and resistor networks having adequate matching and tracking with temperature) can be solved at reasonable cost by purchasing complete units. Nevertheless, there are many applications for which the user has a compelling necessity to build-his-own (because of special form factor, interfacing, or company policy). Until recently, the effort of obtaining selected components was a major source of cost and inconvenience.

To serve the needs of such users, Analog now makes available μDAC components for D/A and A/D Converters. These are low-cost monolithically-matched sets of switches and resistors capable of 12-bit (and better) performance, the "secret ingredients" of our complete modular converters. The μDAC's bring 12-bit converter design – to fit a wide variety of physical forms – within the province of any competent circuit design engineer, at a cost he can afford. The details of the AD555 *voltage* switches and R-2R ladder networks were discussed in *Dialogue*, Vol. 5, No. 2. The details of the AD550 *current* switches have been spelled out in a number of publications.† In the pages that follow, we summarize a number of ideas about low-frequency aspects of the use of current switches in high-accuracy D/A conversion.

WHAT 12- AND 16-BIT ACCURACY MEAN

There are few electrical instruments in routine high-volume use of which we expect and demand the order of resolution and stability nominally specified for D/A and A/D converters (including DVM's, which are BCD A/D's with visual readout). 12-bit converters are now fairly commonplace, and it is with only slight discomfort that we remind ourselves that ½LSB nonlinearity (see Definitions on facing page) in a 12-bit binary converter means 1 part in 2^{13} (=8192), or 120 ppm! And we really have to shift our thinking into overdrive to contemplate 16-bit converters, in which ½LSB amounts to 7.6ppm! Interpreted in more tangible terms, the limits of error for an ideal DAC or ADC operating at 10V nominal full-scale are

	LSB value (resolution)	½LSB (relative accuracy)
16 bits	152μV	76μV
12 bits	2.4mV	1.2mV

The 16:1 difference in magnitude can appear startling: 1.2mV is already small enough, but 76μV can be well into the noise level for many kinds of equipment. At any rate, this brief preliminary consideration may suggest that, while 12-bit resolution is just at the threshold of "multiple influence" (many individually-negligible contributions affecting error), 16-bit resolution requires control (minimization or isolation) of a very large number of influences, including not only the quite stringent tolerances on switch and resistor matching, but also the op amp voltage and current offsets, stray leakages, thermocouples in lead junctions, power supply noise coupling, and other sources of both inherent and induced noise and drift. Even milliohms of series resistance now become important, and one must look with a jaundiced eye at one's calibration standards if "absolute accuracy" is a part of the specification.

We will discuss here new circuit design techniques and components that make 12-bit performance achievable for a careful circuit designer, freeing him from arduous matching; and we will attempt to provide a suggestion as to how limitations of the basic components are overcome in 16-bit converter design (without presuming to suggest that *anyone* can – or should – design his own, because of all the many unimportant characteristics of components, subassemblies, interfaces, wiring, and system environment that suddenly become very formidable.)

ELEMENTS OF 12-BIT-ACCURATE DAC's

Figure 1 is a schematic of the critical portions of a DAC having 4-bit resolution but capable of excellent linearity. The reference supply is shown at the left. Feedback amplifier A1 stabilizes the average resistance of zener diode Z1 by exciting the bridge, Z1, R7, R8, R9 (see *Dialogue*, Vol. 5, No. 1, page 13). With a reasonably good operational amplifier at A1, and metal-film resistors for R7, R8, and R9, the supply rejection of the reference zener will be about 10^6:1, and temperature and time stability will depend primarily on those properties of the zener itself. Performance appropriate for 12-bit DAC performance is not hard to find.

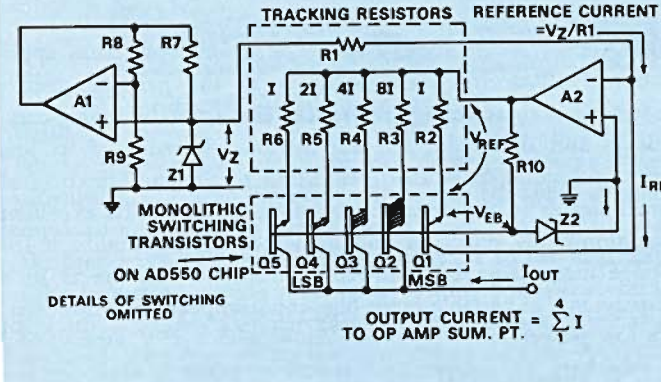


Figure 1. High-accuracy basic 4-bit D/A converter. Resistor chip also contains feedback resistor of output op amp (not shown).

*For data on Analog's complete line of D/A and A/D Converters and conversion accessories, use the reply card. Circle C3

†See: "μDAC AD550 & AD850 *Application Note*" and detailed technical bulletins on μDAC AD550 and AD850. For further information on applications of current-switching D/A Converters, see "*MINIDAC Application Note*." To receive a complete "package" of μDAC information, enveloped in a jacket outlining specific ordering information, use the reply card. Circle C4

Q2, Q3, Q4, and Q5 are switching transistors located close to one another on the chip, and also in close proximity to Q1, a reference compensation transistor. Because of the monolithic construction and identical transistor geometry, all five units are well matched, and they track well with temperature. Thin-film resistors, R1 through R6, are built on a single monolithic chip and are trimmed initially to high accuracy. They have inherently-excellent temperature tracking.

The details of the TTL/DTL-compatible switching circuitry need not be discussed here.* When the switches are open (logic "true"), output leakage current is negligible. When the switches are closed (logic "0"), they contribute binarily-weighted collector currents to the output bus, which is usually connected to the summing point of an operational amplifier. The resistor chip also contains a feedback resistor, which tracks the other resistors, for current-to-voltage conversion with minimal error.

Resistance R_1 is such that V_z produces a current of 1/8 mA through it. This is the basic reference current. Operational amplifier A2 maintains the voltage at its negative input at zero and causes I_{ref} to flow through the collector circuit of Q1. Augmented by Q1's base current, it develops a negative reference voltage across R2. The operational amplifier develops the necessary voltage to supply the baseline voltage, (via Z2) the emitter-base voltage, and the reference voltage developed by the emitter current of Q1. Because the switching transistors all share a common base line, and because their V_{EB}'s track, essentially identical voltages appear across R3, R4, R5, and R6. The resistance of R3 is 1/8 that of R2, and the emitter current of Q2 is thus 8x that of Q1 (i.e., 1mA). Because Q2 has 8 paralleled emitters, the current density is identical to that of Q1, and V_{BE} tracking is near-perfect; hence the voltage across R3 is identical to that across R2, both initially and as a function of temperature. The collector current contributed to the common output summing junction is thus very nearly 8x the reference current, since the base current (8-fold increased) tracks that of Q1. R3 tracks R2, tending to maintain the 8:1 ratio, independently of temperature variations.

Similarly, the collector current of Q3 = 4x that of Q1 (or $4I_{ref}$), and Q4 and Q5 contribute $2I_{ref}$ and I_{ref}, respectively. Thus, depending on which transistors are switched on and which are switched off, a range of precise currents from 0 to $15I_{ref}$ is available (i.e., from 0 to 1.875mA). Finally, because the feedback resistor of the output op amp also tracks R1, the output voltage depends almost solely on the stability of the basic zener reference, and not at all on the initial accuracy of the resistors, so long as they track one another.

To summarize, the absolute value of the output current *does not depend* on either the absolute value of the resistors or the parameters of the switching transistors; only the tracking, or ratio accuracy, is significant. For instance, if age or temperature should cause R2 to increase in value, then all resistors would tend to change in the same ratio, thanks to the monolithic construction. The only observable manifestation of change would be a decrease in reference current (which would be compensated for by an increase in feedback resistance around the output amplifier.

*See second footnote on page 6. See also page 11.

WHAT IS ACCURACY?

All too often, the many parameters which define the performance of a converter are lumped into something called "accuracy." We must be careful to define our terminology, especially when discussing high-performance devices. The following guidelines should be useful to those attempting to relate converter specifications to the intended usage.

Accuracy is the exactness with which any level of a D/A converter's output agrees with the reading of a working standard, which is itself traceable to a primary standard. Most commercially-sold devices are adjustable by the user to any desired zero and full-scale *absolute accuracy,* within the resolution of the adjustment, under a prescribed set of conditions.

Relative accuracy The closeness with which the ratio of the value of a point on the defining function to full scale approaches the ideal ratio. For linear functions, which are characteristic of most D/A converter types, *relative accuracy* and *linearity* are often (but not always) interchangeable. Note that the maximum analog value of the function is not "full scale," but is usually of the order of an LSB below full scale, depending on the digital code and its implementation.

Linearity is the closeness of the function at every point to a "best" straight line. Linearity within ±½LSB insures *uniqueness* and *monotonicity.*

Differential linearity is the closeness of each incremental step to the ideal 1 LSB. If the differential nonlinearity is $< |1$ LSB$|$ at all transitions, uniqueness and monotonicity are guaranteed.

Monotonicity and *uniqueness* mean that each and every consecutive step is in the same direction, that each digital code corresponds to a unique portion of the analog range (D/A's), and that all intermediate codes exist (A/D's).

Resolution. Nominal resolution ("resolution") is the relative value of the "least significant bit (LSB)" or 2^{-n} for binary devices, when n is specified by the manufacturer. It may be expressed as 1 part in 2^n, as a percentage, in parts per million . . . or simply by n. *Useful resolution* (not usually specified) is the smallest *uniquely distinguishable* bit for all conditions of operation (time, temperature, etc.). For example, a "12-bit" converter may have a useful resolution, over its temperature range, of only 10 bits.

Useful resolution is limited by the relative accuracy, but resolution need not limit accuracy. For example, the 4-bit converter used as an example in this article has only 16 levels but each step might have a relative accuracy of 0.01%, which would be useful if one were building a digitally-programmed high-accuracy power supply stepping from 2.5V to 10V in 0.5V increments. Note also that low-cost completely-monolithic "6-bit" (or even "8-bit") DAC's would not have sufficient accuracy for such an application, although their resolution is adequate.

For some inherently-monotonic applications, accuracy need not limit resolution. A good example is the integrating A/D converter, in which the less-significant bits may not be accurate, but they do provide useful resolution.

A single quad switch and resistor set covers a range of 15 increments (i.e., 4-bit resolution). If an 8-bit converter is required, one adds another set of four switches and resistors (Q2′ – Q5′ and R3′ – R6′). Its output is connected to the current output bus via a 16:1 current divider. If 12-bit resolution is needed, a third set of quads is added, with another 16:1 divider in the output, etc. Since the output of these less-significant quads is attenuated, so are their errors; hence, less expensive grades may be used. The entire *resistance* complement needed for a 12-bit DAC is available in a single package, the AD850, to accompany the three graded AD550 quad switches.

For BCD (binary-coded decimal) networks (AD851), the inter-quad attenuators are 10:1 instead of 16:1. That's all!

APPROACHES FOR 16-BIT DAC'S

For a 16-bit converter, one LSB is 15ppm of full scale! Thus, we can allow only a few ppm to account for resistor errors, switch errors, and tracking errors. The AD850L thin-film resistor network is trimmed to relative magnitudes within 0.01% or so, but the tracking is to within 1ppm/°C. A selected AD550 current switch can maintain ½ppm/°C in output linearity in terms of the V_{EB} differential between switches. With these components, (everything else being perfect), a system can be built having linearity within 1.5ppm/°C and gain stability within 2ppm/°C, plus errors due to the reference. Zener aging will be the prime cause of long term drift, typically of the order (for some readily-available units) of 5 to 10 ppm/month.

Two practical problems that arise are these: finding a suitable method of trimming the most significant bit resistors to the needed accuracy, and overcoming errors caused by series resistance and thermal voltage drops in the output leads.

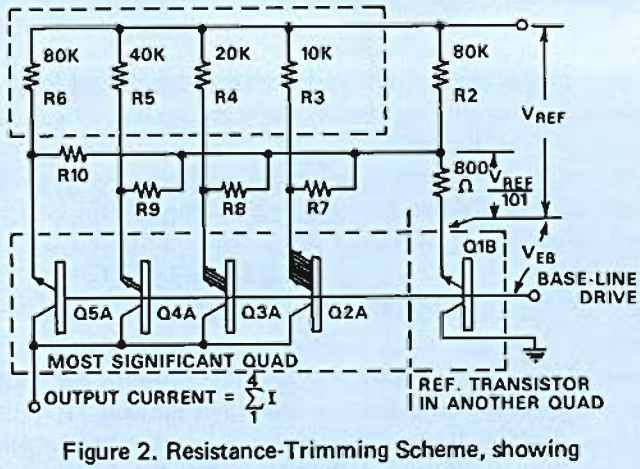


Figure 2. Resistance-Trimming Scheme, showing how practical resistance values can be used by attenuating the reference voltage

Trimming. First of all, it should be pointed out that ordinary lab test equipment, such as digital voltmeters, are almost never adequate in resolution and stability to perform as standards for the calibration. A resistance network of primary standard quality must be used, and ratiometric measurements must be applied.

It's also interesting to note that simple series or parallel trimming of the resistor network is *not* feasible. 5ppm of 10kΩ is only 0.05Ω in series or 2000MΩ in parallel, both impracticable values. Series trimming will be suitable for lower-order quads, since the resistance resolution needed will be of the order of 0.5 to 1. ohm. An interesting approach is shown in Figure 2. It is a shunt scheme in which a voltage divider is used to apply 1/101 of the reference voltage to the trim resistors, R7, R8, R9, and R10, thus magnifying them in the same ratio. In the circuit, as shown, a value of 10 megohms at R7 would cause a current change through Q2 of about 10ppm, equivalent to 1,000MΩ in parallel with R3.

External sensing Consider that, for a 16-bit converter operating with 10 volts full scale, one LSB is 152μV. If the load current is 10mA at full-scale output, and series wiring and connector resistance is only 15mΩ, the output will be in error by 1 LSB. To make output independent of lead resistance, the circuit may be slightly modified to provide sensing of the output (by the op amp's feedback circuit) at the load itself. Figure 3 shows one way of accomplishing this. For the circuit used in Analog's DAC-16QG, typical output impedance is 1 milliohm, low enough to cause negligible error.

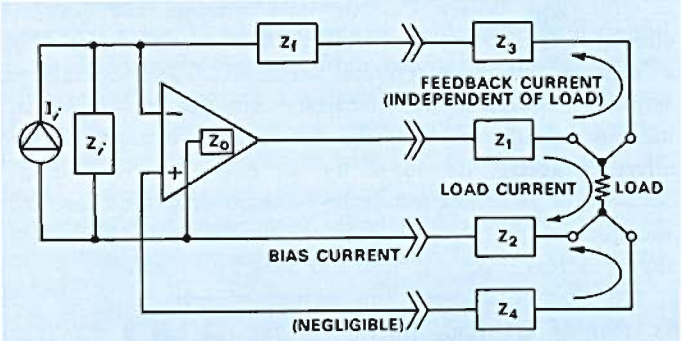


Figure 3. Sensing Scheme for making load voltage essentially independent of lead resistance and connector thermals; output resistance ($Z_o + Z_1 + Z_2$) is reduced by loop gain

The 4-terminal output configuration will have increased capacitance and inductance, which will adversely affect either loop stability or speed or both. Settling time of 25μs is not unfeasible with lead lengths of about 1 meter, in conservatively-designed practical units.

USES OF 16-BIT DAC'S

Uses of 16-bit DAC's are pretty much the same as those of lower-resolution converters, only with greater resolution and accuracy. However, there are some new applications that become feasible:

1. Testing lower-resolution converters by direct comparison
2. High-resolution histograms
3. High-resolution A/D converters to replace the combination of lower-resolution converters and programmable-gain amplifiers
4. High-accuracy programmable current or voltage sources

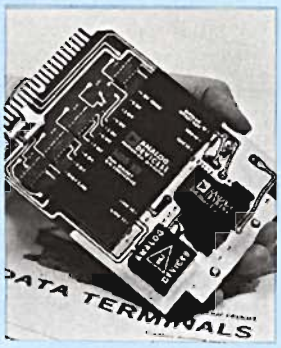
ERRATUM

In Volume 5, No. 2, of *Dialogue*, (green cover), on page 8, there is an error in Figure 3: The Z_{in} (3) and Z_o (9) terminal connections are interchanged. For multiplying, E_o (4) should be connected to Z_{in} (3), the offset trim to Z_o (9).

new products

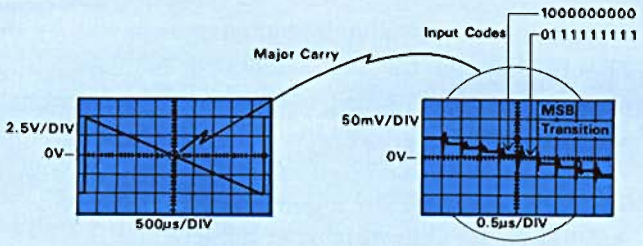
Low–Glitch D/A Converter

10-BIT UNIT IS DESIGNED FOR GRAPHIC-DISPLAY DEFLECTION SYSTEMS, HIGH-SPEED D/A SYSTEMS, DIGITALLY-CONTROLLED POWER SUPPLIES, OTHER APPLICATIONS REQUIRING MINIMAL SWITCHING TRANSIENTS.



Model DAC-10D* is a 10-bit D/A converter designed for high-speed D/A conversion with small switching transients of short duration. Settling time is 200ns (±½LSB) for 1LSB and 500ns for full scale transitions, *including* the settling time of the built-in operational amplifier. Maximum switching transients are ±50mV, only 2% of the ±2.5V full-scale output. At either ±2.5V or (optionally) ±5V, the available output current is at least ±15mA. The DAC-10D's bipolar output is obtained with either offset binary or 2's complement coding, as specified by the user. The logic inputs are compatible with both DTL and TTL. Linearity is ±½LSB, and scale factor TC is 50ppm/°C.

Low-glitch converters are principally used in deflection systems of digitally-operated CRT displays. They provide clean patterns, untroubled by the spikes that may be characteristic of fast D/A conversion, resulting from data skew at major transitions. They are also useful in programmed test instrumentation, protecting the device under test from large unprogrammed transients, and the test itself from spurious results.



Output staircase from DAC with counter input. Minimal transients are seen in magnified region.

BRIEF SPECIFICATIONS OF DAC-10D

Resolution	10 Bits
Settling Time Full-Scale Step (to ±½LSB)	500ns, max
1-LSB Transition at Any Level	200ns, max
Switching Transient	50mV, max
Output Range	±2.5V @ ±15mA
Optional	±5.0V @ ±15mA
Linearity Error	±½LSB, max
Temperature Coefficient	
Offset (Referred to Full Scale)	20ppm/°C, max
Scale Factor	50ppm/°C
Operating Temperature Range	0° to 70°C
Coding Options	Offset Binary
	2's Complement
Price (1–9)	$700.

*For further information on DAC-10D, use reply card. Circle C5

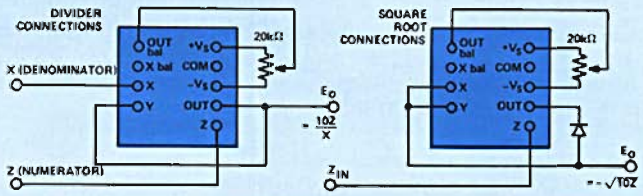
Pre-Trimmed Analog Multiplier

SMALL, COMPACT 1"-SQUARE UNIT HAS FULL-POWER OUTPUT AT 700kHz 45V/µs SLEW RATE, 1–2% ACCURACY MULTIPLIES, DIVIDES, SQUARES, ROOTS, COSTS ONLY $39(1–9), LESS IN QUANTITY

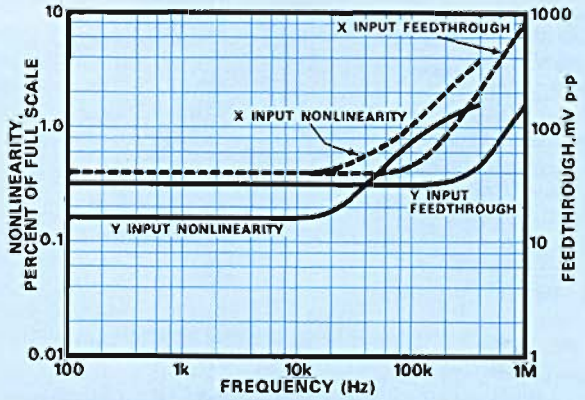
Model 432* will serve as a multiplier divider, squarer, or rooter with no external amplifiers. It is fast, offering full power output to 750kHz, slew rate of 45V/µs, -3dB frequency of 1MHz, and settling time of 1µs to within 2% of a ±10V step excursion.

Specified accuracy of Model 432J is within ±2% max, of full scale, and selected versions (model 432K) are available with ±1% performance at small additional cost. Linearity errors are small, being specified at ±0.3% for the Y input, ±0.8% for the X input.

Its small size (1" x 1" x 0.4"), low cost, and pre-trimmed assembly (no external pots are needed to meet its specifications as a multiplier) make it highly competitive with circuits assembled around partial transconductance-multiplier I.C.'s, in space, cost, and performance.



Divider and Square Root connection of 432, showing Optional Zero Adjust



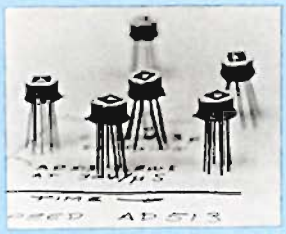
Model 432 Multiplier. Nonlinearity and Feedthrough as Functions of Frequency

*For further information on Model 432, use reply card. Circle C6

new products

FET-Input IC's Can Slew at 50V/μs

FEATURES:
- 5pA BIAS CURRENT
- 50V/μs SLEW RATE
- 1MHz FULL-POWER OUTPUT
- 90dB CMRR
- 15μV/°C DRIFT



The new AD513 family* of FET-input IC op amps in the hermetic TO-99 package are designed to provide the benefits of high input impedance, low bias current, fast slew rate and wide bandwidth for high-speed integrator, sample-hold, peak-detector, and differential-input comparator applications.

Typical dynamic characteristics, with feedforward compensation, include: slew rate of 50V/μs, settling time of 3μs, to 0.1%, -3dB bandwidth of 1MHz at gain of unity and 300kHz at gain of 100.

In addition to these excellent dynamic characteristics, the dc gain is 100,000, bias current is 5pA, and offset voltage is 10mV, adjustable externally to zero. (For internally-trimmed units, consult Analog about the AD516).

The AD513 family is available in three basic models, the low-cost AD513J ($15. in 100's), the high-performance AD513K ($17.50 in 100's), and the MIL-temperature range unit, the AD513S. Specifications are listed below.

SINGLE-CAPACITOR COMPENSATION

Although the AD513 enjoys its greatest speed with feedforward compensation (3 capacitors) and slightly less with "lead" compensation (2 capacitors), a *single capacitor* can be chosen to stabilize the amplifier for any value of closed-loop gain, with the excellent dynamic performance specified below.

BRIEF SPECIFICATIONS

PARAMETERS	AD513J		AD513K	
(T_A = 25°C, V_s = ±15V)	Typ.	Max/Min	Typ.	Max/Min
Input Offset Voltage, mV	20	50	8	20
Drift (μV/°C, 0-70°C)	30	75	15	25
Input Offset Current, pA	5	25	2	10
Input Bias Current, pA	15	50	5	25
Gain, Open Loop	100K	20K	200K	50K
CMR (dB)	90	70	90	70
Input Voltage Range (VCM)		±10		±10
PSRR (μV/V)	30	300	20	150
Output Voltage (R_L = 2kΩ)	±13	±10	±13	±10
3dB Bandwidth,† MHz				
A_{CL} = 1	1		1	
A_{CL} = 100	0.3		0.3	
Slew Rate,† V/μs,				
A_{CL} = 1	5		5	
A_{CL} = 10	20		20	
Settling Time† to 0.1% (μs)				
A_{CL} = 1	3		3	
A_{CL} = 100	20		20	
Price (1-24)	$22.50		$26.30	
100+	$15.00		$17.50	

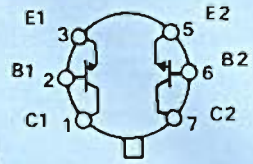
*For complete information on AD513 IC's, use reply card. Circle C7

†Single-capacitor compensation

Monolithic Dual Transistors

AD810 FAMILY ARE IDEAL FOR AMPLIFIER INPUTS, LOG CIRCUITS, ETC. ANALOG OFFERS TO SHARE ANOTHER "SECRET INGREDIENT"

FEATURES:
- NEAR-PERFECT MATCHING (0.5mV)
- LOW DRIFT (2.5μV/°C)
- HIGH CURRENT GAIN (> 400)
- WIDE RANGE OF I_c (10μA to > 5mA)



The AD810 family* of monolithic dual small-signal NPN transistors were originally designed to be (and are widely) used in Analog's operational amplifiers, especially for input stages. They were developed because duals having adequate performance simply were not available at a realistic price for large-scale OEM usage.

Now Analog has broken both the price and performance barriers, with such characteristics as differential voltage less than 0.5mV, temperature drifts below 2.5μV/°C, breakdown voltages greater than 40V, making the devices ideal for critical differential input stage requirements. In addition, the excellent current gain performance, typically 500 over a 10μA to 5mA collector current range, insures high-performance operation in second-stage applications.

In addition to their use in input stages and second stages of operational amplifiers, they are suited to a wide range of applications requiring excellent matching of characteristics, high current gain, and high voltage breakdown – including a wide variety of logarithmic device applications. They are packaged in standard JEDEC TO-71 hermetically-sealed packages and are rated for operation at temperatures from -55°C to +125°C.

They are graded into 4 basic selection categories:
- AD810 for lowest-price high performance
- AD811 for all-around excellent characteristics
- AD812 for highest current gain from 10μA to 5mA
- AD813 for the ultimate in input characteristics

BRIEF SPECIFICATIONS

PARAMETERS	AD810	AD811	AD812	AD813
(T_A = 25°C)				
h_{FE1}/h_{FE2} mismatch, max	20%	10%	10%	10%
$V_{BE1} - V_{BE2}$, max	3mV	1mV	1mV	0.5mV
$\Delta(V_{BE1} - V_{BE2}) / \Delta T$, max -55°C to 125°C	15μV/°C	5μV/°C	5μV/°C	2.5μV/°C
$I_{B1} - I_{B2}$, max	20nA	5nA	2.5nA	5nA
$\Delta(I_{B1} - I_{B2}) / \Delta T$, max -55°C to 125°C	0.6nA/°C	0.3nA/°C	0.3nA/°C	0.3nA/°C
h_{FE} min (I_c = 10μA, V_{cc} = 5V)	100	200	400	200
h_{FE} max	–	600	1000	600
h_{FE} min (I_c = 1mA)	100	200	400	200
h_{FE} min (I_c = 5mA)	85	170	350	170
h_{FE} min (I_c = 10μA, -55°C)	35	75	150	75
$V_{CE(sat)}$ (I_c = 1mA, I_b = 0.1mA)	0.5V	0.3V	0.3V	0.3V
Prices (1-99)	$2.40	$4.50	$6.75	$8.25
100+	$1.60	$3.00	$4.50	$5.50

*For data on AD810 monolithic duals, use reply card. Circle C8

I.C. Insight

Quad-Switch TTL-Compatibility

WHAT IS TTL COMPATIBILITY?

The TTL logic currently accepted as standard calls for gate *outputs* no less than 2.4V for logic "true" ("1") and no greater than 0.4V for logic "false" ("0") at 25°C. Corresponding to these levels are the usually-specified logic *input* levels: a gate must respond to levels greater than 2.0V (or input "1") or to levels less than 0.8V (or input "0").

The difference of 0.4V between input levels and output levels provides a margin of safety usually termed "noise immunity." This is shown graphically in Figure 1. If 350mV of peak noise were superimposed on a gate output level, it would still be inadequate to trip an in-spec gate.

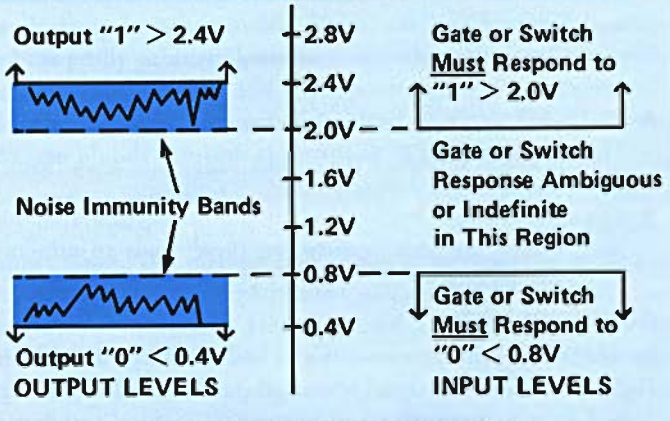


Figure 1. Standard TTL Levels at 25°C

In addition to being applicable to totally-digital systems, this principle also applies to interface components, such as A/D and D/A converters. It is highly desirable that the digital output of A/D Converters that interface with TTL systems conform with TTL output levels and that D/A Converter input sensitivity conform to TTL input levels. This is a principle that has been observed faithfully for the many years that we have been manufacturing conversion products. It is what *TTL-compatible* means.

TTL-COMPATIBILITY OF IC CURRENT SWITCHES

Naturally, we have applied it to our newer IC conversion components, and – in particular – to our AD550 (see page 6) and AD555 (see *Dialogue*, Vol. 5, No. 2) monolithic quad voltage – and current switches.

Recently, there have appeared on the market quad switches from at least two manufacturers (one known for modular op amps, the other for MOS memories), having the same pinout as, and comparable accuracy specs to, the AD550, with somewhat higher speed, and the implication that they are TTL-compatible. Yet their input specifications are 0.4V for "0" and 2.5V for "1". How compatible they are in fact is shown graphically in Figure 2.

Not only are the specifications incompatible with TTL input, but a further qualification on the "Model 4550" data sheets makes this matter of incompatibility seem even a bit more serious: "For these TTL logic levels, the emitters of the switching transistors in the ON condition must (*sic*) be at a level of −5V . . . Input thresholds track the ON-state emitter voltages . . ."

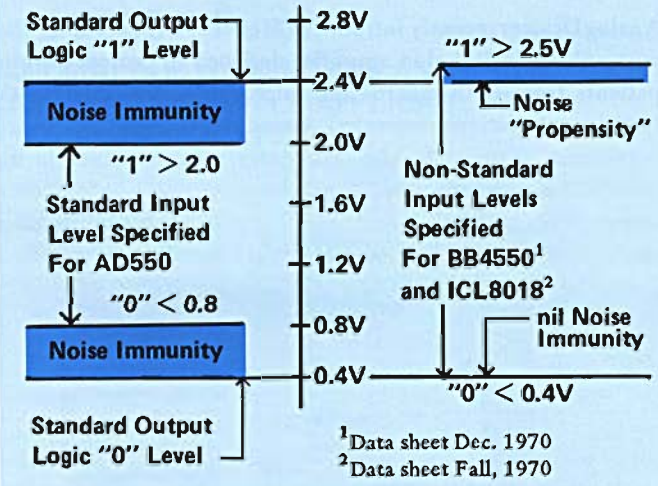


Figure 2. Input Levels of Quad Switches Compared

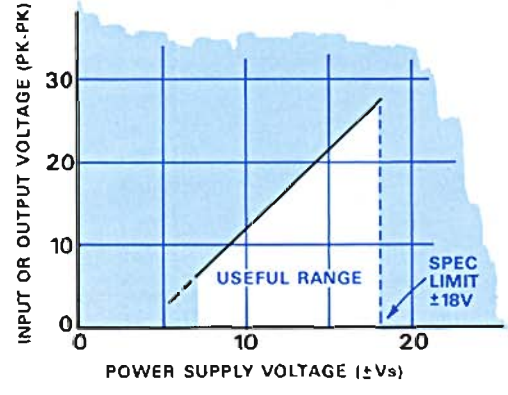
If one were looking at the matter critically, one might infer from the last statement that the marginal and precariously-balanced noise immunity would be irretrievably biased in one direction or the other if the emitter voltages were to shift.

If you use monolithic current-switching and if noise immunity is important in your specification – or if TTL-compatibility is a prerequisite, be sure to look closely at the specifications of the unit you are considering.

AD530 Output vs. Supply

The AD530 complete-on-a-chip monolithic multiplier-divider* will operate over a wide range of supply voltages, from the specified upper limit of ±18V to well below ±10V. Thanks to its internal regulator circuit, its scale factor will be essentially independent of supply voltage, except for normal supply voltage sensitivity (which will tend to increase at very low voltage).

The plot shows typical values of peak-to-peak signal-handling capacity (the smallest of the output or either input), with 2kΩ load, as a function of symmetrically-applied supply voltage.



Full-Scale Input or Output Voltage vs. Power Supply for AD530 as a Multiplier

*For complete information on AD530, use the reply card. Circle C9

Application Briefs

Good Circuit Practice Gets Best Results from Isolation Amplifiers

Analog Devices recently introduced Model 272*, a modular low-cost unity-gain isolation amplifier designed to protect hospital patients from both macro- and microshock, by isolating the outputs of patient-connected transducers from the instrumentation system. It also facilitates measurements in the presence of common-mode signals by providing 120dB CMR at 60Hz. The internal transformer-coupled shielded isolation system of the 272 is shown schematically in Figure 1.

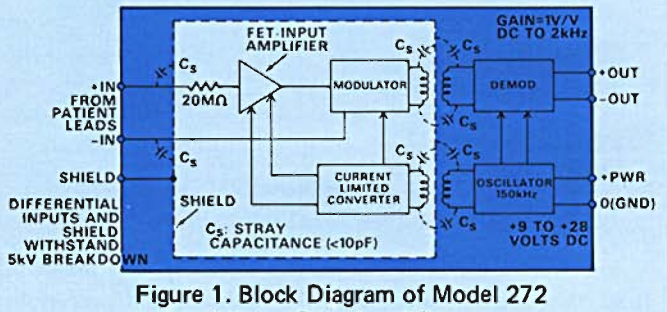


Figure 1. Block Diagram of Model 272 Isolated Unity-Gain Amplifier

SHIELDING AND GROUNDING

A major source of "noise" in instrumentation systems is the coupling of common-mode voltages and currents into the input circuit. This can occur in many ways: through ground potential differences ("ground loops"); electrostatic (i.e., capacitive) coupling from power lines and other high-frequency sources; electro-magnetic induction in closed conductive loops; unbalanced loading of the signal leads, etc.

The 272 allows a number of these sources of common-mode error to be minimized. By connecting its floating guard shield to the cable shield, one can in effect totally enclose the differential signal leads, thus protecting them from electrostatic pickup. With the shield connected at only one point (i.e. at a point representative of the common-mode potential): it acts as a guard to minimize unbalanced capacitive loading; it minimizes magnetic pickup; and it avoids introducing ground difference potentials into the instrumentation system.

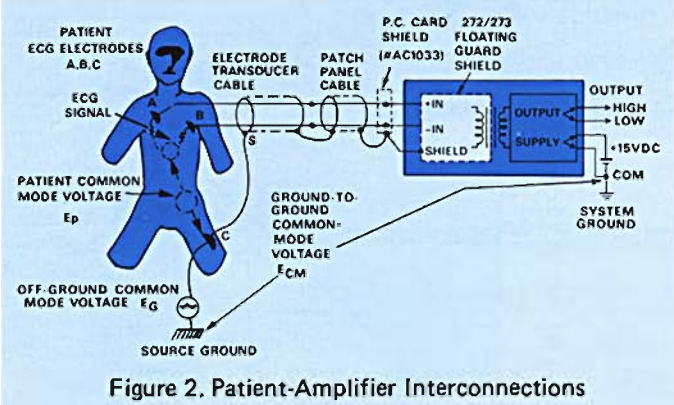


Figure 2. Patient-Amplifier Interconnections

To derive full benefit from the 272's ability to reject noise and "float" input signals, one should employ good shielding practice, as shown in the example of Figure 2. Although a medical application is shown, it is representative of other applications pertaining to guarding millivolt signals in hostile environments, such as voltages developed in current shunts in "heavy industrial" plants.

GUIDELINES

- Use twisted shielded cable to reduce inductive and capacitive pickup.
- Drive the shield(S) with the common-mode signal source (E_G) to reduce effective cable capacitance significantly. This is accomplished by connecting shield point(S) as closely as possible to the signal "low"(B). Often, a "right leg" electrode connection establishes a patient "ground"(C).
- The common-mode rejection of the 272 between the differential inputs and the shield will suppress the effects of common-mode voltages generated within the patient's body (E_P).
- To avoid ground loops and excessive power-frequency "hum," signal low(B) or the cable shield(S) should never be grounded at more than one point.
- Dress unshielded leads short at the connection terminals and reduce the area formed by these leads to minimize inductive pickup.

AC COUPLING

In addition to the common mode, biological signals often include a differential (i.e., "normal mode") offset voltage, caused by an electrochemical bias (i.e., "contact") potential, at the patient-electrode interface, subject to polarity and magnitude variations as the pressure and position of the probes change. Its magnitude is typically in the hundreds of millivolts.

It is the presence of this potential that accounts for the initial choice of design of the 272 as a unity-gain dc follower. The contact potential is usually larger than the signal being measured – therefore built-in gain would produce saturation with limited amplification. If ac components (usually 0.05 to 100Hz) are the signals of interest, an arbitrary choice of characteristics (bandwidth, recovery time, capacitor stand-off voltage) would leave many applications unsatisfied; also it would cost more and might sacrifice CMR or protective capability.

A good solution nowadays is for a general-purpose isolation amplifier to transmit the signal essentially unchanged, allowing the circuit designer to determine the type and degree of signal conditioning to be used in the circuits that follow. Figure 3 shows an example of ac coupling and adjustable gain.

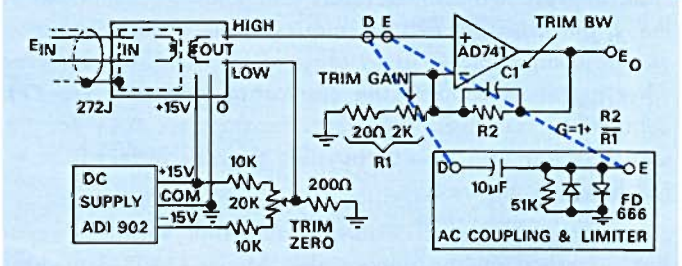


Figure 3. DC Isolation Amplifier with Adjustable Gain and AC Coupling Option

*See *Dialogue*, Vol. 5, No. 2. For further information on Model 272, use the reply card. Circle C10

Application Briefs

Protecting a 1/8-Gigawatt Power Supply (8kA at 16kV) or How Do You Make a 2000A Fuse?

FAST DATA-ACQUISITION SYSTEM DETECTS ANOMALOUS WAVEFORMS AND SHUTS DOWN 2000-AMP IGNITRONS BEFORE THEY CAN MELT. PRINCIPLE APPLICABLE TO MANY OTHER SYSTEMS WHERE FAILURES MUST BE ANTICIPATED.

A new data-acquisition technique provides rapid acquisition, real-time threshold detection, and fast, multi-option response, for a large number of continuously-varying parameters. The concept was first employed in a multi-channel system recently installed in the main power supply for the Bevatron, a particle accelerator at the Lawrence Radiation Laboratory.*

In this application, a fast data-acquisition system samples waveforms at a large number of intervals during each cycle and compares them with stored threshold values (perhaps obtained by processing earlier data). Any significant disparities are noted, analyzed, and appropriate action is automatically taken, ranging from minor adjustments, to a "flag," to shutdown.

Since such anomalies can signal incipient breakdowns, this scheme can prevent the very expensive consequences of loosing large amounts of power under fault conditions, even for very short times. The method also allows observation of aging and so can effect parts replacement at appropriate – rather than arbitrary – intervals. It also provides rather obvious means of feedback of life data to parts manufacturers, who can use the information for product reliability improvement. (It's much more difficult to diagnose the source of failure by investigation of a hardened puddle of metal and glass, or after an explosion of suddenly-vaporized coolant!)

Although most of our readers are unlikely to be designers of control systems for Bevatrons, a little thought will show that the monitoring principles discussed in this practical example are useful for minimizing risks of failure and for failure analysis in any system where catastrophic failure should be unthinkable. Such situations can range through a wide gamut, limited only by the imagination, but are likely to include steel rolling mills, supertankers, intensive-care patients, power-distribution networks, etc.

This approach is facilitated by the increasing availability of memories, minicomputers, and conversion components, and the decreasing cost of building instrumentation networks.

THE PROBLEM

The power supply is a dual-motor-generator installation, feeding a 12-phase mercury ignitron rectifier network. The peak power output of the supply is *128 megawatts: 8,000 amperes dc at 16,000 volts!* There are 48 ignitrons, occupying 8 cubicles, each containing 6 tubes. Typical anode currents are 2,000A. Pairs of tubes, in parallel, share normal conduction currents.

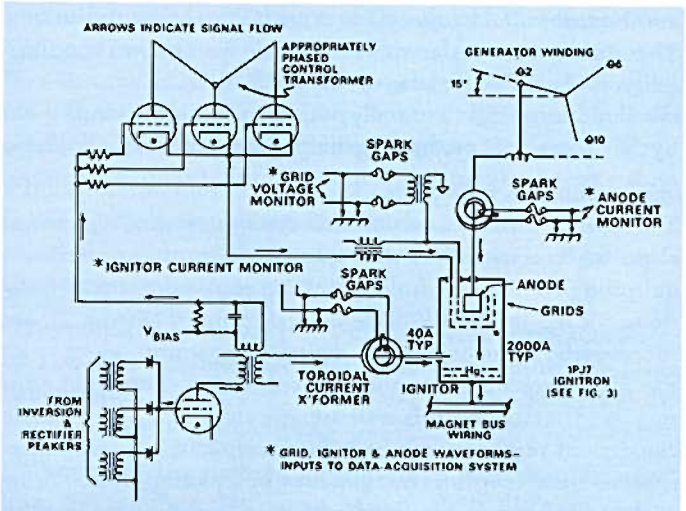


Figure 1. Ignitron Monitoring System

Unusual or spurious operation of elements of this system can result in improper sharing of current, which may result in catastrophic tube failure, due to excessive anode currents. Ignitrons require proper phasing of both grid waveforms and ignitor currents, with respect to the anode voltage. A failure in either of these control signals can prevent a tube from firing, thus creating excessive anode currents in its partner.

The data-acquisition system described here aids in the monitoring, detection, diagnosis, and correction of such occurrences.

Data acquisition, in this case, calls for quantitative measurement of the state of a series of parameters, and an immediate decision of the parameter's relationship to a threshold. This may be handled in a number of ways: by applying an analog comparator directly to the transducer, by applying

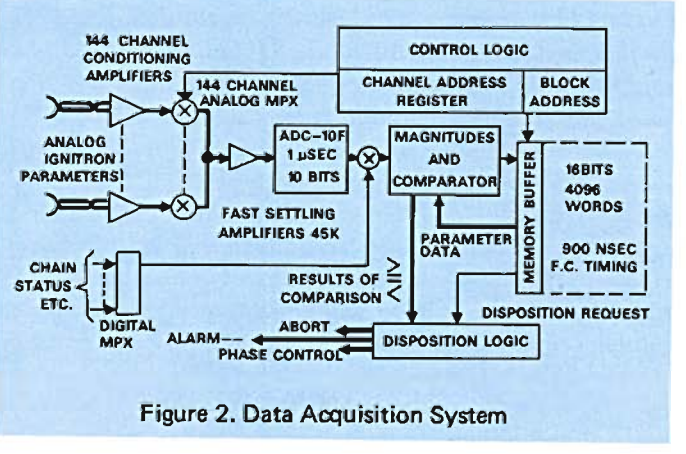


Figure 2. Data Acquisition System

THE AUTHOR

Don M. Evans, who has degrees in Electrical Engineering and Computer Science from the University of California at Berkeley, has been at the Lawrence Radiation Laboratory's Bevatron project for the past 10 years. The Digital Control Facility, for which he is responsible, includes four small computers in a "multi-processor" organization, which exercises control over the guide-field mentioned in this article, 42 200kW energy tracking magnets, a wide-band RF system for particle acceleration, and a Linac and Ion Source.

*Work performed under the auspices of the USAEC

it to the output of the conditioning amplifier, or by using a digital processor, which analyzes the data after it has been acquired and entered into the processor's memory.

Analog comparators respond quickly, but a large number of separate comparators – and possibly even separate power supplies – are required. Also, unless many D/A's or sample-holds are used, threshold adjustment is manual and time-consuming. The digital case is far more desirable because of its flexibility – but the time-lag from acquisition to detection of threshold crossings is usually large. In addition, time spent by the processor in performing comparisons reduces the available on-line time for executing far more important control functions.

The new approach used for this system allows detection as close to "real time" as the repetition rate of parameter acquisition permits. Threshold values and required response functions, stored in a block of peripheral memory (Figure 2), are successively brought into a digital comparator, along with the proximate results of the A/D conversion. Threshold crossing is detected at this earliest possible time, and simple consequent responses can be immediately initiated (e.g., interruption of succeeding firing pulses or "phasing-back" firing angles). Other responses might be to flag the occurrence by incrementing a location in the peripheral core memory, or to interrupt a digital processor, which can then compare it with many earlier data points, stored in the peripheral memory.

FAST ACQUISITION REQUIRED

Forty-eight tubes under surveillance, with three parameters per tube (grid voltage, ignitor current, and anode current), require a total of 144 channels of analog multiplexing. A record of previous data for each of the parameters is maintained, in order to properly diagnose causes of failures and to provide a histogram of fault growth as follows: After comparison of the threshold with the converted parameter value, each value is inserted into its respective location in a block of the peripheral memory. Upon completion of one multiplex cycle of 144 channels, the block address is advanced to a new block. Because ten memory blocks are used, 10 successive values of each parameter are available, allowing reconstruction of the history of each tube in the system for that cycle prior to detection of an impending fault condition.

Since the rectifier system operates at a nominal frequency of 60Hz, a tube's conduction period is about 3ms. The magnitudes of the 144 parameters are sampled 10 times during one conduction cycle, allowing 10 blocks of core to be filled with data and the comparisons to be performed, all of which requires a maximum acquisition cycle duration of 300μs. This requirement establishes a maximum duration of 2μs for each individual sequence. During this time, a parameter must be addressed, the multiplexer settled, A/D conversion accomplished, digital comparison made, and parameter value written into memory. Also, the threshold to be compared is read out of core. While a parameter is being written into memory, the analog multiplexer seeks the address of the next parameter to be accessed. Further time-sharing takes place as the threshold value is read from core and rewritten while the A/D converter digitizes the analog value and provides the results to the

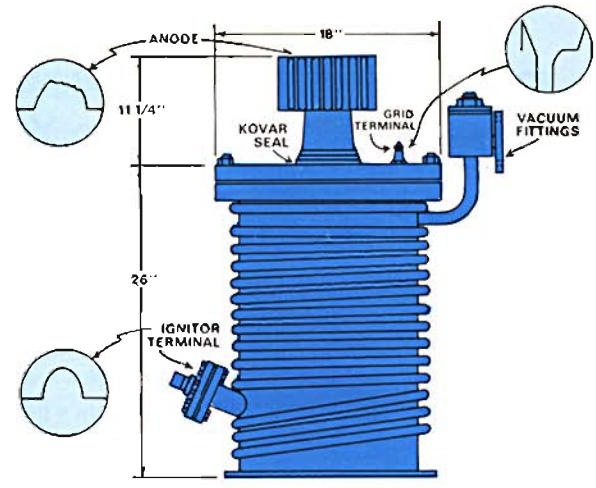


Figure 3. Mercury Ignitron 1PJ7 (One of 48)

comparator. All of these considerations determine the required operating characteristics of the modular devices. To meet the 0.1% system resolution requirement, the Analog Devices ADC-10FB*, a 10-bit successive-approximation converter with a conversion rate of 100ns/bit, is used.

DATA HANDLING

In order to accomplish the tasks required within the 2μs allowed, a core memory module with 900ns full cycle time is included, an Information Control Corporation 16-bit, 4k-word model. The parameter magnitude data occupies 10 bits of each word; the additional bits (6 most significant) are utilized as flag bits, counters for incrementing to keep track of the number of unusual occurrences, or as a code. These indicate the course of action to be taken, depending upon the results of the threshold comparison; i.e., >, =, or <. This flexibility provides for polarity and allows appropriate threshold action, whether an increase or a decrease signals a fault condition.

The net result of these design considerations is a peripheral memory bank synchronized to the conduction periods of mercury ignitrons, address-keyed to sequence through significant parameters in step with the conversions occurring in real time. Rapid cycling through a number of parameters during each repetition cycle, with occasional "side-trips" to acquire static data with no "real-time significance," allows one to obtain a histogram of data by successively inserting each block of data points into adjacent blocks of peripheral memory. Continuing this process over 10 blocks of data provides a rotating list of blocks, each with data from the present cycle of data-taking and the 9 prior readings for comparison.

OTHER ASPECTS

In addition to handling the analog data, the system also includes provisions for status monitoring for safety and control logic chains. This provides a "software" capability for alteration of chains, since all chain points are collected and assembled in parallel words of 10-bits each and then presented to a digital multiplexer. This output is scanned occasionally instead of the converter output. The "threshold" words used to compare against the digital status words allow simple logical decisions to be made, depending on the status of the collected chain data.

* For information on Model ADC-10F, use reply card. Circle C11

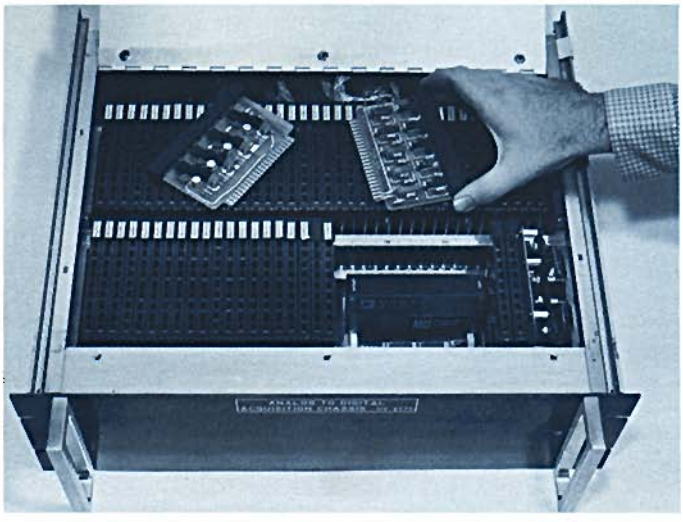


Figure 4. Data Acquisition Chassis. Converter is in Foreground

Low-cost (modest-speed) IC differential-input operational amplifiers are used for signal conditioning. MOS analog switches with internal chip address decoding simplify logic and switch-driver design. Because of the fast transient response required by the buffer amplifier between the multiplexer and the converter, an Analog Devices Model 45K* fast-settling op amp is used.

Communication with the memory is available through both keyboard and digital processor direct-memory-access channel. The system has a capability for set-point control and closed-loop regulation: One may include in the system, as a portion of its cycle, an *output* mode, such that certain address blocks supply digital words to control external devices. By manipulating the feedback element in each of those devices included, (in the control mode) the results of threshold comparison can be utilized to increase or decrease voltages or currents or gains, as a digitally-operated servomechanism.

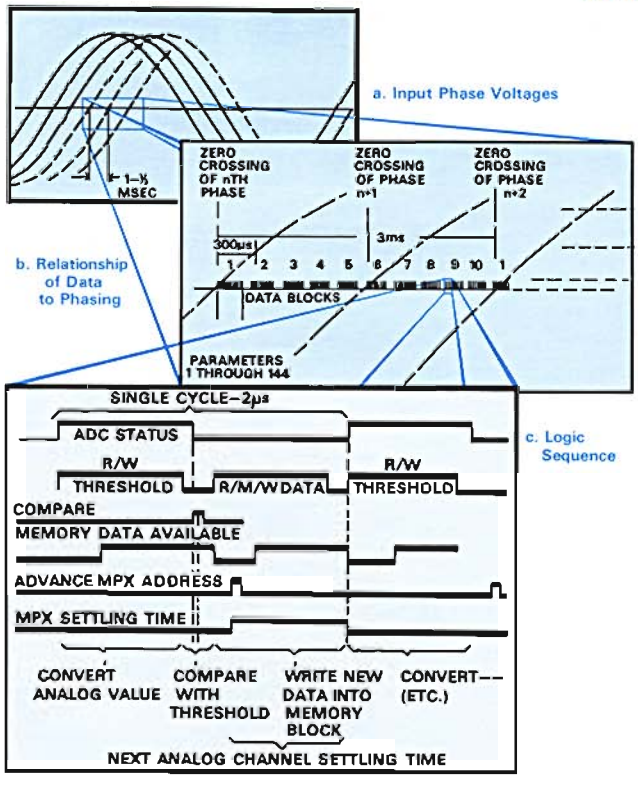


Figure 5. System Timing

*For information on Model 45, use reply card. Circle C12

Worth Reading

The following items should be of interest to readers whose especial interest is in solving problems of input instrumentation in the presence of common-mode signals. They form a logical supplement to the Model 272 bibliography, started in Vol. 5, No. 2, of *Dialogue*. Although not all new, they are relevant.

√*Signal Conditioning*, Applications bulletin No. 101, published by Brush Instruments Division of Gould, Inc., Cleveland, Ohio, a 16-page easy-to-read discussion of how-to-marry a given source configuration (e.g., single-ended–grounded, balanced–floating, etc.) to the appropriate preamp input configuration. Includes a useful "Amplifier Compatibility Summary" matrix.

√*Input Connection Practices for Differential Amplifiers, published by* Neff Instrument Corporation, Duarte, California, a practical discussion that ends with a summary of 6 "Input Connection Rules."

√*Techniques to Analyze and Optimize Noise Rejection Ratio of Low-Level Differential Data Systems*, published by Dana Laboratories, Inc., Irvine, California, includes numerical examples of CMRR calculations.

√"Protecting Hospitalized Patients from Electrical Hazards," from *Hewlett-Packard Journal*, March, 1970, discusses various approaches to isolation, including a floating-input scheme used in some h-p products. Also includes a table listing 9 steps to "Maintaining safe patient electrical environments."

√"Micropower Through the Heart is a Killer," from *EDN*, April 1, 1971 (cover story), a review of the problems of patient isolation and a solution (viz., the 272). **For reprint, Circle C13**

NOTED BRIEFLY

"Putting D-A Converters to Work: 10 Examples Show Versatility," by Rick Spofford, from *Electronics*, October 26, 1970, Uses of fixed-reference and multiplying DAC's in display, radar, servo, and other applications. **For reprint, Circle C14**

"Solid-State Voltage References," from *Instruments and Control Systems*, March, 1971, Compares standard cells and zener diode references, discusses tradeoffs for selection as primary and transfer standards.

Sample-and-Hold module, technical data SHA-1A. **Circle C15**

Isolation Amplifiers for Medical and Industrial Applications, technical data Models 272/273. **Circle C16**

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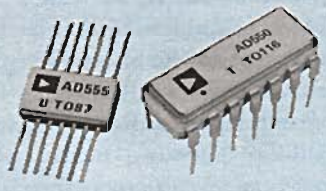

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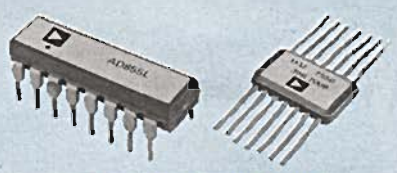

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