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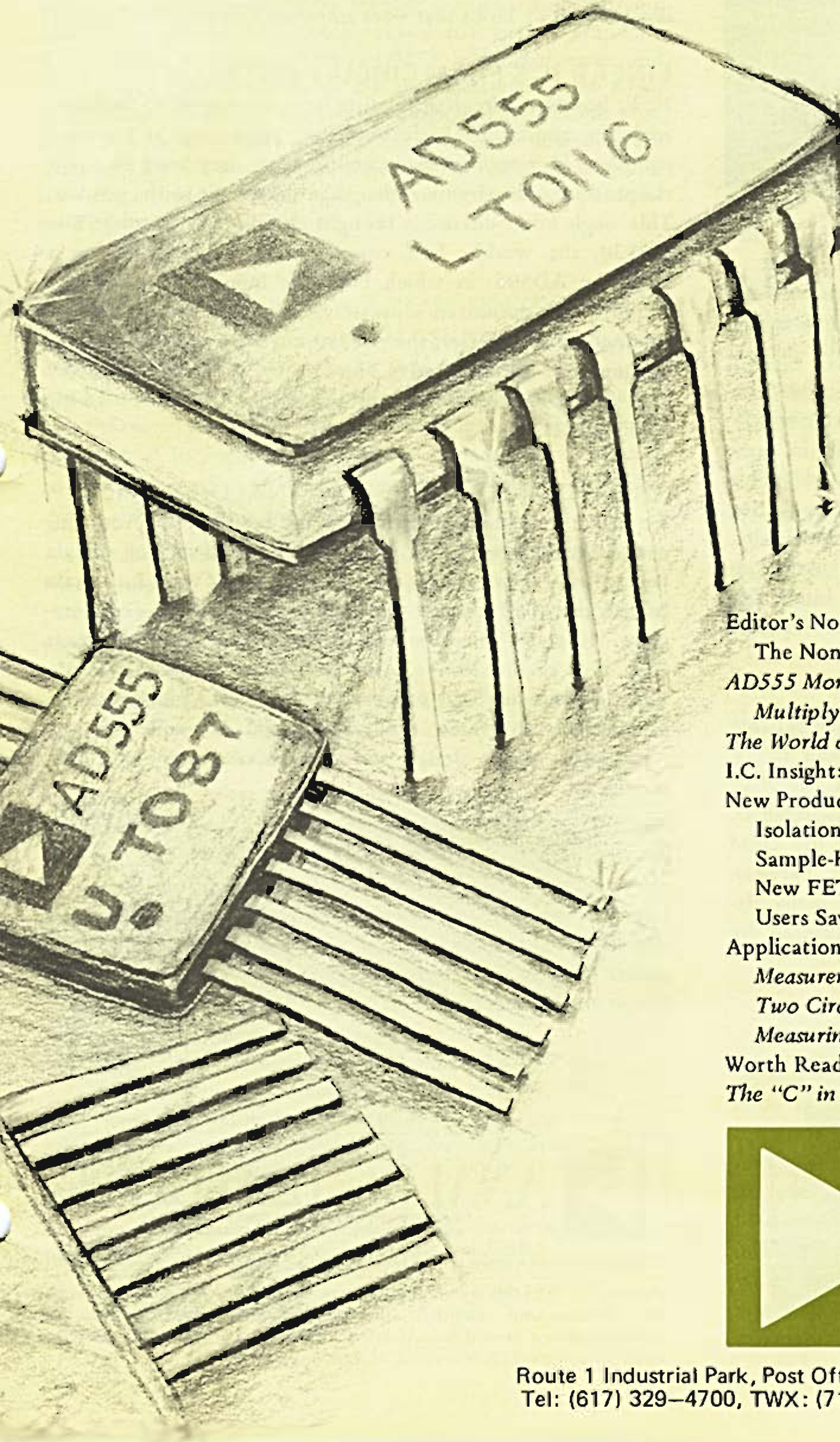

MONOLITHIC SWITCHES FOR 12–BIT MULTIPLYING D/A's – See page 3

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ANALOG DEVICES
Circuit Specialists

Route 1 Industrial Park, Post Office Box 280, Norwood, Massachusetts 02062, USA
Tel: (617) 329–4700, TWX: (710) 394-6577, Cables: ANALOG NORWOODMASS

Editor's Notes

(OUR) IC'S VS. DISCRETES: THE NON-CONFLICT

Ever since the first day that integrated circuits appeared feasible to manufacture, some observers have fostered the notion of a "product battleground," whereon the IC's (and their manufacturers) march to a successive series of victories over the ever-thinning ranks of obsolescent discrete-component circuits (and *their* manufacturers).

There is *indeed* a product battleground, but the issue that is joined is *not* "discretes vs. IC's." Rather it is an older issue, "Make or buy?" reflecting the philosophical differences between "component specialists" and *circuit* specialists. An excellent and revealing example of these differences, in the form of a direct and searching (albeit passionate) comparison between two recently-introduced monolithic multiplier circuits begins on page 8 of this issue of *Dialogue*.

A HISTORICAL DIGRESSION

The packaged-circuit revolution—boon to equipment designers. Whenever circuits occur repeatedly and identically in equipment, there is a strong temptation towards modular design.

The first major step toward modularity came during the '50's, as vacuum tubes in the miniature "noval" form became widely available. Small circuit pieces were offered for sale in octal plug-in packages, running the gamut from uncommitted Vector turrets ("make") to complete plug-in op amps ("buy"). For the first time, with the coming of the early circuit specialists, the OEM designer had the choice between designing-and-building his own circuits *in toto*, and shucking his design headaches for repetitive general-purpose modular circuit pieces having guaranteed specifications, low cost, compact size, and overall performance "good enough" for most jobs.

The trend accelerated during the early '60's, when reliable transistors and etched circuits became available, and epoxy packaging made the proverbial "black box" an everyday reality. These black boxes were designed by circuit specialists, as families of general-purpose devices, and again the OEM designer was offered the opportunity to choose between buying guaranteed-performance packages and the many degrees of freedom (but greater expense-to-use) offered by design with components.

IC's — BOON OR BANE?

Hard on the heels of this trend came the development of monolithic integrated circuits. The monolithics were a boon to everyone: They enabled "make"-oriented OEM's to put together more-complex systems and equipment at lower cost per function. They enabled OEM's oriented toward "buy" to replace—at lower cost—packaged circuits that were being used for simple applications. They enabled circuit specialists, who manufactured modular packaged circuits, to produce more-versatile circuit packages having better performance at lower cost. And, of course, they were a source of tremendous growth for their makers.

LINEAR IC'S FROM COMPONENT SPECIALISTS

The large semiconductor manufacturers, with their specialty of supplying components (starting with transistors) for "make," have entered the linear IC business cautiously, step by step, initiating only circuits that promised to be suitable for *extremely-high-volume production*, with minimal labor content. If this meant that the power supply had to be +12V, −6V, or that 3 passive elements were required for dynamic stabilization, or that 4 tracking resistors were required for the multiplier even to multiply, or that the op amp using all-monolithic FET's had 2000pA bias current, so be it! They have taken the risks of product development; the OEM world, like the weary traveler, should be grateful for its ultra-low-cost Procrustean bed, and for the small degree of interference with sleep caused by limbs that were amputated or stretched to fit!

LINEAR IC'S FROM CIRCUIT SPECIALISTS

Now, however, *circuit specialists* have developed IC facilities, with the continuing basic objective, producing, at low cost, *complete high-performance circuits now*; they tend to adapt the process to the circuit, rather than the circuit to the process. This single-mindedness has brought about (for example): The AD530, the world's first *complete* multiplier/divider on a chip; the AD506, in which both the input FET's and the amplifier are optimized separately, then "married" and laser-trimmed to zero offset; the AD550 quad conversion switches, designed for tracking to 12-bit accuracy, with the bonus that a reasonable fraction of production can meet 16-bit requirements.

IC'S AND DISCRETES FROM ANALOG DEVICES

We view the monolithic process as the best way of producing circuits to get small size, low cost, hermeticity, or certain performance advantages. It's also a good way to obtain critical *pieces of circuits* with which to make our own complete "discrete" circuits with better performance or lower cost, or in smaller size (e.g., as the "secret ingredients" of our 12-bit multiplying D/A converters.) It's all of a piece with our philosophy of "greater user value," and of employing any respectable circuit design and construction technique that provides it.

Our sales engineers, applications engineers, and field engineers (as well as our contributors to *Dialogue*) can now help you solve equipment design problems in terms of *both make and buy*, in terms of *both IC's and discretes*. And we're the *only* supplier of circuits, at present, that can offer you the whole gamut of possibilities, the hundreds of years of successful circuit experience, that "best buy" in circuits.

Dan Sheingold

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AD555 Monolithic "μDAC" Quad Switches Make 4-Quadrant Multiplying DAC's with 12-bit Linearity

by Heinrich Krabbe and Fred Molinari

Voltage switches, operated by DTL/TTL logic, are used with resistive ladder networks to make digital-to-analog converters. Analog Devices has recently introduced the new AD555 series of monolithic quad voltage switches (four to a chip), a unique design having superb tracking (±1mV, 10Ω) and capable of 12-bit (i.e., ±0.0125% = ±½LSB) accuracy. Factors contributing to their excellent accuracy and tracking include the use of dielectric isolation, monolithic matching, and some novel circuit-design approaches. The switches are available in hermetic 14-pin dual-in-line or flat packs, for both commercial and military equipment. Applications include 4-quadrant multiplying DAC's and high-accuracy analog signal distribution. Matching R-2R ladder networks (Series AD855)§ for converter and hybrid multiplier applications are also available. In the pages that follow, we discuss the circuit design and application principles pertinent to these new devices, and show how they are used in 12-bit 4-quadrant multiplication. Analog Devices has used these devices as "secret ingredients" of a new 4-quadrant multiplying D/A converter, the DAC-12M.†*

LOGIC-CONTROLLED VOLTAGE SWITCHES

Each of the four switches in the AD555 is of the single-pole, double-throw, break-before-make type (Figure 1). Two independent analog voltages are applied in common to the respective "throws" of all four switches. Each output, depending on the state of its input logic signal, is independently connected to one or the other of the input reference voltages. The logic inputs are compatible with both TTL and DTL.

Analog input voltages are typically in the range ±4V to common, supplied by operational amplifiers or other low-impedance sources, capable of sinking about 300μA per switch, plus the total load current, to a maximum of 1.5mA per quad. The switches will supply output currents up to 0.5mA, and will comfortably supply load resistances as small as 8kΩ (best accuracy is realized with load resistances of the order of 50kΩ or more).

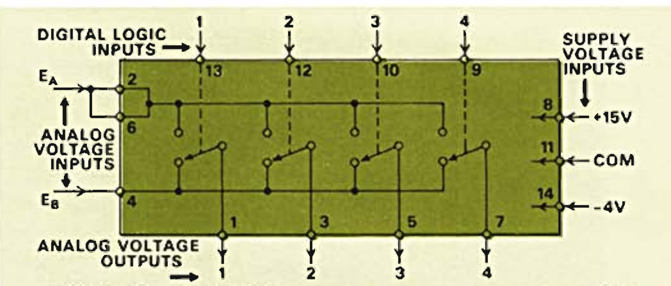


Figure 1. Functional circuit diagram of AD555 quad switch. Logic "0" connects an output to input B; logic "1" connects it to input A. Input voltages are supplied at low impedance (e.g., from operational amplifier outputs).

* For further data on the AD555 switches, use reply card. Circle B1
§ For further data on resistor networks, use reply card. Circle B2
† For further data on complete multiplying D/A's, use reply card. Circle B3
This article is based, in part, on a paper delivered at the 1971 IEEE International Solid-State Circuits Conference.

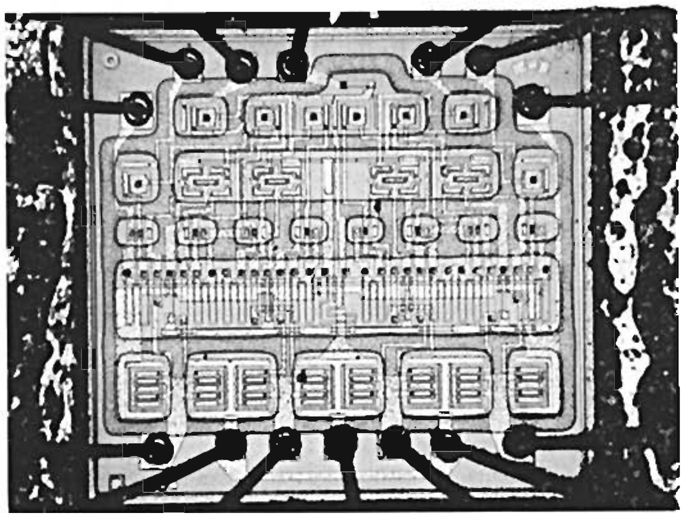


Figure 2. AD555 Chip Photomicrograph

Switch causality may be reversed: the analog input signals may be applied to the switch poles and the outputs taken from the throws. However, the common connection of the throws limits the usefulness of this connection.

Leakage current in the open circuit (< 1.5nA) is generally negligible. In the closed circuit, not only are offset voltage (millivolts) and series resistance (10's of ohms) small to start with, but the symmetry and identical construction of all four switches results in excellent tracking, which further reduces errors in circuit applications.

VOLTAGE SWITCHES IN D/A CONVERSION (Fig. 3)

The major application for which the AD555 switches were designed is digital-to-voltage conversion. In D/A conversion of an n-bit binary number, the first digit (i.e., the *most significant bit:* MSB) corresponds to a voltage equal to one-half of a reference voltage, and each successive digit corresponds to one-half the value assigned to its immediate predecessor. The output voltage can be expressed by the equation

$$E_0 = E_R\left(\frac{1}{2}u_1 + \frac{1}{4}u_2 + \frac{1}{8}u_3 + \ldots + 2^{-n}u_n\right)$$
$$= E_R\sum_{i=1}^{n} 2^{-i}u_i$$

where E_0 is the output voltage, E_R is the reference voltage, and u_i is either 1 or 0, depending on the logic state of the

THE AUTHORS

"Hank" Krabbe (R) is Director of Engineering at Analog. His many successful designs range from circuits to systems, from discretes to monolithic IC's.

Fred Molinari, Marketing Manager—Conversion Products, has both MSEE and MBA, and a proven record in conception and design of conversion IC's.

ith bit. If all bits are "1", the output will be equal to $E_R\ (1 - 2^{-n})$, or the reference voltage minus the *least significant bit* (LSB).

If E_R is fixed, the process is a simple D/A conversion. If E_R can vary, the analog voltage E_R is multiplied by the digital number. (The converter is essentially a digitally-controlled potentiometer.) Since the AD555 switches can accept reference (i.e., analog) signals of either positive or negative polarity, two-quadrant multiplication is an inherently simple matter. Four-quadrant multiplication adds but little complexity.

Mechanizing the conversion is also inherently a simple matter. The number of bits determines the number of switches, and thus the number of switch quads, i.e., $n/4$ (+1 if n is not evenly divisible by 4). The switches provide the u_i function, switching each bit-input between reference and common potentials.

Although one may conceive of many circuits to achieve the weighting and summation, perhaps the simplest to manufacture as a thin-film network, as well as to use, is the "R-2R" ladder network, composed—as its name indicates—of a repetitive pattern involving two sets of equal resistors related by a factor of 2. An example of a 12-bit attenuator in a DIL package is the Analog Devices AD855, designed specifically for use with the AD555 switches. It incorporates the special feature that the resistances in series with the four most-significant-bit switches are less than the nominal $2R$ by 15Ω, thus limiting switch-resistance errors to those caused by mistracking, and allowing lower values of network resistance. It also includes a set of matched and tracking scaling resistors for setting gain of inverting operational amplifiers, or for choice of output attenuation. Figure 3 shows how three switch quads and a 12-bit R-2R ladder network are used together for 12-bit D/A conversion. Unloading and overall scale factor adjustment are provided by an operational amplifier.

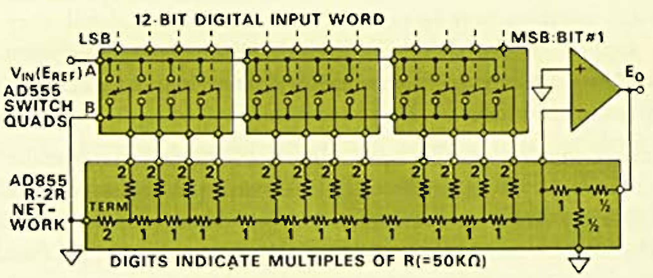
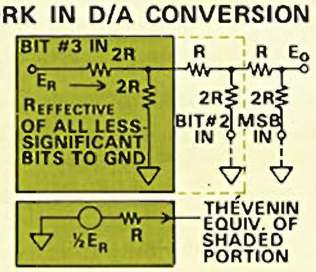


Figure 3. Quad switches and R-2R ladder network used in simple 12-bit D/A converter or 2-quadrant hybrid multiplier. Amplifier may be used either inverting, with tracking gain resistors, as shown, or non-inverting, with external feedback network.

HOW R-2R NETWORKS WORK IN D/A CONVERSION

Since the network is linear, superposition may be used to compute the total effect of all bits. Let us, as an example, compute the contribution of the third bit (2^{-3}). Apply the reference voltage to Bit #3, and ground all the other bits. The situation will be as depicted in the figure.



1. The effective resistance of the less-significant bits to ground will always equal $2R$. This may be seen in Figure 3 above; 2 $2R$'s in parallel form R; 2 R's in series form $2R$; 2 $2R$'s in parallel form R; etc.
2. The Thévenin equivalent of the shaded portion consists of a voltage, $E_R/2$, and a series resistance, R.
3. Applying this Thévenin generator to the next "L"-section, we can develop a new Thévenin equivalent consisting of a voltage, $E_R/4$, and a series resistance, R.
4. Applying this Thévenin generator to the final section, we find that $E_O = ½E_R/4 = E_R/8 = E_R \times 2^{-3}$.
5. If a current output is desired, the output node may be connected directly to the summing point of an operational amplifier. The output current will be equal to open-circuit output voltage divided by R. The feedback resistance should track R.

DESIGN OF THE AD555 SWITCHES

Problems. The designer of high-accuracy monolithic switches must solve two basic problems satisfactorily:

1. The error contribution of the output transistors must be minimized, in terms of both of its major components: intrinsic voltage offset, E_{os}, and saturation resistance, R_S.
2. Parasitic effects due to substrate conduction when the output transistors are in saturation, using conventional junction isolation, must be somehow overcome.

These problems have limited attempts to accomplish a monolithic high-accuracy D/A converter circuit design, using voltage switching. The problems have recently been satisfactorily solved for 12–16-bit *current*-switching conversion, as embodied in the AD550*; now the AD555 represents a successful solution in terms of 12-bit *voltage*-switching conversion and hybrid multiplication.

Process. From the outset, a bipolar approach was chosen in preference to a FET design, because monolithic processing of FET circuits for 0.01% accuracy in production quantity at low price is yet to become a practical matter. The bipolar approach utilizes output transistors capable of low offset voltage and minimal saturation resistance. To accomplish this, dielectric isolation offered a number of advantages:

1. It blocked parasitic substrate conduction effects.

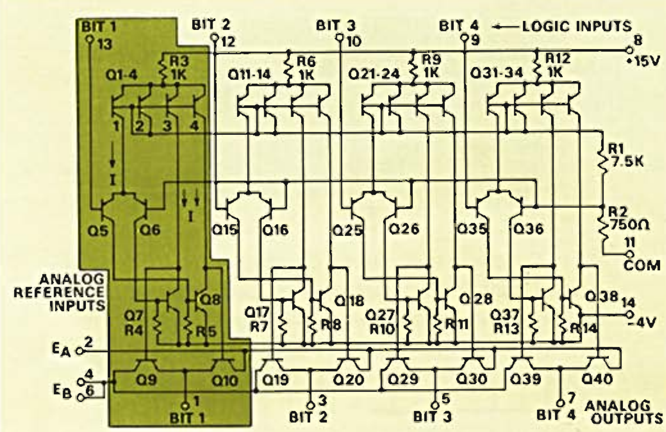


Figure 4. AD555 Quad Current Switch Schematic Diagram

*"A High Performance Monolithic D/A Converter Circuit," Pastoriza, Krabbe, and Molinari, ISSCC, February, 1970: for technical data on the AD550 current switches, use the reply card. Circle B4

2. It permitted the use of a wraparound low-resistivity buried layer that could provide the low saturation resistance needed in the output transistors.

3. It allowed the geometry of the symmetrical NPN output transistors to be tailored for low voltage offset.

Although dielectric isolation is somewhat more costly than junction isolation, it allowed our design objectives to be realized in a producible unit at high yield and low cost *now*.

Circuit. The complete circuit of the AD555 switches is shown in Figure 4. We shall describe its operation in terms of the Bit 1 switch, shaded in color. The other three switches are identical to it.

Q1-Q4 and resistors R1-R3 form a triple current source, which develops 0.3mA in the collectors of Q1, Q3, and Q4. The output-switching transistors are Q9, which connects the output to input B, and Q10, which connects the output to input A. The appropriate switching transistor is activated when sufficient current to saturate it is dumped into its base circuit. The current developed by Q3 flows in the base circuit of Q9 when Q7 is cut off, and the current developed by Q4 flows in the base circuit of Q10 when Q8 is cut off.

Q5 and Q6 form a comparator that switches the current developed by Q1 to the base of either Q7 (saturating it and causing it to deactivate Q9 by "stealing" its base current) or Q8. The base of Q6 is biased at a constant value between "0" and "1". When the logic input is at "0" (below the bias potential), Q5 conducts, turning off Q6 and Q7, thus turning on Q9 and connecting the output to input B. When the logic input is at "1", Q5 is turned off, turning off Q8, turning on Q10, thus connecting the output to input A.

It should be noted that all action in the switch occurs in complementary fashion (i.e., "push-pull"). Because an output-switching transistor cannot conduct until its associated "current-stealing" transistor is out of saturation, the sequence of operation when switched is: the closed circuit opens, then the open circuit closes. That is, the switch is "break-before-make."

The excess base current flows through the collector circuit. For this reason, the analog reference sources should be capable of sinking the base current of all four switches, plus any current drawn by the output load. Since the saturating current is independent of voltage level, the reference input (and hence, the output) voltage may swing positive or negative and is

FOUR-QUADRANT MULTIPLICATION

BINARY NUMBER	LADDER TERMINATION: GROUND	LADDER TERMINATION: $-V_{in}$
111 . . . 11	V_{in} (1 − LSB)	V_{in} (1 − 2LSB)
– –	—	—
100 . . . 01	$+V_{in}$ (3LSB)	$+V_{in}$ (2LSB)
100 . . . 00	$+V_{in}$ (LSB)*	0
011 . . . 11	$-V_{in}$ (LSB)	$-V_{in}$ (2LSB)
– –	—	—
000 . . . 01	$-V_{in}$ (1 − 3LSB)	$-V_{in}$ (1 − 2LSB)
000 . . . 00	$-V_{in}$ (1 − LSB)	$-V_{in}$

Four - Quadrant multiplication can be achieved with a slight modification of the circuit of Figure 3. Just apply an inverted version of input E_A to input B, which had been grounded. Apply offset binary instead of straight binary logic. That's all! The table indicates the range of positive and negative gains available for V_{in} over the range of digital input. Note that zero gain is not available, being bracketed by ±1LSB. If you wish a specific number to correspond to zero gain, connect the termination of the ladder to $-V_{in}$ instead of ground. The result will be as shown in the third column of the table.

*V_{in} (LSB) = $V_{in} \times 2^{-n}$

BRIEF SPECIFICATIONS

(Typical at +25°C and nominal supply voltages unless otherwise noted)

SWITCH OFFSET VOLTAGE, E_{OS}, max ($R_L \to \infty$)	AD555J/S	+10mV
	AD555K/T	+3mV
	AD555L/U	+2mV
OFFSET MISMATCH, ΔE_{OS}, max	AD555L/U	±1mV
SWITCH "ON" RESISTANCE, R_S, max	AD555J/S	100Ω
	AD555K/T	40Ω
	AD555L/U	25Ω
"ON" RESISTANCE MISMATCH, max	AD555L/U	10Ω
ERROR TEMP COEFFICIENT (between switches), max		±5ppm/°C
SWITCH LEAKAGE CURRENT, max (switch "off", V_{REF} = ±4V)		1.5nA
SETTLING TIME (to ±0.01% of final value)		5μs
SWITCH LOAD CURRENT, max		0.5mA
INPUT SIGNALS		
Digital: "0"	< 0.8V @ −500μA	
"1"	> 2.0V @ +100μA	
Analog: REF A and/or REF B in any combination	−3V (min) to +3V (max)	
REF A alone (REF B grounded)	−4V (min) to +4V (max)	
ANALOG INPUT CURRENT (REF A or REF B), max		−1.5mA
OUTPUT VOLTAGE RANGE, max		±4V
POWER REQUIREMENTS	±15V (±20%) @ +7mA	
	−4V (±1%) @ −3mA	
Dissipation	130mW typ, 200mW, max	
OPERATING TEMP RANGE	AD555J/K/L	0°C to +70°C
	AD555S/T/U	−55°C to +125°C
STORAGE TEMP RANGE	All Models	−55°C to +125°C

limited only by the negative bias voltage and the base-emitter breakdown rating of the "off" transistor. A happy feature of the AD555 design is that the switching transistors are both NPN and can therefore be matched to achieve virtually symmetrical performance.

PERFORMANCE

DC errors. The equivalent circuit of an AD555 switch, when conducting, is essentially a dc offset voltage ranging from 0 to +10mV, and a series resistance ranging up to 100 ohms. AD555 units are sold in three grades, depending on desired accuracy. For 12-bit conversion, the three quads would include one from each grade. The most accurate is indicated by the suffix L (commercial) or U (military), and has less than 2mV offset, ±1mV offset mismatch between switches, 25Ω series resistance, and 10Ω mismatch between switches. The nominal offset can be adjusted to zero in the output operational amplifier, and the initial resistance error can be reduced by trimming either the network resistors, the reference, or the gain. (In the AD855 12-bit R-2R ladder, the 100kΩ series-2*R* resistors are made 15 ohms low for the first 4 bits.) The remaining errors contribute less than ±½LSB of 12 bits (±0.0125%). Because the spread from "best" to "worst" is only 10:1, the errors are contributed almost wholly by the "most significant quad." Leakage through the open switch is less than 1.5nA (1.5nA × 100kΩ = 0.15mV). As one would expect, temperature tracking is excellent, and total error with temperature changes by about 5ppm/°C.

Response speed. The limiting source of delay is the time taken by the sinking transistor to come out of saturation and allow the switching transistor to saturate. This delay, in AD555, is typically 5μs, comparable with other available voltage switches. Since other sources of delay are small, the settling time to ±½LSB is of the same order.

Op Amps as Electrometers or — The World of fA

The definition of an electrometer is not graced by complete agreement. However, it would seem to describe a system capable of measuring voltages or currents with leakages or resolutions somewhere in the range below 1 picoampere, and probably closer to 1 femotampere ($1fA = 10^{-15}$ A). The elements of the system include a transducer, a connecting cable (long or short), an input device, and some form of signal conditioning.

Today, op amp designers have introduced a number of devices capable of functioning as input devices and signal conditioners. A good example is the Analog Devices 310/311* family of parametric amplifiers, with bias currents less than 10fA. The essential advantage of such devices to the electrometer designer is that—if well made and capable of predictable performance—they remove one of the headaches from an exercise that exceeds Excedrin dimensions.

We offer here a few notes for the electrometer designer encompassing the factors to be considered in the choice of input device, and the considerations involved in interfacing it to the rest of the problem. A *non*electrometer-designer who is interested in low current measurements may also derive some ideas for extending his frontiers and an appreciation of the scope of the problems involved when performing measurements in "the world of fA."

THE PROBLEM

Nominally, the problem is either to transduce a current into a voltage, or to measure a voltage with negligible current drain. The basic configurations, familiar to all op amp users, are shown in Figures 1a and 1b.

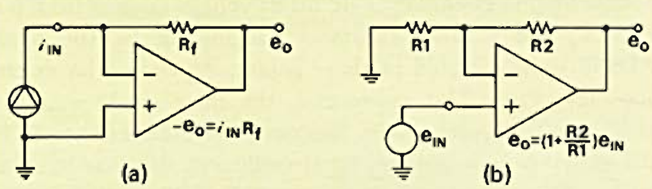


Figure 1. Basic low-level measurement circuits. (a) The operational amplifier develops e_O proportional to i_{IN}, offering a low dynamic impedance to the current source. (b) The operational amplifier unloads the voltage source, and provides voltage gain.

The basic configuration of 1a is typically used for gas chromatographs, ionization gages, and photomultipliers; the configuration of Figure 1b is typical for pH meters. In the limited space available, we will discuss largely the configuration of Figure 1a. Most of the considerations, except for the feedback resistor, will be applicable to Figure 1b, together with the added leakage and common-mode problems introduced by above-ground operation.

Figure 2 shows how the basic configuration of Figure 1a looks in the harsh light of reality. The two major sources of trouble are:

1. The circuit inside the triangle
2. The circuit outside the triangle.

INSIDE THE TRIANGLE

On the opposite page are tabulated a number of approaches to obtaining low-leakage high-impedance inputs for electrometers, with a comparison of their input current, voltage drift, and noise.

For new designs, one may dismiss the vibrating-capacitor electrometer as being too expensive and too good for most jobs, although it approaches the limits of resolution the most closely of all. One may also dismiss electrometer tubes, if any of the solid-state approaches will do the job, on the basis of the mechanical ruggedness and lower drift of the solid-state alternatives.

Among solid state designs, MOSFET's, though not yet proven, are just coming into use, where—if their dc drift and need for protection can be tolerated—they are most useful in applications calling for low leakage current over a wide range of temperature. (However, the increase in leakage of the protection diodes with temperature may somewhat dimish this apparent advantage.) Junction FETs have the virtue of simplicity, and they can be selected for low voltage drift, but even the best have leakage currents (10^{-13}A), that are difficult to offset.

Parametric op amps have stable, readily-predictable characteristics, and they have been reliably used in electrometer circuitry for a number of years. Their major weakness is the bandwidth limitation imposed by their input capacitance, and the excessive current noise developed by the voltage noise across the input capacitance, at frequencies above 1Hz.

OUTSIDE THE TRIANGLE

Feedback Resistor In the world of fA, some resistors look like capacitors; others look like filters: connected in a feedback circuit, their "Boella" effect can cause an amplifier's response (and noise) to peak. Correction of such noise peaking can be accomplished by connecting a short length of stiff wire to the output end of the feedback resistor and bending it to form an angle to the resistor body. The length of the wire and the angle can be experimentally adjusted

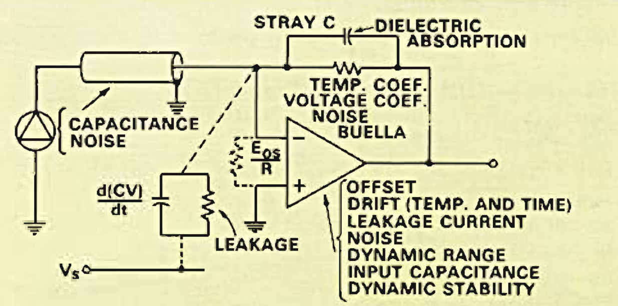


Figure 2. Some sources of error.

BIBLIOGRAPHICAL NOTE

"The Ubiquitous Electrometer," which appeared in *Industrial Research*, March, 1971 (pp 36-39), offers an articulate discussion of the present status of electrometry and a wealth of applications ideas for electrometers and "para-electrometers." Soft on MOSFET's, but well worth reading.

to eliminate peaking effects without appreciable bandwidth limiting (as compared to a single lumped feedback capacitor).

Dielectric absorption (in resistors), 1/f noise (in poor-quality resistors carrying current), resistance temperature- and voltage-coefficients, all can cause errors. Typical high-value resistors with acceptable properties that have been in use in recent years have been manufactured by such companies as Victoreen and Welwyn. A recent development is chip resistors; M.S.I. Company of North Attleboro, Mass., can supply values up to $2 \times 10^9\,\Omega$ and beyond.

Resistive attenuators may be used in the feedback circuit to magnify the effective resistance (a 10:1 attenuation will make $10^{10}\,\Omega$ look like $10^{11}\,\Omega$) and decrease the shunt capacitance somewhat. However, they magnify voltage drift and low-frequency noise.

Input cable capacitance is important, because the amplifier's noise voltage develops a noise current proportional to the shunt capacitance. The cable should be rigid, because capacitance changes will cause changes in its charge (or current: $i = dQ/dt = d(CV)/dt = C\,dV/dt + V dC/dt$). Also, bending of the insulation can develop noise, in the form of friction-induced charges. Graphite-coated insulation helps. The input leads should be fixed in relation to the power supply, again because the motion of vibration can cause substantial $V\,dC/dt$ currents (capacitor microphone effect). e.g., for $dC/dt = 10^{-15}$ F/sec, i = 15fA. And don't forget the $C\,dV/dt$ term for ripple, or the dc leakage resistance.

Guard the input, but watch out for humidity! Especially transient humidity, i.e. "breath storms." Consider galvanic action, developed in the presence of moisture, between the gold plating on a TO-5 case and the Kovar leads.

Dielectric absorption can be important. The circuit time constants are all so long, even with low-pF capacitors, that dielectric absorption, having comparable time constants, can hoodwink the unwary observer.

Teflon insulators should have minimal stresses, because they can generate currents up to 1pA, which gradually subside over periods of *hours*.

Thermal baffling. With $10^{10}\,\Omega$, 0.1pA will develop 1mV of signal. For best accuracy and lowest drift, the circuit should be kept in a thermally-baffled enclosure. It will respond to changes in temperature slowly, and be essentially immune to fast transient changes. Ovens having proportional temperature control may be used, but their elevated temperatures cause increased current offsets, and their regulation may not be close enough to make their use of more than incremental value.

This article is based, in part, on a talk given by Analog's Bob Demrow at CECON '70, entitled "Low-Current Measurements"

KEY PARAMETERS OF INPUT ELEMENTS

INPUT CURRENT

METHOD	IDEAL LIMIT	REAL LIMIT	TYP VALUE Amperes	COMMENTS
Varactor Bridge (Model 310)	$\sim \Delta I_o \left(\frac{V_P}{k}\right)^2$	$\frac{\Delta E_{os}}{k}\,I_o$	2×10^{-15}	Design compromises possible by variation of pump voltage and pump frequency
Metal Oxide Semiconductor (MOS)	0	Insulation Resistance and Protective Device	10^{-14}	Input protection needed
Electrometer Tube	0	Insulation Resistance ION Currents and Photon Currents	10^{-14}	Drifts with time
Junction FET	I_D	Insulation Resistance	10^{-13}	Relatively high input current
Vibrating Capacitor	0	~0	10^{-16}	Expensive

VOLTAGE DRIFT

METHOD	IDEAL LIMIT	REAL LIMIT	TYP VALUES	COMMENTS
Varactor Bridge (Model 310)	Work Function	Cost of Selection	20μV/Day 10/30μV/°C	———
Metal Oxide Semiconductor (MOS)	Work Function and Channel Resistance	Qss (Surface State Charge)	5mV/Day 150μV/°C	Qss is time, temp, bias, previous history, and $ dependent
Electrometer Tube	Work Function	Time Change of Work Function	2mV/HR	Rate is time dependent
Junction FET	Work Function and Channel Resistance	Cost of Selection	50μV/Day 10/50μV/°C	———
Vibrating Capacitor	Work Function	Materials Technology	100μV/Day	———

NOISE

METHOD	i_N (rms) @ 1Hz	C_{IN}	e_N (rms) @ 1Hz
Varactor Bridge (Model 310)	1/4 fA/$\sqrt{Hz}$ (Shot Noise)	30pF	flat – :1μV/$\sqrt{Hz}$ $e_n \propto V_p$
Metal Oxide Semiconductor (MOS)	1/4 fA/$\sqrt{Hz}$ Protective Device and Leakage Resistance	10pF (With Protection)	1/f (3μV/$\sqrt{Hz}$)
Electrometer Tube	1/6 fA/$\sqrt{Hz}$ Leakage Resistance and Shot Noise	5pF	1/f (½μV/$\sqrt{Hz}$)
Junction FET	1 fA/$\sqrt{Hz}$ Shot Noise and Leakage Resistance	6pF	1/f (½μV/$\sqrt{Hz}$)
Vibrating Capacitor	Johnson Noise	10pF	flat – (½μV/$\sqrt{Hz}$)

*For technical data on 310/311 and low-leakage FET's, use reply card. **Circle B5**

A Choice, Not an Echo: The AD530 vs. . . .

COMPLETE-ON-A-CHIP IC MULTIPLIER COSTS MORE, IS NEVERTHELESS A BEST BUY

Analog Devices has recently announced and is delivering from stock the AD530, a 1%-2% monolithic "MDSSR"—Multiplier-Divider-Squarer-Square Rooter—the first and only (as of this writing) such device to be complete on a single chip. It was reported on at length in Analog Dialogue, Vol. 5, No. 1.*

The AD530 is not the only monolithic device having balanced multiplying properties available to users at present. There are on the market two "multiplier cell" designs, hampered by not being complete multipliers, and requiring substantial external paraphernalia to function as MDSSR "black boxes." Since these devices are at the moment individually priced somewhat below the AD530, they will inevitably be considered for use in circuits and systems where AD530 could be more profitably employed, if the designer were acquainted with all the relevant issues.

It is our purpose here to spell out the engineering factors that should be considered, to make known to our readers the results of engineering evaluations on a comparative basis (specs alone won't do it, because we're comparing a complete device with a partial circuit whose system performance cannot be guaranteed by its manufacturer), and to invite you to explore the alternatives – and make an informed choice of the AD530.

THE CONTENDERS IN BRIEF

MC1495/1595 This was the first commercially-available IC to offer a balanced multiplying capability. Its price is very low; however, even if it cost nothing, it would still be expensive to use as an MDSSR. It is a bare multiplying cell, requiring a 32-volt and a −5-volt power supply for ±10V signals, an external component array composed of (for multiplying) 18 resistors/pots (7 of which are 1% tolerance or better), 2 capacitors, and an operational amplifier, plus a degree of design virtuosity (to figure out what kind of overall performance to guarantee). For dividing, it uses 6 additional resistors, 4 additional capacitors, and an additional operational amplifier. The manufacturer's recommended connection of MC1495/1595 is shown in Figure 1. For most new designs its manufacturer has superseded it by MC1494/1594.

MC1494/1594 A recently-announced improved version of the '95, this new multiplying cell, operating from ±15V, includes

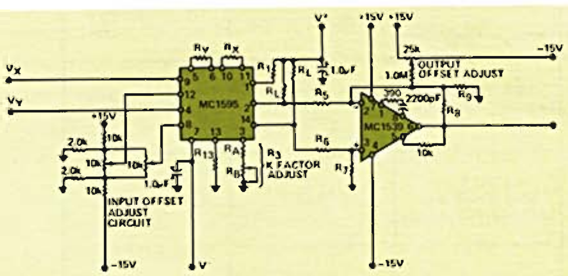
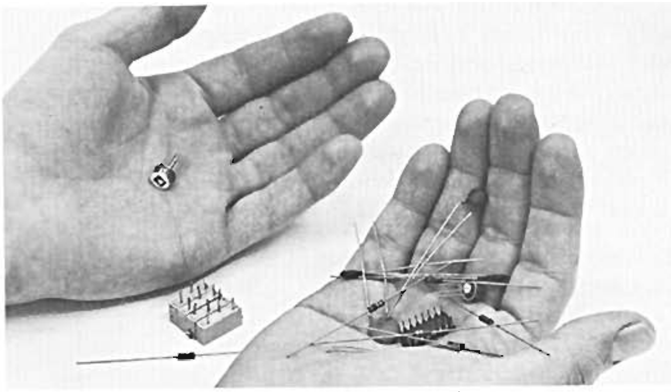


Figure 1. Suggested connection of MC1494/1595L as Multiplier in Motorola Semiconductor Application Note



a current-and-voltage regulator and a differential/current converter. However, to make it multiply, as a low-output impedance MDSSR, you still need 10 resistors/pots, 5 capacitors, and an op amp, and the circuit requires 15 connections to the device itself. The manufacturer's recommended connection of MC1494/1594 is shown in Figure 2.

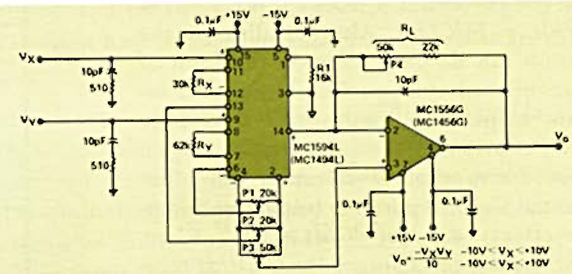


Figure 2. Suggested connection of MC1494/1594L as Multiplier in Motorola Semiconductor Application Note. Simpler than connections in Figure 1.

AD530J/K The AD530 contains on its chip an output op amp and all necessary resistors and other circuit elements to act as a complete multiplier, with *no external paraphernalia*, to a degree of accuracy satisfactory for many purposes (~20dB error referred to full scale). By the use of four externally-connected adjustments, it is set to rated 1% (AD530K) or 2% (AD530J) accuracy. It can be connected for division, squaring or square rooting without any external amplifiers. Only 10 device connections are required, allowing it to be packaged in a hermetically-sealed TO-100 can. The complete circuit of the AD530, connected as a multiplier that meets its specifications, is shown in Figure 3.

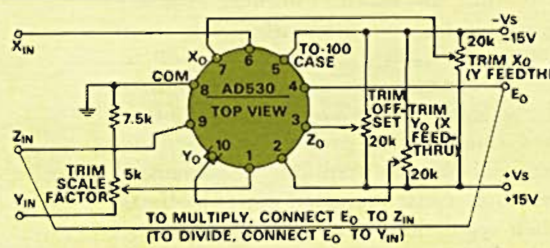


Figure 3. Suggested connection of AD530 as Multiplier, showing maximum number of trims. For Division, simply connect E_O to Y_{IN} and feed numerator signal to Z (X–Negative)

*For further data on AD530, use reply card. Circle 86
See also the advertisment on page 16.

PERFORMANCE EVALUATION

Since the manufacturer does not to our knowledge publish an overall accuracy specification, we put ourselves in the shoes of the user, and performed a circuit performance evaluation, directly comparing a group of ungraded AD530's with some (premium) MC1594 units on a parameter-by-parameter basis, as multipliers. All tests were made with the devices adjusted for the best performance in the manufacturer's recommended circuits. The tests were evaluated by eyes jaded by many years' critical evaluation of modular multipliers of all kinds. Specific comments appear in the table below.

MAKE OR BUY?*

On a side-by-side, package to package, performance-doesn't count, no-other-components-added basis, the price of the MC1494/1594 is less than that of the AD530. However, besides the fact that the '94 won't work without added components, such a price comparison is illogical. The knowledgeable user never buys on apparent price alone. He first looks for the minimum performance required to meet his system objectives, then the lowest cost. To determine his *true cost,* he adds the device price, the additional component prices, the cost of ordering, stocking, testing, handling (etc.), the added components, and the engineering cost required to put the whole thing together and make it work well enough to meet his performance goals, *in production.*

When compared on this basis, the lower performance and absence of guaranteed function accuracy of the 1494/1594 circuits, combined with the considerable extra effort required to make them work, may well outweigh any seeming component-price advantage.

Stan Harris

CHARACTERISTICS OF AD530 & MULTIPLIER-CONNECTED MC1594 — A CRITICAL EVALUATION

CHARACTERISTIC	MC1594	AD530
Accuracy:		
Total Error	No spec, range was 0.9% to 1.3%	80% met K (1%) spec, many below 0.5%
Total Error vs. Supply	10 to 15mV/%ΔV_s	10 to 15mV/%ΔV_s
Total Error vs. Temp. (0°C to +70°C)	0.05%/°C	0.05%/°C
Accuracy Components:		
Linearity	Spec: 0.5%, measured: 0.25%	X-input 0.35%, Y-input 0.15% A definite advantage where linearity must be as low as possible on one input
Offset and Drift	1.5mV/°C	1mV/°C
Feedthrough (V_{out} for one input grounded, 20V p-p at other input)	50mV p-p, both X and Y	35mV p-p (Y input), 20mV p-p (X input)
Feedthrough vs. Frequency	Essentially constant to 10kHz; increasing to 1V p-p @ 100kHz	X essentially constant to 100kHz, Y to 200kHz X: 60mV p-p @ 100kHz. Both increase @ 6dB/octave at higher frequencies
Dynamic Response:		
f_t (small-signal −3dB frequency)	800kHz with MC1741, 1MHz with MC1556	1MHz
f_p (full-power-output break frequency)	10kHz (MC1741), 40kHz (MC1556)	700kHz
Slew Rate	0.6V/μs (MC1741), 2.5V/μs (MC1556)	50V/μs
Character of Response	Exhibits peaking (MC1556)	Smooth: no peaking, no overshoot on step
±1% Amplitude Error	40kHz (MC1741), 50kHz (MC1556)	90kHz
1% Vector Error	6kHz (MC1741), 10kHz (MC1556)	10kHz
Settling Time (to 1%, 0 to 10V step)	20μs (MC1741), 6μs (MC1556)	0.8μs
Output (±10V)	Supplies only ±200μA full-scale output without op amp	Supplies full ±5mA @ ±10V
Input Characteristics	High input resistance (> 1MΩ) Low bias currents (1 to 2μA) Tends to have parasitic oscillation in absence of input damping networks	High input resistance (> 1MΩ) Low bias currents (1 to 2μA, X and Y)
Noise	No spec; 2 to 3mV rms 10Hz to 1MHz	3mV rms, 5Hz to 5MHz, occasional unit has zener noise
Range of Adjustments (referred to f.s. at multiplier input)	±40% (i.e., ±4V)	±20% (i.e., ±15V internally attenuated to ±2V)
External Parts for Best Performance	6 resistors, 5 capacitors, 4 pots	1 resistor, 4 pots
Number of Pin Connections	15 connections (16-pin DIP)	10 connections (TO-100 package)

*See also "FET Op Amps, Make or Buy" P. 11 Analog Dialogue, Vol.4, No.2

new products

Medical Isolation Amplifier

NEW AMPLIFIER HAS 1000V ISOLATION FROM GROUND, 120dB CMR. FLOATING DESIGN PROTECTS HOSPITAL PATIENTS FROM MICROSHOCK. NEW APPLICATIONS PROMISED IN INSTRUMENTATION AND PROCESS CONTROL. ONLY $109 (1–9).

The first of a new series of products for medical instrumentation, Analog's new Model 272* is a compact FET-input unity-gain amplifier that has "total" isolation between input and output signal ground circuits, using modulation techniques and transformer isolation. It will decouple transducers, hospital patients, and other sources of moderate signal levels from signal conditioning and control system wiring, avoiding both overt and "sneak" paths to ground.

It will withstand 5,000 volts across its shielded isolation circuit and has a maximum of 10pF of stray capacitance between *any two terminals*. It will process millivolts of signal having components from dc to 2kHz with 0.2% nonlinearity, even in the presence of 1,000VCM. The input CMR is 120dB (60Hz) with 5kΩ source unbalance, and the input current is limited to less than 10μA for ground faults to a 220VAC line.

Until now, this kind of performance and protection has been available only in rack-and-panel designs priced on the order of $1,000. Now you can get it *for 1/10* of that price, in a convenient plug-in module. Ideal for portable equipment, it will operate from a single dc supply at any voltage from +9V to +28V, deriving all regulated operating power via an internal 150kHz dc-dc converter. Figure 1 shows a block diagram of the 272 isolation amplifier.

SAFETY IN MEDICAL APPLICATIONS

Model 272's isolation was designed to exceed the existing patient-safety specifications of Underwriters' Laboratories, the Veterans' Administration, and other regulatory agencies. Being specifically optimized for use in patient-monitoring equipment (ECG in particular), this amplifier will do its job without exposing the hospital patient to the hazards of microshock and possible electrocution. Such hazards can arise when voltage differentials as little as *10mV* exist between the patient and equipment grounds. If we assume the subcutaneous body resistance to be about 500Ω, exposure of a catheter to 10mV of ground-fault potential will produce a current of 20μA, a level of microshock that could conceivably cause ventricular fibrillation, and possible electrocution.

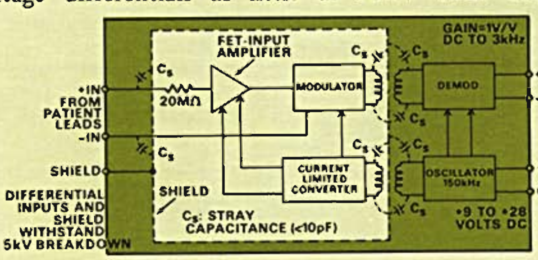


Figure 1. Block Diagram of Model 272 Isolated Unity-Gain Amplifier

*For further information on Model 272, use reply card. Circle B7 Or phone Peter Zicko, Marketing Manager, Analog Products, 617/329-4700

The bibliography lists a number of articles that discuss the serious consequences of hazards resulting from inadequate design, poor grounding, and careless use of electronic equipment in hospitals by untrained personnel. The estimate that 1,100 patients were electrocuted in 1969 underlines the desirability of using isolated circuitry, which will permit the patient to survive a broader range of accidents of malfunctions.

When 272 is used with patient-connected electrodes, the transformer isolation and low-stray intershield and terminal capacitance (10pF) and extremely-high circuit resistance will interrupt these lethal current paths. The amplifier and its input stage are conservatively designed, including fail-safe techniques to prevent the amplifier itself from becoming a source of lethal currents. The amplifier is designed to survive 5kV pulses (as from defibrillators) across its input terminals or across isolation barriers, thereby eliminating the need for (and cost of) such special external protective devices as spark gaps, neon bulbs, zener diodes, etc., as are needed to protect many types of ECG amplifiers.

The amplifier also makes patient-circuit current limiters unnecessary, since the input-current levels are less than 10μA, even with 220VAC applied, either between the inputs or from either input to system grounds. (Such protection devices, when used, usually reduce CMR and degrade instrument performance.) Figures 2 and 3 show examples of the kinds of test circuits that demonstrate protection of patients against ground faults and the 272 against defibrillator pulses.

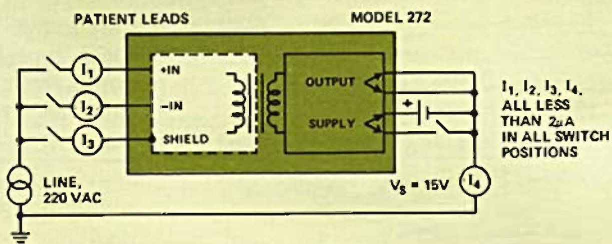


Figure 2. Patient Safety Test Circuit. All currents less than 2μA for any Set of Switch Closures

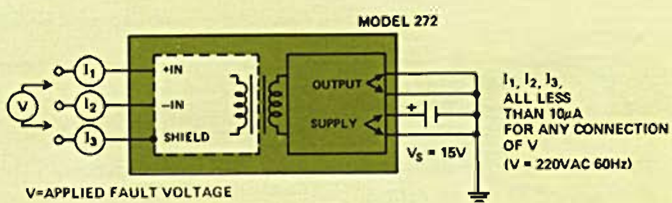


Figure 3. Patient and Equipment Safety Test Circuit. Amplifier must recover within 0.2 seconds for any connection of V = 5kV peak, 10-second rectified 60Hz burst. All currents less than 10μA for V = 220VAC, 60Hz; in any connection.

APPLICATIONS IN INSTRUMENTATION AND PROCESS CONTROL

In the chemical and metal-processing industries, there is often a requirement to measure and control millivolt-level signals in the presence of substantial common-mode voltage (e.g., 1,000V). Not only will Model 272 provide adequate isolation for the electrical signal; it will also protect both the equipment and the operator. In nuclear applications, there is often the need to buffer transducers from system fault modes. Finally, the amplifier is useful for handling floating signals in remote-programming and processing equipment.

NEW HORIZONS

Besides the obvious and urgent medical applications, and the examples of instrumentation applications mentioned above, the 272 may well open up a new range of applications in the monitoring of difficult-to-interface system variables at reasonable cost.

ISOLATION FEATURES & BRIEF SPECIFICATIONS

- Stray Capacitance < 10pF (Note: 22pF leaks 1μA @ 117VAC, 60Hz)
- 5,000-Volt Input-Output Isolation
- Patient-Input Fault Currents < 10μA @ 220VAC Line
- Input Circuit Protected for 5kV Defibrillation Pulse[1]
- Zero dc Bias Current < ±50pA $I_{in}^+ \equiv -I_{in}^-$

Gain	1V/V ±3%, 0.015%/°C
Linearity	0.2% (4V p-p), 0.02% (40mV p-p)
Settling Time	200ms (±100μV) after 1kV-10ms input pulse
CMR (MIN)	115dB @ 60Hz (5kΩ unbalance)
CMV	1,000V (5,000V, 10ms pulse)
Offset Voltage[1]	±50mV, drift 125μV/°C
Input Current	±50pA each input, 0 both inputs (i.e., $I_{in}^+ = -I_{in}^-$)
Input Impedance	$10^{11}\Omega$, linear region 20MΩ resistive, overload region[1] $10^9\Omega$ || 10pF, common mode
Output Voltage Range	±3V (dc to 250Hz), f_t = 2kHz
Output Noise[1]	25μV p-p, 0.05 to 100Hz
Output Impedance	1.5kΩ
Supply Requirement	+9 to +28VDC (14mA @ 15VDC)
Operating Temp	0°C to +70°C

[1] Model 273 Option Available: offers 1 meg instead of 20 meg input resistor. Same specs except Noise—10μV p-p, input protection ±400V, Drift 100μV/°C

BRIEF BIBLIOGRAPHICAL REFERENCES

These are representative examples and do not form an exhaustive or comprehensive listing. However, more-complete bibliographies are published in several of these items.

Electric Hazards in Hospitals, Proceedings of a Workshop, National Academy of Sciences, Washington, D. C., 1970.

"Hazards of Electrical Apparatus," Bruner, J. M. R. *Anesthesiology*, 28 (2), pp 396-425, 1967.

"Monitors that Save Lives Can Also Kill," Stanley, P. E. *Modern Hospital*, 108 (3), p. 121, 1967.

"New Concept: Safe Patient Power Center," Walter, C. W. *Modern Hospital*, 112 (6), pp 152-162, 1969.

Articles on the subject have also appeared recently in many electronics trade publications, such as *Electronics*, *Electronic Design*, *MED*, etc.

Proposed *Safety Standards for Electromedical Apparatus*, American National Standards Institute, Sub-Committee C-105 (Medical Electronic Equipment); Dr. Hans A. Von der Mosel, Cliffside Park, N. J.

new products

Modular Sample-Hold is a "Best Buy"

MAX 0.01% ACQUISITION TIME 5μs
MAX DROOP RATE IN HOLD 5ppm/ms
HIGH-Z INPUT, SEPARATE GROUNDS
FULL OUTPUT RANGE ±15V SUPPLIES
PRICE ONLY $150, LESS IN QUANTITY

Model SHA-IA is a general-purpose unity-gain sample-hold amplifier in a 2" X 2" X 0.4" modular package with dual in-line pin spacing. It is designed to be "used anywhere" in an analog or hybrid data-handling system for resolutions to 12 bits. It can be applied in data acquisition, ahead of an A/D converter, in data distribution, after a D/A converter, in a peak- and valley-following, and as an all-around storage unit for analog signals.

Features that make it unusually useful in these applications include fast signal acquisition (5μs max to 0.01% of a 20V step), and even faster aperture time (40ns max with 5ns peak jitter allows capture of 0.2V/μs-slewing signals with less than 1mV uncertainty), combined with a low droop rate (50μV/ms is 5ppm per ms of full scale).

Three separate grounds allow the analog signal output to be essentially independent of the digital system's common-mode level and ground noise. Tracking errors include maximum nonlinearity of 0.01%, offset drift of 25μV/°C, and gain error of +0, −0.05%. Settling time, sample-to-hold, is less than 300ns to ±1mV, and feedthrough in Hold is less than 50ppm of 20V p-p at 10kHz. The high input impedance ($10^{12}\Omega$, 10nA max) buffers the storage capacitor from the signal source, avoiding source stability problems and providing a controllable slew rate. The logic inputs ("1"—Sample or Track, "0"—Hold are DTL/TTL compatible.

BRIEF SPECIFICATIONS – MODEL SHA-IA

Sample Mode	
Gain	+1 (+0, −0.05% max)
Nonlinearity	1mV max, 0 to ±10V
Slew Rate	4V/μs
Settling Time to 0.01%	5μs max, 20V step
Sample-to-Hold	
Aperture Time	40ns, less than 5ns variation
Settling Time to ±1mV	300ns
Hold Mode	
Droop Rate	50μV/ms, max
Feedthrough Error	50ppm max, 20V p-p @ 10kHz
Input Circuit	
Impedance	10^{12} || 5pF
Bias Current	10nA max, 1nA typ
Offset Voltage & Drift	1mV max, 25μV/°C max
Output Voltage & Current	±10V min @ ±20mA min
Logic Inputs (DTL/TTL)	"1" – Sample, "0" – Hold
Temp Range	0°C to +70°C, rated accuracy
Price (1–9)	$150.

For further information on Model SHA-IA, use reply card. Circle B8

new products

New Modular Op Amps

BOTH HAVE FET INPUTS, FEATURE CMR 80dB OVER FULL ±10V. MODEL 44 SETTLES TO 0.01% IN < 1μs, MODEL 43J HAS F_p = 100kHz, LOW COST, LOW OFFSET AND DRIFT.

*MODEL 44K** is a fast-settling differential amplifier designed specifically for applications as a high-impedance buffer. In addition to its settling time of less than 1μs to 0.01%, it has guaranteed 80dB CMR over the full ±10-volt range, drift less than 15μV/°C, 25pA bias current. Its speed and accuracy are further enhanced by ±10V, ±20mA output capability, 75V/μs slew rate, 1MHz full-power response, 10MHz small-signal bandwidth, and dc gain of 100,000.

In addition to buffer applications, amplifiers having fast settling and commensurate drift and dc-accuracy specifications are useful in current-to-voltage converter networks, fast integrators requiring long hold time, etc. Model 44K is priced at $52 (1–9). Model 44J, a companion unit at lower cost ($42), is available for applications having narrower temperature range that can tolerate somewhat greater drift (50μV/°C). Both models are available from stock.

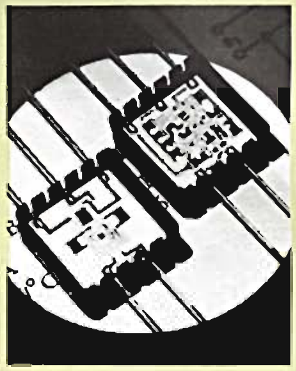
MODEL 43J is designed to provide high circuit accuracy at low cost. The starting point for the unit's superior cost/performance tradeoff is a guaranteed 10,000:1 CMRR over the amplifier's full ±10V input signal range. (Typical common-mode figures for low-cost FET op amps lie in the 1k-5k region, and then usually for limited input-signal swing.)

Not only is its common-mode rejection greatly improved, but bias current of only 10pA means that current drift errors are commensurate with common mode errors. Key specifications, besides the above, include 30μV/°C drift, 4MHz small-signal bandwidth, and 100kHz full-power response. Model 43J is available from stock, and lists at $20 singly.

*For complete information on both Models 44 & 43, use reply card. Circle B9

Users Save 45% with New Hybrid FET Op Amp

AD511B & ADP511B* USE SAME PACKAGE AS 501B & P501B, HAVE SIMILAR SPECS, COST $17 (VS. $31) IN 100'S, HAVE HIGHER RELIABILITY. SECRET IS FEWER CHIPS, LASER TRIMMING



The AD511 and ADP511 FET-input operational amplifiers offer low cost, high performance, and high reliability in modular minipackage form. They are manufactured by combining a dual monolithic FET chip with a specially-designed monolithic amplifier chip on a laser-trimmed thick-film substrate. This technique provides extremely high performance at a significant reduction in assembly cost . . . thereby permitting a low selling price.

Their predecessors, the 501 and P501, with which they are pin- and spec-interchangeable (except for slightly less bandwidth) are constructed with relatively-complex and expensive hybrid assembly techniques, employing a number of discrete transistor, diode, and capacitor chips on a thin- or thick-film substrate. The resulting manufacturing costs—and selling prices—are considerably greater than those of the AD511 and ADP511.

The AD511 and ADP511 are available in A, B, and C specification categories and are pin for pin (in identical packages) low-cost replacements, *not only for 501 and P501*, but also for model numbers 20-008 and 20-108 (Bell and Howell) and 1408 (Teledyne Philbrick).

KEY SPECIFICATION COMPARISON

Specifications (typical @ +25°C except as noted)	Device Type	AD511B ADP511B	501B P501B	20-008 20-108	140801	Units
	Manufacturer	Analog Devices	Analog Devices	Bell & Howell	Teledyne Philbrick	
Open-Loop Gain						
dc rated load (min)		25k	25k	25k	250k	V/V
10kΩ load (min)		100k	100k	100k	500k	V/V
Rated Output, min, Voltage/Current		±10/±5	same	same	same	V/mA
Offset Voltage and Bias Current						
Initial offset voltage (max)		1	same	same	same	mV
Average vs. temp. (−25 to +85°C)		25	same	same	same	μV/°C
Initial bias current (max)[1]		10	10	5	10	pA
Average vs. temp. @ 25°C[1]		1	1	0.5	1	pA/°C
Input Impedance						
Differential or Common-mode		10^{11} ‖ 2	10^{11} ‖ 4	10^{11} ‖ 4	10^{11} ‖ 4	Ω ‖ pF
Common-Mode Input						
Voltage range (min)		±10	±10	±7	±10	V
CMR @ ±5V		86	80	80	80	dB
Differential input (abs. max)		±30	±15	30	30	V
Power Supply						
Voltage (rated spec)		±15	same	same	same	V
Voltage range		±10 to ±22	same	same	same	V
Current, quiescent (max)		±7	±9	±7	±5	mA
Dynamic Response						
Unity-gain frequency (small signal)		1	4	4	4	MHz
Full power response (min)		50	70	70	100	kHz
Slew rate (min)		3	3	3	6 (typ)	V/μs
Temperature Range						
Operating		−25 to +85	same	same	same	°C
Storage		−55 to +125	same	same	same	°C
Price (U.S. $)						
1–9		$24.	$35.	$30.	$33.	
10–24		$21.	$33.	$28.	$31.50	
25–99		$19.	$32.	$26.	$30.50	
100+		$17.				

[1] Doubles every +10°C

*For further information on AD511 & ADP511, use reply card. Circle B10

Application Brief

Long-Term Drift Measurements on Op Amps

DESIGN PREDICTIONS SHOWN TO BE CONSERVATIVE

Representative samples (at least 5 units of each type) of chopper, low-drift chopperless, and low-cost operational amplifiers have been placed on long-term drift test for periods exceeding 1 year. Typical preliminary results are shown here.

The units were kept in oil baths, and buffered by thermal insulation from short-term transient effects. Ambient temperature was measured each time readings were taken, and the readings were individually corrected for temperature, except where noted.

In some cases (the low-drift chopperless units seemed especially susceptible because of the sensitive scale), artifacts turned up that were correlatable between units, but were left in the data for the sake of completeness. These could probably be explained in terms of occasional power outages, and at least two physical moves of the equipment from one location to another.

The data, as a whole, are quite reassuring, in that they seem to indicate that—at least over a year—there is minimal tendency for drift to increase continually. The measurements also indicate that the "long-term drift specifications" on the data sheets were realistic, at worst, and generally quite conservative, and that low long-term drift is characteristic of units designed for low thermal drift.

TEST RESULTS

Model 230 is a chopper-stabilized amplifier whose specified temperature drifts range from 0.5μV/°C down to less than 0.1μV/°C. The data plotted here are uncorrected for temperature coefficient. All units had the same type of history; the "best" and "worst" histories are reproduced here. Although there seems to be a seasonal correlation, the average temperature variation (27°C average in winter to 25°C in summer) does not directly account for it. Nevertheless, *including* temperature effects in the laboratory, the specified typical long-term drift of 5μV/year seems quite conservative. The 231, 232, and 233 families should have long-term drift performance similar to that of the 230 family.

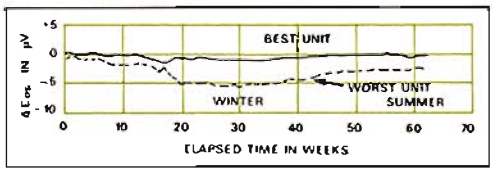
Model 180 is a chopperless unit designed for low offset-voltage and current temperature coefficient (1.5 to less than 0.5μV/°C and 0.05nA/°C to less than 0.02nA/°C). It is also designed for low warmup drift. As can be seen in the curves, the maximum (corrected) voltage drift of this typical unit over a year was less than 10μV, and the maximum (uncorrected) current drift was less than 7nA. Considering an average 5-week "monthly" period, the respective fluctuations are of the order of 4μV and 2nA p-p. The worst unit was about twice as bad. Some of the worst swings seemed to be correlated from unit to unit, and may have resulted from artifacts, as mentioned above.

Model 183 is also a chopperless unit designed for low cost, with about 10X the current drift of the 180 family. Except for the artifacts, as noted above, its long-term (uncorrected) drift is commensurate with that of the 180 family. The 184 and 153 families should have similar performance.

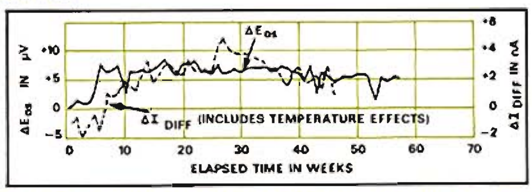
Model 118 is a low-cost unit having a bipolar-transistor input circuit, with average E_{os} TC of 5 to 20μV/°C. The long-term (corrected) drift can be seen to be commensurately larger than that of the low-drift units, but with smaller medium-term (monthly) fluctuations. Its 200μV/month "typical" specification would appear to be grossly conservative. Actually, none of the units tested exceeded 400μV in 9 months. The 119, 163, and 165 families should have similar performance. The design of 118 has been improved since these units were put on test, and should now be—if anything—better.

Model 144 is a low-cost FET-input unit, with average E_{os} TC of 30 and 100μV/°C. Its (corrected) long-term drift over a 4-month period appears to be commensurate with—perhaps less than—that of the 118 family. Again, the 250μV/month specification appears unduly conservative. The 40 family should have comparable performance.

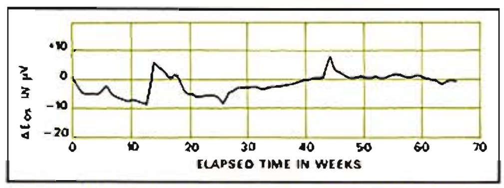
MODEL 230 CHOPPER-STABILIZED AMPLIFIER (UNCORRECTED – INCLUDES TEMPERATURE EFFECTS)



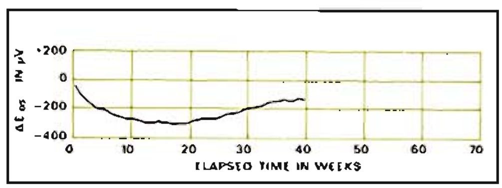
MODEL 180 LOW-DRIFT CHOPPERLESS (COMPENSATED) AMPLIFIER



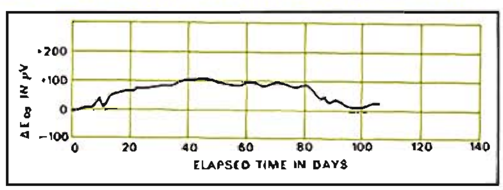
MODEL 183 LOW-DRIFT CHOPPERLESS AMPLIFIER



MODEL 118 LOW-COST BIPOLAR AMP. (DISCRETE)



MODEL 144 LOW-COST FET-INPUT AMPLIFIER

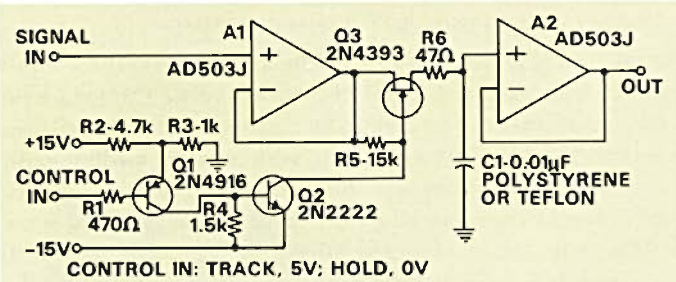


The tests reported on here were designed by and performed under the direction of Bob Demrow, Manager of Applications.

Application Briefs

Two Circuits that Use the AD503* to Advantage

TRACK-HOLD AMPLIFIER



The low input current and high slew rate of the AD503J make it an excellent device for track-hold applications. This circuit will track a ±10V input signal at frequencies up to 4kHz. When the control input changes from Track (+5V) to Hold (0V), the series FET switch, Q3, opens, and the input signal voltage is retained on capacitor C1. The output amplifier, A2, provides a high input impedance to keep C1 from discharging too rapidly.

The drift rate in Hold is determined primarily by the "off" leakage current of Q3, which tends to be greater than that of the amplifier, A2 (25pA max, 5pA typical for AD503J). For example, at 100pA leakage current, the drift rate (for C_1 = 0.01μF) is 10mV/s, and the rate doubles for every 10°C increase of temperature. Lower drift rate and higher accuracy—at the proportional expense of slower acquisition time—can be had by increasing the value of C1. The capacitor should be a type having low dielectric absorption (typically its dielectric would be polystyrene or teflon).

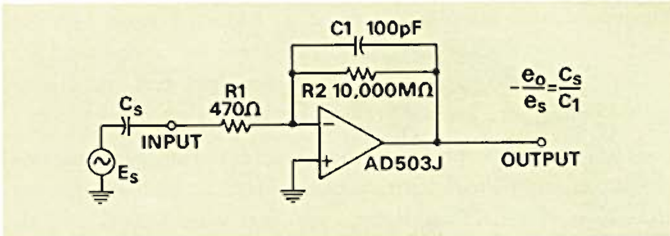
The switching FET, Q3, has low pinchoff voltage, and allows the circuit to handle ±10V signal voltages with standard ±15V supply. In the Track mode, with +5V applied to the control input, Q1 and Q2 are cut off, and the gate of Q3 is at the same voltage as A1's output. Thus, the FET is zero-biased for any value of input and has a resistance less than 100Ω. Resistor R5 adds to the "on" resistance so as to better isolate the capacitive load, C1, from the input follower, A1, to prevent ringing. In the Hold mode, both Q1 and Q2 conduct and pull the gate of Q3 toward -5V. When the gate voltage drops to about 3V below the source (about 100ns after a step change to zero control voltage), the capacitor voltage ceases to track the input. Because of capacitance from the gate to the drain of Q3, the gate swing causes the small transferred charge to produce a small step (offset in Hold) in C1's voltage. Typically less than 10mV over the ±10-volt input range, this step is proportional to the gate voltage swing (15V + V_{in}).

There are also settling transients in A1 and A2, which increase the settling time to within 1mV of final value to about 2μs. For a 10-volt step applied during the Track mode, the settling time to within 1mV of final value is less than 15μs (caused by the limited charging rate of C1, and roughly proportional to C1).

The dielectric absorption of C1 may account for an additional 3mV of error if the input signal is changing rapidly at the time the circuit is gated into Hold.

* For technical data on AD503 FET-input I.C. op amp, use reply card. Circle 811

CHARGE AMPLIFIER



This circuit is designed for use with capacitive sources, such as piezoelectric accelerometers, crystal phono pickups and microphones, and capacitance alarms. The amplifier uses capacitance feedback and has a low input impedance, which makes its scale factor essentially independent of input cable capacitance. Because of the very low input current of the AD503J FET-input amplifier, a very high value of feedback resistance can be used to supply the dc bias current, which results in a low noise level of about 5.5μV rms in a 5Hz to 10kHz bandwidth, when connected to a 100pF source.

The voltage gain is determined by the ratio C_1/C_S. Because the capacitance of an input cable appears essentially across the summing junction of an inverting amplifier, it does not affect scale factor. (The amplifier's contribution to scale factor error is determined by loop gain, $(A\beta)^{-1}$. Since $1/\beta = 1 + C_S/C_1 + C_C/C_1$, and A is open-loop gain, the effect of cable capacitance on gain and linearity error is $C_C/(C_1 A)$. If C_C = 1000pF, its contribution to gain error would be 10/A. For those frequencies at which gain is decreasing at 6dB/octave, the error would be in quadrature, and therefore affect scale factor magnitude insignificantly for $A > 100$.)

With the component values shown, the circuit has a gain of unity when used with a 100pF source. Its frequency response extends from 0.16Hz to 800kHz (within 3dB). Resistor R1 isolates the capacitive source from the feedback loop in the megahertz region and helps to reduce overshoot.

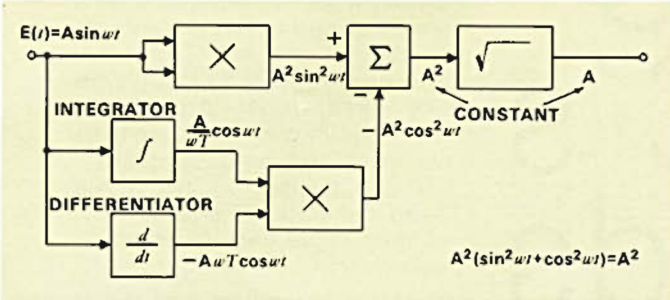
For a gain of 10 with a 100pF source, the feedback capacitance, C_1, can be reduced to 10pF. Because of the very small capacitances in the circuit, care is required in the mechanical construction. Small mechanical vibrations can cause the input capacitance to the supply leads to vary, making in effect, a capacitor microphone, which may have substantial output. Also, because $\Delta Q = \Delta(CE) = C\Delta E + E\,\Delta C$, variations in C1 due to vibration will develop an additive error from the output bias offset voltage.

At low frequencies the response is -3dB at the frequency at which the reactance of C1 is equal to the resistance of R2. The maximum permissible value of R2 is determined by the allowable output offset and the input current of the amplifier at the highest operating temperature $[E_o(\max) = I_b(\max)\,R_2(\max)]$.

Noise in an audio-frequency bandwidth is minimum when R2 has the highest permissible value, because the noise current furnished by R2 is inversely proportional to the square root of its resistance and the gain for amplifier voltage noise is determined (in the operating bandwidth) by $1/\beta$.

Amplitudes Without Delay

MEASURING SINE-WAVE AMPLITUDES WITHOUT FILTERING



Here is an ideal† (that actually works) for making use of trigonometric identities to compute the amplitude of a sinusoid without the usual rectifying and filtering. The advantage is, in concept, that the peak value is computed directly, without one's having to wait for a rectifier-filter to settle down.

What makes it practical, these days, is the availability of low-cost operational amplifiers (such as AD741C) and MDSSR's* (such as AD530—see pages 8 and 16).

For the experimenter who is interested in assembling this system, here are a few practical suggestions:

1. It will work best for a single-frequency input. As frequency varies, the dynamic range depends on the square of the ratio of maximum to minimum frequency.
2. The input amplitude should be scaled to ±10V for the largest signal, to make fullest use of the available dynamic range. For variable frequency, the amplitude should be scaled down by the square-root of the ratio of maximum to minimum frequency.
3. The integrator and the differentiator should be scaled to the input frequency. For variable frequency, they should be scaled to the geometric mean between the highest and lowest frequencies of interest.
4. Both integrator and differentiator should become "non-ideal" at low and high frequencies respectively. The scheme is shown in Figure 1.
5. The assumptions that lead to the indicated results are valid only for clean sine waves. Note that if the amplitude varies rapidly, especially at a frequency comparable to (or approaching) the signal frequency, the sidebands will introduce a ripple component.

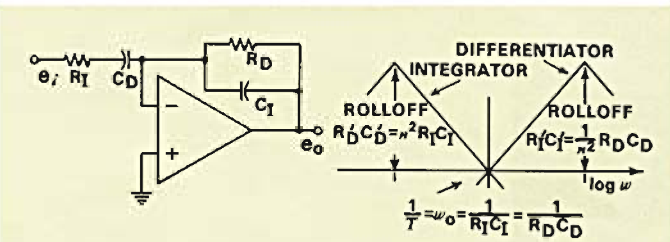


Figure 1. "Nonideal" Integrator/Differentiator circuit. For integrator, use R_IC_I, with $R_D'C_D'$ for rolloff; For differentiator, use R_DC_D, with $R_I'C_I'$ for rolloff; For $n=100$, magnitude error is negligible, phase error is $1.15°$ at ω_0.

†Contributed by Douglas Jolley, McDonnell Douglas Aircraft Company

Worth Reading

"On the Differential Operational Amplifier, a Device that Simulates Almost Anything." From "The Amateur Scientist," *Scientific American,* January, 1971. Here is a practical and introductory treatment of the operational amplifier from the standpoint of the amateur experimenter. The reader is shown how to construct a breadboard/instrument into which an op amp will be plugged, then introduced to basic circuits (inverter, follower, subtractor, current-to-voltage converter). Then, for more practical uses, he is introduced to amplifier, constant-current circuit, constant-temperature circuit, tuning fork circuit, low-frequency oscillator, and power amplifier. Mainly for the amateur, but it can help broaden the outlook of the occasional or routine user.

Applications of dc Constant Current Source, Hewlett-Packard Application Note No. 128. Here is an excellent discussion of the applications of current sources to resistance measurement, semiconductor device measurements, component testing, cryogenics and electrochemicals. A general section discusses desirable characteristics and approximations. This booklet should be of especial interest to op amp users, because of the ready applicability of the op amp—at very low cost—to current generation, and voltage-to-current transconduction, and hence to the banquet of testing suggestions outlined in the Note.

"Extraction of Square Roots . . . A Useful Analog Instrumentation Technique" by Tom Cate, *Electronic Instrument Digest,* January, 1971, page 7. Since square rooting is a ready application of MDSSR's* (AD530, Model 426, etc.), such applications as linearizing square-law signals, two-component vector computation, and rms computing may be of especial interest. The article is interesting when it talks about applications, but it bogs down in the particularities of design of a special purpose module.

NOTED BRIEFLY

New publications available: Use reply card; Circle number

Internally-trimmed MDSSR's: 427, 428, 432 . . .	B12
8-channel expandable multiplexer, MPX-8A	B13
Ultra-low bias current FET-input Op Amps 41, 42 .	B14
Thin-Film Resistor Networks for Conversion . . .	B2
Modular 12-Bit D/A Converters, DAC-QM, DAC-QS .	B15
Dielectrically-isolated low-offset dual transistors 2N4878, 79, 80	B16

*MDSSR: Multiplier-Divider-Squarer-Square Rooter

The "C" in IC's stands for CIRCUITS

Now that the glamour has worn off the integrated circuit, it's become increasingly clear that more than semiconductor know-how alone is needed to further advance the state of the art. Truly outstanding integrated circuits require exceptional processing skills, including thin-film technology, laser trimming, computer-controlled testing . . . the whole "fourth generation" arsenal . . . *and must begin with outstanding circuit design.*

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Here's a specific example: The AD530 Multiplier/Divider/Squarer/Square Rooter. The first and only MDSSR that combines the output op amp with a zener regulated current source and transconductance multiplier, all on a single chip . . . a completely self-contained circuit. With four simple trim pots you get performance unmatched by any other I.C. multiplier, and rivalling that of the black boxes.

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AD 530K accuracy, 1% max.
AD530J accuracy, 2% max.
Full-power bandwidth, 500kHz

Check the ease of operation: write for our Multiplier Engineering Manual that removes the complexities and problems for multiplier use. And while you're at it, get our Fourth Generation Analog I.C. Brochure on the AD530 and six other stunners. Write for them and *join the fourth generation* . . . you can't beat it.

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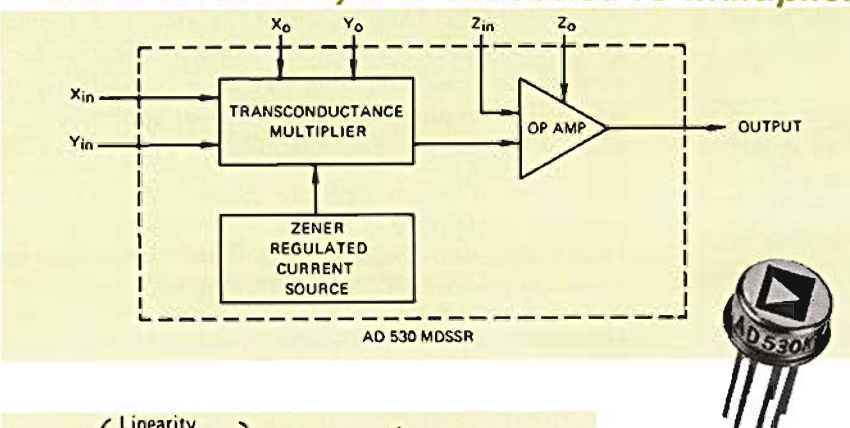

Σ {Linearity, Offset Voltage, Feedthrough, Scale Factor} → Accuracy = {1% Max. . (AD530K); 2% Max. . (AD530J)}

PERFORMANCE PARAMETERS

LINEARITY (X INPUT)	0.5%
LINEARITY (Y INPUT)	0.2%
FEEDTHROUGH (X = 0)	0.5%
FEEDTHROUGH (Y = 0)	0.3%
OUTPUT VOLTAGE (R_L = 2KΩ)	±10V
FULL POWER BANDWIDTH	500 KHz
SLEW RATE	40V/μs
SUPPLY CURRENT	4 mA
OFFSET VOLTAGE VS. SUPPLY VOLTAGE	100 mV/V
SCALE FACTOR VS. SUPPLY VOLTAGE	2%/V
OFFSET VOLTAGE VS. TEMPERATURE	0.2 mV/°C
SCALE FACTOR VS. TEMPERATURE	0.04%/°C
OVERALL ACCURACY VS. TEMPERATURE	0.05%/°C

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