ANALOG SIGNAL PROCESSING

ANALOG COMPUTATION & SIGNAL PROCESSING

This section of our seminar considers techniques for analog computation and signal processing. By exploiting the basic physics of semiconductor devices it is possible to design circuits which perform a wide variety of mathematical operations, including addition, subtraction, multiplication, division as well as trigonometrical, logarithmic, and exponential functions. Such circuits perform in the analog domain and frequently offer real advantages over more conventional digital computation.

BENEFITS OF ANALOG DATA PROCESSING

There is a widely-held belief that analog techniques are obsolete and that every possible electronic operation can today best be performed by digital techniques. This is far from the true state of affairs and there are many operations which are still best performed in the analog domain.

ANALOG SIGNAL PROCESSING
HAS
REAL ADVANTAGES
OVER
DIGITAL COMPUTATION

Analog signal processing can out-perform digital processing on accuracy, cost, complexity, speed, and various combinations of these.

IN SOME CIRCUMSTANCES
ANALOG SIGNAL PROCESSING
IS

- **Faster**
- **Cheaper**
- **More Accurate**
- **Simpler**

THAN
DIGITAL PROCESSING

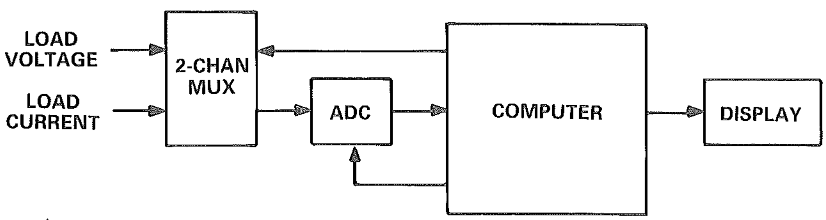
Operations where analog computation is often preferable to digital include those where both the input and output signals must be analog, where limited amounts of processing are required and no digital circuitry is present, where the signal is differentiated to produce a rate signal, where fast signals must be processed in real time, where large dynamic ranges are involved, and where complex or transcendental functions must be evaluated.

ANALOG SIGNAL PROCESSING HAS ADVANTAGES WHERE:

- **Inputs and Outputs Are Both Analog**
- **No Digital Circuitry is Available**
- **Signals are Being Differentiated**
- **Fast Real-Time Results are Required**
- **There are Wide Dynamic Ranges of Input**
- **Complex and Transcendental Functions are Being Computed**

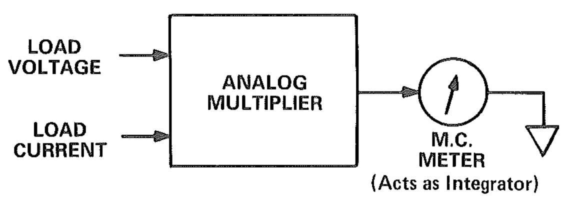
It is important that each case is judged on its merits. The dedicated analog engineer can solve almost any processing problem in a purely analog manner—the dedicated bit-freak in a digital way. The balanced approach is to consider each problem in its own terms and make a decision based on all the factors present. For example a simple ac power meter integrates and displays the product of voltage and current at a load and may be constructed very easily with a single analog multiplier—whereas a digital power meter would require conversion of both voltage and current at the digital form (with considerable attention to the timing of the conversions since the relative phase of the two signals is of critical importance) and then a digital multiplication followed by output to a display. If I had nothing and simply wanted a power meter I should use the analog approach (which needs only a multiplier and a meter), but if I already had a computer, a display driven by it, and a multiplexed ADC with spare capacity, I would be foolish to do so since the cost of adding a power monitor facility to my existing hardware and software would be minimal.

DIGITAL POWER MONITOR



NOTE: CONVERSION TIMING MUST PRESERVE PHASE INFORMATION

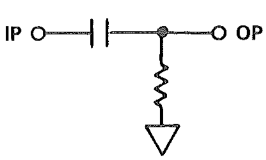
ANALOG POWER METER



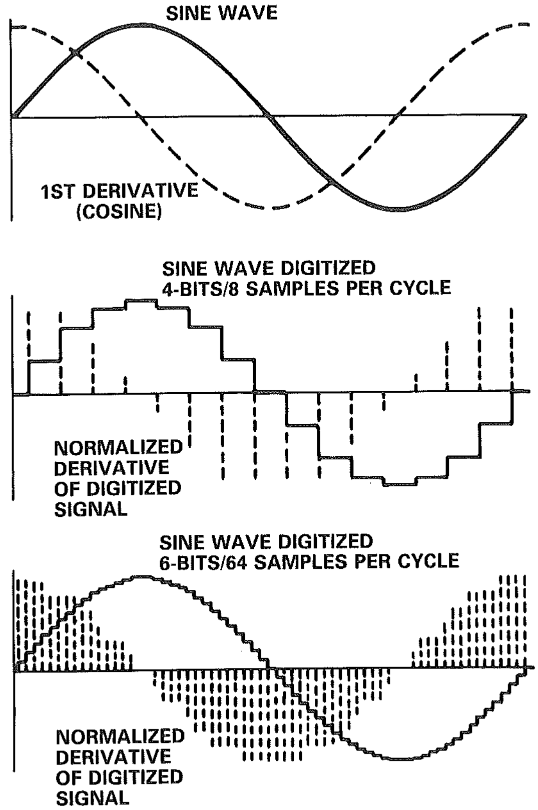
If the computer were busy, though, even if the final result were required in digital form it might make sense to compute the power with an analog multiplier and then convert its output rather than convert and then compute—simply to save CPU time. This is more likely to be useful where more complex computations are involved but could be significant even for simple multiplication.

The first derivative of a varying analog signal can be computed with a capacitor and a resistor.

ANALOG DIFFERENTIATION

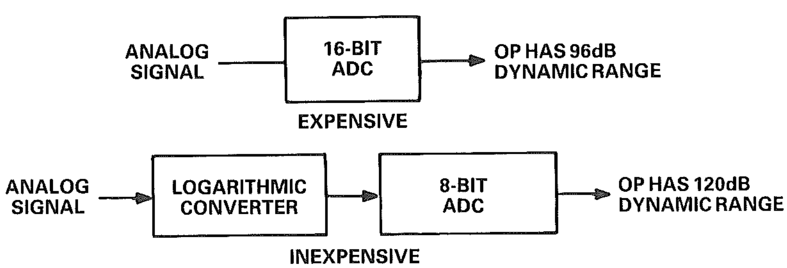


If the analog signal is first digitized not only must a more complex method be used to obtain the derivative but the digitizing must be done at high resolution in both time and amplitude—which is expensive.



Digitizing a signal with wide dynamic range is also expensive. If such a signal is digitized with a 16-bit ADC (which is a very expensive device) the ratio of an LSB to full-scale is 96dB whereas if the signal is first applied to a logarithmic converter (frequently misnamed a logarithmic amplifier) then a dynamic range approaching 120dB is practical with an 8-bit ADC.

ANALOG PROCESSING IS LESS EXPENSIVE AT HIGH DYNAMIC RANGE



ANALOG COMPUTATION CIRCUITRY

The operation of any analog computational circuits depends on the logarithmic properties of silicon junctions. An ideal logarithmic diode has the current voltage relationship:

$$I = I_O\left(e^{\frac{qV}{kT}} - 1\right) \quad \text{or} \quad V = \frac{kT}{q}\ln\left(\frac{I}{I_O} + 1\right) \simeq \frac{kT}{q}\ln\left(\frac{I}{I_O}\right)$$

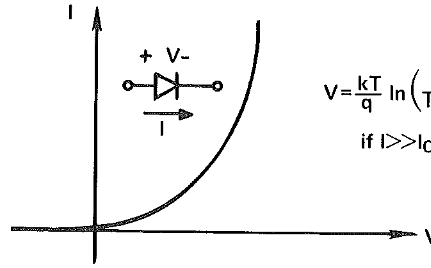
AN "IDEAL LOGARITHMIC DIODE" HAS THE CURRENT-VOLTAGE RELATIONSHIP:

$$I = I_O\ (e^{qV/kT} - 1)$$

where:

- I is the current through the diode
- V Is the voltage across the diode
- q is a constant equal to the unit charge, 1.60219×10^{-19} coulombs
- k is Bolzmann's constant, 1.38062×10^{-23} joules/°Kelvin
- T is the absolute temperature (in °Kelvin)
- I_O is the extrapolated current for $E_O = V = 0$ Volts

DIODE CURVE



$$V = \frac{kT}{q}\ln\left(\frac{I}{I_O}\right)$$

if $I >> I_O$

These equations clearly tell us that the current in a diode increases exponentially with voltage or, conversely, the voltage increases logarithmically with current. What they do not show so clearly is that I_O, the theoretical diode current at zero voltage, is temperature dependent and so the variation of a diode's behavior with temperature is by no means as simple as the equation would suggest— i.e., the voltage is not proportional to absolute temperature at a fixed current. There are several approximations concerning the logarithmic behavior of diodes which are worth remembering:

DIODE APPROXIMATIONS

$$\frac{kT}{q} = 26\text{mV (exact at } 28.58°\text{C)}$$

$$\frac{kT}{q}\ \ln(10) = 60\text{mV (exact at } 29.25°\text{C)}$$

This simplifies the diode expression to:

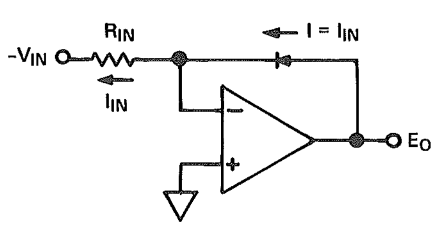
$$V = 60\text{mV} \log \frac{I}{I_O}$$

or V increases 60mV every time current increases by a factor of 10

$$\frac{d}{dT}\frac{kT}{q} = 0.34\%/°\text{C} @ 25°\text{C}$$

If we were to place an ideal logarithmic diode in the feedback path (output to inverting input) of an operational amplifier and apply a current to the inverting input the output voltage would be the logarithm of the input current times a temperature varying constant.

DIODE LOG CONVERTER



$$E_O = \frac{kT}{q}\ \ln\ \left(\frac{I_{IN}}{I_O}\right) \cong 0.06 \log \frac{V_{IN}}{R_{IN}\, I_O}$$

if $I_{IN} >> I_O$

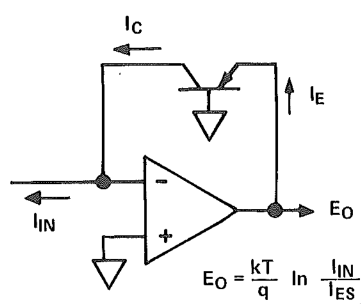
It is unfortunate that real diodes are not ideal logarithmic diodes. In a real diode the bulk resistivity, R_B, of the silicon limits the logarithmic accuracy at high currents and diffusion currents in surface inversion layers and generation-recombination effects in space-charge regions cause a scale factor error, m, at low currents. We thus find that:

$$E_O = m\frac{kT}{q}\ln\left(\frac{I}{I_O}\right) + IR_B \text{ (m varies with current)}$$

Even with similar diodes m can vary (it is never less than 1 and may be as high as 4), as does the value of E_O at which m changes. General purpose diodes are thus impractical as logarithmic diodes for dynamic ranges of more than 100:1 (2 decades).

Luckily we can replace the diode with a grounded-base transistor and get a dynamic range of 1000000:1 (6 decades) or more—the only disadvantage of such a circuit is that the signals can only have a single polarity (so does the logarithm function—but a "bipolar logarithmic converter" might have practical applications).

TRANSISTOR LOG CONVERTER



From the Ebers and Moll equations* it may be shown that:

$$E_O = \frac{kT}{q}\ln\left(\frac{I_{IN}}{I_{ES}}\right) - \frac{kT}{q}\ln(\alpha n) \text{ where } I_{IN} >> I_{ES}$$

Where I_{ES} is the emitter saturation current and αn is the forward current-transfer ratio (αn is NOT the grounded-base current gain).

Since I_{ES} is less than a picoampere, and αn is nearly unity over a wide range of currents, in the silicon planar transistors used to manufacture logarithmic converters the effect of the second term may generally be disregarded and the equation simplifies to:

$$E_O = \frac{kT}{q}\ln\left(\frac{I_{IN}}{I_{ES}}\right)$$

Such logarithmic converters are temperature sensitive. kT/q has a TC of 0.34%/°C around 25°C, and I_{ES} doubles for every 10°C temperature rise (and varies with device size and geometry). However if we have two transistors matched for V_{BE}, which is relatively simple to do on an integrated circuit, the ratio of their αI_{ES} tends to be stable with temperature:

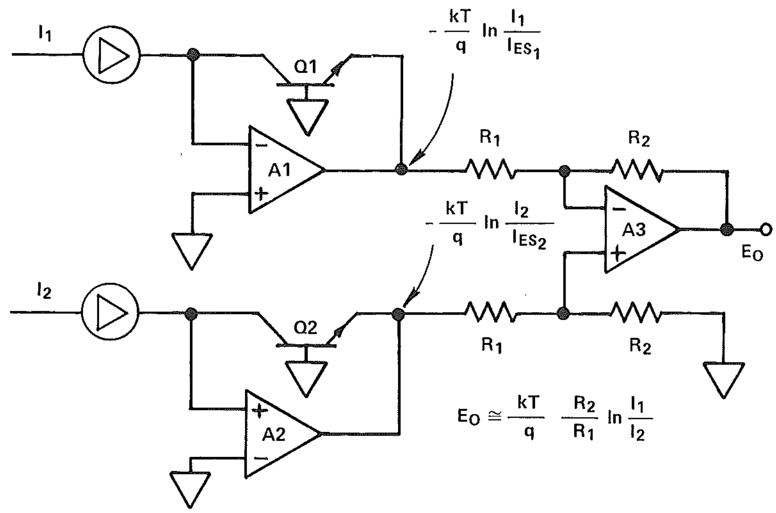
$$\frac{kT}{q}\ln\left(\frac{I1}{\alpha I_{ES1}}\right) - \frac{kT}{q}\ln\left(\frac{I2}{\alpha I_{ES2}}\right) = \frac{kT}{q}\ln\left(\frac{I1}{I2}\right) + \ln\left(\frac{\alpha I_{ES2}}{\alpha I_{ES1}}\right)$$

$\ln\left(\frac{\alpha I_{ES2}}{\alpha I_{ES1}}\right)$ term is zero if $\alpha I_{ES1} = \alpha I_{ES2}$.

*See "Nonlinear Circuits Handbook", Daniel H Sheingold, ed, Analog Devices 1974 for a detailed derivation.

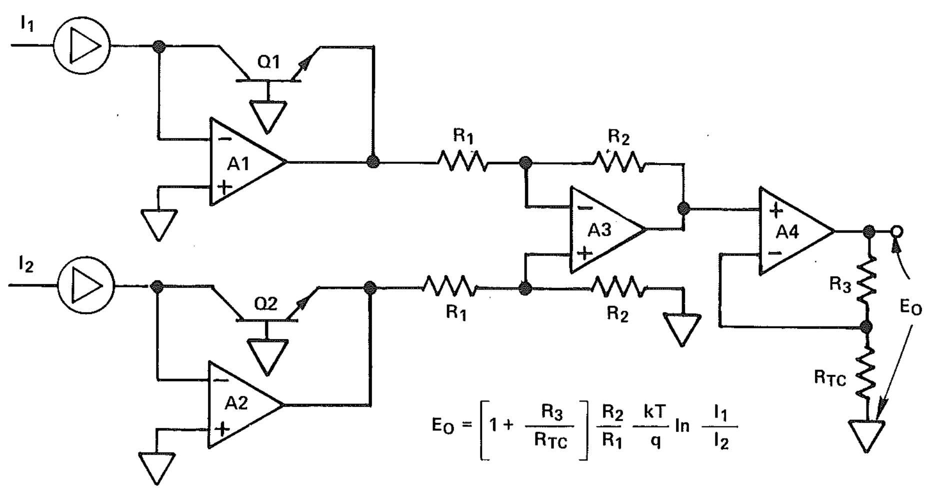
If transistors Q1 and Q2 are identical the circuit below will compensate for I_{ES} variation and give us an output voltage which is proportional to kT/q times the logarithm of the ratio of its input currents.

LOG RATIO CIRCUIT WITH TEMPERATURE-COMPENSATED I_{ES}



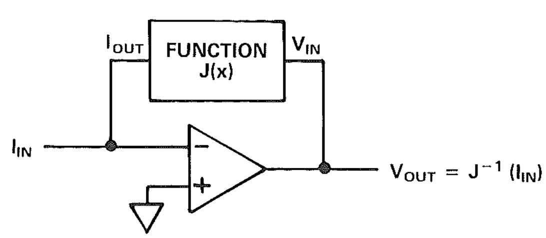
By adding a second amplifier with temperature varying gain it is possible to compensate for kT/q as well—such an amplifier uses a thermistor in its feedback path and if a suitable thermistor is chosen (and it is maintained at the same temperature as the transistors Q1 & Q2) the final circuit shown will have a logarithmic ratio response which is more or less stable over temperature. A suitable thermistor is the Q-81 from Tel Labs Inc. [154G, Harvey Road, Londonderry, NY 03053. (603) 625-8994 Twx: (710) 220-1884]. Like the simple logarithmic converter these log ratio circuits will work with signals of one polarity only.

LOG RATIO CIRCUIT WITH COMPENSATION FOR BOTH I_{ES} AND kT/q



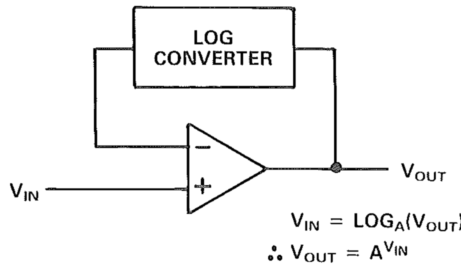
A basic technique in the design of analog computation circuits is that inverse function generators may be produced by using a function generator in the feedback path of an operational amplifier.

INVERSE FUNCTION GENERATOR MAY BE BUILT FROM A FUNCTION GENERATOR AND AN OPERATIONAL AMPLIFIER



This only works for a monotonic function, of course, or over a range of values where a function remains monotonic—otherwise the negative feedback becomes positive feedback and the circuit either oscillates or latches up. The logarithm is a monotonic function and therefore an operational amplifier with a logarithmic converter in its feedback path will act as an antilogarithmic converter, or exponentiator.

LOGARITHMIC CONVERTER IN A FEEDBACK PATH YIELDS AN ANTILOGARITHMIC CONVERTER (OR EXPONENTIATOR)

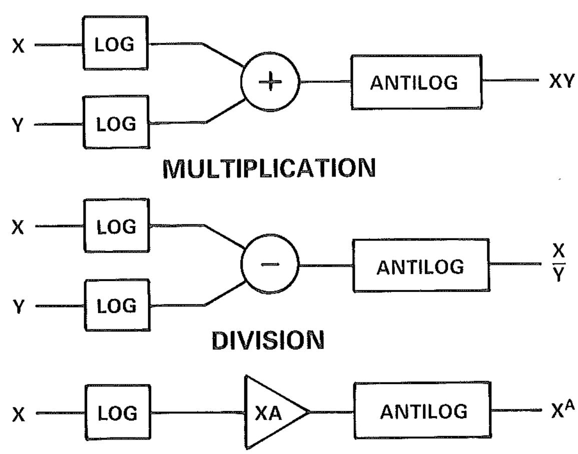


If the logarithm and antilogarithm functions are performed on the same chip it is reasonable to assume that the temperature of the logging transistor in both circuits is the same and therefore not only is I_{ES} compensation unnecessary—so is kT/q compensation. Remember the basic properties of logarithms and antilogarithms:

$$\log(x) + \log(y) = \log(xy) + \log(x) - \log(y) = \log(x/y) + a\ \log(x) = \log(x\ a) + \text{antilog}(\log(x)) = x$$

it is obviously possible to combine several log and antilog circuits on a chip, together with summing circuitry and amplifiers, to make temperature-stable circuitry capable of multiplication, division, and power computations.

COMPUTATION WITH LOG AND ANTILOG CIRCUITS



The AD538 is just such a chip. It contains three logarithmic and one antilogarithmic circuits, together with an adder, a subtractor, and an amplifier whose gain, M, may be set to values between 0.2 and 5. It has three inputs X, Y, and Z and its transfer function is:

$$E_O = Y\left(\frac{Z}{X}\right)^M$$

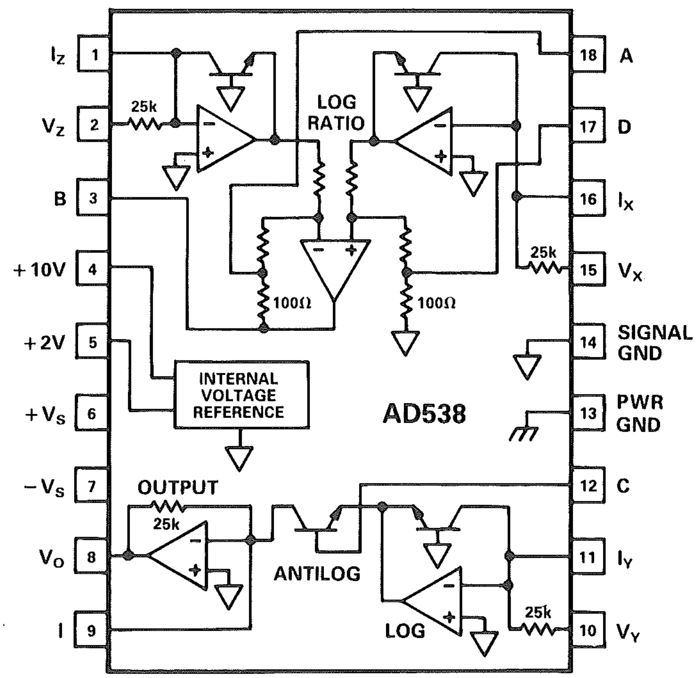
Since any of X, Y, Z, and M may be set to unity it can be used as a multiplier, or a divider, or for raising to powers (squarer, square rooter, cuber, cube rooter, etc.) It is particularly valuable as a divider with a wide denominator range but, of course, will only work with unipolar values of X, Y and Z.

AD538 FEATURES

$V_{OUT} = V_Y \left(\frac{V_Z}{V_X}\right)^m$ **Transfer Function**

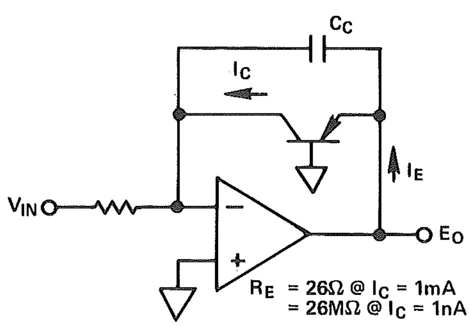
Wide Dynamic Range (Denominator) – 1000:1
Simultaneous Multiplication and Division
Resistor-Programmable Powers & Roots
No External Trims Required
Low Input Offsets <100μV
Low Error ±0.25% of Reading (100:1 Range)
+2V and +10V On-Chip References
Monolithic Construction

AD538 FUNCTIONAL BLOCK DIAGRAM



This type of logarithmic converter has another disadvantage as well as its unipolar inputs (although the AD538 is not very seriously affected by it)—its bandwidth varies with signal amplitude.

LOG CONVERTER COMPENSATION PROBLEM

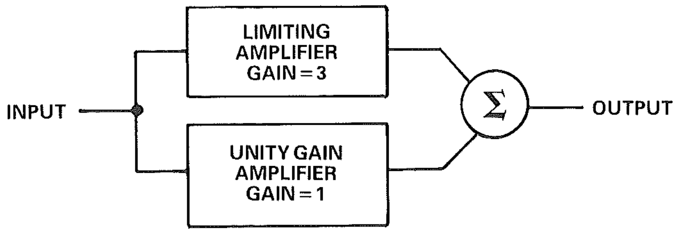


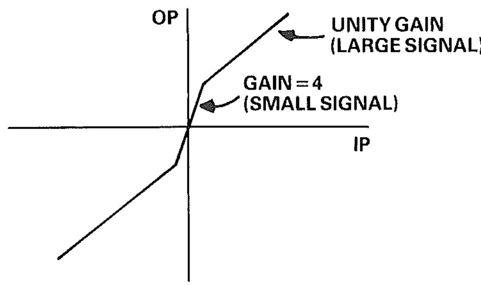
The problem results from the variation of emitter resistance, R_E, with current in the grounded-base transistor. This resistance is inversely proportional to the emitter current, being approximately 26Ω at 1mA. The bandwidth of the circuit is inversely proportional to the product of R_E C_C (which may be an external compensation capacitor or merely stray capacity) and is thus proportional to the transistor current. Thus if the logarithmic converter works over 120dB dynamic range its bandwidth will vary by 1,000,000:1—which can be inconvenient.

Adding some extra resistance (usually a few kilohms) between the amplifier output and the emitter will improve the situation at higher currents and the use of fully integrated logarithmic converters with very low values of C_C can also help but the maximum bandwidth of such circuits is rarely more than 100kHz and frequently much less. At higher frequencies other techniques must be used.

A conceptually simple technique offering wide bandwidth and large dynamic range (and sometimes known as a "true log amplifier") uses a number of similar amplifiers in cascade. Each amplifier consists of a limiting amplifier and a unity gain amplifier whose outputs are summed.

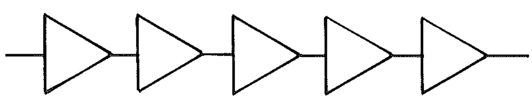
STRUCTURE AND PERFORMANCE OF "TRUE LOG AMPLIFIER" ELEMENT

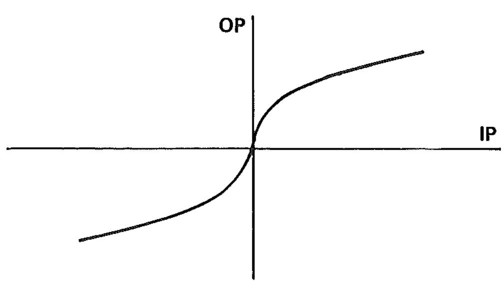




When a small signal is applied to the amplifier both halves contribute to the output and the gain, in the amplifier shown, is 4 times (12dB)—when the signal is increased the limiting amplifier goes into its limiting region and contributes a fixed amount to the output and the gain drops to unity (0dB). If a number of such stages are cascaded the gain of the strip to small signals is the product of the gains of all the stages, but as the input is increased the lasts stage limits and the gain drops by 12dB, then the next stage limits and the gain drops by a further 12dB, and so on until all the stages have limited and the gain is unity. If we plot the transfer characteristic of such a strip of cascaded amplifiers we find that it consists of a series of straight lines but approximates quite closely to a logarithmic law.

LOG AMPLIFIER FORMED BY CASCADING SEVERAL STAGES

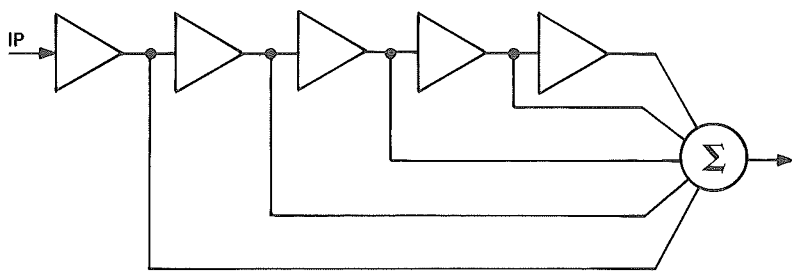




In practice, if the limiting amplifier limits "softly" (i.e., does not step suddenly from full gain to zero gain but does so over a small change of input), the approximation to a logarithmic response can be within 0.1dB. Practical logarithmic converters of this type are quite often used in the intermediate frequency (IF) sections of radar receivers and can be made with dynamic ranges of up to 100dB (100000:1). The response is not truly logarithmic, of course—it is symmetrical about zero, while the function ln (x) is indeterminate for negative values of x, and it is also not logarithmic for very small positive values of input when a true logarithmic function would give rise to a negative output. Such logarithmic converters are also, usually, ac coupled and work for inputs between a few MHz and a few hundred MHz which is fine for radar receivers but less so for analog computation. There are several varients of the basic concept.

If we replace the complex stages consisting of a limiting amplifier, a unity gain amplifier, and a summing circuit with a simple limiting amplifier and use a single summing circuit to sum all the amplifier outputs we have a logarithmic converter which is functionally equivalent to the one we have just described, but simpler.

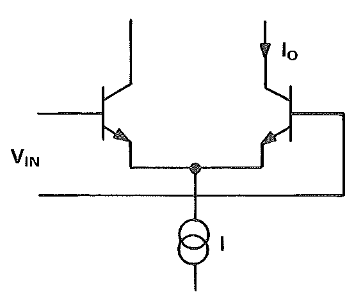
LOG AMPLIFIER FORMED WITH LIMITING AMPLIFIERS



This architecture is also used in radar systems, with ac coupled amplifiers and current output detectors making the connection from each amplifier output to the summing point (it is slightly easier to sum currents than voltages, especially at high frequencies). It is known as a "successive detection" logarithmic amplifier (though the term "logarithmic amplifier" is somewhat of a misnomer and "logarithmic converter" better describes the function). The ac coupling of logarithmic converters developed for radar IF strips makes them almost useless for low frequency or dc analog computation (the use of capacitors integrated on the chips of such limiting amplifiers results in low frequency limits well over 1MHz in many cases), and the difficulties of combining low offsets with high frequency performance have prevented the technique being used at dc or LF.

Logarithms and antilogarithms are by no means the only functions that may be generated with silicon transistors. The basic long-tailed pair of transistors (which consists of two similar transistors with a current fed to their joined emitters and a differential signal applied between their bases) has hyperbolic tangent transconductance. With suitable circuitry this can be modified to approximate a sinusoidal characteristic over the range $-90°$ to $+90°$.

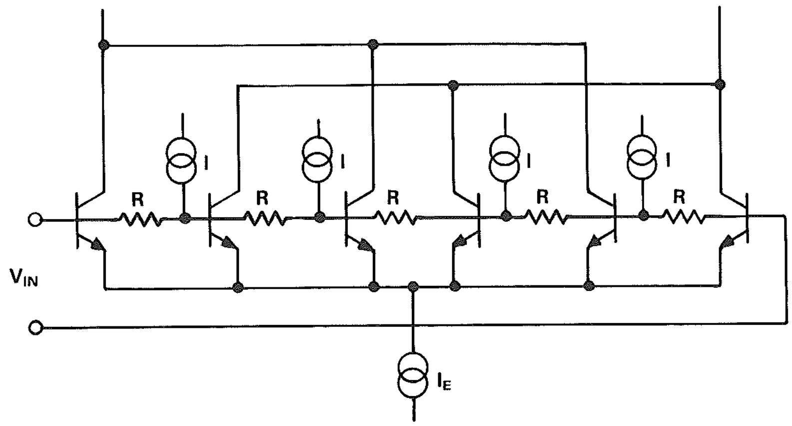
TRANSFER CHARACTERISTIC



A LONG-TAILED PAIR
OF TRANSISTORS HAS
A HYPERBOLIC TANGENT
TRANSFER CHARACTERISTIC

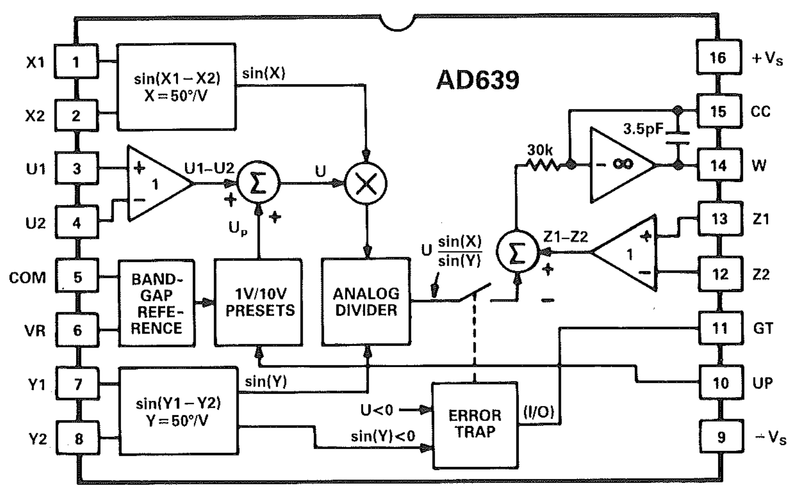
Gilbert* has shown that it is possible to design a temperature-compensated sine-shaper circuit having a dynamic range of over 1000° by cross-connecting six transistors, five equal resistors, and four equal current sources in what is basically an extension of a long-tailed pair.

1000° SINE FUNCTION CIRCUIT



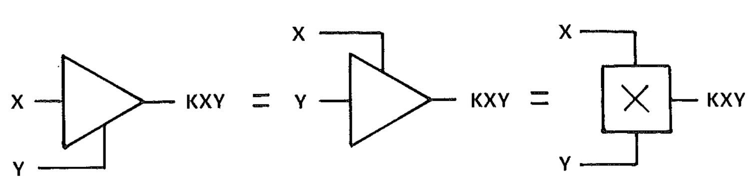
The AD639 contains two such sine-shaper circuits with fully differential inputs having 50°/V scaling, plus a voltage reference of 1.8V (=90°), a divider, a scaling circuit with fully differential input, and an output amplifier. It may be configured to compute sine, cosine, tangent, secant, cosecant, and contangent and, using the function generator in the feedback of the amplifier as mentioned above, to compute the inverse functions as well.

AD639 BLOCK DIAGRAM



The properties of a long-tailed pair of transistors are also used in transconductance multipliers. A transconductance multiplier is a circuit which may be considered as an amplifier with signal-controlled gain—there are two input ports and the gain from one to the output is proportional to the signal at the other. Since the gain from the other to the output is self-evidently proportional to the signal at the one there seems little to be gained by representing the multiplier as an amplifier with a gain-control terminal and so we represent it as a multiplying black box.

A MULTIPLIER MAY BE CONSIDERED AS AN AMPLIFIER WITH GAIN PROPORTIONAL TO CONTROL INPUT



*IEEE Journal of Solid-State Circuits, SC-17 No:6, December 1982: "A Monolithic Microsystem for Analog Synthesis of Trigonometric Functions and their Inverses".

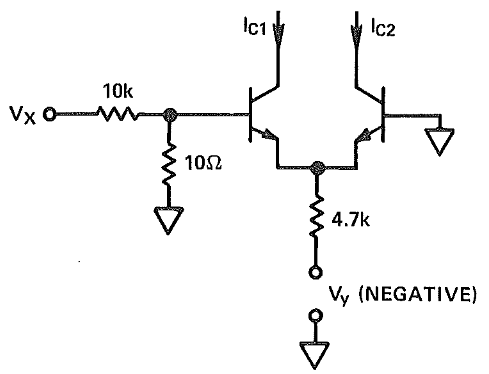
There is a linear relationship between the collector current of a silicon junction transistor and its transconductance (gain) which is given by the following equation:

$$\frac{dI_C}{dV_{BE}} = \frac{q}{kT} I_C$$

where I_C = collector current
V_{BE} = base-emitter voltage
q = electron charge (1.60219×10^{-19})
k = Boltzmann's constant (1.38062×10^{-23})
T = absolute temperature ($K = °C + 273.15$)
q/kT = $1/(25.69\text{mV})$ at 25°C

We can exploit this relationship in a long-tailed pair of transistors to construct a multiplier:

BASIC TRANSCONDUCTANCE MULTIPLIER CIRCUIT



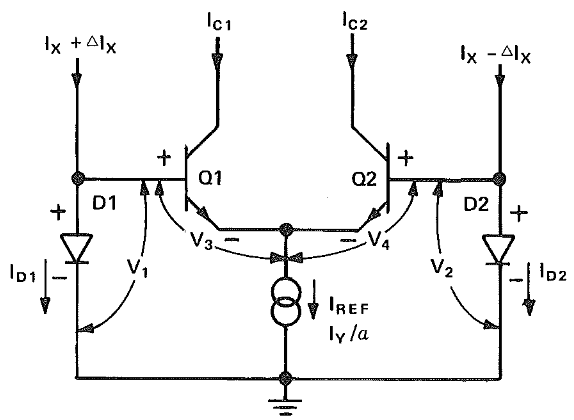
$$I_{C1} - I_{C2} = \Delta I_C = \frac{q}{kT}\left(\frac{V_y + V_{be}}{4.7 \times 10^3}\right)\left(\frac{10}{10{,}010}\right)V_x$$
$$= 8.3 \times 10^{-6}\,(V_y + 0.6)\,V_x \quad @\ 25°C$$

it is, alas, a rather poor multiplier for a number of reasons:

1. The Y input is offset by the V_{BE} which changes nonlinearly with V_Y.
2. the X input is nonlinear because of the exponential relationship between I_C and V_{BE}.
3. The scale factor is temperature variable (-0.34% at 25°C).

We can improve its performance by utilizing the logarithmic properties of diodes in quite a simple circuit.

LINEARIZED 2-QUADRANT MULTIPLIER (PRINCIPLE)



This circuit has a differential current input as its X input and a current as its Y input. Since the differential X currents flow in two diodes the diode voltages, being logarithmic, compensate for the exponential V_{BE}/I_C relationship and, moreover, the q/kT scale factor in the transconductance function is exactly compensated by the kT/q scale factor in the diode response. This gives the circuit, which is known as the "Gilbert Cell" after its inventor*, the linear transfer function:

$$\Delta I_C = \frac{\Delta I_X \cdot I_Y}{2I_X}$$

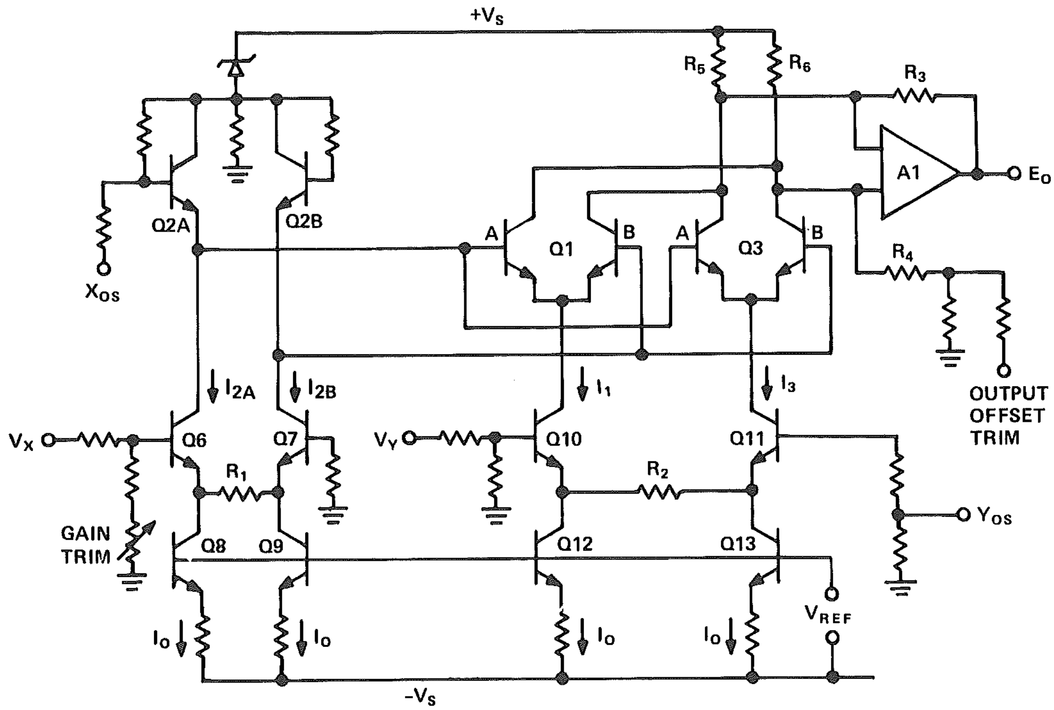
*For further details see the "Nonlinear Circuits Handbook", Analog Devices, pp 217-221.

As it stands the circuit has a number of inconvenient features:

1. Its X input is a differential current,
2. Its Y input is a unipolar current—so the whole cell is only a two-quadrant multiplier, and
3. Its output is a differential current.

By cross-coupling two such cells, by using two long-tailed transistor pairs as input voltage-to-current converters, and by using a subtractor amplifier to convert the differential current output to a single-ended voltage output we can overcome all these inconveniences.

4-QUADRANT VARIABLE-TRANSCONDUCTANCE MULTIPLIER



In the circuit illustrated Q1 and Q3 are the two Gilbert cells, Q2A and Q2B form the linearization diodes (the base-emitter junction of a transistor is, of course, a diode and in applications of this type has a better diode performance than a simple diode), and the amplifier A1 acts as a differential current to single-ended voltage converter. Such transconductance multipliers have many advantages:

ADVANTAGES OF THE VARIABLE-TRANSCONDUCTANCE MULTIPLIER

1) **Good accuracy. Overall errors as small at ±0.1% accuracy (AD534L)**
2) **Wide bandwidth. Bandwidths of 30MHz or more can be realized independent of input signal level (unlike log-antilog)**
3) **Simplicity and low cost.**

The four-quadrant transconductance multiplier relies on the matching of a number of transistors in geometry, temperature, and current. This makes it a practical circuit in monolithic IC form but considerably harder to make with discrete components. Even the best IC process will leave some residual mismatches and these manifest themselves as dc errors—if the circuit is manufactured so that such error may be trimmed a very high performance indeed is possible. The errors, and their effects, are as follows:

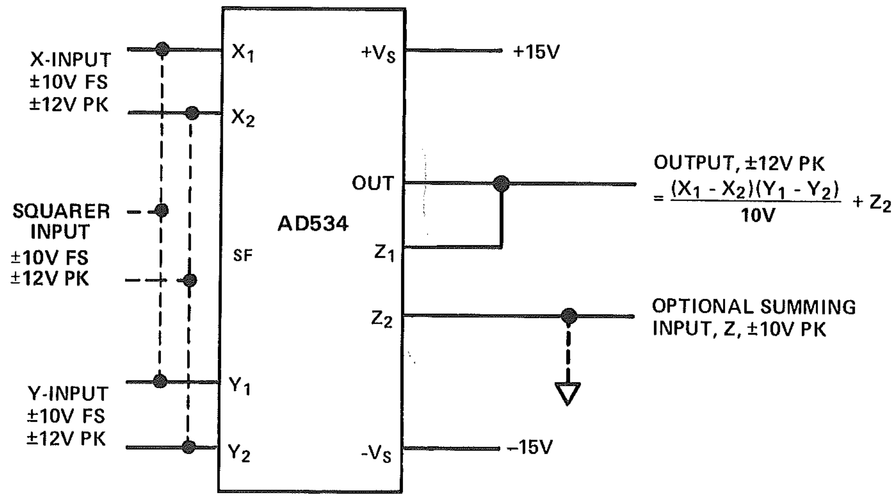
TRIMMABLE ERRORS IN MULTIPLIERS

X-Input Voltage Offset:	Y Feedthrough
Y-Input Voltage Offset:	X Feedthrough
Z-Input (output amplifier) Voltage Offset:	dc output offset voltage
Resistor Mismatch:	Gain Error

Early transconductance multipliers were trimmed by the user but since there are four parameters requiring trimming and second-order effects will always make the trims more or less interactive the process is quite complex. As in so many other products Analog Devices' ability to fabricate integrated circuits with accurate, stable, laser-trimmed resistors reduces customer trims from a dire necessity to an unnecessity. The AD534 is the industry standard transconductance multiplier and is trimmed during manufacture to give very high performance.

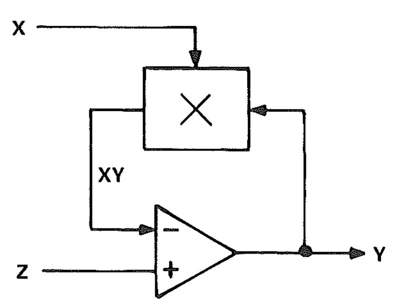
Because the structure of the transconductance multiplier is necessarily differential internally it is generally differential externally (one of a pair of differential inputs can always be grounded if a single-ended input is required) and this enables additional computation to be performed. Applications are better described in the applications section of this part of the seminar but one feature of the AD534 architecture should be explained at once—the summing inputs. It would be quite convenient if the connections to the summing (Z) inputs in the standard multiplier configuration were made internally.

BASIC MULTIPLIER CONNECTION



This would only interfere with multiplier operation when a Kelvin voltage-sensing connection was required—which is not common with computation circuits like multipliers. However such an internal connection would prevent the output of the multiplier being fed back to the Y input and a separate input being applied to Z. This places the multiplier in the feedback path of its own internal amplifier and, as we noticed earlier, when a function generator is placed in a feedback loop the whole system becomes an inverse function generator—in this case a divider.

A MULTIPLIER IN A FEEDBACK LOOP MAKES A DIVIDER



NEGATIVE FEEDBACK FORCES

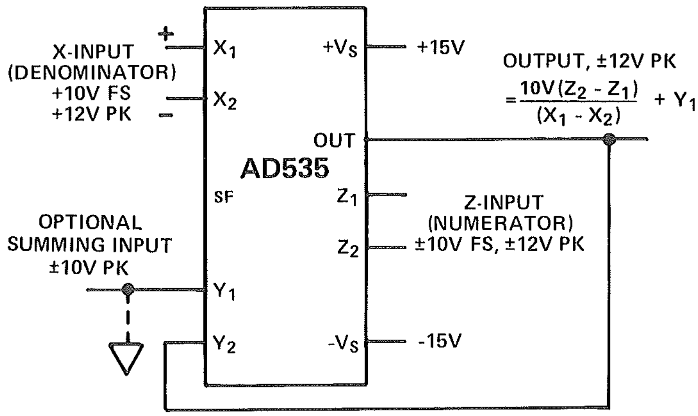
$XY = Z$

$\therefore$ Y (= OUTPUT) $= \frac{Z}{X}$

$\therefore$ CIRCUIT IS A DIVIDER!

The AD534 acts very successfully as a divider but if the error calculations are checked it turns out that a different trim algorithm is required to optimize divider performance from that required to optimize multiplication. The AD534 is therefore manufactured in a version which is factory trimmed as a divider—this circuit is known as the AD535. The chips are identical, only the trim algorithms differ.

THE AD535 DIVIDER



FEATURES

Pretrimmed to ±0.5% max Error, 10:1 Denominator Range
±2.0% max Error, 50:1 Denominator Range
All Inputs (X, Y and Z) Differential
Low Cost, Monolithic Construction

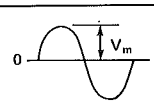
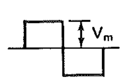
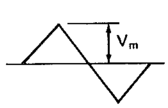
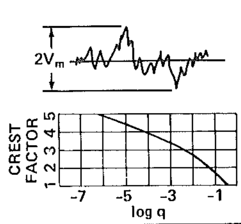
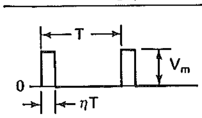
APPLICATIONS

General Analog Signal Processing
Differential Ratio and Percentage Computations
Precision AGC Loops
Square-Rooting

We have considered a number of computational circuits which make use of the logarithmic properties of silicon diodes where some pains are necessary to extend the computational function from one or two quadrants. There is one specific application where this is not necessary—the rms to dc converter.

The rms value of an ac waveform is important to the engineer because it is a measure of the power in the signal. With sine and square waveforms the ratio of rms to peak (or to mean average—strictly mean average deviation (MAD)) value is well-known and measurements may be made by measuring one or the other and applying a correction factor, but where the waveform is unknown or variable the error between the mean and the rms value may be substantial.

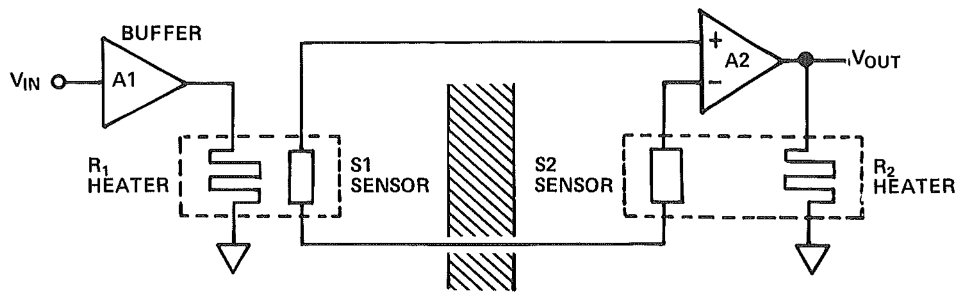
RMS, MAD, AND CREST FACTOR OF SOME COMMON WAVEFORMS

WAVEFORM		RMS	MAD	$\frac{\text{RMS}}{\text{MAD}}$	CREST FACTOR
	SINE WAVE	$\frac{V_m}{\sqrt{2}}$ $0.707\,V_m$	$\frac{2}{\pi}V_m$ $0.637\,V_m$	$\frac{\pi}{2\sqrt{2}} = 1.111$	$\sqrt{2} = 1.414$
	SYMMETRICAL SQUARE WAVE OR DC	V_m	V_m	1	1
	TRIANGULAR WAVE OR SAWTOOTH	$\frac{V_m}{\sqrt{3}}$	$\frac{V_m}{2}$	$\frac{2}{\sqrt{3}} = 1.155$	$\sqrt{3} = 1.732$
	GAUSSIAN NOISE CREST FACTOR IS THEORETICALLY UNLIMITED. q IS THE FRACTION OF TIME DURING WHICH GREATER PEAKS CAN BE EXPECTED TO OCCUR	RMS	$\sqrt{\frac{2}{\pi}}$ RMS = 0.798 RMS	$\sqrt{\frac{\pi}{2}}$ 1.253	C.F. / q: 1 / 32% 2 / 4.6% 3 / 0.37% 3.3 / 0.1% 3.9 / 0.01% 4 / 63ppm 4.4 / 10ppm 4.9 / 1ppm 6 / 2×10^{-9}
 η: "DUTY CYCLE"	PULSE TRAIN	$V_m\sqrt{\eta}$	$V_m\eta$	$\frac{1}{\sqrt{\eta}}$	$\frac{1}{\sqrt{\eta}}$

η	MARK/SPACE	RMS	MAD	$\frac{\text{RMS}}{\text{MAD}}$	CREST FACTOR
1	∞	V_m	V_m	1	1
0.25	0.3333	$0.5V_m$	$0.25V_m$	2	2
0.0625	0.0667	$0.25V_m$	$0.0625V_m$	4	4
0.0156	0.0159	$0.125V_m$	$0.0156V_m$	8	8
0.01	0.0101	$0.1V_m$	$0.01V_m$	10	10

We therefore require circuitry to measure the rms value of ac (or, indeed, dc) waveforms. The earliest circuitry for this purpose used the heating effect of the waveform.

THERMAL RMS-DC CONVERTER

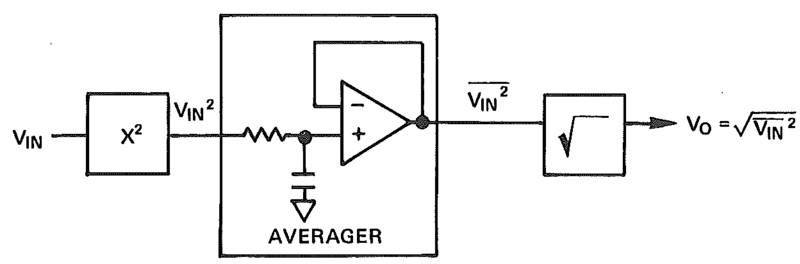


The waveform (probably amplified in a buffer amplifier) is applied to a resistor, R1, which heats a temperature sensor. The output from the sensor is applied to the input to an amplifier whose other input comes from an identical sensor heated by an identical resistor, R2, powered from the amplifier output. If the two sensor/resistor pairs have the same thermal inertia and the same thermal resistance to ambient (and some isolation from each other—although this need not be very good) it is evident that an equilibrium will be reached when the dissipation in R1 is equal to the dissipation in R2. The dc voltage at R2 will then be equal to the rms value of the waveform at R1.

Such an rms computation can be very accurate (errors $< 0.1\%$) and can handle very high frequencies but the dynamic range available is small (auto-ranging can help but only at the expense of system complexity) and therefore the crest factor which can be handled is also relatively small. Also the design of the hardware, with its requirement for identical thermal performance is the two resistor/sensor pairs, is very demanding and the thermal inertia of the system causes long settling times. Quite recently the matched sensors have been fabricated in monolithic form. The accuracy of these monolithic parts is only about 1% and they are still slow to settle but they do have real advantages at RF.

At lower frequencies, however, higher accuracies and wider dynamic range can be achieved at lower cost by a computational circuit. It is quite obvious that the rms value of a signal may be computed with a squarer, an averager (leaky integrator), and a square rooter and that such a circuit could be built with two AD534s and an operational amplifier.

DIRECT COMPUTATION OF THE ROOT-MEAN-SQUARE



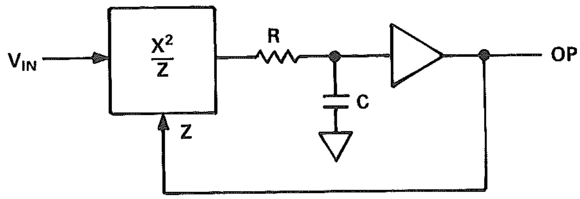
It is not a very good circuit. Because it contains a squarer it has limited crest factor and dynamic range (the output varies by 10,000:1 as the input varies by 100:1) and it also uses three precision components (two multipliers and an op amp). The same function can be realized by a single simple circuit which does what is known as an 'implicit' calculation which is arrived at by manipulating the original expression for rms.

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

$$\therefore V_{rms}^2 = \overline{(V_{IN}^2)}$$

$$\therefore V_{rms} = \frac{\overline{(V_{IN}^2)}}{V_{rms}}$$

We have arrived at our 'inverse function' again—by placing a function generator in a feedback loop we obtain an inverse function. In the present case we require a circuit to perform the X^2/Z function—we apply our input to the X terminal and take the output through an averager and then use it as a feedback signal at the Z terminal as well as the output.

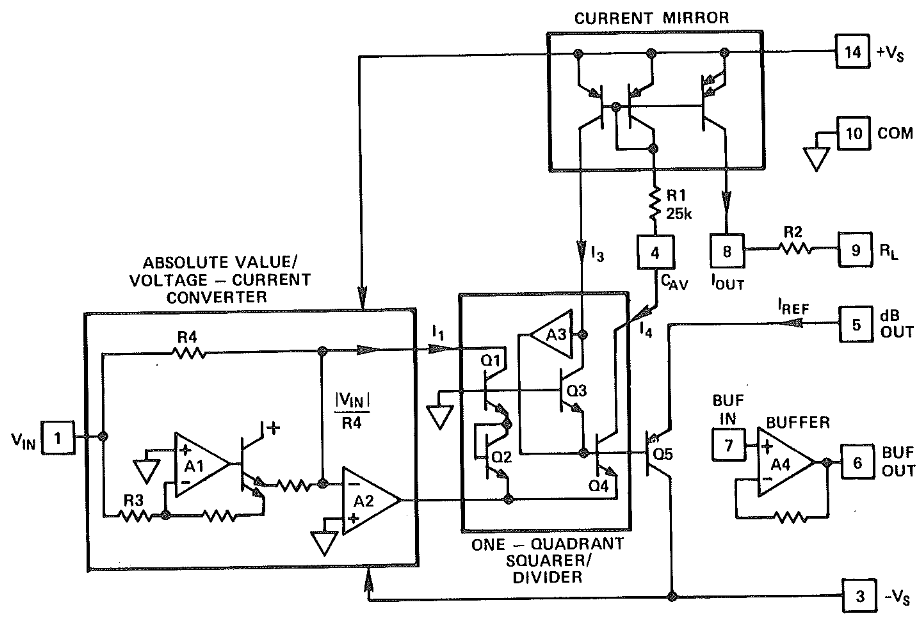


$$V_O = \sqrt{\overline{(V_{IN}^2)}}$$

IF RC>> THE PERIOD OF THE INPUT SIGNAL

While it is possible to construct such an rms circuit from AD534 circuits it is far simpler to design a dedicated rms circuit. The X^2/Z circuit may be current driven and need only be one quadrant if the input first passes through a modulus circuit (a modulus circuit may have either polarity input but although its output has the same amplitude as its input it always has the same polarity, whatever the polarity of the input.) The whole function may be constructed with four transistors, an amplifier, and a current mirror.

RMS TO DC CONVERTER SIMPLIFIED SCHEMATIC



The whole rms/dc circuit is not much more complex. It comprises the circuit mentioned above, a modulus circuit with current output, a buffer amplifier to drive low impedances with a voltage output (the output from the current mirror is a current in a fixed resistor which loses accuracy if loaded), and a single extra PNP transistor to provide a linear dB output (which needs external temperature compensation). Analog Devices* make several such circuits—they all contain laser trimmed resistors and hence provide high accuracy, even at high frequencies, high crest factors and high dynamic ranges, with minimal external trimming—they do, however, require a single external capacitor in the averaging circuit.

*RMS to DC Conversion Application Guide – 2nd Edition", Analog Devices, 1986.

RMS TO DC CONVERTER FEATURES

FEATURES	AD536AJH	AD636JH	AD637JQ	AD736JN	UNITS
Input Range	7	0.200	7	0.200	V_{rms}
Accuracy	±5±0.5	±0.5±1.0	±1±0.5	±0.5±0.6	mV±%rdg.
Bandwidth 1% Accuracy FS Input	100	130	200	50	kHz
Crest Factor $= \frac{V_{peak}}{V_{rms}}$ (±1% Additional Error)	7	6	10	5	–

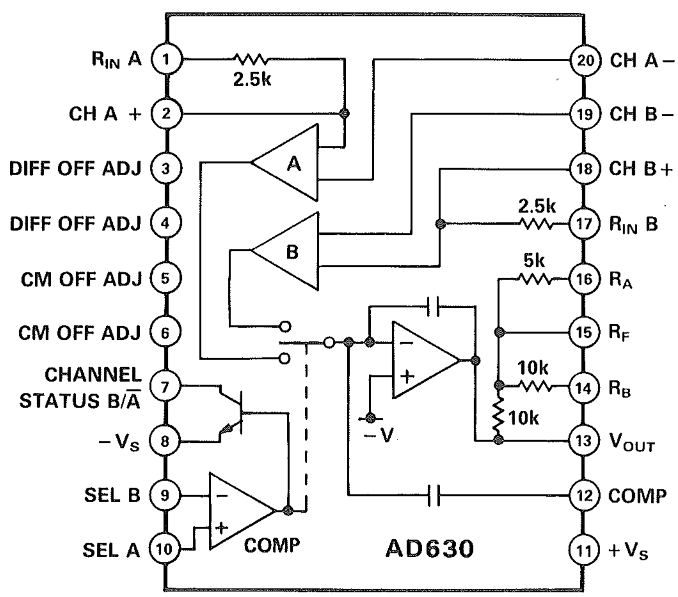
Their operation is quite simple—the two amplifiers in the input stage form an active rectifier which acts as a modulus circuit and drives a unipolar current to the squarer/divider which uses the logarithmic properties of its transistors to perform the function:

$$I_{OUT} = \frac{I_{IN}^2}{I_{FB}}$$

where I_{FB} is the feedback current from the current mirror which is driven by I_{OUT}. If we place a capacitor, C_{AV}, to ground from the I_{FB} port it will form a low-pass filter with R1 and act as an averager. The output of the whole circuit is taken from a second output from the current mirror—this output flows in a resistor to give a voltage output which may be buffered, if required, by the buffer amplifier. It is obvious that this circuit will perform the implicit rms function described above—some applications are described in the next part of this seminar.

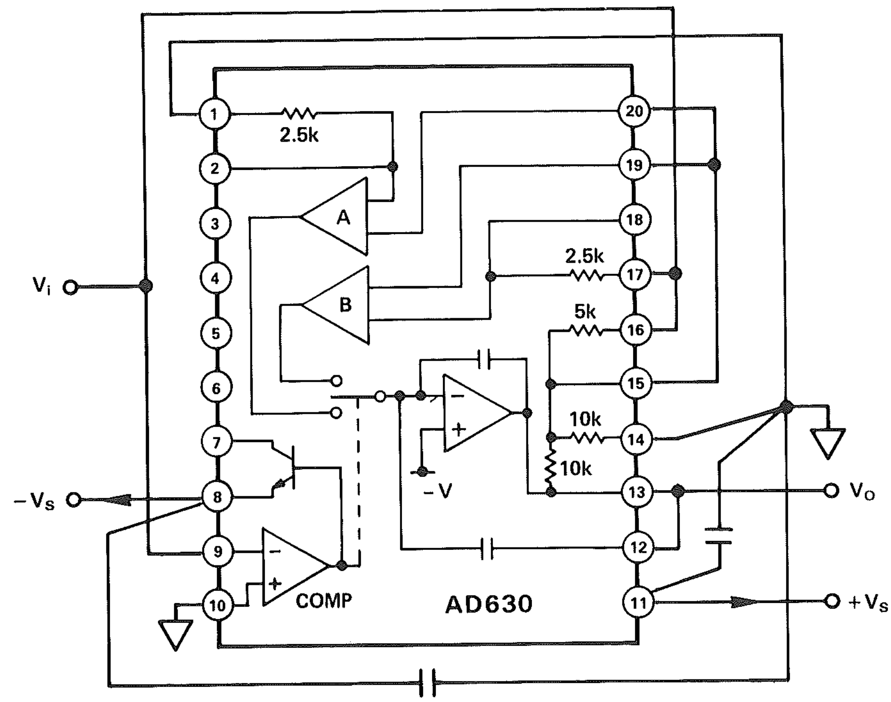
The final member of our collection of analog computational circuits does not use any new or unexpected properties of silicon junctions—it merely reconfigures two operational amplifiers and a comparator to yield a circuit which can be used in many different places. The AD630 contains the input stages of two operational amplifiers, one output stage, a number of laser-trimmed application resistors, and a comparator. The input to the comparator determines which of the two input stages is connected to the output stage.

AD630 FUNCTIONAL BLOCK DIAGRAM

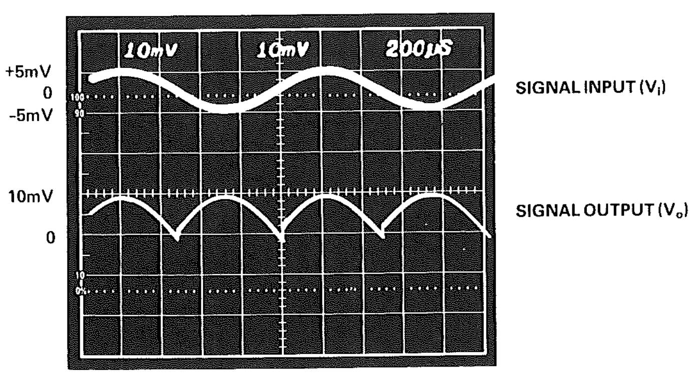


It may, of course, be used as a two-channel multiplexer. If the input stages are configured as inverting and noninverting inputs with the same gain and the same input signal it will act as a modulator if a carrier signal is applied to the comparator (its carrier leak can be as low as − 66dBV at 100kHz and improves at lower frequencies) and as a modulus circuit and a precision low-level rectifier if its input signal is also applied to its comparator.

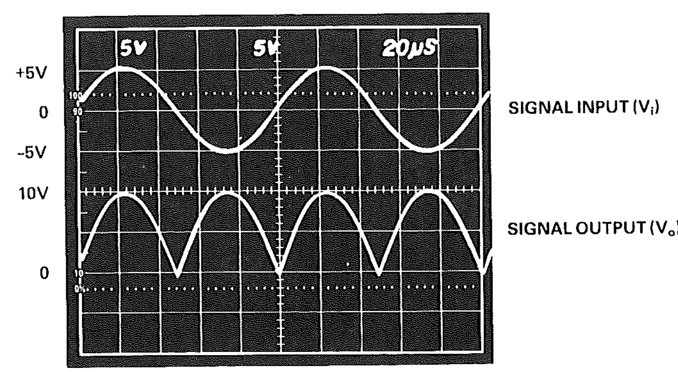
PRECISION RECTIFIER SCHEMATIC DIAGRAM



PRECISION RECTIFIER – LOW LEVEL INPUT



PRECISION RECTIFIER – HIGH LEVEL INPUT



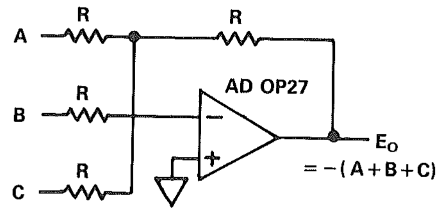
APPLICATIONS OF ANALOG COMPUTATION CIRCUITS

There are innumerable applications of analog computation circuitry. As emphasized in the introduction to this part of our seminar there is no absolute superiority of analog or digital techniques for computation but there are many more places where analog computation is superior than are presently being used.

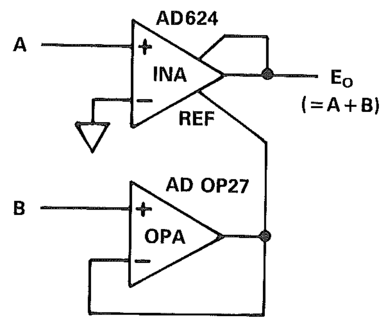
When precision analog computation elements are used (i.e., laser trimmed computation circuitry, op amps with low offset and minimal bias currents, etc.) it will often be found that the theoretical circuit will work almost without modification or trimming—although error budget analysis should always be carried out to ensure that the system performance will remain within specification. We can thus perform the basic algebraic operations of addition, subtraction, multiplication and division on voltages (or currents) with very simple circuitry.

Adders are usually made from an op amp and some resistors but this arrangement is inverting and requires a second inverting amplifier to restore the original polarity. It has, moreover, a relatively low input impedance. Both problems, and the necessity of finding accurate resistors, can be solved by the use of an instrumentation amplifier and an operational amplifier (if the source impedance of the B signal is very low (<1Ω) it may be applied directly to the reference input of the instrumentation amplifier).

ADDERS



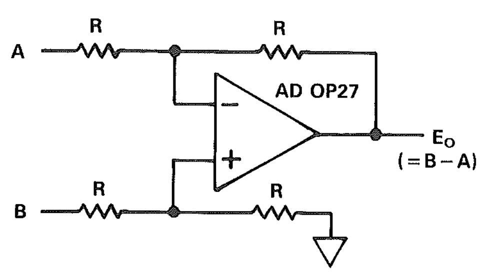
BASIC INVERTING ADDER



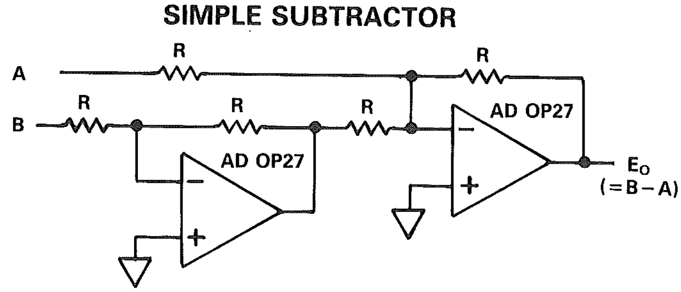
NONINVERTING ADDER USING AN INSTRUMENTATION AMPLIFIER AND AN OPERATIONAL AMPLIFIER

A subtractor is merely an adder with the sign of one of its inputs reversed. It can be made with an operational amplifier and four equal resistors (but this type must be driven from a very low source impedance), from two operational amplifiers and five resistors (with the same disadvantages as an adder of the same general type), or from an instrumentation amplifier.

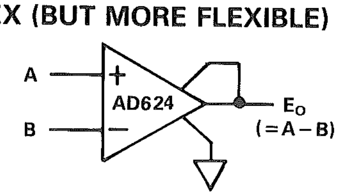
SUBTRACTORS



SIMPLE SUBTRACTOR



MORE COMPLEX (BUT MORE FLEXIBLE) SUBTRACTOR



HIGH IMPEDANCE SUBTRACTOR WITH INSTRUMENTATION AMPLIFIER

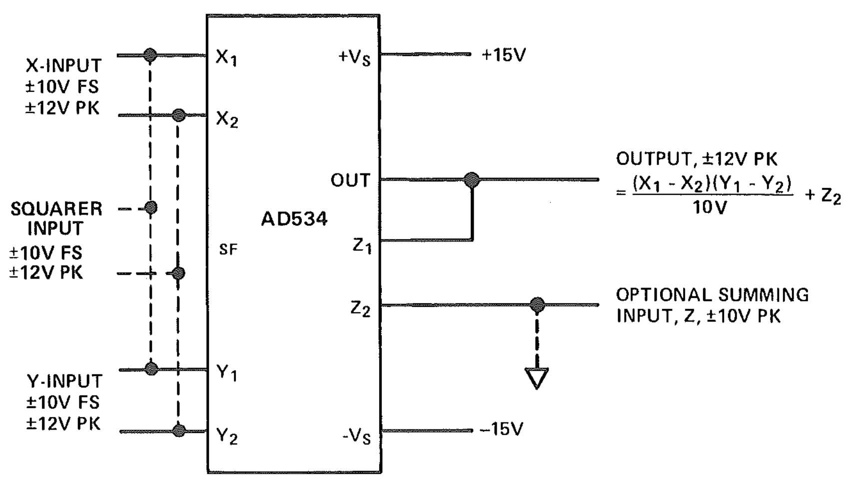
The transconductance multiplier has fully differential inputs. Its transfer function is therefore:

$$E_O = \frac{(X1-X2)(Y1-Y2)}{SF} + Z2$$

$(X1-X2)$ is the differential voltage on the X inputs and $(Y1-Y2)$ is the differential voltage on the Y inputs. SF is the scale factor and is preset to 10V but may be set to any value from 3V to 10V. The scale factor is expressed in volts for dimensional reasons: both inputs to the multiplier are voltages, as is its output—if the equation is to be dimensionally consistant the scale factor must have the unit of volts. It makes no difference to the arithmetic, but the mathematics and physics would be incorrect.

Because the inputs are differential the multiplier will compute the products of differences and will add to the output any voltage applied to the Z2 terminal. It is thus quite a powerful analog computer without any external components whatsoever, and will return its full specified performance of better than 0.25% FS accuracy without external trimming. If the same input is applied to both X and Y ports the basic multiplier functions as a squarer with about half the errors of the XY/10 multiplier.

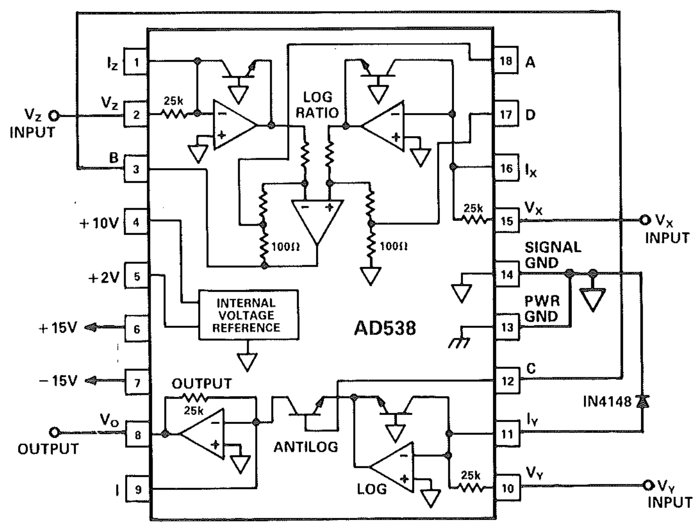
BASIC MULTIPLIER CONNECTION



The AD538 analog computational circuit computes $Y\left(\frac{Z}{X}\right)^M$ using logarithmic techniques. This implies that if X and M are set to unity it functions as a simple multiplier. In many applications it is more accurate than a transconductance multiplier since (due to its logarithmic computation method) its output error is a percentage of reading (plus an offset) rather than a percentage of full scale but is inputs are not differential and will only accept positive signals (although it does have both current and voltage input terminals) so it is unsuitable for

ONE-QUADRANT COMBINATION MULTIPLIER/DIVIDER

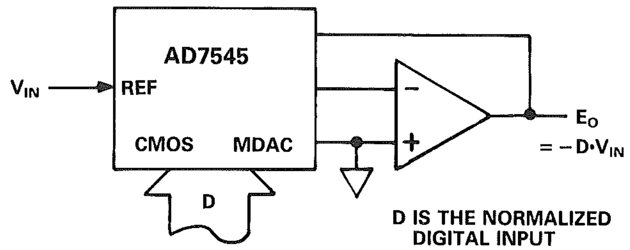
$$V_{OUT} = V_Y\left(\frac{V_Z}{V_X}\right)$$



bipolar applications. It also has a lower bandwidth than the fastest transconductance circuits. Its application circuit is almost embarassingly simple: there are two inputs, an output, and three power pins to connect and two pairs of pins must be short-circuited and one pin must be grounded via a diode.

While not strictly a pure analog technique we should not overlook the multiplying digital/analog converter (MDAC). These circuits multiply an analog input by a digital input and have an analog output, and are very convenient for many computations involving mixed analog and digital variables. A full description of their uses in computation, attenuators, programmable oscillators and filters, and many other applications is given in the data conversion section of this seminar—here we shall merely point out that it is possible to multiply bipolar analog voltages of up to ±30V by unipolar or bipolar digital numbers with resolutions between 8 and 14 bits by using CMOS MDACs in the AD752X, AD753X and AD754X ranges.

MDAC PERFORMS MIXED ANALOG/DIGITAL MULTIPLICATION



If we connect the output of a transconductance multiplier, such as the AD534, to its Y2 input (with Y1 grounded) and apply external inputs to X and Z (both, being fully differential, may be used with differential or single-ended signals) we find that, because of the negative feedback in the system,

$$\frac{X \cdot (Y1 - Y2)}{10V} = Z$$

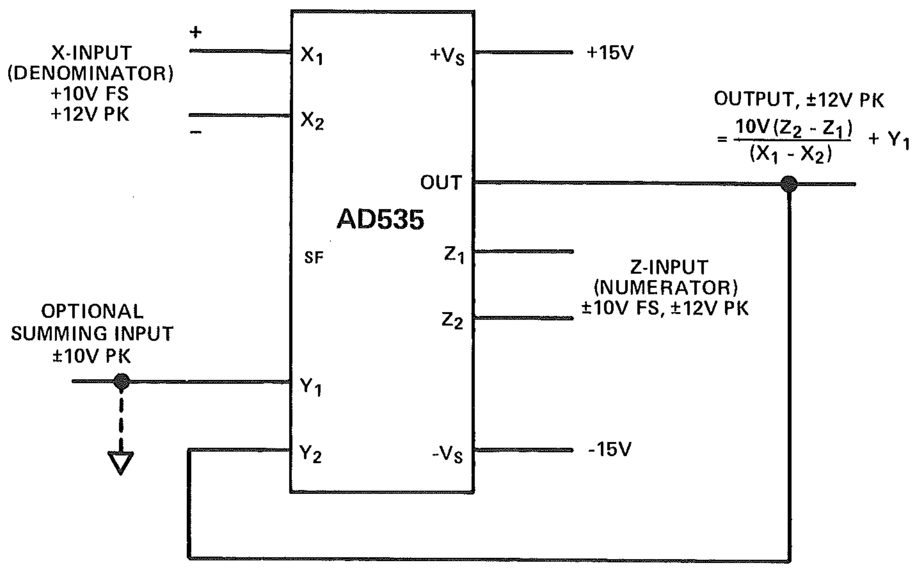
but $Y2 = E_O$ and $Y1 = 0$

$$\frac{X \cdot E_O}{10V} = Z$$

and $E_O = -10V \cdot \frac{Z}{X}$ so that the circuit is a divider.

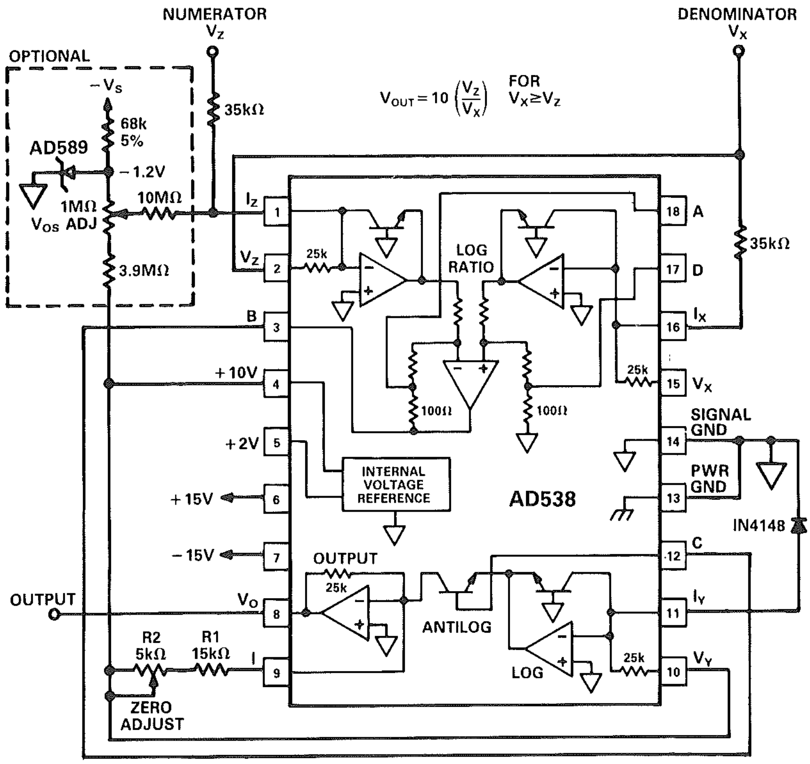
If Y1 is not grounded but has a voltage applied to it that voltage will be added to the output voltage.

BASIC DIVIDER CONNECTION



The AD538 can, moreover, be used as a two-quadrant divider, even though its inputs accept only unipolar signals. This is done by using the current X and Z input ports and external 35K resistors to enter the numerator and denominator and adding the denominator to the numerator via the numerator voltage input port.

TWO-QUADRANT DIVISION WITH 10V SCALING



This changes the transfer function from:

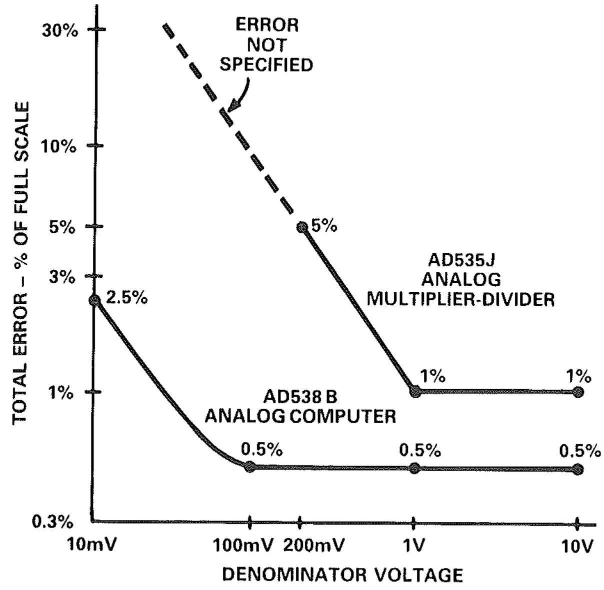
$$V_O = 10V\left(\frac{Z}{X}\right)$$

to $V_O = 10V\,\frac{(Z+BX)}{X} = 10V\left(B+\frac{Z}{X}\right) = 10V{\cdot}B + 10V\left(\frac{Z}{X}\right)$

Where $B = \frac{35K}{25K}$ (the ratio of the input resistors).

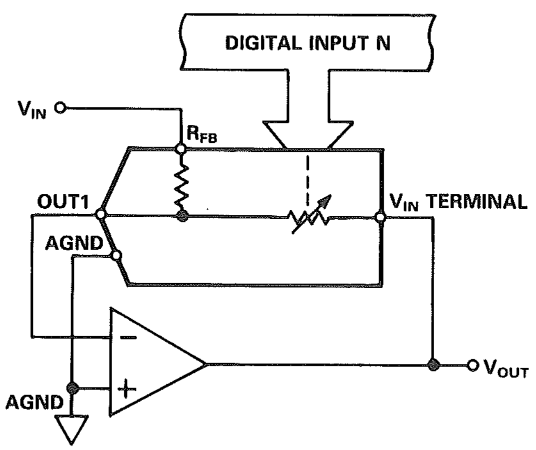
The additional 10V•B is subtracted in the output amplifier using the AD538 reference and an external resistor, R1 (trimmed by a potentiometer, R2, since the absolute values of the internal resistors of the AD538 are not very accurate, even though their matching is very accurate indeed—a penalty of the laser trim technique). The consequence of this operation is that, provided the magnitude of the denominator is greater than the magnitude of the numerator, the numerator may have either polarity and therefore the circuit is a two-quadrant divider.

COMPARISON OF AD538 AND AD535 AS DIVIDERS



A final technique for division uses a CMOS MDAC as the multiplier in a feedback loop—again it is not a true analog divider but should not be overlooked in systems where both analog and digital signals are present.

MDAC IN FEEDBACK LOOP MAKES A DIVIDER



EQUATIONS

$$V_{OUT} = -\frac{V_{IN}}{N}$$

$$A_V = \frac{V_{OUT}}{V_{IN}} = -\frac{1}{N}$$ where: A_V = Voltage Gain

and where:

$$N = \frac{BIT\ 1}{2^1} + \frac{BIT\ 2}{2^2} + \frac{BIT\ 8}{2^8}$$

EXAMPLES

$N = 00000000$, $A_V = -A_{OL}$ (OP AMP)
$N = 00000001$, $A_V = -256$
$N = 10000000$, $A_V = -\frac{256}{128} = -2$
$N = 11111111$, $A_V = -\frac{256}{255}$

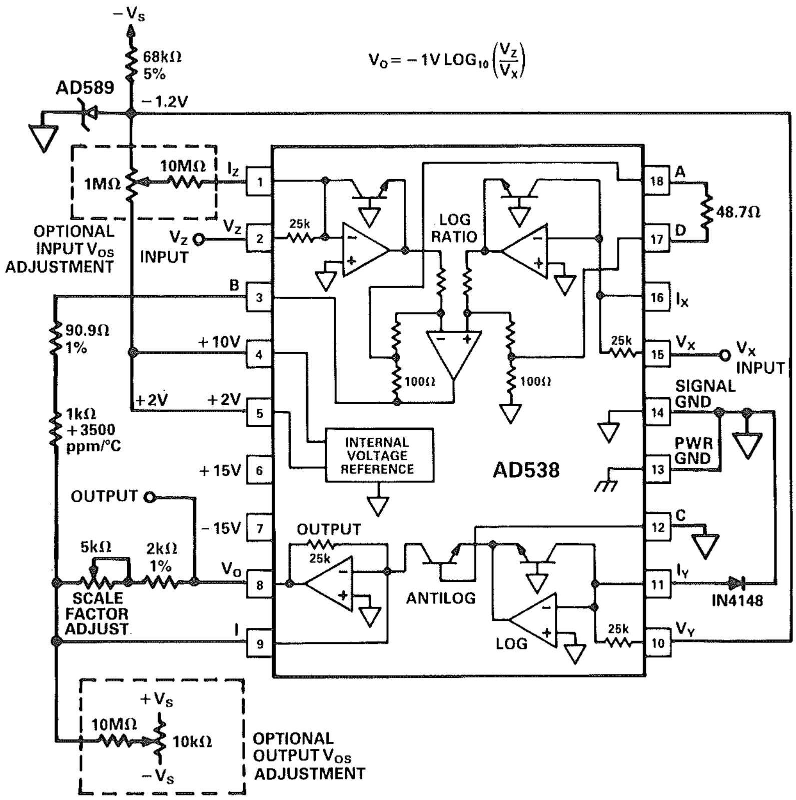
Everyone knows that division by zero yields infinity. When mathematics is performed with a pencil and paper this is acceptable—when it is performed with electronics and the output is a voltage we might expect division by zero to yield sparks and a smell of ozone. In fact division by too small a denominator will result in the divider output limiting near to one or other of the supplies, and the use of larger, but still small, denominators will tend to reduce accuracy and system bandwidth. It is important to calculate the expected accuracy at the extremes of the input ranges from the device data and to choose a device which will give the performance required. It is also important to choose a realistic performance, consistant with the total system, rather than demand the highest possible performance because it is within the state-of-the-art.

Many more complex mathematical operations are barely more difficult to perform than the basic $+$, $-$, $\times$, and $\div$. Logarithmic circuits will easily compute logarithms and powers and the AD538 will also compute arc tangents more quickly than most digital techniques and more accurately than any other analog ones.

The most serious drawback of the diode logarithm circuit, as we discussed earlier, is that the output contains a kT/q term and is temperature dependent. This temperature variation may be compensated by the use of a +3500ppm/°C temperature variable resistor*. The log ratio circuit shown has better than 0.5% accuracy for over 60dB dynamic range of inputs (10mV to 10V).

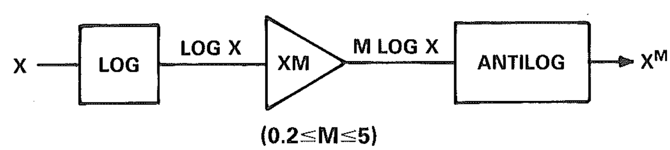
*Available from Tel Labs Inc., 154G, Harvey Road, Londonderry, NH 03053. Tel: (603) 625-8994 Twx: (710) 220-1844.

LOG RATIO CIRCUIT

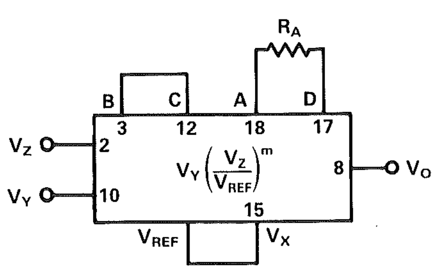


In a logarithmic circuit like the AD538, which contains log and antilog circuits and an amplifier, all that needs to be done to raise an input X to the Mth power (0.2<M<5) is to set the amplifier gain to M, which can be done with one or two resistors, depending on whether M>1 or M<1 (if M = 1 no external resistors are necessary but who wants a unity power circuit?)

HOW AN AD538 COMPUTES POWERS

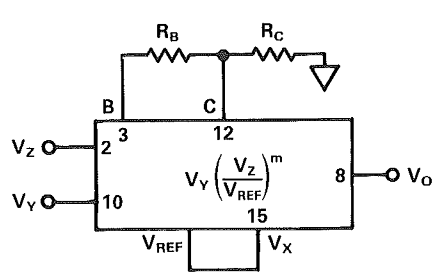


HOW TO SET VALUES OF M



POWERS

m	R_A
2	196Ω
3	97.6Ω
4	64.9Ω
5	48.7Ω



ROOTS

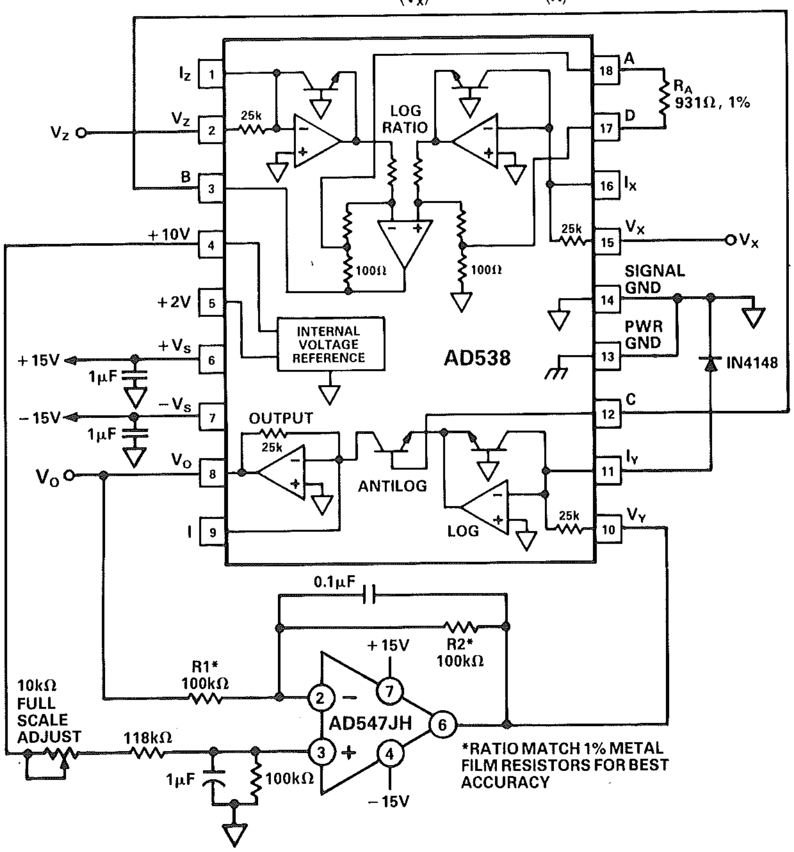
m	R_B	R_C
1/2	100Ω	100Ω
1/3	100Ω	49.9Ω
1/4	150Ω	49.9Ω
1/5	162Ω	40.2Ω

The arc-tangent circuit shown is typical of AD538 applications where Y is 1 so that $V_O = \left(\frac{Z}{X}\right)^M$ for M<1.

In an approximation to the arc-tangent function the AD538 may be made to compute the angle represented by two rectangular coordinates which, since they are applied to the X and Z inputs, we shall call X and Z rather than the more usual X and Y. If X and Z are within the range 100μV and 10V the error in the computed angle is under 1° (the AD639 can do a similar computation with fewer components but cannot work over as wide a dynamic range).

THE ARC-TANGENT FUNCTION

$V_\theta = [V_\theta\mathrm{REF} - V_\theta] \times \left(\frac{V_Z}{V_X}\right)^{1.21}$ $\theta = \mathrm{TAN}^{-1}\left(\frac{Z}{X}\right)$



The circuit exploits the fact that

$$T = \frac{(\tan T)^{1.21}}{1 + (\tan T)^{1.21}}$$

Where T is the angle normalized to 90°.

The AD538 and the external amplifier calculates log(tan T) from X and Z, amplifies it by a factor of 1.21 to raise to the 1.21 power and performs an implicit calculation to calculate the angle (which is expressed in terms of the reference voltage). Under these conditions the output voltage tends to V_{REF} as the angle tends to 90° although, in fact, the circuit cannot be used much above 89.5° because at 90° tangents become infinite and before then the circuit becomes unstable. R1 and R2 must be matched for highest accuracy, and the circuit is stablized by the 0.1μF integrating capacitor in the amplifier feedback path. The circuit works in a single quadrant since both X and Z must be positive.

Trigonometrical functions are more normally calculated by the AD639. No external components (other than supply decoupling capacitors) are required to compute sines, cosines, tangents and cotangents, and secants and cosecants with the AD639. Little more than an extra reference voltage (which may be generated from the internal reference with an operational amplifier and a couple of resistors) are required for versines and coversines. For details of the less common functions the reader should consult the AD639 datasheet and various application notes—the sine, cosine, and tangent will be described here.

THE AD639 WILL COMPUTE

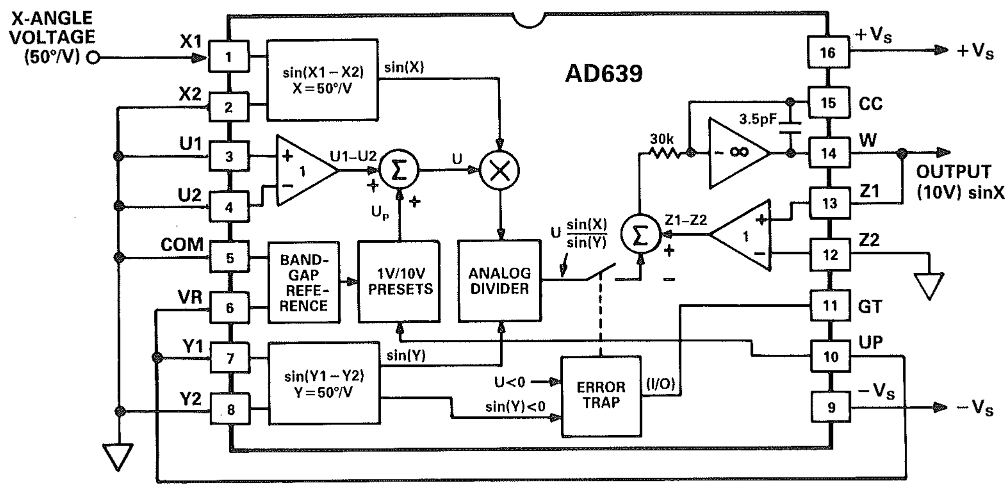
- Sine
- Cosine
- Tangent
- Cotangent
- Secant
- Cosecant
- Versine
- Coversine
- Exsecant
- Etc . . .

AND THEIR INVERSE FUNCTIONS

MOSTLY WITHOUT EXTERNAL COMPONENTS

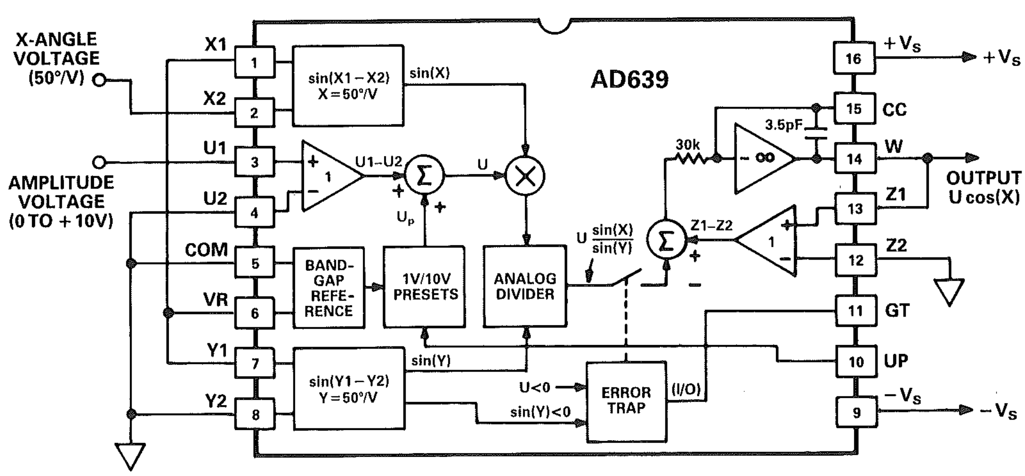
The sine circuit functions from −500° to +500° with best accuracy between −90° and +90°. Different scale factors may be selected by different connections of U1, U2 and UP.

CONNECTIONS FOR THE SINE MODE WITH AMPLITUDE PRESET TO 10V



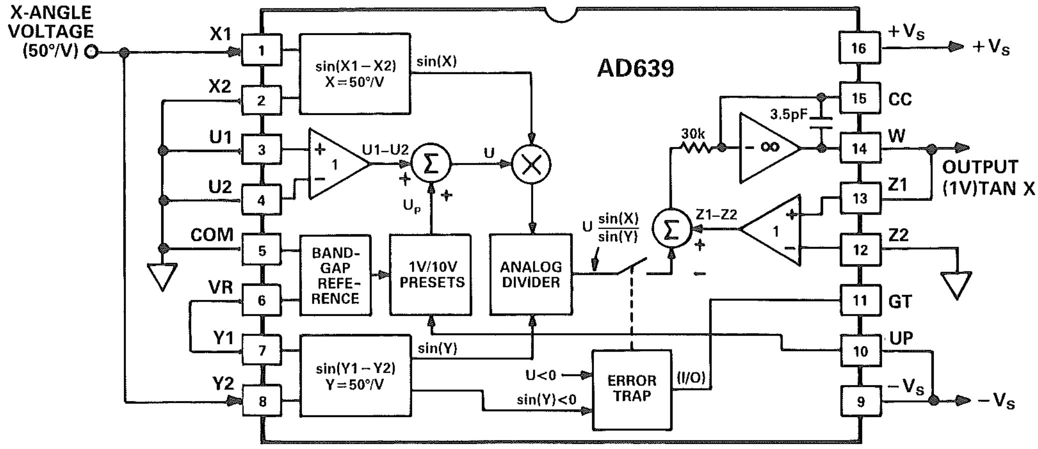
The cosine circuit differs from the sine only in that the X1 input is offset by the internal reference (preset to 1.8V (=90°)) and the input is applied to the inverting X input −X2. Its angular range reflects the offset, being −400° to +600°, and its best accuracy is between 0 and 180°.

CONNECTIONS FOR THE COSINE MODE WITH EXTERNAL AMPLITUDE CONTROL



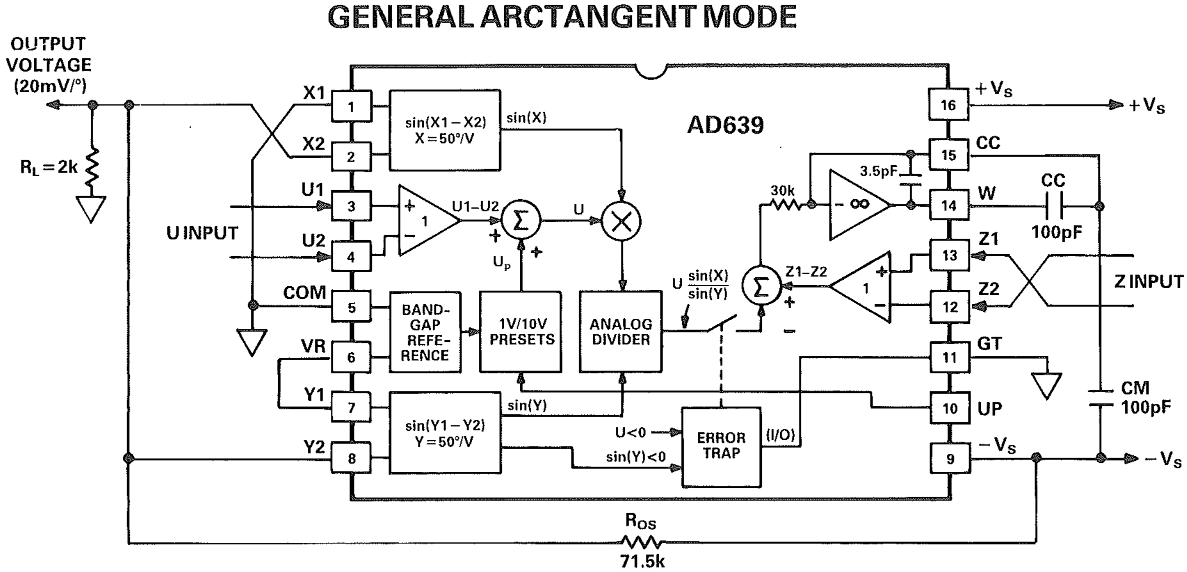
The computation of the tangent requires sine input to the X input and cosine (i.e., negative sine offset 90°) to the Y inputs. For inputs above ±85° the outputs, even if scaled to 1V, will go out of the operating range of the amplifier with a ±15V supply—instead the internal error trap operates and the output returns to zero while a warning flag is set on pin 11.

CONNECTIONS FOR TANGENT MODE WITH AMPLITUDE PRESET TO 1V



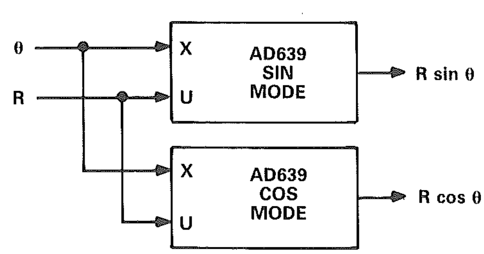
With the AD639 inverse trigonometrical functions are almost as easily computed as direct ones, although normally only the principal angular range is available (−90° to +90° for arc sine, 0° to +180° for arc cosine, etc.). Although the AD639 contains an operational amplifier so that the inverse mechanism we keep explaining—a function generator in a negative feedback loop generates an inverse function—may be used yet again, some extra circuitry, mostly resistors, is necessary because the AD639 has low resistance inputs (only 3.6K to common) and therefore loads any circuitry driving it. The basic arc-tangent circuit is shown in the diagram and its operation, and that of the arc-sine and arc-cosine circuits, is described in the data sheet.

CONNECTIONS FOR THE GENERAL ARCTANGENT MODE



The AD639 tangent and act-tangent circuits may be used, with other circuitry such as the AD630 and the AD534, to perform coordinate conversion from Cartesian (X, Y) to polar (R, τ) or vice versa.

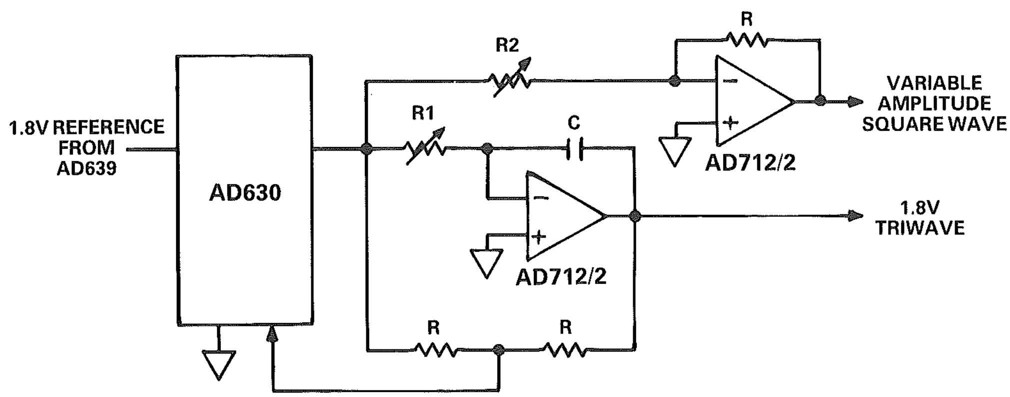
POLAR-CARTESIAN COORDINATE CONVERTER



When the AD639 is described on its data sheet as a function generator the descripton is accurate, but it can be misleading since the term 'function generator' is frequently applied to a piece to testgear—an oscillator having sine, triangular, and squarewave outputs. The AD639 is not an oscillator.

Nevertheless the AD639 may be used to convert a triangular waveform into a sinewave and, if a squarewave is available in quadrature to the tri-wave, the phase of the output sinewave may be varied continuously merely by varying the squarewave amplitude. The tri-wave required for this application has an amplitude of ±1.8V and may be generated very easily with an AD630 and an op amp, using the 1.8V reference from the AD639 itself. The frequency of oscillation is set by R1 and C—if digital control of frequency is required R1 might be replaced with a CMOS MDAC—and the amplitude of the squarewave (which controls the phase of the AD639 output) is set by R2—which again could be replaced by a CMOS MDAC for digital control of phase.

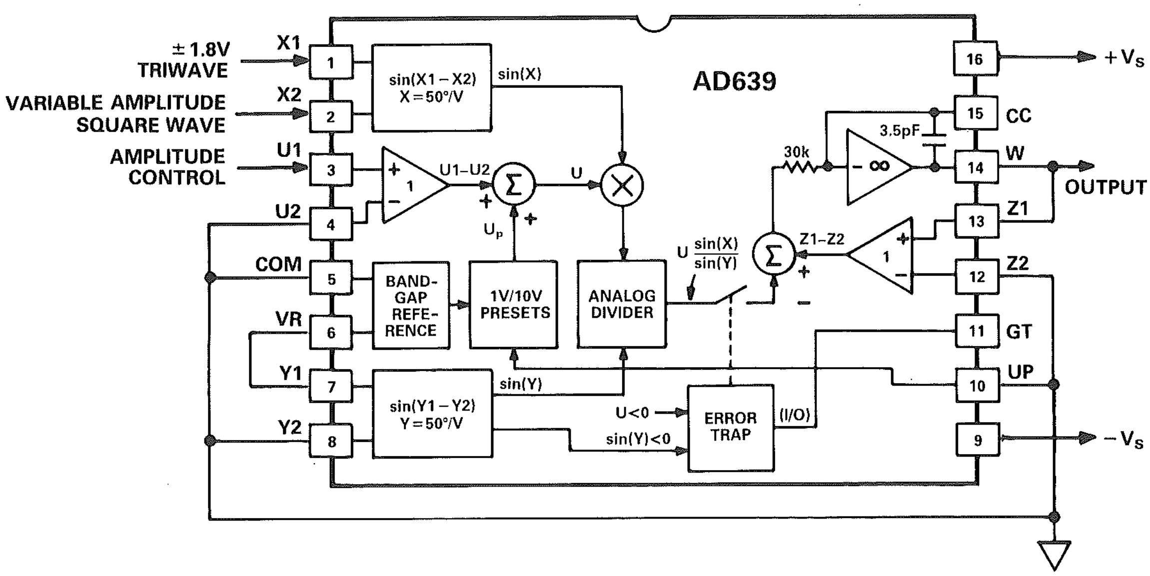
AD630 TRIWAVE GENERATOR



(R1 AND R2 MAY BE REPLACED WITH MDACS)

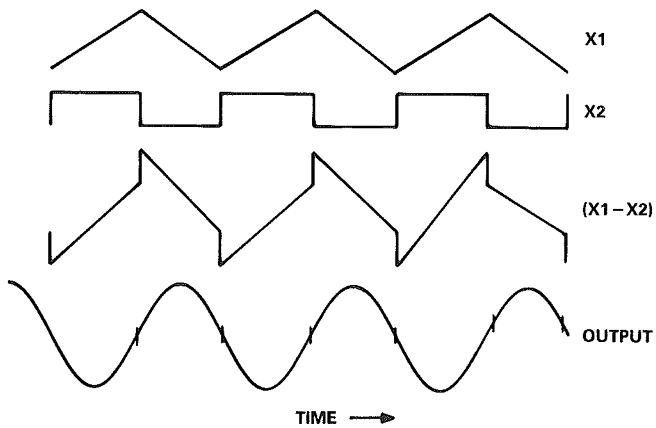
The triwave is applied to the X1 terminal of the AD639 and the squarewave to the X2 terminal. The Y port is connected to the reference (Y1) and ground (Y2) to bias the internal divider, but is otherwise unused. The output amplitude may be preset with the UP terminal or may be varied by applying a control voltage to U1 (again if digital control of amplitude is required this terminal might be driven by a DAC).

VARIABLE-PHASE SINEWAVE GENERATOR

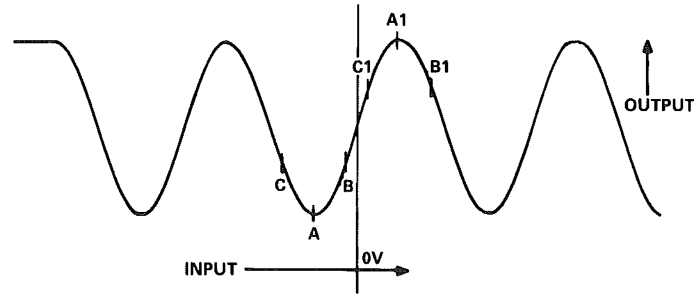


The operation of the circuit may be appreciated by first considering the case where the squarewave input has zero amplitude (i.e., X2 is grounded). As the triwave goes from −1.8V to +1.8V the output goes from A to A1 on the transfer characteristic (on the bottom line of the diagram) and then, as it goes from +1.8V to −1.8V, the output goes back from A1 and A. Thanks to the sinusoidal transfer function of the AD639 this linear cycling of the input produces a sinewave at the output.

WAVEFORMS IN CIRCUITRY



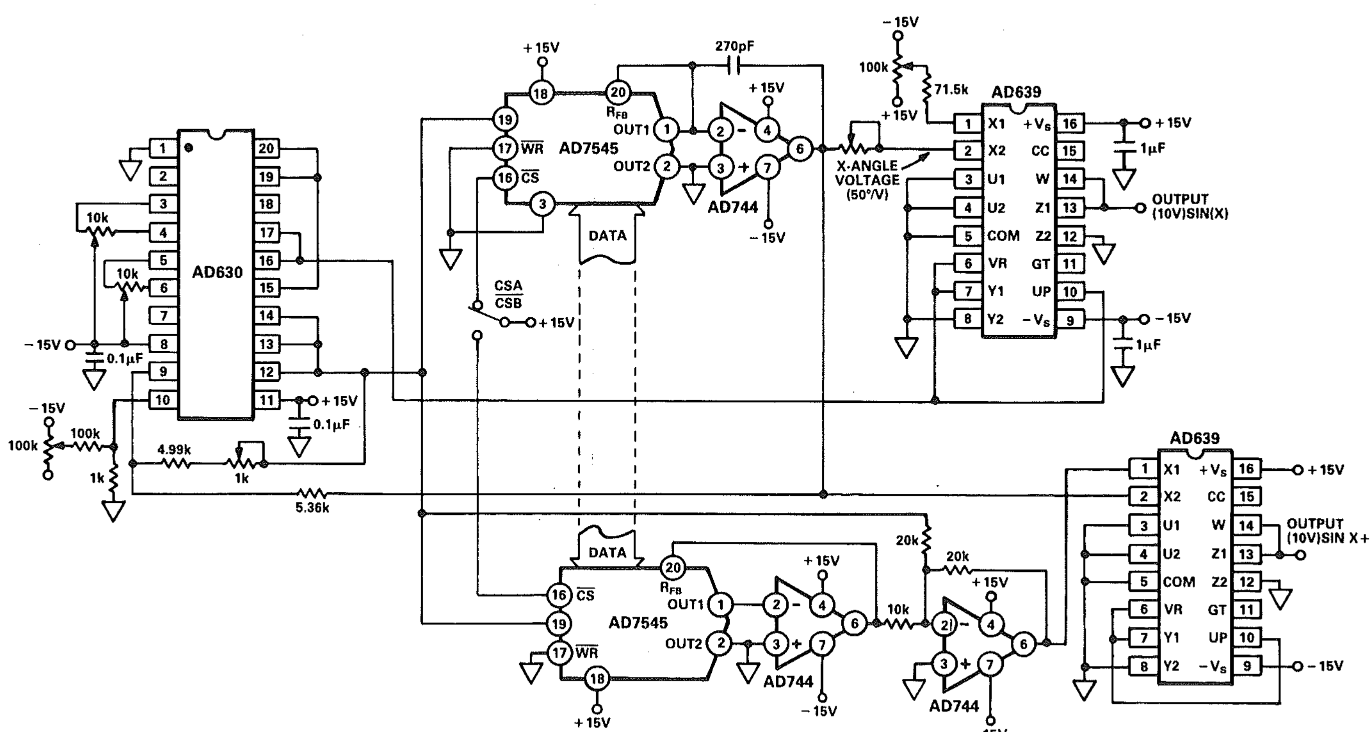
AD639 TRANSFER CHARACTERISTIC



If we now restore the squarewave to the X2 input, the output cycles from B to B1 and then, as the squarewave switches, jumps to C1 and cycles to C on the negative-going half of the triwave. At C the squarewave switches again and the output cycles back from B to B1 again. The effect at the output is a sinewave formed by sections B1 to B and then C to C1 of the AD639 transfer characteristic. Since there are jumps from B to C and from C1 to B1 when the squarewave switches there are small glitches on the output waveform where the jump occurs but they are generally insignificant because if the squarewave is symmetrical (equal positive and negative amplitudes) the values of B and C and or C1 and B1 are identical so the glitch is only the spike which gets past the output amplifier filter, not an overall change of level.

With the waveforms shown in the diagram increasing the squarewave amplitude increases the phase delay (lag) in the output waveform—if the phase of the squarewave is reversed the sinewave will, of course, lead the triwave, the squarewave in either case is 90° out of phase with the triwave—in the diagram it leads it.

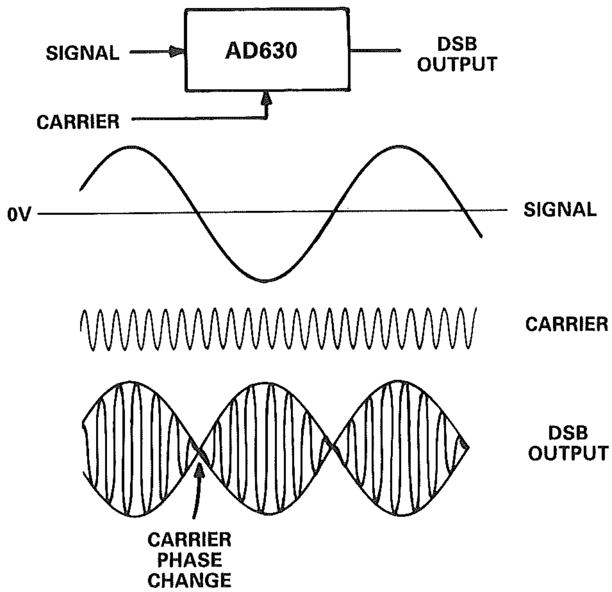
PROGRAMMABLE FREQUENCY AND PHASE SINEWAVE GENERATOR



The AD630 will do a lot more than merely generate triwaves to drive the AD639. As mentioned above it will function as a modulus circuit giving a unipolar output with the same magnitude as its input (and a logic output indicating the polarity of the input), it will act as a two-channel multiplexer/buffer, and it will act as a high-performance modulator/demodulator.

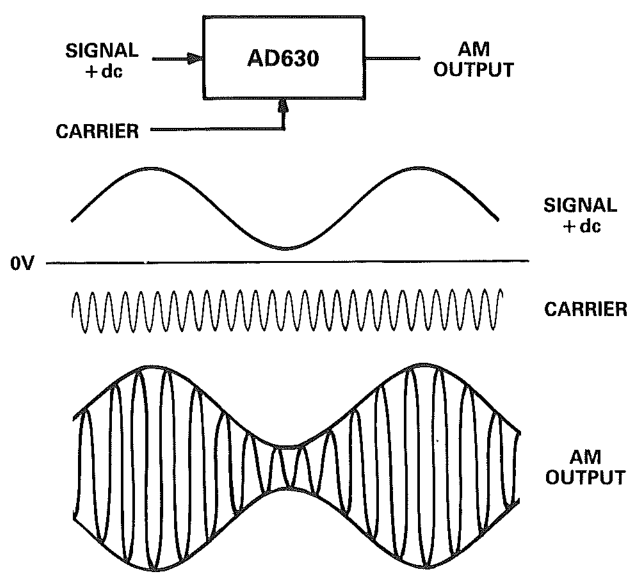
Modulator/demodulator circuits are quite common, and are available working at much higher frequencies than the AD630. Its contribution to the technology is its laser-trimmed precision—it has very low offset (and hence low carrier leak) and accurate gain. It is unsuitable for higher speed applications, or ones where cost is important and precision is not (the industry standard 1496 and 1596 were intended for such applications) but wherever precision is needed the AD630 has real advantages. As a modulator it has low carrier leak and high accuracy. If a signal is applied to the AD630 signal port and about 50mV rms of carrier to the comparator the output will be a double sideband signal (DSB or 'suppressed-carrier' AM).

BALANCED MODULATOR PRODUCES DSB SIGNAL



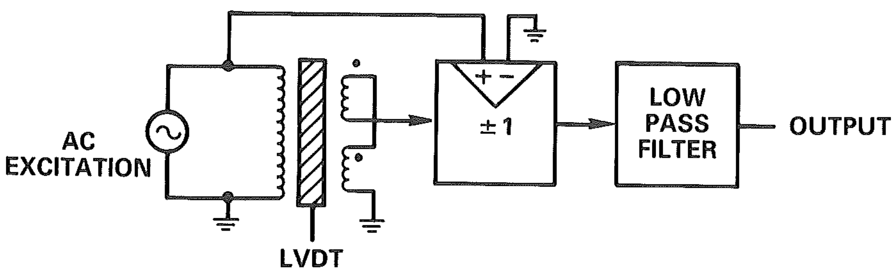
Many people expect amplitude modulation from such a circuit but in fact to obtain AM either the signal input must have a dc offset or carrier leak must occur (possibly by adding carrier to the signal input).

CARRIER LEAK CAUSES DSB TO BECOME AM



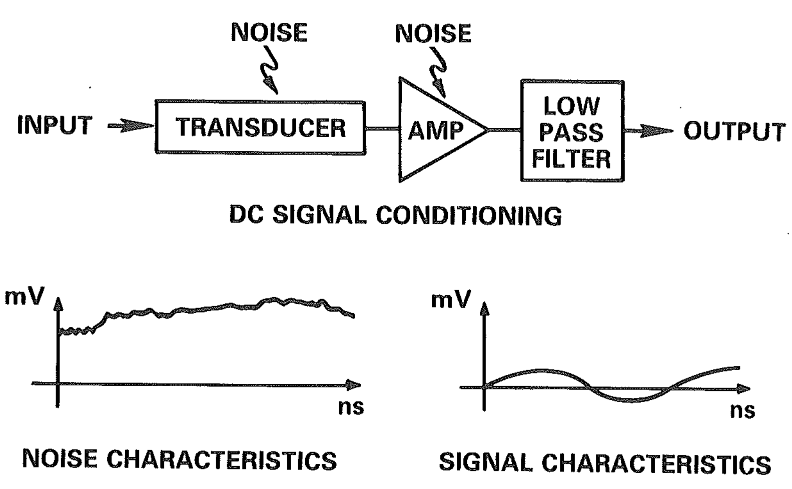
A modulator/demodulator will also function as a phase-sensitive rectifier. An application where a very accurate phase-sensitive rectifier is required is the demodulation of LVDT signals. A linear variable displacement transducer (LVDT) is a device for sensing position which consists of three coils, a primary and two secondaries connected in series antiphase. The secondaries are positioned on either side of the primary and the amount of flux in each depends on the position of a moving core—when the core is central there is equal flux in each secondary and the total output is zero. If the core is displaced one secondary will be linked by more flux than the other and will have greater output—the output amplitude and phase of the whole thing depends on the position of the core and an AD630 is ideal for converting this to a bipolar dc position signal.

LVDT SIGNAL DEMODULATOR



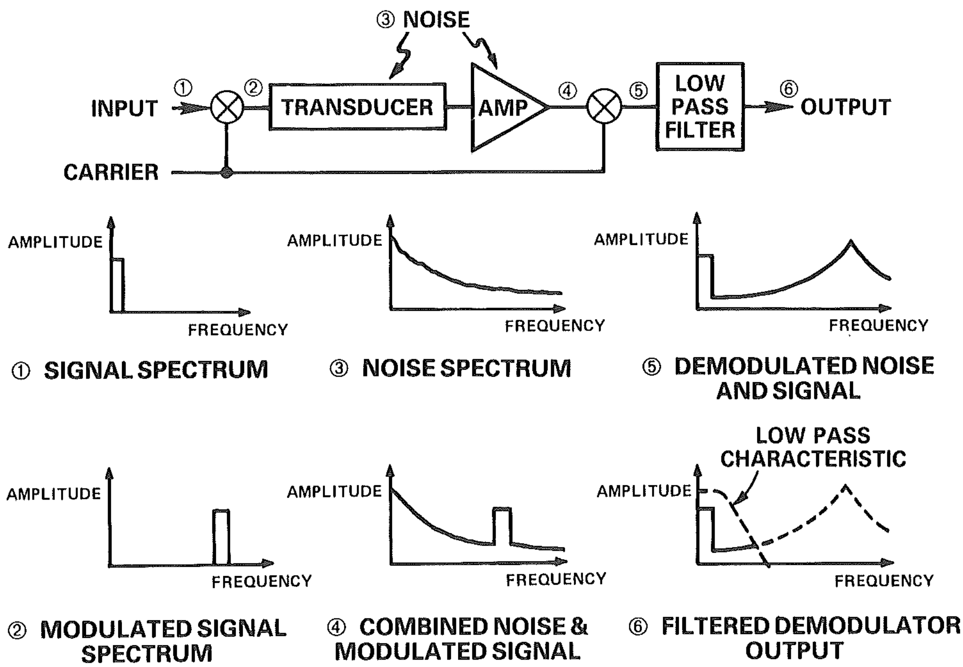
Whole libraries have been written on applications of modulators and demodulators which we do not have space to review here, but there is one more application where the exceptional accuracy of the AD630 is particularly useful—synchronous demodulation. Many transducers have relatively small outputs and are often in locations where they are vulnerable to electrical noise, and dc amplifiers are liable to drift and to cause low frequency noise on their own account. In such circumstances noise may swamp a wanted signal.

SMALL DC SIGNALS ARE EASILY LOST IN NOISE



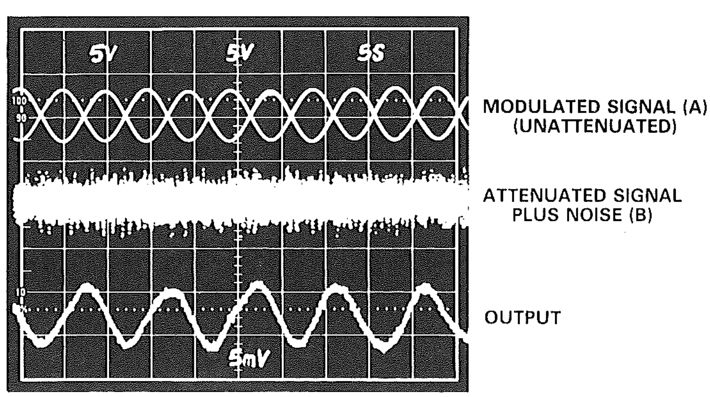
If we can use the signal to modulate a carrier—perhaps by energizing a transducer with ac rather than dc, perhaps by chopping the dc transducer output, or perhaps even by chopping the physical input to the transducer. We can then amplify the resultant ac signal, filter dc and LF noise, and, by synchronous demodulation, recover the original signal. It is more helpful to consider the operation in the frequency domain than the time domain.

AC SIGNAL CONDITIONING FREQUENCY DOMAIN ANALYSIS



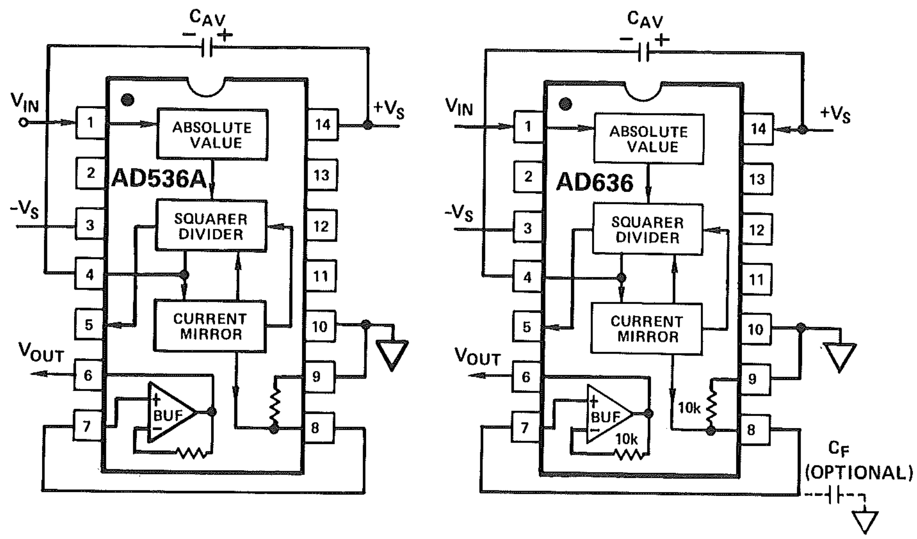
Finally the oscilloscope photograph shows how this technique may be used to recover information modulated on a carrier and buried in white noise 100dB stronger than itself.

LOCK-IN AMPLIFIER WAVE FORMS

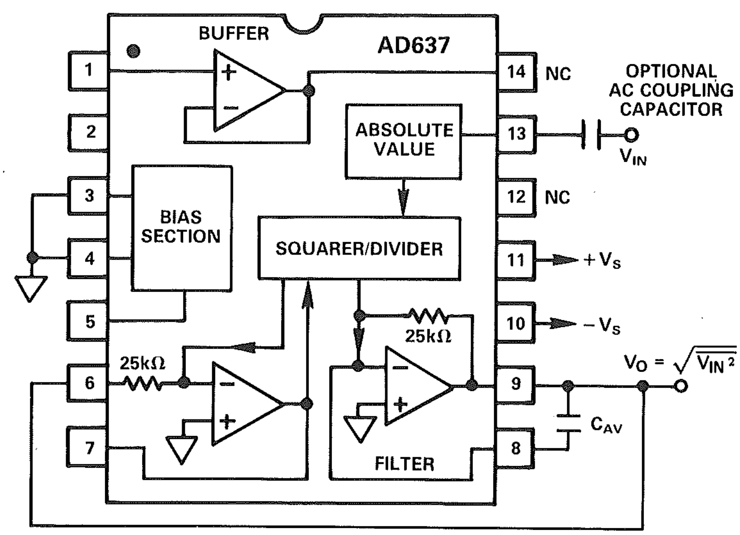


The final computational circuit whose applications we shall consider is the rms/dc converter*. Obviously its main use is the computation of the rms value of an input signal which is generally ac but may be modulated dc of some sort. Although offset trims will improve performance slightly the only external component required is the integrating capacitor.

AD536A/AD636 STANDARD RMS CONNECTION

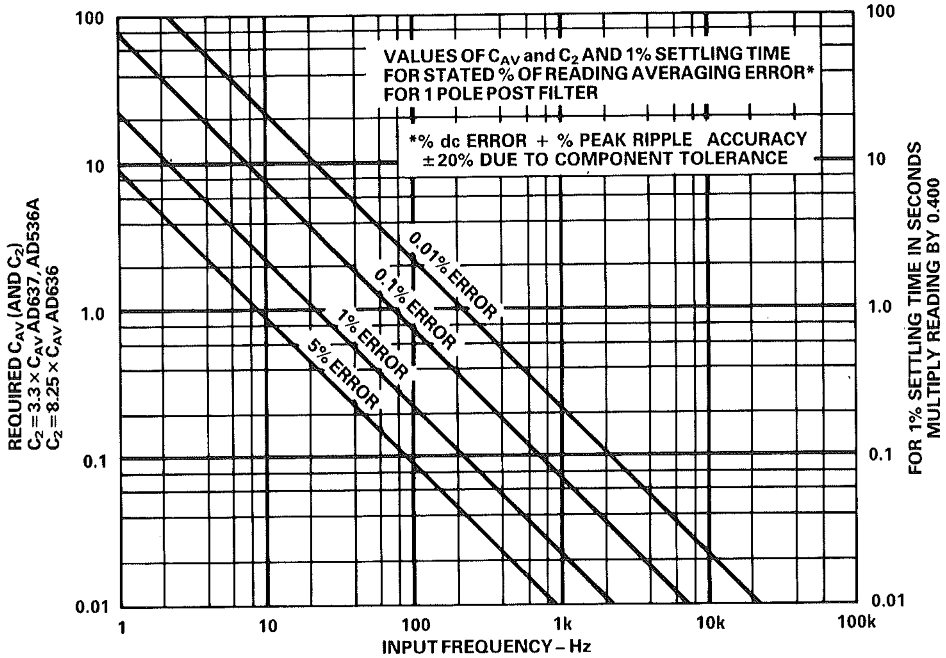


AD637 STANDARD RMS CONNECTION



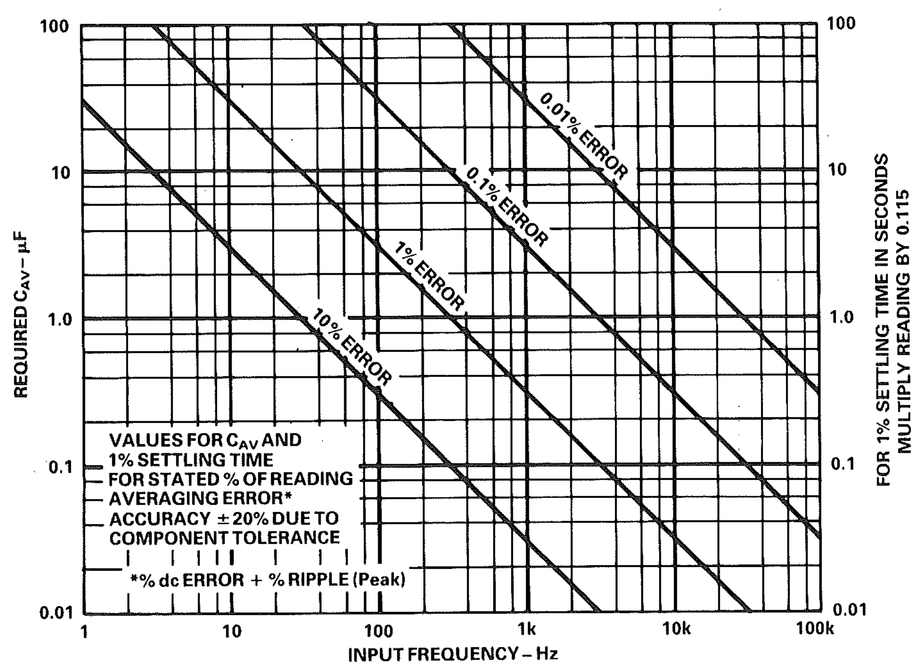
*For detailed discussion see "RMS to DC Conversion Application Guide – 2nd Edition" – Analog Devices, 1986.

ERROR/SETTLING TIME GRAPH FOR USE WITH 1-POLE OUTPUT FILTER CONNECTION

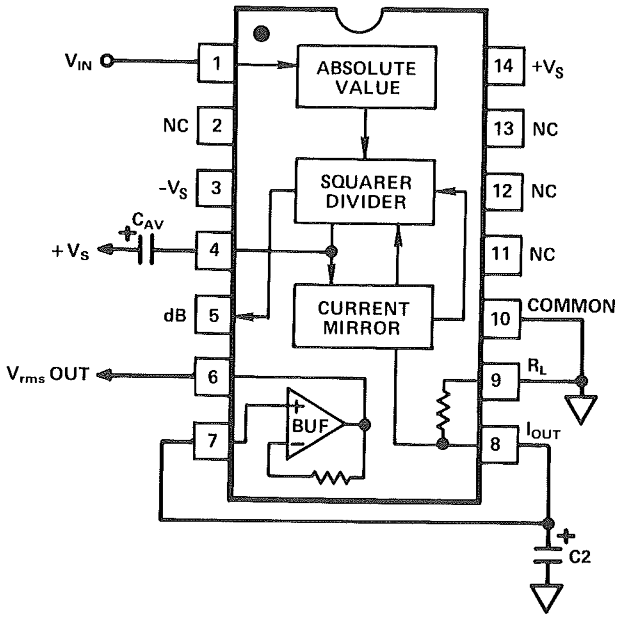


An extra stage of output filter will improve the ripple and the settling time (note that the settling time is twice as long after step reductions of input than after step increases—this is due to the averaging capacitor charging from a current source but discharging through a resistor).

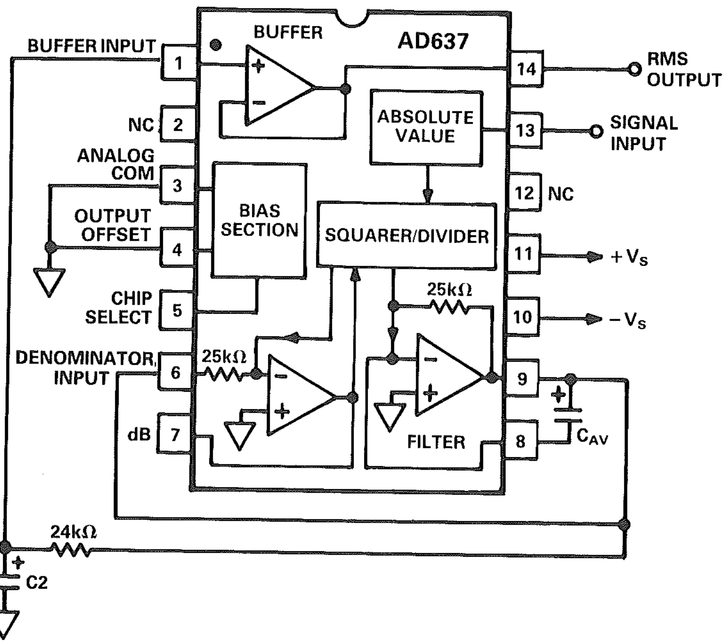
ERROR/SETTLING TIME GRAPH FOR USE WITH THE STANDARD RMS CONNECTION



AD536A/AD636 WITH A 1-POLE OUTPUT FILTER

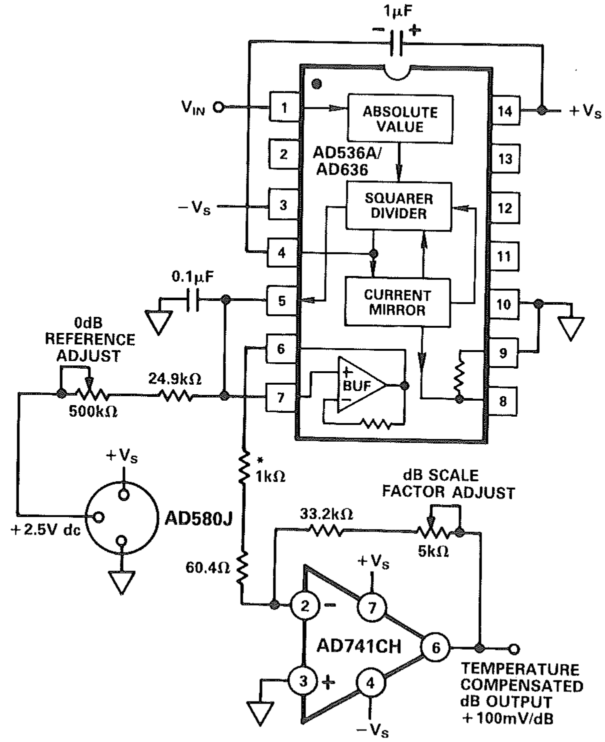


AD637 WITH A 1-POLE OUTPUT FILTER



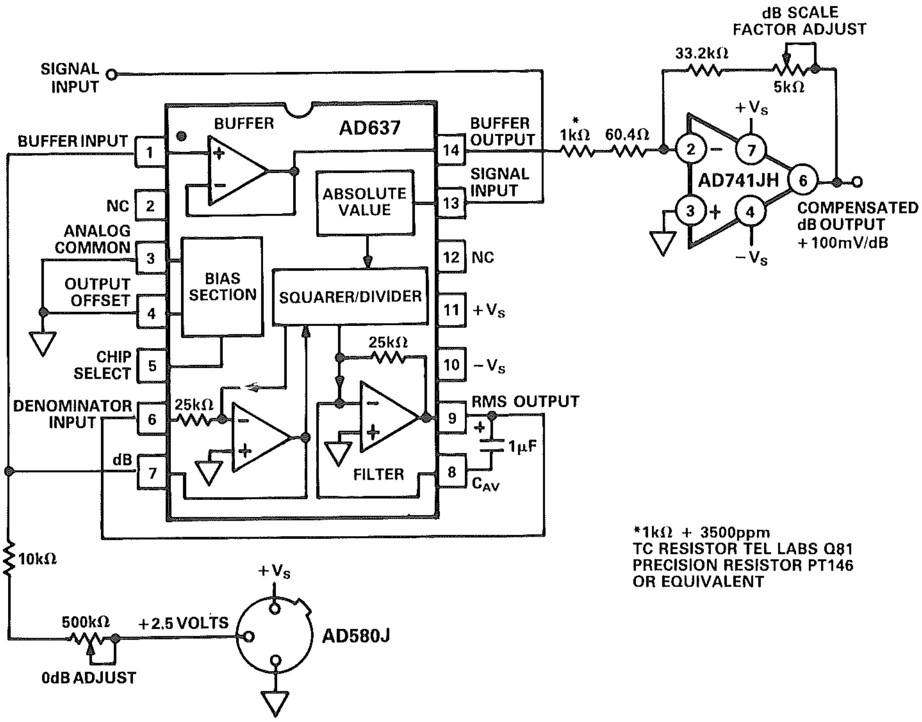
The dB output of the rms/dc converters is another useful circuit function. The output voltage changes by 3mV/dB if the dB terminal is driven by a constant current (which provides an I_{ES} reference for the dB stage). This voltage has a very high output impedance and it is also proportional to kT/q so it must be buffered with a buffer whose gain has a q/kT characteristic to compensate. This is quite easily done with the +3500ppm/°C resistor type mentioned earlier.

AD536A/AD636 TEMPERATURE COMPENSATED dB OUTPUT CIRCUIT



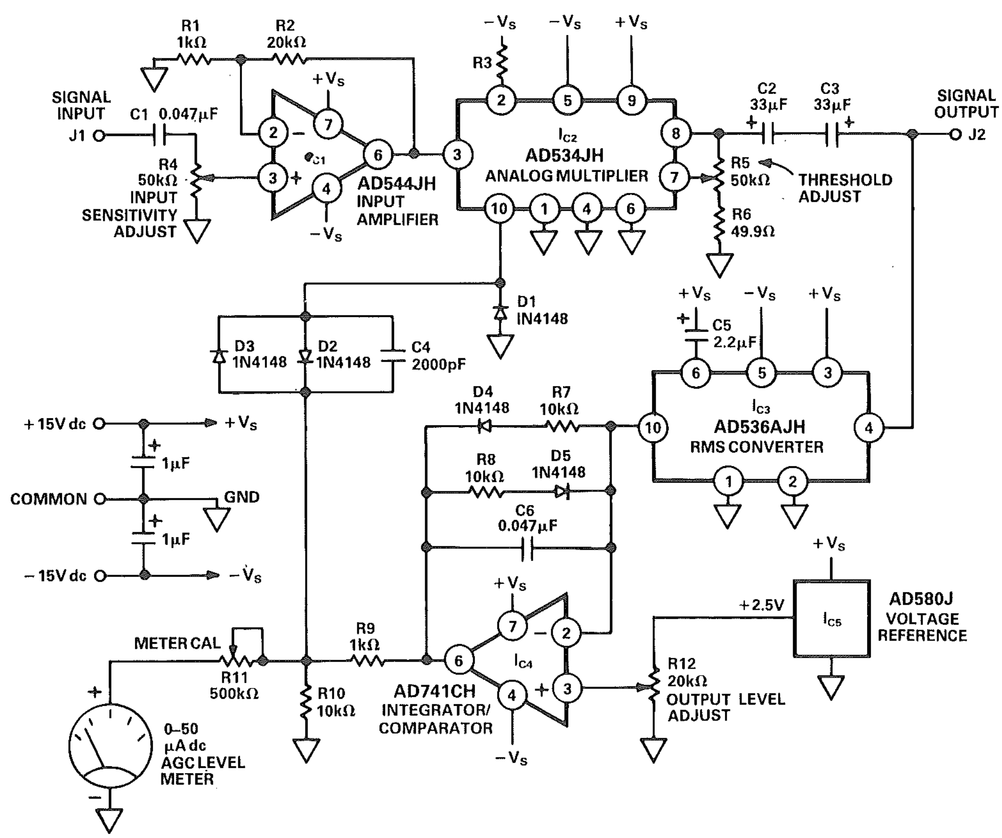
*SPECIAL TC COMPENSATION RESISTOR +3500ppm 1% TEL LABS Q-81
PRECISION RESISTOR CO #PT146 OR EQUIVALENT

AD637 TEMPERATURE COMPENSATED dB CIRCUIT

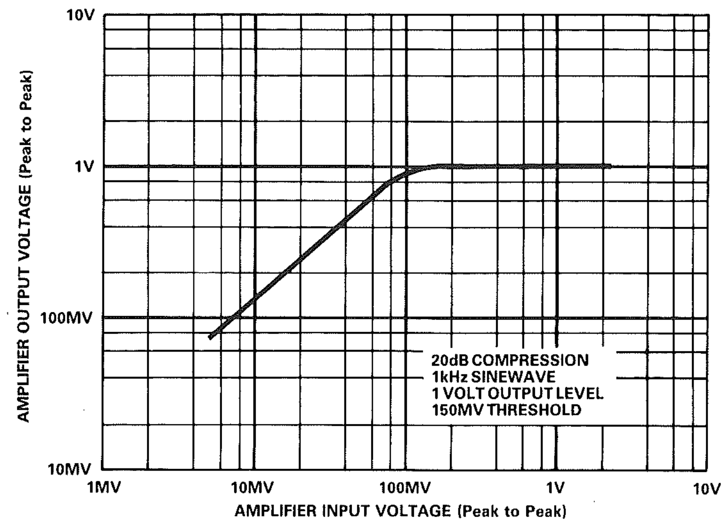


The output of an rms/dc converter may be used to set the gain of an amplifier in an automatic gain control (agc) system so that the output amplifier runs at constant output power.

AN AUDIO RMS AGC AMPLIFIER

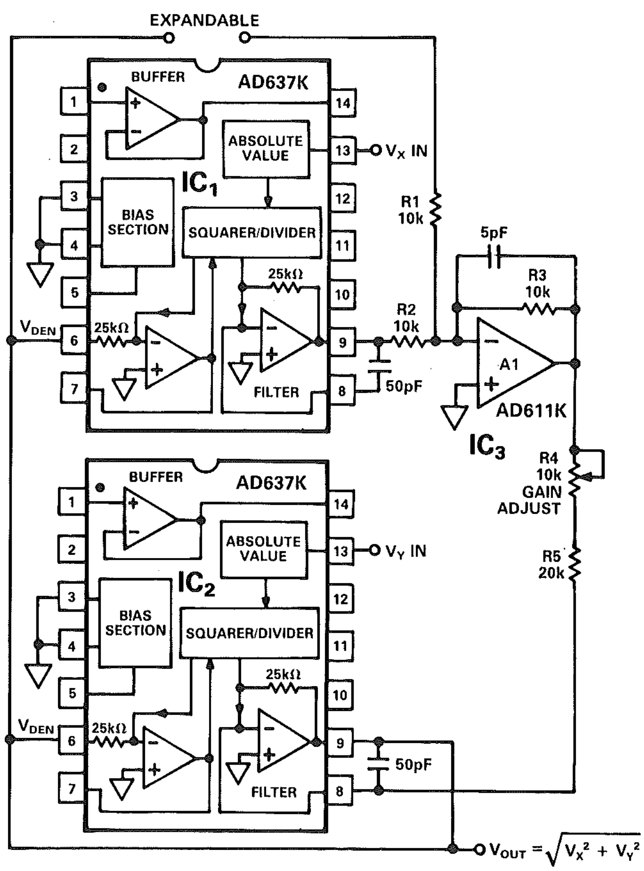


INPUT VS OUTPUT—AUDIO RMS AGC AMPLIFIER



The last application of rms/dc converters is in vector summation. If we connect two or more AD637s in the circuit shown and their inputs are X1, X2 etc. the output of the system is $\sqrt{X1^2 + X2^2 + X3^2 + \ldots}$. This is another implicit calculation (the output is fed back to all the denominator inputs) and gives better accuracy and a wider dynamic range (60dB) than the explicit calculation.

A VECTOR SUMMATION CIRCUIT



CONCLUSION

This presentation on the uses of analog computation cannot be exhaustive. It has merely touched the surface of what computation and other circuit techniques are possible with precision laser-trimmed analog computation circuits. It if has given a slight insight into what can be done, however, it will have achieved its purpose—to encourage engineers to consider analog as well as digital solutions to computational problems. Once this is done there will be no need to urge more use of analog circuitry—it will be the obvious and cost-effective solution to many circuit and system problems.

The author would like to close, however, by offering a small gift to cut out and stick in your auto window:

THINK ANALOG