Current-compensated op amp improves OTA linearity

by Jacob Moskowitz
Raytheon Co., Portsmouth, R. I.

The control-input resistance of operational transconductance amplifiers which is nonlinear at low bias voltages, must be made linear if the device is to be used in certain voltage-controlled amplifier or automatic gain control applications. Here is a control scheme that provides a linear gain characteristic while preserving the full output-voltage swing of the OTA.

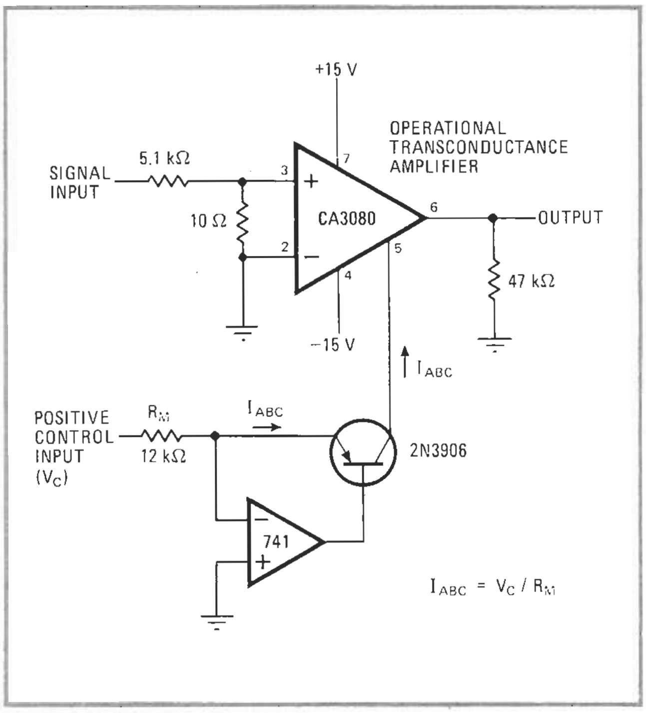
In OTAs such as the RCA CA3080, a forward-biased pn junction between the bias-current terminal and the chip substrate (negative supply) causes the nonlinearity of the control input resistance. Placing a resistor between the control voltage source and the bias terminal swamps nonlinear effects for large bias voltages, but fails to linearize the gain for biases of less than a volt or so.

A better solution, proposed by Walter Jung in the April 1975 Journal of the Audio Engineering Society, provides good linearity by means of a controlling op amp. This solution is practical for OTAs operating in an op-amp summing mode, but since the negative supply pin of the OTA is tied to the output of the controlling op amp, the output-voltage swing of the OTA is limited to values between the output voltage of the op amp and the positive supply voltage.

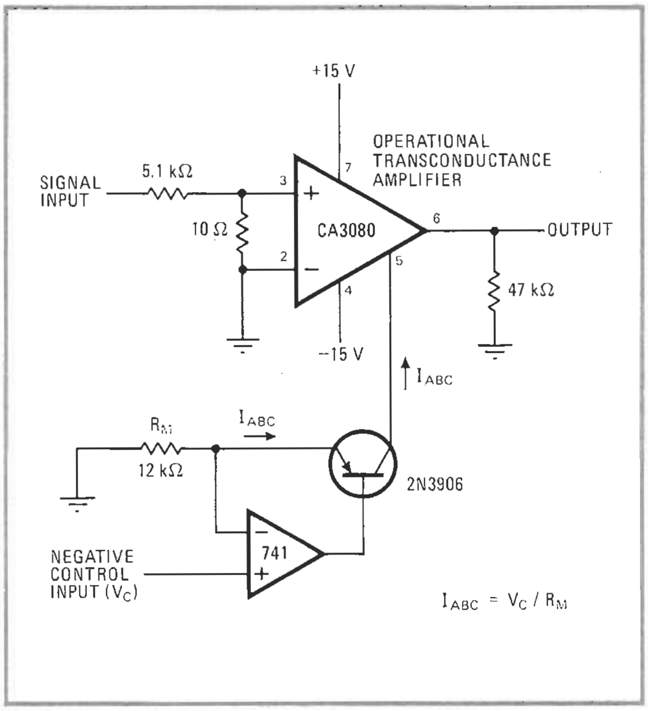
A linearizing scheme permitting full output voltage swing is shown in Fig. 1. In this version, a pnp transistor in the feedback path of the controlling op amp produces a current source linearly proportional to bias voltage V_C. This provides both gain-to-control-voltage linearity over a large control range (almost four decades), and minimum gain (in the linear portion) when V_C is zero. It is also much better suited for vca or agc applications than are four-quadrant multipliers, since control voltages below 0 will not produce negative outputs. Instead this circuit only turns off harder, which is most practical in audio work. In addition, the circuit of Fig. 1 requires no adjustment or trimming.

Figure 2 shows a similar circuit for use with negative control voltages. Since the control voltage is applied to the noninverting input of the op amp, the circuit exhibits a much higher input impedance than the circuit of Fig. 1, while the linearity characteristics are much the same. Both circuits offer a minimum linear gain at $V_C = 0$ and will only turn off harder for control voltages of the opposite polarity. Both allow full bipolar output voltage swing. □

1. Linearizing the OTA. A pnp transistor provides current feedback for controlling op amp to linearize operational transconductance amplifier under bias of less than 1 V. The higher the transistor's current gain, the better.



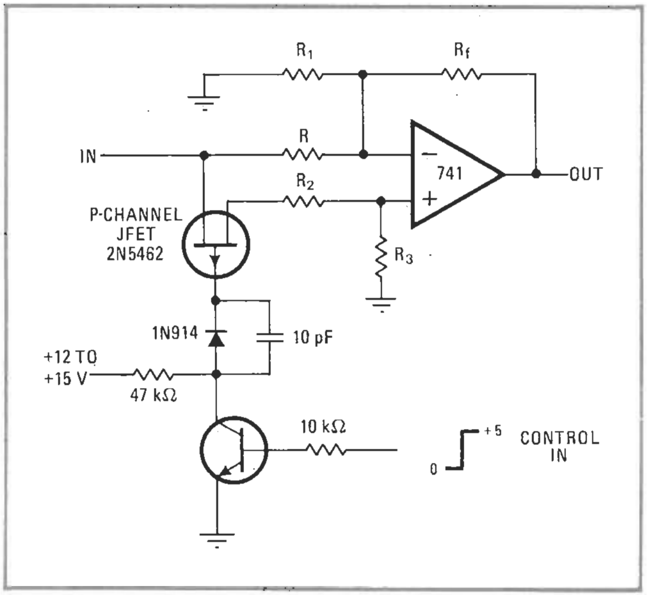
2. Negative OTA controller. For biasing OTAs with negative voltages, transistor/op-amp network is reconfigured as shown. Note controlling voltage is applied to noninverting input of op amp; input impedance is thus much greater than in circuit at left.



FET programs op amp for invertible gain

by Ken A. Dill and Mark Troll
Revelle College, University of California, La Jolla, Calif.

With only a few inexpensive components, an amplifier can be built with a gain of either +N or –N, depending on whether a field-effect transistor is turned off or on. Such a circuit is useful for programable inversion of analog signals or for programable phase-shifting of 180°



FET inverts op amp. Amplifier gain can be programmed either positive or negative, depending on whether the field-effect transistor is conducting or not conducting. Gain is the ratio of R_f to R; for gains of ±1, R_f, R_2, and R_3 are all equal value, and R_1 is half the value.

for signals that are symmetrical with respect to ground. When a comparator is added to program the inverter, the circuit becomes a precision rectifier, the output of which is:

$$V_{out} = |V_{in} - V_{ref}|$$

When the FET is off, the input signal goes only to the inverting input terminal of the operational amplifier; the gain is:

$$V_{out}/V_{in} = -R_f/R$$

But when the FET is on, the gain is:

$$V_{out}/V_{in} = \eta A/[1+(ARR_1)/(R_1R_f+RR_f+RR_1)]\theta \times [f-)R_1R_f)/(R_1R_f+RR_f+RR_1)]$$

where A is the open-loop gain of the op amp, and

$$f = R_3/(R_2+R_3)$$

Since A is large, this reduces to:

$$V_{out}/V_{in} = (f-1)(R_f/R) + f[(R_f/R_1)+1]$$

To make +N and –N numerically equal, choose the resistance values so that $R_f/R = N$. From that, it follows algebraically that:

$$N = (f-1)(N) + f[(NR/R_1)+1]$$
$$2N = fN + (fNR/R_1) + f$$
$$2NR_1 = fNR_1 + fNR + fR_1$$
$$2NR_1 - fNR_1 - fR_1 = fNR$$
$$R_1 = NRf/[2N-(N+1)f]$$

For the simplest case—a gain of ±1—all amplifier input and feedback resistors have the same value, except R_1, which is half that value.

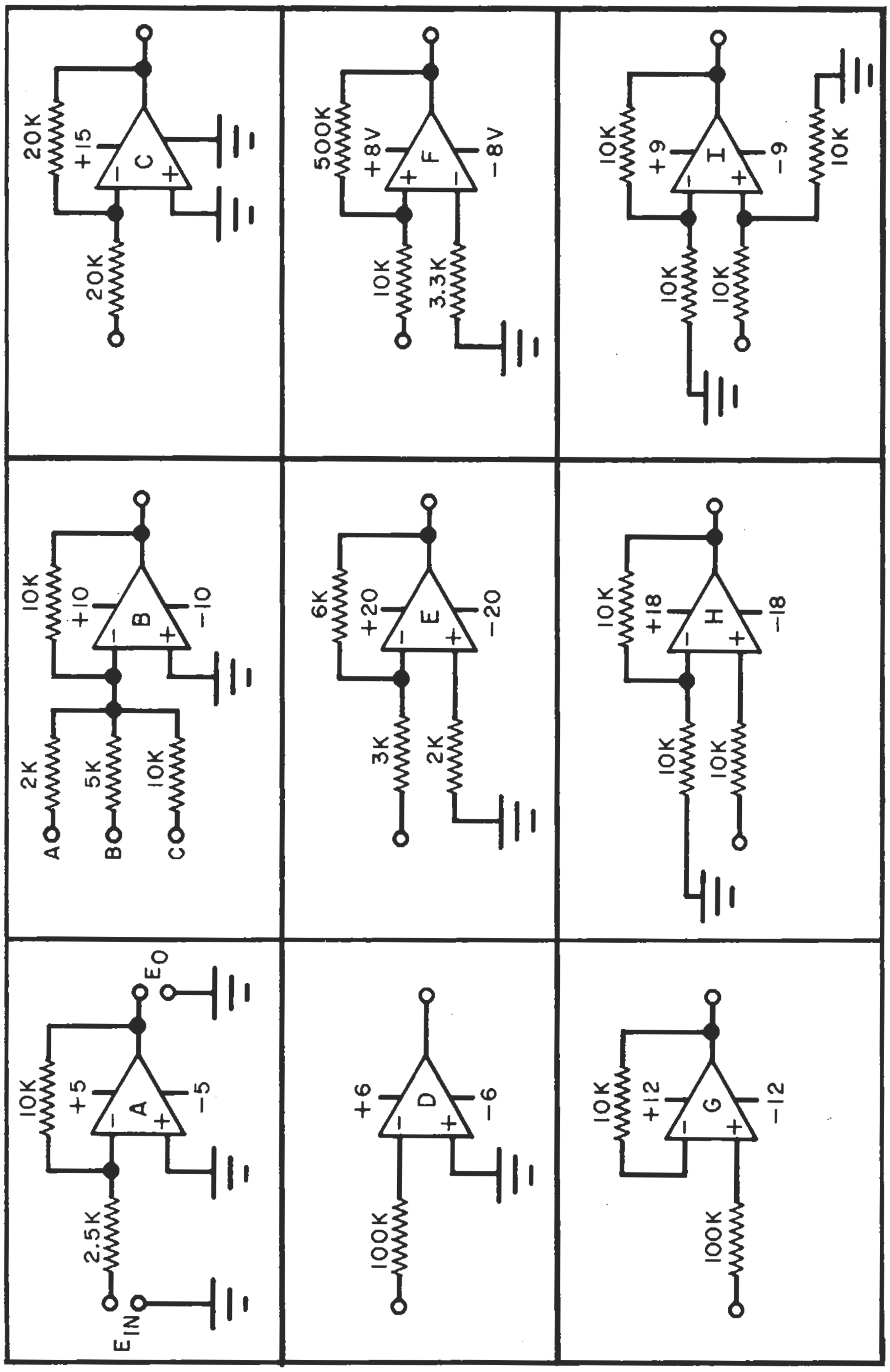
The gate of the FET is controlled by a standard analog switch configuration, which allows the inputs to be 0 or +5 volts, compatible with TTL. □

Designer's casebook is a regular feature in Electronics. We invite readers to submit original and unpublished circuit ideas and solutions to design problems. Explain briefly but thoroughly the circuit's operating principle and purpose. We'll pay $50 for each item published.

OPERATIONAL AMPLIFIER QUIZ

BY WILLIAM E. PARKER

Assume each of these circuits is an ideal op amp and each input is +1.0 volt dc. Determine the output voltages.



Answers A: −4; B: −8; C: 0; D: −6; E: −2; F: +8; G: +1; H: +2; I: +1;

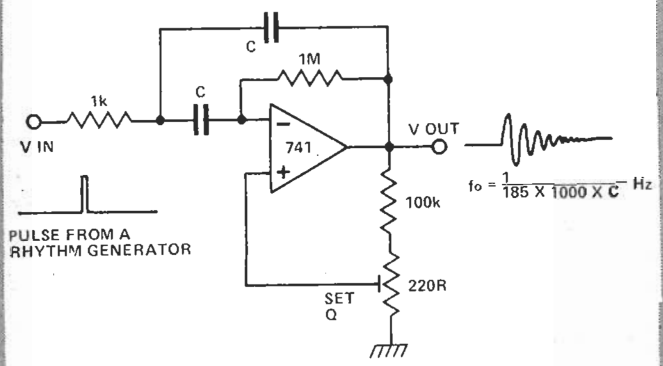
FEATURE:

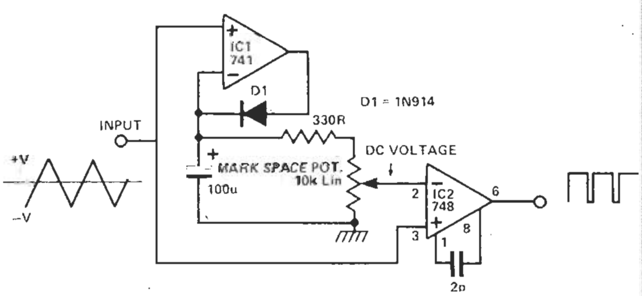
OP~AMPS PART 3

Continuing along Tim Orr's circuit strewn path through op-amps

Simple Musical Chime Generator

The circuit shown is that of a multiple feedback bandpass filter. The preset is used to add some positive feedback and so further increase the Q factor. The principle of operation is as follows. A short click (pulse), is applied to the filter and this makes it ring with a frequency which is its natural resonance frequency. The oscillations die away exponentially with respect to time and in doing so closely resemble many naturally occurring percussive or plucked sounds. The higher the Q the longer the decay time constant. High frequency resonances resemble chimes, whereas lower frequencies sound like claves or bongos. By arranging several of these circuits, all with different tuning, to be driven by pulses from a rhythm generator an interesting pattern of sounds can be produced. There may be some stability problems when high Q or high frequency operation is involved. To achieve better performance, an op-amp with a greater bandwidth than the 741 should be used. Alternatively, a different structure, such as a state variable filter could be used. Qs of up to 500 can be obtained with this latter circuit.



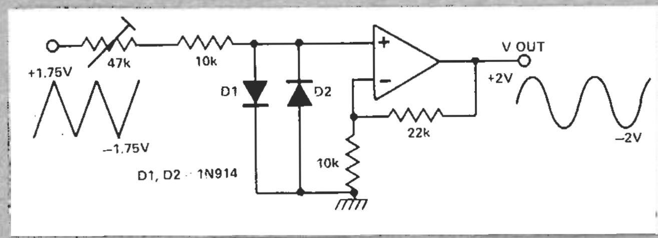


Variable Markspace Squarewave Generator with Automatic Level Adjustment

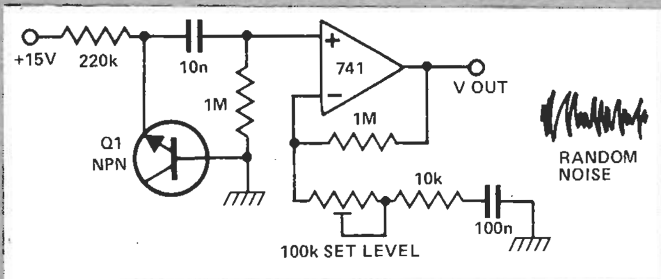
By putting a triangle wave into one input of a comparator and a manually controlled DC level into the other, it is possible to generate a variable ratio mark/space square wave. However, if the amplitude of the triangle varies then so will the markspace ratio. Alternatively, if you want the manual control to produce a very thin waveform at one end of its travel, then you will probably need a preset and a very stable triangle amplitude. However, this circuit solves these problems. The DC voltage is generated by a peak voltage follower, IC1, driven by the triangle itself. Thus the circuit tracks the peak voltage level. Secondly, only 97% of this voltage is ever fed to the comparator, IC2, and so at the end of the markspace pot, a 60:1 ratio square wave is generated. At the other end of the pot the ratio is 1:1. A 748 is used as the comparator because it has more bandwidth than the 741. As the frequency of the triangle increases, it may be necessary to use an even faster op-amp for IC2 or even a comparator.

Simple Triangle to Sinewave Converter

A simple way of converting a triangle to a sinewave is shown. The logarithmic characteristic of the diodes is used to approximate to that of a sine curve. The distortion to be expected is of the order of 5%. However, the distortion may be tolerable if the sinewave is only used to generate audio tones.



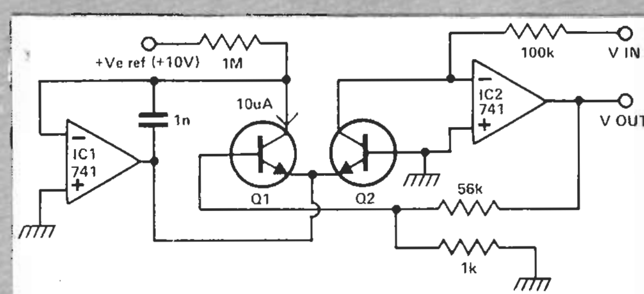
Noise Generator

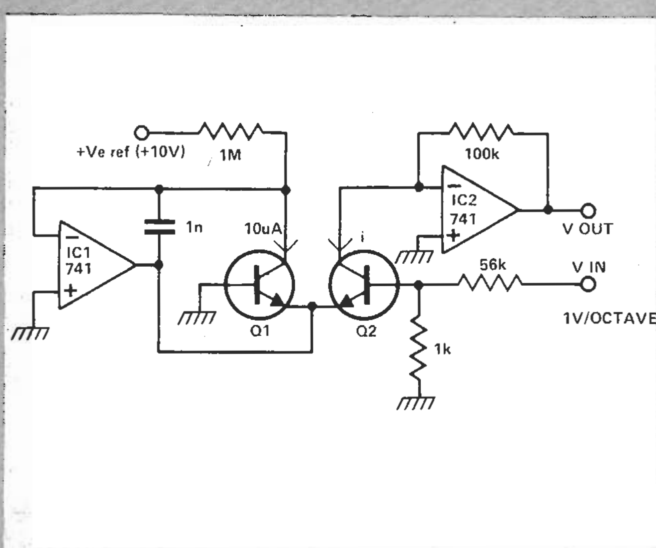


The zener breakdown of a transistor junction is used in many circuits as a noise generator. The breakdown mechanism is random and so generates a small noise voltage. Also this voltage has a high source impedance. By using the op-amp as a high input impedance, high AC gain amplifier, a low impedance, large signal noise source is obtained. The preset is used to set the noise level by varying the gain from 40 to 20 dB.

Logarithmic Voltage to Voltage Converter

The output voltage is logarithmically proportional to the input voltage. The difference between this circuit and the previous is that the exponentiator is in the feedback loop of the op-amp and hence the mathematical function has been inverted. The circuit is useful for performing true logarithmic compression or for converting linear inputs into dBs.





Exponential Voltage to Current/Voltage Converter

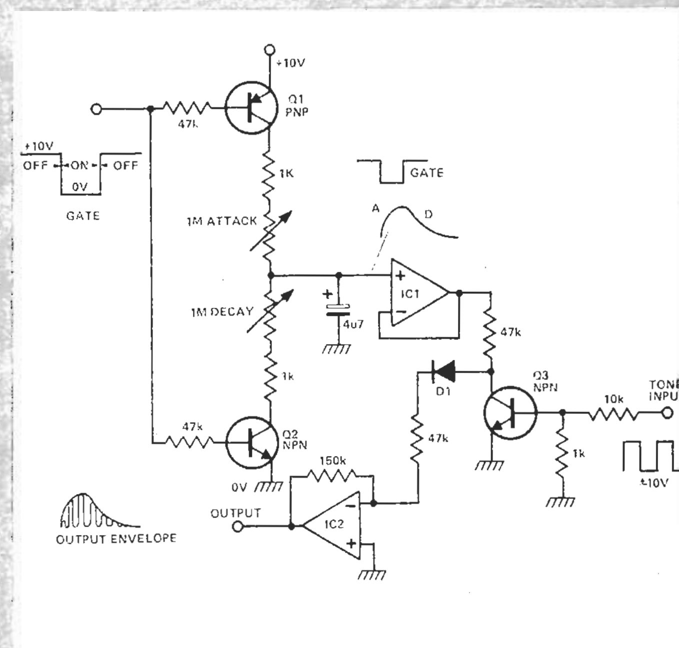
The circuit shown converts a linear input voltage into an exponential current or voltage. This type of circuit is used in music synthesisers to change linear control voltages into musical intervals. That is, if the circuit were used to control an oscillator, input increments of 1 V would change the pitch by one octave. The exponential characteristics of a transistor are exployed to generate the correct transfer function. Q1 and Q2 are a matched pairs of transistors, preferably a transistor dual. IC1 maintains Q1 at a constant current. Thus, the op-amp serves only to bias the emitter of the second transistor Q2 into a suitable operating region. The purpose of Q1 is to generate this bias voltage. The base emitter junction of a transistor has a high temperature coefficient (−1.9 mV/° C) and so the reason for using a matched pair is to use the first transistor, Q1, to provide temperature compensation for the second.

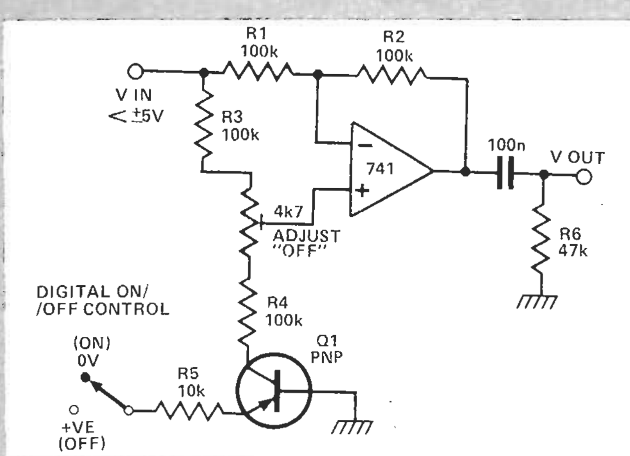
Musical Envelope Generator and Modulator

A gate voltage is applied to initiate the proceedings. When the gate voltage is in the ON state, Q1 is turned on, and so the capacitor C is charged up via the attack pot in series with the 1K resistor. By varying this pot, the attack time constant can be manipulated. A fast attack gives a percussive sound, a slow attack the affect of 'backward' sounds. When the gate voltage returns to its off state, Q2 is turned on and the capacitor is then discharged via the decay pot and the other 1K resistor, to ground. Thus the decay time constant of the envelope is also variable.

This envelope is buffered by IC1, a high impedance voltage follower and applied to Q3 which is being used as a transistor chopper. A musical tone in the form of a squarewave is connected to the base of Q3. This turns the transistor on or off and thus the envelope is chopped up at regular intervals, the intervals being determined by the pitch of the squarewave.

The resultant waveform has the amplitude of the envelope and the harmonic structure of the squarewave. IC2 is used as a virtual earth amplifier to buffer the signal and D1 ensures that the envelope dies away at the end of a note.



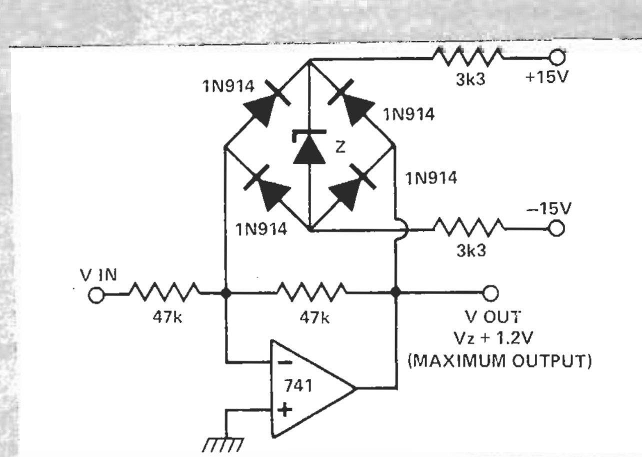


Transistor Used To Turn An Op Amp On Or Off

When transistor Q1 is switched off, the circuit behaves as a voltage follower. By applying a positive voltage to the emitter of Q1 via a 10K resistor, the transistor is made to turn on and go into saturation. Thus the lower end of R4 is shorted to ground. The circuit has now changed into that of a differential amplifier (see fig. 7), but where the voltage difference is always 0v. Now as long as the resistor ratios in the two branches around the op amp are in the same ratio then there should be zero output. A 4K7 preset is used to null out any ratio errors so that the 'OFF' attenuation is more than 60dB. The high common mode rejection ratio of a 741 enables this large attenuation to be obtained.

Fast Symmetrical Zener Clamping

The problem with using two zeners, back to back in series to get symmetrical clamping are one, the knee of the zener characteristics is rather sloppy. Two; charge storage in the zeners causes speed problems and Three; the zeners will have slightly different knee voltages and so the symmetry will not be all that good. The circuit overcomes these problems. By putting the zener inside a diode bridge then the same zener voltage is always experienced. The voltage errors due to the diodes are much smaller than those due to the zener. Also the charge storage of the bridge is much less. Lastly by biasing the zener on all the time, the knee appears to be much sharper.



To be continued

Keyboard programs the gain of an operational amplifier

by P. A. Benedetti
LAFBIC-CNR, Pisa, Italy

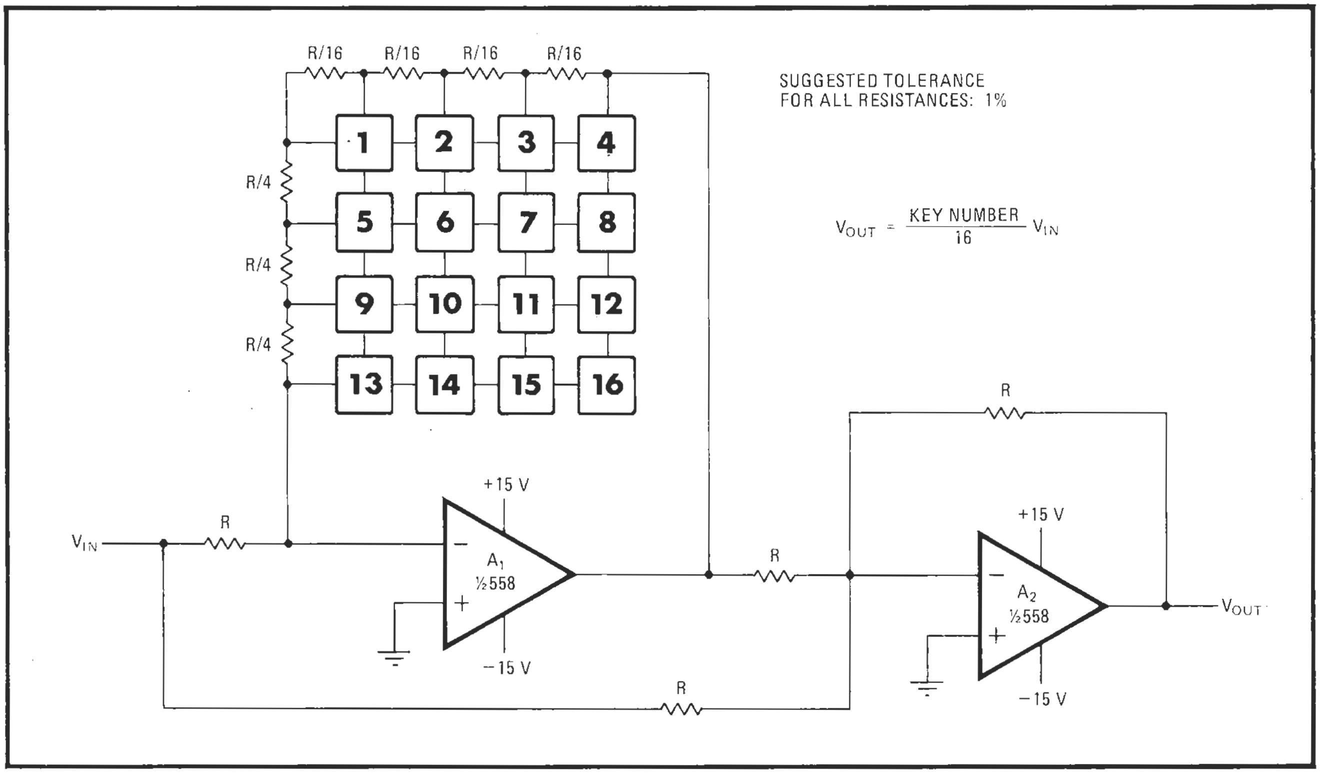
Placing a standard keyboard and a few precision resistors in the feedback loop of an operational amplifier produces a handy gain-programmable amp, useful for generating any one of several equally spaced voltages at the push of a button. Applications vary from testing components to controlling a computer program that employs an analog-input channel.

As the figure indicates, depressing 1 of 16 keys on the normally open contacts of the keyboard selects the value of the feedback resistor placed across the 558 operational amplifier A_1, to which a fixed input voltage V_{in} has been applied. A_1's gain varies with feedback resistance, of course, and so the output voltage also varies and assumes 1 of 17 equally spaced values (including 0), depending on which button has been depressed. The resistance values in the feedback loop have been selected so that the output voltage at A_2 is:

$$V_{out} = \left[\frac{\text{Key number depressed}}{16}\right] V_{in}$$

As might be expected, the programmable-gain principle applies to a keyboard of any size. Resistor precision must vary accordingly, however, becoming greater as the number of keys increase.

Key-bounce effects are not a problem except in some computer-based applications. A solution is to include double-testing of contact points in the software. □



Digital control A standard keyboard and a few precision resistors in op amp's feedback circuit generate an output voltage proportional to the number on the key depressed. Circuit applications vary from component testing to analog-voltage control of computer systems.

Offsetting op amp forms low-threshold detector

by William D. Kraengel
Valley Stream, N.Y.

The offset-voltage option on many operational amplifiers can generate a low-voltage reference for millivolt-threshold comparator and detector circuits. As shown in the figure, offsetting the op amp slightly provides a reference voltage for the transducer amplifier circuit, which would otherwise require additional components to form a separate reference source.

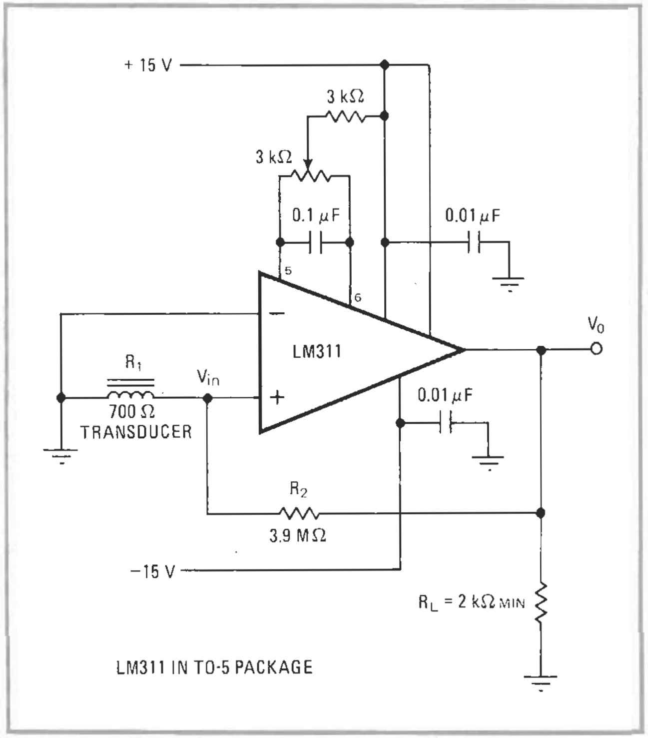
The desired reference voltage for the proper triggering point in this circuit is determined by the equation:

$$V_{ref} = \frac{V_{utp}(R_2) + V_{ns}(R_1)}{(R_1 + R_2)} = \frac{V_{ltp}(R_2) - V_{ps}(R_1)}{(R_1 + R_2)}$$

where V_{utp} is the desired upper threshold (switching) point of the circuit, V_{ns} is the negative supply voltage, R_2 is the feedback resistor value, R_1 is the input resistance of the transducer, V_{ltp} is the lower switching point, and V_{ps} is the positive supply voltage. For example, if we desire a V_{utp} for 10 mv with a hysteresis of 4.5 mv (hysteresis = $V_{utp} - V_{ltp}$), V_{ref} would equal 7.3 mv at the inverting port of the op amp. Thus, triggering occurs when the noninverting port signal exceeds 7.3 mv.

However, if the inverting port were grounded, an additional −7.3 mv would be required at the noninverting port of the comparator for the switching action described above. This may be done by adjusting the 3-kilohm potentiometer between pins 5 and 6 of the op amp to offset the noninverting input by −7.3 mv.

The desired switching point can be verified by observing a constant 10 mv rms at the input of the op amp with an oscilloscope. The triggering points are clearly visible on the signal as the circuit switches.



Saving components. Reference of 7.3 mV is developed by offset control for low-threshold trigger. If input bias and offset voltages are not excessive, op amp data sheet specifications will not be greatly affected. Current compensation is accomplished internally.

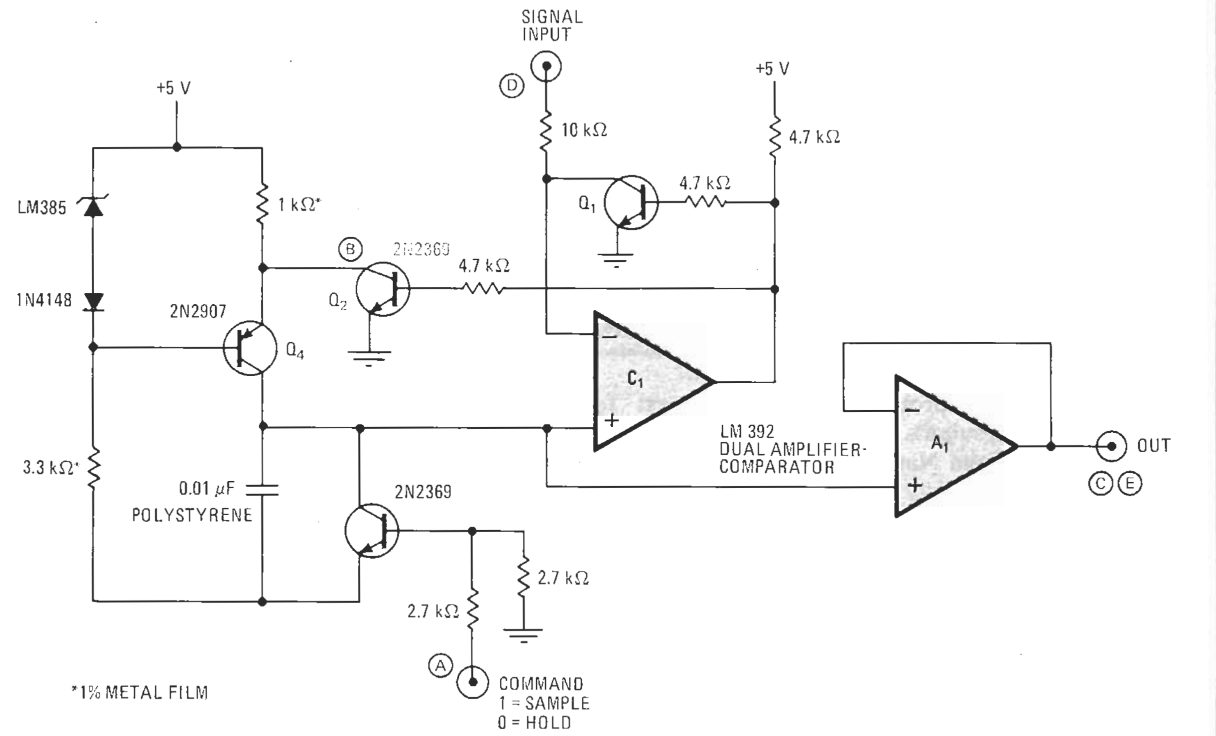
Another way to calibrate the circuit is to perform a static check. A −7.3-mv source can be derived from a resistor voltage divider and connected to the inverting input of the op amp. The transducer is shorted and the output voltage of the op amp is brought to zero by adjusting the offset potentiometer. This procedure yields the approximate switching point. □

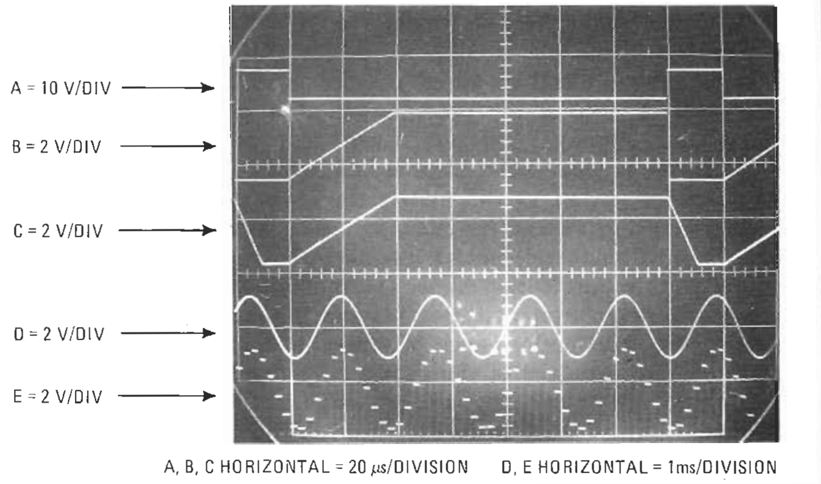
Engineer's notebook is a regular feature in *Electronics*. We invite readers to submit original design shortcuts, calculation aids, measurement and test techniques, and other ideas for saving engineering time or cost. We'll pay $50 for each item published.

Dual-function amp chip simplifies many circuits

by Jim Williams
National Semiconductor Corp., Santa Clara, Calif.

Various circuits that combine low cost, single- or dual-supply operation, and ease of use can easily be built with comparators and operational amplifiers like National Semiconductor's LM339 and LM324 because of their general applicability to a wide range of design problems. Now circuit complexity can be reduced even further with up-and-coming dual-function devices like the LM392, which put both a comparator and an op amp on one chip. Besides allowing a degree of flexibility in circuit function not readily implemented with separate chips, this device retains simplicity at low cost. The building of such circuits as a sample-and-hold circuit, a feed-forward low-pass filter, and a linearized platinum thermometer is





1. Compact sampler. A comparator and op amp chip consolidates a sample-and-hold circuit, which utilizes only one supply, has virtually zero gain error, and will not self-trigger. Owing to the configuration, the device does not have to rely on the hold cycle, so that the circuit is simplified.

discussed here in the first of two articles.

The circuit in Fig. 1 is an unusual implementation of the sample-and-hold function. Although its input-to-output relationship is similar to standard configurations, its operating principle is different. Key advantages include no hold-step glitch, essentially zero gain error and operation from a single 5-volt supply.

When the sample-and-hold command pulse (trace A) is applied to transistor Q_3, it turns on, causing Q_4's collector to go to ground. Thus the output sits at ground. When the command pulse drops to logic 0, however, Q_4 drives a constant current into the 0.1-microfarad capacitor (trace B). At the instant the capacitor ramping voltage equals the signal input voltage, comparator C_1 switches, thereby causing transistor Q_2 to turn off the current source. Thus the voltage at Q_4's collector and A_1's output (trace C) will equal the input.

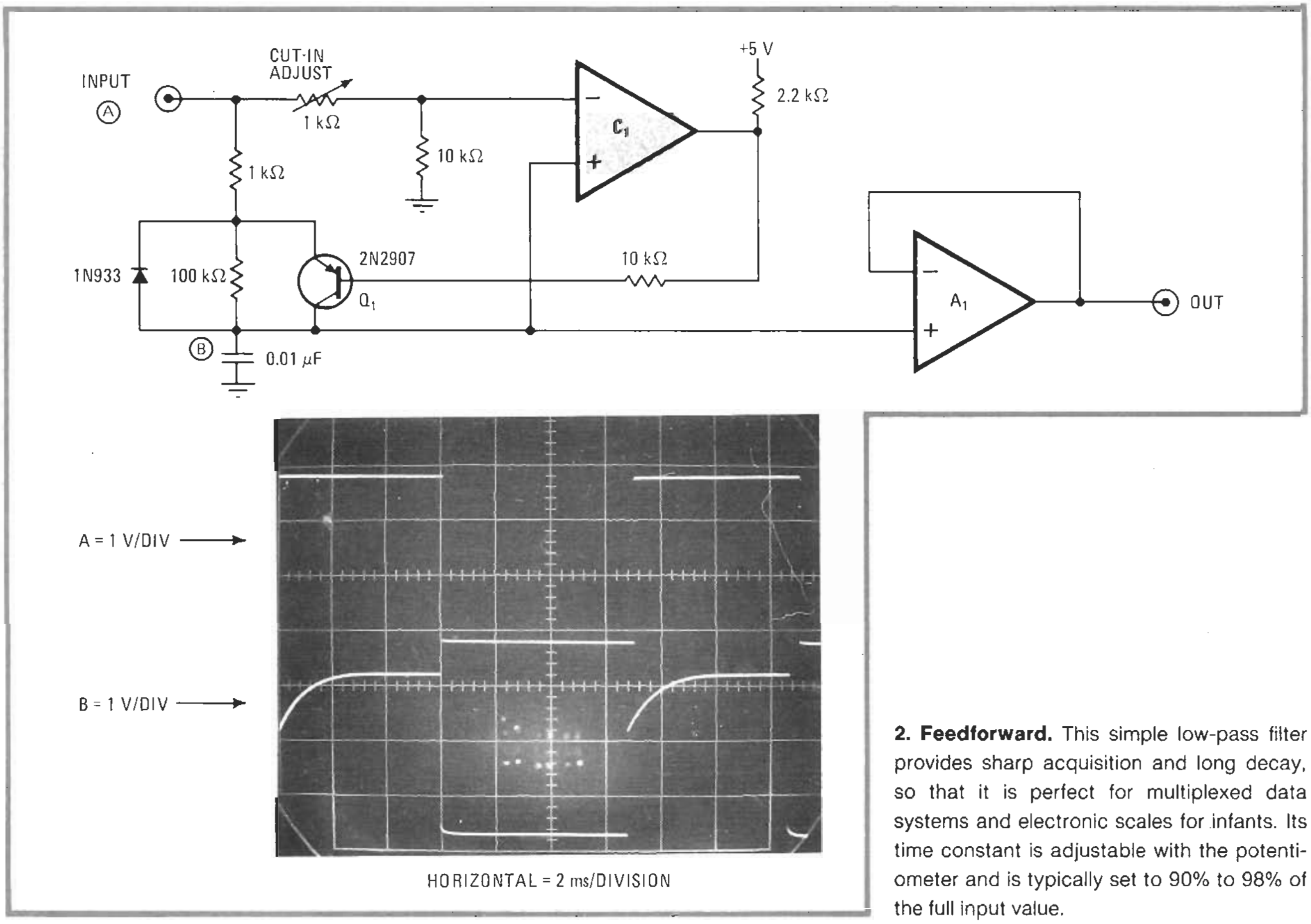
Q_1 ensures that the comparator will not self-trigger if the input voltage increases during a hold interval. If a dc-biased sine wave should be applied to the circuit (trace D), a sampled version of its contents will appear at the output (trace E). Note that the ramping action of the current source, Q_4, will just be visible at the output during sample states.

In Fig. 2, the LM392 solves a problem common to filters used in multiplexed data-acquisition systems, that of acquiring a signal rapidly but providing a long filtering time constant. This characteristic is desirable in electronic scales where a stable reading of, for example, an infant's weight is desired despite the child's motion on the scale's platform.

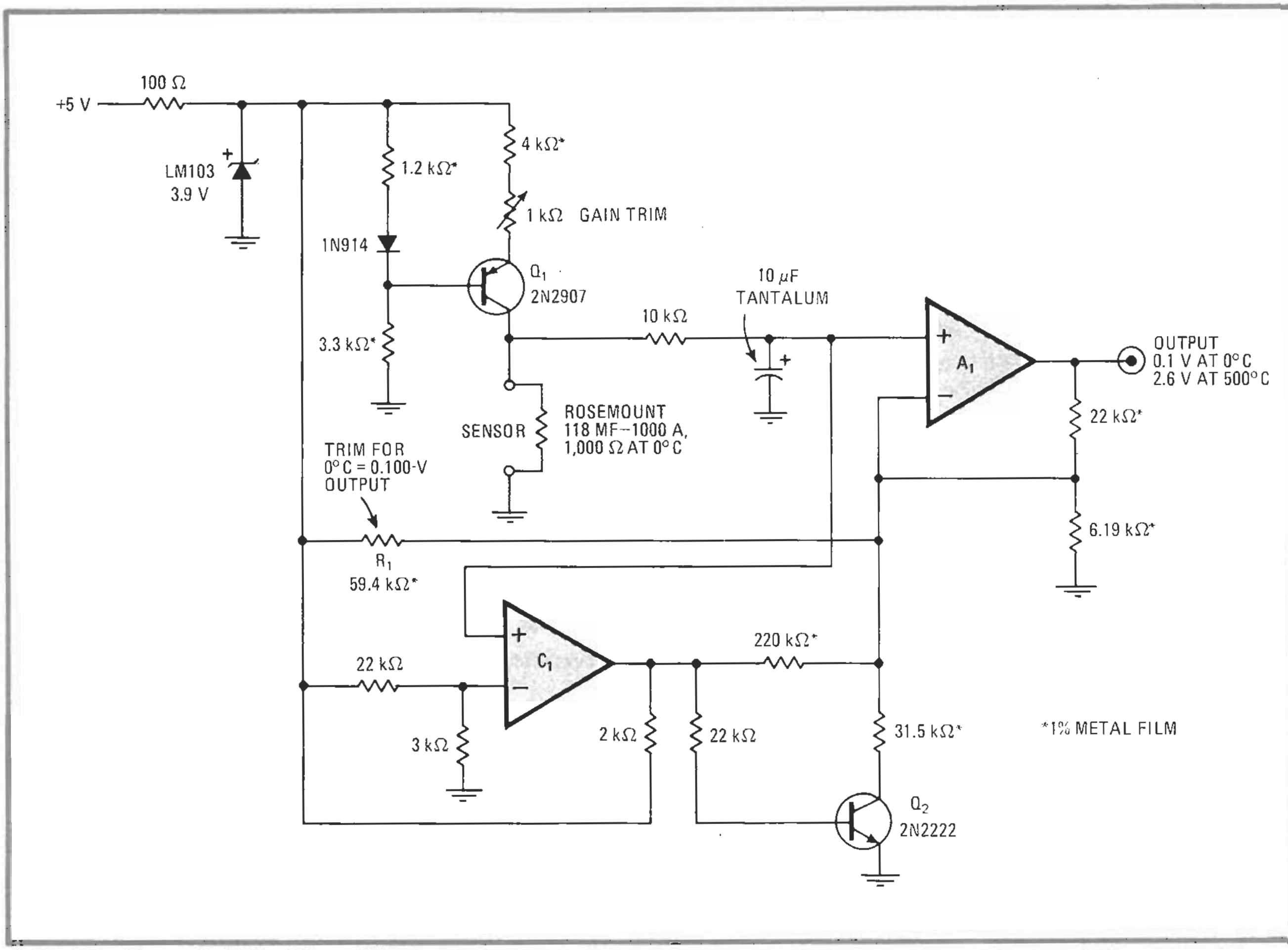
When an input step (trace A) is applied, C_1's negative input will immediately rise to a voltage determined by the setting of the 1-kilohm potentiometer. C_1's positive input, meanwhile, is biased through the 100 K $-$0.01 F time constant, and phase lags the input. Under these conditions, C_1's output will go low, turning on Q_1.

This action causes the capacitor (trace B) to charge rapidly up to the input value. When the voltage across the capacitor equals the voltage at C_1's positive input, C_1's output will go high, turning off Q_1. Now, the capacitor can only charge through the 100-kΩ resistor and the time constant must therefore be long.

The point at which the filter switches from the short to the long time constant is adjustable with the potentiometer. Normally, this pot will be set so that switching occurs at 90% to 98% of the final value (note that the trip point is taken at about the 70% point in the photo so that circuit operation may be easily seen). A_1 provides a buffered output. When the input returns to zero, the



2. Feedforward. This simple low-pass filter provides sharp acquisition and long decay, so that it is perfect for multiplexed data systems and electronic scales for infants. Its time constant is adjustable with the potentiometer and is typically set to 90% to 98% of the full input value.



3. Tracking thermals. This platinum RTD thermometer has 99% accuracy over the 0°-to-500°C range. C_1 derives the breakpoint change in A_1's gain for sensor outputs exceeding 250°C, compensating for the sensor's nonlinearity. Current through the 220-kΩ resistor shifts A_1's offset voltage, in effect preventing glitches at the breakpoint. The instrument is calibrated only at two points with a decade resistor box.

1N933 diode (a low forward-drop type), provides rapid discharge for the capacitor.

In Fig. 3, the LM392 is used to provide gain and linearization for a platinum resistor-temperature device in a single-supply thermometer circuit. This one measures from 0°C to 500°C with ±1° accuracy.

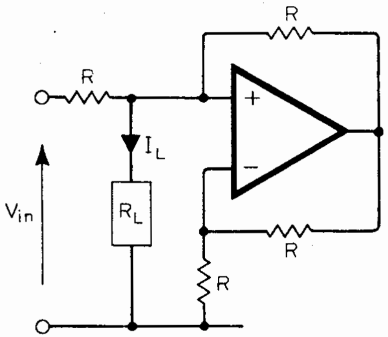
Q_1 functions as a current source that is slaved to the 3.9-V reference. The constant-current–driven platinum sensor consequently yields a voltage drop that is proportional to its temperature. A_1 amplifies the signal and provides the circuit output.

Normally, the slightly nonlinear response of the sensor would limit the circuit accuracy to about ±3°C. C_1 compensates for this error by generating a breakpoint change in A_1's gain at sensor outputs corresponding to temperatures exceeding 250°C. Then, the potential at the comparator's positive output exceeds the potential at the negative input and C_1's output goes high. This turns on Q_2, which shunts A_1's 6.19-kΩ feedback resistor and causes a change in gain that compensates for the sensor's slight loss of gain from 250° to 500°C. Current through the 220-kΩ resistor shifts the offset voltage of A_1 so no discernible glitch will occur at the breakpoint.

A precision decade box should be used to calibrate this circuit. Once inserted in place of the sensor, it is adjusted for a value of 1,000 ohms and a 0.10-V output by means of resistor R_1. Next, its resistance is set to 2,846 Ω (500°C) and its gain trim control adjusted for an output of 2.6 V. These adjustments are repeated until the zero and full-scale readings remain fixed at these points. □

Earth referenced V-to-I

The circuit idea for Earth referenced V-to-I which appeared in *Wireless World* December 1977, required one op-amp and three transistors. This simpler circuit uses one op-amp to produce a current source referenced to ground.



It can be shown that the load current $I_L = V_{in}/R$. If R_L is replaced by a capacitor, a ground referenced integrator is formed and the resulting voltage ramp may be taken via a buffer stage across C or alternatively from the output of the o-amp. For the last mentioned connection, the output ramp is modified to become

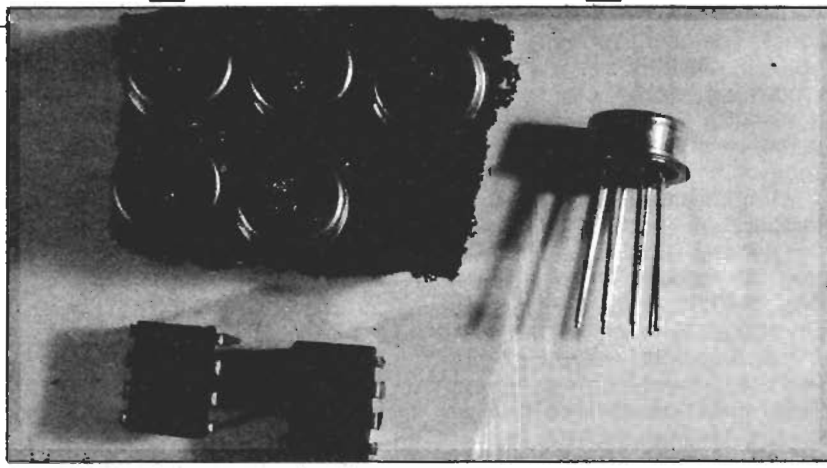
$$V_o = \frac{2}{CR} V_{in}, dt.$$

The maximum load resistance that may be used is $\leqslant V_{OH}/2I_L$ where V_{OH} is the peak output or the op-amp before saturation commences.

M. Applebaum,
Chubb Alarms Manufacturing,
Middx.

Using biFET & biMOS Op Amps

The availability of BiFET and BiMOS op-amps has revolutionised circuit design since they appeared on the scene five or so years ago. While we've used devices like the CA3140 op-amp in projects we've not got around to describing practical applications circuitry. This 'Lab Notes' fills that gap.



Modern BiMOS and BiFET op-amps come in both can and DIL packages.

THE AVAILABILITY of BiFET and BiMOS devices in various packages with one to four operational amplifiers per package has revolutionised the operational amplifier market. Apart from the relatively expensive hybrid FET input devices, other FET input operational amplifiers had been available for some considerable time, so why should BiFET and BiMOS devices be so important?

The first point to note is that amplifiers with FET input stages can offer far higher input impedances than devices with ordinary bipolar transistors in their input stages. For example, the well-known 741 has an input impedance of the order of 1M and a maximum input bias current of 500 nA. The use of bipolar transistors to obtain a high input impedance has been pushed to the limit in devices such as the LM108, using supergain input devices to provide a typical input impedance of 70M and an input bias current of just under 1 nA. These values may be compared with those of some of the economical BiFET and BiMOS devices, where typical input impedances are of the order of 1 Tera-ohm (one million Megohms!) and input currents are some tens of picoamps (pA) at room temperature.

Thus if one connects the input of one of these BiFET or BiMOS amplifiers to almost any circuit, it will impose a very small load on that circuit. This can be a vital consideration when one is designing such high-impedance circuits as those used in pH meters or in ionisation chamber smoke detector circuits, whose output current is inadequate to drive devices such as the 741.

INTRODUCTION TO THE BIMOS AND BIFET OP-AMP

The first BiFET products were announced by National Semiconductor in 1975 (the LF155, LF156 and LF157 series, where LF signifies Linear FET device). The main advantages of these products is that the junction FET devices used in their input stages are fabricated on the same silicon chip as the remainder of the operational amplifier. Although hybrid operational amplifiers with FET input stages had been available for some considerable time previously, all of these hybrid devices contained the junction FET devices fabricated on a separate silicon chip from the remainder of the operational amplifier. Such hybrid devices can be made to have a very good performance if adequate trouble is taken in their design, but the extra labour costs involved in the testing of the separate chips for appropriate matching characteristics and in connecting the two chips in a single hybrid package inevitably resulted in a price tage far above that of modern BiFET devices. The general type of construction of a BiFET device is shown in Figure 1, the channel between the source and the drain electrodes of the FET input devices being fabricated by ion implantation.

Although National Semiconductor produced the first BiFET products, it was not long before other manufacturers entered the BiFET market, and

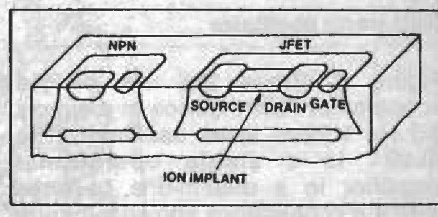


Fig. 1 Construction of a BiFET device.

such products are now available from Advanced Micro Devices, Analog Devices, Fairchild, Harris Semiconductor, Motorola, Intersil, Precision Monolithics, Raytheon and Texas Instruments, although National Semiconductor still offer the widest range of BiFET products, details of which can be found in their Linear Databook.

Very soon after National Semiconductor had announced the first BiFET products, RCA introduced their first BiMOS product, the economical CA3130 operational amplifier. This has some similarities to the BiFET amplifiers, but employs MOSFET transistors in the input stage rather than junction FET devices. RCA soon introduced further BiMOS devices, one of the best known type being the CA3140, which can be used as a pin-for-pin replacement for the 741 when a higher performance is required. More recently the CA080 series has been introduced as pin-for-pin replacements for the Texas Instruments series of TL080 BiFET types.

If one considers the very early types of monolithic FET input operational amplifiers (such as the Fairchild uA740), they do have the desired high input impedance, but their disadvantage is that their input offset voltage and its temperature coefficient are so high that they do not approach the high standard of performance required by the modern professional engineer. Modern BiFET and BiMOS devices provide a very high input impedance with relatively good stability and temperature performance — although the input impedance of any of these devices at 25°C is much greater than over the full temperature range.

In general BiFET and BiMOS economical devices offer a comparable performance. If anything, BiMOS devices tend to offer the lower input bias currents and BiFET products the lower noise levels. However, premium devices of both types are available with performances far above the average for the type of device concerned.

Half-Hertz oscillator

Figure 2 shows the use of the economical TL081 device in a simple 0.5 Hz square wave oscillator. The TL081 is a single operational amplifier in a dual-in-line package with the connections shown in Figure 2; the pin connections are the same as those of the well-known 741 devices, internal frequency compensation being employed so that no external compensating capacitor is required. External offset adjustment can be made when required by means of an external variable resistor. The TL071 is a similar low-noise device with the same connections, and is quite suitable for use in this circuit, but its low-noise characteristics are not needed. The TL061 is a low-power device with the same connections.

The frequency of oscillation of the Figure 2 circuit is given by

$$f = 1/(2\pi R_F C_F)$$

or about 0.5 Hz with the values shown. The high input impedance of the circuit enables a relatively high value of feedback resistor, R_F, to be employed, so the value of C_F can be reasonably small for a given frequency of operation. About nine-tenths of the output voltage is fed back to the non-inverting input to provide positive feedback to maintain oscillation. The capacitor C_F charges and discharges through R_F according to whether the

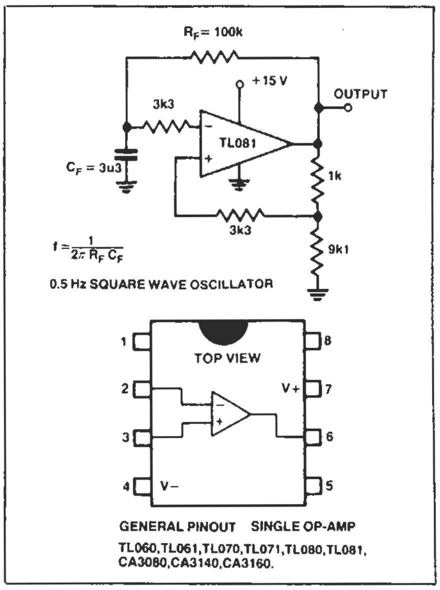


Fig. 2 Half-Hertz oscillator using a TL081 — pinout below.

state of the output voltage is 'high' or 'low' at the time concerned.

The circuit of Figure 2 generates square waves which are approximately symmetrical. However, if a circuit which generates waves with an unequal mark-to-space ratio is required, it is only necessary to connect a resistor of perhaps 10k to 50k in series with a diode across R_F. The direction in which the diode is connected determines whether the output spends the greater part of its time in the 'high' or in the 'low' state.

100 kHz oscillator

Figure 3 shows the circuit of a 100 kHz oscillator prividing two outputs which are 90° out of phase with each other. Although the TL081 is perfectly satisfactory for use in this circuit, it is more convenient to use the dual TL072 low noise version device so that this one device is all that is needed. The connections of the 8-pin dual-in-line TL082 device are shown in Figure 3; it employs internal frequency compensation, but has no external offset adjustment facilities.

BiMOS generator

A function generator which produces square and triangular waveforms is shown in Figure 4. It employs a CA3140 BiMOS device together with a CA3080A and CA3080. A particular feature of this circuit is that a frequency range of one million to one can be obtained by the use of a single variable resistor, or alternatively by the use of an auxiliary sweeping signal.

A CA3130 device may be employed instead of the DA3140 shown, but in this case a frequency compensating capacitor (about 56p) must be connected between pins 1 and 8, since the CA3130 is not internally compensated. The CA3160, which does not require any external frequency compensation, is also suitable for use in this circuit.

The high frequency linearity of the ramp is adjusted by the 7-60p variable capacitor connected between the output of the CA3140 and the output CA3080 device. The triangular wave output level is determined by the four 1N914 level-limiting diodes in the output circuit and the network connected to pin 2 of the CA3080.

It is important to minimise lead length and parasitic coupling capacitance in this circuit by careful layout.

Notch filter

The circuit of Figure 5 shows the use of a TL071 low-noise amplifier in a notch filter circuit. This is the normal 'twin-T' filter in the input circuit, in

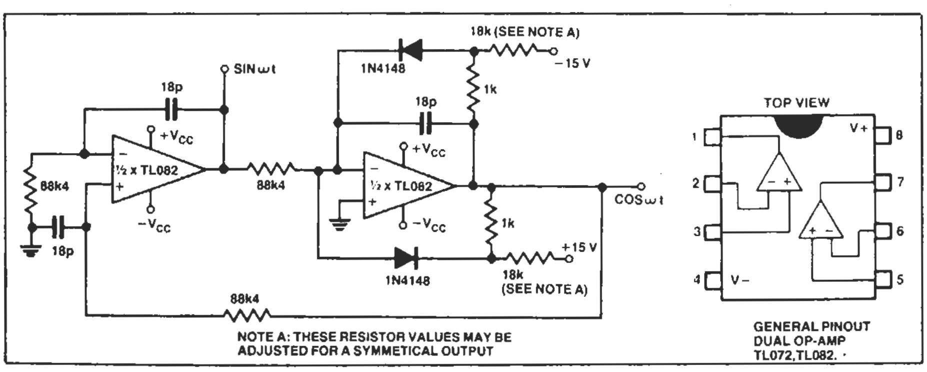


Fig. 3 Quadrature oscillator producing two outputs 90° out of phase, using a TL082 dual op-amp — pinout at right.

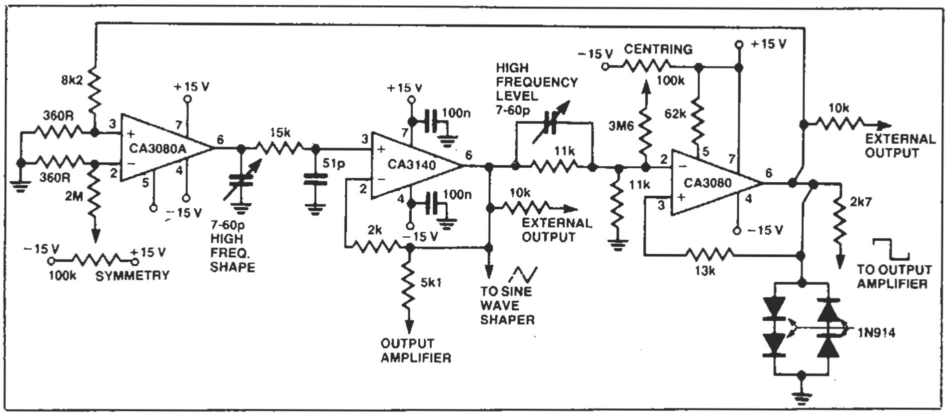


Fig. 4 Function generator circuit. Sourcing or sinking current from pin 5 of the left hand CA3080 will vary the frequency.

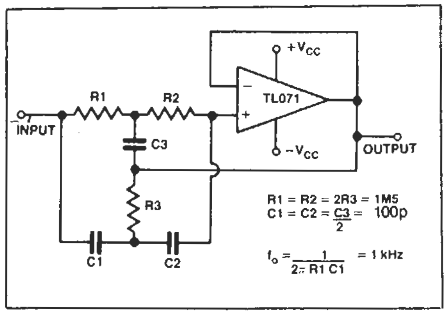


Fig. 5 Notch filter using a 'Twin-T' filter section on the input of a TL071 op-amp.

which one of the 'T' sections consists of R1, R2 and C3 and the other part of C1, C2 and R3. It is designed to reject signals of one particular frequency (the notch frequency), whilst passing signals of any other frequency virtually unattenuated.

For optimum performance, when a sharp notch in the frequency response is required, the components should have matched values (to within 1% or 2%). When the values shown are employed, the notch frequency occurs at approximately 1 kHz. An advantage of using a high input impedance device such as the TL071 is that relatively large values may be employed for R1, R2 and R3 and, therefore, for any given frequency, C1, C2 and C3 can have a relatively low value. Large value, close tolerance capacitors are expensive, so the ability to employ devices of low value is important.

Tone controls

Two tone control circuit using the CA3140 are shown in Figures 6 and 7. Figure 6 is of the Baxandall type, which provides a gain of unity at the mid-frequencies and uses standard linear potentiometers. The high input impedance of the CA3140 enables low-value (and therefore cheap) capacitors to be employed in a circuit which has an impedance great enough to avoid excessive loading of the stage feeding this circuit.

Bass/treble boost or cut are about +15 dB at 100 Hz and 10 kHz respectively. Full peak-to-peak audio output is available up to at least 20 kHz, since the CA3140 has a relatively high slew rate (about 7 V/us). The gain falls by about 3 dB at a frequency of around 70 kHz.

The circuit of Figure 7 provides similar boost and cut facilities, but the gain of this circuit is about

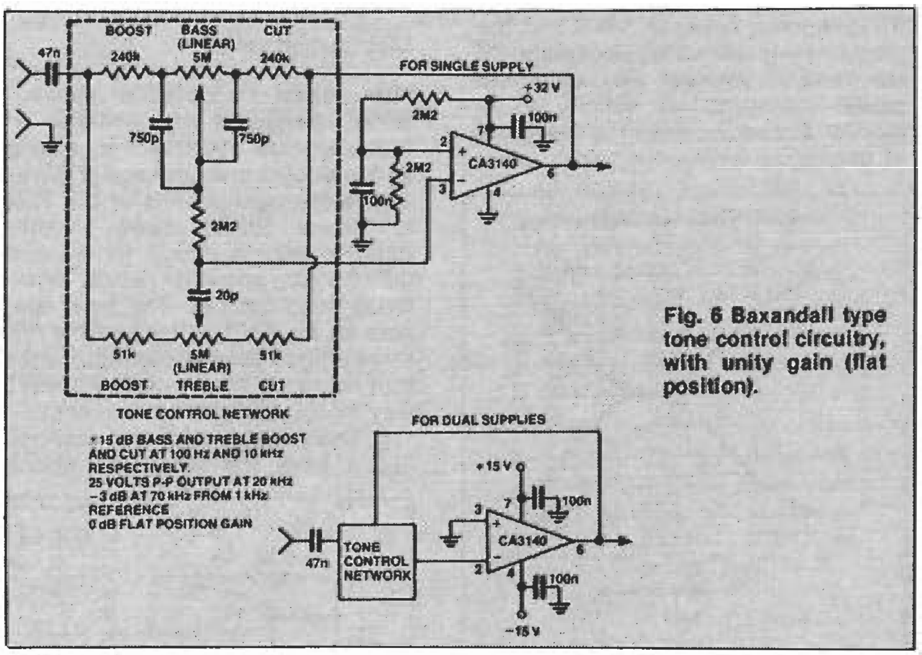


Fig. 6 Baxandall type tone control circuitry, with unity gain (flat position).

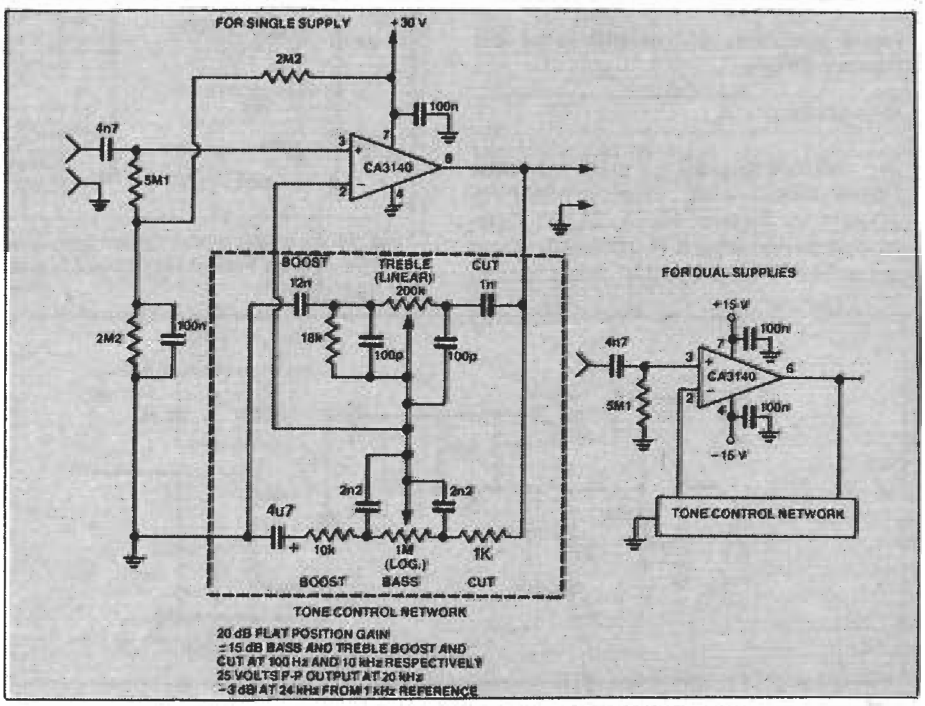


Fig. 7 Tone control circuit with 20 dB of gain, flat position.

eleven. The input impedance is basically equal to the resistor from pin 3 to ground.

A tone control circuit using the TL060 low-noise BiFET devices is shown in Figure 8. The TL060 is not internally compensated and therefore requires the 10p external frequency compensation capacitor shown connected in the circuit of each device. Similar circuits can, of course, be made using the TL080 devices at the expense of a higher power level. A further alternative is the use of TL066 programmable BiFET device without any compensating capacitors, but with a suitable value of the programming resistor between pin 8 and the negative line (about 1k, depending on the trade-off between bandwidth and power consumption which is required). Figure 9 shows the response of the Figure 8 circuit.

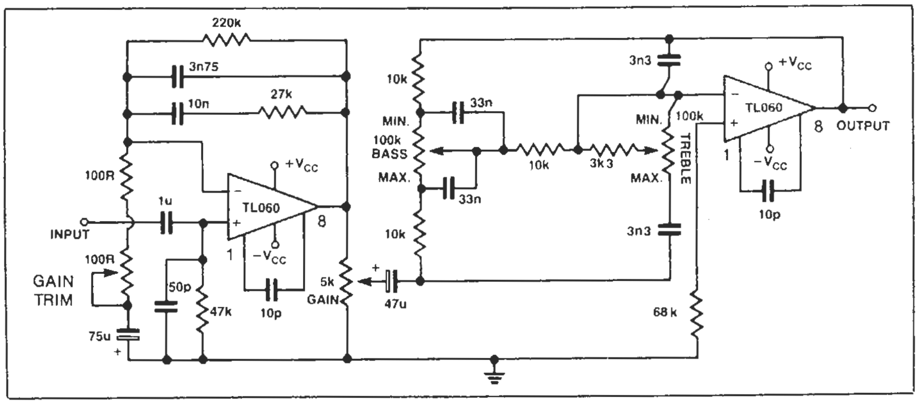


Fig. 8 An RIAA preamp and tone control circuit using TL060 devices. (Same pinout as TL070).

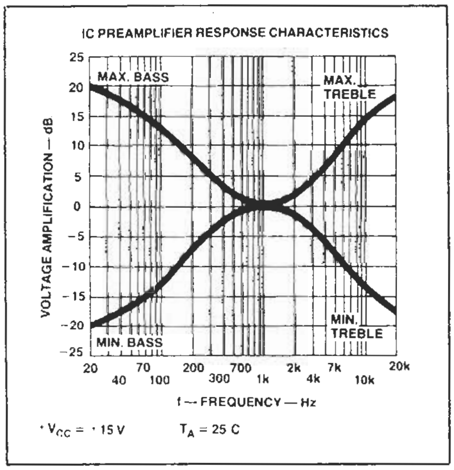


Fig. 9 Response characteristics of the Figure 8 circuit.

Mic preamp

A moving-coil microphone preamplifier with tone control is shown in Figure 10. A TL061 low-power device which is internally compensated is employed in this circuit.

Distribution amp

The Texas Instruments series of BiFET devices is also available with four separate amplifiers in a single 14-pin dual-in-line package. Figure 11 shows the connections of the TL064 low-power BiFET quad amplifier, together with a circuit for an audio distribution amplifier using one of these quad devices. The input stage acts as an input buffer and the other three stages act as output buffers, so that no signal from output A finds its way into any of the other outputs.

The TL084 and the low-noise TL074 have the same pin connections, whereas the TL085 and the low-noise TL075 devices are quad types with connections similar to the RC4136. There is no TL065 at present.

Offset Nulling

Tolerances within the amplifier cause a small DC error voltage to appear on the output. This can usually be neglected for most applications, but may cause problems in sensitive DC circuits such as meter drives. The circuit shown in Fig. 12 includes a potentiometer for trimming the output to zero volts.

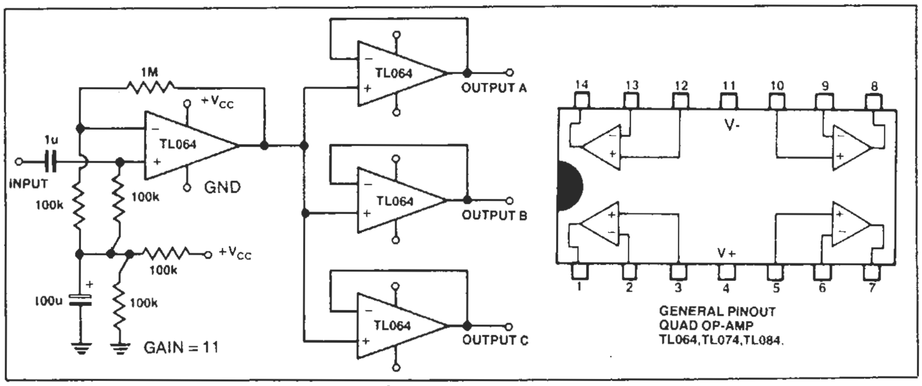


Fig. 11 An audio 'distribution' amplifier for 'slaving' several pieces of equipment from a single source. Pinout for the quad op-amp is shown at right. Circuit is for single — supply operation.

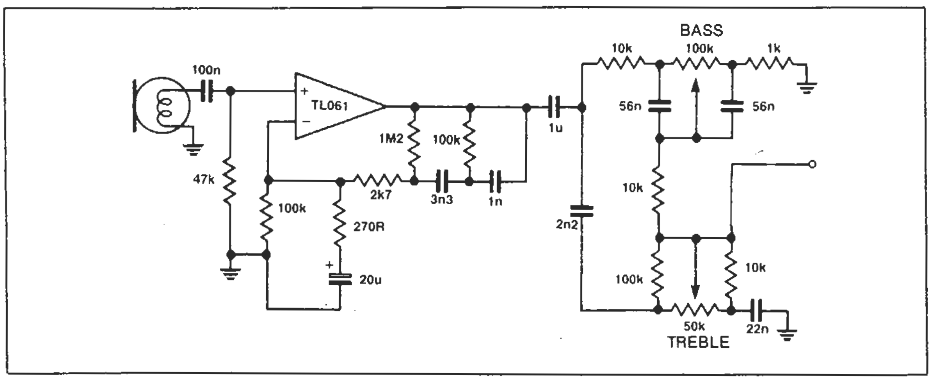


Fig. 10 Moving-coil mic preamp with tone controls, using an internally compensated TL061 device (same pinout as TL071).

Ice warning

The circuit of Figure 13 employs three of the four amplifiers of the TL084 device in an ice warning detector. It is especially suitable for use in vehicles to warn the driver when the temperature of the thermistor (placed outside the vehicle) falls below 0 °C.

When the temperature of the thermistor falls, its resistance rises and the current flowing through the thermistor decreases. Thus the inverting input of the TL084 connected to this thermistor receives less current from the positive supply line and its

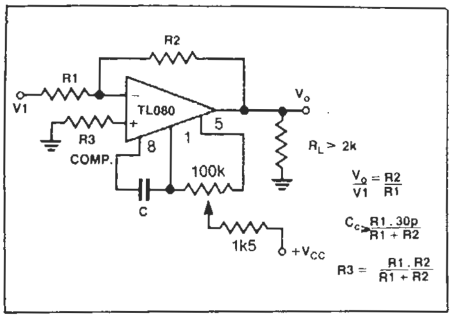


Fig. 12 Offset adjusting using the TL080.

output voltage tends to rise. This output voltage is fed to the TL084 output amplifier and produces a voltage across the LED, which lights, providing the required warning.

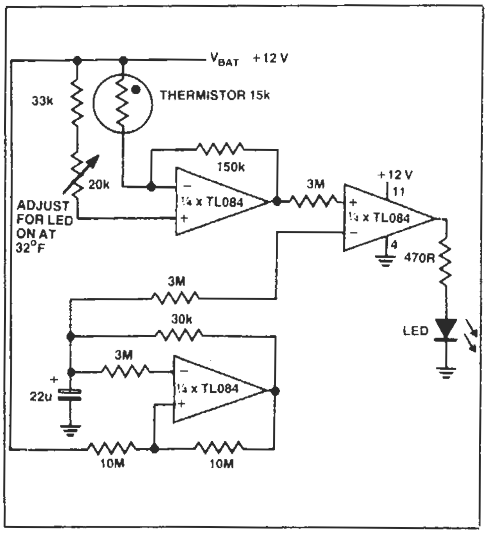


Fig. 13 An ice warning indicator.

Light detector

The circuit of Figure 14 is a low-level light detector preamplifier using the low-power TL061 device with a TIL601 or similar phototransistor. The variable resistor can be used to balance the output at any particular value of light level.

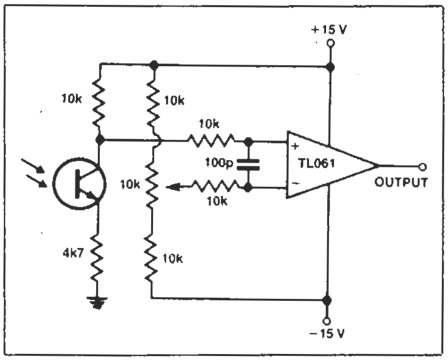


Fig. 14 Low-level light detector using FPT100 or similar phototransistor.

Sine shaper

the circuit shown in Figure 15 uses a CA3140 as a voltage follower device in combination with diodes from the CA3019 array to convert the triangular signal from a function generator into a sinewave output, which has typically less than 2% harmonic distortion.

The circuit is best adjusted using a distortion analyser, but a fairly good adjustment can be made by comparing its output signal on an oscilloscope with that from a good sinewave signal generator. The initial slope is adjusted by R1, followed by an adjustment of R2. The final slope is established by adjusting R3, thereby adding additional segments that are contributed by these diodes. Repetition of the adjustments may be necessary, since there is some interaction between the adjusting potentiometers.

Wien bridge

A CA3140 BiMOS amplifier is used in the circuit of Figure 16, together with a CA3019 diode array, to form a Wien bridge oscillator. One of the package diodes is used as a zener diode to shunt the 7k5 feedback resistor and, as the output signal amplitude increases, the zener diode impedance rapidly decreases so as to produce more feedback, with a consequent reduction in gain. This action stabilises the output signal amplitude. This combination of a monolithic zener diode and the bridge rectifier tends to provide a zero temperature coefficient for this regulating system.

As the output circuit contains no RC time constant, there is no lower frequency limit for operation. If C1 = C2 = 1u (polycarbonate) and R1 = R2 = 22M, the operating frequency can be about 0.007 Hz. At high frequencies, as the frequency is increased the amplitude of the signal must be reduced to prevent slew rate limiting from taking place. An output frequency of about 180 kHz will reach a slew rate of about 9 V/us when the output voltage amplitude is about 16 V peak-to-peak.

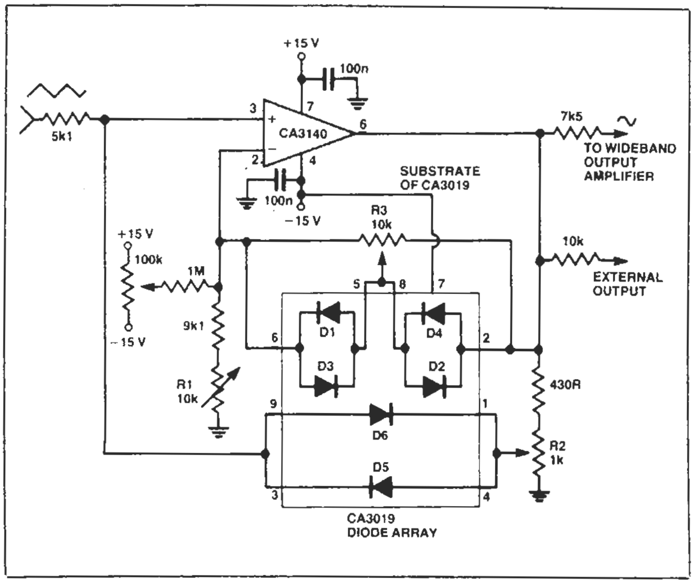


Fig. 15 A triangle-to-sine waveshaping circuit employing a CA3140 op-amp and a CA3019 diode array.

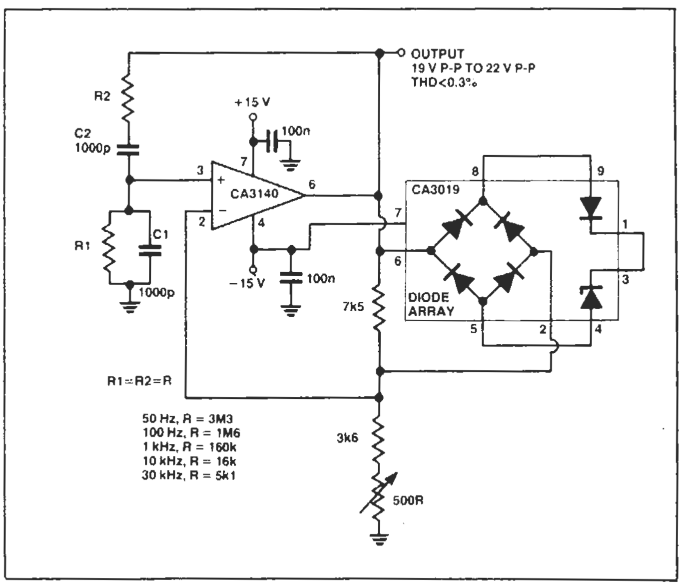


Fig. 16 A Wien bridge oscillator featuring amplitude stabilisation via the zener action from the CA3019 diode array.

Meter

The high input impedance of BiFET and BiMOS devices has led to their use in many voltmeters of high input resistance and also in meters to measure very small currents.

The circuit of Figure 17 was designed by Texas Instruments for the measurement of voltages in the range ±0.6V to ±600V, where the source resistance may be quite high, and to measure currents from 6 nA to 6 uA. The instrument was required to accept inputs of either polarity and be inexpensive, robust and reliable. It also had to have a long battery life, so a TL061 low-power operational amplifier device was selected. An inexpensive centre zero meter is considerably cheaper than a liquid crystal display and would provide adequate accuracy for the purpose.

When the switch is in one of the positions A to D inclusive, the instrument is set for the measurement of voltages. The amplifier has a non-inverting gain of 10 and range selection is achieved by a simple potential divider network with a fixed input impedance of 1000 megohm. A panel-mounted 'centre zero' control is included in the circuit to facilitate corrections for the mechanical movement of the meter zero and for the change in the operational amplifier input voltage offset (for example, with temperature).

In the current measuring mode of switch positions E to H inclusive, the amplifier operates as a current-to-voltage converter. For the most sensitive range of 6 nA, a transimpedance of 1 Gigaohm is required to produce a full-scale deflection of the meter. Rather than use a resistor of such a high value, a resistance multiplier arrangement was devised with a 100M feedback resistor for the most sensitive range.

The two diodes across the input of the operational amplifier in conjunction with R6 provide protection against any gross overloading of the instrument. A suitable arrangement incorporating a fullwave rectifier into this circuit would allow alternating input signals to be measured, but arrangements would have to be made to allow for frequency roll-off of the response at high frequencies.

3 pA meter

A CA3160 and a CA3140 are used in the circuit of Figure 18 to construct a picoammeter with ±3 pA full scale deflection (one picoamp = 10^{-12}

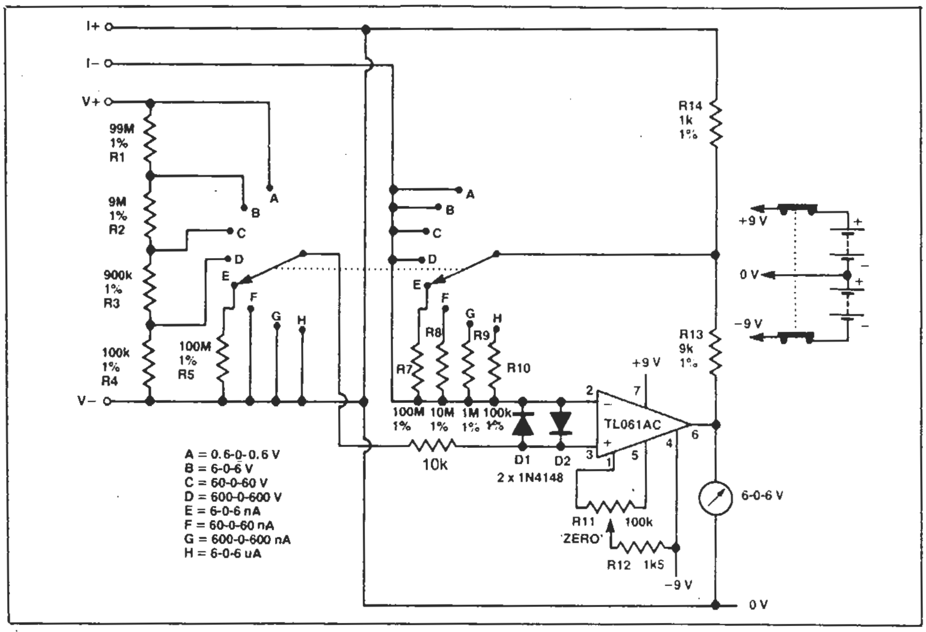


Fig. 17 A multi-range voltmeter with high impedance input plus multi-range low-current meter.

amps). Pins 2 and 4 of the CA3160 are connected to ground, so the input pin 3 between them is effectively 'guarded'. If slight leakage resistance is present between terminals 3 and 2 or 3 and 4, there would be zero voltage across this leakage resistance and this would reduce the leakage current by a large factor.

It is preferable to operate the CA3160 with its output pin 6 near the ground potential, so as to reduce the dissipation by reducing the device supply current. The CA3140 serves as a x100 gain stage to provide the required plus and minus output voltage swing for the meter and feedback network. A 100:1 voltage divider network consisting of a 9k9 resistor in series with a 100 ohm resistor sets the voltage at the 10 kMohm resistor to ±30 mV full-scale deflection. This 30 mV signal results from ±3 V appearing at the top of the voltage divider network, which also drives the meter circuitry.

It is possible to switch the 9k9 and 100 ohm network in the output circuit so that current ranges from 3 pA to 1 nA can be handled using the single 10 kM resistor.

The writer has seen circuits using BiMOS devices published for use in measuring currents down to 100 femtoamps (0.1 pA), but obviously extreme care is required to ensure the insulation is adequate when such small currents are being measured.

Voltmeter

A further voltmeter circuit covering the range 10 mV to 300 V is shown in Figure 19, which also uses a CA3160 device. The range switch SW1 is ganged between the input and output circuitry to enable the proper output voltage for feedback to terminal 2 through the 10k resistor to be selected.

This circuit is powered by a single 8.4 V mercury battery, the power supply current being

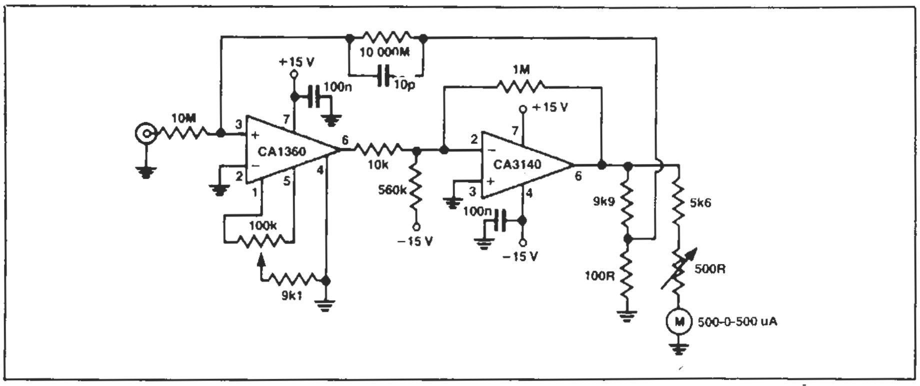


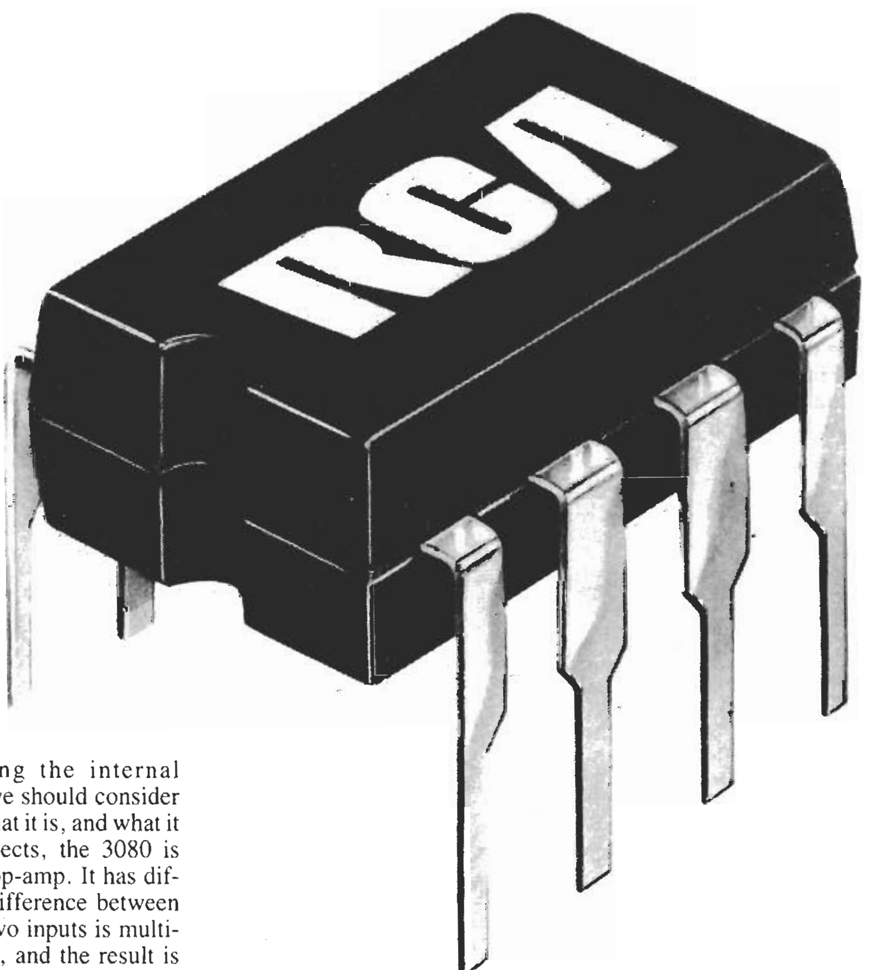
Fig. 18 This circuit will measure very low currents — full-scale deflection is ± three picoamps!

Continued on page 75

Often ignored by beginners, operational transconductance amplifiers are useful and easy to work with. This article will give you a good start toward designing and building your own projects using these versatile devices.

THOMAS HENRY

How to Use Transconductance Operational Amplifiers



IT SEEMS HARD TO BELIEVE THAT THE first integrated-circuit operational amplifiers (op-amps) were introduced less than twenty years ago. The extremely low price of IC op-amps, and the almost unlimited number of applications for them, have served to make them one of the mainstays of modern electronic-circuit design. Indeed, it is difficult to flip through the pages of any electronics publication without seeing some reference to those useful IC's.

Actually, the term "op-amp" doesn't describe one integrated-circuit, but rather is a generic term for a whole family of linear circuits. There are compensated and uncompensated types, single-supply types, current-differencing types, BiFET types, and so on. But one breed of op-amp, known as the operational transconductance amplifier (OTA) hasn't received the amount of attention that it deserves. This article details both the theoretical and practical aspects of the OTA. By the time you are done reading it, you should feel confident enough to attempt your own design with this interesting type of IC.

One of the earliest OTA's was the 3080. There are now several others available and, while they offer several interesting additional features, they essentially obey the same rules and operate just like their predecessor. Hence, designing with the 3080 is emphasized in this article, but keep in mind that switching over to other OTA's is easy.

Before considering the internal makeup of the 3080, we should consider in general terms just what it is, and what it can do. In many respects, the 3080 is much like a common op-amp. It has differential inputs. The difference between the voltages at those two inputs is multiplied by a certain gain, and the result is available at an output pin. Also, the gain can be altered simply by changing the values of certain resistors. And, finally, the 3080 needs a bipolar power-supply.

What sets it apart from the common op-amp, however, is the inclusion of another pin that allows the user to change the gain (or more properly, the transconductance) of the amplifier. That control pin, pin 5, is a current-type input. The more current, I_{ABC} that flows into the pin, the greater the IC's gain. In other words, that input current varies the transconductance of the device. In the January, 1983 issue of **Radio-Electronics**, in the series on analog-circuit design, the op-amp was modeled as a voltage source in series with an output resistance. (See Fig.3 of that article.) The transconductance amplifier, on the other hand, is modeled as a current source in parallel with an output resistance. So, yet another difference between an OTA and other op-amps is that an OTA features a current output. Speaking very generally, then, the 3080 is a current-controlled amplifier. If you consider the input for I_{ABC} (the control current) to be a programming input, then the 3080 is a programmable OTA.

What can such a device be used for? There are countless applications, but some of the more interesting ones are voltage-controlled amplifiers, voltage-controlled oscillators, sample and hold circuits, analog switches, a trianglewave-to-sinewave converter, and so on. Several of those circuits will be discussed later in this article.

Internal structure

Now that we know basically what a 3080 is and what it can do, we can start to consider its internal makeup. Unlike some other integrated circuits,

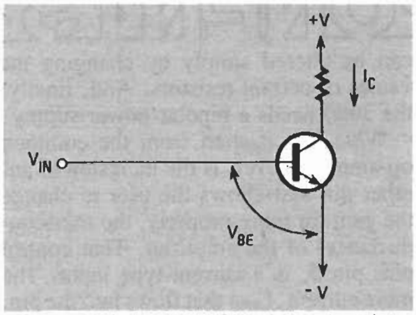


FIG. 1—THE OPERATION of the transistor differential-pair amplifier is based on the relationship of the input voltage to the collector current

where no specific knowledge of the internal circuitry is needed to use them, the unusual nature of the 3080 makes such knowledge very important.

The usual model used to demonstrate the internal structure of the 3080 is the differential-pair amplifier, although there are a few differences between them. Let's take a closer look at such an amplifier.

To understand the operation of the transistor differential-pair amplifier let's first look at the relationship of input voltage to collector current in a single transistor (see Fig. 1). That relationship is exponential and is given by the equation:

$$I_C = I'(e^{qV_{BE}/K_BT} - 1)$$

Several physical constants appear in that equation. The so-called emitter saturation-current, I prime, depends on the particular transistor used. Its value will generally be between 1 and 0.01 picoamperes. Other constants that appear include K_B, the Boltzmann constant, and q, the charge of a single electron. The equation can be considerably simplified by letting $V_T = K_BT/q$, where T is the temperature of the transistor in degrees Kelvin. The value of V_T is then about 26 mV at room temperature. Finally, the -1 term can be ignored if the transistor is forward biased. The revised equation is then:

$$I_C = I'e^{V_{BE}/V_T}$$

That revised equation is considerably easier to work with.

Figure 2 shows a transistor differential-pair amplifier. The input of the amplifier is at the base of Q1, while the base of Q2 is grounded. A control current is drawn from the two tied emitters, and it is that control current that is used to alter the gain of the circuit. The output is taken from the two collectors.

In examining the operation of this amplifier, it is convenient to assume that the beta of the transistors is large. In that case, the emitter currents are approximately equal to the collector current. Hence, $I_O = I1 + I2$. Using the transistor equation we previously discussed, and Ohm's law, the equations for I_1 and I_2 can be derived. They are:

$$I1 = \frac{I_O}{(1 + e^{-V_{IN}/V_T})}$$

$$I2 = \frac{I_O}{(1 + e^{+V_{IN}/V_T})}$$

Note the symmetry of those equations; They must always sum to I_O. That relationship is shown clearly in Fig. 3. The asymptotes of that curve are quite

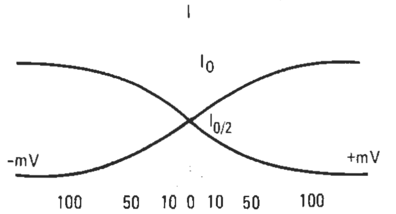


FIG. 3—THIS PLOT of emitter currents I1 and I2 shows that their sum is never greater than I_O. In fact, the sum of I1 and I2 is a constant and is equal to I_O.

important; neither I1 or I2 can be less than zero or greater than I_O. At $V_{IN} = 0$, the currents both equal $I_O/2$.

The curves are, of course, exponential, but for input voltages of less than 10 mV or so, the relationship between I1 and I2, and V_{IN}, is more or less linear. Therefore, to keep distortion to an acceptable level in linear applications, V_{IN} should be held to 10 mV or less. Also, once the input voltage exceedes 100 mV, raising V_{IN} farther has no additional affect. One of the transistors will be cut-off, while the other will be saturated.

A simplified schematic of the 3080 is shown in Fig. 4. The transistor differential-pair is quite apparent, as is the absence of resistors. In their place, current mirrors are used. In a current mirror, the output current "mirrors" the input current, hence the name. Current mirrors CM1 and CM4 are current-sinking types, while CM2 and CM3 are current-sourcing types.

Current mirrors CM3 and CM4 mimic the collector currents of the two transistors of the differential pair. The sum of those currents is thus presented to the output. If the two currents are equal, indicating equal potentials at the inverting and noninverting inputs, the currents balance and there is no current at the output. If, however, CM3 sources more than CM4 can sink, the surplus is made available at the output. Similarly, if CM4 is sinking more than CM3 can provide, the difference must be provided through the output pin.

Figure 4 also shows a pinout of the 3080. The inverting and non-inverting inputs are at pins 2 and 3 respectively; those are voltage inputs. Pin 6 is the current output and may source or sink current depending on the conditions described above. Pin 5 is the input for the amplifier control-current, I_{ABC}. Finally, pin 7 is the positive supply pin and pin 4 is the negative supply pin.

Some practical design-equations

Having described the internal structure

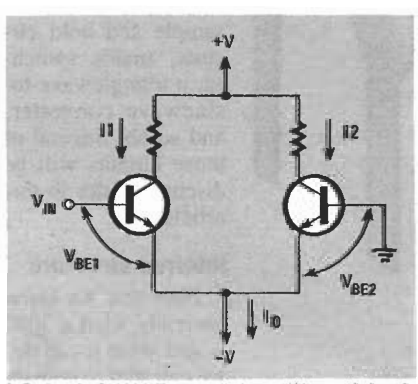


FIG 2—A SIMPLE transistor differential-pair amplifier. In analyzing it, it is convenient to assume that the beta of the transistors is very large.

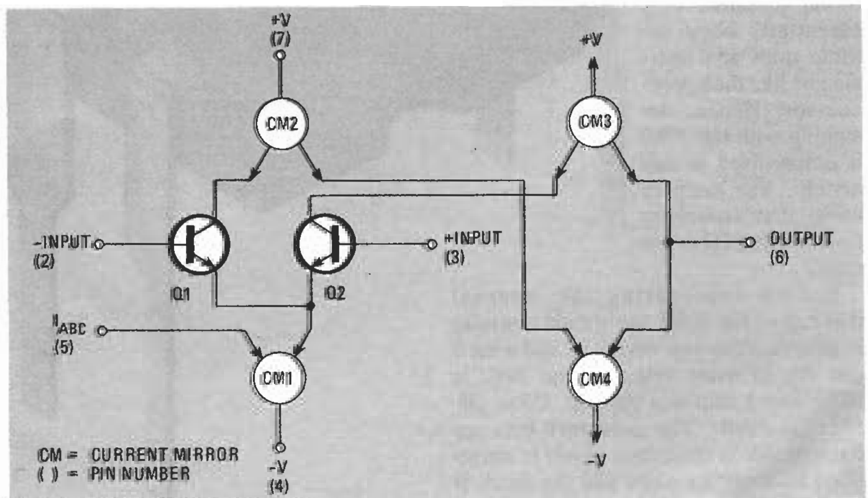


FIG. 4—SIMPLIFIED SCHEMATIC DIAGRAM of the 3080. The pin-out of the device is also shown here.

of the 3080 we can derive some practical design-equations. The most important is the so-called general transconductance equation that relates the output current to the input voltage and control current. It is: $I_{OUT} = 19.2 I_{ABC} V_{IN}$, where I_{OUT} and I_{ABC} are measured in milliamps and V_{IN} is measured in volts.

Eventually, we will want to convert I_{OUT} to a voltage, so that the unit will operate as a voltage amplifier, but for the moment, note that if V_{IN} is some fixed input-signal, we can vary the output amplitude simply by modulating the control current I_{ABC}.

Before the foregoing equation can be put to good use, some practical limits must be specified. In general I_{ABC} should always lie between 0.5 μA and 0.5 mA for best results. While it is possible to increase that upper limit somewhat, it is best not to do so since the 3080 can go into thermal runaway.

The inputs at pins 2 and 3 also have certain limitations that must be respected for good results. As we previously saw, the relationship between the input voltage and the output current is exponential and once the input reaches 100 mV any further increase will have no affect on the output current. Obviously, then, the input must be limited to less than 100 mV (or 200-mV peak-to-peak) for any sort of normal amplifier-response. But for true linear-response, the input voltage must be limited even more—a maximum value of 10 mV (20-mV peak-to-peak) is usually best. There is a trade-off here too, however, as the lower the input voltage, the lower the signal-to-noise ratio. Thus, keeping the inputs at the high end of the linear range—as close to 10 mV as possible—is desirable.

Let's now consider the supply voltage. The 3080 will work well with any power supply between ± 2 volts and ± 18 volts. Those high and low limits are extremes; best results are obtained with voltages that are somewhat between those. In many modern designs, a bipolar 15-volt supply is used, and that seems about right.

Before we move on, let's consider two other points. First of all, pin 5, the control-current input, is usually at a potential that is about one diode drop above the negative supply-voltage. Thus, if a bipolar 15-volt supply is used, the potential at pin 5 is $-$ 14.4 volts. When calculating resistor values for that input, be sure to take the negative potential into account.

Secondly, even though the 3080 is an uncompensated-type op-amp, compensation is not usually needed since most applications use an open-loop design. Compensation is only needed when negative feedback is introduced. And simplifying things still farther, the two most common negative-feedback applications for the 3080, the voltage-controlled lowpass filter and the sample-and-hold, already use capacitors in their designs. Thus no additional compensation is needed.

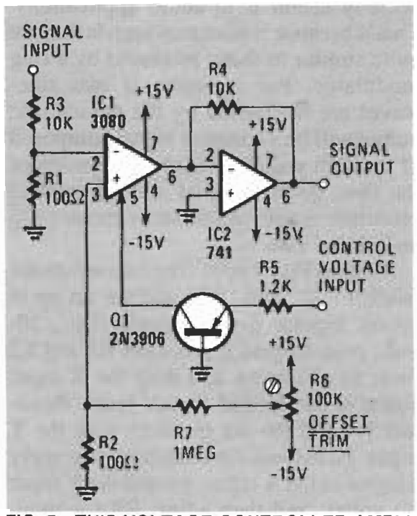


FIG. 5—THIS VOLTAGE-CONTROLLED AMPLIFIER uses just two IC's and a transistor. Though simple, this circuit works surprisingly well.

Of course, there is much more to working with the 3080, but that can be picked up with experience. For now, the rules we've presented are all that you need to begin designing. Let's see how to use them in practice.

Some practical circuits

Figure 5 shows a common 3080 application, a voltage-controlled amplifier, (VCA). That circuit is common nowadays, and shows up in everything from noise-reduction units and computerized recording-studio mixing consoles to electronic music-synthesizers. (See the May, 1983 **Radio-Electronics** for a discussion of how VCA's are used in such synthesizers.) We'll examine that circuit first because it involves a very straightforward application of the design formulas and constraints we've discussed.

Suppose that our VCA is going to process a signal with a peak amplitude of ± 1 volt. As we've said, however, the 3080 works best when the input-level is limited to 10 mV. Hence, resistors R1 and R3 are used to drop the voltage to that level. Since R1 is 100 ohms, a similar resistor, R2 is placed at the other input of the 3080. In theory the 3080 should now be properly balanced, and there should be no DC feedthrough. In practice, however, offsets can still occur and when those are modulated by the amplifier severe "thumps" will result. Therefore, trimmer potentiometer R6 is added to the circuit so that any offsets can be nulled out. To adjust that trimmer, modulate the control voltage input rapidly while watching the output on an oscilloscope. Adjust R6 for minimum DC-feedthrough.

It was seen above that I_{ABC}, fed to pin 5, should be no greater than 0.5 mA under most circumstances. Resistor R5 and transistor Q1 provide a linear current that meets that requirement. At the maximum control-voltage of + 15 volts, Ohm's law shows that R5 will conduct a current of about 0.5 mA. (Remember, pin 5 is at $-$ 14.4 volts).

A 741 op-amp, IC2 is configured as a current-to-voltage converter and will provide a low-impedance output as well. To calculate the value for R4, we apply the general transconductance equation. We know that I_{ABC} is a maximum of 0.5 mA, and we know that the input voltage is 10 mV (thanks to the attenuator—R1 and R3). Substituting those numbers into the transconductance equation yields an output current of 96μA. Now using Ohm's law, a value for R4 can be calculated. For unity gain, divide 1 volt (the original peak input-voltage) by 96μA and the result is 10.4K. Pick 10K as the nearest standard value.

That VCA, while very simple, works quite well. Perhaps its main fault is the non-linear response of the control input when the control voltage is small. A better circuit can easily be realized with the addition of a few parts. Such a circuit is shown in Fig. 6.

That circuit is actually a linear voltage-to-current converter. It will produce a current that is linearly dependent on the input voltage. In addition, since the transistor is within the feedback loop, a very precise response is guaranteed whether the control voltage is small or large. The design equation is:

$$I_{ABC} = \frac{V_{IN}\ (R1/R2)}{R2||R}$$

where R2||R3 is the parallel combination of R2 and R3.

In designing this circuit, you determine the desired output-current and then select appropriate values for R1 and R2. Those three values are then used in the equation to determine the value of R3. The circuit, using the values shown, is set up to output a maximum current of 0.5 mA for a 15-volt input. That current is fed to pin 5 of the 3080. Diode D1 is included to protect the circuit from large negative input-

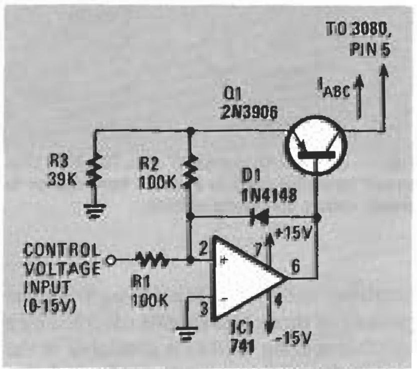


FIG. 6—FOR BEST RESULTS, this high-precision current source can be used with the circuit shown in Fig. 5. Its output is fed to pin 5 of the 3080.

voltages.

The VCA just described is actually a two-quadrant multiplier. It is a multiplier in the sense that the input signal is multiplied by a certain gain; that gain is determined by the control-voltage input. And it has two-quadrant operation because the signal is allowed to be bipolar; the control voltage, though, can only be positive. For that reason, the graph of the product of the two inputs, which is the output of the circuit, can lie only in one or the other of two quadrants of the four-quadrant Cartesian plane so familiar to most of us from elementary algebra. A four-quadrant multiplier, on the other hand, allows both the control voltage and the signal to be bipolar. Thus the output of that circuit can fall in any of the four quadrants.

Four-quadrant multiplier

Perhaps one of the most interesting circuits to come along in quite a while is a four-quadrant multiplier that uses a single 3080 and a 741 op-amp. Figure 7 shows such a circuit; its simplicity is quite striking. Before describing the circuit in detail, a few things should said about four-quadrant multiplier applications. As mentioned above, either input of the amplifier will accept bipolar signals. The product of those two signals (divided by a suitable scaling factor) is available at the output. But, most important, the polarity of the output will be correct. For example, if two negative signals are multiplied, the output will be positive. That, of course, makes such a circuit quite interesting for, among other things, analog-computer applications. (Actually the circuit given here is inverting; that means that the output will be the opposite of the true product. That can easily be corrected, if needed, by adding an additional inverting-stage).

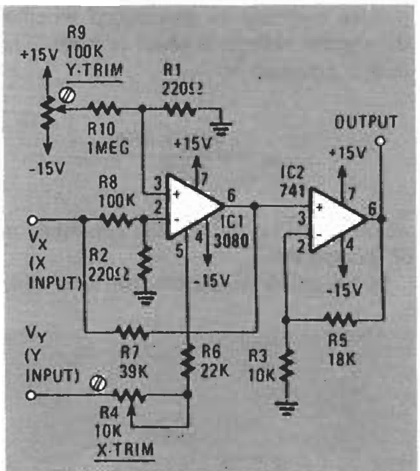


FIG. 7—A FOUR-QUADRANT MULTIPLIER. This circuit could be used in a music synthesizer to create chime and gong effects.

Another place that the circuit is especially useful is in audio applications. That's because it produces sounds that are quite similar to those produced by a ring modulator. For example, if two sinewaves are multiplied by the device, the output will be a complex signal composed of the sum and difference frequencies of the two. Such a signal can be used in electronic music-synthesis to create gong and chime effects.

Refer to Fig. 7 now. The two inputs are labeled "X" and "Y" and are set up to accept bipolar 5-volt signals (i.e., 10-volts peak-to-peak). Resistors R8 and R2 form an attenuator and drop the X-input signal to the desired 10-mV level. Resistors R4 and R6 are in series with the Y input. To balance the multiplier you apply a signal to the X input, ground the Y input (O volts), and then adjust R9 for minimum feedthrough. Then, reverse the procedure—apply a signal to the Y input, ground the X input, and adjust R4 for minimum feedthrough.

The output is converted to a voltage by IC2, a 741 op-amp. For more demanding applications, that IC should changed to a BiFET-type op-amp, such as the LF351. Note that this stage not only buffers, but also scales the output suitably—since the circuit is set up to accept bipolar 5V signals the output is scaled so that it equals $-V_XV_Y/5$. That puts the output in the same range as the inputs.

One drawback of this circuit is that only very-low-impedance input sources can be used. That is easy enough to correct, though, by buffering the two inputs. Additionally, the driving sources must be DC coupled. If those limitations are respected, however, the circuit performs very well and is far cheaper to build than any equivalent.

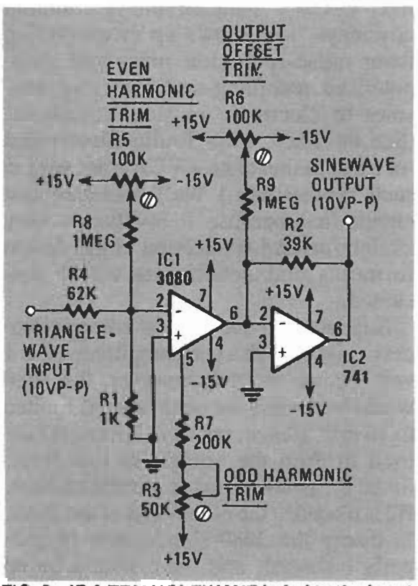


FIG. 8—IF A TRIANGLEWAVE is fed to the input of this circuit, the output will be a sinewave of the same amplitude. The total harmonic distortion will range from 2% to 4%.

Trianglewave-to-sinewave converter

When we looked at the differential pair, we said that the input signal must always be at or below 10 mV for lowest distortion. A circuit that deliberately violates that rule is shown in Fig. 8.

In that circuit, which is a triangulwave-to-sinewave converter, a triangular wave with a value of 10-volts peak-to-peak is applied to the input at R4. Resistors R4 and R1 drop the voltage to about 160-mV peak-to-peak, which is applied to the 3080. Resistor R5 is used to trim the symmetry, which reduces the even-order harmonics. Resistors R7 and R3 form the current source for the 3080, and adjusting R3 has the effect of rounding or flattening the output; the result is that the odd harmonics are reduced. By adjusting R5 and R3, a very close approximation of a sinewave can be obtained. The total harmonic distortion of the circuit will typically range from 2% to 4%. Resistor R6 is used to adjust the output offset.

The output of this circuit will be a sinewave with the same amplitude as the input triangular wave. An important thing to note about the circuit is that it is non-reactive—it uses no capacitors or inductors. Thus it will work over a wide range of frequencies.

As this article has shown, the 3080 operational transconductance amplifier is not only versatile, but quite easy to work with. The equations we've presented, and Ohm's law, are really all it takes to get circuits using that device up and running. Obviously there are many refinements that can be made—correcting for temperature effects, for instance—but they can be tackled later on when you've had more experience with the device. **R-E**

Analog ICs divide accurately to conquer computation problems

Housed in dual in-line packages, the hybrids can multiply, divide, or take the square root

by Yu Jen Wong, *Burr-Brown Research Corp., Tucson, Ariz.*

☐ Although analog dividers are basic building blocks in a wide variety of applications, until recently they remained bulky, very limited in operating range, and prohibitively expensive. Within the past two years, though, they have profited from the kinds of technological and design advances that have characterized the progress of integrated circuits in other areas.

Now, dedicated analog dividers are available in dual in-line packages, and their low price—typically less than \$20—has gone hand in hand with performance that has improved by orders of magnitude. Burr-Brown Research Corp., for instance, makes a hybrid precision divider, the 4291, with a guaranteed maximum error of less than 0.25% over a 100/1 denominator voltage. With optional external trims, the error may be held to 0.10% over a 1,000/1 range.

What is an analog divider?

Analog dividers are widely used in such applications as ratiometric measurements, percentage computations, transducer and bridge linearization, automatic level- and gain-control systems, voltage-controlled amplifiers, and analog simulations. They may be thought of as black boxes having two inputs and one output and the transfer function given by the equation:

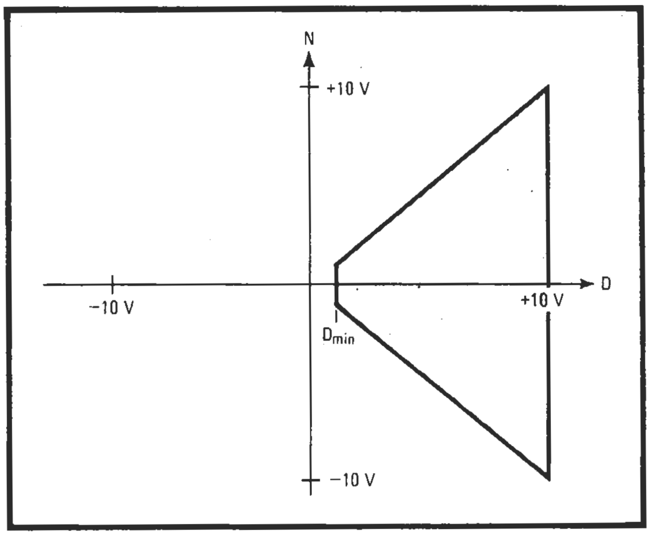
$$E_o = K(N/D)$$

where:

E_o = output voltage
K = a constant
N = numerator input
D = denominator input

For most commercial packaged dividers, K is internally set at 10. Since the divisor can never pass through zero, D is always unipolar. Because N can be bipolar, the divider will operate in two of the four quadrants, as shown in Fig. 1; it is therefore called a two-quadrant divider. Dividers that are designed for operation with N of one polarity are called one-quadrant dividers. At this point, no commercial four-quadrant divider exists, because it is impractical, though not impossible, to design one that would accept bipolar denominator voltages with a dead zone around zero.

There are two limiting conditions for every divider. First, the absolute value of N must be smaller than that of D to prevent the output from saturating beyond 10 volts. Second, a lower limit, D_{min}, is always specified for the denominator below which the divider will exhibit unacceptably large errors. These two conditions define the operating region of a divider (the shaded area in Fig. 1). For one-quadrant dividers, the operating region is either the top or the bottom half of the operating region of a two-quadrant divider.



1. Operating region. The shaded area represents the operating region of a two-quadrant divider. A one-quadrant divider will perform in either the top or bottom half of the shaded area. Below D_{min}, the denominator exhibits unacceptably large errors.

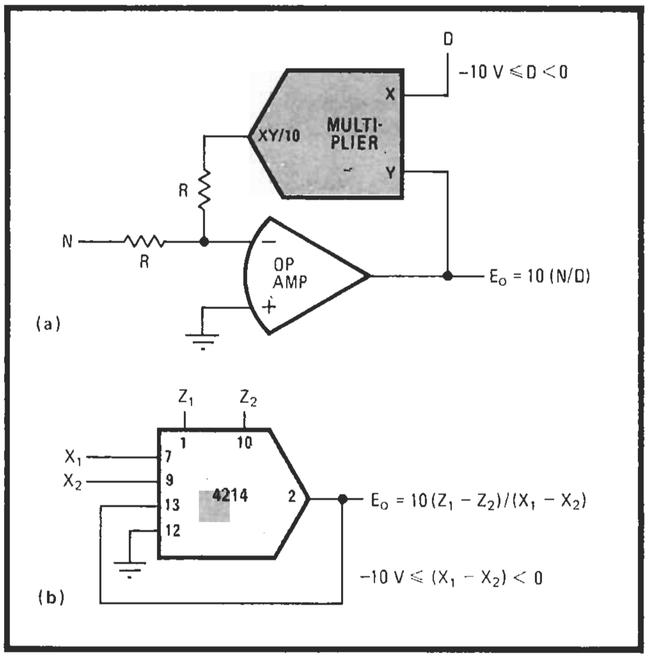
Performing division with multipliers

The oldest and perhaps still the most common method of performing analog division is to connect a multiplier in a feedback loop of an operational amplifier (Fig. 2a). An extra op amp is not needed with commercial packaged multipliers, since their output op amps can be employed through external pin connections.

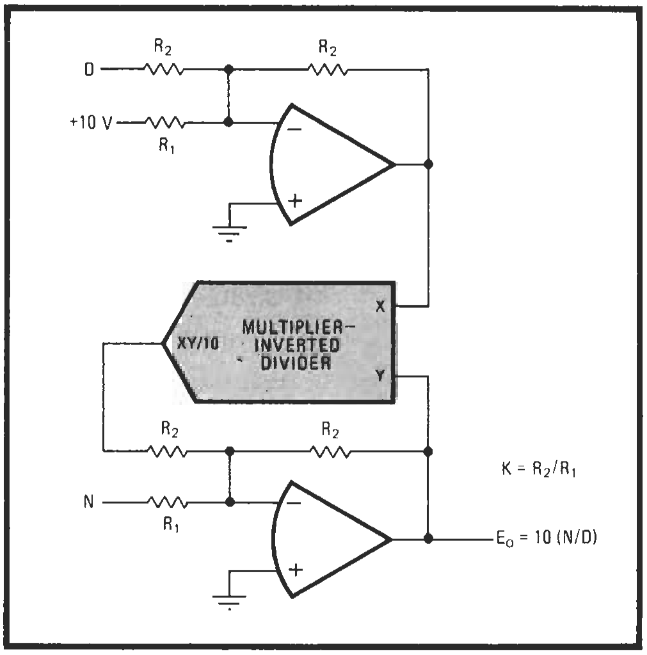
Figure 2b shows a 4214 transconductance multiplier connected as a differential divider. One limitation of the multiplier-inverted divider (MID) is its limited divisor range. The divider error that limits the ranges can be estimated by:

$$\epsilon_d \doteq 10(\epsilon_m/D)$$

where ϵ_m is the multiplier error specified by the manufac-



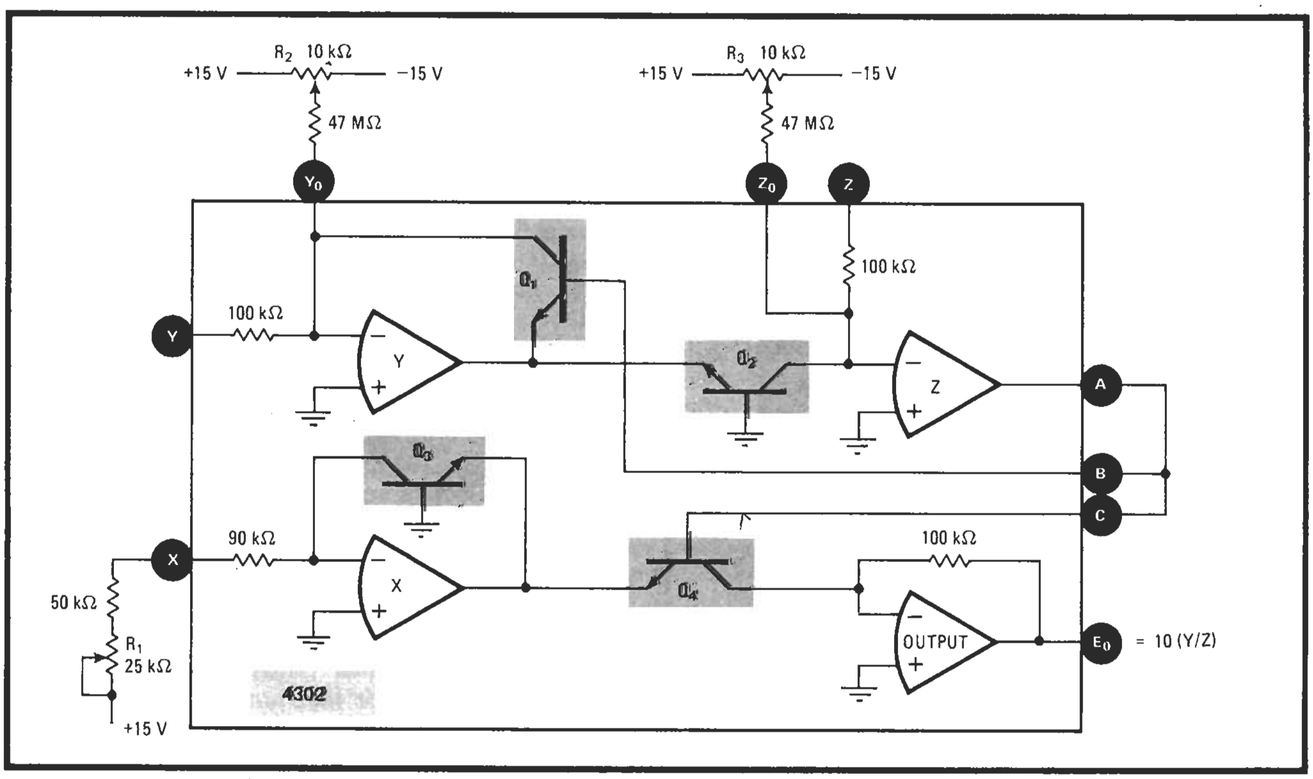
2. Economizing space. An analog divider can be constructed by connecting an analog multiplier in the feedback loop of an op amp (a). Packaged multipliers (b) do not require that extra op amp to be pin-programmed for performing division.



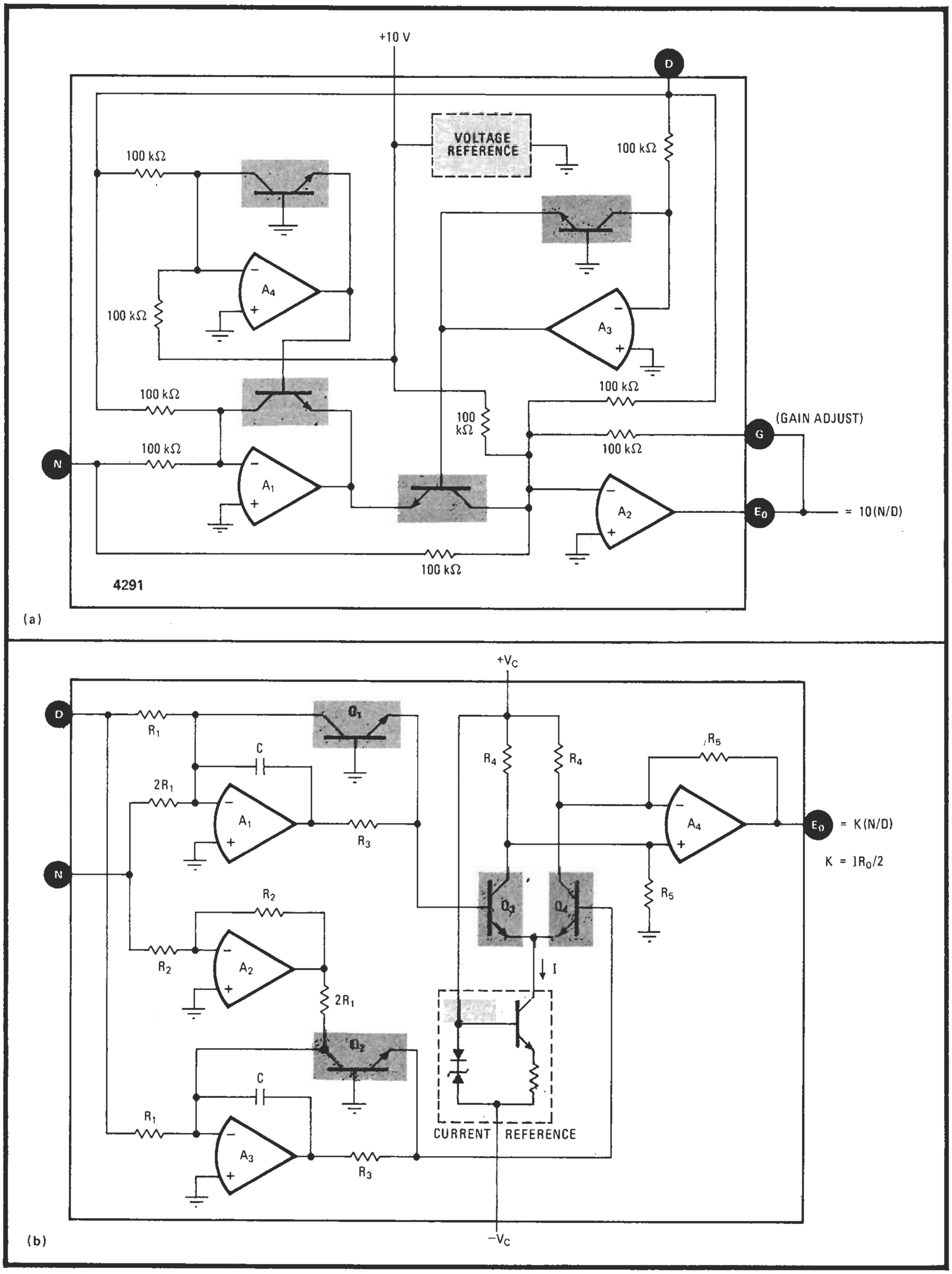
3. More accurate. The multiplier–inverted divider exhibits improved accuracy when connected in the manner shown. The error equation is given approximately by $\epsilon_d = 10\epsilon_m/(KD)$, where K is the ratio of R_2 to R_1 and can be used to optimize the divisor's range.

turer and D is the denominator voltage. With a 0.5% transconductance multiplier, the divider error will go as high as 5% when D goes down from 10 to 1 V. Hence, for practical purposes, these dividers are accurate only over a 10/1 denominator range.

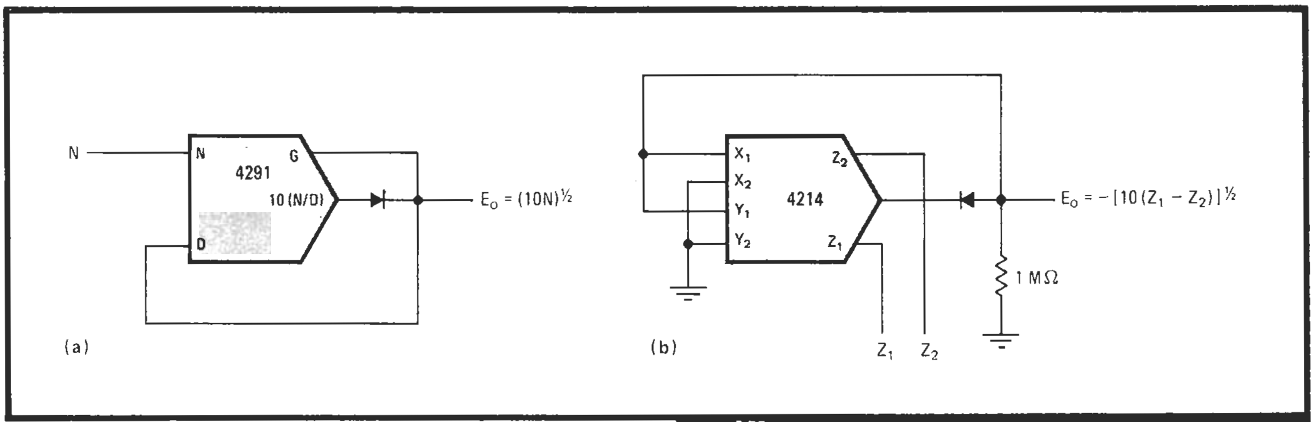
The divider error can be reduced by shifting the level of and preamplifying the divisor input and then shifting back at the output stage. In Fig. 3, with K defined by the ratio R_2/R_1, the divider error given by $\epsilon_d \doteq 10\epsilon_m/D$ will be reduced by a factor of K. However, the divisor input is thus limited to $-20/K$ volts. When $K = 2$, the divisor will swing within the same range of 0 to -10 V. In other

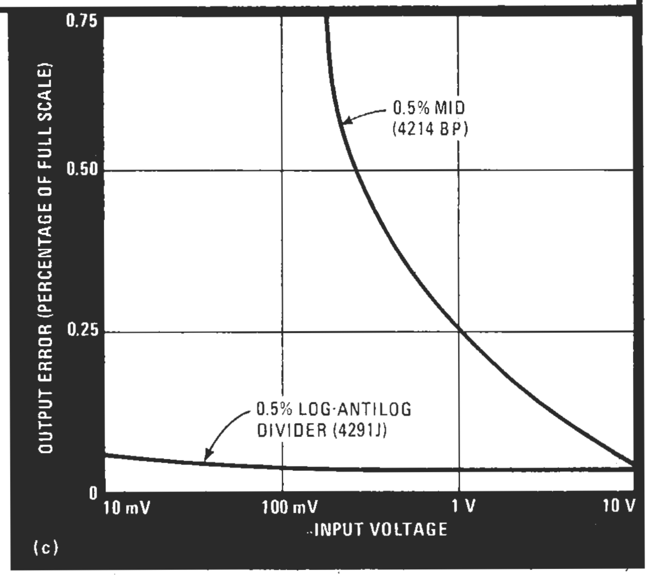


4. Converted converter. The multifunction converter may be used as a precision one-quadrant divider with a maximum error of 0.25% over a 100/1 denominator range. The maximum error can be reduced to 0.1% over a 1,000/1 range by using potentiometers R_2 and R_3.



5. Divider IC. The four op amps and the four logging transistors shown in the functional diagram (a) are rearranged and the voltage reference is replaced by a current reference to form a log-antilog divider that maintains a constant-level bandwidth with decreasing divisor voltages (b).





6. Taking the square root. Implementing either the multiplier–inverted divider (a) or the log-antilog divider (b) for finding the square root is a matter of the degree of accuracy wanted. Typical error curves for the two types are shown in the graph (c).

words, the divider error can be cut in half without sacrificing the divisor's dynamic range. With K greater than 2, say, K = 10, the divisor is limited to the range of 0 to −2 V.

One-quadrant divider

The well-known multifunction converter (MFC), through different external connections, can be used as a precision divider whose accuracy and dynamic range greatly exceed that of a multiplier–inverted divider. It is, however, good for one-quadrant operation only, whereas the MID is a two-quadrant divider.

The functional diagram of this converter is shown in Fig. 4. Its transfer function is given by:

$$E_o = X(Y/Z)^m$$

where m is determined by two external resistors and can range from 0.2 to 5. The circuit can be analyzed by applying to each of the four transistors used to achieve the logarithmic relationship, Q_1–Q_4, the Ebers-Moll equation:

$$V_{be} = (KT/q)\ln(I_c/I_s)$$

where:

V_{be} = base-to-emitter voltage
K = Boltzmann's constant (8.62×10^{-5} electron-volt/K)
T = absolute temperature
q = charge of an electron (1 ev)
I_c = collector current
I_s = emitter saturation current

Solving the equation for each of the four transistors simultaneously yields the converter's simple transfer function. This procedure assumes that the four transistors are matched, so that I_s and T are the same for all four equations.

The multifunction converter is capable of operating over a 100/1 denominator range with an error of less than 0.25%. At low input-signal levels, the offset voltages and bias currents of the Y and Z op amps contribute most of the errors. By trimming them out with potentiometers R_2 and R_3, the maximum error can be reduced to 0.1% over a 1,000/1 dynamic range. R_1 is used to trim out gain errors.

The 4291 analog divider has been optimized as a log-antilog divider. It is specified to be the most accurate two-quadrant, self-contained divider available in IC form. It operates in principle very similarly to a multifunction converter, but has several additional features. For one, it contains an internal level-shifting circuit for two-quadrant operation. For another, it is laser-trimmed to hold total error to less than 0.25% over a 100/1 dynamic range. In addition, both linearity compensation and an on-board temperature-compensated reference are provided.

Precise for two quadrants

The divider's functional circuit diagram is given in Fig. 5a. Q_1–Q_4 are the four logging transistors, which are always laid out on a monolithic chip along a thermal equilibrium line. Their geometries are specially designed for maximum conformity to a logarithmic output. In fact, log-conformity error is less than 0.05% over four decades of collector current from 100 microamperes to 10 nanoamperes. Thus, the divider can maintain its accuracy over many decades of denominator voltages.

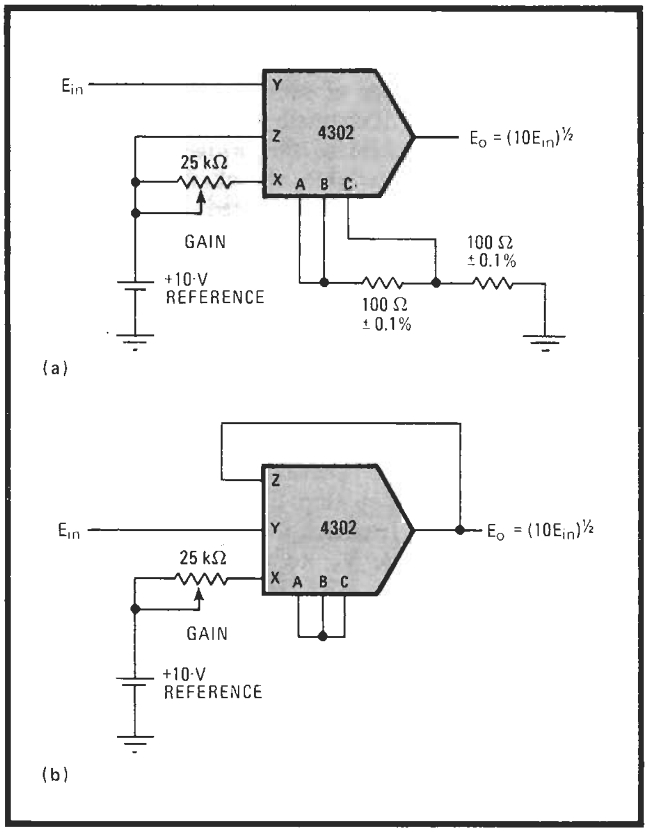
The error sources at low input levels are mainly due to the offset voltages and bias currents of the numerator and denominator input op amps, and not to the logging transistors. Optional trims are usually provided by manufacturers in order to eliminate the offsets and bias currents that are inherent in all op amps.

As with the multifunction converter and the multiplier–inverted divider, the bandwidth of the log-antilog divider decreases almost linearly with divisor voltage level; for example, a 400-kilohertz divider at a 10-V divisor voltage will become a 4-kHz divider at a 100-microvolt divisor voltage. By rearranging the four logging transistors and the four op amps and replacing the voltage reference by a current reference, a log-antilog divider whose bandwidth remains constant at high level even with decreasing divisor voltages can be realized (Fig. 5b).

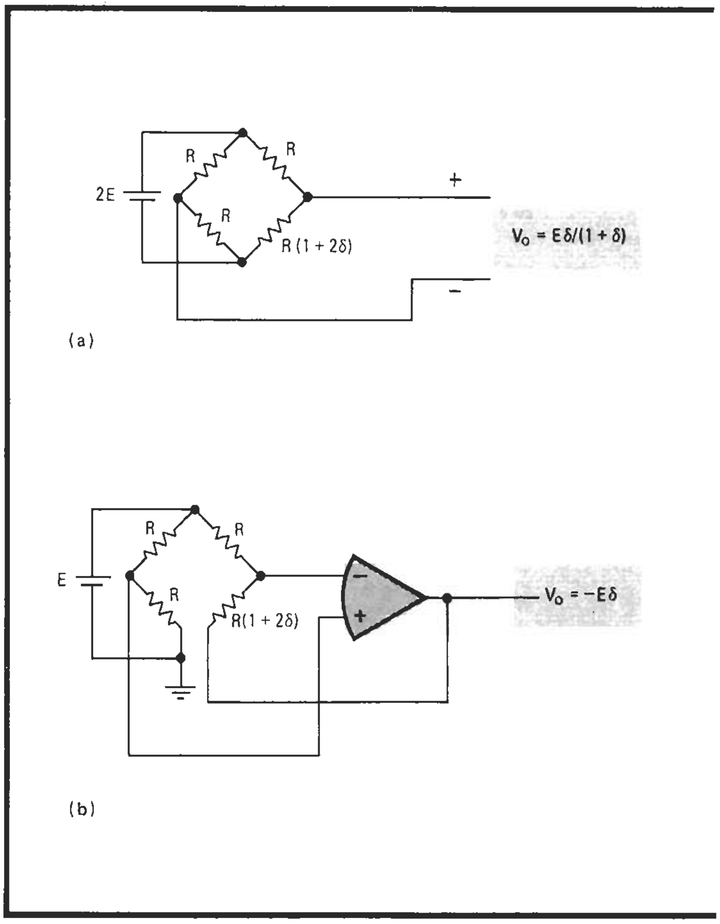
Notice that the current through the output stage (Q_3, Q_4, and A_4) is determined by the reference current, I, and remains constant. If I is set high, the divider's bandwidth will stay fairly flat from a 10-V divisor voltage down to 100 mV and then start to drop gradually, at a much slower rate than the circuit in Fig. 5a. Using 741-type op amps and setting I equal to 200 μA, typical component values are:

R_1 = 50 kilohms	R_4 = 10 kΩ
R_2 = 10 kΩ	R_5 = 100 kΩ
R_3 = 33 kΩ	C = 33 picofarads

As mentioned before, the offset voltages and bias currents of the op amps should be nulled out for low-signal operations. Unfortunately, with a reference current in place of a reference voltage, this divider circuit cannot be readily used as a three-input multiplier-divider to perform $E_o = XY/Z$.



7. Another approach. The multifunction converter (here, the 4302) may also be used as a square-rooter. To implement the transfer function, $E_o = X(YZ)^m$, m can be set equal to ½ (a) or to 1 (b), with the other connections appropriately made.



8. Linearizing bridges. The Wheatstone bridge has a nonlinear output dependent on the input variable (a). It may be linearized by using an op amp in a feedback loop (b), a multiplier–inverted divider (c), or an MID with an instrumentation amplifier (d).

One application of a precision divider is computing the square root of an input signal, often required in process-control systems. If the divisor's input is connected to its output terminal, the divider's transfer function, that is, $E_o = 10N/D$, becomes:

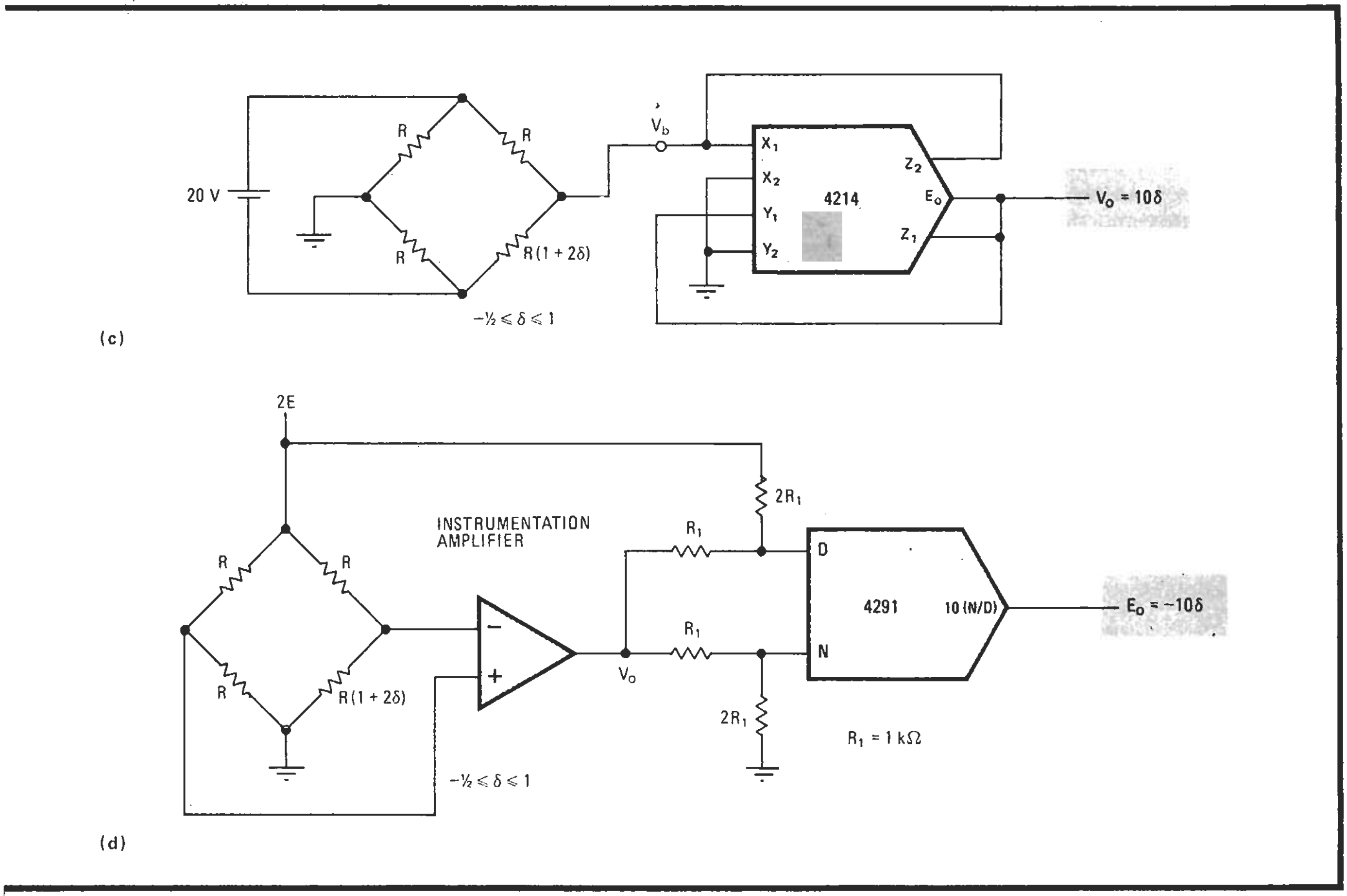
$$E_o = 10(N/E_o) = (10N)^{1/2}$$

The output is now proportional to the square root of the input, N.

Square-rooters

Square-rooters employing a multiplier–inverted divider and a log-antilog divider are shown in Fig. 6a and 6b, respectively. Since E_o is always unipolar, adding a diode at the output of the divider will help prevent the square-rooter from saturating to the opposite supply voltage, which is occasionally caused by power-supply transients. In Fig. 6b, a 1-megohm output load may be necessary to turn on the diode, because the input impedances of the divider are so high (about 10 MΩ) that, without the load, practically no current will flow through the diode.

The square-rooter's accuracy is strictly dependent upon the accuracy of the divider employed. With a multiplier–inverted divider, the accuracy is poor at low input voltages. The error-versus-signal voltage can be estimated from:



$$E_o = (10E_{in} + 10\epsilon_m)^{1/2}$$

where E_o and E_{in} are the square-rooter's output and input voltages, respectively, and ϵ_m is the multiplier error specified by the manufacturer. For example, for a 0.5% multiplier, ϵ_m = 50 mV maximum, and therefore the square-rooter's error would be 25 mV maximum at E_{in} = 10 V, but would be 109 mV maximum at E_{in} = 500 mV.

Figure 6c compares the typical error curves of square-rooters built with a multiplier–inverted divider and those made with a log-antilog divider. Typical errors would normally be much lower than in the graph. As can be seen, if small-signal accuracy is critical, a precision divider like the log-antilog type should be used.

With an external voltage reference, a multifunction converter may also be used to build a square-rooter. There are two ways to implement this function. The straightforward method is to set m = ½ with two matched resistors and connect X and Z to a 10-V reference (Fig. 7a). Then the output voltage becomes:

$$E_o = 10(Y/10)^{1/2} = (10Y)^{1/2}$$

Alternatively, m can be set to 1 as in Fig. 7b and the X input connected to a +10-V reference. By shorting Z to the output, E_o, the transfer function becomes:

$$E_o = 10(Y/E_o) = (10Y)^{1/2}$$

The accuracy of this square-rooter is about equal to that of a log-antilog divider.

The familiar Wheatstone bridge is widely used in measuring the resistance of sensors like strain gauges, pressure transducers, thermistors, and servo motors. Unfortunately, the output of the bridge is a nonlinear function of the input variable, the change in the resistance being measured. As illustrated in Fig. 8a, the output voltage, V_o, is related to the input variable, δ, by:

$$V_o = E\delta/(1 + \delta)$$

where 2E is the bridge supply voltage.

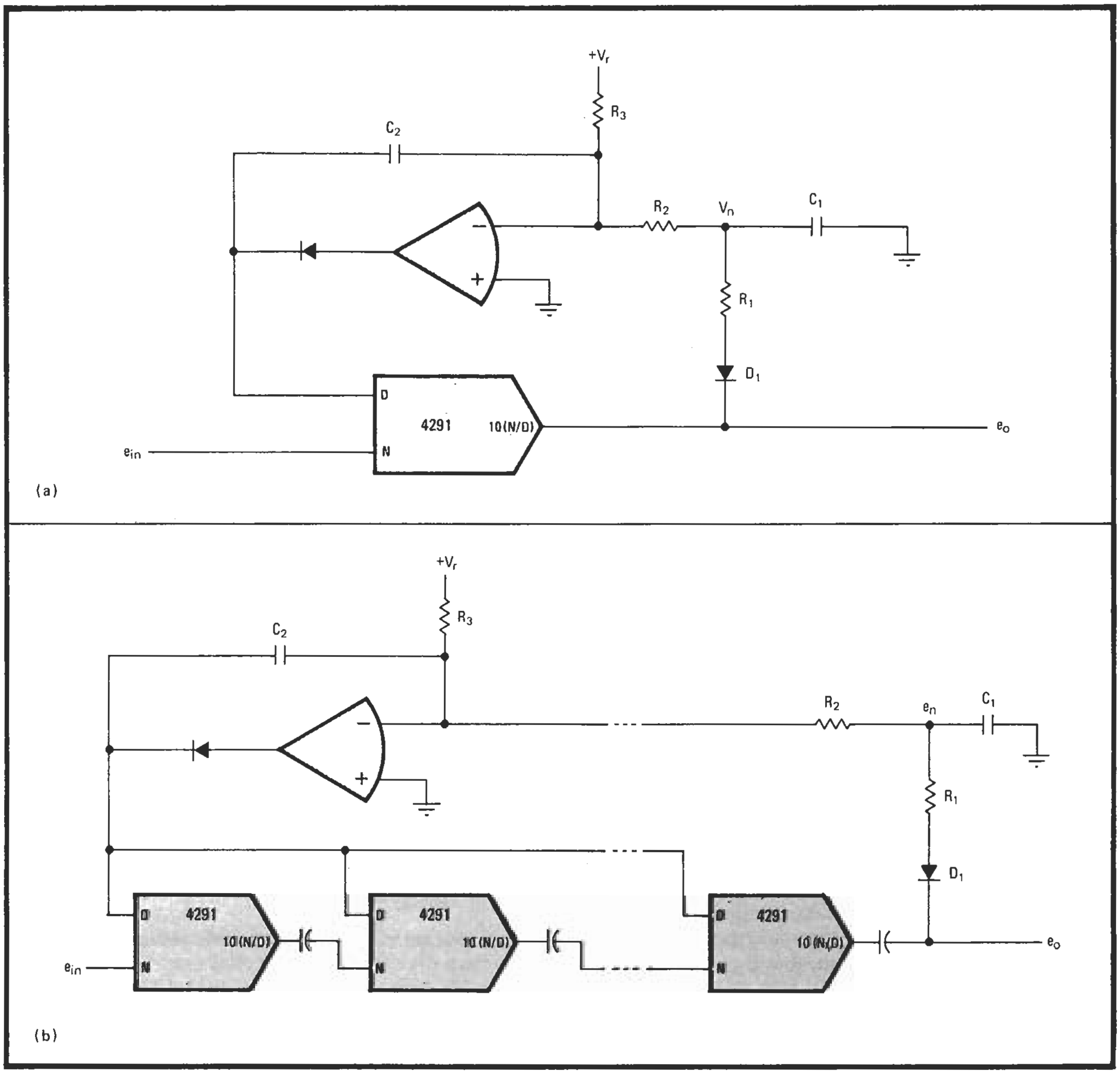
Linearizing the bridge

Because direct measurement and manipulation of nonlinear data is often undesirable, a circuit is needed to first linearize the bridge function. The simplest method of linearization uses an op amp. Connecting the variable-resistance arm in the feedback loop (Fig. 8b) causes the output of the op amp, V_o, to vary linearly with the variable, δ. Thus, $V_o = -E\delta$. However, some inexpensive bridges are packaged in four-terminal boxes and therefore will not work with this method, which requires five terminals.

A low-cost multiplier–inverted divider with differential Z inputs can, however, implement the inverse of the bridge function and linearize it. In Fig. 8c, the output voltage of the bridge, V_b, is given by:

$$V_b = 10\delta/(1 + \delta)$$

and the multiplier–inverted divider provides the transfer function, $V_o = 10V_b/(10 - V_b)$. The series connection of these two nonlinear circuits results in a linear function,



9. Controlling gain. The bandwidth of a 40-dB automatic-gain-control circuit using the 4291 (a), will be increased by n times, or its tracking range by n times, by cascading n dividers in the feedback loop shown in (b). The 400-kHz bandwidth, however, cannot be exceeded.

that is, $V_o = 10\delta$.

If the bridge supply voltage is single-ended, rather than floating as in Fig. 8c, an instrumentation amplifier is needed to convert the two output terminals of the bridge to a single output. The amplifier can be used effectively to compensate for bridge voltage variations. By inverting the signal such that $V_o = -E\,\delta/(1+\delta)$ and using four resistors to sum the bridge voltage with, and divide it by, V_o (Fig. 8d), the divider's denominator and numerator voltage become:

$$D = \frac{2R_{iD}}{3R_{iD}+2R_1}(E+V_o)$$

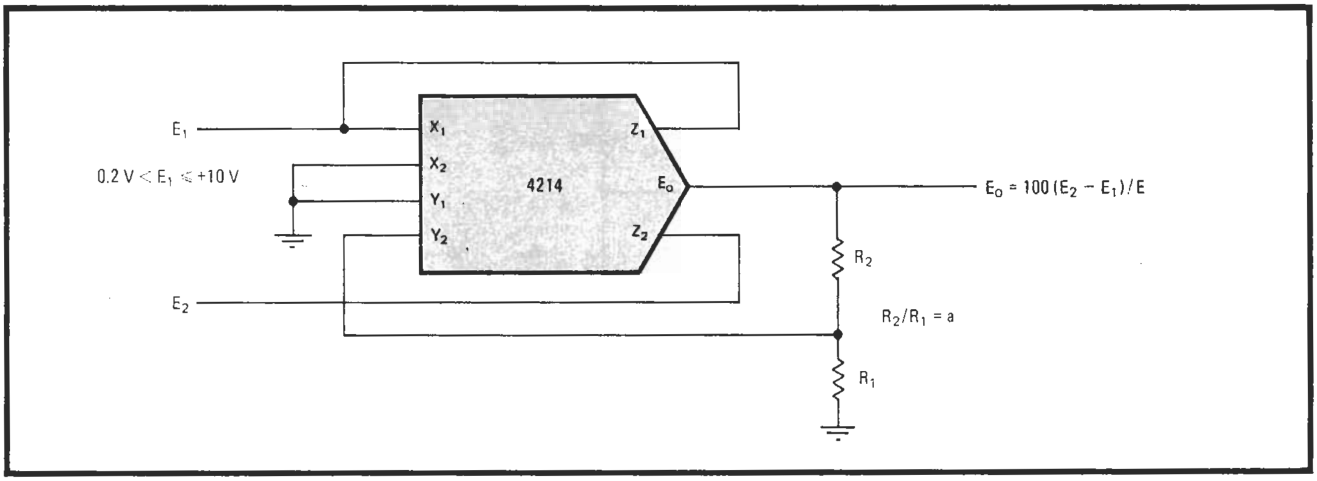
$$N = \frac{2R_{iN}}{3R_{iN}+2R_1}V_o$$

respectively, where R_{iD} is the input impedance of the divider's denominator input and R_{iN} is that of the numerator input.

The cleverness of this circuit becomes clear when D and N are substituted into the divider's simple transfer function, $E_o = 10N/D$. The bridge voltage, 2E, and the input impedance, R_i ($= R_{iN} = R_{iD}$), cancel out, resulting in $E_o = -10\delta$. Therefore, the output is independent of the bridge supply voltage. When R_1 is much smaller than R_i, the circuit is insensitive to the value of R_i.

Controlling the gain automatically

To compensate for amplitude fluctuations of any given signal, nothing less than a well-designed automatic-gain-control circuit will do. A good AGC circuit is one that can keep the output constant over a wide dynamic range of



10. A natural application. Direct readouts in percentage of such parameters as efficiency, distortion, gain/loss, and error are easily obtained by connecting a 4214 in a configuration that provides an output of 1 V = 1% with deviations measured up to ±10%.

input signal levels (tracking range). Analog dividers are excellent candidates for such appplications.

The tracking range of an AGC is directly related to the denominator's operating range of the divider employed. For example, if a divider has a divisor operating range from 10 V down to 100 mV, the AGC circuit associated with it will track ac signals over a 40-decibel range.

Figure 9a shows an AGC circuit using a two-quadrant log-antilog divider. The divider serves as a voltage-controlled amplifier whose output increases with a decrease in divisor voltage. Diode D_1 rectifies the output voltage, e_o. Low-pass filter R_1–C_1 produces a negative voltage, V_n, proportional to the negative peak of e_o. The integrator, comparing V_n with a positive reference voltage, V_r, determines the divisor voltage of the divider.

Automatic gain control is achieved thus: as the input signal, e_{in}, increases, e_o tends to increase, pushing V_n further negative. This increases the integrator's output voltage, which is connected to supply the divisor's input. As the divisor voltage goes up, it will pull e_o back down until it reaches an equilibrium.

Typical values for audio applications are:

$R_1 = R_2/10 = R_3/10 = 1\ k\Omega$
$C_1 = 10C_2 = 10\ \mu F$
$V_r = 0.3\ V$

These values will provide a 2-V peak-to-peak output amplitude, which can be reset by adjusting either R_3 or V_r. For subaudio frequencies, an increase in the values of both C_1 and C_2 is necessary.

The upper frequency limit is determined by the bandwidth of the divider, and the bandwidth of most dividers decreases with decreases in divisor voltage. This means that the bandwidth of the AGC will decrease with input signal voltages.

As an example, with a divider's 3-dB bandwidth specified for 400 kHz at a 10-V divisor voltage, the bandwidth will be, as a rule of thumb, 40 kHz at a 1-V divisor voltage and 4 kHz at a 100-mV divisor voltage. With these specifications, a 40-dB AGC circuit will operate over a 40-dB signal range only up to 4 kHz and over a 20-dB range up to 40 kHz. Although it can function up to 400 kHz, the circuit will have no practical tracking range at those frequencies.

The bandwidth of the AGC circuit can be expanded by cascading two or more dividers in the feedback loop (Fig. 9b). For the 40-dB example, each of the dividers operates actually over 40/n dB, where n is the number of dividers cascaded. The bandwidth of the AGC circuit will thus be increased by n times, but will, of course, never exceed the divider's maximum bandwidth of 40 kHz. The cascading technique will also increase the tracking range of the AGC circuit by n times; that is, two 40-dB dividers will yield an 80-dB AGC circuit. Ac coupling at the output of each divider is recommended to eliminate unwanted divider offset voltages.

Taking ratios

For ratiometric applications, a divider naturally comes to mind. Percentage measurement:

$$E_o = 100(E_2 - E_1)/E_1$$

is just another version of ratiometric measurement, but it requires a divider with differential numerator inputs and adjustable gain (from the nominal 10 to 100). With low-cost IC dividers, it is practical to provide direct readout in percentages of such parameters as efficiency, distortion, gain/loss, error, and so on.

Figure 10 shows a percentage measurement circuit employing a differential multiplier–inverted divider. The circuit, which provides 1 V = 1%, is capable of measuring ±10% deviations. Wider deviations can be measured by decreasing the ratio of R_2/R_1, and narrower variations by increasing the ratio. If the dynamic range of E_1 is too wide for a multiplier–inverted divider to handle, a log-antilog divider may be employed, but an extra operational amplifier will be needed to take the difference $E_2 - E_1$.

The percentage circuit can also be used to sort components by first converting the component's parameter into a voltage and comparing it with a reference. A comparator at the output of the percentage circuit may then be set to separate units beyond a specified limit. □

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