WORKING WITH OTA's

How to use operational transconductance amplifiers in your designs and projects.

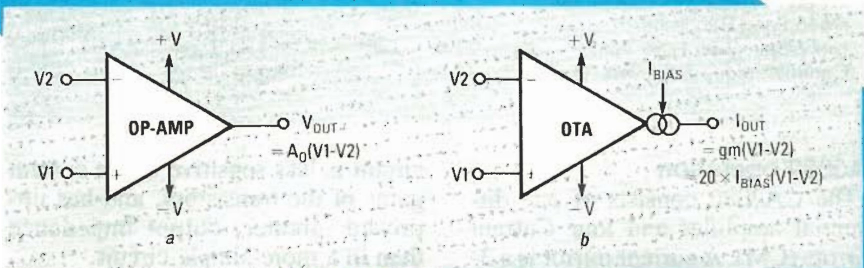


FIG. 1—A CONVENTIONAL OP-AMP (*a*) is a fixed-gain voltage-amplifying device, whereas an OTA (*b*) is a variable-gain voltage-to-current amplifier..

RAY MARSTON

THERE ARE MANY DIFFERENT TYPES OF OPerational amplifiers in use today, but an *O*perational *T*ransconductance *A*mplifier, or OTA, is one that you may not be familiar with. This month we'll introduce you to the CA3080 OTA.

Op-amps and OTA's

Conventional op-amps are essentially voltage-amplifying devices. As figure 1-*a* shows, a conventional op-amp has differential input terminals and produces an output voltage of $A_O \times (V1 - V2)$, where A_O is the open-loop voltage gain (that gain is typically 100,000), V1 is the signal voltage at the non-inverting input, and V2 is the signal voltage at the inverting input. Also, a conventional op-amp has a fixed open-loop voltage gain, a high input impedance, and a low output impedance.

Like a standard op-amp, an OTA has differential input terminals, but, as shown in figure 1-*b* it is a voltage-to-current amplifier, as indicated by the constant-current symbol at its output. The input voltages produce an output in the form of a high-impedance current with a value of $gm \times (V1 - V2)$, where gm is the transconductance in mhos, or the *voltage-to-current gain* of the device. The transconductance is directly proportional to an external bias current (I_{BIAS}) fed into the amplifier's bias input. In an OTA, that current can be varied from 0.1 μA to 1 mA, providing a 10,000:1 gain-control range.

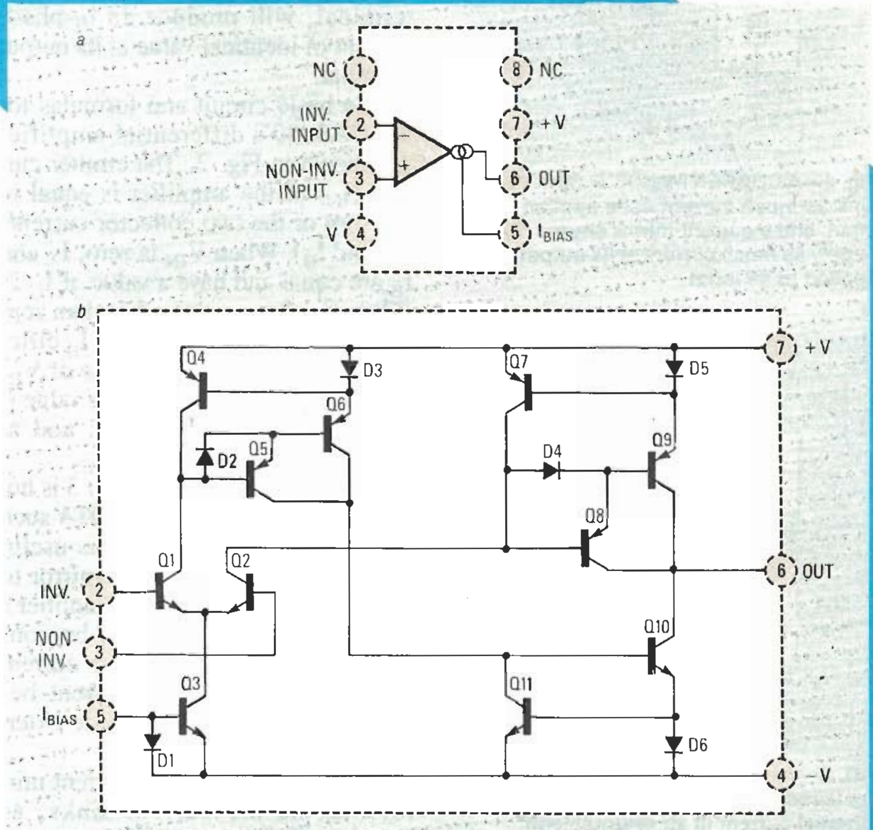


Fig. 2—THE PIN CONNECTIONS of the 8-pin dip version of the CA3080 are shown in (*a*), and the internal circuitry is shown in (*b*).

An OTA can be made to operate like a conventional voltage-amplifying op-amp by connecting a suitable load resistance to its output terminal so that its output current is converted to a proportional voltage. The total current consumed by an OTA is double the value of I_{BIAS}, which may be as low as 0.1 μA. That means that the device can be used in true micropower applications. The amount of I_{BIAS} can be controlled easily using an external voltage and a series resistor. An OTA can be used as a *V*oltage-*C*ontrolled *A*mplifier (VCA), *V*oltage *C*ontrolled *O*scillator (VCO), or *V*oltage *C*ontrolled *F*ilter (VCF).

One of the best known OTA's is the CA3080. Figure 2-*a* shows its pin connections, and Fig. 2-*b* shows its internal circuitry. Table 1 lists the basic parameters of the device.

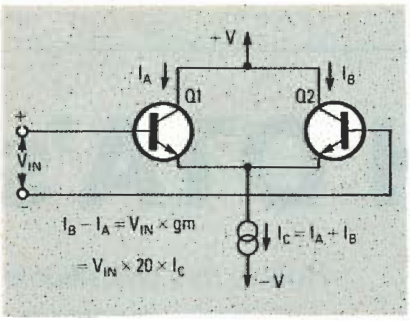


FIG. 3—A DIFFERENTIAL AMPLIFIER is at the heart of the CA3080.

TABLE 1

CHARACTERISTIC	LIMITS
Supply Voltage Range	+4V to +30V DC or ±2V to ±15V
Max Differential Input Voltage	±15V
Power Dissipation	125 mW MAX
Input Signal Current	1 mA MAX
Amplifier Bias Current	2 mA MAX
Output Short-Circuit Duration	INDEFINITE
Forward Transconductance, gm	9600 μmho typ
Open Loop Bandwidth	2 MHz
Unity-Gain Slew Rate	50 V/μs
Common-Mode Rejection Ratio	110 dB typ

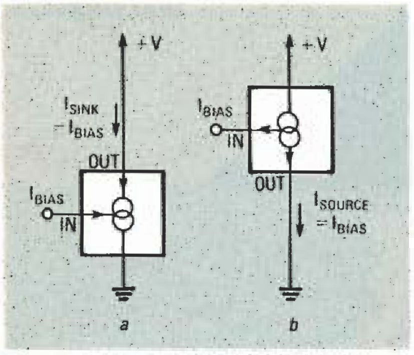


Fig. 4—A CURRENT-MIRROR SINK (*a*) will sink as much current as is applied to its input, and a current-mirror source (*b*) will supply as much current at its output, as is applied to its input.

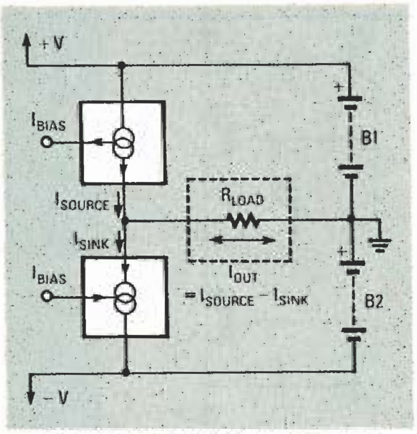


FIG. 5—WHEN TWO CURRENT MIRRORS are wired as shown, they generate a differential current in an external load.

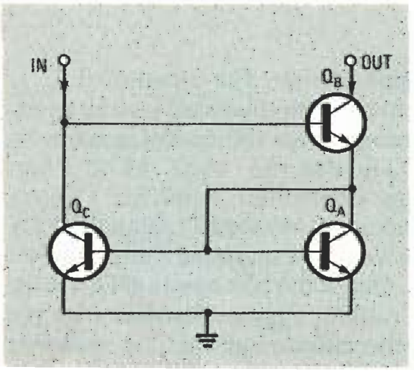


FIG. 6—A SINK-TYPE CURRENT MIRROR is made up of the circuitry shown here.

CA3080 operation

The CA3080 consists of one differential amplifier and four *C*urrent *M*irrors (CM). A current mirror is a 3-terminal circuit that, when an external bias current is provided at its input terminal, will produce an in-phase current of identical value at its output terminal.

The basic circuit and formulas for the CA3080's differential amplifier are shown in Fig. 3. The emitter current (I_C) of the amplifier is equal to the sum of the two collector currents (I_A and I_B). When V_{IN} is zero, I_A and I_B are equal and have a value of $I_C/2$. When V_{IN} has a value other than zero (±25 mV maximum), I_A and I_B differ and produce an $I_B - I_A$ value of $V_{IN} \times$ gm. The transconductance value is directly proportional to I_C, and at 25°C roughly equals $20 \times I_C$.

By itself, the circuit in Fig. 3 is not very useful. However, in an OTA such as the CA3080, the circuit is useful because by using a current mirror to externally control I_C, the amplifier's transconductance can then be controlled. By using three more current mirrors, the difference current between I_A and I_B can be made externally available.

There are two types of current mirrors. Some are current sinks, as shown in Fig. 4-*a*, and others are current sources, as shown in Fig. 4-*b*. When a current mirror source and current mirror sink are connected as shown in Fig. 5, and powered from a bi-polar power supply, they generate a differential current ($I_{SOURCE} - I_{SINK}$) in any load that is connected between the junction point and the circuit ground.

Figure 6 shows the actual circuit of a sinking-type current mirror. Transistor Q_A, which operates like a diode, is wired across the base-emitter junction of a second, closely-matched transistor. The mirror accuracy of that circuit is less sensitive to the current gains of the transistors, and has improved (greater) output impedance than in a more-simple circuit.

Figure 7 shows how the differential amplifier and four current mirrors are connected in the CA3080 to make a practical OTA. Bias current (I_{BIAS}) controls the emitter current, and thus the transconductance of the Q1/Q2 differential amplifier via current-mirror C. The collector current of Q1 is mirrored by current-mirror A and fed to the bias terminal of current-mirror D, and the collector current of Q2 is mirrored by current-mirror B and fed to the sink terminal of current-mirror D, so that the externally available output current is equal to $I_B - I_A$.

If you refer back to Fig. 2-*b*, you will notice that Q1 and Q2 form the differential amplifier, D1 and Q3 make up current-mirror C, and current-mirror D is comprised of D6, Q10, and Q11. Current-mirror A (Q4–Q6, D2, and D3), and current-mirror B (Q7–Q9, D4, and D5) are slightly more complex than the others, using Darlington pairs of transistors and speed-up diodes to improve their performance.

Some finer points

All of the major operating parameters of the CA3080 are adjustable and depend on the value of I_{BIAS}. The maximum output current is equal to I_{BIAS}, and the total operating current of the IC is double the I_{BIAS} value. The input bias currents drawn by pins 2 and 3 when the IC is operating in the linear mode are each equal to approximately $I_{BIAS}/200$, with the actual values depending on the current gains of Q1 and Q2 within the chip.

The transconductance (Fig. 8-*a*) and the input and output impedances (Fig. 8-*b*) vary with I_{BIAS}. Figure 8 shows typical parameter values when the IC is driven from a bi-polar 15-volt

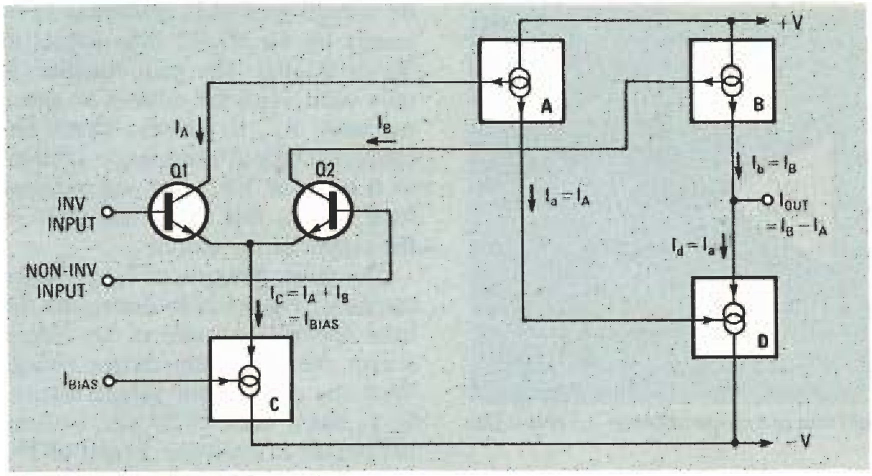


FIG. 7—THE CA3080 IS COMPRISED of one differential amplifier and four current mirrors.

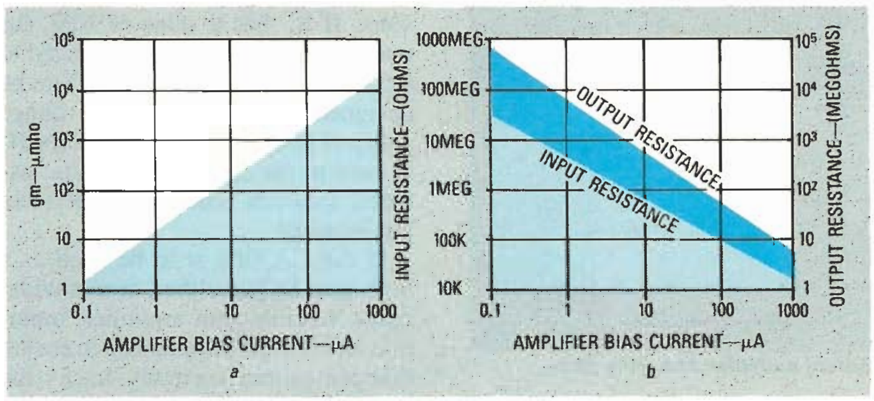


FIG. 8—THE TRANSCONDUCTANCE (*a*) and the input and output resistances (*b*) of the CA3080 vary with the bias current.

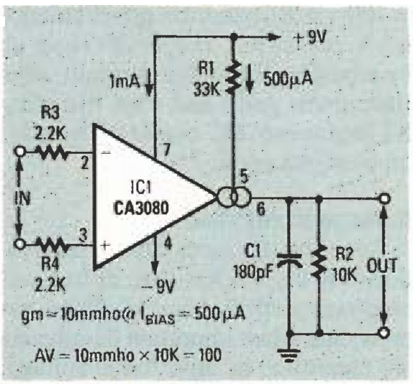


FIG. 9—THIS DIFFERENTIAL AMPLIFIER has 40-dB voltage gain.

supply at an ambient temperature of 25°C. Therefore, at a bias current of 10µA, gm is typically 200 µmho, with an input resistance of 800K and an output resistance of 700 megohms. At 1-mA bias current, those values change to 20 mmho, 15K, and 7 megohms respectively.

The output voltage of the IC depends on the values of I_{BIAS} and an external load resistor connected to the output (pin 6) of the device. If the load impedance is infinite, the output can swing to within 1.5 volts of the positive supply and within 0.5 volt of the negative supply. If the impedance is not infinite, the peak output swing is limited to $I_{BIAS} \times R_L$. Thus at a 10-µA bias with a 100K load, the output swing is 1 volt.

The slew rate and bandwidth of the IC depend on the value of I_{BIAS} and any external loading capacitor connected to pin 6. The slew-rate value,

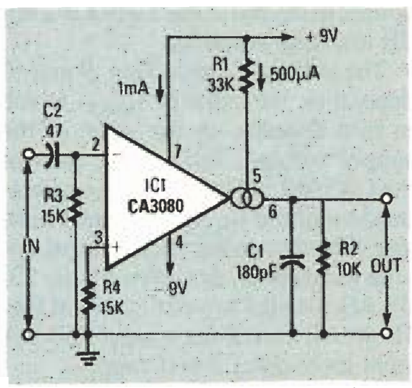


FIG. 10—AN AC-COUPLED 40-dB inverting amplifier.

in V/µs, equals I_{BIAS}/C_L, where C_L is the loading capacitance value in pF, and I_{BIAS} is in µA. With no external loading capacitor connected, the maximum slew rate of the CA3080 is about 50V/µs.

Basic circuits

The CA3080 is very easy to use. Its I_{BIAS} terminal (pin 5) is internally connected to the negative supply (pin 4) by a base-emitter junction, so the biased voltage of the terminal is about 600 mV above that of pin 4. I_{BIAS} can be obtained by connecting pin 5 to either the ground line or the positive supply via a current-limiting resistor of suitable value.

Figures 9 and 10 show two ways of using the CA3080 as a linear amplifier with a voltage gain of about 40 dB. The circuit in Fig. 9 is a direct-coupled differential amplifier, and Fig. 10 shows an AC-coupled inverting amplifier. Both designs operate from bi-polar 9-volt supplies, so 17.4 volts is generated across bias-resistor R1, which feeds about 500 µA into pin 5 causing each IC to draw another 1 mA from their supply.

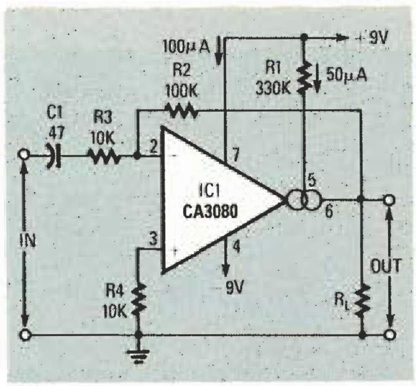


FIG. 11—THIS 20-dB MICRO-POWER inverting amplifier consumes very little power.

At a bias current of 500 µA, the transconductance of the CA3080 is approximately 10 mmho. The outputs of Figs. 9 and 10 are loaded by a 10K resistor (R2), and therefore provide an overall voltage gain of 10 mmho × 10K = 100, or 40 dB. The peak current that can flow into the 10K load is 500 µA (equal to I_{BIAS}), so the peak output is 5 volts. The output is also loaded by a 180-pF capacitor (C1), giving the circuit a slew-rate limit of 500 µA/180pF = 2.8V/µs. The output impedance of each circuit equals the R2 value of 10K. Note that in those two circuits the IC is used in the

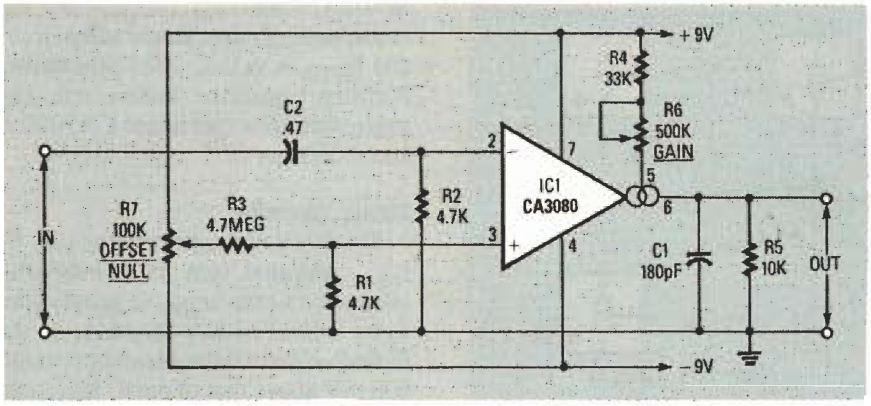


FIG. 12—A VARIABLE-GAIN AC amplifier has a gain that can range between × 5 and × 100.

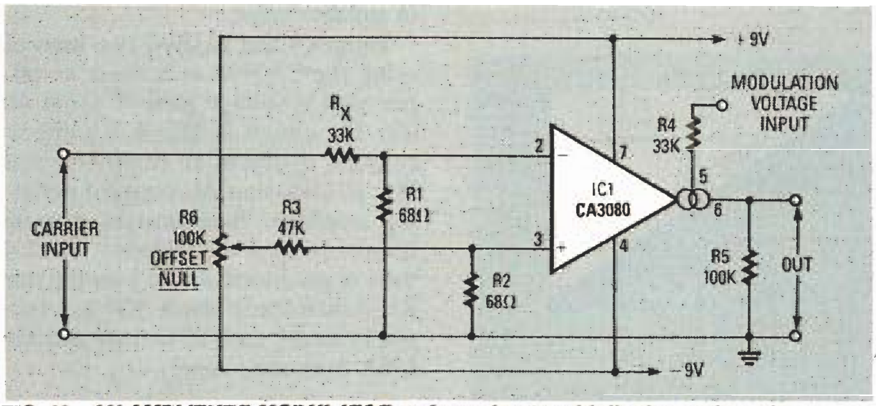


FIG. 13—AN AMPLITUDE MODULATOR or 2-quadrant multiplier has unity gain.

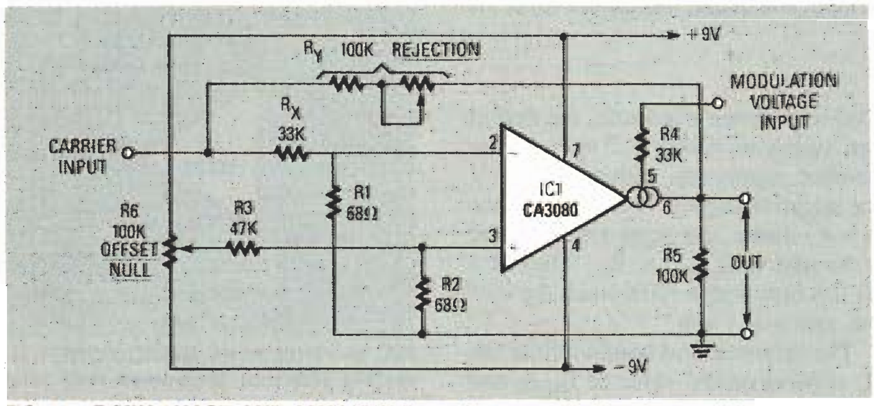


FIG. 14—BOTH PHASE AND AMPLITUDE can be controlled by the modulation signal in this ring modulator or 4-quadrant multiplier.

open-loop mode, and that if the slew rate is not externally limited by C1, the IC will operate at its maximum bandwidth and slew rate. Under those conditions the CA3080 may be excessively noisy.

In the circuit in Fig. 9, the differential inputs are applied via series resistors R3 and R4, which help equalize the source impedances of the two signals and maintain the DC balance of the IC. The circuit in Fig. 10 has both inputs tied to ground via 15K resistors and the input signal applied to one terminal only. With the input connected to pin 2, the circuit is a 40-dB inverting amplifier.

The voltage gains in Figs. 9 and 10 depend on the value of I_{BIAS}, which in turn depends on the value of the supply voltage. The voltage gain of the CA3080 can be made almost independent of the I_{BIAS} and supply-voltage values by using conventional op-amp techniques, as shown by the 20-dB AC-coupled inverter circuit of Fig. 11, which consumes a mere 150 μA from its bi-polar 9-volt supply.

The circuit in Fig. 11 is wired like a conventional inverting amplifier, with its voltage gain (A_V) determined primarily by the R2/R3 ratio (equal to 10, or 20-dB). The gain equation is only valid when the value of an external load, R_L, is infinite. That's because the output impedance is equal to $R2/A_V$, or 10K, and any external load lessens that value and reduces the output of the circuit.

The main function of I_{BIAS} in the circuit of Fig. 11 is to determine the total operating current of the circuit and/or the maximum output swing. With the component values shown, I_{BIAS} has a value of 50 μA, causing the circuit to consume a total of 150 μA. When R_L is infinite, the output is loaded only by R2, which has a value of 100K, so the maximum output is 5 volts. If R_L has a value of 10K, the maximum output voltage is limited to 0.5 volt. That circuit can therefore be designed to have any desired voltage gain and peak output, and since the IC is used in the closed-loop mode, external slew-rate limiting is therefore not required.

If the CA3080 is to be used as a high-gain DC amplifier, or as a wide-range variable-gain amplifier, input-bias levels must be balanced to ensure that the output correctly tracks the input signals at all values of I_{BIAS}. Figure 12 shows how to bias an inverting AC amplifier in which the voltage gain is variable from roughly × 5 to × 100 via R6, and the offset balance is pre-set via R7. The circuit is set up by adjusting R6 to its minimum value (maximum gain) and then trimming R7 to give zero DC output with no AC input signal applied.

Voltage-controlled gain

Some of the most useful applications for the CA3080 are in true micropower amplifier and oscillator circuits, and when important parameters are controlled by an external voltage. In the latter category, one major application is as a VCA or amplitude modulator, in which a carrier signal is fed to the input of the amplifier, and the output amplitude is controlled or modulated by another signal fed to the I_{BIAS} terminal. Figure 13 shows a practical version of such a circuit.

The circuit in Fig. 13 is a variable-gain inverting amplifier. Input-bias resistors R1 and R2 have low values to minimize the noise levels of the IC and eliminate the need for external slew-rate limiting. Offset biasing is applied to the non-inverting input via

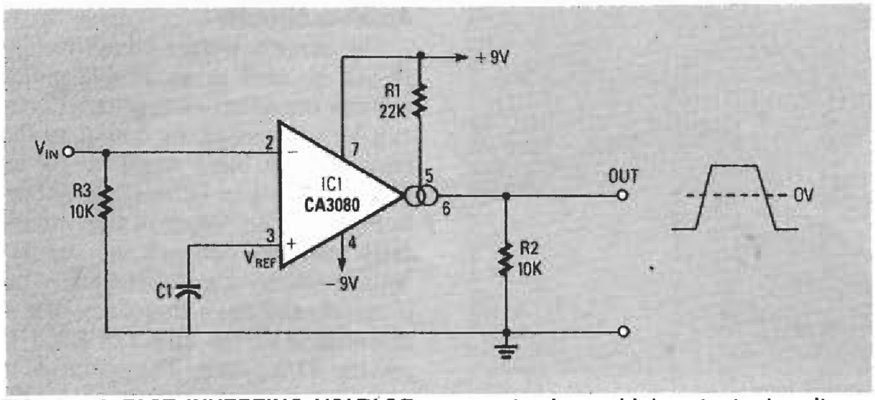


FIG. 15—A FAST INVERTING VOLTAGE comparator has a high output when its input falls below V_{REF}.

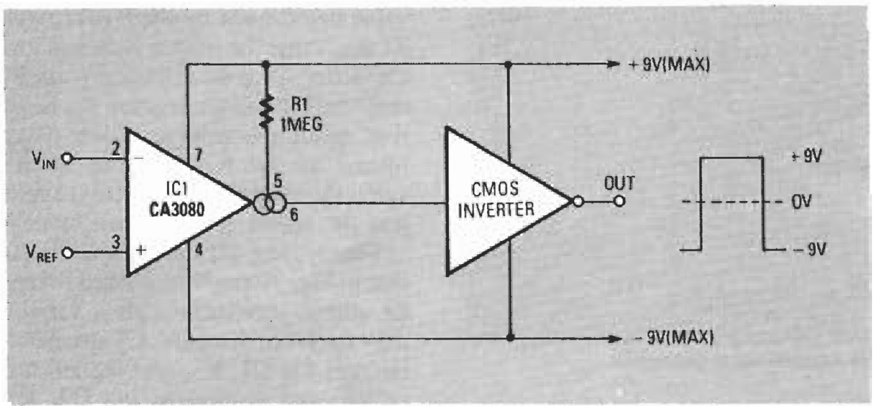


FIG. 16—A NON-INVERTING micropower voltage comparator has inputs that are sensitive to small changes.

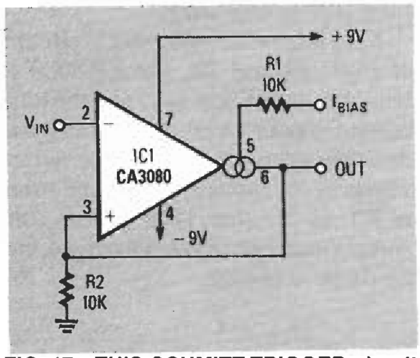


FIG. 17—THIS SCHMITT-TRIGGER circuit has programmable trigger-thresholds and peak-output.

R3/R6. The carrier signal is applied to the inverting pin of the CA3080 by the voltage divider Rx/R1. When R_X has a value of 33K as shown, and the modulation input terminal is tied to ground, the circuit basically has unity gain. The gain doubles when the modulation terminal is tied to +9 volts (+V), and when the modulation terminal is tied to −9 volts (−V) the circuit has roughly 80 dB of signal rejection.

The instantaneous polarity of the output signal of the circuit in Fig. 13 is determined entirely by the instantaneous polarity of the input signal. The amplitude of the output signal is determined by the product of the input and the gain-control values. That type of circuit is known as a 2-quadrant multiplier.

Figure 14 shows how the circuit in Fig. 13 can be modified so that it can be used as a ring-modulator or 4-quadrant multiplier, in which the output-signal polarity depends on the polarities of both the input signal and the modulation signal.

The circuits of Figs. 13 and 14 are identical, except that in Fig. 14, resistor-network R_Y is connected between the input and output terminals. When the modulator input is tied to ground, the inverted signal flowing into R5 from the OTA's output is balanced by the non-inverted signal flowing into R5 from the input signal via R_Y. Therefore, zero volts is generated across R5. If the modulation input goes to +V, the output of the OTA exceeds the current of the R_Y network, and an inverted gain-controlled output is obtained. If the modulation input is −V, on the other hand, the output current of R_Y exceeds that of the OTA, and a non-inverted gain-controlled output is obtained. So, both the phase and the amplitude of the output signal of the 4-quadrant multiplier circuit are controlled by the modulation signal. The circuit can be used as a ring modulator by feeding independent AC signals to the two inputs, or as a frequency doubler by feeding identical sine-wave signals to the two inputs.

Note that with the R_X and R_Y values shown in Fig. 14, the circuit has a voltage gain of 0.5 when the modulation terminal is tied to +V or −V. The gain doubles if the values of R_X and R_Y are halved. Also note that the Fig. 13 and Fig. 14 circuits each have a high output impedance, and that in practice an output buffer must be added between the output terminal and the outside world.

Comparator circuits

The CA3080 can easily be used as a programmable or micropower voltage comparator. Figure 15 shows the basic circuit of a fast, programmable, inverting comparator, in which a reference voltage (V_{REF}) is applied to the non-inverting terminal and the test input is applied to the inverting terminal. The circuit's operation is such that the output is driven high when the test input is below V_{REF}, and is driven low when the test input is above V_{REF}. The circuit can be used as a non-inverting comparator by reversing the input connections of the IC.

With the component values shown in Fig. 15, the I_{BIAS} current is several hundred μA, so the device has a slew rate of about 20V/μs, and operates as a fast comparator. When the test voltage and V_{REF} are almost identical, the IC operates as a linear amplifier with a voltage gain of gm × R2 or about 200. When the two input voltages are significantly different, the output voltage is limited to values determined by the values of I_{BIAS} and R2. In Fig. 15, the output is limited to about 7 volts when R2 has a value of 10K, or about 700 mV when R2 has a value of 1K.

The circuit in Fig. 15 can be modified so that it is an ultra-sensitive micropower comparator, as shown in Fig. 16. That circuit typically consumes only 50 μA but has an output that fully swings between the +V supply and the −V supply, and can provide drive currents of several mA. In Fig. 16 the CA3080 is biased at about 18 μA via R1 but has its output fed to the near-infinite input impedance of a CMOS inverter stage. That

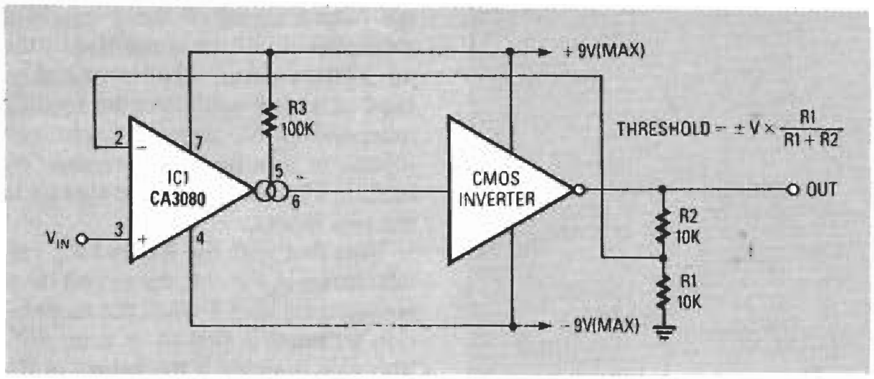


FIG. 18—A MICRO-POWER Schmitt trigger is shown here.

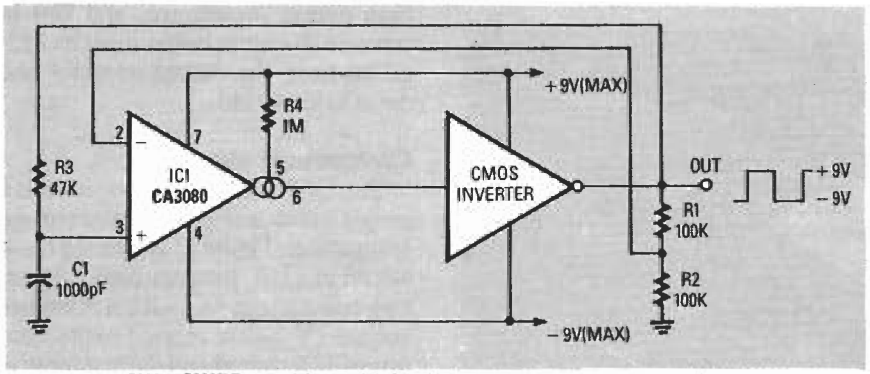


FIG. 19—A LOW-POWER astable multivibrator or square-wave generator.

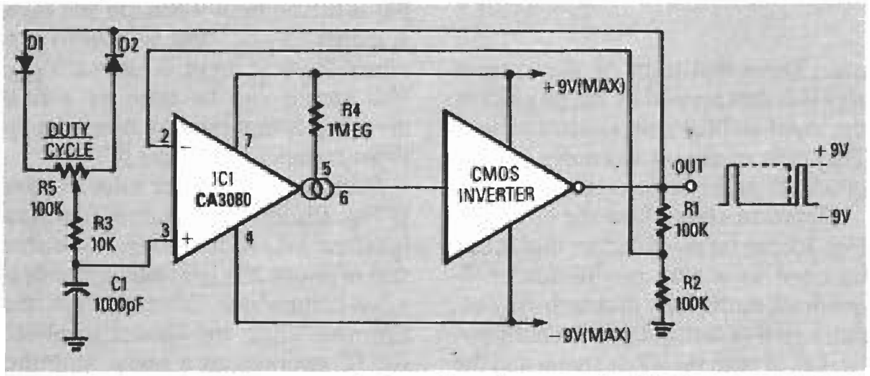


FIG. 20—THIS VARIABLE DUTY-CYCLE oscillator has an output that can be varied from 10:1 to 1:10. D

combination gives the circuit an overall voltage gain of about 130 dB, so that input-voltage changes of only a few μV are enough to switch the output from one supply level to the other.

Schmitt-trigger circuits

The voltage comparator circuit of Fig. 15 can be used as a programmable Schmitt trigger by connecting the non-inverting reference terminal directly to the output of the CA3080, as shown in Fig. 17. In that case, when the output is high, a positive reference value of $I_{BIAS} \times R2$ is generated. When V_{IN} exceeds that value, the output regeneratively switches low and generates a negative reference voltage of $I_{BIAS} \times R2$, and when V_{IN} falls below that value, the output is regeneratively switched high and once more generates a positive reference voltage of $I_{BIAS} \times R2$. Therefore, the trigger thresholds, and also the peak output voltages of the Schmitt circuit, can be precisely controlled or programmed by changing the value of either I_{BIAS} or R2.

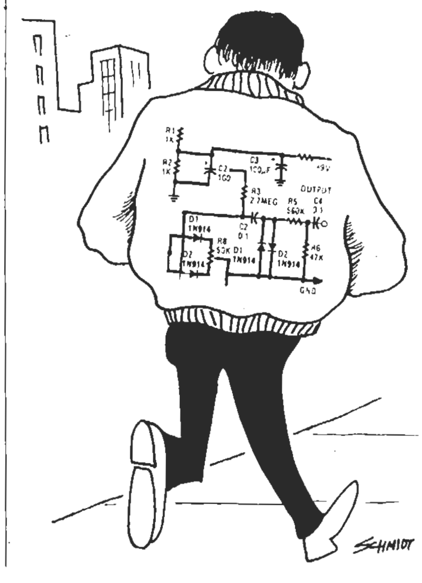
Figure 18 shows an another type of Schmitt trigger, in which the output fully switches between the supply-voltage values. The switching-threshold values are determined by the R1/R2 ratio and the supply-voltage values, and is equal to $+V \times R1/(R1+R2)$.

Astable circuits

The Schmitt-trigger circuit in Fig. 18 can be used as an astable multivibrator or square-wave generator circuit by connecting its output to the non-inverting input terminal via an RC time-constant network, as shown in Fig. 19. The output of that circuit fully switches between the supply-voltage values, is approximately symmetrical, and has a frequency that is determined by the values of R3, C1, and the R1/R2 ratio. The operation is such that, when the output is high, C1 charges via R3 until the C1 voltage reaches the positive reference-voltage value determined by the R1/R2 ratio. At that value the output switches low. Capacitor C1 then discharges via R3 until the C1 voltage reaches the negative reference-voltage value determined by the R1/R2 ratio. at that value the output switches high again, and the whole process then repeats.

Finally, Fig. 20 shows how the circuit in Fig. 19 can be modified to have an output waveform with a variable duty cycle. In that case, C1 alternately charges via D1, R3, and the left half of R5, and discharges via D2, R3, and the right half of R5, to provide a duty-cycle ratio that is fully variable from 10:1 to 1:10 via R5.

Note that in the two astable circuits of Figs. 19 and 20, the CA3080 is biased at only a few μA, and the total current consumption of each design is determined primarily by the series values of R1 and R2, and by the value of R3. In practice, total current consumption of only a few tens of μA can easily be achieved. **R-E**



CIRCUITS

WORKING WITH OTA's

The LM13600 is a second-generation dual Operational Transconductance Amplifier that can be used as a voltage-controlled amplifier, resistor, filter, or oscillator. In this article we'll show you how.

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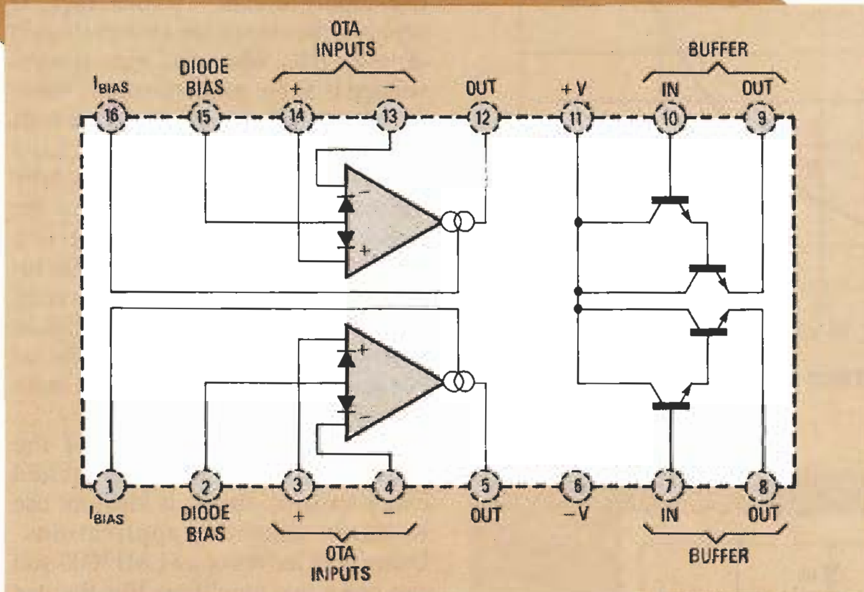

FIG. 1—PIN CONNECTIONS of the LM13600 dual OTA are as shown. You can also see how each IC contains two OTA's.

LAST TIME (MAY 1988) WE DISCUSSED THE basic operating principles of the CA3080 *O*perational *T*ransconductance *A*mplifier (OTA), and showed how it can be used to make various types of voltage- or current-controlled amplifiers and micro-power Schmitt triggers, comparators, and oscillators. However, the CA3080 has relatively high distortion and an unbuffered high-impedance output.

In this article we will introduce you to an improved second-generation OTA; the LM13600. That device is actually a dual OTA, as shown by the pinout of Fig. 1. The package also incorporates linearizing diodes that greatly reduce signal distortion, and a coupled output-buffer stage that can provide a low-impedance output.

The two OTA's of the LM13600 share common supplies, but are otherwise fully independent. All elements are integrated on a single chip, so both OTA's have closely matched characteristics (transconductance values are typically matched within 0.3 dB), making the IC ideal for use in stereo applications. The commercial version of the LM13600 can be powered from a split supply of up to 18 volts, or a single-ended supply of up to 36 volts.

Linearizing diodes

Figure 2 shows a basic connection diagram for one of the OTA's in the LM13600. Basically, that OTA is almost identical to the one in the CA3080, except for the addition of linearizing-diodes D1 and D2. Those are integrated with Q1 and Q2, and have characteristics that are matched to the base-emitter junctions of Q1 and Q2. In use, equal low-value resistors R1 and R2 are wired between the inputs of the differential amplifier and ground. Bias-current I_D is fed to those resistors from the positive supply via R3, D1, and D2. Since D1 and D2 have identical characteristics, and since R1 and R2 have equal values, bias current, I_D, divides equally between R1 and R2.

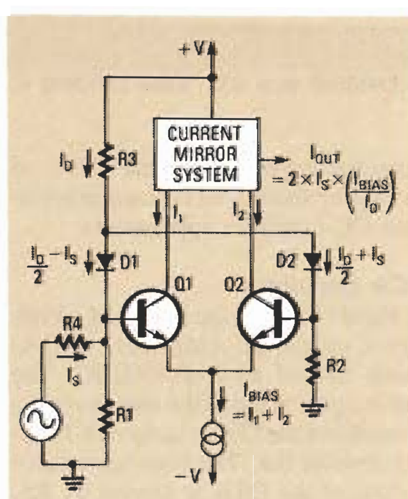

FIG. 2—SIMPLIFIED circuit of an LM13600 OTA shows one of the more basic ways of using the IC.

The circuit's input voltage is applied via R4, which has a large value relative to R1, and generates an input signal current of I_S. That signal current feeds into R1 and generates a voltage across it, reducing D1's current to a value of $(I_D/2) - I_S$. Since I_D is constant, the D2 current rises to $(I_D/2) + I_S$. Consequently, the linearizing diodes apply a heavy negative feedback to the differential amplifier and substantially reduce the signal distortion. If I_S is small relative to I_D, the output current of the circuit is equal to $2 \times I_S \times (I_{BIAS}/I_D)$. The gain of the circuit can be controlled either by I_{BIAS} or I_D. In use, I_D and I_{BIAS} should both be limited to a maximum of 2 mA.

The graph of Fig. 3 shows typical distortion levels of the LM13600 at various peak-to-peak input voltages, with and without the use of the linearizing diodes. At a 30-mV input, the distortion is below 0.03% with the diodes, and 0.7% without them. At a 100-mV input, the distortion is roughly 0.8% with the diodes, and 8% without them.

Controlled-impedance buffers

Figure 4 shows the internal circuit of each half of the LM13600. The two

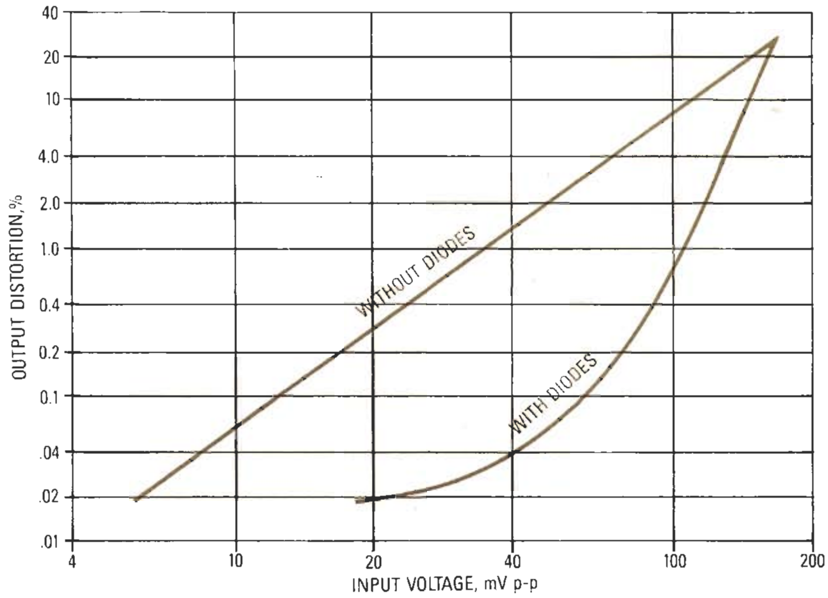

FIG. 3—TYPICAL DISTORTION LEVELS of the LM13600 OTA, with and without the use of the linearizing diodes.

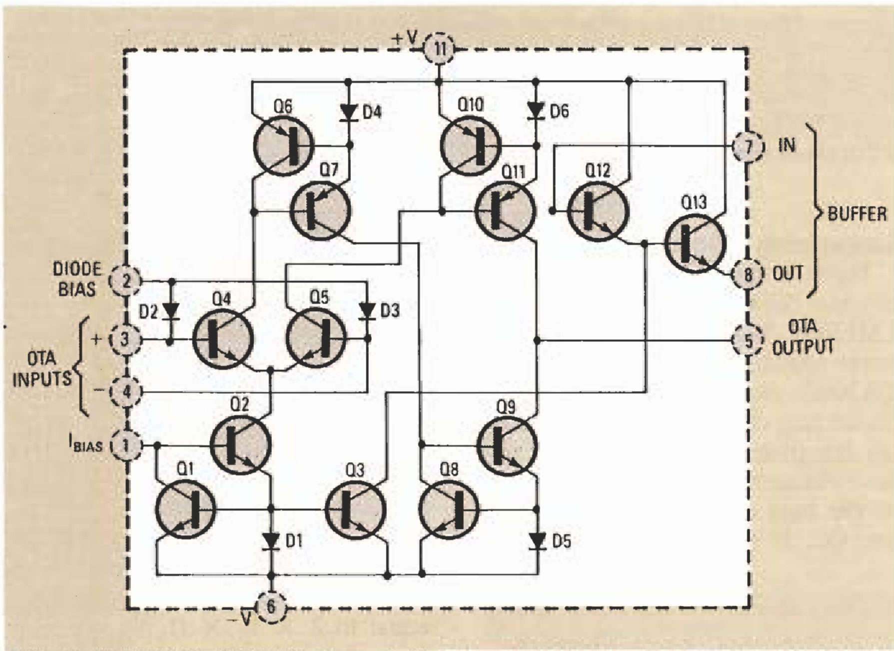

FIG. 4—INTERNAL CIRCUIT of one half of the LM13600 dual OTA. Each LM13600 IC contains two of these circuits.

output transistors (Q12 and Q13) are connected for use as a controlled-impedance Darlington emitter-follower buffer stage. When the base of Q12 is connected to the output of the OTA, and the emitter of Q13 is connected to the negative supply (−V) via a suitable load resistor, the buffer makes the OTA's high-impedance output signal available at a low-impedance level. The output current of each buffer stage should be limited to 20 mA maximum. Note that the output of the buffer stage is about 1.2 volts (two base-emitter voltage drops) below the output-voltage level of the OTA, so the buffer should not be used in precision DC-amplifier applications.

VCA circuits

Figure 5 shows the circuit of a *V*oltage-*C*ontrolled *A*mplifier (VCA) using half of an LM13600 IC. The input signal is fed to the non-inverting terminal of the OTA via current-limiting resistor R4. The high-impedance output of the OTA is loaded by R5, which determines the peak amplitude of the output signal. The output signal is available at a low-impedance level via the buffer stage, which is loaded via R6.

The circuit in Fig. 5 is powered from dual 9-volt supplies. The current I_D is fixed at about 0.8 mA via R1, but I_{BIAS} is variable via R7 and an external gain-control voltage. When the gain-control voltage is at the negative-supply level of −9 volts, I_{BIAS} is zero and the circuit has an overall gain of −80 dB. When the gain-control voltage is at the positive-supply value of +9 volts, I_{BIAS} reaches a value of roughly 0.8 mA, and the circuit has a gain of roughly 1.5. The gain is fully variable within those limits via the gain-control input. The circuit is a non-inverting amplifier, since the input signal is fed to the non-inverting input of the OTA. It can be used as an inverting amplifier by feeding the input signal to the OTA inverting input instead.

Because the two halves of the LM13600 have closely matched characteristics, the IC is ideal for use in stereo-amplifier applications. Using both halves of an LM13600 you can make two amplifiers like the one in Fig. 5. Then, if you connect both gain-control inputs together, and feed them from a single gain-control voltage and current-limiting resistor, you'll have a voltage-controlled stereo amplifier.

The VCA circuit of Fig. 5 can be used as an amplitude modulator or 2-quadrant multiplier by simply feeding the carrier signal to the OTA input, and the modulation signal to the gain-control input. If desired, the gain-control pin can be DC-biased so that a carrier output is available while no AC-input signal is applied. Figure 6 shows a practical example of an inverting-amplifier circuit of that type. The AC-modulation signal modulates the amplitude of the carrier-output signal.

Figure 7 shows how one half of an LM13600 can be used as a ring modulator or 4-quadrant multiplier. In that circuit, there is no carrier output when the modulation voltage is at ground level, but increases when the modulation voltage moves positive or negative with respect to ground. When the modulation voltage is positive, the carrier-output signal is inverted relative to the carrier input; and when the modulation voltage is negative, the carrier output is non-inverted.

The circuit in Fig. 7 is similar to the circuit in Fig. 6, except that the component values shown are suited for operation from a dual 15-volt supply, and that I_{BIAS} is adjustable via R7. The OTA's output (inverted relative to the input signal) feeds into the one end of R5, and at the same time the input signal feeds directly into the other end of R5. Potentiometer R7 is adjusted so that when the modulation input is tied to ground, the overall gain of the OTA is such that its output current exactly balances (cancels) the carrier input current to R5. Under that condition the circuit has no carrier output. When the modulation input goes positive, the OTA's gain increases and its output signal exceeds that caused by the carrier input to R5, so an inverted output signal is generated. Conversely, when the modulation input goes negative, the OTA's gain decreases and the carrier input to R5 exceeds the output of the OTA; therefore, a non-inverted output signal is generated.

Offset biasing

The circuits in Figs. 5–7 are shown with the OTA's input biased by fixed-value 470-ohm resistors wired between the two input terminals and ground. In practice, that simple arrangement may cause the circuit's DC level at the output to shift slightly when the gain-control input (I_{BIAS}) is varied between its minimum and maximum value. If desired, that level-shifting effect can be eliminated by adding a presettable offset-adjust control, as shown in Fig. 8. Potentiometer R4 enables the relative values of the biasing resistors, R2 and R3, to be varied over a limited range. To adjust the offset bias, reduce I_{BIAS} to zero, note the DC level of the output signal, and then increase I_{BIAS} to maximum and adjust R4 for the same DC-output level.

AGC amplifier

Figure 9 shows how to make an Automatic Gain Control (AGC) amplifier in which a 100:1 change in the input-signal amplitude causes only a 5:1 change in the output amplitude. In that circuit, I_{BIAS} is fixed by R4, and the output signal is available directly across R5. The output buffer is fed from the output of the OTA and is used as a signal rectifier. The rectified output of the buffer is smoothed via R6 and C2, and used to apply the I_D current to the linearizing diodes of the OTA. However, no significant I_D current is generated until the OTA's output goes high enough to turn on the Darlington buffer and the linearizing diodes. An increase in I_D reduces the OTA's gain, and negative feedback holds the output at a steady level.

The basic gain of the amplifier in Fig. 9, with no I_D current, is 40. Therefore, with an input signal of 30 mV p-p, the OTA's output of 1.2 volts p-p is not enough to generate an I_D current, so the OTA operates at full

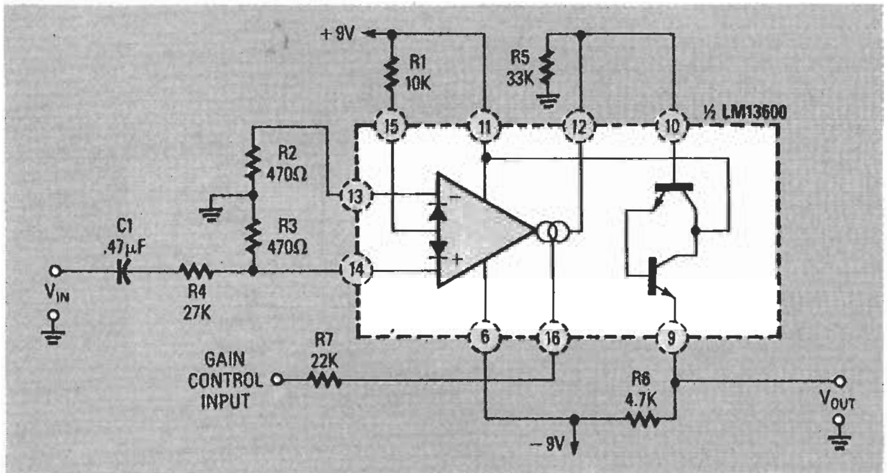

FIG. 5—A VOLTAGE-CONTROLLED amplifier (VCA) is one application for the LM13600. The LM13600 is also well suited for use as a stereo amplifier because it contains two matched amplifiers.

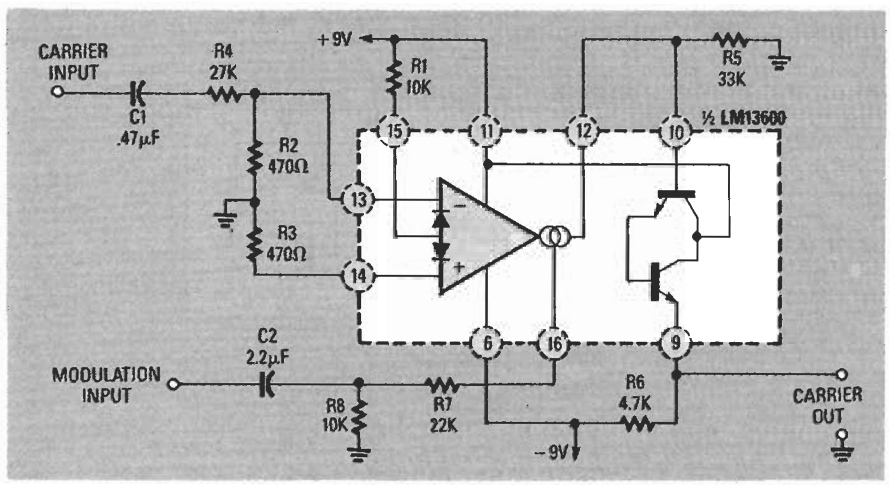

FIG. 6—AN AMPLITUDE MODULATOR or 2-quadrant multiplier can be made using one half of the LM13600.

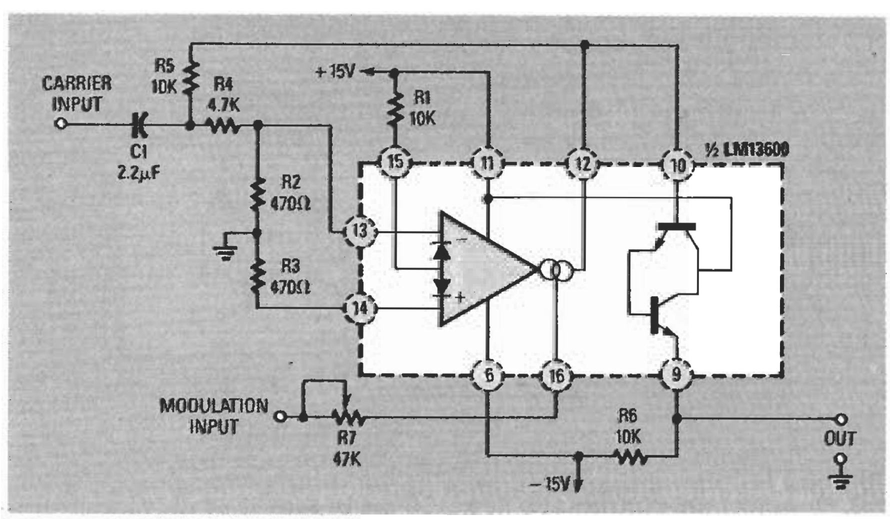

FIG. 7—THE CIRCUIT SHOWN HERE is a ring modulator or 4-quadrant multiplier.

gain. With an input of 300 mV, however, the OTA's output is enough to generate a significant I_D current, and the circuit's negative feedback automatically reduces the output level to 3.6 volts p-p, giving an overall gain of 11.7. With an input of 3 volts, the gain falls to 2 (an output of 6 volts p-p). The circuit thus has a 20:1 signal compression over that range.

Voltage-controlled resistors

One unusual application of the LM13600 is as a *V*oltage-*C*ontrolled *R*esistor (VCR), using the circuit shown in Fig. 10. The basic theory is as follows: An AC signal applied to the R_X terminals feeds into the inverting terminal of the OTA via C1, the output-buffer transistors, and the R5/R_A attenuator. The OTA will then generate an output current that is proportional to V_{IN} and I_{BIAS}. Therefore, because R = V/I, the R_X terminal functions as an AC resistor whose value is determined by I_{BIAS}.

The effective resistance between the R_X terminals of the circuit in Fig. 10 equals $(R5+R_A)/(gm \times R_A)$, where gm (transconductance) is approximately $20 \times I_{BIAS}$. That formula can be approximated as $R_X = R5/(I_{BIAS} \times 20R_A)$. Using the component values shown, R_X equals approximately 10 megohms when it has an I_{BIAS} current of 1 μA, and 10 kilohms when it has an I_{BIAS} current of 1 mA.

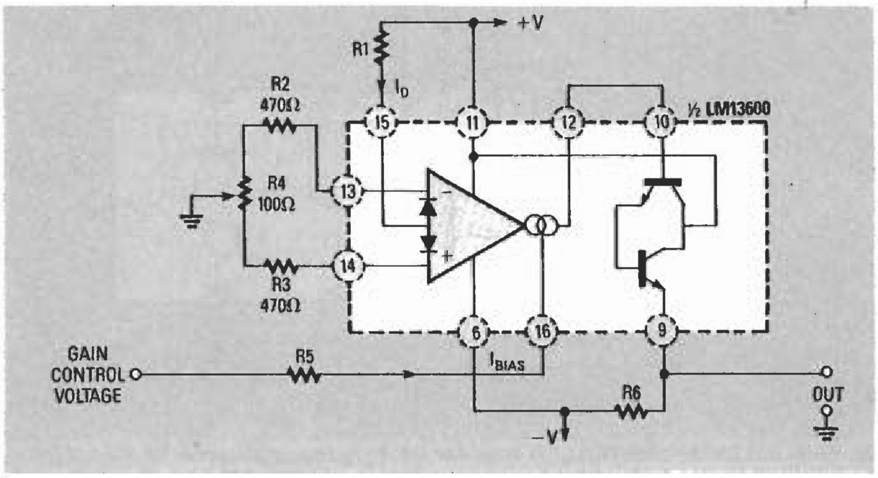

FIG. 8—A METHOD OF APPLYING offset biasing to the LM13600.

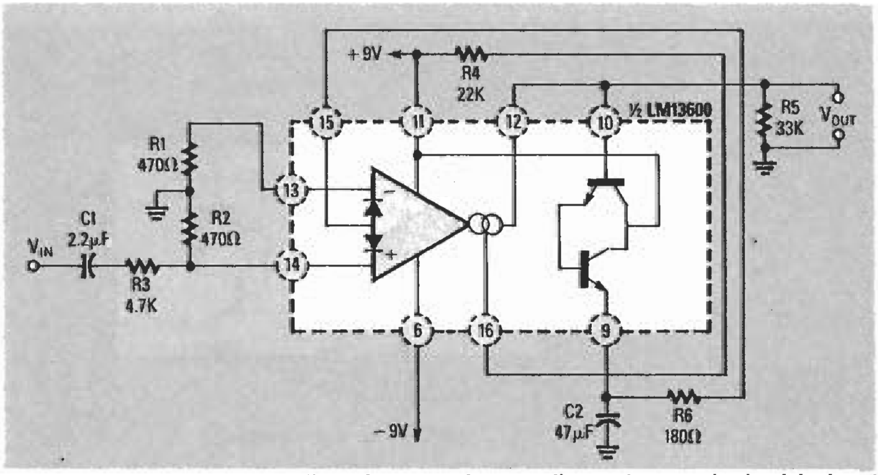

FIG. 9—AN AGC AMPLIFIER adjusts its own gain according to the magnitude of the input signal.

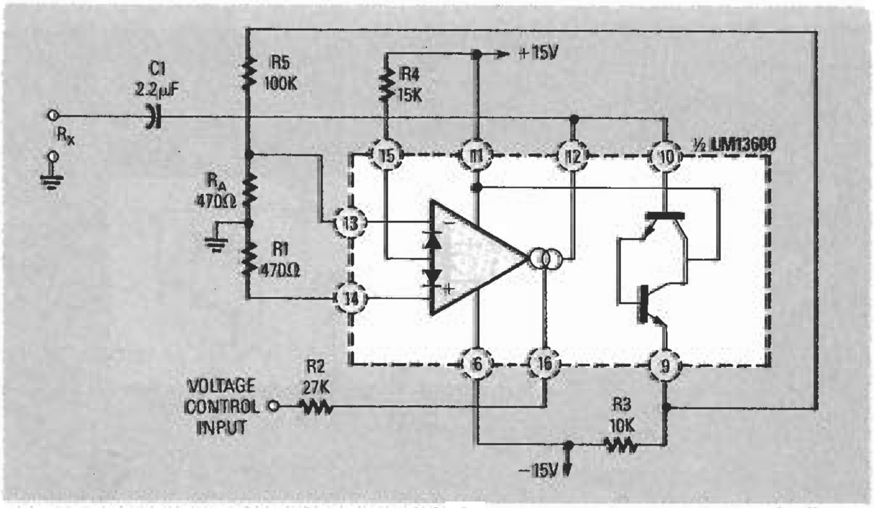

FIG. 10—A VOLTAGE-CONTROLLED RESISTOR can be used as an electronically-controlled resistor.

Voltage-controlled filters

A voltage-controlled low-pass filter can be implemented by using one half of an LM13600 in the configuration shown in Fig. 11. In that circuit, the values of R5, C2, and I_{BIAS} control the cut-off frequency (f_O) of the filter. The input signal is applied to the non-inverting terminal of the OTA via voltage-divider network R1/R2. The OTA's output signal is "followed" by the buffer stage and fed back to the inverting terminal via an identical voltage-divider network, R5/R_A. The basic OTA operates as a non-inverting amplifier with a gain of R5/R_A, but because the input signal to the OTA is applied via a voltage divider with a value equal to R5/R_A, the overall circuit operates as a unity-gain voltage follower.

At low frequencies, C2 has a very high impedance and is able to be fully charged by the OTA's output current, so the circuit operates as a voltage follower as was previously described. As the frequency increases, C2's impedance decreases and is no longer able to be fully charged by the OTA's output current, so the output signal starts to attenuate at a rate of 6-dB-per-octave. The cut-off point of the circuit, defined as the point where the output falls by 3 dB, occurs when $X_C/20I_{BIAS}$ equals R5/RA, as shown by the formula in the diagram (gm is approximately equal to $20 \times I_{BIAS}$). With the component values that are shown in Fig. 11, the filter's cut-off

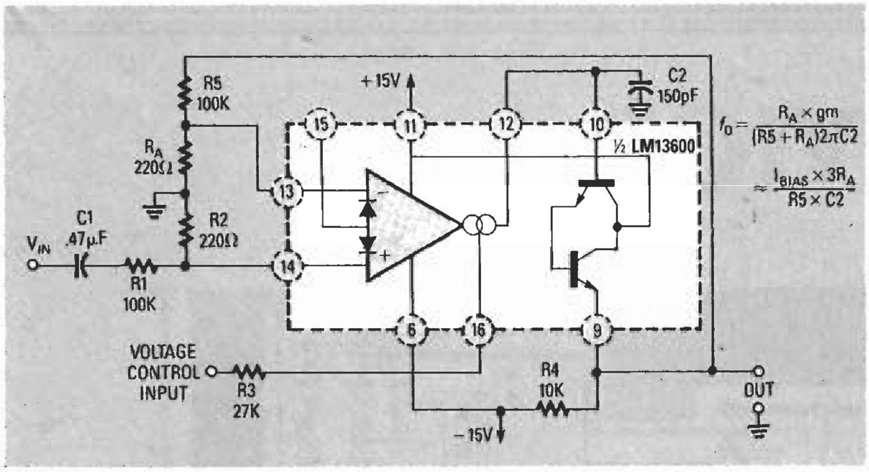

FIG. 11—A VOLTAGE-CONTROLLED LOW-PASS filter covering 45 Hz to 45 kHz.

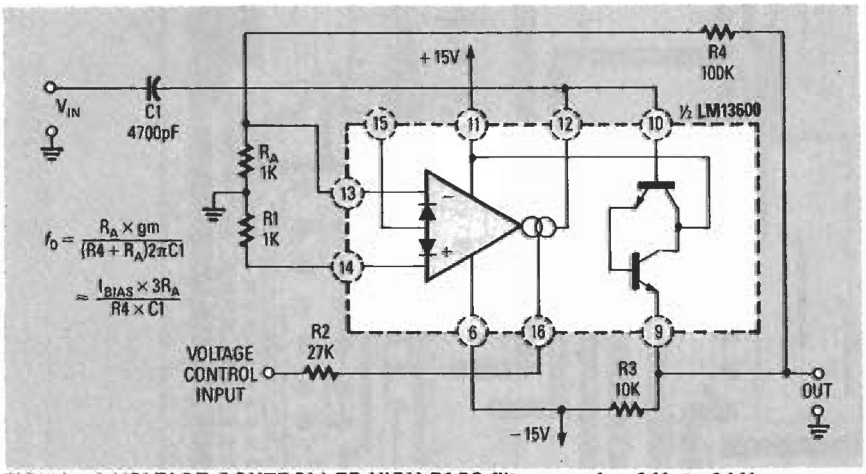

FIG. 12—A VOLTAGE-CONTROLLED HIGH-PASS filter covering 6 Hz to 6 kHz.

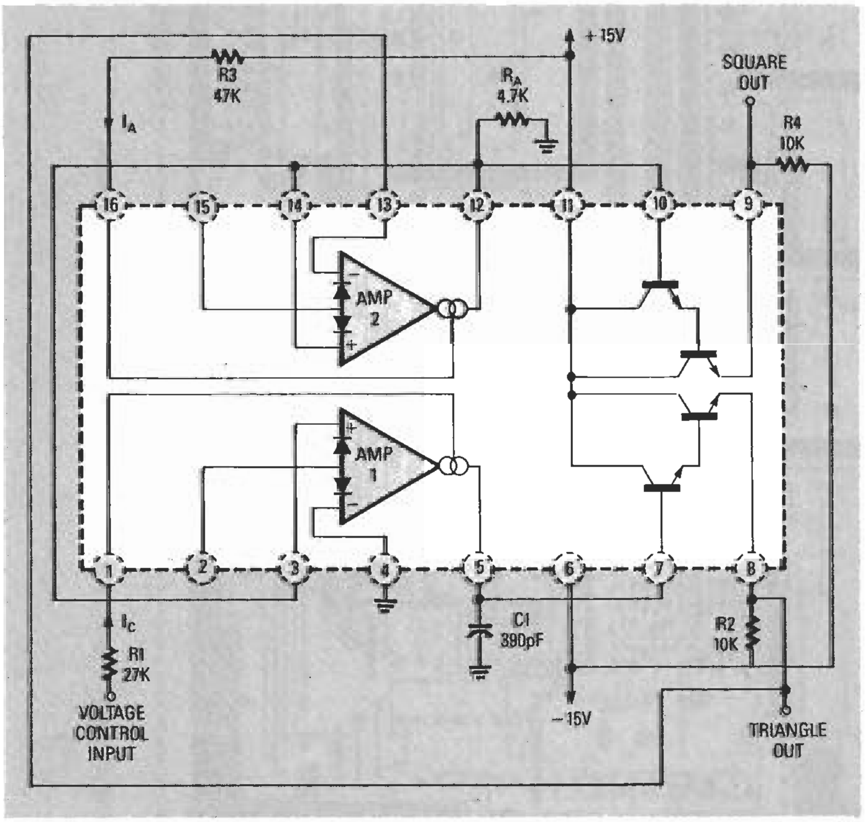

FIG. 13—A COMBINATION TRIANGLE-SQUARE-WAVE VCO covering 200 Hz to 200 kHz.

point occurs at about 45 Hz with an I_{BIAS} of 1 µA, and at 45 kHz with an I_{BIAS} of 1 mA.

A similar principle can be used to make a voltage-controlled high-pass filter. As shown in Fig. 12, that circuit has cut-off frequencies of 6 Hz when it has an I_{BIAS} current of 1 µA, and 6 kHz when it has an I_{BIAS} current of 1 mA.

Voltage-controlled oscillators

To conclude this look at the LM13600 operational transconductance amplifier, Fig. 13 shows how to use the IC as a *V*oltage-*C*ontrolled *O*scillator (VCO). The circuit uses both halves of the LM13600, and simultaneously generates both triangle and square waves.

To understand the operating theory of the circuit, assume initially that capacitor C1 is negatively charged and that the square-wave output signal has just switched high. Under that condition a positive voltage is developed across R_A, which is fed to the non-inverting terminals of the two amplifiers. That voltage causes amp 1 to generate a positive output current, equal to bias current I_C, that flows into C1 and generates a positive-going linear ramp voltage.

The ramp voltage is then fed to the inverting terminal of amp 2 via the Darlington buffer stage, until it eventually equals the voltage on the non-inverting terminal, at which point the output of amp 2 starts to swing in a negative direction. That initiates a regenerative switching action, and at that moment, the signal at the square-wave output terminal abruptly goes negative.

In that new state, a negative voltage is generated across resistor R_A, causing amp 1 to generate a negative output current equal to I_C, causing capacitor C1 to discharge until its voltage equals that of R_A, at which point the square-wave output switches high again.

The process repeats over and over again, making available a triangle waveform at R2 and a square wave at R4. The frequency of those waveforms is variable via the voltage-control input; that input is what controls the value of I_C. With the component values shown, the circuit then generates a frequency of about 200 Hz when the I_C bias current is 1 µA, and 200 kHz when the bias current is 1 mA. **R-E**