

UNDERSTANDING AND USING 'OTA' OP-AMP ICs

by Ray Marston

Part 1

Take a look at Operational Transconductance Amplifier (OTA) op-amp principles and at practical CA3080 OTA circuits in this opening episode of a two-part mini-series.

Most popular op-amp (operational amplifier) ICs — such as the 741, CA3140, and LF351, etc. — give an output voltage that is proportional to the difference between the IC's two input pin voltages, and are thus known as voltage-differencing amplifiers or 'VDAs.' There are, however, two other basic types of op-amps that are in common use; one of these is the type that gives an output voltage that is proportional to the difference between the *currents* applied to its two input terminals, and is thus known as a current-differencing amplifier or 'CDA.' This type of op-amp was described in depth in the "Understanding and Using 'Norton' Op-Amp ICs" two-part mini-series published in this magazine. The third type of op-amp is known as an operational transconductance amplifier or 'OTA,' and acts as a variable-gain voltage-to-current amplifier.

One of the best known OTA ICs is the CA3080, which is of particular value in making voltage- or current-controlled amplifiers, or micro-power voltage comparators or oscillators, etc. OTAs have operating characteristics very different from conventional op-amps, and *Figure 1* illustrates the major differences between these two types of devices.

Figure 1(a) shows the basic symbol and formulas of the conventional op-amp, which is essentially a voltage amplifying device. It has differential input terminals and gives an output of $A_O \times (e_1 - e_2)$, where A_O is the open-loop voltage gain of the op-amp and e_1 and e_2 are the signal voltages at the non-inverting and inverting input terminals, respectively. Note that the open-loop voltage gain of this op-amp is fixed, and the device has a high input impedance and a low output impedance.

Figure 1(b) shows the basic symbol and formulas of an OTA, which is essentially a voltage-to-current amplifier. It has differential voltage input terminals (like a conventional op-amp) but — as indicated by the constant-current symbol on its output — these input voltages produce a high-impedance output in the form of a current with a value of $g_m \times (e_1 - e_2)$, where g_m is the transconductance or voltage-to-current gain of the device and can be controlled by (and is directly proportional to the value of) an external bias current fed into the I_{bias} terminal. In the CA3080, I_{bias} can be varied from $0.1\mu A$ to $1mA$, giving a 10,000:1 gain-control range.

An OTA is a very versatile device. It can, for example, be made to act like a normal op-amp by simply wiring a suitable load resistance to its output terminal (to convert its output current into voltage). Again, since the magnitude of I_{bias} can easily be controlled by an external voltage and a series resistor, the OTA can easily be used as a voltage-controlled amplifier (VCA), oscillator (VCO), or filter (VCF), etc. Note that the total current consumption of the CA3080 OTA is only twice the I_{bias} value (which can be as low as $0.1\mu A$), enabling the device to be used in true micro-power applications.

The best known current-production versions of the OTA are the CA3080 and the LM13700. The LM13700 is a dual second-generation OTA with built-in output-buffer stages, and will be fully described next month. The CA3080 is a first-generation OTA, and is the exclusive subject this month. *Figure 2(a)* shows the connections of the eight-pin DIL 'E' version of the CA3080. *Figure 2(b)* shows its internal circuit, and *Figure 3* lists its basic parameter values.

CA3080 BASICS

The CA3080 is a fairly simple device and consists of one differential amplifier and four current mirrors. *Figure 4* shows the basic circuit and formulas of its differential amplifier. Its emitter current (I_C) 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 (up to $\pm 25mV$ maximum), the I_a and I_b currents differ and produce an $I_b - I_a$ value of $V_{in} \times g_m$, where g_m is the OTA's transconductance, is directly proportional to I_C , and has a typical mho value of about $20 \times I_C$. The *Figure 4* circuit is of little value on its own and, in the CA3080, it is turned to good use by using a simple current mirror to externally control its I_C value (and thus the g_m of the OTA), and by using another three current mirrors to extract the difference between the I_a and I_b currents and make this difference current available to the outside world. A current mirror (CM) is a three-terminal circuit that, when pro-

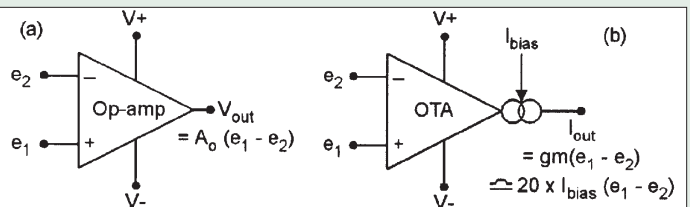


Figure 1. A conventional op-amp (a) is a fixed-gain voltage-amplifying device. An OTA (b) is a variable-gain voltage-to-current amplifier.

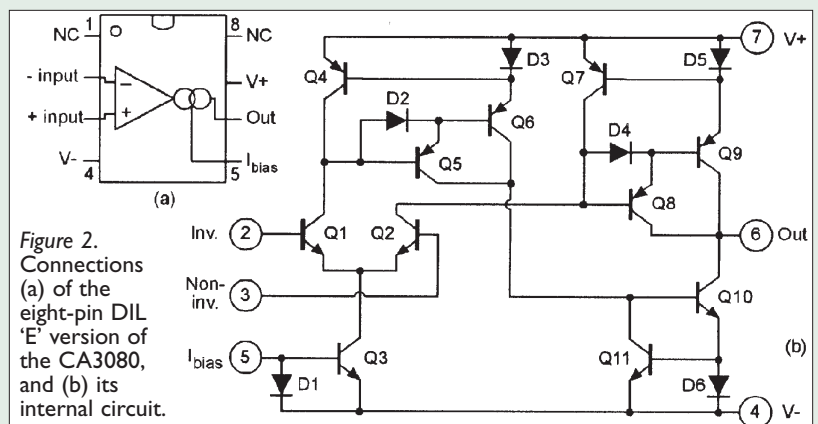


Figure 2. Connections (a) of the eight-pin DIL 'E' version of the CA3080, and (b) its internal circuit.

vided with an external input bias current, produces an in-phase current of identical value at its output terminals, as shown in Figure 5. Some CMs act as current sinks, as shown in Figure 5(a) and others as current sources, as in Figure 5(b). When a CM source and a CM sink are connected as shown in Figure 6 and powered from split supply rails, they generate a differential ($I_{\text{source}} - I_{\text{sink}}$) current in any external load connected to the 0V rail. Figure 7 shows the actual circuits of two sink-type current mirrors. In the simplest of these (Figure 7(a)), a diode-connected transistor (QA) is wired across the base-emitter junction of a second, closely matched, transistor that is integrated on the same chip. The input current is fed to the bases of both matched transistors and thus divides equally between them. Suppose these transistors have current gains of $\times 100$ and are each drawing base currents of $5\mu\text{A}$. In this case, they each draw collector currents of $500\mu\text{A}$. Note, however, that the QA collector current is drawn from the circuit's input current, which thus equals $500\mu\text{A}$ plus ($2 \times 5\mu\text{A}$), or $510\mu\text{A}$, and that the QB collector current is the output or mirror current of the circuit.

The input and output currents of this circuit are thus almost identical (within a few percent), irrespective of the input current magnitude. In practice, the input/output current ratio of the above circuit depends on the close gain matching of the two transistors, and this can actually vary by several percent. Figure 7(b) shows an improved current mirror circuit that is less sensitive to current gain variations and also gives an improved (greater) output impedance.

Figure 8 shows how the differential amplifier and four current mirrors are interconnected in the CA3080 to make a practical OTA. Bias current I_{bias} controls the emitter current, and thus the gm, of the Q1-Q2 differential amplifier via CMC. The collector currents of Q1 and Q2 are mirrored by CMA and CMB respectively, and then fed into the bias and sink terminals respectively of CMD, so that the externally available output current of the circuit is equal to $I_B - I_A$. Looking back to Figure 2(b), which shows the actual internal circuit of the CA3080, the reader should now have little difficulty in working out the functions of individual circuit elements. Q1 and Q2 form the differential amplifier, with D1-Q3 making up CMC of Figure 8, and CMD comprising D6-Q10-Q11. Current mirrors CMA (Q4-Q5-Q6-D2-D3) and CMB (Q7-Q8-Q9-D4-D5) are slightly more complex, using Darlington pairs of transistors, plus speed-up diodes, to improve their performances.

SOME FINER POINTS

All the major operating parameters of the CA3080 are adjustable and depend on the value of I_{bias} . The maximum (short circuit) output current is equal to I_{bias} , the total operating current of the OTA is double the I_{bias} value, and the input bias currents drawn by pins 2 and 3 typically equal $I_{\text{bias}}/200$.

The transconductance (gm) and the input and output impedance values also vary with the I_{bias} value, as shown in the

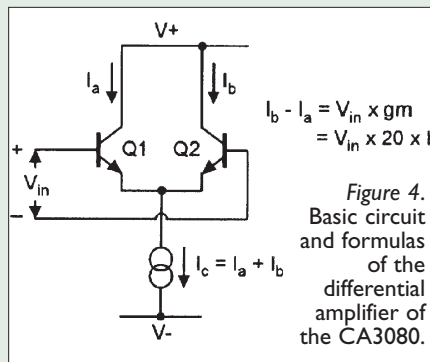


Figure 4. Basic circuit and formulas of the differential amplifier of the CA3080.

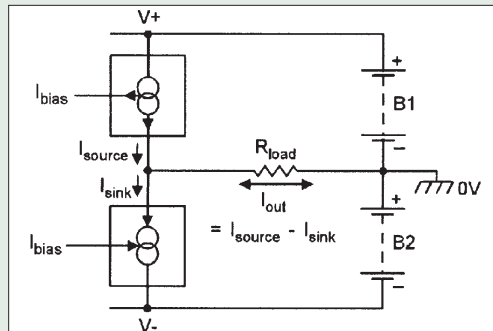


Figure 6. When two current mirrors are wired as shown, they generate a differential current in an external load.

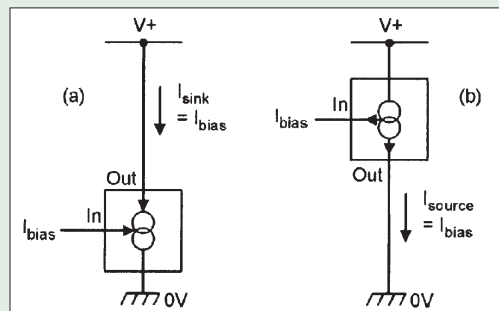


Figure 5. Symbols representing (a) a current-mirror sink and (b) a current-mirror source.

graphs in Figure 9, which show typical parameter values when the IC is driven from split 15V supplies at an ambient temperature of $+25^\circ\text{C}$. Thus, at a bias current of $10\mu\text{A}$, gm is typically $200\mu\text{mho}$, and input and output impedances are 800k and 700M , respectively. At 1mA bias, the values change to 20mmho , 15k , and 7M , respectively.

The available output voltage swing of the IC depends on the values of I_{bias} and any external load resistor connected to the OTA output. If the load impedance is infinite, the output can swing to within 1V_5 of the positive supply rail and to within 0V_5 of the negative rail. If the impedance is finite, the peak output voltage swing is limited to $I_{\text{bias}} \times R_L$. Thus, at $10\mu\text{A}$ bias with a 100k load, the available output voltage swing is a mere $\pm 1\text{V}_0$.

The slew rate (and bandwidth) of the IC depend on the value of I_{bias} and any external loading capacitor connected to the output. The slew rate value, in volts per microsecond, equals I_{bias}/C_L , where C_L is the loading capacitance value in pF, and the I_{bias} value is in microamps. With no external loading capacitor connected, the maximum slew rate of the CA3080 is about $50\text{V}/\mu\text{S}$.

BASIC CIRCUITS

The CA3080 is a very easy IC to use. Its pin 5 I_{bias} terminal is internally connected to the pin 4 negative supply rail via a base-emitter junction, so the biased voltage of the terminal is

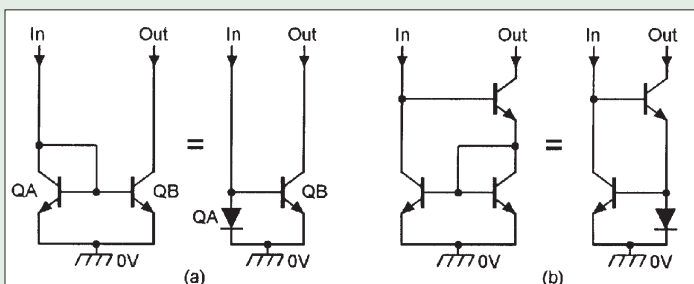


Figure 7. Examples of simple sink-type current mirrors.

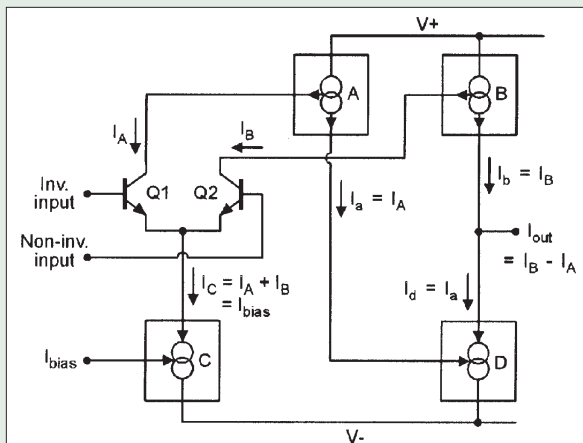


Figure 8. The CA3080 comprises one differential amplifier and four current mirrors.

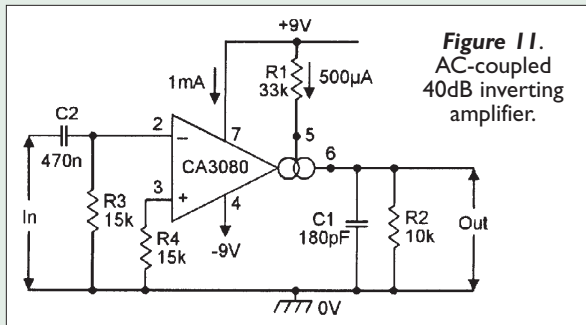


Figure 11. AC-coupled 40dB inverting amplifier.

about 600mV above the pin 4 value. I_{bias} can thus be obtained by connecting pin 5 to either the common rail or the positive supply rail via a suitable current limiting resistor.

Figures 10 and 11 show two simple ways of using the CA3080 as a linear amplifier with a voltage gain of about 40dB. The Figure 10 circuit acts as a DC-differential amplifier, and Figure 11 as an AC-coupled inverting amplifier. Both designs operate from split 9V supplies, so 17.4V is generated across bias resistor R1, which thus feeds an I_{bias} of 500 μ A into pin 5 and thus makes each IC consume another 1mA of supply current.

At an I_{bias} value of 500 μ A, the g_m of the CA3080 is roughly 10mmho, so since the outputs of the Figure 10 and 11 circuits are loaded by a 10k resistor (R2), they give an overall voltage gain of 10mmho x 10k = x100, or 40dB. The peak current that can flow into the 10k load is 500 μ A (I_{bias}), so the peak available output voltage is $\pm 5V_0$. The output is also loaded by 180pF capacitor C1, giving the circuit a slew rate limit of 500 μ A/180pF, or 2.8V/ μ S. The output impedance of

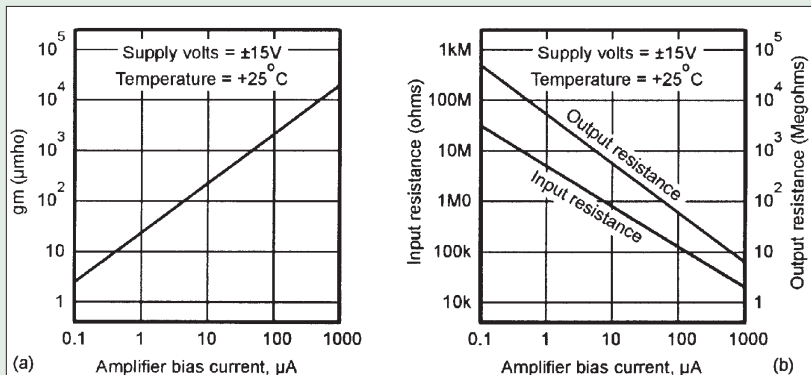


Figure 9. The transconductance (a) and the input and output resistances (b) of the CA3080 vary with the bias-current value.

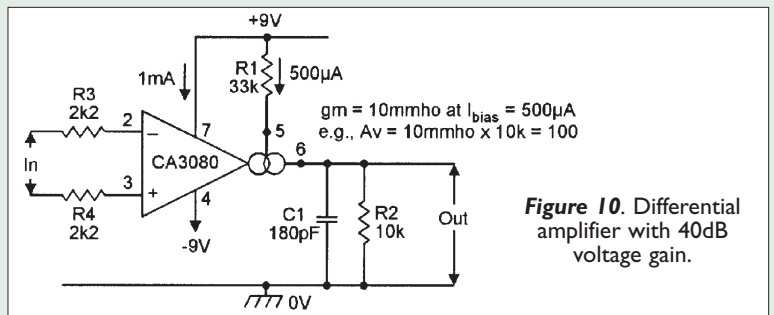


Figure 10. Differential amplifier with 40dB voltage gain.

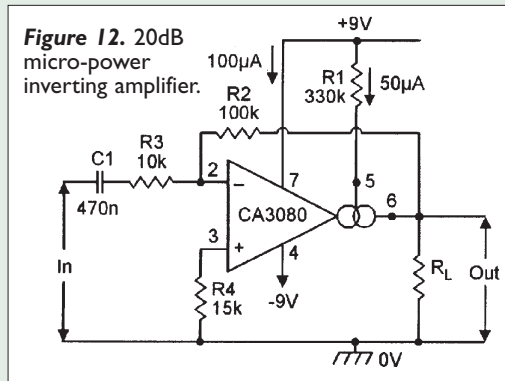


Figure 12. 20dB micro-power inverting amplifier.

each circuit equals the R2 value of 10k.

Note in these two circuits that the IC is used in the open-loop mode, and that if the slew rate of the IC is not externally limited via C1, the OTA will operate at its maximum bandwidth and slew rate. Under such a condition, the CA3080

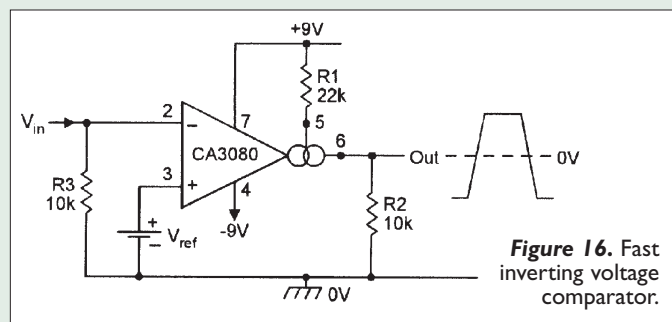
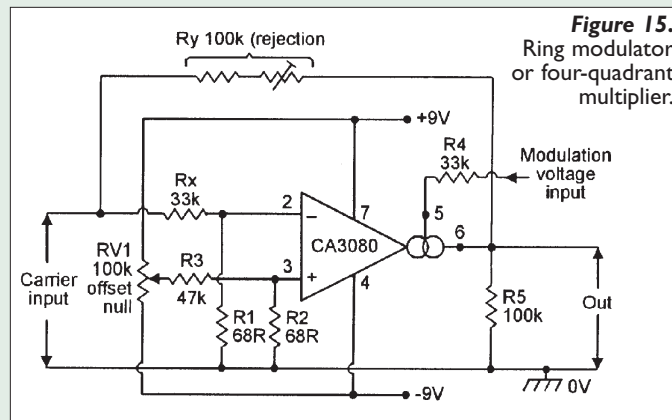
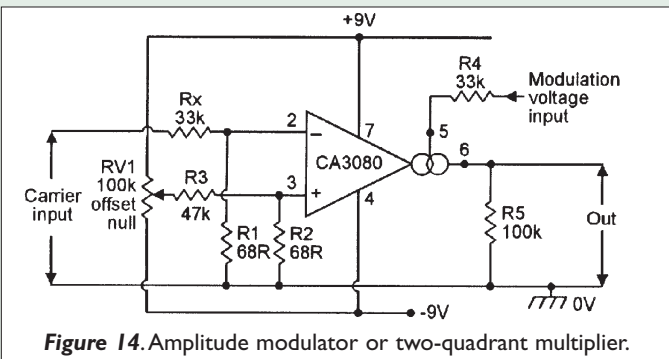
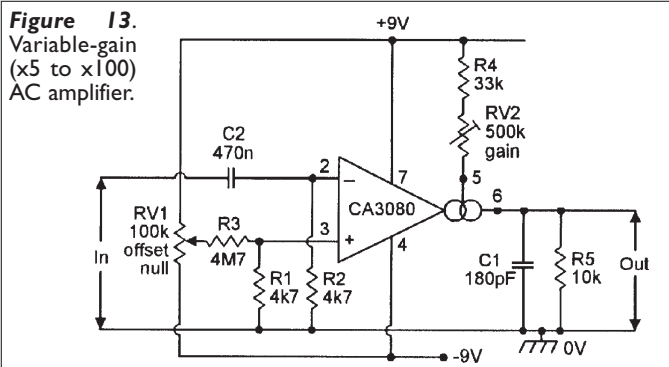
may be excessively noisy, or may pick up unwanted RF signals. In the Figure 10 circuit, the differential inputs are applied via series resistors R3 and R4, which help equalize the source impedances of the two signals and thus maintain the DC balance of the OTA. The Figure 11 circuit is a simple variation of the above design, with both inputs tied to the common rail via 15k resistors and with the input signal applied to one terminal only. With the input fed to pin 2 as shown, the circuit acts as a 40dB inverting amplifier; alternatively, non-inverting action can be obtained by connecting the input signal to pin 3 via C2.

CLOSED-LOOP OPERATION

The Figure 10 and 11 circuits are used in the open-loop mode and their voltage gains thus depend on the value of I_{bias} which, in turn, depends on the value of supply rail voltage. The voltage gain of the CA3080 can be made almost independent of the I_{bias} and supply voltage values by using conventional closed-loop op-amp techniques, as shown in the micro-power 20dB AC-coupled inverting amplifier in Figure 12.

Figure 12 is wired like a conventional op-amp inverting amplifier, with its voltage gain (A_V) determined primarily by the R2/R3 ratio (x10, or 20dB). This gain equation is, however, only valid when the value of an external load resistor (R_L) is infinite, since the output impedance of this design is equal to R_2/A_V , or 10k, and any external load causes this impedance to give a potential divider action that reduces the output of the circuit.

In Figure 12, the main function of I_{bias} is to set the total operating current of the circuit and/or the maximum available output swing. With the component values shown, I_{bias} has a value of 50 μ A. When R_L is infinite, the output is loaded only by R2, which has a value of 100k, so the maximum available output voltage swing is $\pm 5V_0$. If R_L has a value of 10k, on the other hand, the maximum output voltage swing is $\pm 0.5V$. This circuit can thus be designed to give any desired values of voltage gain and peak output voltage. Note that, since the IC is used in the closed loop mode, external slew rate limiting is not required.



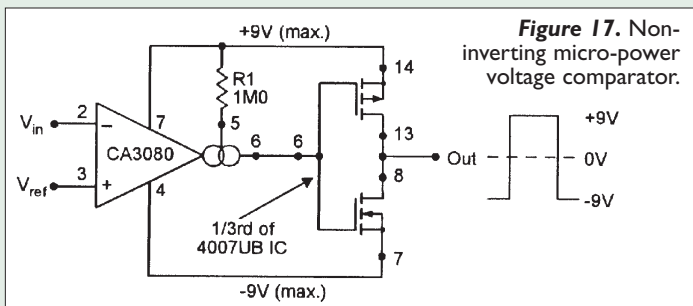
OFFSET BALANCING

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 prevailing I_{bias} values. *Figure 13* shows how suitable bias can be applied to an inverting AC amplifier in which the voltage gain is variable from roughly x5 to x100 via RV2, and the offset balance is pre-set via RV1. The circuit is set up by adjusting RV2 to its minimum (maximum gain) value and then trimming RV1 to give zero DC output with no AC input signal applied.

VOLTAGE-CONTROLLED GAIN

The most important uses of the CA3080 are in true micro-power amplifier and oscillator applications, and in applications in which important parameters are variable via an external voltage. In the latter category, one important application is as a voltage-controlled amplifier (VCA) or amplitude modulator, in which a carrier signal is fed to an amplifier's input, and its output amplitude is controlled or modulated by another signal fed to the I_{bias} terminal. *Figure 14* shows a practical version of such a circuit.

The *Figure 14* circuit acts as 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, and offset biasing is applied to the non-inverting pin of the IC via R3-RV1. The carrier input signal is applied to the inverting pin of the CA3080 via potential divider Rx-R1. When Rx has the value shown, the circuit gives roughly unity gain when the modulation input terminal is tied to the zero volts rail. The gain is x2 when the modulation terminal is at +9V, and the circuit gives roughly 80dB of signal rejection when the modulation terminal is tied to the -9V rail. Note that the instantaneous polarity of the output signal in the *Figure 14* circuit is determined entirely by the instantaneous polarity of the input signal, which has only two possible 'states'. This type of circuit is thus known as a two-quadrant multiplier. The amplitude of the



output signal is determined by the product of the input and gain-control values.

Figure 15 shows how the above circuit can be modified so that it acts as a ring-modulator or four-quadrant multiplier, in which the output signal polarity depends on the polarities of both the input signal and the modulation voltage. The *Figure 15* circuit is identical to that of *Figure 14*, except that resistor network Ry is connected between the input and output terminals. The action here is such that, when the modulator input is tied to the zero volts rail, the inverted signal currents feeding into R5 from the output of the OTA are exactly balanced by the non-inverting signal currents flowing into R5 from the input signal via Ry, so that zero output is generated across R5. If the modulation input goes positive, the output of the OTA exceeds the current of the Ry network, and an inverted gain-controlled output is obtained. If the modulation input is negative, on the other hand, the output current of Ry exceeds that of the OTA, and the non-inverted gain-controlled output is obtained.

Thus, both the phase and the amplitude of the output signal of this four-quadrant multiplier 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 sinewaves to the two inputs. Note that, with the Rx and Ry values shown, the *Figure 15* circuit gives a voltage gain of x0.5 when the modulation terminal is tied to the positive or negative supply rail; the gain doubles if the values of Rx and Ry are halved. Also note that the *Figure 14* and *15* circuits each have a high-output impedance, and

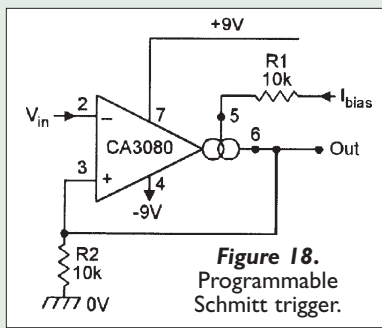


Figure 18.
Programmable Schmitt trigger.

that, in practice, a voltage following buffer stage must be interposed between the output terminal and the outside world.

COMPARATOR CIRCUITS

The CA3080 can easily be used as a programmable or micro-power voltage comparator. *Figure 16* shows the basic circuit of a fast, programmable, inverting comparator, in which a reference voltage is applied to the non-inverting input terminal, and the test input is applied to the inverting terminal. The circuit action is such that the output is driven high when the test input is below V_{ref} , and is driven low when test is above V_{ref} (the circuit can be made to give a non-inverting comparator action by transposing the input connections of the IC). With the component values shown, the I_{bias} current in the *Figure 16* circuit is several hundred microamps and under this condition, the CA3080 has a slew rate of about $20V/\mu S$ and thus acts as a 'fast' comparator. When the test and V_{ref} voltages are almost identical, the IC acts as a linear amplifier with a voltage gain of $gm \times R2$ (about $\times 200$ in this case). When the two input voltages are significantly different, the output voltage limits at values determined by the I_{bias} and $R2$ values. In *Figure 16*, the output limits at about $\pm 7V$ when $R2$ has a value of $10k$, or at $\pm 0.7V$ when $R2$ has a value of $1k$.

Figure 17 shows how the above circuit can be modified so that it acts as an ultra-sensitive micro-power comparator which

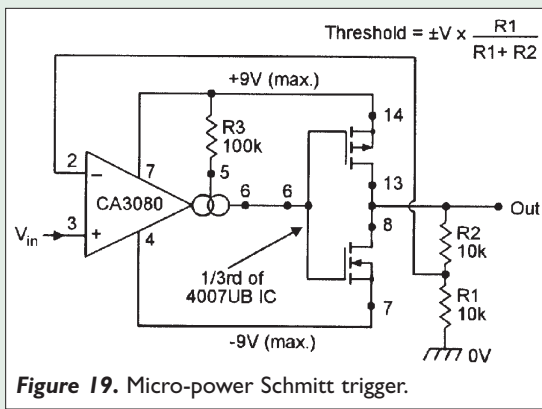


Figure 19. Micro-power Schmitt trigger.

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SCHMITT TRIGGER CIRCUITS

The simple voltage comparator circuit in *Figure 16* can be made to act as a programmable Schmitt trigger by connecting the non-inverting reference terminal directly to the output of the CA3080, as shown in *Figure 18*. In this case, when the input is high, a positive reference value of $I_{bias} \times R2$ is generated. When V_{in} exceeds this value, the output regeneratively switches low and generates a negative reference voltage of $I_{bias} \times R2$.

When V_{in} falls below this new value, the output switches high again and once more generates a positive reference voltage of $I_{bias} \times R2$. Thus, the trigger thresholds (and also the peak output voltages) of this Schmitt circuit can be precisely controlled or programmed via either I_{bias} or $R2$. *Figure 19* shows an alternate type of Schmitt, in which the output transitions fully between the supply rail values, and the switching threshold values are determined by the $R1$ and $R2$ ratios and the values of supply voltage, V , and equal $\pm V \times R1 / (R1 + R2)$.

ASTABLE CIRCUITS

The *Figure 19* Schmitt trigger circuit can be made to act as an astable multivibrator or squarewave generator by connecting its output back to the non-inverting input terminal via an R-C time-constant network, as shown in *Figure 20*. The output of this circuit switches fully between the supply rail values, is approximately symmetrical, and has a frequency that is determined by the values of $R3$ and $C1$, and by the ratios of $R1$ and $R2$. The circuit action 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 which point, the output switches low. $C1$ then discharges via $R3$ until the $C1$ voltage reaches the negative reference voltage value determined by the $R1$ - $R2$ ratio, at which point, the output switches high again, and the whole process then repeats *ad infinitum*.

Finally, to complete this look at the CA3080 OTA, *Figure 21* shows how the above circuit can be modified to give an output waveform with a variable mark-space ratio. In this case, $C1$ alternately charges via $D1$ - $R3$ and the left half of $RV1$, and discharges via $D2$ - $R3$ and the right half of $RV1$, to give a mark-space ratio that is fully variable from 10:1 to 1:10 via $RV1$.

Note in the above two astable circuits that the CA3080 is biased at only a few microamps and that 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 figures of only a few tens of microamps can easily be obtained. **NV**

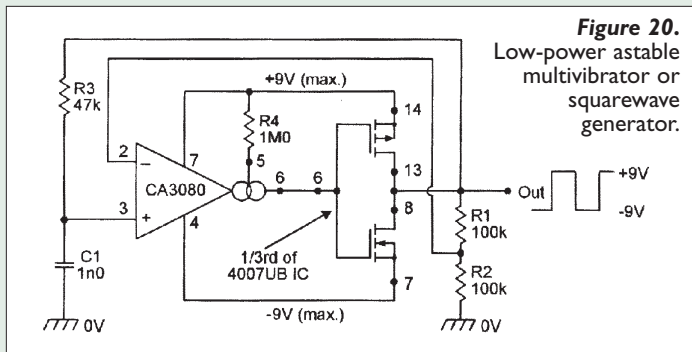


Figure 20.
Low-power astable multivibrator or squarewave generator.

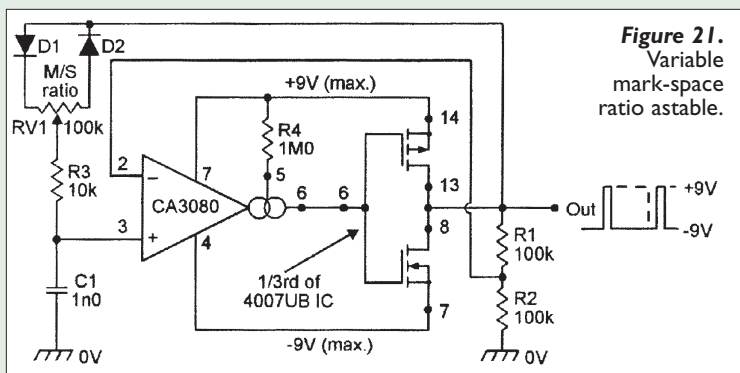


Figure 21.
Variable mark-space ratio astable.

UNDERSTANDING AND USING 'OTA' OP-AMP ICs

by Ray Marston

Part 2

Ray Marston looks at the operating principles and practical applications of the LM13700 dual Operational Transconductance Amplifier (OTA) IC in this concluding episode of a two-part mini-series.

Last month's opening episode of this two-part mini-series described basic 'OTA' operating principles, and took a close look at the popular and widely available CA3080 OTA IC. The CA3080 is actually a simple, first-generation OTA that generates fairly high levels of signal distortion and has a high-impedance unbuffered output. This month's concluding episode describes an improved second-generation OTA IC – the LM13700 – which does not suffer from these snags.

The LM13700 is actually a dual OTA, as indicated by the pin connection diagram in Figure 1. Each of its OTAs is an improved version of the CA3080, and incorporates input linearizing diodes that greatly reduce signal distortion, and have an optional buffer stage that can be used to give a low impedance output. The LM13700 is, in fact, a very versatile device, and can easily be made to act as a voltage-controlled amplifier (VCA), voltage-controlled resistor (VCR), voltage-controlled filter (VCF), or voltage-controlled oscillator (VCO), etc.

LINEARIZING DIODES

The CA3080 OTA consists of (as described last month) a differential amplifier plus a number of current mirrors that give an output equal to the difference between the amplifier's two collector currents, as shown in the simplified circuit in Figure 2. A weakness of this circuit is that its input signals must be limited to 25mV peak-to-peak if excessive signal distortion is not to occur. This distortion is caused by the inherently non-linear V_{be} -to- I_c transfer characteristics of Q1 and Q2.

Figure 3 shows the typical transfer characteristics graph of a small-signal silicon transistor. Thus, if this tran-

sistor is biased at a quiescent collector current of 0.8mA, an input signal of 10mV peak-to-peak produces an output current swing of +0.2mA to -0.16mA, and gives fairly small distortion. But an input swing of 30mV peak-to-peak produces an output swing of +0.9mA to -0.35mA, and gives massive distortion. In practice, the CA3080 gives typical distortion figures of about 0.2% with a 20mV peak-to-peak input, and a massive 8% with a 100mV peak-to-peak input.

Figure 4 shows the basic 'usage' circuit of one of the improved second-generation OTAs of the LM13700, which is almost identical to that of the CA3080, except for the addition of linearizing diodes D1 and D2, which are integrated with Q1 and Q2, and thus have characteristics matched to those of the Q1 and Q2 base-emitter junctions. In use, equal, low-value resistors – R1 and R2 – are wired between the inputs of the differential amplifier and the common supply line, and bias current I_D is fed to them from the positive supply rail via R3 and D1-D2 and, since D1-D2 and R1-R2 are matched, divides equally between them to give R1 and R2 currents of $I_D/2$.

The circuit's input voltage is applied via R4 (which is large relative to R1) and generates input signal current I_s , which feeds into R1 and thus generates a signal voltage across it that reduces the D1 current to $(I_D/2) - I_s$. The I_D current is, however, constant, so the D2 current rises to $(I_D/2) + I_s$. Consequently, the linearizing diodes of the Figure 4 circuit apply heavy, negative feedback to the differential amplifier and give a large reduction in signal distortion. If I_s is small relative to I_D , the output signal current of the circuit is equal to $2 \times I_s \times (I_{bias}/I_D)$. Thus, the circuit's gain can be controlled via either I_{bias} or I_D .

The OTAs of the LM13700 can be used as simple OTAs of the CA3080 type by ignoring the presence of the two linearizing diodes, or can be used as low-distortion amplifiers by using the diodes as shown in Figure 4. The graph in Figure 5 shows typical distortion levels of the LM13700 at various peak-to-peak values of input signal voltage, with

and without the use of linearizing diodes. Thus, at 30mV input, distortion is below 0.03% with the diodes, but 0.7% without them, and at 100mV, input is roughly 0.8% with the diodes, but 8% without them.

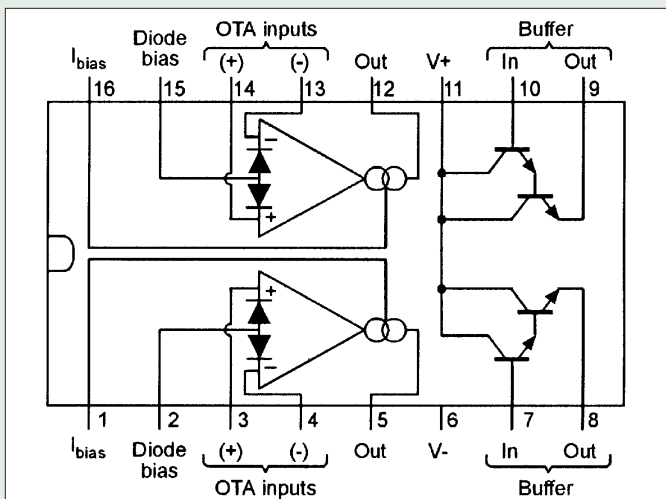


Figure 1. Pin connections of the LM13700 dual OTA IC.

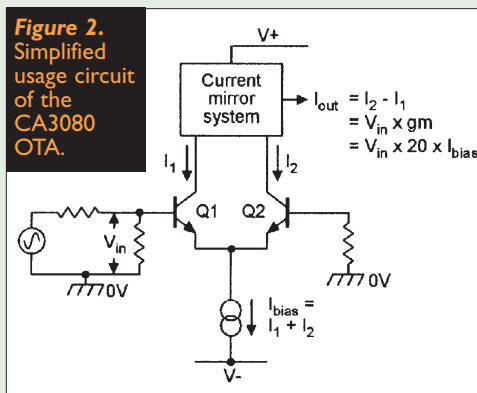


Figure 2. Simplified usage circuit of the CA3080 OTA.

INTERNAL BUFFERS

Figure 6 shows the internal circuit of each half of the LM13700 IC package. If this circuit is compared with that of the CA3080 shown last month, it will be seen to be broadly similar except for the addition of linearizing diodes D1-D2 to the inputs of the OTA's Q1-Q2 differential amplifier, and the addition of output transistors Q11-Q12, which are configured as a Darlington emitter follower buffer stage and can (by wiring its input to the OTA output and connecting Q12 emitter to the negative rail via a suitable load resistor) be used to make the high-impedance output of the OTA available at a low-impedance level. Note in this latter case that the output of the buffer stage is two base-emitter volt drops (about 1.2V) below the output voltage level of the OTA, so this buffer is not suited for use in high-precision DC amplifier applications.

The two OTAs of the LM13700 share common supply rails, but are otherwise fully independent. All elements are integrated on a single chip, and the OTAs have closely-matched characteristics (gm values are typically matched within 0.3dB), making the IC ideal for use in stereo VCA and VCF applications, etc. The standard commercial version of the LM13700 can be powered from split supply rails of up to $\pm 18V$, or single-ended supplies of up to 36V. In use, I_D and I_{bias} should be limited to 2mA maximum, and the output current of each buffer stage should be limited to 20mA maximum.

VCA CIRCUITS

Figure 7 shows a practical voltage-controlled amplifier (VCA) made from half of an LM13700 IC. Here, the input signal is fed to the non-inverting terminal of the OTA via current-limiting resistor R4, and the high-impedance output of the OTA is loaded by R5, which determines the peak (overload) amplitude of the output signal in the way described last month. The output signal is made available to the outside world at a low-impedance level via the buffer stage, which is loaded via R6.

The Figure 7 circuit is powered from dual 9V supplies. The I_D current is fixed at about 0.8mA via R1, but I_{bias} is variable via R7 and an external gain control voltage. When the gain-control voltage is at the negative rail value of -9V, I_{bias} is zero and the circuit gives an overall 'gain' of -80dB. When the gain-control is at the positive rail value of +9V, I_{bias} is about 0.8mA, and the circuit gives a voltage gain of

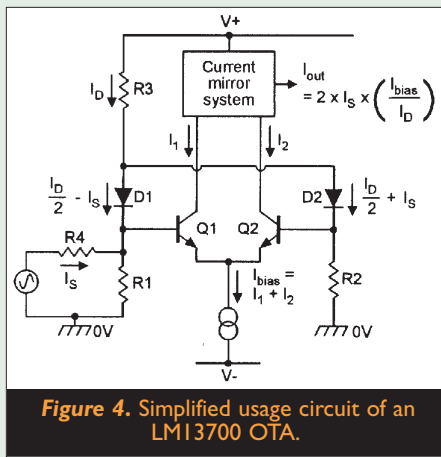


Figure 4. Simplified usage circuit of an LM13700 OTA.

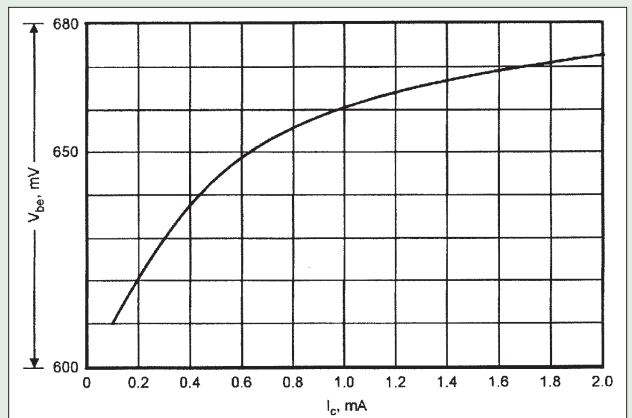


Figure 3. Typical transfer characteristics of a small-signal silicon transistor.

roughly $\times 1.5$. The voltage gain is fully variable within these two limits via the gain-control input. The two halves of the LM13700 have closely-matched characteristics, making the IC ideal for use in stereo amplifier applications. Figure 8 shows how two amplifiers of the Figure 7 type can be used together to make a voltage-controlled stereo amplifier. Note, in this case, that the I_{bias} gain-control pins of the two OTAs are shorted together and fed from a single gain-control voltage and current-limiting resistor. The close matching of the OTAs ensures that the gain-control currents divide equally between the two amplifiers.

Note that the Figure 7 and 8 circuits act as non-inverting amplifiers, since their input signals are fed to the non-inverting pins of the OTAs. They can be made to act as inverting amplifiers by simply feeding the input to the inverting pins of the OTAs.

The VCA circuit in Figure 7 can be used as an amplitude modulator or two-quadrant multiplier by feeding the carrier signal to the input terminal, and the modulation signal to the gain-control input terminal. If desired, the gain-control pin can be DC biased so that a carrier output is available with no AC input signal applied. Figure 9 shows a practical example of an inverting amplifier of this type. The AC modulation signal modulates the amplitude of the carrier output signal.

Figure 10 shows how half of an LM13700 can be used as a ring modulator or four-quadrant multiplier, in which zero carrier output is available when the modulation voltage is at zero (common supply rail) volts, but increases when the

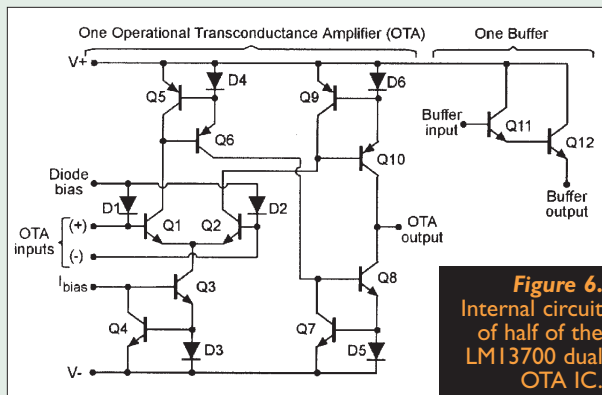


Figure 6. Internal circuit of half of the LM13700 dual OTA IC.

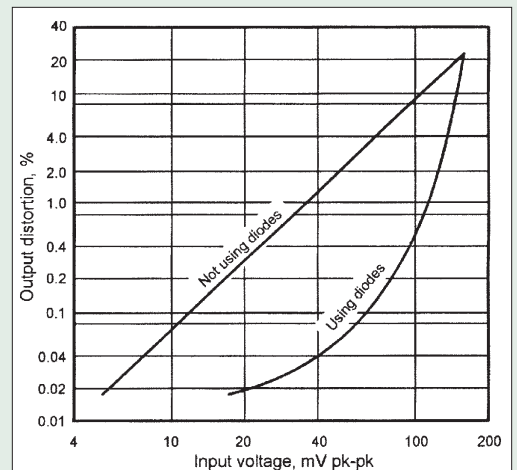


Figure 5. Typical distortion levels of the LM13700 OTA with and without the use of the linearizing diodes.

Figure 7.
Voltage-controlled amplifier (VCA).

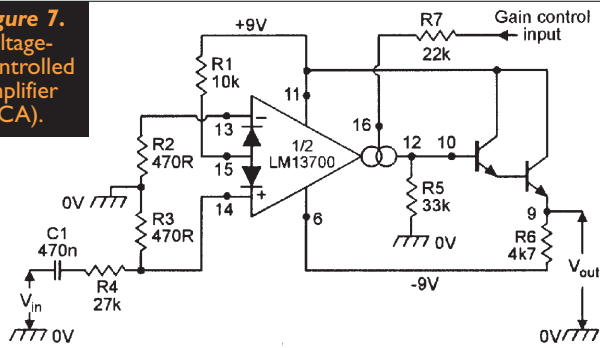


Figure 8.
Voltage-controlled stereo amplifier.

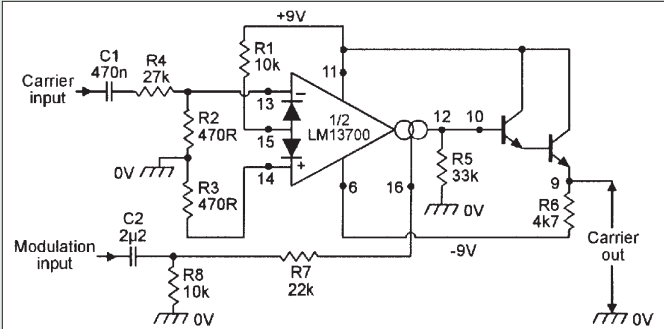
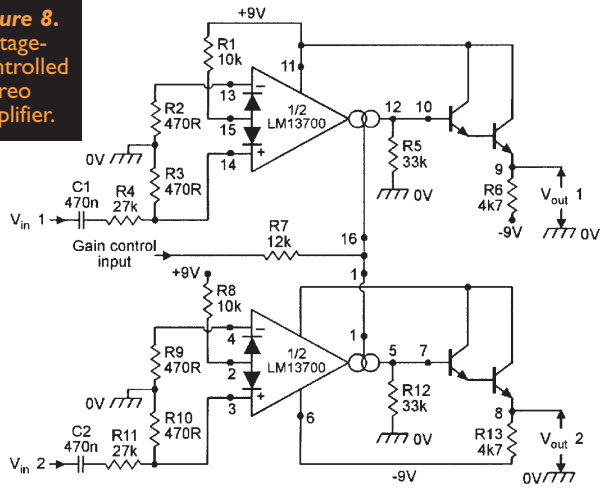


Figure 9. Amplitude modulator or two-quadrant multiplier.

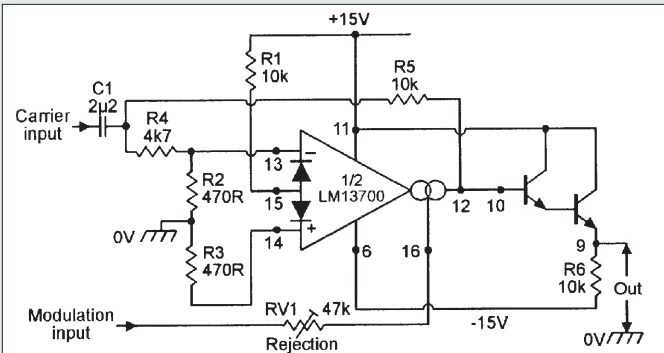


Figure 10. Ring modulator or four-quadrant multiplier.

modulation voltage moves positive or negative relative to zero. 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 is non-inverted.

The Figure 10 circuit is shown with values suitable for operation from dual 15V supplies, but is essentially similar to the Figure 9 circuit, except that R5 is connected between the input signal and the output of the OTA, and I_{bias} is “pre-settable” via RV1. The basic circuit action is such that the OTA feeds an inverted (relative to the input) signal current into the bottom of R5, and at the same time, the input signal feeds directly into the top of R5. RV1 is pre-set so that when the modulation input is tied to the zero volts common line, the overall gain of the OTA is such that its output current exactly balances (cancels) the direct-input current of R5, and under this condition, the circuit gives zero carrier output.

Consequently, when the modulation input goes positive, the OTA gain increases and its output signal exceeds that caused by the direct input into R5, so an inverted output carrier is generated. Conversely, when the modulation input goes negative, the OTA gain decreases and the direct signal of R5 exceeds the output of the OTA, and a non-inverted output signal is generated.

OFFSET BIASING

The circuits in Figures 7 to 10 are shown with OTA input biasing applied via 470R resistors wired between the two input terminals and the zero volts rail. In practice, this simple arrangement may cause the DC output level to shift slightly when the I_{bias} gain-control is varied from minimum value. If desired, this shifting can be eliminated by fitting the circuits with a pre-settable offset adjust control

as shown in Figure 11, enabling the biasing resistance values to be varied slightly. To adjust the offset biasing, reduce I_{bias} to zero, note the DC level of the OTA output, then increase I_{bias} to maximum and adjust RV1 to give the same DC output level.

AN AUTOMATIC GAIN CONTROL AMPLIFIER

In the Figures 7 to 10 circuits, the amplifier gain is varied by altering the I_{bias} value. A feature of the LM13700, however, is that its gain can be varied by altering either the I_{bias} or the I_D current, and Figure 12 shows how the I_D variation can be used to make an automatic gain control (AGC) amplifier in which a 100:1 change in input signal amplitude causes only a 5:1 change in output amplitude.

In this circuit, I_{bias} is fixed by R4, and the output signal is taken directly from the OTA via R5. The output buffer is used as a signal rectifier — fed from the output of the OTA — and the rectified output is smoothed via R6-C2, and used to apply the I_D current to the OTA's linearizing diodes. Note, however, that no significant I_D current is generated until the OTA output reaches a high enough amplitude ($3 \times V_{be}$, or about 1.8V peak) to turn on the Darlington buffer and the linearizing diodes, and that an increase in I_D reduces the OTA gain and — by negative feedback action — tends to hold V_{out} at that level.

The basic zero I_D gain of this amplifier is x40. Thus, with an input of 30mV peak-to-peak, the OTA output of 1.2V peak-to-peak is not enough to generate an I_D current, so the OTA operates at full gain. At 300mV input, however, the OTA output is enough to generate significant I_D current, and the circuit's negative feedback automatically reduces the output level to 3V6 peak-to-peak, giving an overall gain of x 11.7. With an input of 3V, the gain falls to x2, giving an output of 6V peak-to-peak. The circuit, thus, gives 20:1 signal compression over this range.

VOLTAGE-CONTROLLED RESISTORS

An unusual application of the LM13700 is as a voltage-controlled resistor (VCR), using the basic circuit in Figure 13. The basic theory here is quite simple – if an AC signal is applied to the R_x terminals, it will feed to the OTA's inverting terminal via C1 and the buffer stage and the R/RA attenuator, and the OTA will then generate an output current proportional to the V_{in} and I_{bias} values. Thus, since $R = E/I$, the circuit's R_x terminal acts like an AC resistor with a value determined by I_{bias} .

The effective resistance value of the R_x terminal actually equals $(R + RA)/(gm \times RA)$, where gm is roughly $20 \times I_{bias}$. This formula approximates to $R_x = R/(I_{bias} \times 20RA)$, so, using the component values shown in the diagram, R_x equals roughly $10M$ at an I_{bias} value of $1\mu A$, and $10k$ at an I_{bias} of $1mA$. Figure 14 shows a similar version of the VCR, where the linearizing diodes are used to effectively improve the noise performance of the resistor, and Figure 15 shows how a pair of these circuits can be used to make a floating VCR in which the input voltage is direct-coupled and may be at any value within the output voltage range of the LM13700.

VOLTAGE-CONTROLLED FILTERS

An OTA acts basically as a voltage-controlled AC current source, in which an AC voltage is applied to the amplifier's input, and the magnitude of the output current depends on the value of I_{bias} . This fact can be used to implement a voltage-controlled low-pass filter by using half of an LM13700 in the configuration shown in Figure 16, in which the values of R , C , and I_{bias} control the cut-off frequency, f_c , of the filter. The operating theory of this circuit is as follows.

Assume, initially, that capacitor C is removed from the circuit. The input signal is applied to the OTA's non-inverting terminal via potential divider $R1-R2$, and the OTA's output is followed by the buffer stage and fed back to the inverting input via an identical divider made up of R and RA . The basic OTA thus acts as a non-inverting amplifier with a gain of R/RA but, since the input signal is fed to the OTA via a potential divider with a value equal to R/RA , the circuit acts overall as a unity-gain voltage follower.

Assume now that capacitor C is fit into place. At low frequencies, C has a very high impedance and the OTA output current is able to fully charge it, causing the circuit to act as a voltage follower in the way already described.

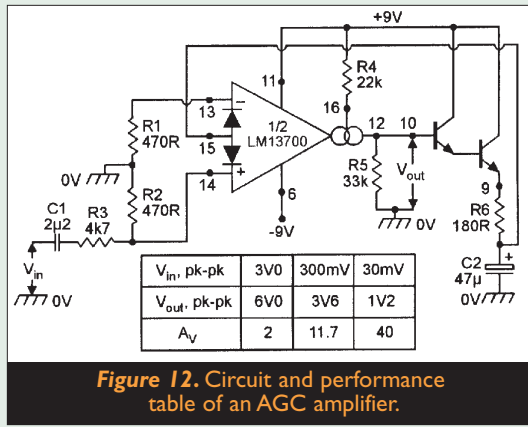


Figure 12. Circuit and performance table of an AGC amplifier.

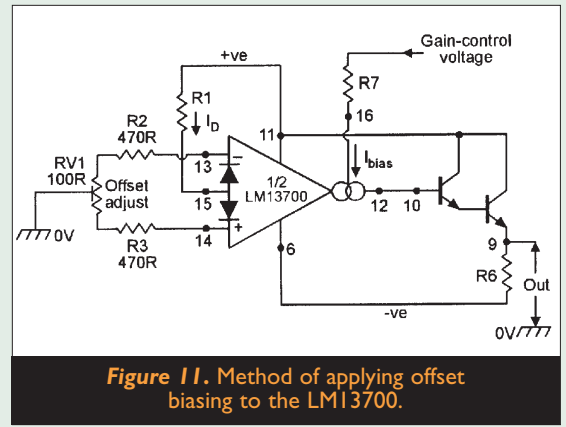


Figure 11. Method of applying offset biasing to the LM13700.

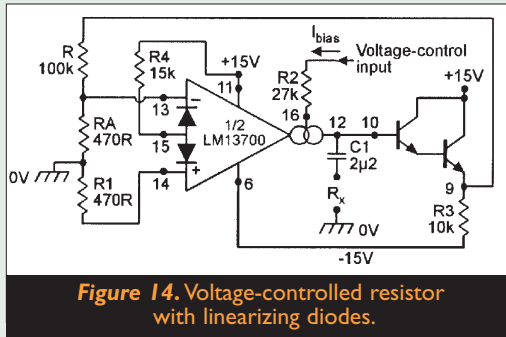


Figure 14. Voltage-controlled resistor with linearizing diodes.

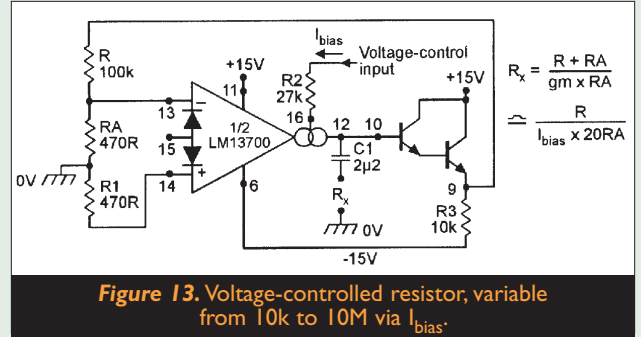


Figure 13. Voltage-controlled resistor, variable from 10k to 10M via I_{bias} .

As the frequency increases, however, the impedance (X) of C decreases and the OTA output current is no longer able to fully charge C , and the output signal starts to attenuate at a rate of 6dB per octave. The cut-off point of the circuit – at which the output falls by 3dB – occurs when $XC/20 \times I_{bias}$ equals R/RA , as implied by the formula in the diagram. With the component values shown, cut-off occurs at about 45Hz at an I_{bias} value of $1\mu A$, and at 45kHz at an I_{bias} value of $1mA$. A similar principle to the above can be used to make a voltage-controlled high-pass filter, as shown in Figure 17. This particular circuit has, with the values shown, cut-off frequencies of 6Hz and 6kHz at I_{bias} currents of $1\mu A$ and $1mA$, respectively.

Numbers of filter stages can easily be cascaded to make multi-pole voltage-controlled filters. The excellent tracking of the two sections of the LM13700, make it possible to voltage-control these filters over several decades of frequency. Figure 18 shows the practical circuit of a two-pole (12dB per octave) Butterworth low-pass filter having cut-off frequencies of 60Hz and 60kHz at I_{bias} currents of $1\mu A$ and $1mA$, respectively.

VOLTAGE CONTROLLED OSCILLATORS

To conclude this look at applications of the LM13700,

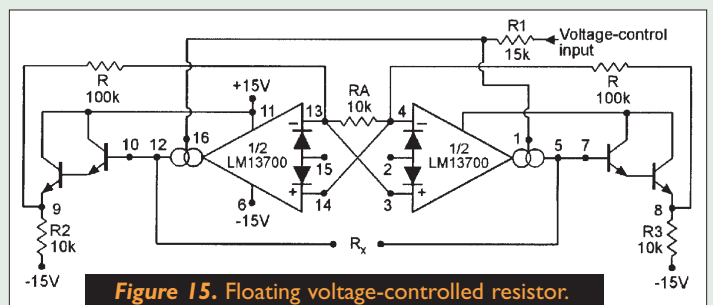


Figure 15. Floating voltage-controlled resistor.

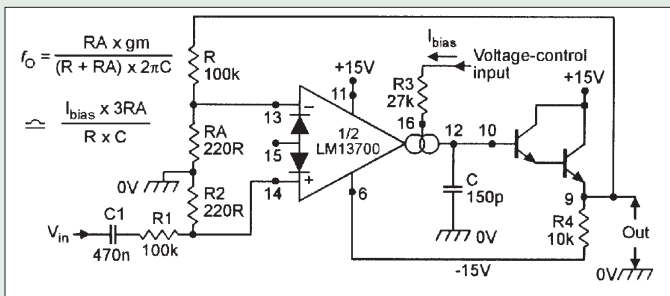


Figure 16. Voltage-controlled, low-pass filter covering 45Hz to 45kHz.

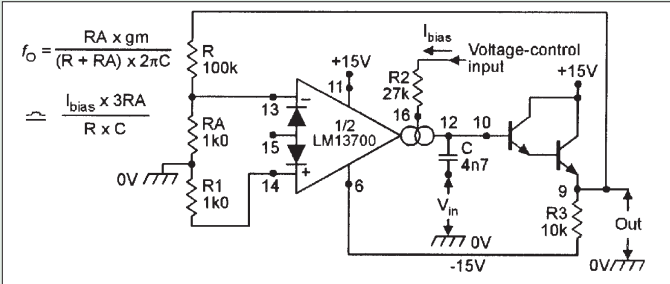


Figure 17. Voltage-controlled, high-pass filter covering 6Hz to 6kHz.

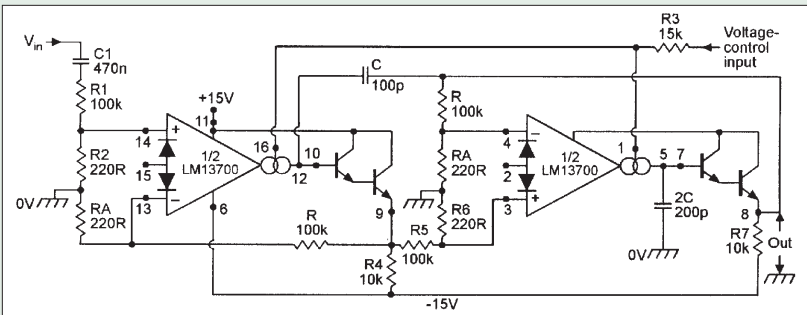


Figure 18. Voltage-controlled two-pole Butterworth low-pass filter covering 60Hz to 60kHz.

Figures 19 and 20 show two ways of using the IC as a voltage-controlled oscillator (VCO). The Figure 19 circuit uses

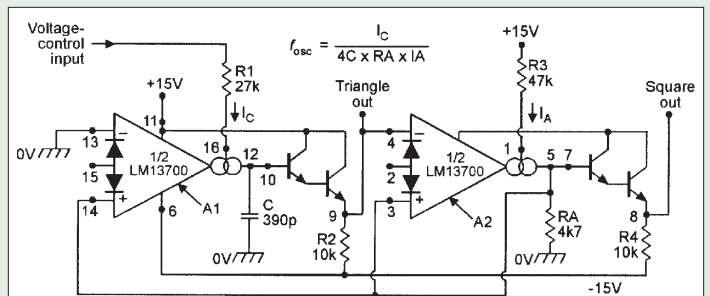


Figure 19. Triangle/squarewave VCO covering 200Hz to 200kHz.

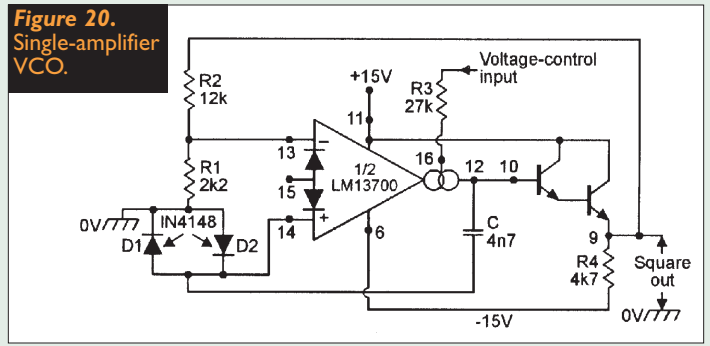


Figure 20. Single-amplifier VCO.

both halves of the LM13700, and simultaneously generates both triangle and squarewaves. The Figure 20 design uses only half of the IC, and generates squarewaves only.

To understand the operating theory of the Figure 19 circuit, assume initially that capacitor C is negatively charged and that the squarewave output signal has just switched high. Under this condition, a positive voltage is developed across RA and is fed to the non-inverting terminals of the two amplifiers, which are both wired in the voltage comparator modes.

This voltage makes amplifier A1 generate a positive output current equal to the bias current, I_C , and this flows into capacitor C, which generates a positive-going linear ramp voltage that is fed to the inverting terminal of A2 via the Darlington buffer stage until, eventually, this voltage equals that on the non-inverting terminal, at which point the output of A2 starts to swing negative. This initiates a regenerative switching action in which the squarewave output terminal switches abruptly negative.

Under this condition, a negative voltage is generated across resistor RA, causing amplifier A1 to generate a negative output current equal to I_C . This current causes capacitor C to discharge linearly until, eventually, its voltage falls to a value equal to that of RA, at which point the squarewave output switches high again. This process repeats *ad infinitum*, causing a triangle waveform to be generated on R2 and a squarewave output to be generated on R4.

The waveform frequency is variable via the voltage-control input, which controls the value of I_C . With the component values shown, the circuit generates a frequency of about 200Hz at an I_C current of 1 μ A and 200kHz at a current of 1mA.

Finally, Figure 20 shows a single-amplifier VCO circuit which generates a squarewave output only. The circuit operates in a similar manner to that described above, except that 'C' charges via D1 and discharges via D2, which generates a 'polarity' signal on the non-inverting terminal of the amplifier. **NV**