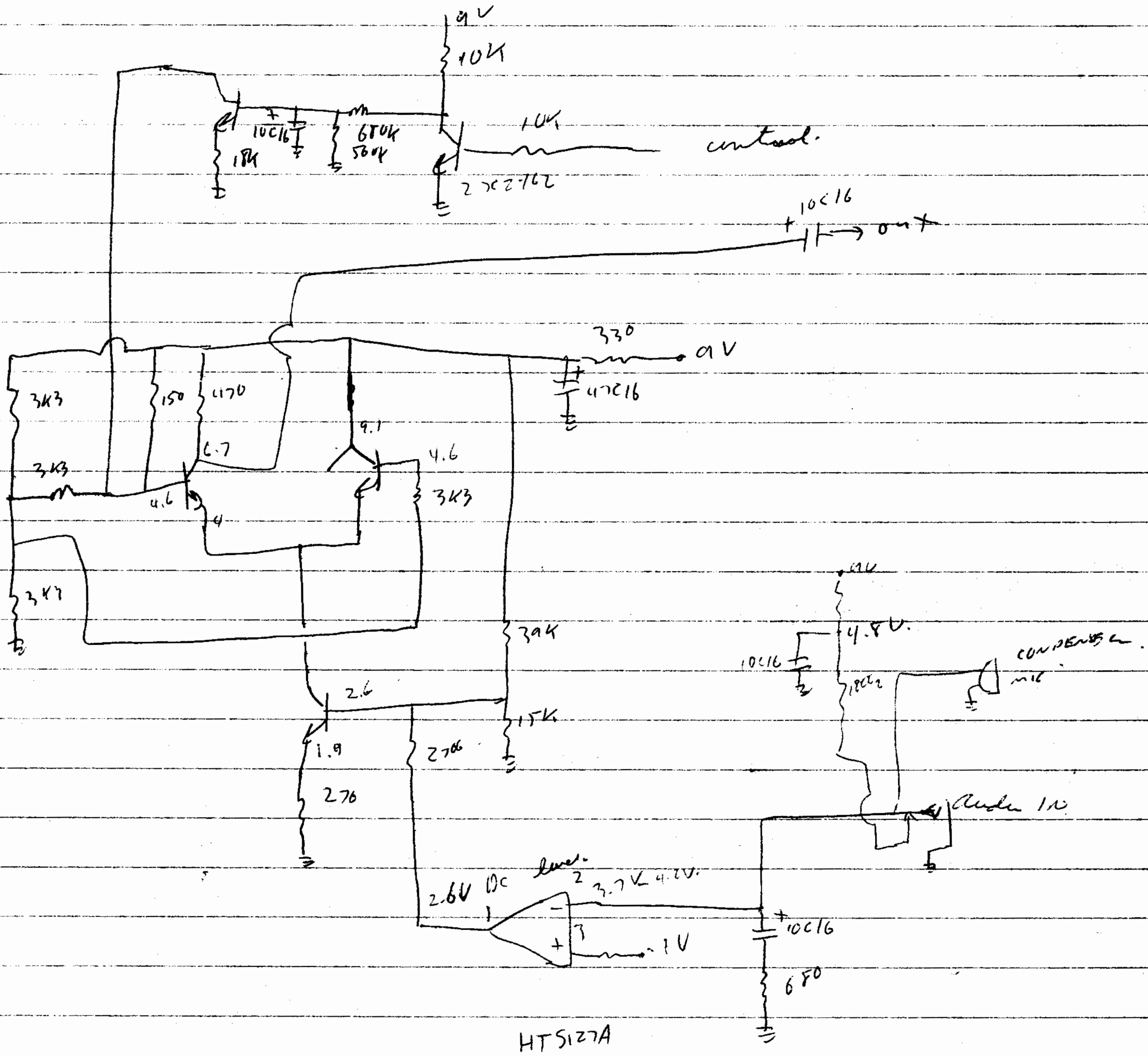
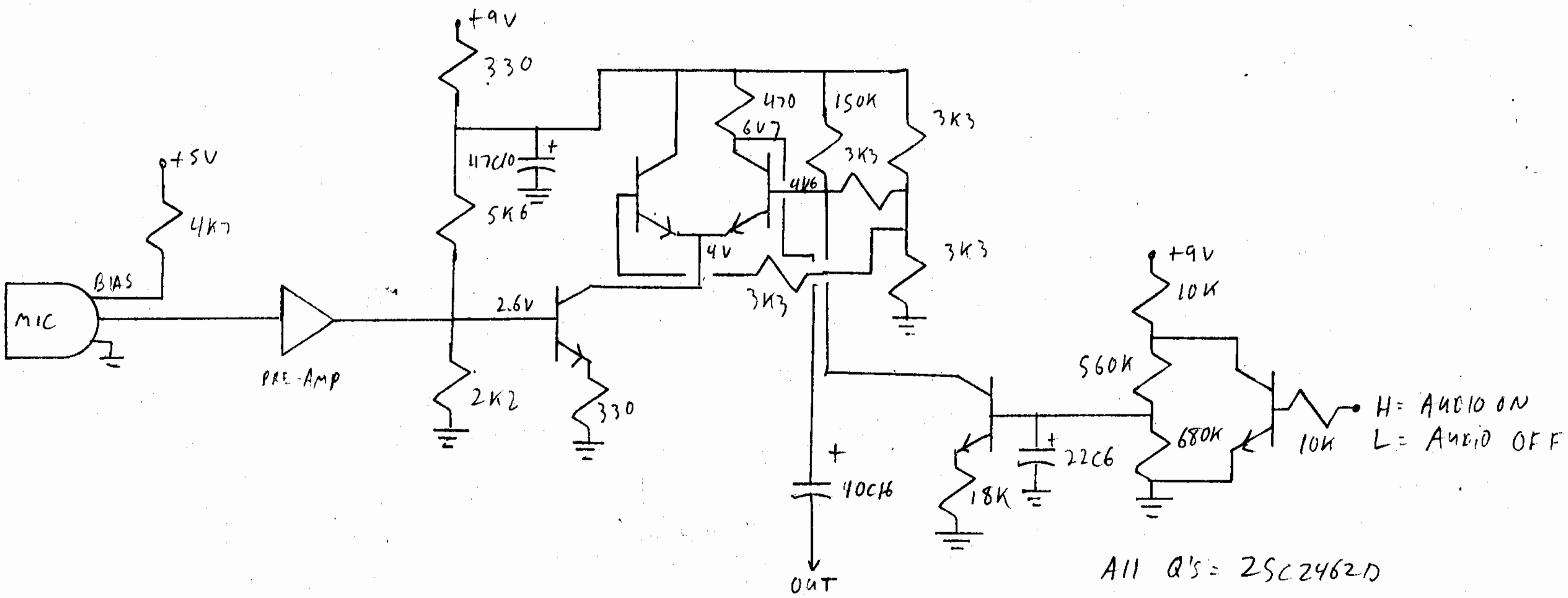
Audiofade circuit CC017.

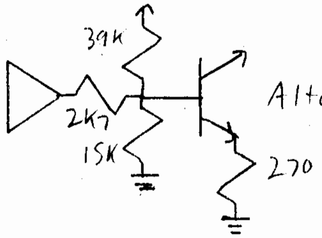
all 2SC2462





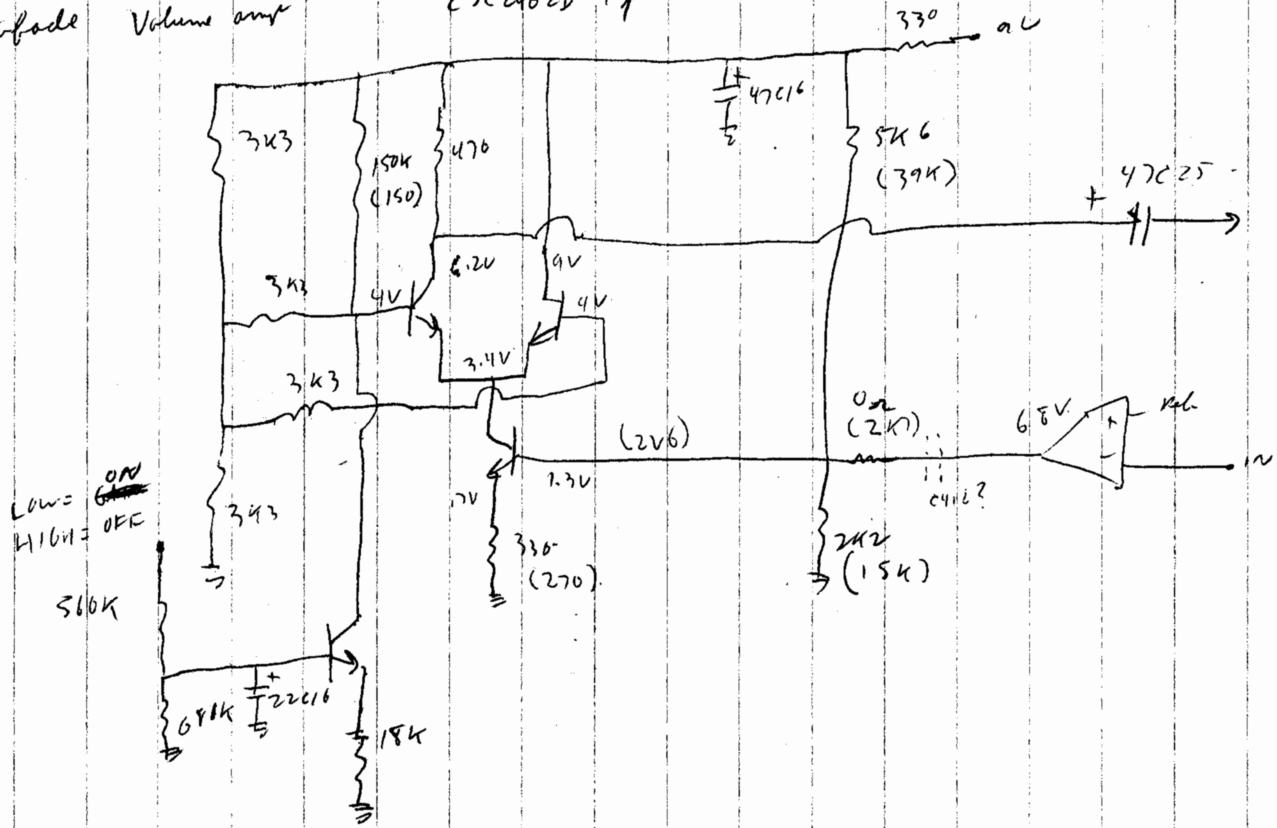
All Q's = 2SC2462D

CC017/CC030 Auto fade circuit



Alternate Configuration

Autofade Volume amp 2SC2462D X4



THE 1537 VCA

Our Stereo Image Coordinator made use of a useful device. The 1537A offers some very impressive specs. Keith Brindley explains.

THERE IS always a great deal of excitement generated in electronics on the arrival or introduction of a new circuit, concept or chip, particularly if the system is potentially a field leader. The 1537A chip is just that! The specifications which the device can offer in situ are well above those of any similar preceding systems. Table 1 gives a listing of specifications, which can be obtained in the correct applications.

Parameter	Specification
Bandwidth	DC-200kHz
T.H.D., 20Hz-20kHz	0.004%
I.M.D. (SMPTE TEST)	0.03%
Noise	-90dBv, ± 1dB (worst case, unity gain)
Overshoot and Ringing	None
Slew Rate	> 10v/usec, symmetrical & constant
Input Impedance	20KΩ
Maximum Input Level	+20dBv
Gain	0dB (Unity)
Maximum Attenuation	>94dB
Control Voltage	0 to +10V
DC shift vs. Attenuation	≤ 5mV
Power Requirements	Regulated ± 15V at +25, -33mA

Table 1. The maximum possible specifications available from a 1537A system.

With harmonic distortion of 0.004% and a signal/noise ratio of over 90 dB the system is of course well suited to studio applications, although use in this environment is by no means its only area of involvement. The IC itself seems at first glance, somewhat highly priced at around $22, but nevertheless, it requires few extra components to produce a VCA system of the superb quality (suggested in the specifications of Table 1) and overall represents good value for money to the amatuer and professional engineer alike.

Amplifier Or Attenuator

The term VCA is normally used as an abbreviation of the phrase Voltage Controlled Amplifier, but in its simpler modes the 1537A is, strictly speaking, a voltage controlled attenuator ie with a maximum gain of unity. The inventors do, however, stress that connection of the 1537A into the feedback loop of an amplifier (such as an op amp) produces a voltage controlled amplifier. The applications section of this article show how this can be achieved.

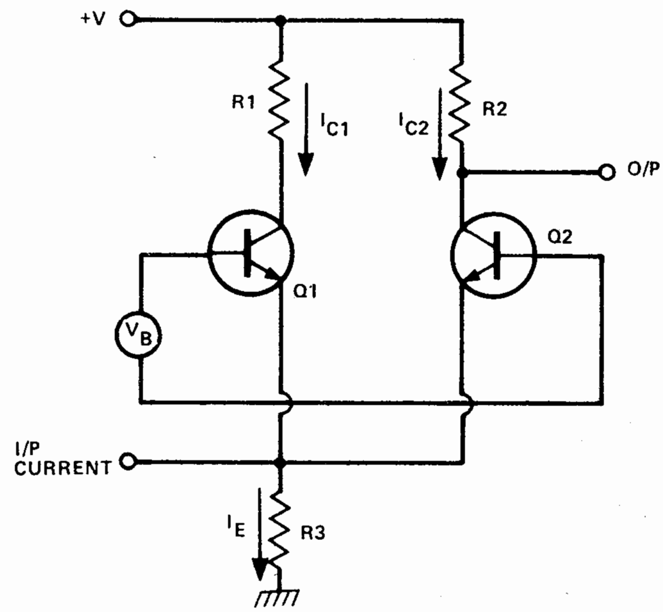


Fig. 1. A differential pair of transistors — the basis of a VCA.

The operation of the 1537A VCA depends upon the gain control function of a differential pair of transistors as in Fig. 1. The transistors in Fig. 1 are connected at thei. emitters. The current through R3 (I_E) is, therefore, approximately equal to the sum of their two collector currents I_{C1} and I_{C2} through R1 and R2 respectively. The relative bias voltage, V_B, between the two bases determines the relative collector currents. If we now apply an input signal current to the joined emitters we obtain output signal currents through R1 and R2, the sizes of which are determined by the bias voltages. In other words, by altering this bias voltage we alter the size of the output signals.

Figure 2 shows a simplified internal circuit of the 1537A chip giving pin numbers and external load and emitter resistors necessary for operation. There are two basic gain control circuits within the chip, similar to that in Fig. 1 (built around Q1, 2 and Q5, 6) except for three main differences:
— the diode connection of the transistor pair not used for signal output ie Q1 and Q6, which reduces the distortion due to transistor gain differences.
— the addition of buffers around Q4 and Q8 to reduce loading of the output collectors of the gain transistors, in turn allowing idealised characteristics over the full gain range.
— the use of transistors Q3 and Q7 as voltage to current converters enabling the input to be applied as a voltage rather than as a current.

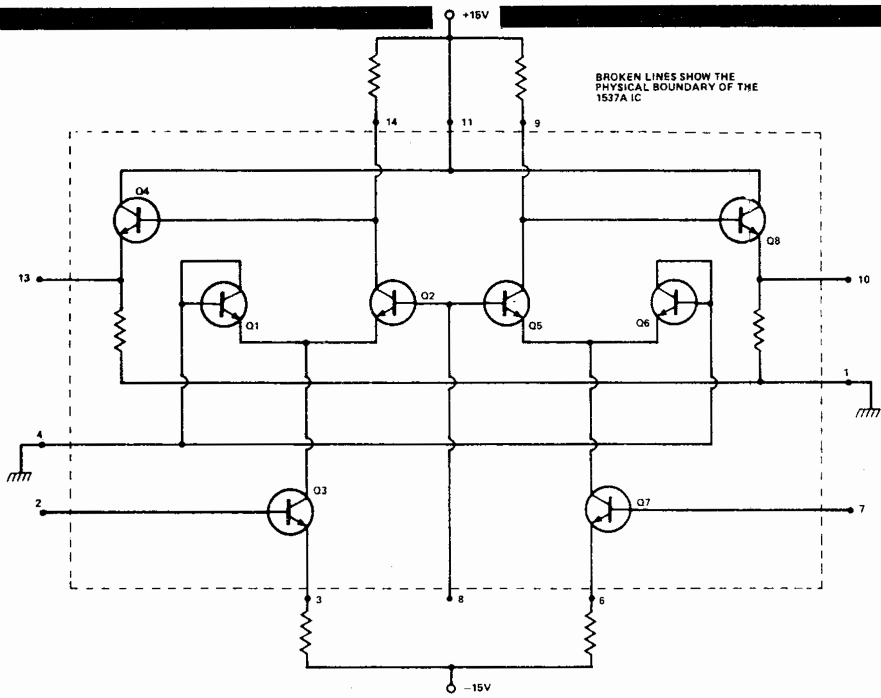
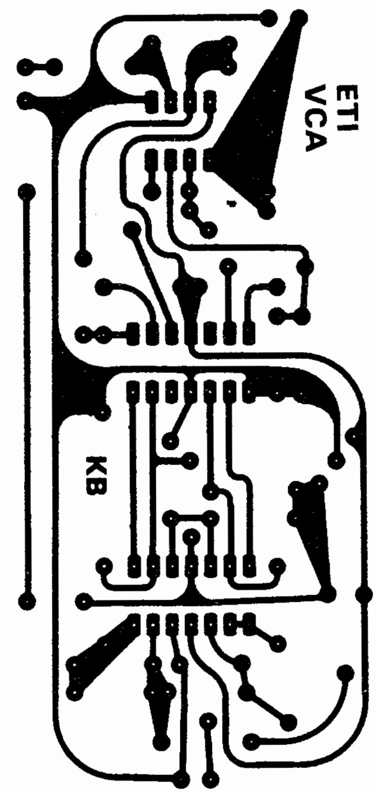


Fig. 2. A much simplified internal circuit of the 1537A IC, showing external load and emitter resistors.



There is, however, a much more subtle difference, on top of this and that is the use of large geometry transistors. The effect of larger geometry transistors can improve second order intermodulation by as much as ten times for a tenfold increase in transistor size. Noise can also be reduced by about 10 dB for a similar increase in geometry.

This leads us now to the simplest mode of operation of the 1537A using each gain control circuit individually, although the control voltage affects the gain of each circuit simultaneously (Fig. 3).

The ratio of R9 and R10 is calculated to allow a control voltage range of 10 volts (ie 0 to minus 10 V), altering the gain of the system from 0 dB to about —90 dB. The input impedance of the circuit to applied signal is low and ideally buffers should be placed before this circuit. Although this circuit does not give studio quality specifications it will, however, still produce results in the "high fidelity" range, providing impedance matches are considered.

Figure 4 shows a circuit application which gives a higher impedance input. Also included is an inverting stage in the control voltage link which allows a voltage of 0 to + 10 volts to be used for controlling attenuation.

Although any operational amplifier could be used for ICs 1, 2 and 3 in the previous circuit, it should be fairly apparent that the noise, distortion and bandwidth specs of the circuit are limited to those of the op amps used.

Either of the two circuits of Figs. 1 and 2 can be adopted as the voltage controlled gain heart of a stereo system. Their outputs are about 10 dB down on the inputs so necessary amplification should be given before or after the attenuator.

Coming Up To Scratch

Now, three more developments to the circuitry can be undertaken to improve the specifications to those of Table 1. Figure 5 shows the circuit of the ideal system capable of these high specs.

Firstly, actively linearised voltage to current sources (op amp 3 and 4 in Fig 5) improve distortion figures when using a wide range of input signal voltages.

Secondly, parallelling of the two individual gain control circuits (ie the same input signal is fed to both devices at their inputs and mixed at their outputs) gives a 3 dB improvement in S/N ratio.

Finally, a technique is utilised which is complementary to the previous development of parallel devices, whereby the same input is applied to both gain control devices but 180 degrees out of phase. The two outputs are combined in a differential amplifier to give a single ended output. The differential amp is formed around op amp 6. This technique has the effect of reducing DC shift caused by bias and control voltages and with careful adjustment of RV1, the minimal DC shift now left at the output can be reduced even.

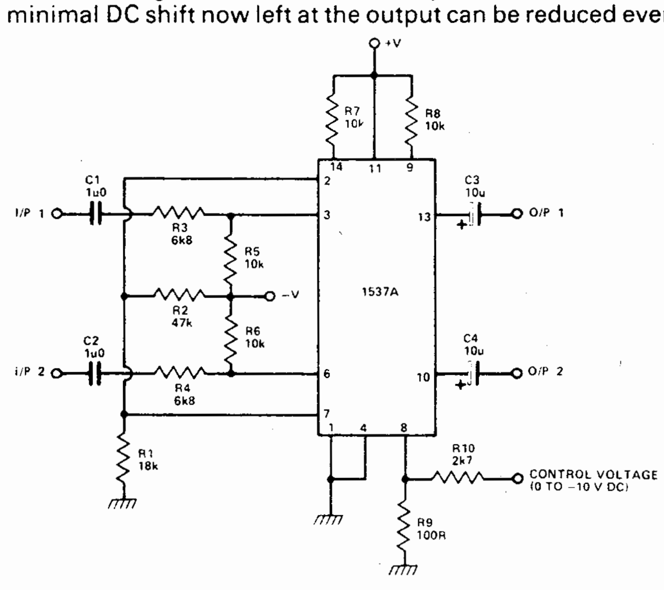
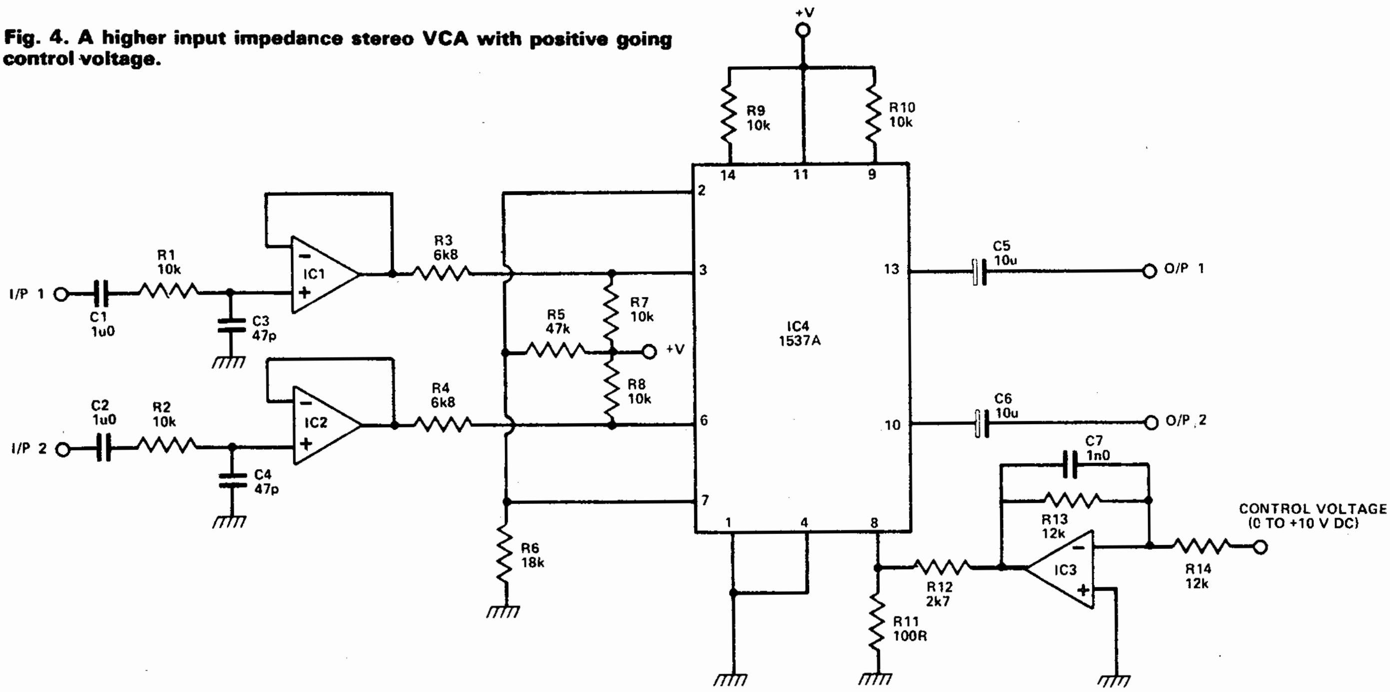


Fig. 3. The simplest mode of operation of the 1537A — a low input impedance stereo VCA with a negative going control voltage.

Fig. 4. A higher input impedance stereo VCA with positive going control voltage.



further to near (if not actually) zero. The prototype circuit shown, upon testing, actually gave no DC shift at all (or at least none measureable on our test equipment).

The complete circuit can be used as an exceptionally high quality VCA whose signal input can be anything from a few millivolts through to about 20 volts pk to pk without distortion. The lack of DC blocking capacitors at the input and output means that the system can be used to control a DC voltage applied to the input. AC signals up to well over 200 kHz are easily catered for, due to the system's wide bandwidth.

The overlay in figure 6 shows the component layout on printed circuit board of the circuit. As far as we know this article is the first of its kind to present a circuit in a form where "experimenters" can benefit easily and directly from the written text while simultaneously using the device in a tried and tested form.

Construction

If the circuit board layout is followed then there should be no problems. IC holders are advisable though by no means necessary. RV1 should be a good quality type (cermet), to assist in setting up the output offset shift to zero, cheaper quality presets can sometimes be tricky to adjust in low voltage DC applications of this nature. Op amps 1 to 4 in the circuit are combined in IC1 and can be of a wide range of types from a quad 741 type (3403) upwards. Obviously, if you wish to obtain the best specs the quality of the op amps are critical. LF 347 or TL 074 will give the best results.

Similarly op amps 5 and 6 are included in IC3 and LF 353 or TL 072 are of optimal quality.

Setting Up

The system should work without any adjustment for an AC signal and varying the control voltage from 0 to 10 volts should give total control over the output amplitude. Some setting up will be required if the input is to be DC, though. This is best achieved by earthing the input. Measure the output voltage using a high impedance voltmeter (it should only be the order of a few millivolts). Adjust RV1 until a complete sweep of control voltage ie from 0 to 10 volts produces only minimal change in DC output voltage. The circuit is now completely set up to accept an input signal in the frequency range DC to 200 kHz. At minimum attenuation the system operates as a unity-gain wide range, high quality buffer, with a reasonably high input impedance and low output impedance. Variation of the DC control voltage over the range 0 to 10 volts will produce over 90 dB of attenuation of the output signal.

If an overall gain is required in the circuit, resistors R18 and R20 can be changed as in Table 2.

GAIN	R18 & R20
0dB	10k
6dB	22k
10dB	33k
15dB	56k

Table 2. The values of R18 and R20 to give the required overall gain in the VCA system of Fig. 5.

The control voltage range of 10 volts can be altered as required simply by changing the ratio of resistors R13 and R14 to suit.

To our knowledge, there is no officially recognised standard symbol for a VCA and rather than redraw the whole circuit of figure 5 upon every reference to the circuit we thought it better to invent a symbol for the purposes of this article. A horizontal trapezoid shape appeared to be the ideal symbol, as shown in Fig. 7. It symbolizes the system as a modular buffer amplifier, whose output (symbolized by the top line), decreases as the control voltage (the bottom line). increases. We shall use the modular sumbol of a VCA whenever reference is made to the circuit of Fig. 5, although any VCA module of another design should function in the applications which we give.

Fig. 5. Full specification mono VCA (showing component numbers and values of a practical circuit).

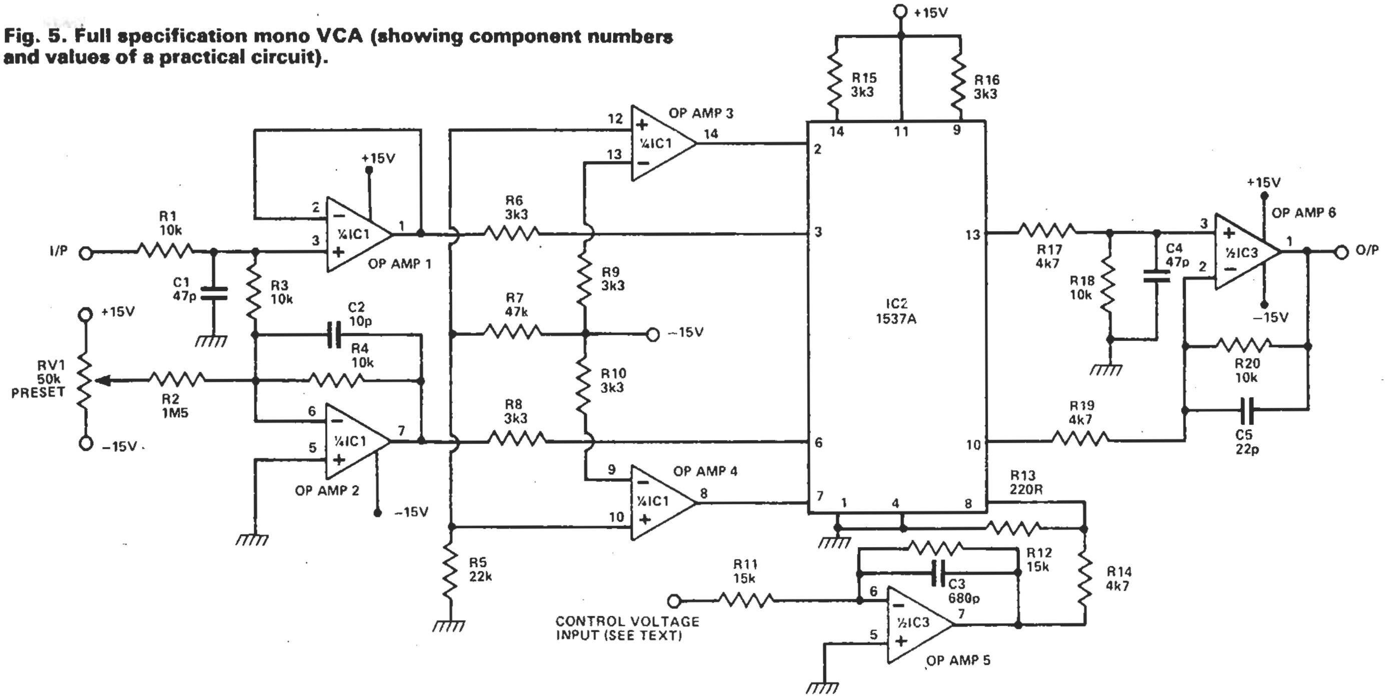
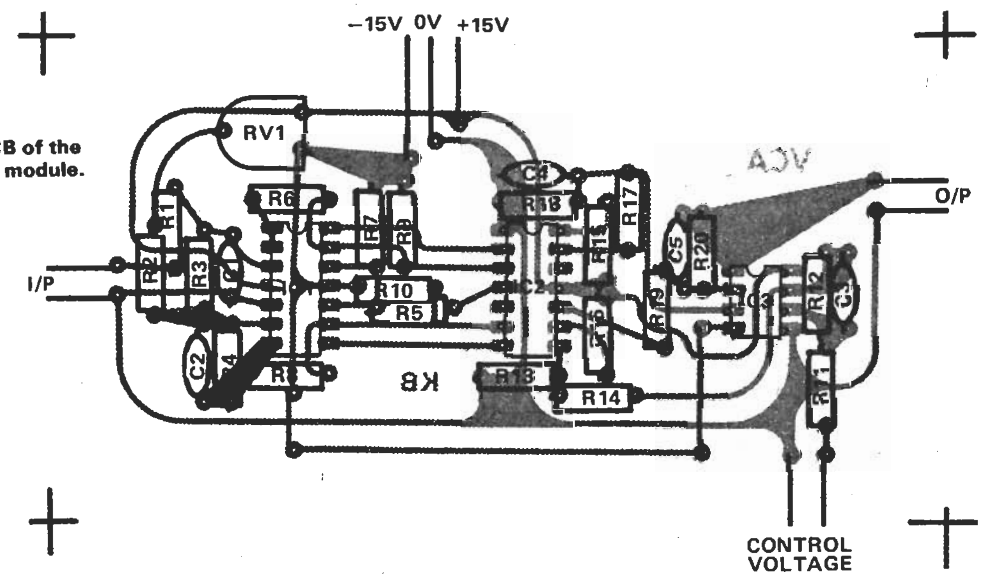


Fig. 6. Overlay of PCB of the 1537A VCA module.



Use of the 1537A system module as a DC controlled analogue gate can produce many effects. Amplitude modulation of the signal occurs and the usual associated effects are observed. For instance, in Fig. 8 we can see a simple but high quality tremelo unit. Transistors Q1 and Q2 are connected as a phase shift oscillator and buffer, with speed and depth controls whose varying DC output is connected directly to the control port of the 1537A module. The frequency range of the oscillator is approximately 2 to 5 Hz. Altering the values of all three capacitors will change the main frequency, though that stated will give the best results.

The control voltage in the last application was varied as a sine wave of course, but there is no reason why other waveforms eg square, could not be used for control purposes. Figure 9 shows a 555 operating in the astable

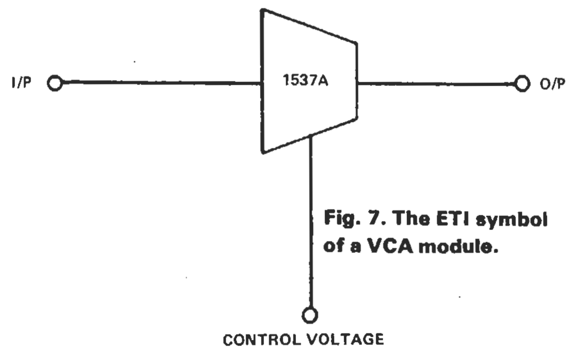


Fig. 7. The ETI symbol of a VCA module.

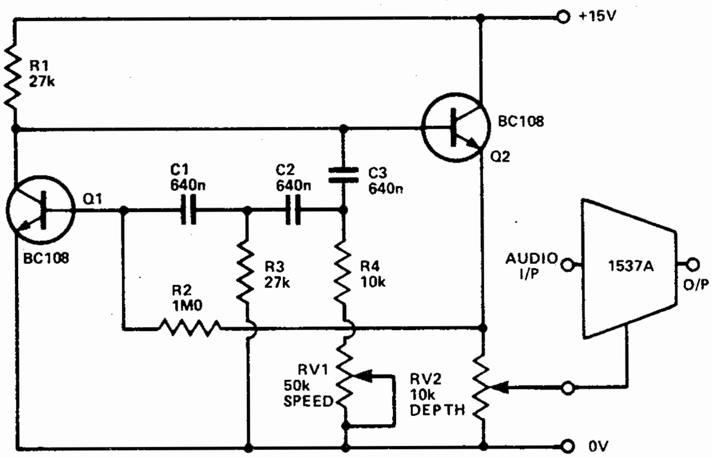


Fig. 8. A simple tremolo circuit.

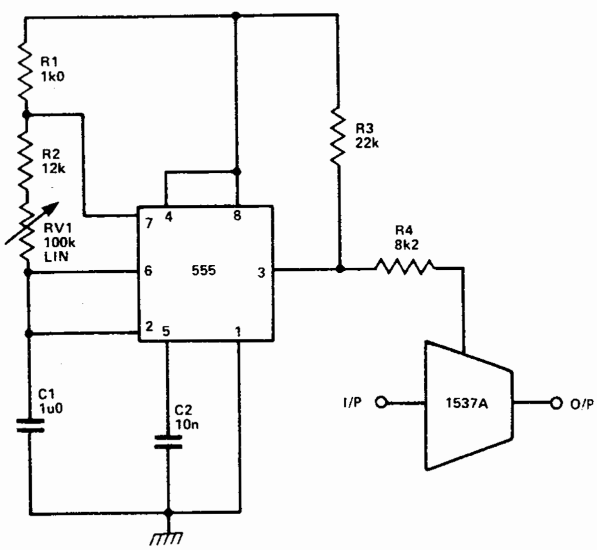


Fig. 9. Ring modulator.

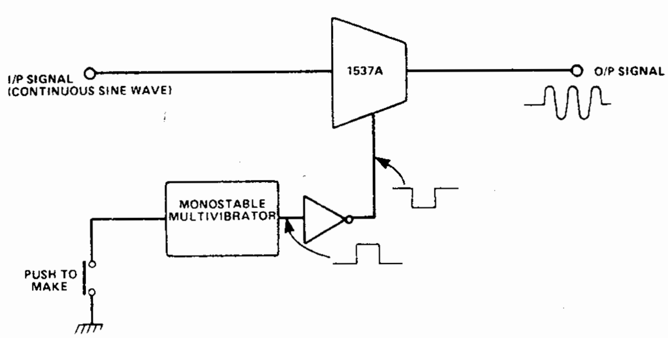


Fig. 10. A simple system enabling the construction of a tone burst generator.

mode with a frequency range of approximately 5-50 Hz. The output signal will be modulated with the square wave and the overall product is a computerized type sound if a vocal signal is applied to the 1537A module.

This square wave control can be taken one stage further if the control voltage is the output from a monostable as in Fig. 10. A tone burst generator can be very easily constructed with this mode of operation. In a tone burst generator, a rectangular envelope 50-500 uS long is formed around a single sine wave frequency of normally 1 kHz. Tone burst generators are useful for testing the transient response of speakers, A push to make switch is used to provide the trigger to fire the multivibrator, producing the correct length pulse which in turn is inverted to form the control voltage pulse, applied to the control port of the 1537A.

The previous applications have all used automatic waveform control of the applied signal to produce the required attenuation characteristics, but this is not a necessary trait. The control voltage can be simply tapped off a variable resistor having the maximum control voltage range (ie 10 volts) across it. In this way, altering the position of the wiper alters the attenuation of the applied signal. The pot acts quite simply as a volume or level control. Ordinary non-DC volume controls can suffer from pick-up problems because the signal itself is being rotated through the pot. As only DC is applied to the pot in this application no pick-up can occur and the control can be remotely mounted from the module with no screened cable being necessary. Figure 11 shows such a volume control.

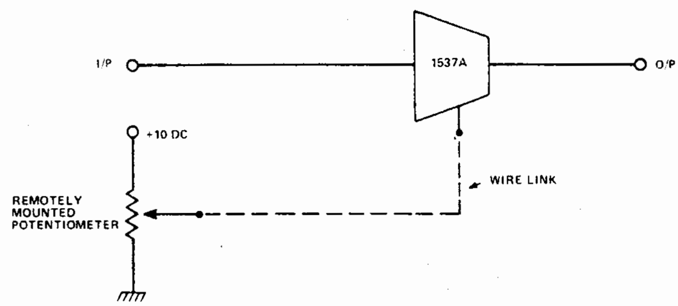


Fig. 11. Remotely (wire-linked) controlled volume control.

This remote control facility can be utilised in an audio mixer which includes remote faders for each channel. Figure 12 shows the general idea of such a circuit, An op amp is used as a summing amplifier into which the output of each channel's VCA is fed and mixed. The mix is relative to the control voltage applied from the remote faders to each VCA. The circuit allows for up to N inputs, where N to practical limits will probably be a maximum of about 12, but with careful layout techniques, there is no reason why this cannot be increased further.

Figure 13, shows an interesting outline to enable digital control of the VCA, say from a computer link. In order that the computer can operate in real-time, ie control of the VCA is not just its only job, it is necessary for the interface to provide a latch for the digital word. The output of this latch is changed to a linear DC voltage by the D/A (digital to analogue) convertor whose output is taken to the control port of the VCA.

The digital latch, once set by a strobe pulse, provides the facility that after the volume required has been found, the computer is free to perform other tasks, When the volume is to be altered, the latch is reset to the new digital input.

The last six applications of the 1537A VCA system have simply shown methods of providing a control voltage (automatically, manually or digitally) to control the module in its function as an analogue gate. The following section begins with the assumption that the control voltage is already present, perhaps by one of the previous methods.

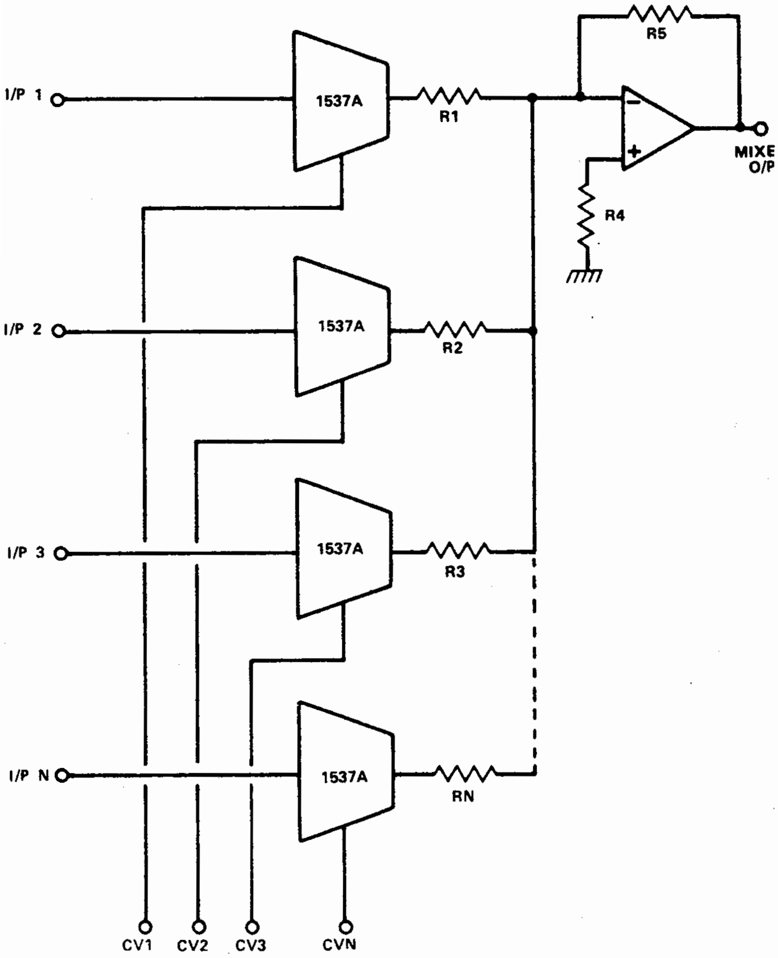


Fig. 12. High quality, remote fader controlled mixer.

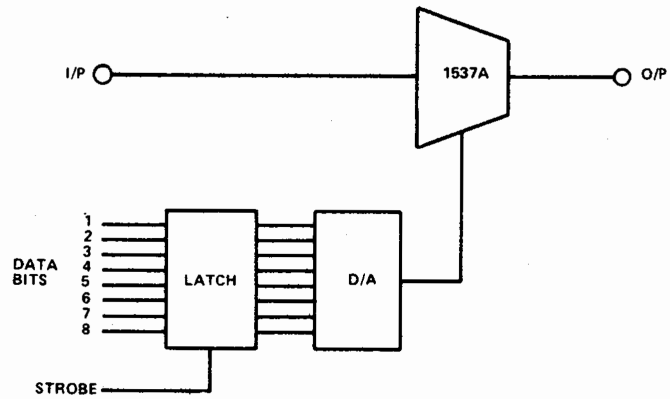


Fig. 13. Main components of a digitally controlled attenuator.

PARTS LIST

RESISTORS ALL ¼W, 5%

R1,3,4,18 20	10K
R2	1M5
R5	22k
R6,8,9,10 15,16	3k3
R7	47k
R11,12	15k
R13	220R
R14,17,19	4k7

PRESET

RV1	50k min horiz cermet

CAPACITORS

C1,4	47p polystyrene
C2	10p polystyrene
C3	680p polystyrene
C5	22p polystyrene

SEMICONDUCTORS

IC1	TL074, LF347 etc.
IC2	1537A
IC3	TL072, LF353 etc.

MISCELLANEOUS

IC Holders
PCB

Applications

Consequently the next few circuits show the system in a much more versatile role — not just as an analogue gate, but one in where the system itself becomes part of a larger system. Figures 14 and 15 give details of circuit in which the 1537A module is used in the feedback loop of conventional operational amplifiers to allow voltage controlled amplifiers to be constructed. The resistance values used give gains of approximately 1 to 100 over the VCA control voltage range and an inverting VCAmp and a non-inverting VCAmp can be easily built as shown.

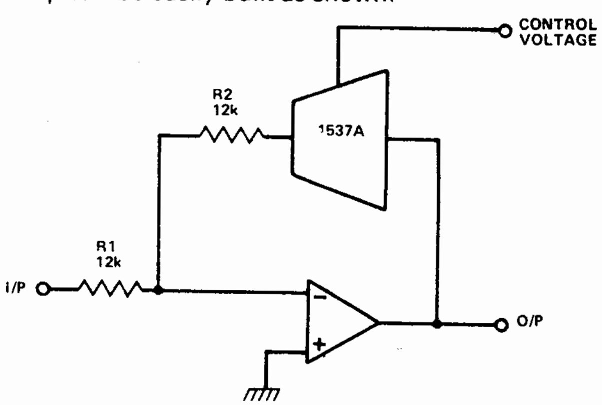


Fig. 14. A non-inverting controlled attenuator.

BUYLINES

The Aphex 1537A is available only from Octopus Audio, Suite 315, 69 Sherbourne St, Toronto, Ontario M5A 3X7. Cost is $22.00 each postpaid (Ontario residents add 7% P.S.T.).

Note the Motorola MC1537 and its second source variants will not work. The Motorola device is a dual 709 op amp.

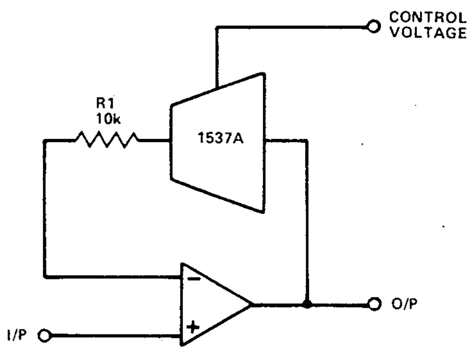


Fig. 15. An inverting voltage controlled amplifier.

A voltage controlled resistor is shown in the application of figure 16. The apparent resistance, R1, is given approximately by the formula

$$\frac{R1}{1-A}$$

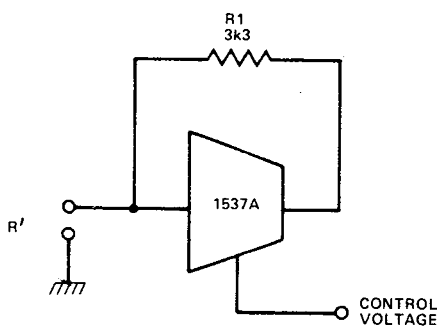


Fig. 16. A voltage controlled variable resistor.

where A is the gain of the VCA module (remembering that it has a maximum gain of unity). The value of R1 shown gives an apparent voltage controlled resistance of 7 k to 100 k over the ten volt control voltage range.

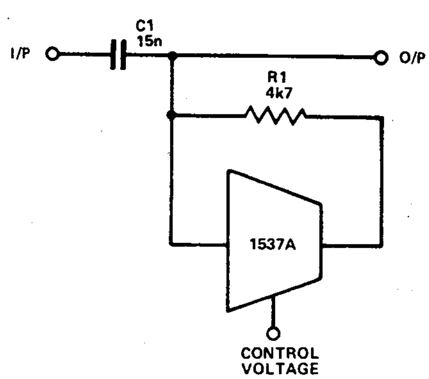


Fig. 17. A voltage controlled High Pass Filter.

The effect of a VCR (voltage controlled resistor) is used in the final two applications as the control element in filter circuits. Figure 17 shows a simple voltage controlled high pass filter. The component values shown filter out all frequencies below the variable limit of 1-2 kHz. Adjustment of the control voltage alters the lower cutoff point.

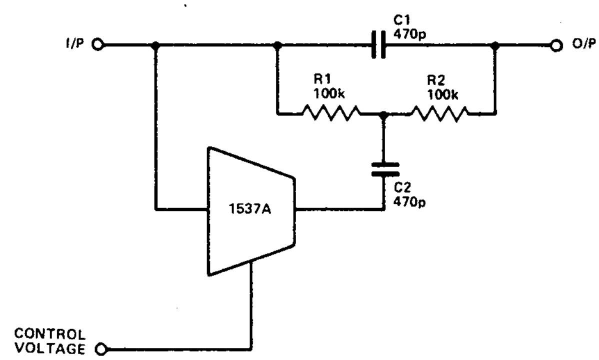


Fig. 18. A voltage controlled Band Reject (Notch) Filter.

Figure 18 consists of the circuit of a voltage controlled band reject or notch filter whose depth of notch is adjusted by the control voltage. The component values shown set the frequency at about 300 Hz and depth of notch is variable from 0 dB to about —15 dB.

Conclusions

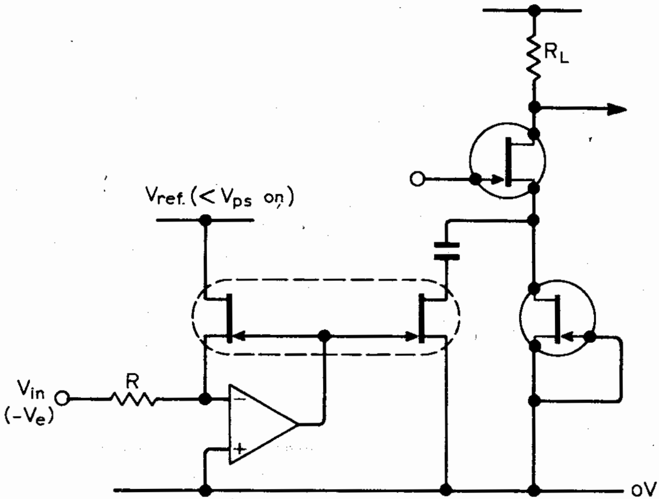
The applications given in this article show the 1537A chip to be a very versatile device. It is remarkably easy to work with, a fact which is borne out by the quality (in technical terms) of the circuitry in the breadboarded fashion of our experimental design work, let alone in the modular fashion allowed by the use of our PCB layout.

THIS IS A VERY GOOD LOGIC STATE ANALYSER AND IT ALSO SOLVES CROSSWORD PUZZLES.

Linear v.c.a.

This circuit is based on dual matched f.e.ts. and can be used in applications such as linear voltage-controlled amplifiers and two-quadrant multipliers. The resistance of the f.e.t. in the feedback loop is adjusted automatically to source the current demanded by V_{in}. This resistance, and therefore the resistance of the other f.e.t., varies inversely with input voltage. The gain of the amplifier stage is given by $R_L/V_{ref}R \times V_{in}$ and is variable over about 80dB.

B. Turner & J. Custo,
Tewkesbury,
Glos.



Automatic gain control quells amplifier thump

by Paul Brokaw
Analog Devices Inc., Semiconductor Division, Wilmington, Mass.

If an audio amplifier with automatic gain control makes a thumping noise when the input signal level changes quickly, the cause may be unwanted feedthrough of the gain control signal to the amplifier output. A simple solution is the addition of a resistor to prevent variations in the control voltage from being fed through to the output.

In the "thumpless" agc circuit of (a), transistors Q_1 and Q_2 form a differential amplifier that has a gain determined by the emitter current of the pair, I_E. This emitter current varies the transconductance and therefore the gain of transistors Q_1 and Q_2. But if gain changes too quickly, a thump may be heard. Inserting resistor R_1 in the emitter-current control circuit eliminates the thump.

Emitter current I_E is made nearly equal to the current (I_2) flowing through resistor R_2 by using identical same-substrate transistors for Q_3 and Q_4. When the base-emitter voltages of these two devices are equal, their collector currents (I_E and I_2) are also equal.

Since the base and collector of transistor Q_4 are shorted together, this device's base-emitter voltage will rise until its collector current becomes equal to $(1 - 2/\beta)I_2$, where β is the common-emitter current transfer ratio. Since transistor Q_3 is identical to transistor Q_4, Q_3's collector current will also rise to the same value. If current transfer ratio β is large and the reverse voltage feedback ratio of the transistor is small, Q_3's collector current (I_E) will nearly equal resistor current I_2. The value of current I_2 is:

$$I_2 = (E_{\text{control}} - V_{B4})/R_2$$

where V_{B4} is the voltage at the base of transistor Q_4.

Because the collector currents of transistors Q_3 and Q_4 are approximately equal, the transconductance of the differential pair (transistors Q_1 and Q_2) will vary in direct proportion to the control voltage. If Q_1 and Q_2 are identical, emitter current I_E will divide equally between them. Each transistor will have a collector current of $\alpha I_E/2$, where α is the common-base current gain.

If α is approximately equal to 1 and I_E is approximately equal to I_2, the collector currents of transistors Q_1 and Q_2 become:

$$I_{C1} = I_{C2} \approx I_2/2$$
$$I_{C1} = I_{C2} \approx (E_{\text{control}} - V_{B4})/2R_2$$

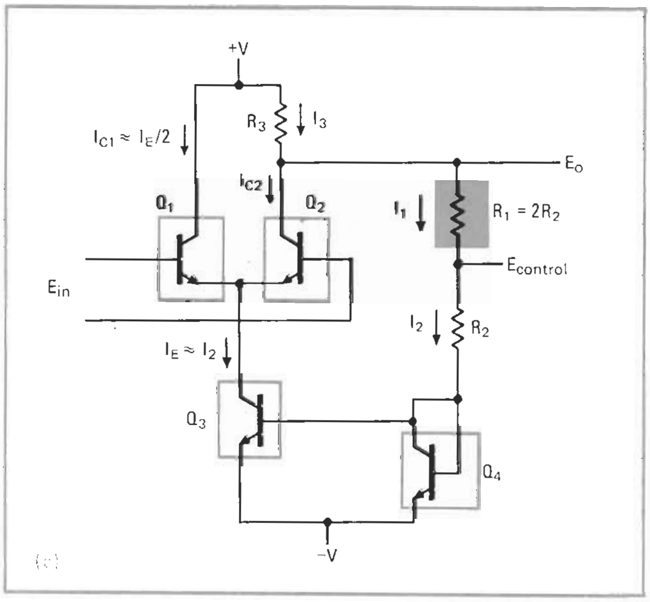
where I_{C1} is the collector current of transistor Q_1 and I_{C2} the collector current of transistor Q_2. The current (I_3) through resistor R_3 is due to both resistor current I_1 and collector current I_{C2}. Current I_1, which flows through resistor R_1, is given by:

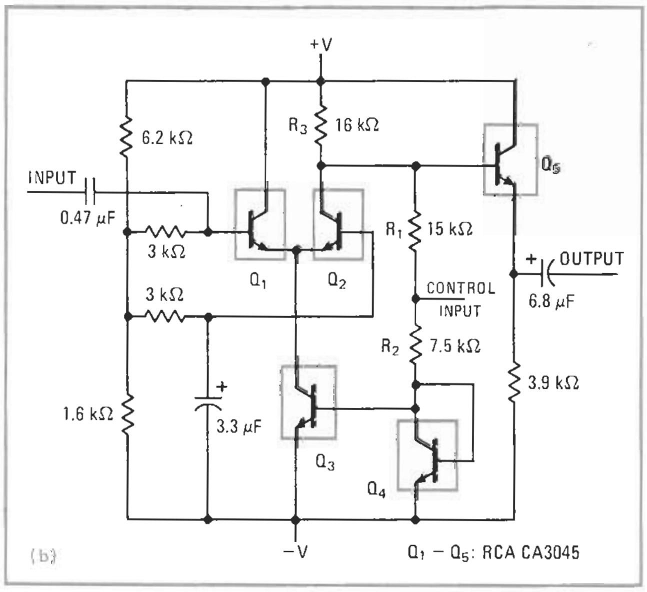
$$I_1 = (E_o - E_{\text{control}})/R_1$$

Resistor current I_3 is the sum of collector current I_{C2} and resistor current I_1:

$$I_3 = I_{C2} + I_1$$
$$I_3 = \frac{E_o}{R_1} - \frac{V_{B4}}{2R_2} + \left(\frac{1}{2R_2} - \frac{1}{R_1}\right)E_{\text{control}}$$

If $R_1 = 2R_2$, the last term in this equation drops out, making current I_3 independent of the control voltage,





Improved agc. Automatic-gain-control circuit (a) for audio amplifier applications eliminates unwanted thumping that may be heard when the input-signal level changes abruptly. Resistor R_1 prevents sudden variations in the control voltage from reaching the output as an audible thumping. An audio amplifier using this agc scheme is shown in (b); amplifier gain is 30 for a control voltage of 15 volts.

except for a small contribution caused by the dependence of V_{B4} on $E_{control}$. Since the output voltage is proportional to resistor current I_3, and not to the control voltage, variations in the control voltage will not be fed through to the output.

To implement a complete audio amplifier (b) with agc requires only a single monolithic array of five matched transistors. Two transistor pairs are used as indicated in (a), while the fifth remaining transistor is used as an output signal buffer.

The base current error introduced by transistor Q_4 can be reduced by making resistor R_2 slightly less than what the half-value approximation calls for. If resistors R_1 and R_2 are made variable, the performance of the circuit can be optimized by adjusting them for minimum feedthrough. For the component values indicated, the amplifier's voltage gain is about 30 when the control voltage is 15 volts. Circuit gain is directly proportional to the control voltage minus V_{B4}. (Voltage V_{B4} can be approximated as 0.55 v.)

Naturally, amplifier performance is limited by component tolerances. With components having 5% tolerances, the feedthrough signal can typically be suppressed by 20 to 30 decibels. Tighter tolerances will, of course, improve feedthrough suppression, but at some point, the various approximations made (like neglecting the transistor base current error) will limit performance. For a large control voltage, amplifier gain becomes inversely proportional to absolute temperature. At room temperature, this variation in amplifier gain amounts to about 0.03 dB/°C, which is not objectionable for most automatic-gain-control applications. □

Transistor array cuts cost of algebraic inversion

by Pavel Ghelfan
M.G. Electronics Ltd., Rehovot, Israel

Monolithic operators for algebraic inversion are convenient, but a reliable algebraic inverter can be built quite simply and at less cost from an integrated five-transistor array and two operational amplifiers. The circuit first converts the input signal to a logarithmic equivalent and then takes the antilog of this.

The output voltage (V_L) of amplifier A_1 is a logarithmic function of the input current (I_{in}) and the current (I_R) that the transistor array sinks at pin 13:

$$V_L = \frac{2kT}{q} ln\left(\frac{I_R}{I_{ES}}\right) - \frac{kT}{q} ln\left(\frac{I_{in}}{I_{ES}}\right) = \frac{kT}{q} ln\left(\frac{I_R{}^2}{I_{in} I_{ES}}\right)$$

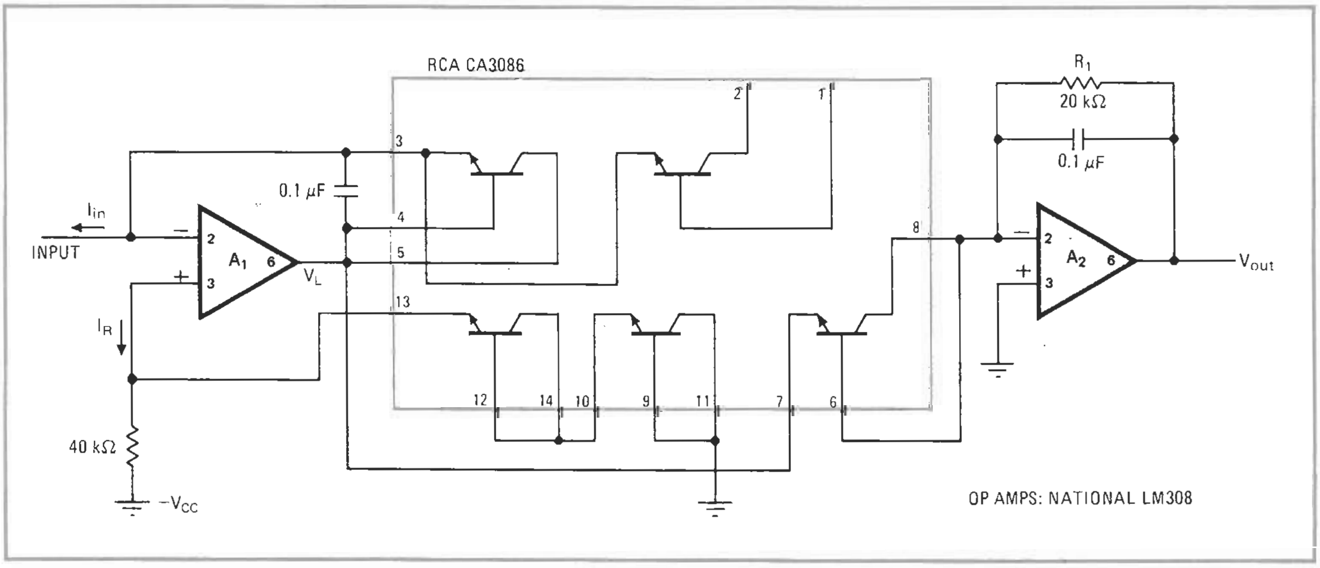
where I_{ES} is the emitter saturation current (with collector shorted to base) of the array's transistors, k is Boltzman's constant, q is the charge of an electron, and T is absolute temperature. The antilogarithmic operation is performed by amplifier A_2. The circuit's output signal can be expressed as:

$$V_{out} = I_{ES} R_1 \, exp(qV_L/kT) = I_R{}^2 R_1 / I_{in}$$

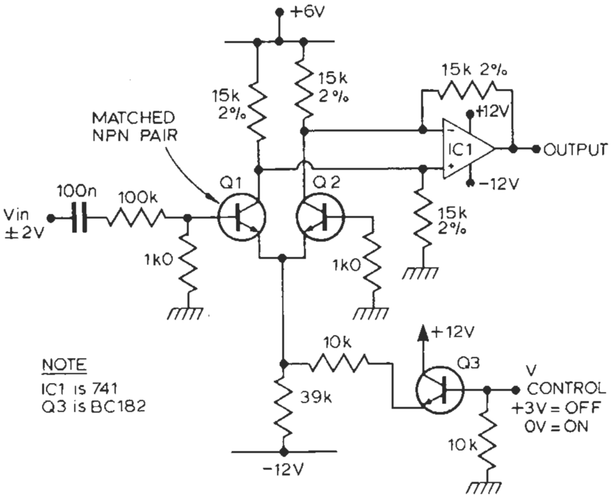
Trimming the value of constant current I_{IR} will adjust the numerator of this equation so that the output voltage of the circuit is brought to the desired value and kept there.

This inversion operator maintains good stability over a 50°C temperature range, as well as over three decades of signal amplitude variation. Its amplitude range can be significantly broadened by using low-bias-current operational amplifiers. □

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.



Taking the reciprocal. Algebraic inverter employs IC transistor array to keep costs low and to provide good temperature stability. The circuit converts the input signal to a logarithmic voltage and then takes the antilogarithm of this voltage to develop the output signal. The output, of course, is indirectly proportional to the input and can be brought to the desired value by adjusting resistor R_1.



Transistor VCA

A circuit similar in operation to a CA3080 can be constructed with a matched pair of transistors and an op amp. Transistors Q1, 2 form a differential transistor pair which is used to steer whatever current is available between the two collectors, just as in the CA3080. the difference between the collector currents is equal to the product of the input voltage times the current I_{EE} times a constant. This difference is extracted by the differential amplifier IC1. The current I_{EE} is controlled by Qe. As the control voltage goes positive, Qe robs most of the current flowing down the 39k resistor, and hence I_{EE} and the output of IC1 decrease.

FET-controlled op amp permits wide dynamic range

by Henry E. Santana
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When a field-effect transistor is operated as a voltage-controlled resistor, it is usually limited to a relatively small dynamic signal-voltage range. This is due to the nonlinearity of its drain-source resistance over a wide range of drain-source voltage.

But a wide-range voltage-controlled amplifier can be realized if a pair of FETs is connected in the bridge configuration shown in the diagram. The inverting terminal of the operational amplifier is kept at virtual ground, permitting the range of each FET's drain-source voltage to remain small, regardless of how broad the actual signal-voltage range is. This also assures that the excursions of V_{DS} will remain well within the FET's pinch-off region.

The circuit's voltage-transfer function can be written as:

$$A_V = -(R_2/R_1) + N(R_1+R_2)/R_1 + NR_2r_{on}[1 - (V_{GS}/V_P)]$$

where r_{on} is the on-resistance of the right-hand FET, V_{GS} is the gate-source voltage, and V_P is the pinch-off voltage. Variable N represents a resistance ratio:

$$N = r_{on}/(r_{on}+R_1)$$

If N is very small, and r_{on} is much less than R_1, then:

$$AV = -(R_2/R_1)(V_{GS}/V_P)$$

Although N must be small, it must, nevertheless, be greater than zero for the circuit to work. The control voltage for the circuit can range from 0 to V_P, and the peak ac input-signal voltage is determined by $I_{DS}R_1$.

Applications for this voltage-controlled amplifier include automatic gain control, true rms conversion, amplitude compression, and signal modulation. □

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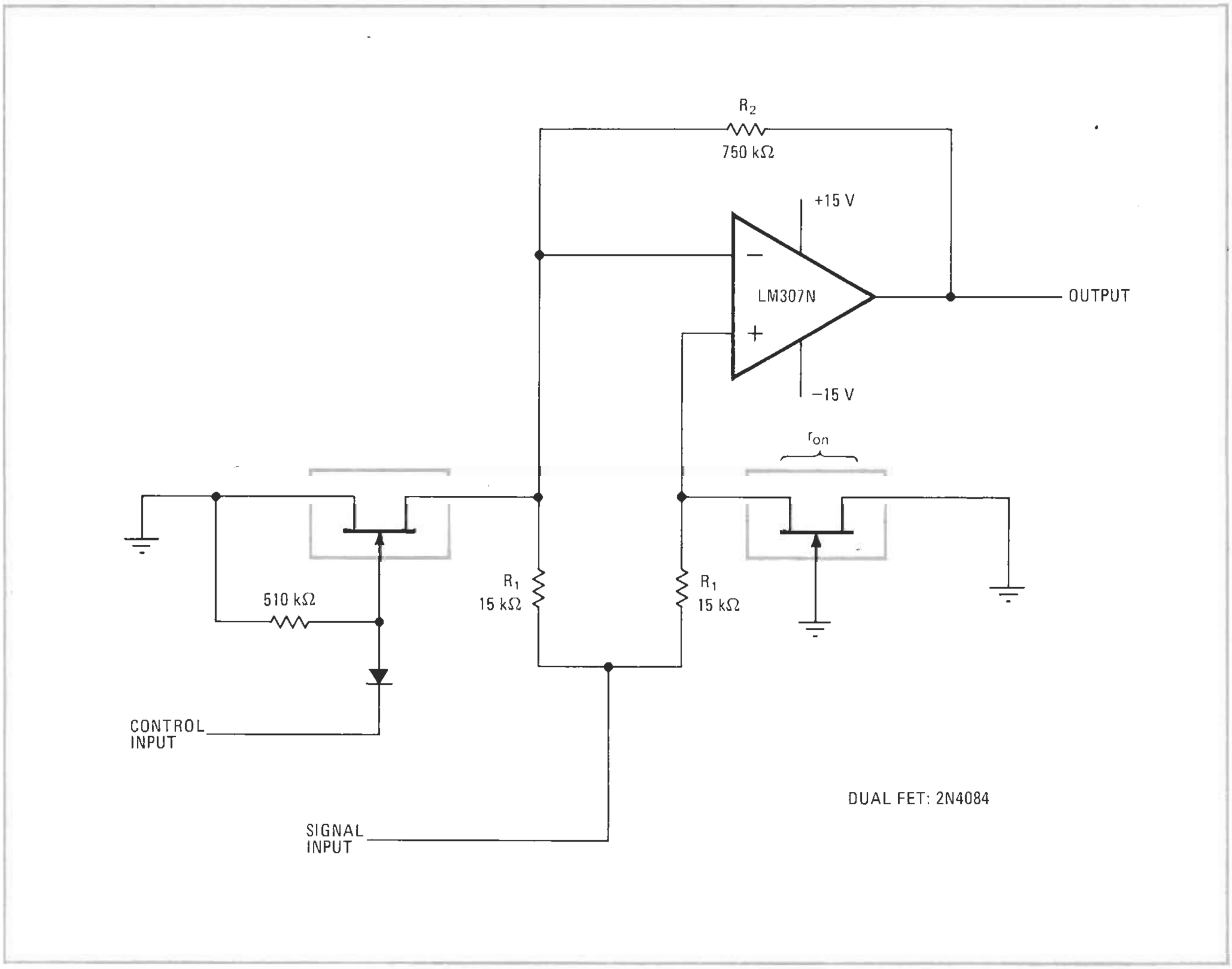
Wide-ranging. Voltage-variable amplifier can operate over a broad range of input-signal voltages. The FETs, which function as voltage-controlled resistors, are wired in a bridge configuration. Their inherent resistance nonlinearity is avoided by limiting each FET's drain-source voltage range, no matter how large the signal voltage becomes. The op amp's inverting input is held at virtual ground.



DESIGNING WITH LINEAR IC'S

This month, we'll learn about voltage-controlled amplifiers, integrators, and differentiators.

JOSEPH J. CARR

Part 7 IN THIS INSTALLMENT of "Designing with Linear IC's we will examine three circuits that we've previously overlooked but are nonetheless very important. Those are the voltage-controlled amplifier (VCA), integrator, and differentiator. As for the latter two circuits, interestingly most of the designs that are usually published do not work. We'll show you why, then provide a design that *does* work.

Voltage-controlled amplifier

A voltage-controlled amplifier (VCA) allows you to set the amplifier's gain via a control voltage. In a way, the VCA is much like the automatic gain-control (AGC) amplifiers found in receivers and certain electronic instruments. One of the most common AGC and/or VCA circuits is shown in Fig. 1. The circuit can be built either from discrete components as shown, or be obtained in IC form (the RCA CA3028, for example).

In the circuit shown, Q1 and Q2 form a differential pair. The output of the circuit can be taken from either collector, or differentially between the collectors. The input designations assume single-ended output from the collector of Q2.

The collector-emitter currents for Q1 and Q2 (I1 and I2, respectively) are derived from I3. We can thus control the gain of the circuit by controlling the collector current of Q3. Of course, that current is determined by the voltage at the base of Q3, which is applied via terminal V_C and divided down by the voltage divider consisting of R1 and R2.

Fig. 2 shows a VCA built from an operational transconductance amplifier (OTA), the RCA CA3080. Recall from our earlier discussion that an OTA has a transfer function that relates an output current to an input voltage. The gain is expressed in units of transconductance (G_M). The voltage gain (A_V) is determined by the product of G_M and the load resistance R_L. The value of G_M, on the other hand, is set by a bias current I_{ABC}, where $G_M = 19.2I_{ABC}$.

The bias current is set by control voltage V_C and resistor R6. The current is maximum (i.e. 0.5 mA) when $V_C = 0$, and minimum when $V_C = -15$ volts. The maximum transconductance, then, will be $(19.2)(0.5 \text{ mA}) = 9.6$ millisiemans = 0.0096 siemans. The voltage gain, therefore is $G_M R1 = (0.0096 \text{ siemans})(10^4 \text{ ohms}) = 96$. The voltage gain will therefore change from 0 to almost 100 as V_C varies from -15 volts up to 0 volts.

Integrators

An integrator is a circuit that will produce an output that is proportional to the time average of the input signal. In other words, the circuit performs the mathematical operation known as integration.

The simplest form of integrator is the R-C network shown in Fig. 3. That circuit may be more familiar to you as a low-pass filter. While it does perform that function, it also does, as we'll soon see, a bit more.

In the circuit of Fig. 3, when voltage V_{IN} is applied, a current flows in the resistor to charge the capacitor. Assuming that any load resistors connected across the capacitor are extremely large compared with R1, output voltage V_O reflects the accumulated capacitor charge, and is proportional to the integral of V_{IN}.

There are a number of problems with the simple circuit of Fig. 3, but most of those are solved by the Miller integrator of Fig. 4.

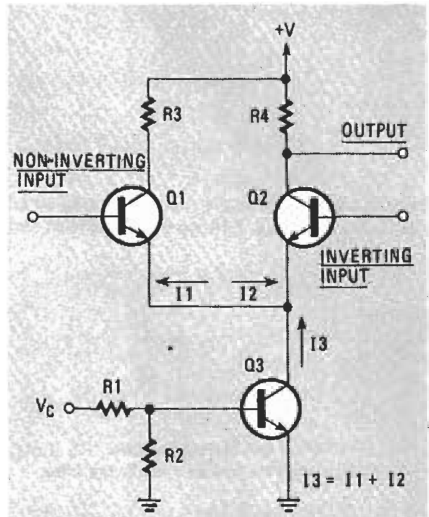


FIG. 1—ONE OF THE MOST COMMON VCA circuits. The differential amplifier can be formed from discrete components as shown, or obtained in IC form.

The Miller integrator consists of a resistor in series with the inverting input of an op-amp, and a capacitor in the feedback loop of the IC. The capacitor charges under the influence of output voltage V_O. The transfer function for the circuit is given by:

$$V_O = \frac{-1}{RC} \int V_{IN}\, dt$$

where V_O is the output voltage, V_{IN} is the input voltage, R is the resistance in ohms, C is the capacitance in farads, and dt denotes integration over time.

The "gain" of the integrator is controlled by the product RC, which is called the time-constant of the integrator. The general rule is to make RC much larger than the period of the waveform applied to V_{IN}.

There is a problem associated with the time-constant, however. Notice that the R-C (time constant) term is in the denominator. Since that product can be very low, gain can be very high. Consider, for example, the case where R = 100,000 ohms and C = .001 µF (i.e. 10^{-9} farads). There the time constant is equal to (10^5

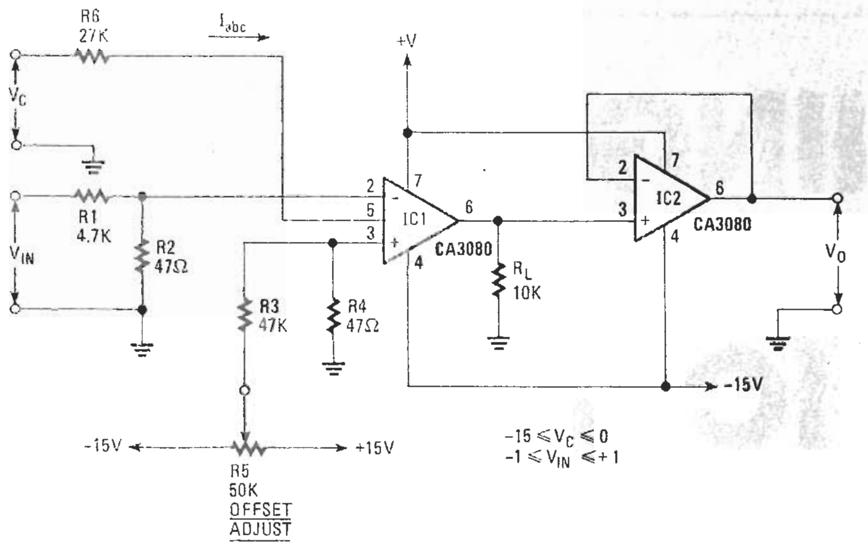


FIG. 2—THIS VOLTAGE-CONTROLLED AMPLIFIER is built using operational transconductance amplifiers. By varying V_C, the gain of the circuit can be made to vary from 0 to 96.

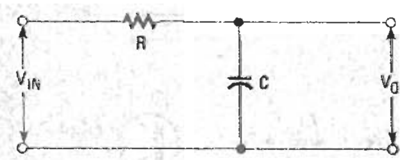


FIG. 3—SIMPLE INTEGRATOR CIRCUIT. It may be more familiar to you as a low-pass filter.

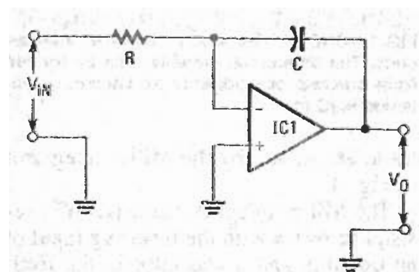


FIG. 4—MANY OF THE DEFICIENCIES of the simple integrator are solved by the Miller integrator shown here.

ohms)(10^{-9} farads) = 10^{-4} seconds. That also results in a gain of 10,000 ($1/10^{-4}$ = 10,000).

What does such a high gain mean in practical terms? It means that very small values of V_{IN} can saturate the integrator in short order! The maximum output permitted, assuming positive and negative supply voltages of 12, will be about 10 volts. If we applied a 10-millivolt DC signal to V_{IN}, therefore, the output voltage rising at a rate of $V_{IN}/RC = 0.01/10^{-4} = 100$ V/S will hit the 10 volt saturation limit in 0.1 second! If there is an input offset potential on the op-amp, or if the signal erroneously contains a DC offset (e.g. from the offset voltage of a previous stage), then the integrator output will rapidly rise to the saturation limit. Of course, shorter time constants than 10^{-4} seconds (0.1 milliseconds) will make the integrator saturate even more quickly.

There seems to be several rules for designing op-amp integrators. Those are to use an op-amp with low input-offset voltages, remove (where possible) erroneous DC offset potentials in V_{IN}, provide offset nulling for the integrator, and use the longest R-C time constant practical in designing the integrator.

The integrator in Fig. 4 is the circuit usually published in texts, and it does not work nicely for the reasons given above. With a typical 741 op-amp for example, typical input offsets cause V_O to rise to V_O(max) so rapidly that you might think the op-amp was shorted! With judicious selection of an op-amp, the modified cir-

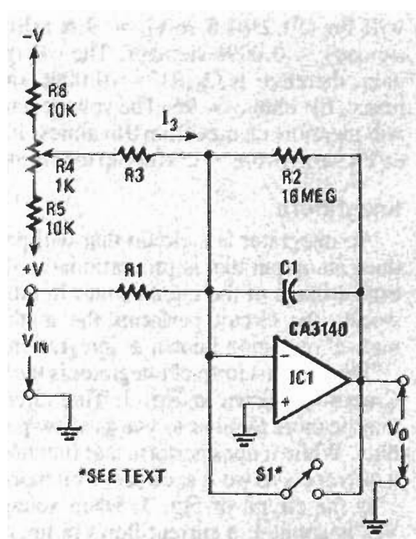


FIG. 5—A PRACTICAL INTEGRATOR. With careful op-amp selection, the circuit shown here will perform well.

cuit shown in Fig. 5 works a lot better.

The selection of an op-amp can be critical, and not all is as it appears in the data sheets. While working on another project, the author ran a series of tests on op-amps to find those best suited for use in intergrator circuits. A test circuit was built that allowed different op-amps to be plugged-in for the test. The integrator used a time constant of 0.1 second, and an input voltage, V_{IN}, of 0. In an ideal integrator, V_O should remain zero. It was found that 741 devices saturated in an average of 2 seconds. The so-called "premium" 725 devices saturated in an average of 5 seconds (still too fast). Other high-priced premium-grade devices saturated in 2-8 seconds. Those rates, it was found, were too fast to easily counteract with the usual null circuit (R3–R6 in Fig. 5). Devices with MOSFET or JFET input transistors behaved themselves much better. Saturation times with those tended to be 20–30 seconds, or more. In the end, we selected the non-premium, low-cost, CA3140 BiMOS op-amp (RCA) as the best device for the integrator design.

The "integration" components in Fig. 5 are R1 and C1; everything else in the circuit is there to "fix problems." The electronic CMOS switch, S1, for example, allows us to dump charge from C1. That charge comes from two sources: previous integrations and output offsets. Switch S1 must be momentarily closed immediately prior to each operation. That switch may be electronic CMOS, mechanical, or a relay. If C1 is very large, however, beware of exceeding the current rating of CMOS switches.

Resistor R2 shunts the integrator capacitor. The purpose of R2 is to keep C1 from being charged by certain offset voltages. Without R2, the output signal zero-baseline will rise to saturation. In one test, we applied a 1-Hz sinewave to V_{IN} and watched the output sinewave climb off the oscilloscope screen. Normally, a symmetrical sinewave will not show any DC component at all at the output of an ideal integrator.

The value of R2 is found by experimentation. If it is too high, then the circuit won't work; if it is too low, then the integrator will act like an ordinary inverting follower with a frequency compensating capacitor across R2! The value 16 megohms was found reasonable with CA3140 op-amps.

What output drift exists is easily counteracted using the null circuit (R3–R6). The input voltage V_{IN} should be zero when that circuit is adjusted. Each time R4 is adjusted, close S1 momentarily to discharge C1. Use either an oscilloscope or sensitive analog DC voltmeter to monitor V_O; select ever more sensitive ranges as you adjust R4 in order to maximize the change in V_O.

In some cases, R4 may have to have a lower value (100–1000 ohms) in order to obtain better control resolution. In all cases, R3–R6 should have a low temperature coefficient.

Integrator calibration

There are cases where we need some means of scaling or calibrating the output of the integrator. That is done by inputting a function whose integral has a uniform slope over time and then calculating that slope. Two functions have integrals that are appropriate for that purpose: a squarewave and a constant voltage.

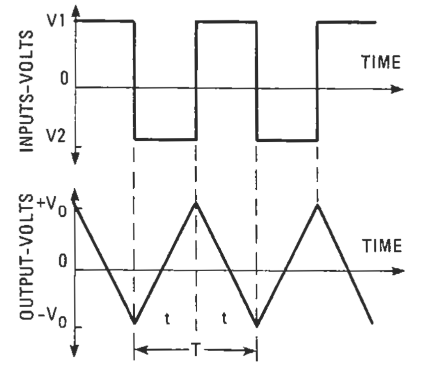


FIG. 6—IF A SQUAREWAVE is input to an integrator, the output will be a triangular wave.

When a squarewave (see Fig. 6-a) is input to an integrator, the resulting output is a triangular wave (see Fig. 6-b). (Note that for a positive input, the output is the negative integral, as is indicated by the circuit's transfer function.) Assuming a symmetrical input, that is $|V1| = |V2|$, the slopes of the ramps that make up the triangular wave are equal to $2V_O/t$, where t is equal to ½ the period (T) of the input squarewave. It is then a simple matter to measure V_O and calculate that slope.

Figure 7 shows a calibrating circuit for use when the input is a constant voltage; the resulting output is then a ramp. Initially, the inverting input of the op-amp is grounded via S2. Before starting the test, close S1 to discharge C1; that is done at time t1 (see Fig. 8), but don't allow a lot of time to pass before starting the test. A voltage is then applied to the inverting input via S2 for a fixed period of time (t2 to t3). That will cause the output to decrease uniformly (ramp) from zero down to $-V_O$, which is the output voltage at t3. The input is then grounded via S2 and V_O is measured. The slope of the ramp can then be found from $V_O/(t3 - t2)$.

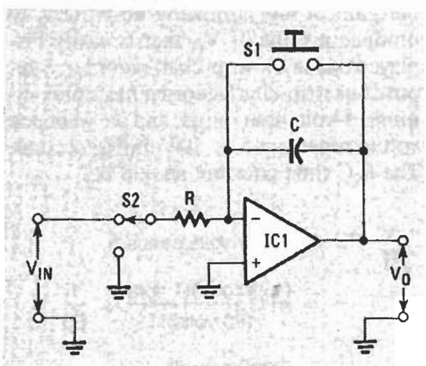


FIG. 7—WITH THIS CIRCUIT, a constant voltage can be used to calibrate an integrator.

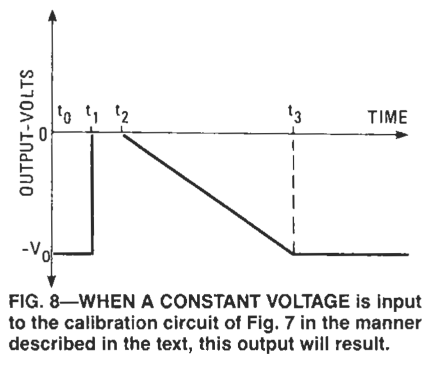


FIG. 8—WHEN A CONSTANT VOLTAGE is input to the calibration circuit of Fig. 7 in the manner described in the text, this output will result.

Differentiators

A differentiator is a circuit whose output is a derivative of the input. In other words, the output voltage is proportional to the rate-of-change of the input signal.

Differentiation and integration are inverse functions of each other. If we apply a time-varying signal, $V_{IN}(t)$, to the input of an integrator, and then apply the integrator output signal to the input of an equivalent differentiator, we should find the differentiator output to be the same as $V_{IN}(t)$—with a little propagation-delay phase-shift.

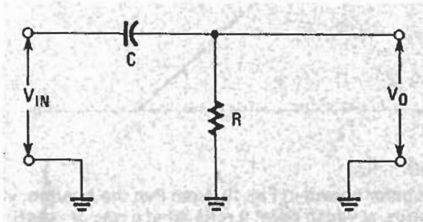


FIG. 9—A SIMPLE DIFFERENTIATOR. The circuit may be more familiar to you as a high-pass filter.

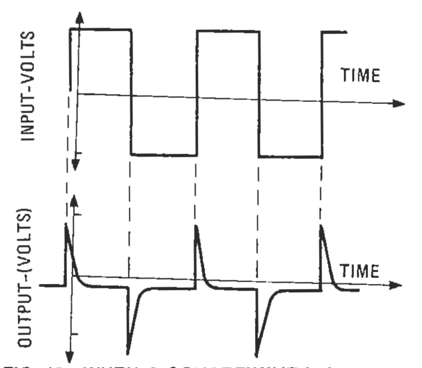


FIG. 10—WHEN A SQUAREWAVE is input to a differentiator, the output is a series of voltage spikes.

The simplest form of differentiator is the R-C network shown in Fig. 9. Note that that circuit may be more familiar to you as a high-pass filter.

The operation of the differentiator on squarewaves is shown in Figure 10. The squarewave is characterized by areas of constant amplitude (zero rate-of-change) sandwiched between edges with extremely rapid rates of change. The result is a spike-like output wave that is positive for positive-going edges, and negative for negative-going edges. Those spikes will be very broad for long time constants (compared with signal periods), and very thin for very short time constants. A differentiator time constant should be short compared with signal periods.

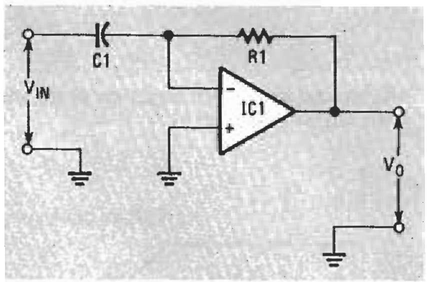


FIG. 11—ALTHOUGH THIS IS a classic differentiator, the circuit tends to be unstable under some circumstances.

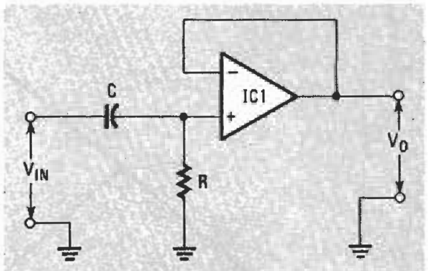


FIG. 12—WHILE THIS CIRCUIT performs much better than the one shown in Fig. 11, it is used only in limited applications.

Two forms of active operational amplifier differentiators are shown in Figs. 11 and 12. The inverting version shown in Fig. 11 is the classic differentiator, and will produce an output potential of:

$$V_O = -R1C1\frac{dV_{IN}}{dt}$$

Unfortunately, the circuit of Fig. 11 tends to be a little unstable (i.e. it will "ring") under some circumstances. In a moment, we will see how to "fix" that problem.

The circuit of Fig. 12 merely uses a noninverting operational amplifier (unity gain) to buffer the R-C differentiator output. That circuit is simple, produces a low output impedance for the R-C differentiator, and is generally well-behaved.

Even so, the circuit is not frequently used. More common is a "fixed" version of the circuit in Fig. 11; that circuit is shown in Fig. 13. Two extra components are used to stabilize the circuit: R2 and C2.

A frequency-response plot of that circuit is shown in Figure 14. As is shown, the frequency response of the amplifier (A_{VOL}) is flat from DC to some frequency, at which the gain begins to roll off at a rate of −6 dB/octave. In order to achieve stability, we will want the curve 1/β to

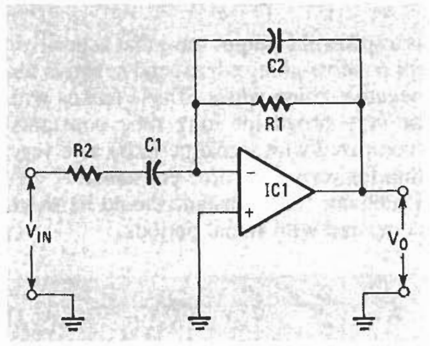


FIG. 13—AN IMPROVED VERSION of the differentiator shown in Fig. 11.

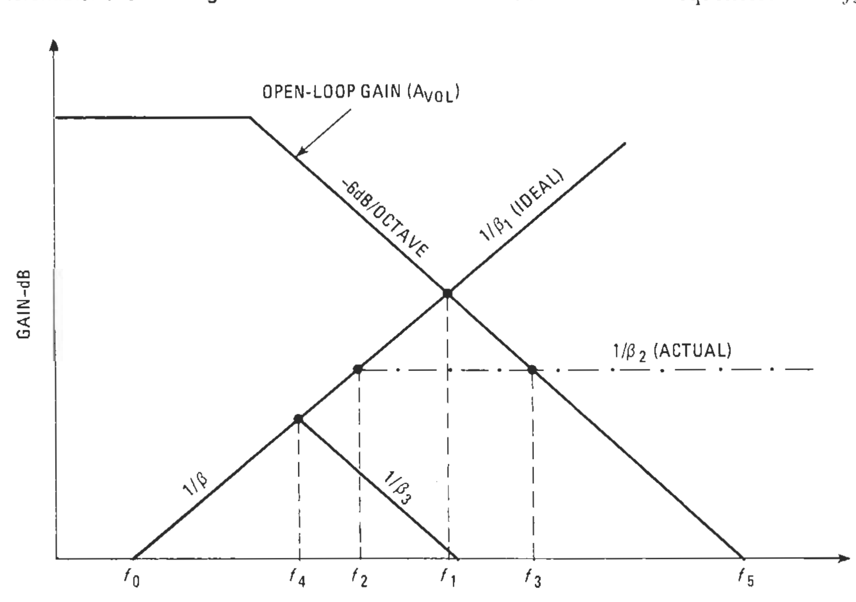


FIG. 14—FREQUENCY RESPONSE plot of the differentiator shown in Fig. 13. Note that the frequency response of the circuit is flat from DC to some frequency, at which point it rolls off at a rate of −6dB/octave.

intersect A_{VOL} with a net slope between them of less than −12dB/octave. (Note: β is the feedback attenuation factor introduced by R1 and C1). Since the net slope between A_{VOL} and $1/\beta$ is high, the circuit may tend to oscillate. We must modify the feedback frequency response to make the feedback curve like $1/\beta 2$; that is the function of resistor R2, which is in series with C2. We want R2 to introduce a frequency response breakpoint at f_2. If f_1 is the frequency at which the ideal curve $1/\beta$, intersects A_{VOL}, then f_2 must be 3.16 times lower than f_1 (i.e. $f_2 = f_1/3.16$). The minimum value of R2 that will accomplish that trick is given by:

$$R2 = \frac{3.16}{2\pi f_1 C1}$$

Combining constants yields:

$$R2 = \frac{0.503}{f_1 C1}$$

In general, minimum values for R2 fall in the range of 40 ohms to 500 ohms (although that is not absolute).

Another problem often seen in active differentiators is that the high-pass filter nature of the R-C networks means that gain increases with frequency. Hence, high frequency noise in active differentiators can be vicious. Capacitor C2 is used to prevent that problem, which creates curve $1/\beta 3$. The value of that capacitor should be found using the following formula:

$$C_2 = \frac{1}{2\pi f_4 R1}$$

Capacitor C2 will provide integrator action at frequencies above f_4, and differentiator action at frequencies below f_4.

Differentiator calibration

The output signal from the differentiator is proportional to the rate of change of the input signal in volts per second. We can calibrate the differentiator using a function with constant-slope ramps.

One such function is a triangular wave. A triangular waveform has two constant-slope edges, one positive and the other negative. Since the usual op-amp differentiator is an inverting circuit, the output will be a positive voltage for a negative-going ramp, and a negative voltage for positive-going ramps. Thus, a triangle input signal will produce a squarewave output. If the amplitude of the input signal is V1, and the period T, then the rate-of-change of each slope is V1(½ T), assuming a symmetrical triangle wave. The amplitude of the output squarewave is 0 to $+V_O$ for the negative ramp, and 0 to $-V_O$ for the positive ramp.

A linear ramp can also be used for differentiator calibration. Such a ramp could be generated by inputting a known constant voltage to a Miller integrator. The differentiator output will be a constant voltage that is proportional to the ramp slope.

If we want the output signal quantified, then we will have to provide some means to vary V_O for calibration purposes; a variable-gain inverting follower will do the trick nicely and will also flip the polarity so that positive outputs are obtained for positive inputs and negative outputs are obtained for negative inputs.

Let's look at a practical example. Differentiators are used in a wide variety of biomedical applications. One of those is as an arterial pressure amplifier. Assume that the leading edge of a human arterial blood-pressure waveform, P(t), has a rate of change (dP/dt) that's on the order of 5 × 10^3 mmHg/second (mmHg is millimeters of mercury). A typical arterial pressure amplifier has an output voltage scale factor of 10mV/mmHg, so it will have a rate-of-change output, dV_O/dt, that's equal to:

$$\frac{dv_o}{dt} = \frac{S \times 10^3 \text{ mmHg}}{\text{second}} \times \frac{10\text{mV}}{\text{mmHg}} \times \frac{1\text{V}}{10^3\text{mV}}$$

$$= \frac{5 \times 10^3 \times 10 \times 1}{10^3} = 50 \text{ volts/second}$$

The differentiator output voltage is:

$$V_O = -RC\frac{dV_o}{dt}$$

$$V_O = -RC \text{ (50 volts/second)}$$

We can vary the value of R and C (those are the components used to determine the time constant of the circuit, such as R1 and C1 in the differentiator of Fig. 13), and the gain of any following amplifiers, to produce a value of V_O that is easily displayed on, say, a strip-chart recorder. Suppose our strip-chart recorder has a plus-or-minus 1-volt input range, and we want one volt to represent 5 × 10^3 mmHg/second. The R-C time constant should be:

$$RC = V_O/(50 \text{ volts/second})$$

$$= \frac{(1 \text{ second})(1 \text{ volt})}{(50 \text{ volts})} = \frac{1}{50}$$

$$= 0.02 \text{ seconds}$$

Once we have a value for our R-C time constant, we need to find a combination of R and C that is appropriate. Let C = 0.1 µF (a tentative guess) and calculate R:

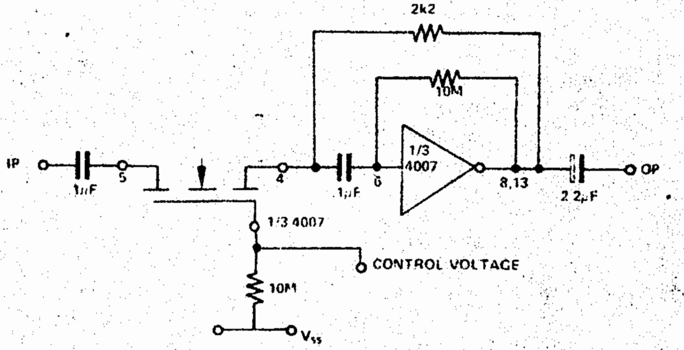
$$R = \frac{.02}{C}$$

$$= \frac{.02}{1 \times 10^{-7}} = 200{,}000 \text{ ohms}$$

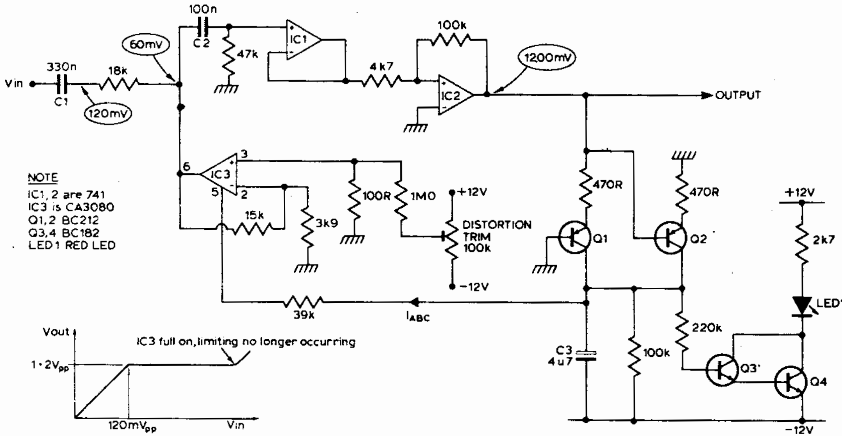
We can, therefore, build our differentiator using a 200,000-ohm resistor and a 0.1 µF capacitor. **R-E**

VOLTAGE CONTROLLED AMPLIFIER

When the voltage at the gate of a n-channel MOSFET is varied from 0V – supply volts its resistance varies from about 1kΩ to several tens of megohms. This fact is utilised in the following VCA. The inverter is biased into linear operation by the 10mΩ resistor. When feedback is applied the gain is set by $\frac{RF}{RIN}$. By allowing a MOSFET to be RIN and RF fixed, with the values shown as the control voltage varies from $V_{DD} - V_{SS}$ the gain of the amplifier varies from cut-off to just over unity.



Voltage Control Of Gain



STATE OF SOLID STATE

Analog IC's and new MOSFET's

ROBERT F. SCOTT
SEMICONDUCTOR EDITOR

MANY INTERESTING IC'S HAVE RECENTLY been developed for audio applications. For example, the MTA1537A is a precision AGC/Voltage-controlled attenuator developed by Aphex Systems. It is specially suited for high-speed precision control of signal level, dynamic range, and phase and amplitude equalization in applications such as high-quality audio controllers, analog computers, precision oscillators, robots, video-effects generators, servo controllers and precision phase detectors. The MTA1537A comes in a standard 14-pin DIP package; typical specifications are:

- 110-dB dynamic range
- 120-dB attenuation
- less than 0.05% THD
- 2-mv control-voltage feedthrough
- 200-MHz bandwidth

Figure 1 shows how the MTA1537A can be used as a stereo attenuator. Typical specifications for that circuit include maximum input and output levels of 7.75 volts rms, maximum attenuation of 100 dB over a frequency range of DC to 200 kHz, and a control voltage feedthrough of 5 mV at 100 dB attenuation. Note that at high frequencies, circuit layout will affect attenuation.

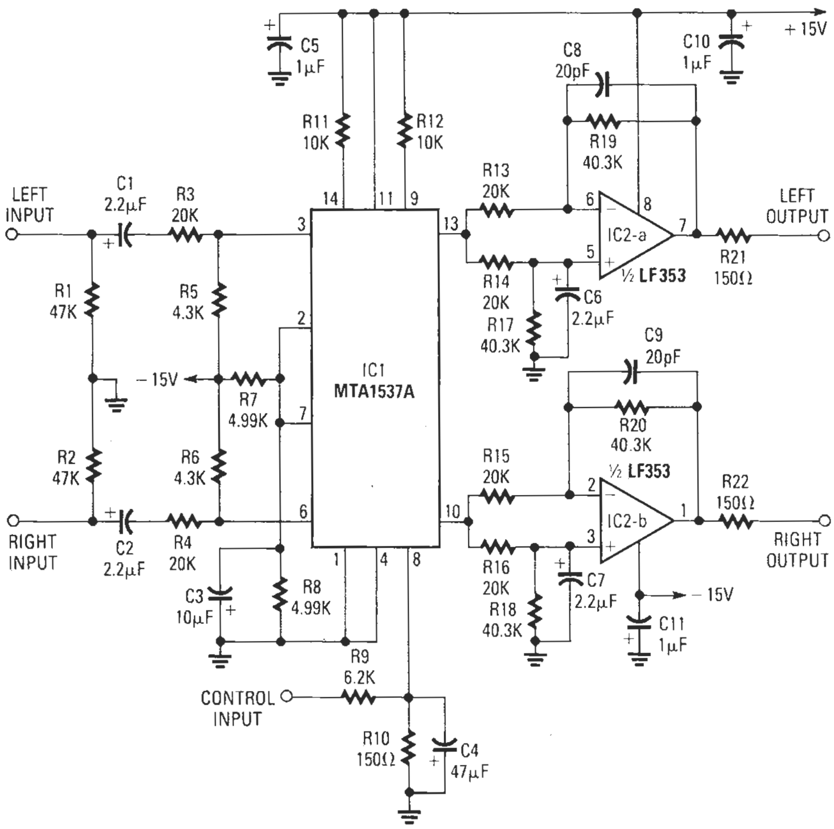


FIG. 1

The op-amps used in that circuit may be LF353's (or equivalents); NE5534's could be used for low-noise applications. The MTA1537's data sheets include basic circuit diagrams for applications such as a voltage-variable resistor, voltage-tunable high-pass filter, voltage-variable inductor, voltage-controlled band-reject filter, and a voltage-controlled graphic equalizer.

For two samples of the MTA1537A, and a copy of the data sheet, send $17.00 to the manufacturer.—**Aphex Systems Ltd.,** 13340 Saticoy St., North Hollywood, CA 91605

New JFET op-amp family

The MC34080 series is a new generation of JFET-input op-amps from Motorola. They are available in single, dual, and quad versions (both compensated and uncompensated), and they offer bandwidth and slew rates as much as four times greater than previously available types.

A combination of JFET and bipolar technologies, along with an all-NPN output stage, has yielded a fully-compensated op-amp family with a gain-bandwidth product of 8.0 MHz and slew rates in excess of 30 V/µs. For greater speed, uncompensated versions are offered with a gain-bandwidth product of 16 MHz and slew rates of 60 V/µs.

The all-NPN output stage provides a peak-to-peak output voltage swing that is 33% greater than standard op-amps that use NPN/PNP output stages. Other features include: input impedance of 10^{12} ohms, open-loop output impedance of 30 ohms at 1.0 MHz, THD of 0.01%, phase/gain margins of 55°/7.6 dB for fully compensated devices, and 30 nV/$\sqrt{\text{Hz}}$ input noise voltage.

Prices range from $0.59 for the single op-amp MC34080 (uncompensated) and MC34081 (compensated) to $2.80 for quad versions. For data sheets and complete pricing information, contact **Motorola Semiconductor Products,** PO Box 20912, Phoenix, AZ 85036.

Low-power quad op-amps

National's LP124 family of quad op-amps is a pin-compatible, low-power version of that company's LM124 family. The new series consists of the LP124, LP2902 and LP324 high-gain, internally-compensated micropower op-amps. They are well suited for CMOS applications, battery-powered equipment and other circuits that require good DC performance and low supply current. The LP124 joins National's LP139 and LP165 low-power quad comparator families.

Maximum power consumption is 125 µA—one-tenth that of National's LM series, and lower than many CMOS amplifiers. In addition, input bias current has been reduced by a factor of ten to a maximum of only 4 nA. Available in 14-pin plastic and ceramic DIP's, the LP324N sells for $0.75 in quantities of 100 or greater. For more information, contact **National Semiconductor Corporation,** 2900 Semiconductor Drive, Santa Clara, CA 95051.

New temperature sensors

The LM34, LM34C, and LM34D make up a series of precision temperature sensors with output voltages linearly proportional to the Fahrenheit temperature scale. Those devices, from National Semiconductor, require no external calibration or trimming to provide typical accuracies of ½°F at room temperature and ±½°F over a full range of −50 to +300°F. Temperature stability is aided by the LM34's low self-heating, which is typically less than 0.2°F in still air.

Interfacing to readout or control circuits is easy. The output impedance is only 0.1 ohm at 2 mA and the output is linear at 10 mV/°F. It can be used with single power supplies or with split (plus and minus) supplies delivering from 5 to 30 volts and draws less than 60 mA. The LM34 can be glued or cemented to a surface and its temperature will be within 0.02°F of the surface reading.

The LM34 is rated from −50 to +300°F, the LM34C from −40 to +230°F, and the LM34D from +32 to +212°F. All are available in TO-46 packages. The LM34C and LM34D are also offered in the TO-92 packages. The LM34 is $1.55 each in 100 and up lots.—**National Semiconductor Corp.**, 2900 Semiconductor Drive, Santa Clara, CA 95051. **R-E**