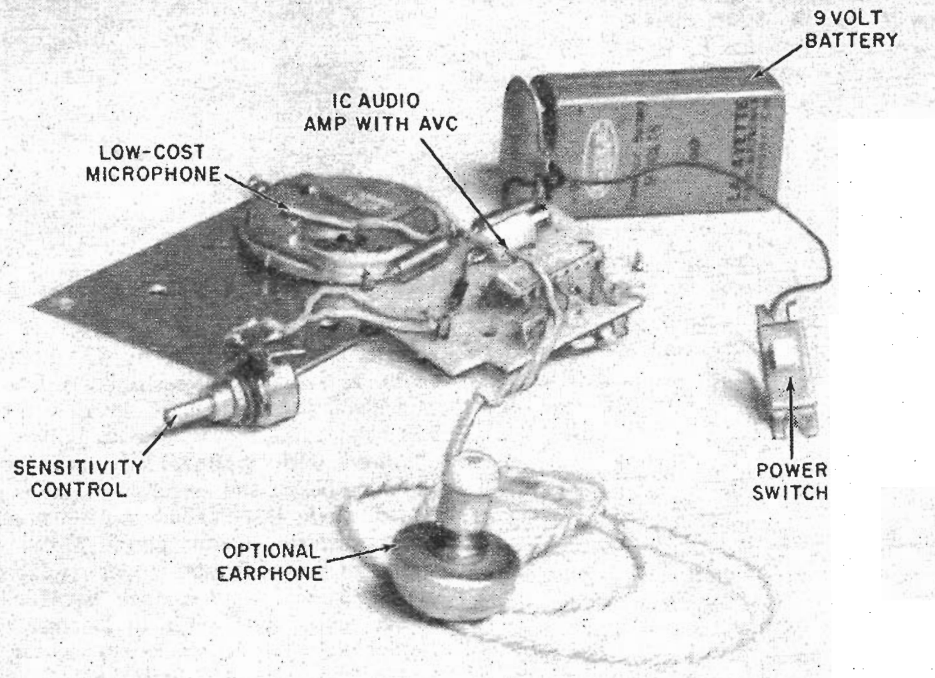
BUILD YOUR OWN...
AMPLIFIED MICROPHONE WITH AUTOMATIC VOLUME CONTROL



Build special hearing aid, or amplified mobile or base station CB mike for extra talk power

By W. H. Sandford Jr.

EVER wonder why television commercials seem to be so much louder than the average program they accompany, or why some very low-level CB signals are more readable than others? In both cases, you can bet that the audio is passed through some kind of volume compressor or peak limiter before being applied to the modulator. In plain terms, it is an example of what more talk power can do!

Here is a volume-controlled microphone-preamp that does much of what the professional limiting circuits do—but at a much lower cost, and it's one you can build yourself. Any sound loud enough to overmodulate a transmitter automatically reduces the output of the mike to prevent overmodulation, while softer sounds benefit from the normally high mike gain and modulate the transmitter with a much higher percentage than they normally would. Hence, you get a higher *average* level of modulation and more talk power.

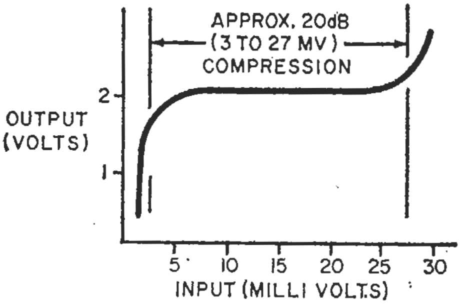
But this is really a universal circuit; instead of using it to modulate a transmitter, replace the level control with a high impedance earphone, and presto—a hearing aid with AVC.

(Turn leaf)

e/e AMPLIFIED MICROPHONE

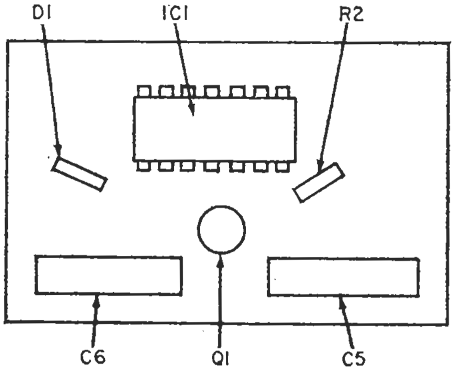
If this circuit is used in a homemade hearing aid as shown in the photographs, the sensitivity (volume) control can be turned way up to catch distant or soft sounds, but loud or very close sounds will be lowered to prevent eardrum overload.

How it Works. Sound waves picked up by the crystal microphone are converted to electrical signals and applied to the input of the integrated circuit amplifier. Called an *operational* amplifier by manufacturers and engineers, it amplifies signals to a sufficient power level to drive the dynamic earphone.



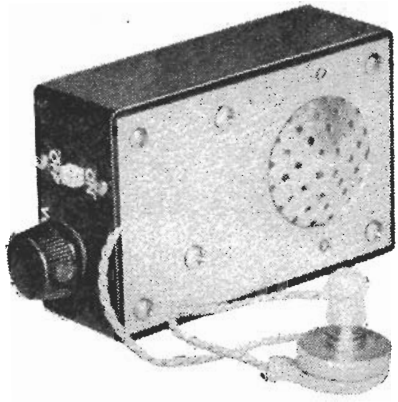
Curve illustrates ideal compression action.

It can put out a sound level of such strength to be uncomfortably loud for a person with normal hearing. When used as a microphone preamp, up to 2-volts are available at the output. The gain of the amplifier (and the resulting output level) is controlled by the amount of signal fed back, by R5, from the output to the inverting input of the amplifier. The more signal fed back, the lower



Location of components on top of perf board. IC1 comes in three different packages. Mark pin numbers of package you use on diagram.

the amplifier gain. Assume a case where the volume control is adjusted for maximum gain (R3 set for zero resistance). In the absence of sound waves falling on the microphone, Q1 acts as a low resistance compared to the resistance of R2, and the gain of the amplifier is approximately equal to the



Now in actual use, the author built this compact hearing aid for under 20 dollars.

ratio of R2 + R5 divided by R2, about 667 times. A 3-millivolt input signal will result in a 2-volt output signal. Converted to decibels (dB), this is a gain of about 57dB.

Lowers Gain. When sound waves fall on the microphone and an output signal is generated, diode D1 conducts on the negative going portions of the output signal and charges C4. This negative bias on the gate of Q1 causes the resistance between the source and drain of Q1 to increase. The greater the voltage developed across C4, the greater the increase in resistance Q1 adds in series with R2. This reduces the overall gain to about 75, which means a 27-millivolt input is required for a 2-volt output. Converting again to dB gives us a gain of about 37dB.

Thus if the FET receives a voltage on the gate great enough to cause its drain-to-source resistance to increase to about 12,000 ohms, the amplifier gain is seen to be reduced by 20 dB. Tests with an audio oscillator replacing the microphone and an oscilloscope observing the output waveforms have shown that after the output signal reaches a peak amplitude, the input signal can be increased another 20 dB without changing the output level. Further increases in input signal level result in distortion of the output waveform and this condition must be taken care of by reducing the amplifier gain with the introduction of resistance into the circuit at

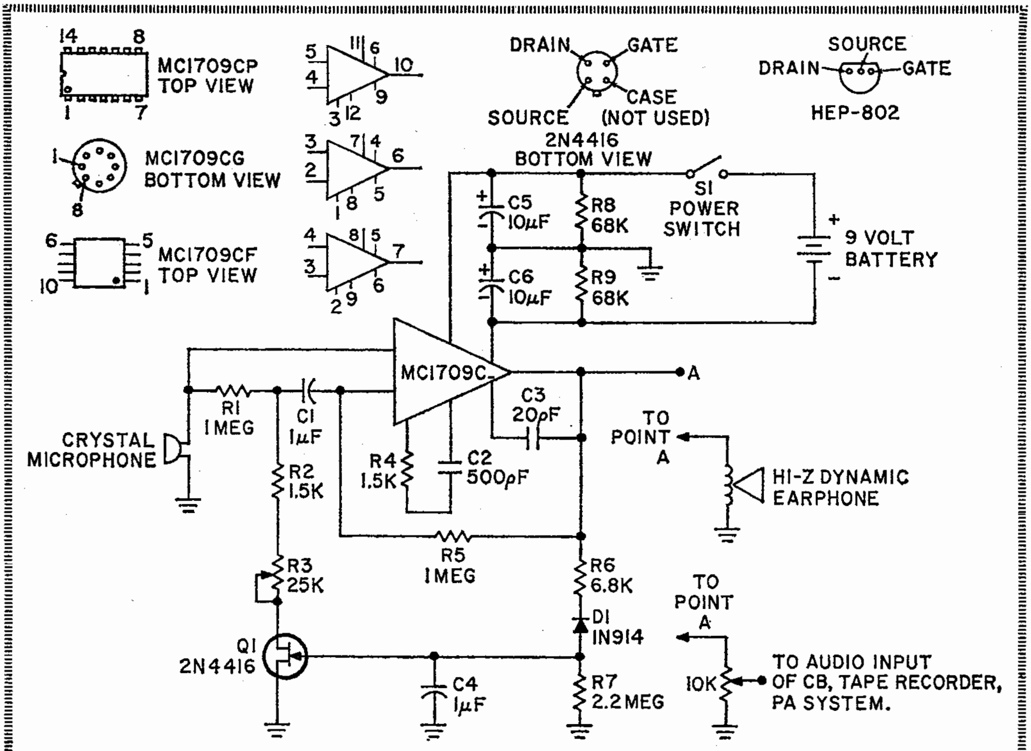
R3 (the sensitivity control). R3 is of such a value to permit 30dB of volume control action.

Bounces Back. The AVC (automatic volume control), or gain reduction, has a fast attack since the diode charges C4 rapidly. The gain recovery takes several seconds after the sound level is reduced due to the discharge of C4 through R7. This is standard practice in compressor circuits.

The standby power drawn by the amplifier from the battery is very low. The required positive and negative voltages needed to operate the integrated circuit are obtained by the low current divider R8 and R9. These resistors are bypassed for signals by the two capacitors C5 and C6.

A desirable feature of the circuit is the *bootstrapping* of the positive amplifier input by returning R1 to the junction of R2 and C1 instead of to ground. This increases the apparent input impedance of the input and reduces loading on the microphone.

The hearing aid is constructed in a 3¼-in. x 2⅛-in. x 1⅛-in. bakelite case with an aluminum panel. All the parts except the battery and switch are mounted on a small piece of perf board which is then fastened to the panel. A hole is cut in the panel for the microphone, and the switch and sensitivity control are mounted through one end of the case. No provision was made to secure the battery since, in my case, the parts placement permitted wedging the battery into the case in such a way that the rest of the components kept it from moving about. ■



PARTS LIST FOR AMPLIFIED MICROPHONE

- **B1**—9-volt battery, Everyready 216 or equiv.
- **C1,C4**—1uF electrolytic capacitor, 3 VDC or better
- **C2**—500 pF disc capacitor, 50 VDC or better
- **C3**—20 pF disc capacitor, 50 VDC or better
- **C5,C6**—10uF tantalum capacitor, 6 VDC or better
- **D1**—Diode, silicon, 1N914 or HEP-156
- **EP1**—Earphone, dynamic, high impedance (Lafayette 40-78010)
- **IC1**—IC amplifier, MC1709CF, G or P (Motorola)
- **M1**—Microphone, crystal (Lafayette 99-45103; Radio Shack 270-095)
- **Q1**—N-channel FET, 2N4416 or HEP-802
- **R1, R5**—1-megohm, ¼ watt resistor
- **R2, R4**—1500-ohm, ¼ watt resistor
- **R3**—25,000-ohm linear taper potentiometer
- **R6**—6800-ohm, ¼ watt resistor
- **R7**—2.2-Megohm, ¼ watt resistor
- **R8, R9**—68,000-ohm, ¼ watt resistor
- **S1**—Slide or toggle switch, SPST
- **Misc.**—Hardware, knobs, perforated board, push-in clips, wire, solder, battery connector, etc.

Compress-O-Phone

Roar into it like a lion or whisper like a church mouse but the output from this microphone will remain constant within 3db.

By JOSEPH RITCHIE

A PERENNIAL hang up when making live recordings is keeping the record level constant (or nearly so). You know what it's like. After setting up to tape, say, a group discussion, you set the level on one person's voice. After the talking starts people speak louder; now and then they quiet down. The result: a tape with overload distortion and passages where you can just about (but not quite) hear what someone is saying. To solve this you must ride gain constantly, which means full attention to the machine's record-level indicator and gain control.

The Compress-O-Phone makes this unnecessary. It will give you hands-off uniform-level tape recordings, crushing CB and ham talk power, a super-loud PA system without feedback and hidden-mike tape recordings of top clarity.

The Compress-O-Phone is a dynamic microphone with a built-in speech compressor the likes of which hasn't been seen outside of a James Bond thriller. The compressor is so effective it almost wipes out dynamic range. Regardless of the sound level into the microphone—a soft whisper, a shout or murmurs 20 ft. away—the microphone's output remains constant within 3db.

Used with a ham or CB transmitter the compressor can add nominally 30db of talk power (and that's not *up to*, it is 30db). In fact, some SSB transmitters designed for peak voice power may not be able to handle the sustained constant level signal from the compressor.

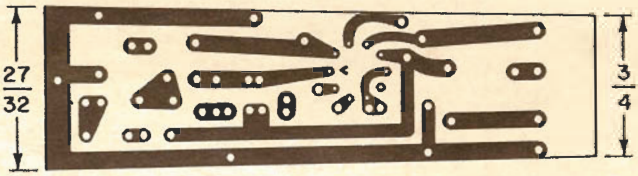
When used with a PA system the tendency to break into microphonic howling is reduced some 20 to 30db, allowing that much of an increase in volume level. The only time you should not use the compressor is when recording music because it will eliminate the desirable dynamic range of the instruments; everything will come out sounding like a player piano—no expression.

The heart of the compressor is the op-amp (operational amplifier) shown in the schematic in Fig. 9. Unlike clippers, which chop down the dynamic range by clipping waveform peaks (thereby creating distortion) or standard compressors, which go into compression only above a reference sound level, our compressor introduces no distortion and goes into maximum compression at signal levels representing a very low whisper. Regardless of the sound level into the microphone, the compressor's output level is essentially flat.

Compress-O-Phone

How it Works. Integrated circuit IC1 is an op-amp whose gain is determined by the ratio of resistance of feedback resistor R3 to the resistance represented by FET Q1, which is in parallel with resistor R1. Since R1's resistance is greater than Q1's static resistance, R1 is effectively out of the circuit and the gain depends on R3 and Q1. At very-low-level inputs (from the microphone) Q1 appears as a low resistance and the op-amp's gain is maximum. As the mike input level increases part of the output at terminal 5

Fig. 1—At right is template for circuit board. Cut it out and trace outline of black areas with carbon paper on foil side of board. Fine line at top, right and bottom, right does not have to be traced.



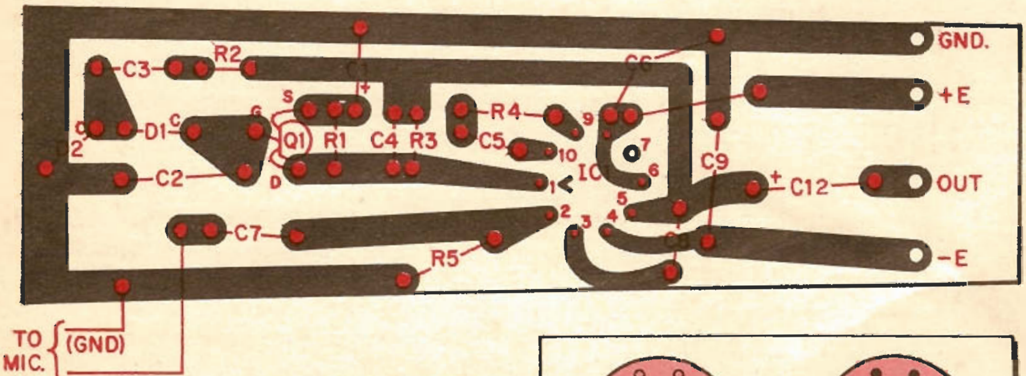


Fig. 2—Diagram above is an X-ray view of board from side on which parts are installed. Designations show location of parts. Note foil from pin 3 on IC1 to bottom edge; foil should be trimmed slightly at edge so it won't touch mike case.

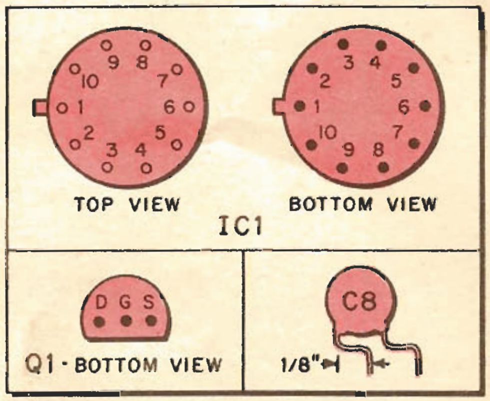


Fig. 3—Diagrams at right show top- and bottom-view numbering of IC1's pins. Get these right when installing the IC or it will be destroyed. Q1's leads are drain, source and gate. Bend leads on C8 so it will be in from edge of mike case.

Fig. 4—Compressor is built on a printed-circuit board which just fits in mike case. Note that the electrolytic capacitor (white) in foreground is folded flat to board. Leave capacitor's leads a trifle long so it can be folded this way. Output of compressor is at left. Circuit's input is at the right.

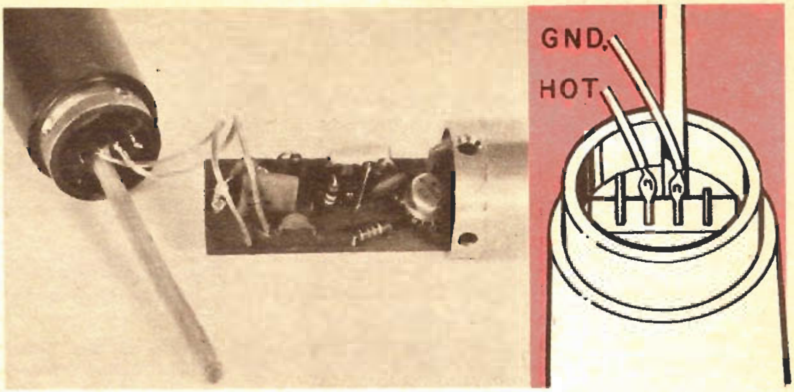


Fig. 5—Hollow tube sticking out of back of front half of mike is air-pressure relief tube for mike element. Tube slips under board during final assembly; don't damage it. Sketch at right shows connections to lugs in mike.

on IC1 is rectified by D1 and D2; the resultant DC is applied to Q1's gate, which effectively increases Q1's resistance, thereby lowering the gain of the op-amp.

When the input increases the DC feedback to Q1's gate similarly increases, thereby holding the op-amp's output at a fixed level. Compression is lost only when the input signal exceeds 200 mv (rms) which is considerably more than the maximum output level of a dynamic microphone, even when you shout into it. For all practical purposes, therefore, the op-amp's output is constant regardless of the sound level into the microphone, until the input signal becomes so weak it is unusable anyway.

Normally, when heavy compression is employed, the higher voice frequencies representing sibilant sounds are highly articulated because they are amplified with respect to the more powerful lower voice frequencies. Part of this excess sibilant articulation is attenuated by C4 and C8, which provide a slight high frequency roll-off.

The compressor is built on a printed, circuit board that fits in the case of an Allied Radio Model 3311 dynamic microphone, making the entire unit self-contained except for the battery box which is located at the tape recorder, transmitter or amplifier. The circuit is specifically designed for the Allied mike and may not work with other dynamic mikes with which we haven't tried it.

Since the gain of IC1 is extremely high the circuit must be assembled as shown using the exact parts specified in the Parts List. Any departure from layout or component type can cause complete instability and the chance of destroying IC1. As long as the unit is assembled on the board shown in Figs. 1 & 4 it will fit in the mike case or in a box at the end of a mike cable no longer than 20 ft. If the circuit breaks into oscillation when mounted at the end of the cable, shorten the cable until oscillation stops. We suggest the microphone installation shown.

Construction. First, disassemble the microphone. At about in the center of the case there's a metal band secured by a single screw. Remove the screw and the band and you'll see three screws that hold together the front and back halves of the case. Remove the three screws and gently pry the halves apart. As the sections separate you will see two wires connected to lugs on the back of the front section. Note which wire connects to which lug and unsolder them. Separate the halves. You will see a long length of sleeving (spaghetti) attached to the back of the front section. Be very careful because this is the mike element's air-pressure relief tube and must not be damaged. Set the front section aside and remove the switch and the connector at the back of the rear section.

Cut a section of copper-clad circuit board 2⅞ in. long which measures 27/32 in. wide at the front and ¾ in. at the rear as in Fig. 1. The front of the board will be near the front section of the mike when the unit is completed.

Slide the board into the rear section; it should go in easily and sit slightly *below* the diameter of the case from front to back. If the back or front lies on the diameter of the case, file the edges of the board so that it fits correctly. Insert a ½-in. rubber grommet in the rear of the case where the jack used to be and set the case aside.

Making the Board. Remove any protective plastic cover which may be on the board and scrub the copper foil with a cleanser such as Ajax; wash thoroughly and dry. Place a piece of carbon paper face down on the foil and tape the template in Fig. 1 on top of the carbon paper. Using an ice pick or pointed tool, indent the copper foil at the drilling points (small holes) by pushing the tool through the template and carbon paper. Using a fine-tip ball-point pen, carefully trace the template outline. And make certain you trace the arrowhead which is opposite pin 1 on IC1. Remove the template and carbon and using a Kepro RMP-700 resist pen (Allied 47 C 1102) fill in the outline traced on the foil. Be very careful the foil outlines don't touch. It is better to make the resist fill-in lines thinner rather than thicker. Don't worry about the drilling circles; pass the pen directly over the indents in the foil. When the excess copper is etched away the indents will indicate the drilling location.

Don't forget to place a drop of resist on the arrowhead pointing to pin 1 on IC1. Also put a dot of resist for the pin 7 lead in IC1. While pin 7 isn't used you will have to have a hole for the lead. Make certain the resist line from pin 3 doesn't touch the edge of the board, because if it does it will short to the mike case.

Allow about 15 minutes for the resist to dry and then immerse the board in at least ¼ in. of etchant solution. After about 20 minutes of agitation inspect the board to see if all the excess copper is dissolved. If any trace of unwanted copper remains, immerse the board again until all areas not covered with resist are free of copper. Thoroughly rinse the board and remove the resist with steel wool or resist solvent.

Inspect the board carefully to make certain no foil areas are touching. Also, check that there is at least 1/32-in. between the foil going to pin 3 and the edge of the board. If foils touch, or there is no clearance trim away some of the excess with a very sharp utility

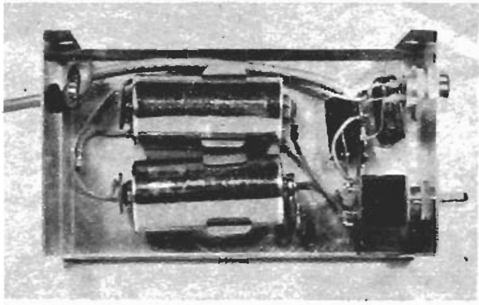
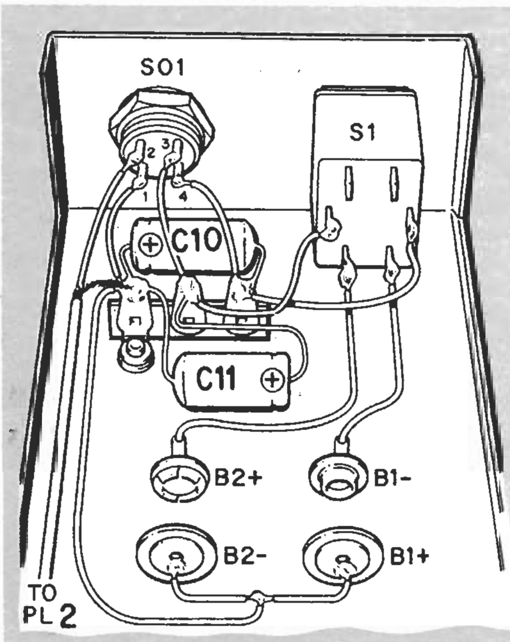


Fig. 6—Power box contains two batteries, on-off switch, input connector and output cable. Secure the batteries with clamps or wire loops.



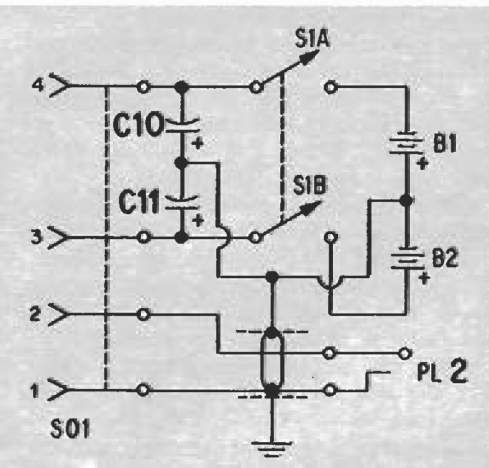


Fig. 7—Schematic above is of power supply. Circuit supplies positive, negative DC to compressor.

Fig. 8—Pictorial, left, shows parts layout in our power box. Watch polarity connections of C10,C11.

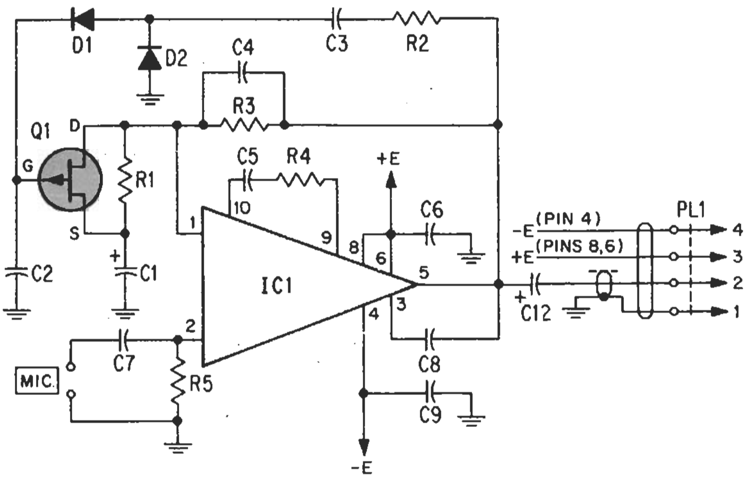


Fig. 9—Compressor schematic. Gain of op-amp (IC1) is determined by ratio of resistance of R3 to the resistance represented by Q1 in parallel with R1. At low input levels to op-amp, Q1 represents low resistance and op-amp's gain is maximum. As input signal increases, part of output at pin 5 is rectified by D1,D2 and DC is applied to Q1's gate, increasing Q1's resistance. This lowers gain of op-amp.

knife or an X-acto knife.

Drill the board with a No. 57 (0.043 in.) drill at the indents in the foil. With the parts in position it is more than likely that attempting to solder with a standard-size iron tip will result in shorts. The board must be soldered with a fine-point soldering tip. We suggest the Ungar 4037 heating element with a PL-111 pencil tip.

Except for R5 all resistors are mounted on end as shown in the photo in Fig. 4. All components must be inside the edge of the board or the board won't fit into the mike case. Except where shown, all parts must be flush with the board. Solder capacitor C1 in place with ¼-in. leads between its body and the board so that it can be folded flat against the board. Capacitor C9 should have about 1/32 in. of lead between its body and board so it can be tilted inward to allow clearance in the case.

Form the leads of capacitor C8 with a ⅛-in. jog as shown in Fig. 3 so C8 will be inside the edge of the board. The right side of C8 (Fig. 3) faces the edge of the board.

Install transistor Q1 with at least ¼ in. leads between it and the board. Integrated circuit IC1 is the last component to be installed and should have at least ⅛-in. clearance between its bottom and the board.

When installing IC1 make certain pin 1, which is opposite the tab on the case, is opposite the arrowhead on the foil. You don't get a second chance; if power is applied with IC1 incorrectly installed, it's bye-bye IC1.

Connect about 20 ft. of 3-conductor/single-shielded cable (Belden 8734 or equiv.) to the board. Connect the shielded hot wire to the connection from C10. Connect the shield to the foil ground, the red wire to the $+E$ foil and the black wire to the $-E$ foil.

Solder the original microphone wires to the board's input terminals. Slip the free end of the 20-ft. cable through the microphone case and slide the board into the case. Then slide the front of the microphone with the relief tube *under* the board toward the rear of the case. Connect wires to their matching front-half terminals and reassemble the case.

[*Continued on page* 112]

PARTS LIST

B1,B2—12.6 V mercury battery (Mallory TR-289 or equiv.)
C1—10 μf, 6-V electrolytic capacitor, single-ended leads (Lafayette 99 T 6073)
C2,C6—.22 μf, 25-V capacitor (Sprague 5C224, Allied 43 C 6698)
C3,C7—.1 μf, 3-V capacitor (Sprague HY-120, Allied 43 C 6671)
C4,C8—50 μμf, 500-V disc capacitor
C5—.002 μf, 500-V disc capacitor
C9—.05 μf, 500-V disc capacitor
C10,C11—100 μf, 15-V electrolytic capacitor
C12—2 μf, 6-V electrolytic capacitor
D1,D2—1N60 germanium diode
IC1—MC1433G integrated circuit (Motorola, Allied 50 F 26 MC1433G-MOT. $8.25. Not listed in catalog)
PL1—Four-contact male cable plug (Amphenol 91-MPM4L, Allied 47 C 0326 or equiv.)
PL2—Phone plug
Q1—2N3820 field-effect transistor (Texas Instruments)
Resistors: ¼ watt, 10%
R1—100,000 ohms
R2—4,700 ohms
R3—1 megohm
R4—100 ohms   R5—47,000 ohms
S1—DPST toggle or slide switch
SO1—Four-contact female chassis connector (Amphenol 78-PCG4, Allied 47 C 0331)
Misc.—Microphone (Allied Model 3311, Stock No. 12 C 7156), 5¼ x 3 x 2⅛-in. Minibox

Compress-O-Phone

Continued from page 33

After the mike is checked, pack the rubber grommet at back of handle with silicon adhesive to secure the cable. Install PL1.

The battery power supply is assembled in a 5¼ x 3 x 2⅛-in. Minibox; the layout isn't critical. Just make certain the polarity of C10 and C11 is correct and be sure of the connections to SO1. Plug PL2 is connected to a length of shielded cable—any length necessary.

Using the Compressor. The Compress-O-Phone can be connected to any input with an impedance of 10,000 ohms or higher. While it can be used with input impedance of less than 10,000 ohms (down to 2,000), the distortion increases sharply below 10,000 ohms. The compressor's output level is approximately 1 V (rms). This is just about right for a high-level (tuner or auxiliary) input, and we suggest using a high-level input rather than a mike input to avoid overload.

Turn on power with S1, speak into the mike and adjust the recorder's or amplifier's volume control for normal meter indication or sound level. That's the whole bit. As you record you will note that the level meter always peaks at the same spot. With transmitters, adjust the modulation level control for 100 per cent (actually 85 per cent is correct) modulation—it will be 100 per cent constantly. With PA amplifiers, adjust the level control until the system starts to howl, then back off slightly on the gain for maximum volume level at the verge of howling. Or, simply adjust the volume control for the desired volume level. Even if the speaker turns his head while talking the sound level will remain constant.

New Fun From Olden, Golden Radios

Continued from page 59

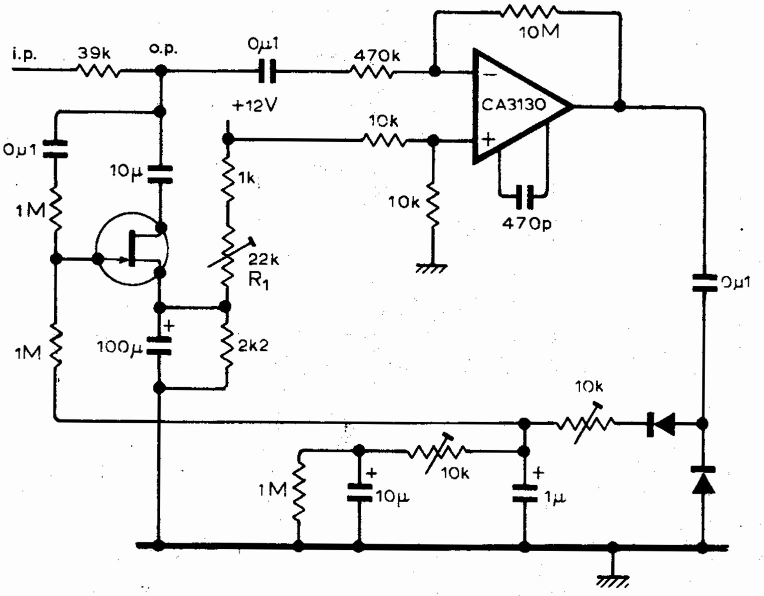
An Investment? While we recommend this hobby primarily for your personal enjoyment, it can also be an investment. Many newer collectors who are anxious to obtain early models of certain makes are now paying five to ten times what the original collector paid, and a few models are almost unobtainable unless a collector passes on and his estate is liquidated by the executor, or he retires to a smaller home and disposes of his collection piecemeal or as a job lot.

Speech compressor/limiter

This simple compressor/limiter, which was developed for p.a. applications, uses the voltage-controlled attenuator designed by D. Self, *Wireless World*, December 1975. Resistor R_1 sets the threshold voltage and the compression law. The output signal from the attenuator is made as large as possible by the inverting CA3130 before being applied to the rectifier and low-pass filter. This minimizes the effects of diode non-linearities and capacitor leakage. The low-pass filter is necessary to obtain a fast attack time of around 500μs and long decay time of about 1 min.

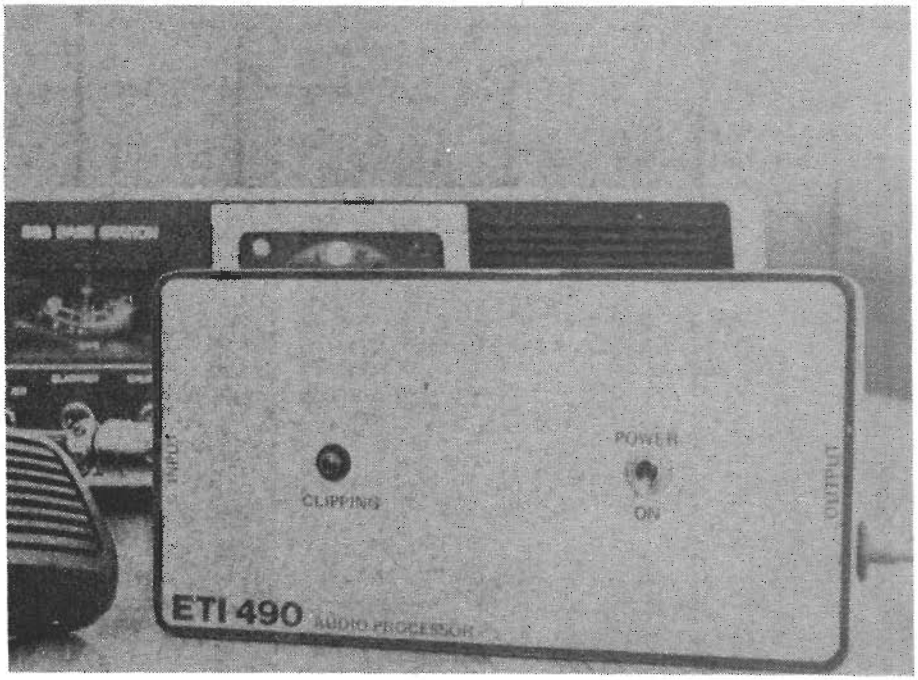
The circuit was used successfully with a microphone in a p.a. system with no noticeable distortion. Bandwidth of the circuit is 15Hz to 25kHz.

M. B. Taylor,
Kingston-upon-Hull.



Audio Compressor

Increase your talk power with this circuit.



THE HUMAN VOICE, being expressive at its best, varies considerably in level, even when one is speaking in a normal conversational voice. The peaks are considerably higher than the lower levels, which can give rise to problems when the speech waveform is being modulated onto a carrier by a transmitter. For example, if the mic gain control is set so that the peaks are just giving 100% modulation, then soft sounds can barely be heard whereas if the gain is turned up to give a higher level on vowel sounds, etc., then plosives (p-sounds) will give overmodulation and consequent splattering and poor speech quality.

A higher ratio of average power to peak voltage can be achieved by several methods, including compression or clipping of the audio signal and compression or clipping of the radio frequency signal. Radio frequency compression or ALC (automatic level control) is often used in the final stages of SSB transmitters.

Radio frequency clipping is the most effective method of increasing the average power; however it requires complex circuitry, since it is necessary to generate an SSB signal, clip, and then insert this signal into the transmitter IF chain.

Almost as effective as RF clipping is a combination of audio compression, clipping and filtering, which is relatively simple and can realise an improvement in signal to noise ratio of up to 5 dB on weak signals.

COMPRESSION

When speaking into a microphone it is desirable to keep the voice level as constant as possible. This can be quite difficult as any change in the distance to the microphone will cause a drastic change in its output. To overcome this a variable gain amplifier can be used which senses the average speech level and adjusts its gain accordingly for a constant output voltage. The compressor operates with a fast attack (gain reduction) and a slow decay (gain increase), to quickly respond to the voice while remaining at this level to prevent amplification of background noise during speech pauses.

CLIPPING

The average power contained in a speech waveform is quite low compared to the peak voltage, and much less than the average power of a sine wave of the same amplitude. If the low energy high voltage peaks are cut off at a preset level the remaining signal can be increased without overdriving the transmitter. The average power is therefore increased. Clipping will slightly change the sound of the voice but will increase the intelligibility of a weak signal, as well as preventing the transmitter from being overdriven by limiting the maximum signal voltage.

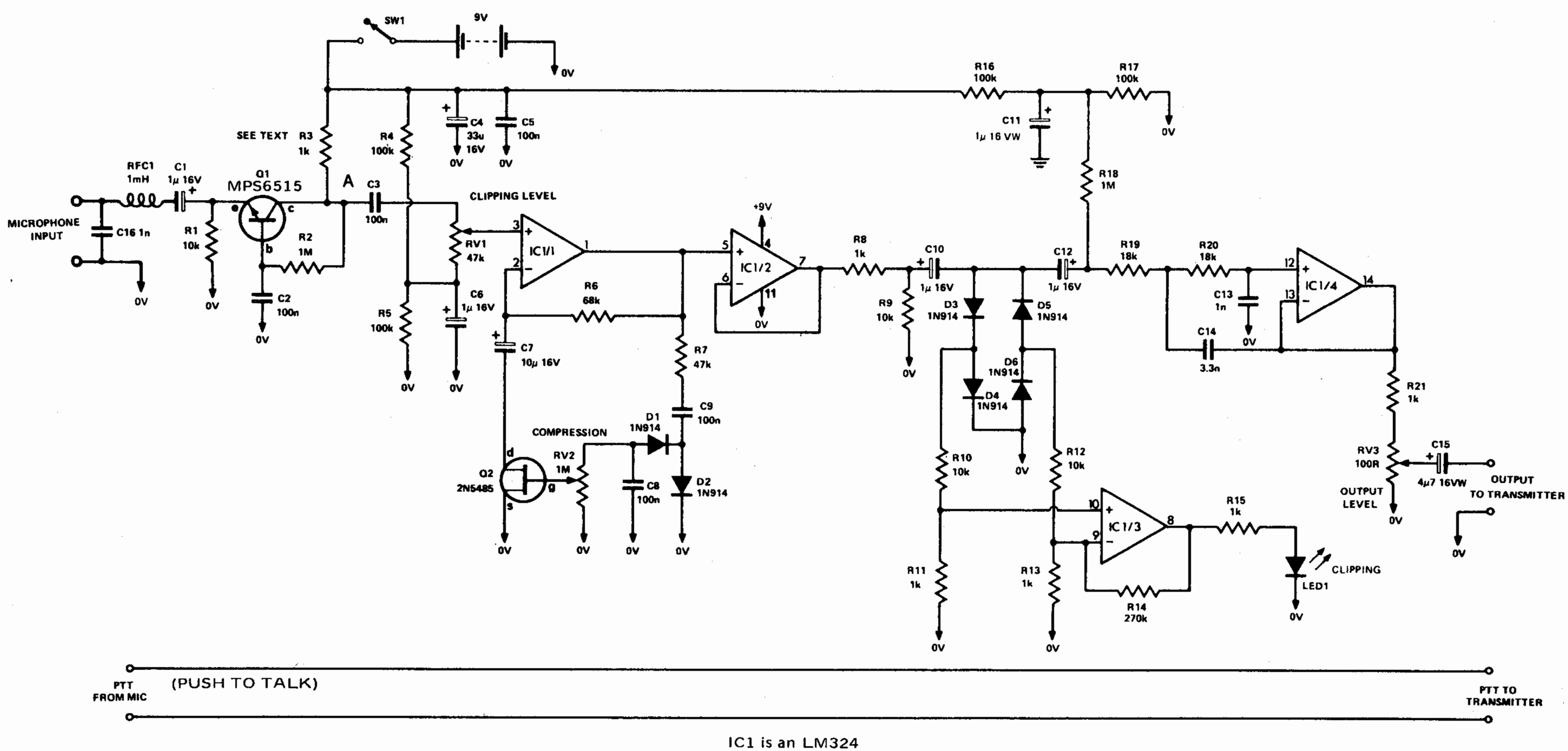


Fig. 1. Circuit of the Audio Processor.

HOW IT WORKS

The input is fed to a common base amplifier (Q1) and then to the gain control, RV1. The signal is then further amplified by IC1/1. Some of the output from IC1/1 is rectified and negatively charges C8. This voltage is then fed to Q2, a depletion mode N – channel FET. As the output of IC1/1 increases the voltage on the gate increases negatively and the impedance of Q2 increases. This increases the ratio of the feed back signal applied to the negative input of IC1/1 and the overall gain is reduced. The attack time is set by the time constant of R7 and C8, while the decay time is set by RV2 and C8.

IC1/1 is a buffer to isolate the peak limiter from the compressor input. R8 limits the output current of IC1/2 on peaks while R9 provides output bias current to prevent crossover distortion in the LM324 when driving capacitive loads. The diodes D3-D6 form the peak limiter by shorting any signal over about 1.5 V. When clipping occurs the voltage across D4 and D6 rises to 0.7 V. This voltage is used to turn on IC1/3 to give an indication of clipping by lighting LED 1.

The active low pass filter, IC1/4, removes the unwanted harmonics produced by clipping. RV3 sets the output level. The low frequency response is limited by the value of the coupling capacitors and C2.

FILTERING

When a waveform is clipped, high order harmonics are produced which, if allowed to reach the transmitter, would cause splatter and interference to neighbouring stations. A filter must be used after the clipper to rapidly attenuate all frequencies above 3 kHz, which are unnecessary for intelligibility. This is achieved by using an active filter with 12 dB/octave attenuation above 2.5 kHz.

CONSTRUCTION

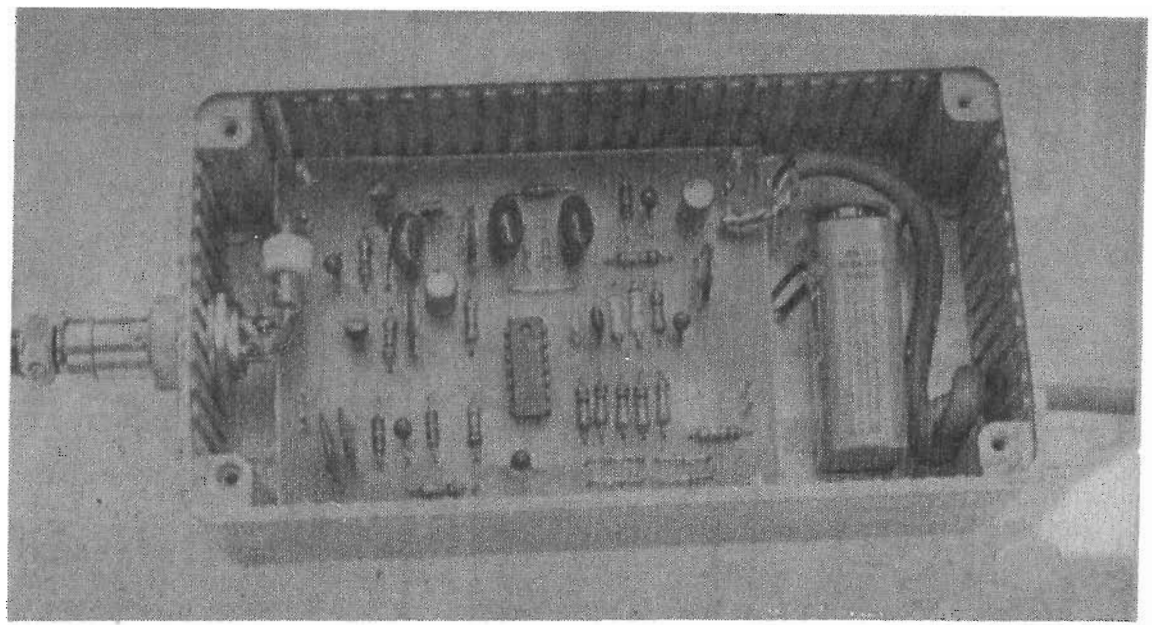
The speech processor is mounted in a diecast aluminium box to guard against feedback which can be caused by strong RF fields. Our box measured 150 mm x 80 mm x 50 mm deep. Either an internal 9V battery or the 12V transceiver supply can be used. The processor is designed to be used in the line from the microphone to the transmitter without any modification to either. A matching socket to the mic plug is used for the input and the output taken via a lead with a matching plug. The connections for the plug and socket vary between makes of transceivers and will have to be taken from the circuit diagram of the transceiver. The clipping indicator (LED 2) and the power switch are mounted on the front panel.

SETTING UP

Turn the compressor control to maximum and speak into the microphone at the greatest distance you are likely to use, (say 30 cm). Increase the gain control until the clipping LED flashes. If this point cannot be reached decrease the compression control and try again. The setting of these two controls is best determined by on-air tests. The output level control should be set so the RF indicator on the transmitter reaches the same peak as with only the microphone plugged in.

For high output, high impedance microphones, such as crystal types, Q1 can be omitted, RV1 replaced with a 1M trimpot and the input fed to point A on the circuit.

The gain of Q1 is proportional to the value of R3. Increasing its value increases the gain. To guard against feedback the lowest value possible should be used.



Inside view of the Processor. *The RF choke should be mounted as close as **possible to the input socket.***

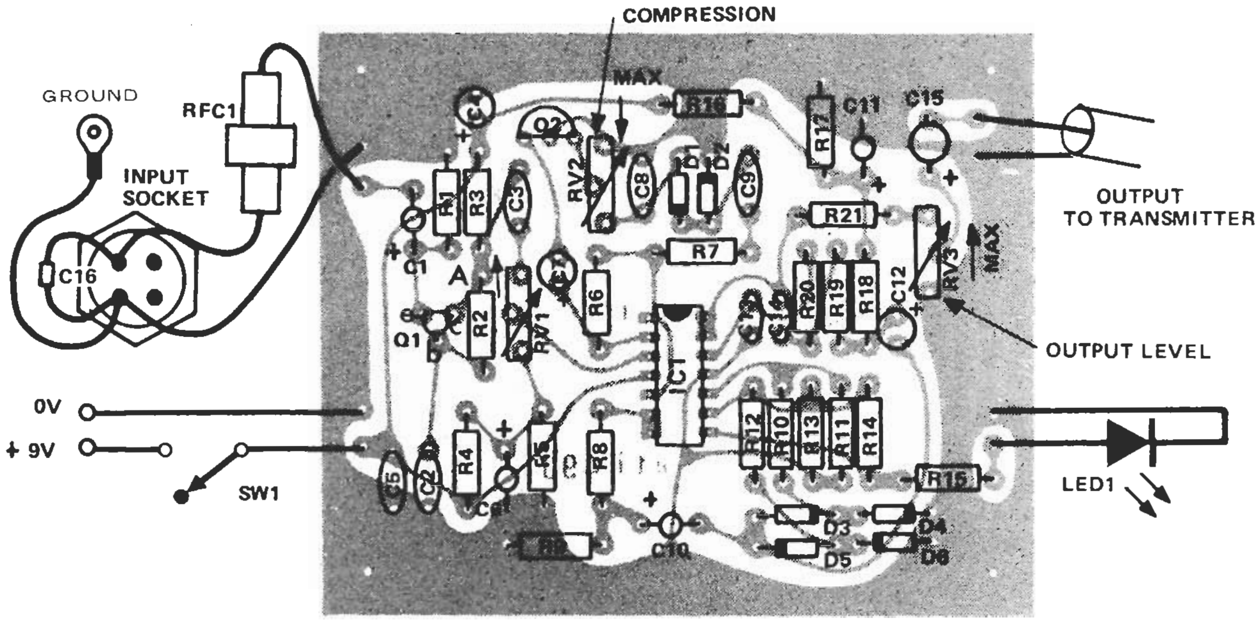
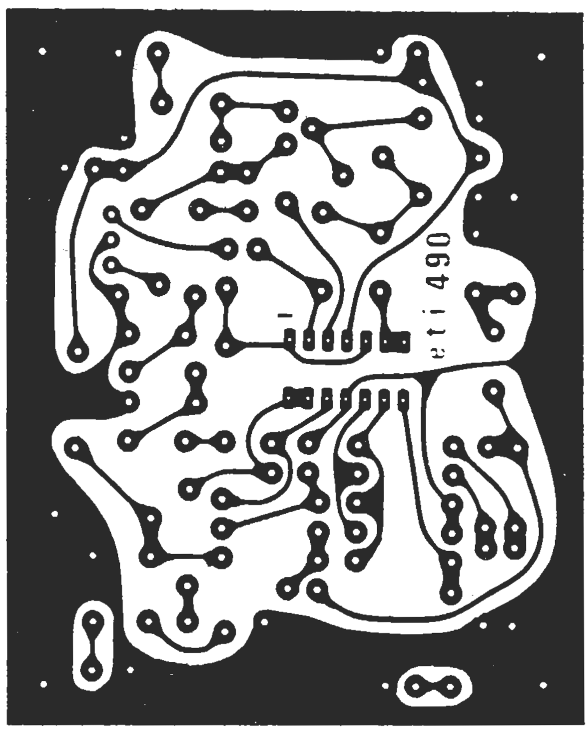


Fig. 2. Component overlay of the Audio Processor. Note the RF choke and capacitor mounted between the PCB and input socket.

Two source are available for the pcb for this project: B & R Electronics, P.O. Box 632F, Hamilton, Ont. L9C 6L9. Also Spectrum Electronics, 38 Audubon St. S. Hamilton Ont L8J 1J7.



PARTS LIST

RESISTORS all ¼W, 5%

R1 . . . . . 10k	R12. . . . . 10k
R2 . . . . . 1M	R13. . . . . 1k
R3 . . . . . 1k	R14. . . . . 270k
R4, 5 . . . . 100k	R15. . . . . 1k
R6 . . . . . 68k	R16, 17 . . 100k
R7 . . . . . 47k	R18. . . . . 1M
R8 . . . . . 1k	R19, 20 . . 18k
R9, 10 . . . 10k	R21. . . . . 1k
R11. . . . . 1k	

POTENTIOMETERS

RV1 . . . . 47k lin mini trimpot
RV2 . . . . 1M lin mini trimpot
RV3 . . . . 100Ω lin mini trimpot

CAPACITORS

C1. . . . . . 1μ 16 V electro
C2, 3 . . . . 100n
C4. . . . . . 33μ 16 V electro
C5. . . . . . 100n
C6. . . . . . 1μ 16 V electro
C7. . . . . . 10μ 16 V electro
C8, 9 . . . . 100n
C10–C12 . 1μ 16 V electro
C13. . . . . 1n
C14. . . . . 3.3n
C15. . . . . 4μ7 16 V electro
C16. . . . . 1n

SEMICONDUCTORS

Q1 . . . . . MPS6515
Q2 . . . . . 2N5485 FET
IC1 . . . . . LM324

D1–D6. . . 1N914 or sim
LED1. . . . Red LED

MISCELLANEOUS

RFC1. . . . 1mH or higher Radio Frequency Choke
SW1. . . . . SPST miniature toggle
Metal box to suit, 9V battery and holder, microphone plug and socket to suit, length multi wire shielded cable.

ressor/limiter

...n attenuator incorporating second ...monic cancellation circuitry

by D. R. G. Self, B.A.
University of Sussex

Compressor with Harmonic Cancellation Circuitry

Compression and limiting play an increasingly important role in the resources of a modern sound studio. The conventional function of signal level control is to avoid overload, but it can be used in the realm of special effects. To date, however, relatively few designs for high-fidelity compressor/limiters have been published.

The main design problem is the voltage-controlled attenuator, v.c.a., which increases attentuation of the input signal in response to a voltage from a control loop as shown in Fig. 1. In limiting, this circuit block continuously monitors the peak output level from the v.c.a. and acts to maintain an almost constant level if it exceeds a threshold value, or, in compression, allows it to increase more slowly than the v.c.a. input signal. This is illustrated in Fig. 2., which shows the input-amplitude/output-amplitude characteristic for both compression and limiting. Note that limiting makes use of a much tighter slope to ensure that the output voltage cannot exceed the chosen limit, and that the threshold (point of onset of attenuation) takes place at a higher level than for compression.

Traditionally, studio-quality compressor/limiters (as the two functions are so similar it is logical to produce a system that can be used for either compression or limiting) used one of two types of v.c.a. Either the audio signal was chopped at an ultrasonic frequency by a variable mark/space square wave – which requires complex circuitry and careful filtering of the audio output to avoid beats with tape-recorder bias frequencies – or it was attenuated by an electronic potential divider, one arm of which was a photoresistor, the control signal being applied via a small filament bulb. The last-mentioned has disadvantages because photoresistors are non-linear devices, therefore noticeable distortion is introduced into the audio signal, and the thermal inertia of the bulb filament limits the speed of attenuation onset.

Most moden compression systems use field-effect transistor operated below pinch-off as a voltage-variable resistance in a potential divider. This technique has many advantages; it is a simple, cheap, and fast-acting configuration that can provide an attenuation variable between 0 and 45dB. The only problem is that an f.e.t. is a square-law device, and tends to generate a level of second-harmonic distortion that increases rapidly with signal amplitude. A typical arrangement is shown in Fig. 3 – R_2, R_3 and C_2 allow the source of the f.e.t. to be set at a d.c. level above ground, so that a control-voltage that moves positive with respect to ground can be used, to avoid level-shifting problems in the control loop. This d.c. level is isolated from the input and output by C_1.

The distortion introduced by this circuit is at its worst for the 6dB attenuation condition, because at this point the drain-source resistance equals R_1, and the maximum power level exists in the f.e.t. Table 1 shows the level of second-harmonic distortion introduced into a sine-wave signal of 100mV r.m.s. amplitude, under the 6dB attenuation condition, for three different f.e.t. types. Measurements were made with a Marconi TF2330 wave analyser, higher-orders of harmonic distortion proved to be negligible amplitude in all cases. These measurements were made on one sample of each type of f.e.t. and, because production spreads are large, the results should be treated with some caution. However, it is clear that these levels of distortion are unacceptable for high-quality applications.

Fortunately, a technique* exists for reducing f.e.t. distortion to manageable levels, if the control-voltage is applied to the f.e.t. gate and summed with a signal consisting of one-half the voltage from drain to source, then the distortion level is dramatically lowered. The configuration in Fig. 4 shows a simple way of realising this; the signal fraction fed back is not critical and 10% resistors can be used for R_4 and R_5. Surprisingly, this distortion cancellation procedure leaves the attenuation/control-voltage characteristic almost unchanged. Table 1 shows the new maximum distortion values for 100mV r.m.s. input. (Note that the maximum no longer occurs at 6dB attentuation, but at a point that

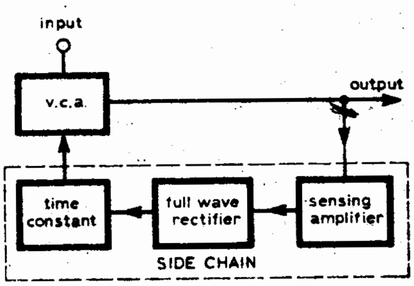


Fig. 1. Voltage-controlled attenuator with d.c. control loop.

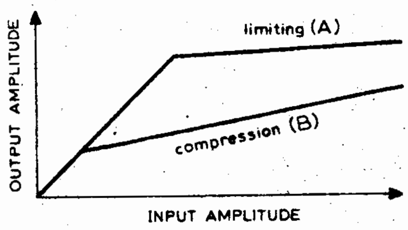


Fig. 2. Amplitude characteristics for compression and limiting—the last-mentioned uses an almost zero slope to prevent the output exceeding a preset level.

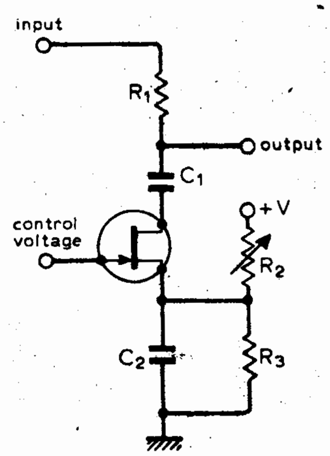


Fig. 3. Basic v.c.a. circuit providing up to 45dB of attenuation. This configuration introduces second-harmonic distortion which is greatest at 6dB of attenuation.

varies with the f.e.t. type, where cancellation is least effective.) From these results the 2N5457 and 2N5459 are superior, the 2N5459 was used in the final version of the v.c.a.

To determine appropriate signal levels in the v.c.a., measurements were made of maximum distortion generated, ie the v.c.a. was set to 2dB attenuation, against r.m.s. input voltage; results are shown in Table 2. The question now arises as to whether this distortion performance is adequate for a high-quality compressor/limiter. There is no general agreement as to the amount of second harmonic distortion that can be introduced into a program signal before it becomes aurally detectable, but 0.1% is a figure that is quoted. This means that the permissible input voltage to the v.c.a. would be restricted to below 100mV r.m.s. In practice, however, the attenuation level will be constantly changing, and because distortion level peaks fairly sharply with attenuation change, this level of distortion will only be present for a very small percentage of the time. In any case, second harmonic distortion alone has a relatively low "objectionability factor". The proof of the pudding is in listening to the compressor output signal; inputs of music around 200mV r.m.s. produced no trace of audible distortion. (Good class A power amplifiers and headphones were used for monitoring).

The control loop consists of an amplifier which senses the v.c.a. output level. A full-wave rectification system is normal practice because program waveforms have positive and negative peaks that can vary by as much as 8dB, and an 8dB uncertainty in the output level is usually unacceptable. A time-constant arrangement is used with the rectification circuit to control the attack and decay rates.

The output sensing amplifier in the system is a non-inverting op-amp which allows a high input impedance because the output impedance of the v.c.a. stage reaches a maximum of about 39kΩ at zero attenuation. The full-wave rectification system consists of a transistor phase-splitter driving two op-amp precision-rectifier stages in antiphase. The principle of a precision rectifier is illustrated in Fig. 5. The rectifying element is placed in the feedback loop of an op-amp, so that the effect of the forward voltage drop on the output voltage is divided by the open-loop gain. During positive half-cycles, if the input voltage exceeds the d.c. level stored on the capacitor C, the op-amp output swings positive and C is charged through diode D until its stored voltage is equal to the input voltage. Thus C takes up a voltage across it equal to that of the positive peak of the input signal. During negative half-cycles, and while the input is less than the voltage on C during positive half-cycles, the op-amp saturates negatively and D remains firmly reverse-biased. Obviously this is only a half-wave rectification circuit, the

Table 1. Second-harmonic distortion level introduced into a sine-wave of 100mV r.m.s.

Device	2N3819	2N5457	2N5459
2 nd harmonic at −6dB	13%	10%	8.9%
2nd harmonic with cancellation attenuation shown	0.39% 2dB	0.12% 10dB	0.12% 2dB

Table 2. Maximum distortion generated by various input voltages at 2dB attenuation.

Input (mV, r.m.s.)	2nd harmonic (%)
20	0.005
50	0.10
100	0.12
200	0.19
500	0.34
1,000	0.56

Table 3. Prototype calibration data and compression ratios

VC_2 (V)	Threshold (mV, pk)	Compression ratio
2.9	10	2.3
3.5	20	5.1
5.0	50	10
6.7	100	20
8.5	200	35
9.8	500	50

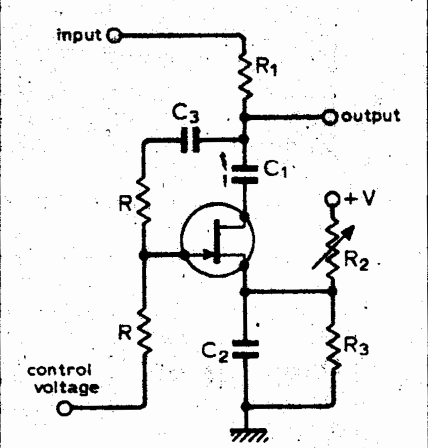


Fig. 4. Standard circuit technique for reducing f.e.t. distortion by summing half of the drain/source voltage with the control voltage.

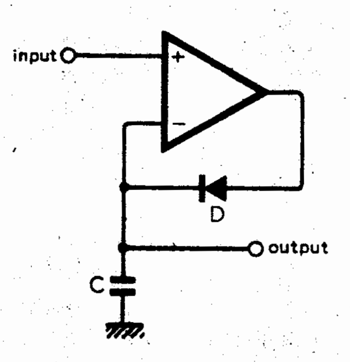


Fig. 5. Basic precision rectifier circuit where the rectifying element is in the feedback loop of an op-amp.

full-wave version uses two of these driven in antiphase, and charging a common capacitor. A resistance through which the charging currents flow determines the attack time, and another in parallel with C defines the decay time-constant.

The complete circuit is shown in Fig. 6. The v.c.a. is essentially as described above and the attenuation threshold is set by the variable resistance R_2. As the resistance is increased the level of control voltage required for attenuation to begin is reduced, and the system's input/output characteristic moves smoothly from A to B on Fig. 2. The threshold decreases and the compression slope becomes less flat as the system turns slowly from a limiter into a compressor by the manipulation of a single control. The output sensing amplifer consists of IC_1 and has a gain of 19 over the audio band. This is rolled off to unity at d.c. by C_5. Transistor Tr_2 and its associated components form a conventional phase-splitter driving IC_2 and IC_3, the precision rectifiers. The rectifier circuitry is more complex than implied above, three modifications have been made improve the performance. Firstly, IC_2 and IC_3 charge C_9 via current amplifier stages Tr_5 and Tr_6 otherwise the current-limited 741 outputs would be unable to provide enough current for the faster attack times (less than 1mS). Secondly, the feedback loop from C_9 to the inverting unputs of IC_2 and IC_3 is completed via a f.e.t. source-follower. Without this, C_9 would be loaded by the two 741 inputs, and this would severely limit the maximum decay times available. Incorporating the source-follower allows decay times of several minutes by using large resistance values for R_{27}. The conventional source-follower has a large negative offset voltage and is unusable in this application because due to their rectifying action IC_2 and IC_3 are unable to provide a voltage on C_9 that is negative of ground. This would be required to allow the source-follower output to be at ground when there is no input to the rectifers. However, if a modified source-follower is used, with a constant-current source and resistance combination in the source circuit, the offset voltage can be varied on either side of zero by manipulation of R_{24} which varies the driving current. The offset voltage is arranged to be plus 0.3V, to allow a large safety margin for thermal variations, component ageing, etc. This means that under no-signal conditions C_9 takes up a standing quiescent voltage of plus 0.3V. The effect of this is taken up in the calibration of R_2.

The third modification is the addition of R_{21}, D_3, and R_{22}, D_4. These two networks prevent IC_2 and IC_3 from saturating negatively, during negative half-cycles of their input voltage, by allowing local negative feedback through D_3 and D_4. This limits the negative excursion of the IC outputs to

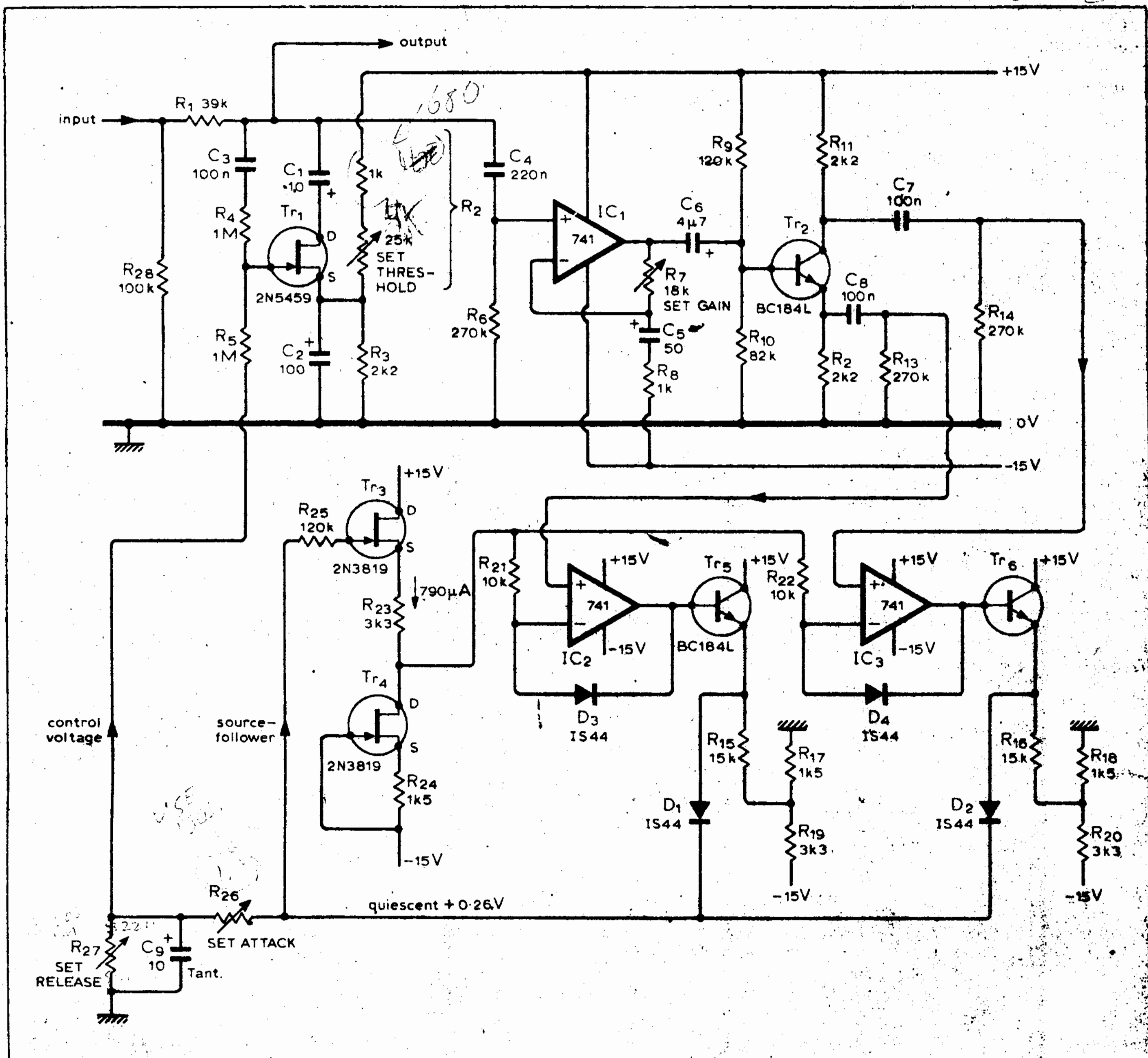


Fig. 6. Complete circuit where the output is taken directly from the v.c.a.—this may be buffered for loads greater than 100kΩ.

about two Volts. The prevention of saturation is necessary because the recovery time of the 741s causes the frequency response of the precision rectifier circuit to drop off at about 1kHz. The addition of the anti-saturation networks provides a frequency response that starts to fall off significantly above about 12kHz which is ample for our purposes as program signals have very little energy content above this frequency.

The final part of the circuit defines the attenuation time constants. Resistor R_{26} sets the attack time constant and R_{27} the decay time constant; these can range between zero and 1MΩ(220µs and 10s) for R_{26}, and 1kΩ and infinity(10mS and 20min) for R_{27}. They can be either switched or variable resistances, depending on the range of variation required.

The circuit in Fig. 6 shows the compressor output being taken directly from the v.c.a. This is only suitable if the minimum load to the output is greater than 100kΩ, otherwise the v.c.a. attenuation characteristic will be distorted by excessive loading. If lower resistance loads are to be driven a buffer amplifier stage must be interposed. The IC_1 amplifier stage is suitable for most applications, and its gain is $(R_7+R_8)/R_8$. For the unity gain case R_8 & C_5 can be eliminated and R_7 replaced by a direct connexion.

The compressor should be driven from a reasonably low impedance output (less than 5kΩ).

Construction is straightforward; the layout is not critical and the prototype was assembled on 0.1in matrix Veroboard. To set up the circuit R_{24} is adjusted so that the voltage across C_9 is about plus 0.3V with no signal input. The value required will vary due to production spreads in the f.e.t.s. To calibrate R_2 it is necessary to relate the level of input signal at which attenuation commences, with the voltage across C_2. This can be done with an oscilloscope, or preferably an a.f. millivoltmeter. As a guide the calibration data for the prototype is shown in Table 3, along with the values of the compression ratio (number of dBs the input must increase by to increase the output by 1dB). This data must be regarded as only a guide. It is worth noting that as the controlling factor setting the compression/limiting function is the voltage acrss C_2 R_2 could be replaced by a 1kΩ resistor connected to a remote voltage source.

The compressor/limiter is quite straightforward in use, provided a few points are kept in mind. Firstly, if it is being used in the limiting mode to prevent overload of a subsequent device, the fastest possible attack time should be used, to catch fast transients, and a decay time (say 100ms; $R_{27} = 10k\Omega$), allow the system to recover rapidly when the transient has passed. Secondly, if a noisy programme signal is being compressed a long decay time should be employed, "otherwise the noisy background will be faded up during quiet passages, and the familiar compressor breathing noises" will be heard. Finally, signals with a large v.l.f. content should be avoided or filtered, otherwise v.l.f. modulation of the signal will result, if a fast decay time is in use.

If a stereo compressor/limiter is constructed from two of the systems described above it is necessary to gang together R_2, R_{26} and R_{27} between the two channels. A direct connexion between the non-grounded sides of the two C_6s is also needed. It might be necessary to select matched f.e.ts to avoid stereo image shift during compression, due to differing attenuation characteristics in the two v.c.as. A well-smoothed p.s.u. providing ±15V should be used to power the compressor/limiter.

Components list

$IC_{1,2,3}$	741	
$Tr_{2,5,6}$	BC184L or equivalent	
Tr_1	2N5459	
$Tr_{3,4}$	2N3819	
$D_{1,2,3,4}$	IS44 or low-leakage equivalent	
R_1	39k	
R_2	25k variable, with 1k in series	
R_3	2.2k	
$R_{4,5}$	1M	
R_6	270k	
R_7	18k	
R_8	1k	
R_9	120k	
R_{10}	82k	All resistors
$R_{11,12}$	2.2k	(except R_2) ¼W
$R_{13,14}$	270k	
$R_{15,16}$	15k	
$R_{17,18}$	1.5k	
$R_{19,20}$	3.3k	
$R_{21,22}$	10k	
R_{23}	3.3k	
R_{24}	see text	
R_{25}	120k	
$R_{26,27}$	see text	
R_{28}	100k	
C_1	10μF 25V electrolytic	
C_2	100μF 25V electrolytic	
C_3	100nF 250V polyester	
C_4	220nF 250V polyester	
C_5	50μF 40V electrolytic	
C_6	4.7μF 40V electrolytic	
$C_{7,8}$	100nF 250V polyester	
C_9	10μF 16V tantalum bead	

Printed circuit boards

Wireless World has arranged a supply of stereo glass fibre p.c.bs. One off price is £3 inclusive; make cheques or postal orders payable to M. R. Sagin, 11 Villiers Road, London NW2.

"Electronic circuit calculations simplified"

We apologize that once again it has been necessary to postpone publication of Part 6 of this series, on LC circuits. The seventh, and final, part will be on active devices.

Literature Received

A British Standard, BS E 9111, on the quality assessment of low-power, fixed-value, non-wirewound resistors has recently been published, being the English text of a European Standard CEEC 40 100, with additions. Copies are available from BSI Sales Department, 101 Pentonville Road, London N1 9ND at £2.70.

Television distribution equipment from Wolsey is briefly described in their new short-form catalogue, available from Wolsey Electronics, Cymmer Road, Porth, Mid Glamorgan WW401

Full descriptions of a range of analogue and digital thermometers, recorders and associated equipment, thermocouples and application information are given in a new catalogue from Comark Electronics Ltd, Brookside Avenue, Rustington, Sussex BN16 3LF WW402

Moore Reed have sent two new leaflets, which give technical data and general descriptions of the company's ranges of stepping motors and rotary contact encoders. The leaflets contain useful descriptions of general interest on each of the classes of device. Moore Reed & Company Ltd, Walworth, Andover, Hants SP10 5AB..... WW404

The Annual Report and Accounts of the Independent Broadcasting Authority are now published, giving details on the financial position, technical developments, programmes and programming, advertising and engineering information. The Report is obtainable from H.M. Stationery Office or booksellers at £1.00.

General transducer techniques are described and specific information is given relating to a range of transducers for the measurement of pressure, displacement, acceleration and force in a new brochure from Sales Department, S.E. Labs (EMI) Ltd, Feltham, Middx. The publication is entitled "A guide to your transducer requirements" .. WW405

We have received a copy of the new catalogue of gears from Davall, which, in addition to data on a vast range of gear products, contains a technical section providing tabular information, conversions, glossary and bibliography. Davall Gear Company Ltd, Welham Green, Hatfield, Herts AL9 7JB WW406

Hewlett-Packard have prepared an eight-page guide to their range of optoelectronic devices, including red, green and yellow l.e.ds, alphanumeric displays and opto-couplers. P.i.n. diodes are also included. The brochure is obtainable from GDS Sales Ltd, Michaelmas House, Salt Hill, Bath Road, Slough, Bucks WW407

Self-balancing chart recorders which use fan-fold chart paper and feature ±1% accuracy, the SM range from Channel Electronics can provide up to six-point dotting with colour and a wide range of speeds. Channel Electronics (Sussex) Ltd, Cradle Hill Industrial Estate, Seaford, Sussex BN25 3JE WW408

A brochure on the E.E.V. range of travelling-wave tubes is now available, which gives descriptions of t.w.ts for high-capacity microwave links, including 10W and 20W types working at 4, 6, 8 and 11GHz. E.E.V., Chelmsford, Essex CM1 2QU...... WW409

A book on the design and use of heat pipes has been produced by Solek. Costing £17.50, the publication includes information on the theory and design of heat pipes, testing, wick materials and applications in 300 pages. The price includes one 12in, ⅜in diameter heat pipe with its data sheet. Solek Ltd, 16 Hollybush Lane, Sevenoaks, Kent WW410

Unitrode have published a 32-page semiconductor selection guide which presents information, in tabular form, on rectifying devices, transistors, diodes and i.cs. There is also a section on reliability, a list of application notes and mechanical details. Walmore Electronics Ltd, 11-15 Betterton Street, Drury Lane, London WC2 9BS WW411

We have received a copy of Pye Ether's new brochure on their range of transducers for industrial measurement. Descriptions are offered of devices for the electrical measurement of displacement, pressure, force, acceleration, vibration, speed, torque and temperature, and associated signal-conditioning, display and recording equipment is illustrated. Pye Ether Ltd, Caxton Way, Stevenage, Herts WW412

Highland Electronics have sent us a leaflet on an over-voltage trip circuit breaker, designed to operate within 2% of the setting in four ranges centred on 25V d.c., 50V d.c., 118V a.c. and 242V a.c. Contact rating of load-switching controls is 50A at 250V a.c. Highland Electronics Ltd., 33-41 Dallington Street, London EC1V 0BD WW413

A booklet from Fairhurst Instruments forms an introduction to a logic tutor and Karnaugh mapper, describing the construction and use of the equipment. It is available, on payment of 15p for postage and packing, from Fairhurst at Dean Court, Woodford Road, Wilmslow, Cheshire WW414

A booklet from Marconi presents specifications and applications information on two precision signal sources, employing signal generators and associated digital synchronizers, with which frequencies locked in 10Hz steps to 88MHz and 100MHz steps up to 520MHz can be generated at crystal stabilities. Marconi Instruments Ltd, St. Albans, Herts AL4 0JN WW415

Speed detection alternative

An alternative method of speed detection on the roads has been proposed, based on the Doppler effect in vehicular noise.* The method correlates the noise frequency spectrum as the vehicle approaches an observer with the spectrum as it moves away. The results of empirical investigation demonstrate that the Doppler shift can be extracted from a motor vehicle's noise and related to the vehicle's speed. Although sources of inaccuracy are significant at lower speeds, a resolution of ±5% was easily achieved at 60 m.p.h. Such a technique might be found useful in large scale traffic speed and density monitoring systems and may prove to be practical with the use of dedicated mini- or micro-processors. A single computation centre could simultaneously serve a large number of inexpensive microphone sensors. There is considerable interest in computer controlled traffic systems, and it's possible that the acoustic speed measuring technique could become economically competitive with the widely used radar method.

*Jakus, K. & Coe, D. S. "Speed Measurement Through Analysis of the Doppler Effect in Vehicular Noise," *IEEE Trans. on Vehicular Technology*, Vol. VT-24, Aug. 1975.

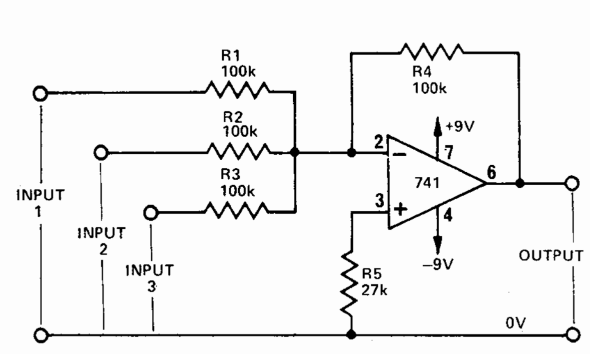


Fig. 14 Unity-gain inverting d.c. adder, or 'audio mixer'.

FIG. 15 shows how two unity-gain inverting d.c. amplifiers can be wired in series to make a precision unity-gain balanced d.c. phase-splitter. The output of the first amplifier is an inverted version of the input signal, and the output of the second amplifier is a non-inverted version.

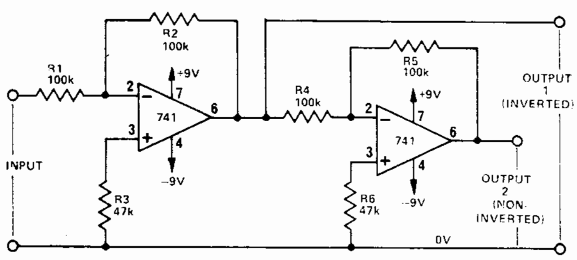


Fig. 15 Unity-gain balanced d.c. phase-splitter.

FIG. 16 shows how a 741 can be used as a unity-gain differential d.c. amplifier. The output of this circuit is equal to the difference between the two input signals or voltages, or to e_1-e_2. Thus, the circuit can also be used as a subtractor. In this type of circuit the component values are chosen such that $R_1/R_2=R_4/R_3$, in which case the voltage gain $A_v=R_2/R_1$. The circuit can thus be made to give voltage gain if required.

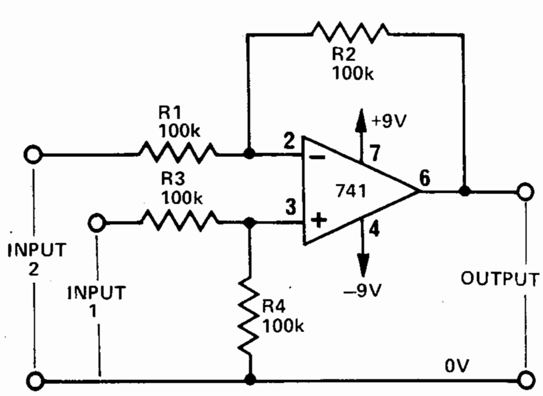
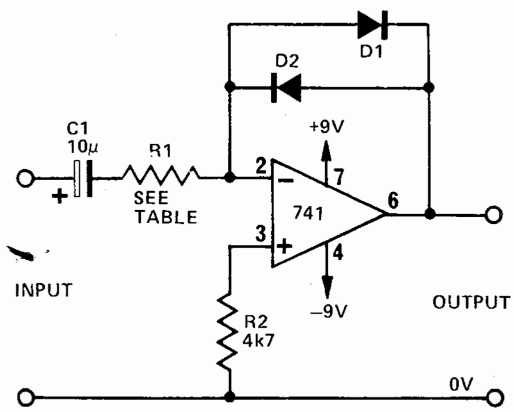


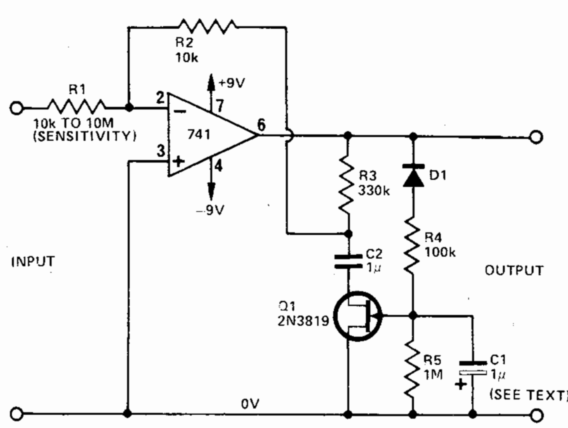
Fig. 16 Unity-gain differential d.c. amplifier, or subtractor.

FIG. 17 shows the amp can be made to act as a non-linear (semi-log) a.c. voltage amplifier by using a couple of ordinary silicon diodes as feedback elements. The voltage gain of the circuit depends on the magnitude of applied input signal, and is high when input signals are low, and low when input signals are high. The measured performance of the circuit is shown in the table, and can be varied by using alternative R_1 values.



INPUT VOLTS (R.M.S)	R_1 = 1kΩ		R_1 = 10kΩ	
	V_{OUT} (R.M.S)	GAIN	V_{OUT} (R.M.S)	GAIN
1mV	110mV	X110	21mV	X21
10mV	330mV	X33	170mV	X17
100mV	450mV	X4.5	360mV	X3.6
1V	560mV	X0.56	470mV	X0.47
10V	600mV	X0.07	560mV	X0.056

Fig. 17 Circuit and performance table of non-linear (semi-log) a.c. voltage amplifier.



V_{IN} (R_1 = 10KΩ)	V_{IN} (R_1 = 100KΩ)	V_{IN} (R_1 = 1MΩ)	V_{IN} (R_1 = 10MΩ)	V_{OUT}
50 mV	500 mV	5 V	50 V	2.85 V
20 mV	200 mV	2 V	20 V	2.81 V
10 mV	100 mV	1 V	10 V	2.79 V
5 mV	50 mV	500 mV	5 V	2.60 V
2 mV	20 mV	200 mV	2 V	2.03 V
1 mV	10 mV	100 mV	1 V	1.48 V
500 μV	5 mV	50 mV	500 mV	0.89 V
200 μV	2 mV	20 mV	200 mV	0.40 V
100 μV	1 mV	10 mV	100 mV	0.20 V
50 μV	500 μV	5 mV	50 mV	0.10 V

Fig. 18 Circuit and performance details of constant-volume amplifier.

DEVLIN M. GUALTIERI

An FET Audio Limiter

An fet core makes possible high s/n, symmetrical/asymmetrical function, frequency selection, control metering, and stereo coupling.

AUDIO LIMITING has always been important in broadcasting and recording to maintain a high signal-to-noise ratio by allowing a higher average signal level without fear of overmodulation or distortion. The heart of any audio limiter is its gain control element, which responds to a d.c. control signal derived from the audio peak amplitude above a reference level to reduce the channel gain and "limit" the audio signal. Various solid state devices can operate as gain control elements.

The operational transconductance amplifier (ota), such as the CA3080, has a linear control voltage/gain response, but suffers from a low s/n ratio. This is because the ota input signal must be attenuated to less than 100 mV to prevent excessive distortion and a typical ota limiter would have barely 50 dB s/n.

The diode as a variable resistance element suffers from the same input limitation. A light-emitting diode/light-dependent resistor (led/ldr) combination is a better quality gain control element.[1] The led eliminates the thermal response lag of older ldr units in which a filament bulb was used as a light source, but ldrs have a "memory" and are not wholly suitable for professional applications.

The field effect transistor (fet) has many advantages in comparison to other variable resistance elements. The d.c. control signal applied to the fet gate to vary the drain-source resistance (R_{DS}) is well isolated from an audio signal applied between the drain and source, eliminating d.c. "thumps." Fets also have a very high ratio of off-to-on resistance, but R_{DS} is linear only over a small range of drain-source voltages (V_{DS}). We might, therefore, expect the same trouble as for a diode limiter, but the fet is a three-terminal device with a separate control input. Consequently, we can supply a negative audio feedback signal that can be mixed with the d.c. control signal at the gate to reduce distortion by an order of magnitude. [2,3,4]

THE CIRCUIT

A realization of an fet limiter is shown in FIGURE 1. The fet, Q5, is in the negative feedback loop of low noise amplifier IC1A so that high values of R_{DS} lead to reduced amplifier gain. Transistor Q6 buffers the fet V_{DS} in an emitter follower configuration and applies this audio signal to the fet gate through a 1M resistor. This negative feedback connection reduces the signal distortion. The d.c. control signal is derived from the output of IC1A by the threshold rectifiers Q1 and Q2.[5] These Darlington transistors conduct heavily when their base-emitter voltage exceeds 1.2 volts. Q1 produces a d.c. control signal proportional to the level of positive peaks above the threshold, and Q2, which is driven by an inverted audio signal produced by IC1B, responds to the negative audio peaks. When Q1 and Q2 conduct, their collector voltage becomes more negative and the 0.47 capacitor, C, charges through the 4.7k resistor, R2, to this voltage. The d.c. voltage across C is buffered by the Q3/Q4 source follower and is applied to the IC2 meter driver and the IC3 level shifter. IC2 drives a 1 mA meter to give the conventional "backwards" metering of the limiting action, with Q7 altering the amplifier response in a piecewise linear approximation to match the Q5 fet limiting characteristics. IC3 allows the fet to be biased at its proper operating point and reduces the 20 volt range of the control signal across C to the one volt range needed to control the fet.

ADJUSTMENT AND OPERATION

Circuit power can be supplied by any well regulated ± 12 volt supply. Since d.c. levels are important, a dual tracking supply is preferred. Initially, the three 20k trim pots (meter, zero, bias) should be adjusted to mid-range. The zero pot should then be adjusted to give zero volts

Table I. Resistance and capacitance value for various limiter release times. The attack time is held constant at about 2 ms.

Release Time seconds	R2 ohms	C mFd
0.1	4.7 k	0.47
0.3	1.3 k	1.47*
1.0	390	4.7
3.0	120	14.7*

*Parallel capacitors.

Table II. Meter calibration for the author's limiter. The meter has the conventional "backwards" response.

Meter mA	Limiting dB
1.0	0
0.8	1.5
0.6	2.0
0.5	4.0
0.4	6.0
0.2	11.0
0	16.0

Dr. Gualtieri is with the University of Pittsburgh, Pittsburgh, Pa.

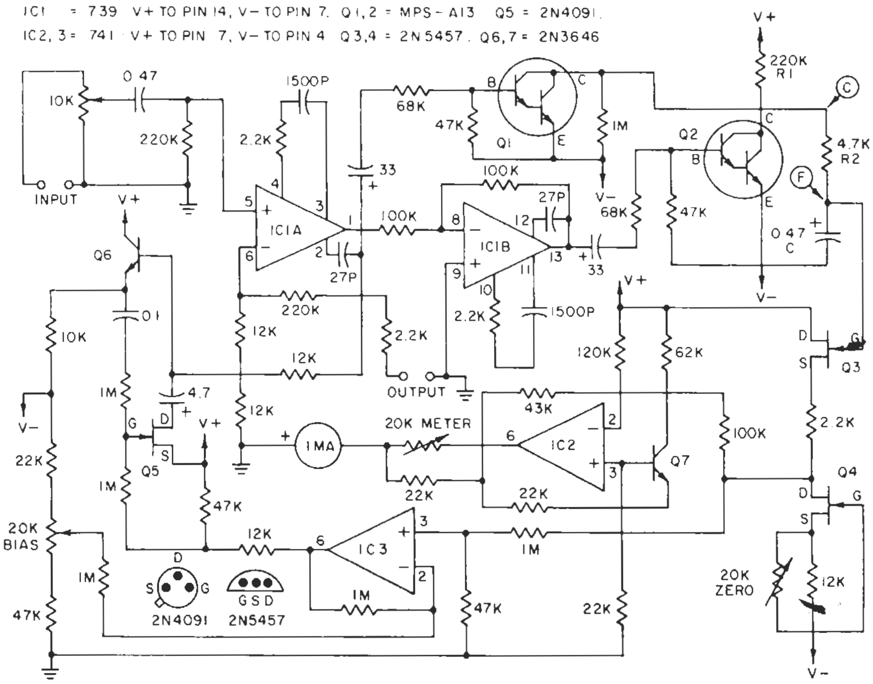


Figure 1. Schematic of the fet audio limiter. Attack and release times are set by R1, R2 and C. Points "C" and "F" allow stereo coupling of two limiters, as explained in the text.

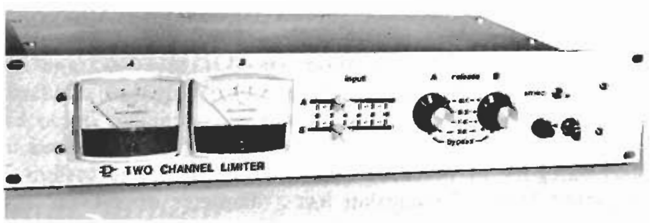
at the drain of Q4 when point F is shorted to ground. Use a vtvm or other high impedance voltmeter for this adjustment. The meter pot is adjusted for full-scale deflection of the meter. The bias pot is then adjusted to give maximum negative voltage at its wiper, biasing Q5 fully *on* and giving maximum gain for IC1A. A low level audio signal, small enough to prevent the output voltage from reaching the limit threshold (about —30 dB) is then applied to the limiter input, and an oscilloscope or sensitive a.c. meter is connected to the limiter output. The bias pot is then adjusted to give a small, but reproducible, gain reduction, about one or two dB. The limiter is now adjusted for proper operation.

The limiter attack time is set by R2C, and the release time is basically R1C. For the values given, the attack time is 2 ms and the release time is 0.1 seconds. Changes in these times can be accomplished by changes in R2 and C (see Table I), leaving R1 as 220k for proper circuit operation. Two milliseconds is judged by most limiter manufacturers as an optimum attack time. Table II shows the meter calibration for my unit. The 3 dB points of the limiter are below 20 Hz and above 40 kHz, and the input equivalent noise is —75 dB for no limiting, to —90 dB at full limiting.

MODIFICATIONS

Several modifications of the basic fet limiter can be made. The input impedance can be reduced by lowering the resistance of the input pot to give a better noise figure for low impedance sources. For operation as a mic channel limiter, a microphone preamp can be added before the input pot to bring the signal up to line level. The 220k resistor in the feedback loop of IC1A, which controls the gain of the controlled stage, and thus the limiting threshold, can be increased to give a larger control range at the expense of s/n in the quiescent state. The gain of IC1B can be decreased by changing the feedback resistor to allow the asymmetrical limiting desirable for a.m. broadcasting. Decreasing the gain of IC1B to 0.80 would give 125 per cent limiting on positive peaks and 100 per cent limiting on negative peaks, referred to the input phase.

Frequency selective limiting, which is desirable for f.m. broadcasting, can be easily accomplished by reducing the 33 mFd capacitors in the base circuits of Q1 and Q2 to give the 75 or 25 μ time constant of pre-emphasis. NAB equalization circuitry can also be inserted at this point to give a "recording limiter" response. For a constant R1, various values of R2 and C can be switched by a selector switch to give variable release time, as in Table I. Stereo coupling of two limiters to preserve L + R balance is easily accomplished by tying the points C and F of one unit to the corresponding points on the other. In this way, the channel specifying the greatest gain reduction will dominate.



The two channel limiter with variable release time. Stereo coupling is accomplished by a dpst switch which connects the C and F tie points of the two limiters.

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Automatic gain control systems

Design considerations and parameters

by N.A.F. Williams, B.Sc. M.I.E.E.

The purpose of all automatic gain systems is to control a variable gain amplifier so that its output voltage stays approximately equal to a reference voltage for all values of input signal within certain limits. These limits define the working range of the system. To carry out this function, negative feedback is used. It is therefore worthwhile considering the parameters which define the operation of a negative feedback amplifier, as shown in Fig. 1.

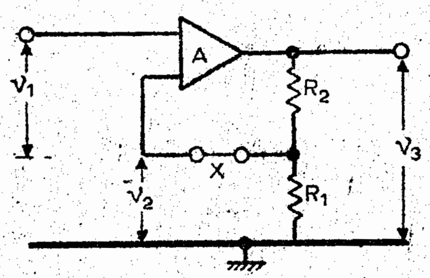


Fig.1. Basic negative feedback amplifier.

The differential amplifier has a gain A, and the output voltage V_3 is equal to $A(V_1 - V_2)$. Voltage V_2 is that fraction of V_3 defined by the potential divider R_1 and R_2. If $R_1/(R_1 + R_2) = B$, then $V_2 = BV_3$ and a simple calculation shows that provided $AB \gg 1$ the magnitude of the gain V_3/V_1 is approximately equal to $1/B$. It should be noted that open loop gain A is the ratio V_3/V_1 when feedback link X is broken. The closed loop gain is approximately equal to $1/B$, and is the ratio of V_3/V_1 when the link is closed. The loop gain, of magnitude AB, is the gain around the feedback loop which determines the stability and precision of the amplifier.

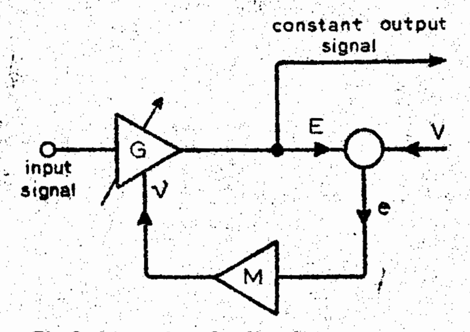


Fig.2. Negative feedback arrangement used in automatic gain control systems.

The negative feedback arrangement used in automatic gain control systems differs from Fig.1, and is shown in Fig.2. The input signal passes through an amplifier of variable gain G and, usually after rectification, is compared with the reference voltage V. The error voltage e is then passed through an amplifier of gain M whose output is control voltage v. Loop gain is determined by M multiplied by the transfer functions of any networks present in the loop. For example, a rectifier converting the output of amplifier G to the direct voltage E before comparison with V, and the factor relating v to G. Let us assume that these are all constants, so that the loop gain $L = KM$ where K is a constant. Besides being responsible for the stability and transient response of a negative feedback system, the loop gain decides what error may exist in the loop under steady state conditions, or under varying input signal conditions where the frequency of variation lies within the bandwidth of the feedback system. In the case of a.g.c. systems, it determines the accuracy of control as shown by the following equations. In Fig.2, $E = Le$ and $e = V - E$. Therefore, $e = V - Le$ or $e\ (1 + L) = V$. From the last equation, if the loop gain $L = 100$ then $e = V/101$ so the actual output differs from that required by only about one per cent. Changes in loop gain will cause corresponding changes in the accuracy of control. For example, reducing the gain to ten reduces the accuracy to within ten per cent. Also, the loop gain is not independent of frequency because all practical systems include frequency sensitive components. In general L has the characteristic of a low pass filter which has a constant amplitude C up to frequency F. Beyond this point the frequency sensitive components begin to take effect and reduce the magnitude of L. The a.g.c. system will respond with an accuracy determined by loop gain $L = C$ for variations of input signal which occur within the frequency range 0 to F. For frequencies greater than F the system will respond with a reduced accuracy. In operation the output of amplifier G is nearly constant for all values of input signal. Hence, for constant loop gain a constant absolute change of output voltage from amplifier G for a given change of y is required for all values of G. If the relationship is considered to be linear, as shown in Fig.3 (a) a change of v gives a constant

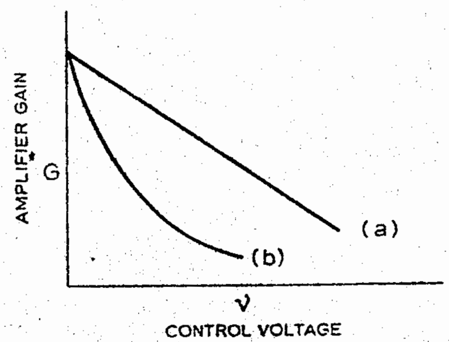


Fig.3. Relationships of amplifier gain G versus control voltage v. Linear trace (a) will not provide a constant loop gain but exponential curve (b) produces a constant loop gain for all values of G.

change of G. Numerically however, it does not provide the desired output voltage for all values of G. For example, let G vary from 100 to 1000 and let the required output voltage be 10V. When the gain is 1000, the input voltage is $10/1000 = 0.01$V, and when the gain is 100, input voltage is $10/100 = 0.1$V. In each case let v change by an amount which causes G to change by say 20 while the input voltage remains constant at either of the two values corresponding to a gain of 100 and a gain of 1000. When the gain is 1020 the output voltage is $0.01 \times 1020 = 10.2$V, and when the gain is 120 the output voltage is $0.1 \times 120 = 12$V. Thus when G is 1000 a given change of v alters the output voltage by 0.2V, but when G is 100 the same change of v alters the output voltage by 2V. This means that the loop gain has changed by a factor of ten, and is greater at the lower value of G. It should be noted that this is a variation in the low frequency flat part of the loop gain characteristic. For any given setting of this zero-frequency-response, reactive elements that may exist within the loop will modify this curve in the usual way as it extends into the higher frequency region.

As a linear relation between v and G will not provide a constant loop gain the preceding calculation shows that a constant percentage change of G is required, that is $dG/dv/G$ = a constant, or $dG/dv = KG$ where K is a constant. Curve (b) of Fig.3 shows such a characteristic. If $G = Ke^{-av}$ then $dG/dv = -Kae^{-av}$ and $dG/dv/G = -Kae^{-av}/Ke^{-av} = -a$. This indicates that the relationship between v and G should be exponential if the loop gain is to remain constant for all values of G. Because $G = Ke^{-av}$, $\log_e G = \log_e K - av = K_1 - av$ where K_1 is another constant, and as $\log_n m = \log m / \log n$ to any base of logarithms, $\log_{10} G = \log_{10} e(K_1 - av) = K_2 - K_3 v$ where K_2 and K_3 are two more constants. This is the equation of the straight line shown in Fig.4 and shows that G in decibels versus v produces a straight line with the desired characteristic.

Variations in the zero-frequency loop gain not only cause changes in the accuracy of the a.g.c. system but can cause instability at settings of G that

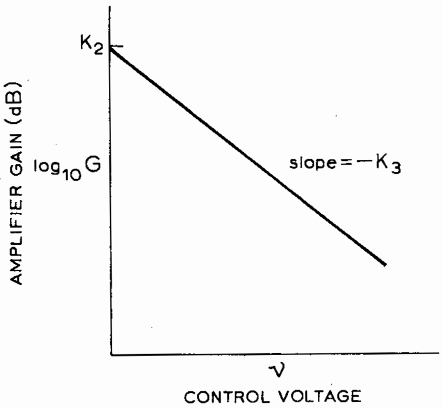


Fig.4. Gain in dB versus control voltage v produces a straight line with the desired characteristic.

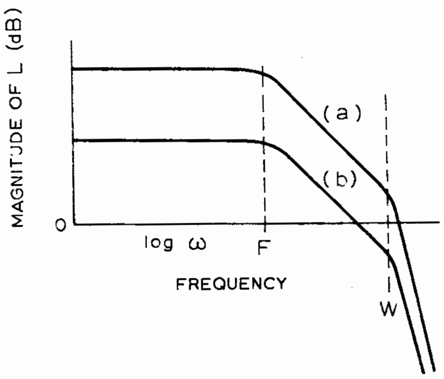


Fig.5. Graphs illustrating that variations in zero-frequency loop gain can cause instability. Curve (a) crosses the 0dB point (unity-loop gain) with a slope of 12dB per octave corresponding to a loop phase shift of 180 degrees. Curve (b) is stable because the loop phase shift is 90 degrees at unity loop gain.

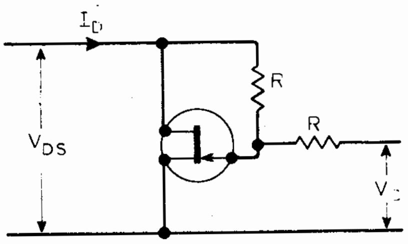


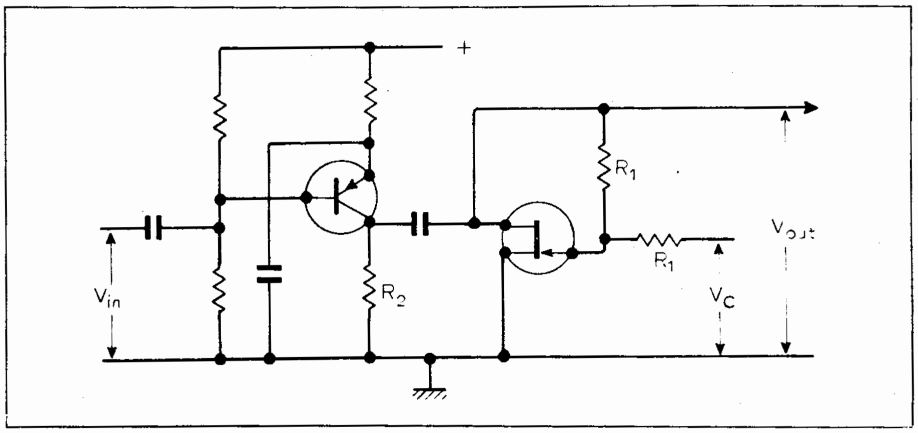
Fig.6. Variable resistor using a f.e.t. The feedback resistor linearises the effective resistance.

give the highest value of loop gain. This is demonstrated in Fig.5 where curves (a) and (b) have the same form but different zero-frequency gain. The amplitude falls off at 6dB per octave from frequency F to frequency W, and at 12dB per octave from frequency W onwards. The system represented by curve (b) is stable because unity loop gain (0dB) occurs with a phase shift around the loop of only 90 (+180) degrees, as indicated by the 6dB per octave rate of change of amplitude assuming a minimum phase network. The system of curve (a), however, is unstable because the 0dB line is crossed at a slope of 12dB per octave, corresponding to a loop phase shift of 180 (+180) degrees. It is difficult to maintain the loop gain constant, and in some systems considerable variations may be permissible. Knowing the extent of the variation allows its effect to be calculated, and gain controlled amplifier circuits which approximate to an exponential relation between G and v will therefore be suitable.

Integrated circuit amplifiers, intended mainly for r.f. or i.f. amplification, are available from several manufacturers. Some of these amplifiers give an approximately straight line characteristic when their gain in decibels is plotted against their a.g.c. control voltage, at least over most of their working range. These are very suitable for applications requiring high constancy of loop gain. Considering simple bipolar transistor and field effect transistor amplifiers, neither has an inherent suitable relationship between gain and some easily controllable parameter such as emitter or drain current. However, if the gain of the common-emitter bipolar transistor amplifier is plotted in decibels against emitter current it is found that the gain varies approximately linearly with emitter current in the low emitter current region. The gain of a common-source field effect transistor amplifier is proportional to the square root of the drain current, and this relationship also approximates to the desired characteristic for low values of drain current. An alternative use for the f.e.t. is as a voltage controlled variable resistor, and Fig.6 shows a well known arrangement of feedback from drain to gate which linearises the effective resistance of the f.e.t. The drain to source resistance R_{ds} of this circuit is given by the expression $R_O/(1 - V_c/2V_p)$ where R_O is the drain to source resistance when the voltage between gate and source is zero, V_p is the pinch off voltage, and V_c is the control voltage shown in Fig.6. For a given device, R_O and V_c are constants, and the expression can be written as $R_{ds} = k_1/1 - k_2 V_c$ where k_1 and k_2 are constants. Plotting this equation gives a curve which, although not an exponential, does approximate to one and is suitable for some applications. The maximum possible slope of the R_o versus V_c graph is fixed by the values chosen for the feedback resistors in Fig.6 although for clarity the effect of these resistors has not been included in the previous expression for R_{ds}. By adjusting the values of R the degree of approximation to an exponential curve can be altered. To make use of this voltage controlled variable resistor the controlled amplifier gain must be made proportional to R_{ds}. This can be achieved by letting R_{ds} form the collector load resistor of a grounded emitter transistor amplifier, as shown in Fig.7, in which R_2 is very much greater than R_{ds}.

Another method of maintaining roughly constant loop gain for varying amplifier gain is to make straight line approximations to the desired response curve by using diodes to provide the break points in the slopes of the straight lines. No doubt readers will visualise other possibilities.

Fig.7. A.g.c. system where a f.e.t. used as a variable resistor forms the collector load of a grounded emitter amplifier.



Bi-FET op amps simplify AGC threshold design

by John H. Davis
Warm Springs, Ga.

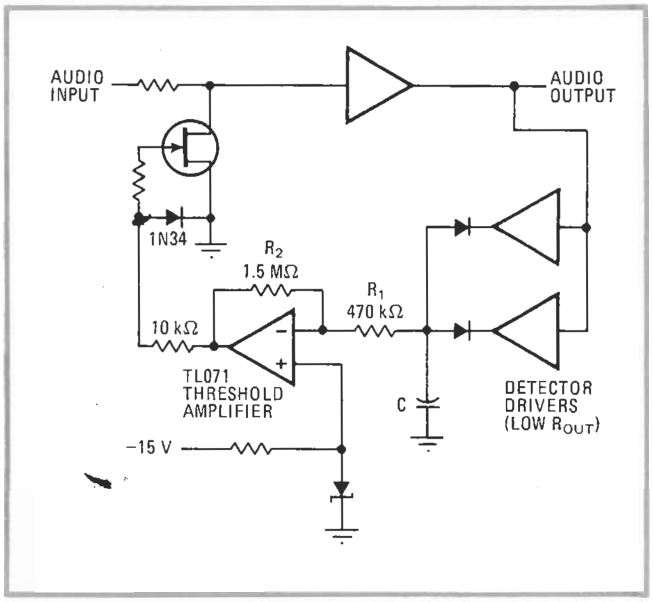
Operational amplifiers with the bandwidth and input impedance available using bipolar–field-effect-transistor (bi-FET) technology are well suited for integrating the threshold detection and automatic-gain-control amplification functions in audio limiters or receiver AGC circuits. Generally, such circuits are implemented with discrete components. But this often entails component selection and critical trimming adjustments, or both. An op amp approach makes an AGC design more predictable, stable, and easier to troubleshoot.

The circuit of Fig. 1 requires only one adjustment, to zero the output of the TL071 op amp under no-signal conditions. In this circuit, a control voltage is required over the range from zero (at full gain) to the negative value corresponding to the FET's cutoff voltage. A zener diode supplies the reference voltage. It is connected in a way that makes use of the common-mode rejection properties of the op amp; thus, R_3 nulls the static output, which thereafter is quite stable.

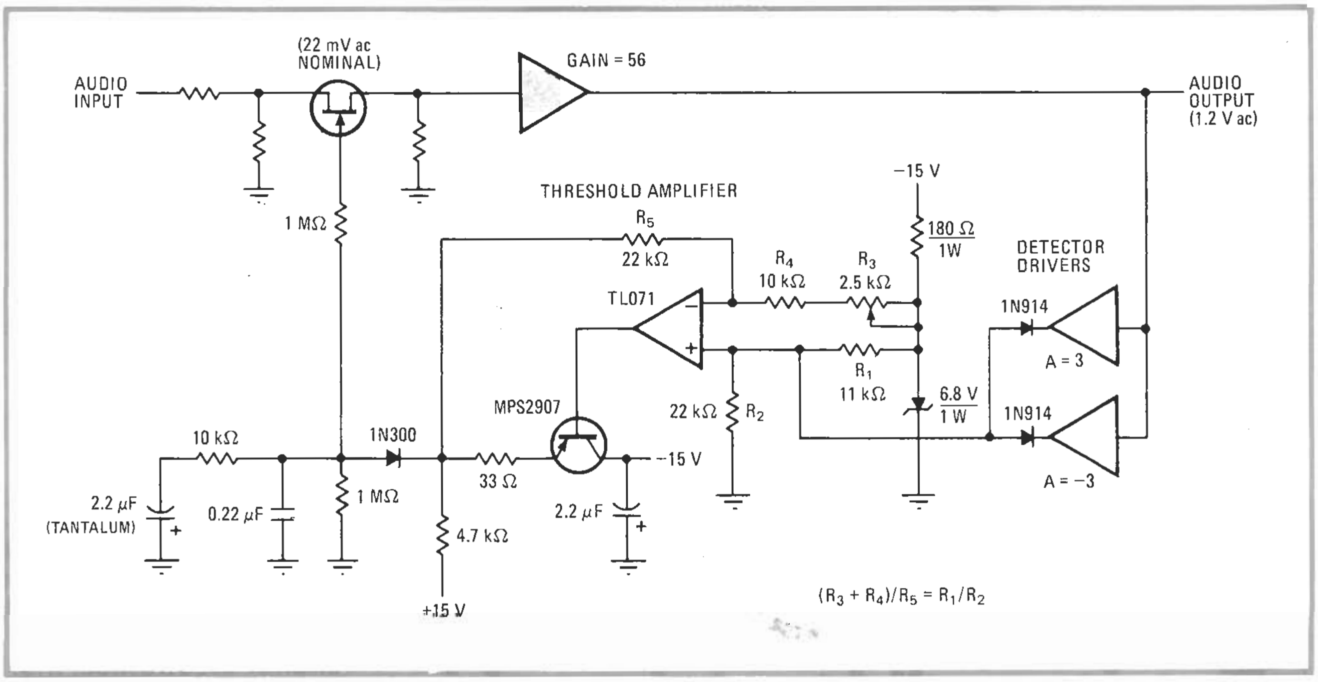
The threshold is the voltage appearing at the junction of R_1 and R_2, plus the forward drop of the detector diodes, and can be readily computed for any desired limiting level. For the detected peaks, the threshold detector has a voltage gain of:

$$A_{det} = (R_1 + R_2)/R_1$$

Not much gain is ordinarily required; too much imposes tighter tolerances on driver gain, diode properties, and



2. Simpler. Threshold detection, time constants, and amplification are consolidated in this single stage. For a receiver's i-f strip, an emitter follower is recommended, however. The control voltage here varies from a fixed negative value toward zero.



1. Easy play. Only one adjustment to zero the output of the TL071 op amp under no-signal conditions is needed in this AGC threshold amp. Good performance is achieved by using the amp's common-mode properties. The control voltage must vary from zero to a negative value.

trimmer adjustment. Driver gain can be adjusted, within output swing limits, to tailor limiting slopes.

The emitter follower improves the attack time of the time constant network. The dual set of time constants shown prevents short-duration peaks from depressing system gain longer than necessary.

Further simplification (Fig. 2) is possible if the driver amplifiers have low output impedance. Here the threshold detection, time constants, and amplification are consolidated in a single stage. This consolidation around one op amp means that little additional circuitry is needed when an FET is the voltage-controlled element.

The circuit assumes the control voltage must vary from a fixed negative value toward zero as gain reduction is needed. With no signal, the threshold op amp is referenced to the desired voltage by the zener diode. As long as no signal peaks are applied to time constant capacitor, C, the op amp acts as a voltage follower. Detected peaks charge C more negative than the reference, and the difference is amplified by a gain of $R_3 \div R_1$. This shifts the control voltage toward zero.

The release time constant is determined by C and R_1 and R_2. (Although only a single capacitor is shown, a dual arrangement as in Fig. 1 can be used.) The simplified circuit shown in Fig. 2 can also provide a fixed positive voltage that ranges toward zero for gain reduction if all the diodes and the reference-voltage polarity are reversed. □

Agc prevents noise build-up in voice-operated mike

by Russell S. Thynes
Kirkland, Wash.

Hands-free operation of intercoms has several advantages over push-to-talk intercom systems. Constantly keyed "live" microphones, however, have the disadvantage of receiving undesirable environmental noise in the absence of speech. When such mikes are used in conjunction with intercoms having automatic gain control in the microphone mixing stages, this environmental noise will produce a swelling tide of sound each time normal communication is interrupted.

Shown here is an agc-VOX (voice-operated switch) scheme that allows constantly keyed microphones to be used in noisy environments without suffering from the effects of noise build-up.

Although the operation of the circuit is twofold, the function is primarily that of a gain-clamped agc circuit. Part (a) of the figure shows the transfer function of an agc with gain clamping and that of typical configuration. For input levels below those of normal speech, clamping the gain to a fixed value limits the area of the gain curve, reducing noise susceptibility—but without placing restrictions on the dynamic range of the agc itself.

The circuit is shown on the right (b). The agc section consists of operational amplifier A_1 and transistor Q_1, with diode D_1 and capacitor C_1 deriving the feedback control voltage. Q_1 is placed in a T configuration to achieve a wide control range and to ensure low levels of distortion. Distortion is further reduced by the gate-biasing resistors R_6 and R_7. As configured, this agc should provide 30 decibels of gain control with less than 0.5% distortion for most of the audio range.

A_3 is arranged as an adjustable noninverting amplifier, the gain of which can be varied from 20 to 40 dB. R_{12} (also in a T configuration) allows the user to set the VOX sensitivity to offset environmental noise conditions. A_4 simply compares the detected output of A_3 with a reference and switches to either a high or a low output limit depending on the VOX input level.

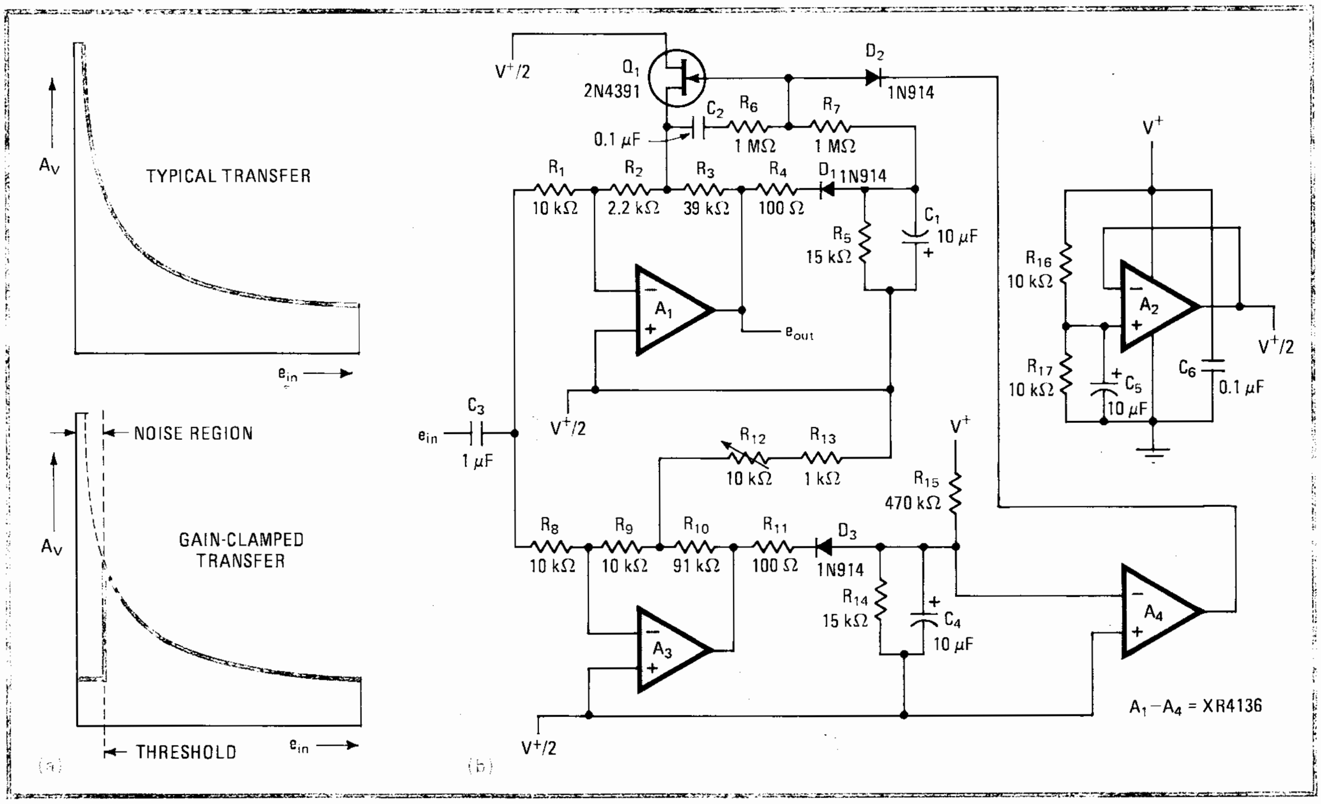
When input levels to the agc are below the VOX sensitivity setting, the output of A_4 will be at its lower limit, biasing Q_1 off and thus clamping the gain of A_1 to $(R_2+R_3)/R_1$.

When the VOX sensitivity level is exceeded, however, the comparator output swings to its upper limit and effectively disconnects the VOX from the agc feedback loop through blocking diode D_2. The gain is then expressed as:

$$-\frac{R_2+R_3}{R_1} \le A_v \le -\left(\frac{R_2+R_3}{R_1} - \frac{R_2R_3}{R_{on}R_1}\right)$$

where R_{on} is the practical on-resistance of the transistor.

This entire circuit can be configured using one quad op amp (such as an XR 4136) and requires no special considerations other than attention to the basic rules of grounding and supply bypassing. □



Hands off. If the gain is clamped to a minimum at input signal levels below the noise threshold, the surrounding noise is filtered out of the amplifier network (a), whereas speech kicks in the amplifier's automatic gain control. Both functions are performed by the circuit shown in (b).

—ETI project 446

AUDIO LIMITER

This simple but effective unit can be used as a limiter, automatic volume control or voltage controlled amplifier.

THE AUDIO COMPRESSOR EXPANDER project described in the May 1976 issue of ETI has proved to be very popular with readers and we have since had many requests for a simpler limiter circuit. Whilst limiters and compressors are similar in operation they are used in completely different ways.

A compressor is normally used in a linear compression mode. That is, for say every 10 dB of input signal level change the output is arranged to change by, for example, 6 dB. The output will change this fixed amount of 6 dB for every 10 dB increment of input. The reverse of this procedure is called expansion. That is, for a 6 dB change in input signal level the output is caused to change by 10 dB.

A compressor/expander is typically used for improving the dynamic range (and hence signal-to-noise ratio) of tape recorders. The signal is first compressed so that its dynamic range can be handled by the tape. On subsequent replay the signal is expanded by a corresponding amount to restore the original dynamic range. As the amount of noise on the tape is constant and the level of signal has been effectively increased, the signal-to-noise ratio has also been increased.

A limiter is a form of compressor which operates only when the signal exceeds a certain predetermined level. For example signals which do not exceed say 80% of the predetermined maximum are not compressed at all and are amplified with their full dynamic range. For signals above the 80% level the limiter begins to operate and very large input signals are required to obtain the extra 20% of output.

Another use of a limiter is in the continuous-limit mode such that it acts as an automatic volume control (AVC). In this mode a 60 dB change in input level can be limited to say, a 6 dB change in output level.

Finally the limiter may also be used as a voltage controlled amplifier having a range of about 55 dB. A typical application of such a device would be a remote volume control. It should be noted, however, that although the transfer function of such a voltage controlled amplifier is fairly sharp, two of them may not necessarily track perfectly due to differences in the FETs in the ICs. Thus on our prototype the difference between channels when used as a stereo volume control was up to 5 dB at some points with any given input.

DESIGN FEATURES

The first decision to be made when designing a limiter is what type of controlled resistive element to use. Common alternatives are FETs, LDRs, base-emitter junctions of transistors, thermistor or balanced modulator ICs. All of these have their respective advantages and disadvantages and all have been tried in our laboratory at one time or another. We selected FETs because we considered them the most cost effective.

When FETs are used in voltage controlled amplifiers it is essential that the voltage across them is kept as low as possible if the distortion is also to be kept low. This means that the FET must be used as an attenuator where the voltage across the FET can be kept low irrespective of input voltage. The most suitable type of FET for this purpose is the enhancement-mode device but these are not readily available. The commonly available types require a negative voltage to turn them off. However, there is a suitable alternative, the 4049 CMOS IC which contains six inverting buffers. By suitable interconnection the IC may be made to provide six enhancement-mode FETs and this is the approach we decided to use.

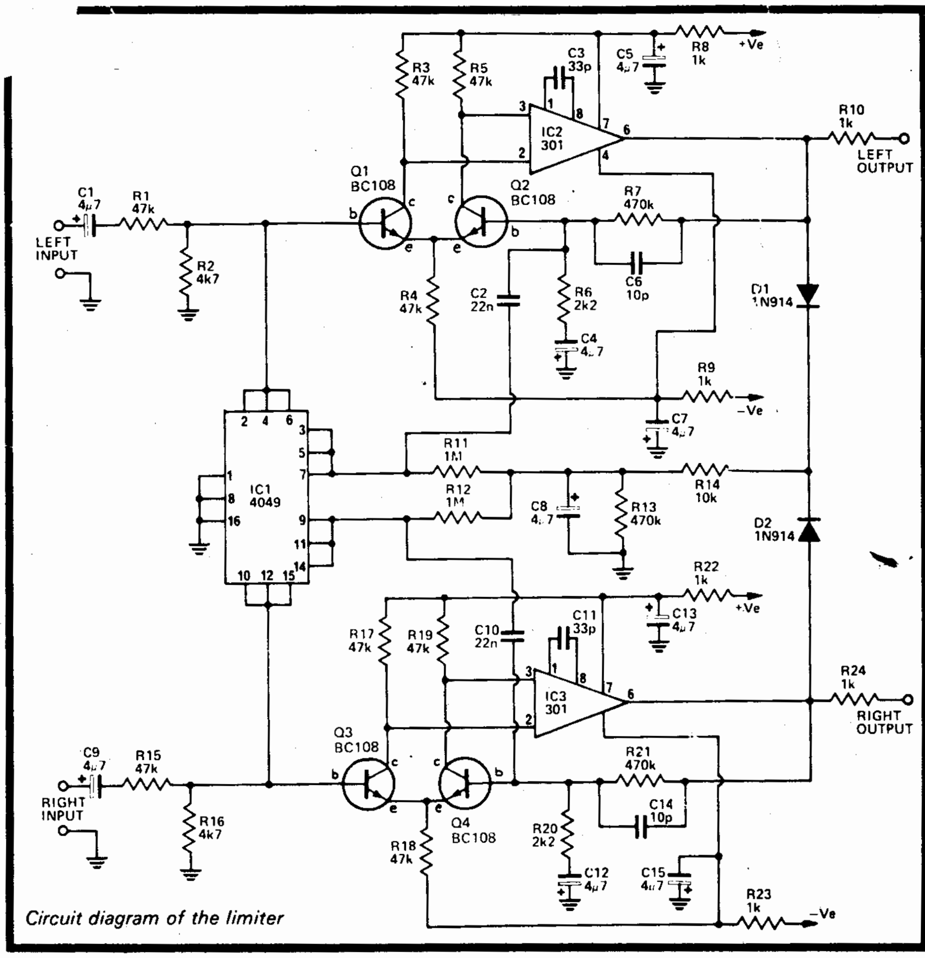
To restore the signal level an amplifier is required and originally we intended to use the LM382 but, because of cost and availability considerations, we finally decided to use an LM301 or 741 operational amplifier together with a transistor pair at the front end. The noise performance of this arrangement was found to be as good as the LM382's and supply voltage to be less critical (although a dual supply is required). If only a single-ended supply is available then a 382 may be used, although a different board layout would be required.

CONSTRUCTION

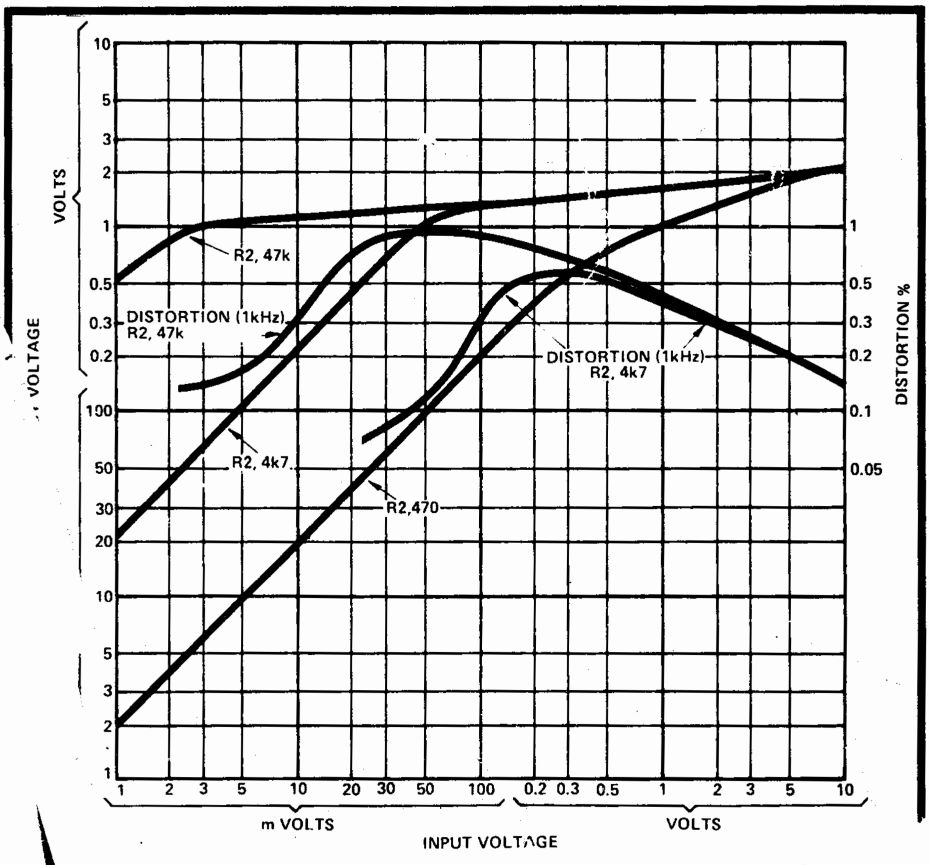
Although a printed-circuit board is not essential it certainly makes construction very much easier. Before assembly decide whether a limiter or an AVC is required as the

Specification ETI 446

Input voltage range	**1 mV – 10 V**
Frequency response	**± 3 dB 10 Hz – 20 kHz**
Limiting point	
set by R2/16	**3mV**
Equivalent signal-to-noise ratio	**70 dB re 1 V out**
Distortion	**see graph**
Input impedance	**47 k**
Maximum gain	
R2/16 = 4k7	**26 dB**
R2/16 = 47k	**40 dB**
Maximum attenuation	
as voltage controlled amplifier	**55 dB**
Supply voltage	**± 8 V to ± 16 V dc at 5 mA**



Circuit diagram of the limiter



Input versus output voltage for various values of R2 (and R16) Distortion at 1kHz for R2=4K7 and R2=47K are also shown.

values of R2 and R16 will vary accordingly. Use 47k for R2 and R16 in the AVC mode and in limit mode, depending on limit point, between 470 and 4k7. The transistor type specified is available from a number of different manufacturers but pin connections are different. If a different brand is used the transistor should be reversed (emitter and collector interchanged). The overlay also shows the arrangement for using the LM301 ICs — these may be directly replaced by 741s simply by omitting the 33 pF capacitors.

Although the CMOS ICs 4449 and 4009 are electrically similar to the 4049 and are interchangeable with it when the devices are used as hex-inverters, they cannot be used as replacements in this circuit. The 4049 must be used. The 4449 and 4009 have different circuitry and will not work in this mode.

How it works

The circuit basically consists of a voltage-controlled attenuator followed by a low-noise amplifier with a gain of 46 dB. The output of this amplifier is rectified to generate a dc voltage which is used to control the attenuator.

The variable element in the attenuator is an enhancement mode FET. This is made from a CMOS hex-inverter IC, the 4049, by special interconnection. The difference between enhancement mode FETs and the normally available depletion-mode junction FETs is as follows: The enhancement mode FET has a high resistance between source and drain when the gate is at zero volts, but this decreases as the gate is taken more positive. A JFET (N type) is hard-on with the gate at zero volts and turns off as the voltage is taken negative.

The amplifier is required to have high open-loop gain and have fairly low noise. The gain requirement is provided by an LM301 operational amplifier and the low-noise requirement by a pair of transistors (connected as a differential pair) placed before the operational amplifier. The gain is set, by the combination of resistors R6 and R7, to 215 (or 46 dB). The lower 3 dB point is set at 15 Hz by C4 and R6 whilst the upper 3 dB point is set at 33 kHz by C6 and R7.

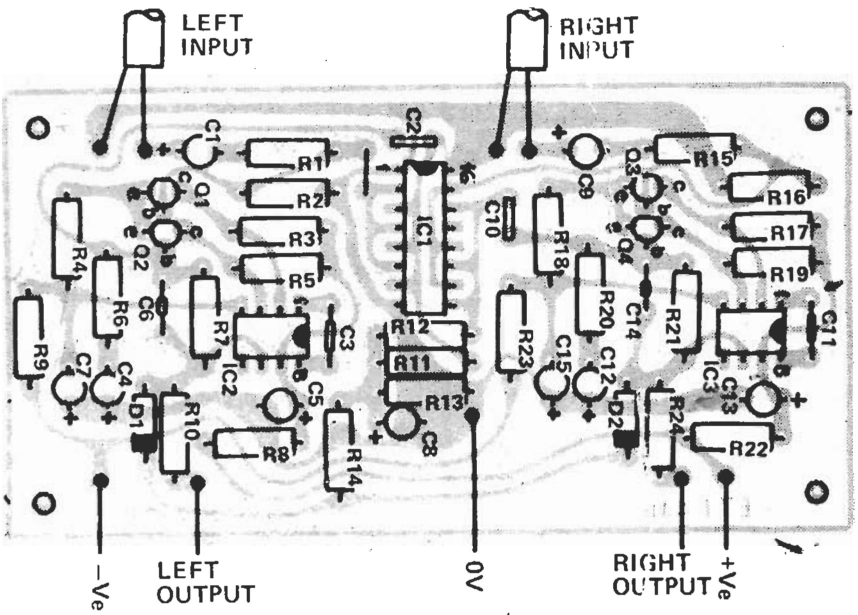
The outputs of both channels are summed and rectified by diodes D1 and D2 to charge C8 via R14. The voltage on C8 is coupled to the gate of the FETs (three in parallel on each channel) via R11 and R12.

As the input voltage increases the output also tends to increase and voltage on capacitor C8 also increases and this increase is applied back to the gates of the FETs. This reduces the resistance of the FETs and thus increases the attenuation, tending to prevent the output from changing as much as the input does.

With all FETs the resistance changes with applied voltage and this gives rise to distortion. However by modulating the gate voltage with a signal equivalent to the voltage across the FETs the distortion is greatly reduced (3.5% down to 0.8%).

The attack and release times can be adjusted by varying R14 for attack and R13 for release.

AUDIO LIMITER



Component overlay.

Parts List

Resistors

R1	47k	½ W	5%
R2	4k7	"	"
R3-R5	47k	"	"
R6	2k2	"	"
R7	470k	"	"
R8-R10	1k	"	"
R11,12	1M	"	"
R13	470k	"	"
R14	10k	"	"
R15	47k	"	"
R16	4k7	"	"
R17-R19	47k	"	"
R20	2k2	"	"
R21	470k	"	"
R22-R24	1k	"	"

Capacitors

C1	4μ7 25 V electrolytic
C2	22n polyester
C3	33p ceramic
C4,5	4μ7 25 V electrolytic
C6	10p ceramic
C7-C9	4μ7 25 V electrolytic
C10	22n polyester
C11	33p ceramic
C12,13	4μ7 25 V electrolytic
C14	10p ceramic
C15	4μ7 25 V electrolytic

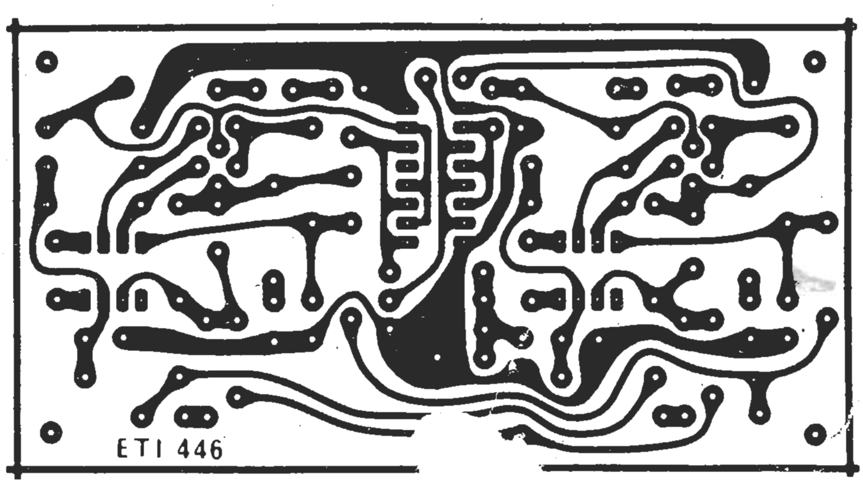
Semiconductors

Q1-Q4 Transistors BC108
D1,2 Diode 1N914
IC1 Integrated circuit 4049 *
IC2,3 " " LM301

Miscellaneous

PC board ETI 446
9 PC board pins

**Do NOT substitute a 4009 or 4449 as the input protection is different.*



Printed-Circuit layout for the limiter. Fu... ...m.

As this unit will normally be used in association with another piece of equipment, and most likely built in to it, a case has not been described. When installing the unit make sure that the input cables are coaxial or shielded cable — outputs are not important and can be normal hookup wire.

USES OF A LIMITER

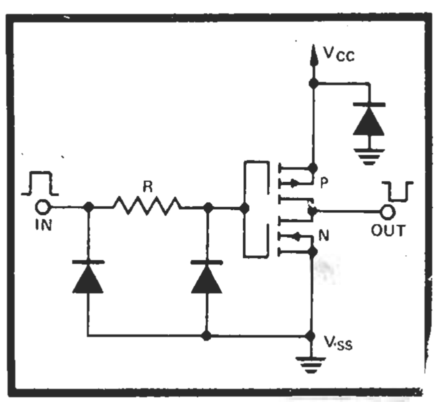
Peak Limiting. In this mode only signals above 85% of maximum level are attenuated. This is useful for preventing amplifier clipping (for pop groups or other live shows) which gives rise to objectionable distortion. It may also be used when tape recording the same type of programme material as above, to prevent the tape being saturated, which again would give rise to distortion.

AVC. In this mode, the limiter is used typically to drastically reduce the dynamic range of a programme being recorded. For example, when recording a lecture the 60dB dynamic range of lecture room speech may be compressed to 6dB.

Voltage Controlled Amplifier. As a voltage-controlled amplifier the unit lends itself to a variety of remote or automatic control applications. For example, it may be used as a remote control for stereo amplifier volume. Alternatively, it may be adjusted to increase car radio volume as ambient noise level rises.

Special Effects. The limiter may also be used to modify the sound of musical instruments. For example, such a limiter is often used to eliminate the attack transient on a bass guitar to give a smoother mellower sound.

The uses of such a circuit are wide indeed, and we are sure our readers will think of many more applications for this interesting circuit. ●



Internal circuit diagram of one of the six inverter stages in the CMOS 4049 IC

Automatic gain control has 60-decibel range

by Neil Heckt
The Boeing Co., Seattle, Wash.

An automatic-gain-control circuit with an input range of 60 decibels (20 millivolts to 20 volts) can be built using a junction FET as a voltage-controlled resistor in a peak-detecting control loop. The circuit exhibits a quick response of 1 to 2 milliseconds and a delay time of 0.4 second.

As shown in the schematic of Fig. 1, the 2N4861 n-channel field-effect transistor Q_2, connecting the noninverting input of the operational amplifier to ground, determines the closed-loop gain of the system. Negative base-voltage peaks from the output of the op amp, beyond V_{BE} of Q_1, turns Q_1 on, and its collector current then charges capacitor C_1.

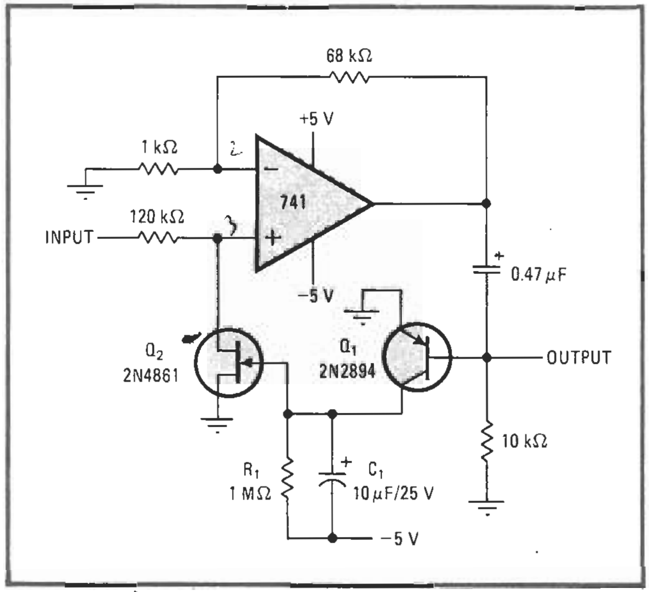
The voltage across C_1 determines the channel resistance of Q_2. Since the range of this resistance is 120 ohms to more than 10^8 ohms, the 60-dB range of the circuit is easily realized.

In the absence of an input signal, capacitor C_1 discharges through resistor R_4, cutting off Q_2. It is the C_1-R_4 combination that determines the delay time of the circuit. The collector current of Q_1 and the value of C_1 determine the circuit's attack time.

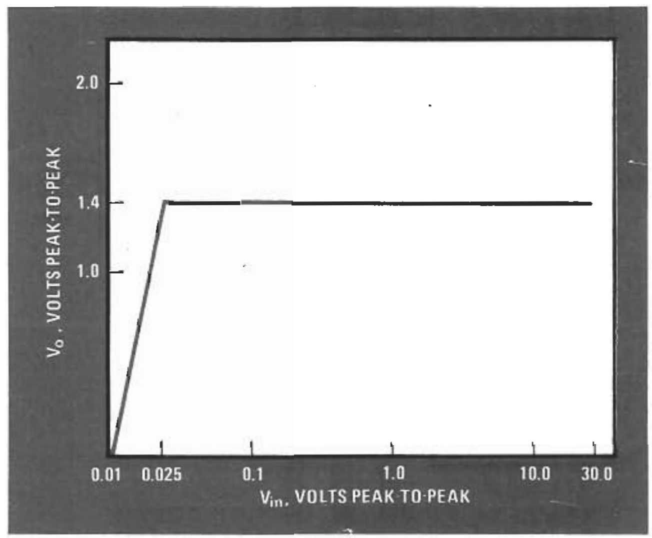
The op amp can be a 741 or any general-purpose device. Even with input signals of 20 V peak to peak, the maximum signal at the device's input is 25 mV peak to peak. Thus it is possible for the input voltage to be greater than the supply voltage.

The op amp's output is ac-coupled to the base of Q_1 because the dc operating point of its output varies with the changing output impedance of Q_2. To avoid dc bias difficulties when coupling to subsequent stages, the circuit's output is taken from the base of Q_1. Figure 2 shows the gain-control behavior of this circuit throughout its dynamic range. ☐

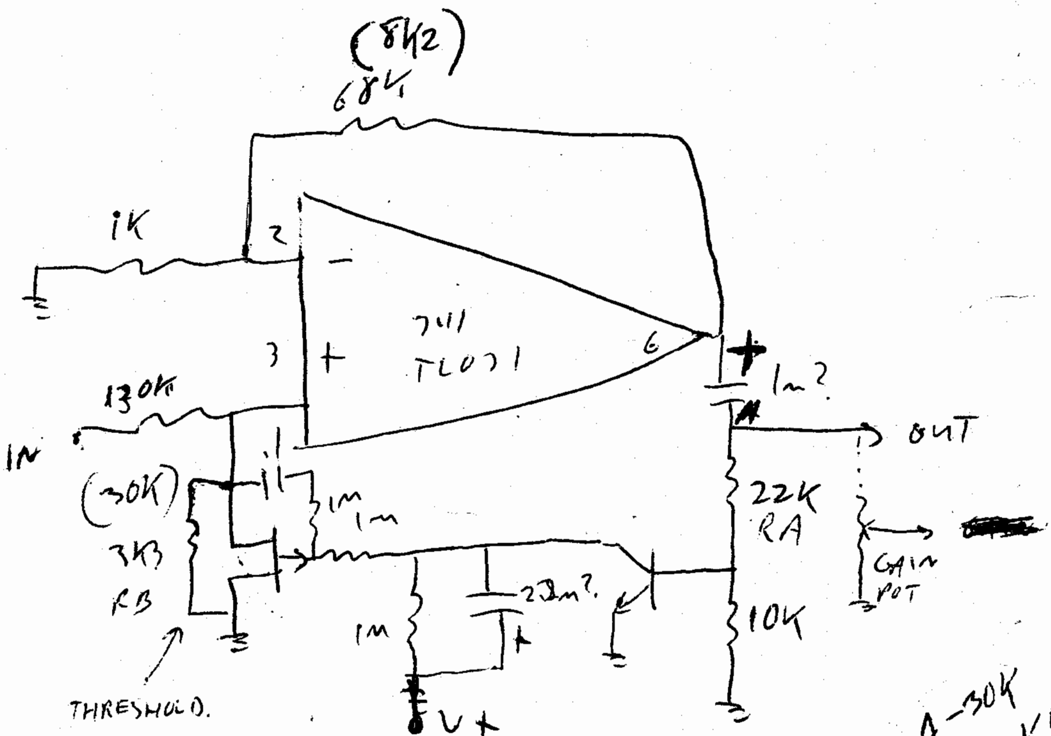
Designer's casebook is a regular feature in *Electronics*. We invite readers to submit original and unpublished circuit ideas and solutions to design problems. Explain briefly but thoroughly the circuit's operating principle and purpose. We'll pay $50 for each item published.



1. Audio AGC: Input voltage for distortionless output is from 20 millivolts to 20 volts in this quick-response AGC circuit. The input signal can have greater magnitude than the supply voltage because the maximum signal across the FET is 25 millivolts.



2. AGC characteristics. The output voltage is approximately 1.4 volts over a 60-dB range. A wider dynamic range would be possible if the off/on-resistance ratio of the FET were greater.



RA-30K
RB-3K9

33K

RA
o/p 47K THRESHOLD = 0VU ~ .775V
V 1.75V o/p= 2.3V OUT = 1.3VRMS
1.4V

V 1.75V 4.4V

/ " .44V

V " .45V

4V " 1.2V

± 15V supply