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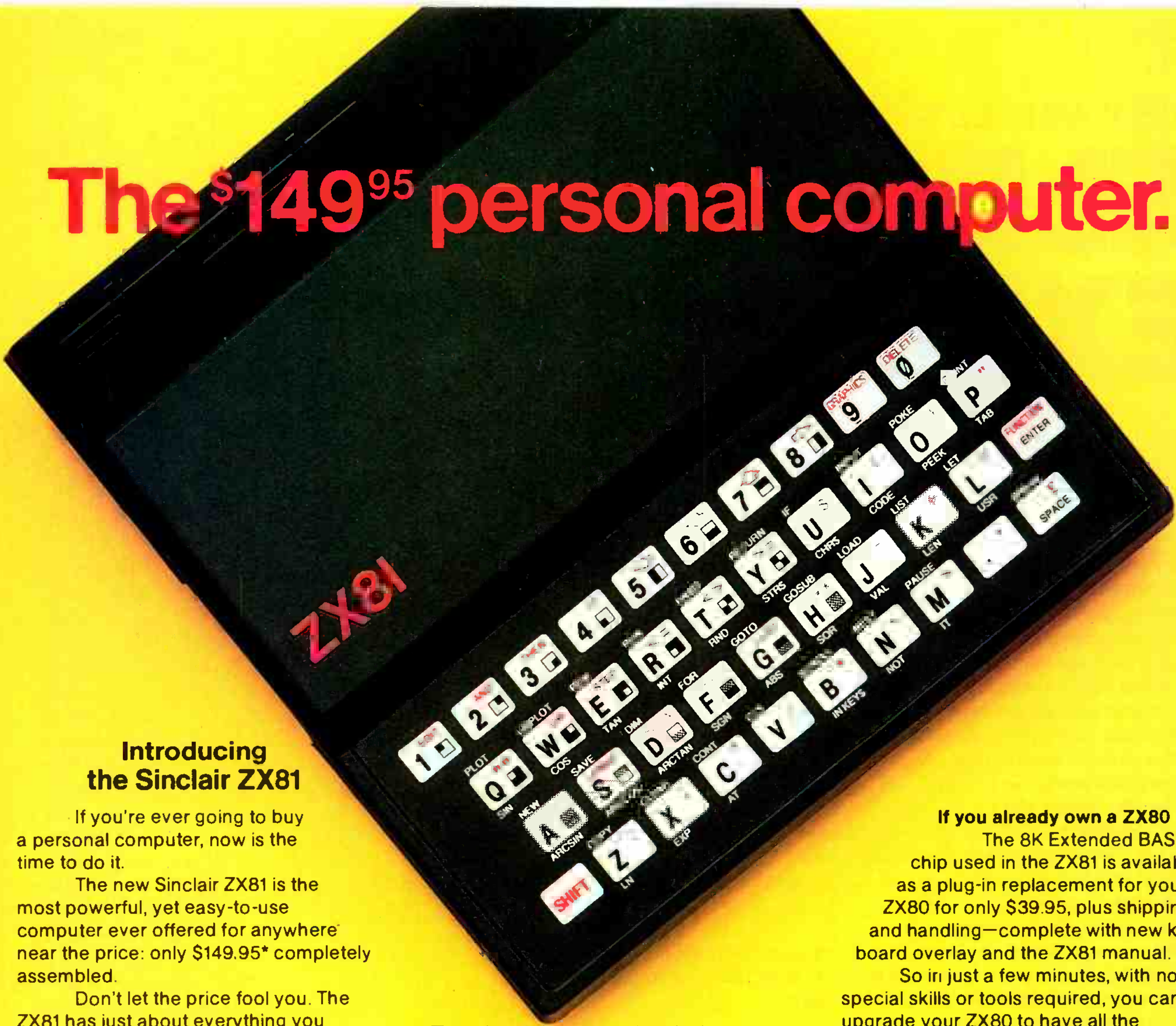

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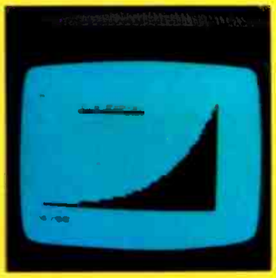

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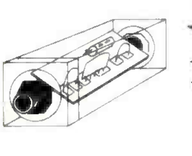

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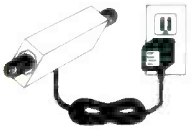

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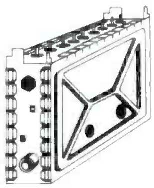

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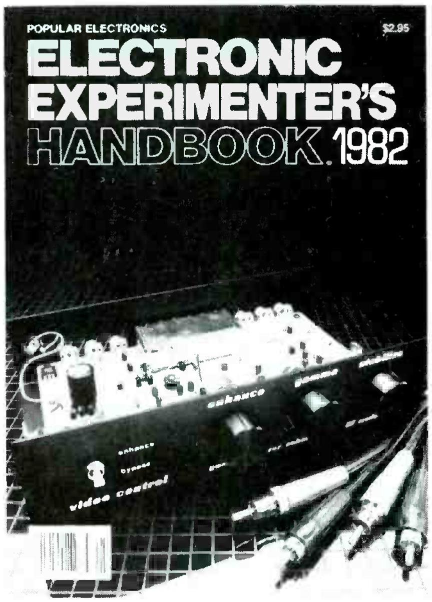

THE COVER

The Video Enhancer project featured on the cover is designed to allow you to make copies of programs on videocassettes that are virtually indistinguishable in quality from the original program. As a bonus, the project also includes a copy-guard stabilizer that permits TV receivers that lack external horizontal hold controls to display steady pictures from copy-guarded videocassettes, without the usual rolling and/or tearing of the picture. For complete information on how to build this inexpensive project, refer to the construction article that begins on page 5.

COVER PHOTO: Jay Brenner

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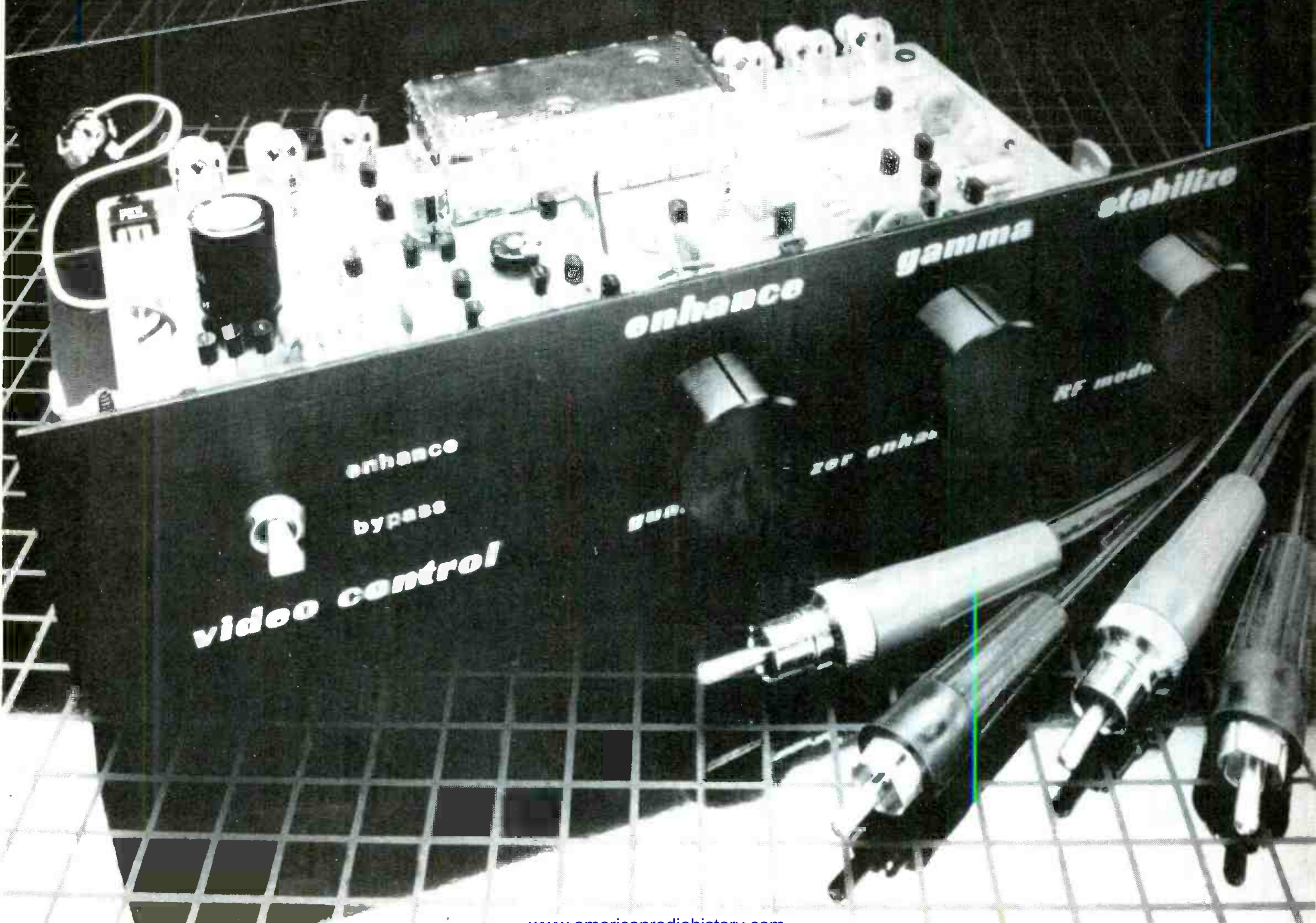
BUILD THIS

VIDEO ENHANCER WITH COPY GUARD STABILIZER

BY ROGER COTA

Inexpensive video control unit eliminates troublesome copy guard and recovers picture detail lost through videocassette recording



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Owners of videocassette tape machines soon realize that there are some problems to overcome. One is the expense of video tape, which motivates the user to record at a slower speed. This, however, degrades picture quality. Another consideration is that many prerecorded movies, concerts, and special programs available for sale or rent are "copy guarded." Accordingly, some television receivers will not play these tapes well because the guard signal makes the picture roll, jitter, or disappear altogether.

To overcome these challenges, here is a low-cost professional unit that will allow you to record video at slower speeds or copy any tape with improved picture quality. The unit also provides a distribution power amplifier for driving more than one tape machine and permits use of an r-f modulator for real-time enhancement while viewing.

Copy Guarding. Video is made up of two components: sync pulses and picture information. Sync pulses are as important as picture information because they format the picture on the screen. Television is made up of fields of pictures traced on the screen of a picture tube by an electron beam. A vertical oscillator controls the picture tracing from top to bottom of the screen. Every 1/60 second, vertical sync pulses in the video signal reset the oscillator, which starts the trace at the top of the screen again. If the vertical sync pulses were missing, the picture would appear to roll uncontrollably.

Tracing action of the beam for one field is illustrated in Fig. 1. Normal vertical sync pulses in the video signal are illustrated in Fig. 2. These pulses are stripped out of the video signal by circuitry in the VCR or TV receiver and then integrated to create a ramping voltage. When this voltage reaches a set threshold, the vertical oscillator driving the picture tube is reset, starting the beam at the upper left of the picture tube screen.

The path of the vertical sync pulses is shown in Fig. 3. Most TV receivers have the designation "vertical hold" for the threshold control, accessible as either a front- or rear-panel control. Some TV receivers and especially videocassette recorders have automatic or fixed thresholds.

When vertical sync pulses are altered, the picture will roll and, therefore, be unviewable. Most manufacturers of prerecorded video tapes, especially of motion pictures, are processing the vertical sync pulses to prevent buyers and renters from copying the tapes. The guard process, however, alters the width of the vertical sync pulses, making them narrower than normal, as shown in Fig. 2B. When integrated, these sync pulses will not produce enough voltage to reach the fixed threshold of the vertical timing circuit in VCRs, "confusing" the VCR's circuitry and preventing recording. The original tape can be viewed normally on TV receivers equipped with vertical-hold controls because vertical hold can be adjusted to compensate for the guard. A problem occurs, though, when a tape is viewed on a TV receiver that has no vertical-hold control. For these receivers, this outboard guard-defeating circuit is needed.

Picture Enhancing. The picture portion of the video signal carries the visual scenes that are actually viewed. The picture is made up of a luminance signal (the black-and-white portion) and a chrominance signal (the color portion). Picture clarity and detail or sharpness is carried by the luminance signal, while color and tone are added by the chrominance signal. As in audio, the luminance signal has a frequency range, though a much wider one. As shown in Fig. 4, the standard luminance signal's bandwidth ranges from dc to approximately 4 MHz, whereas audio ranges from dc to 20 kHz. The highest frequencies of 2 to 4 MHz correspond to the smallest details in the picture. Without these high frequencies, the picture appears fuzzy-soft and fine detail is lost.

High frequencies are lost due to the

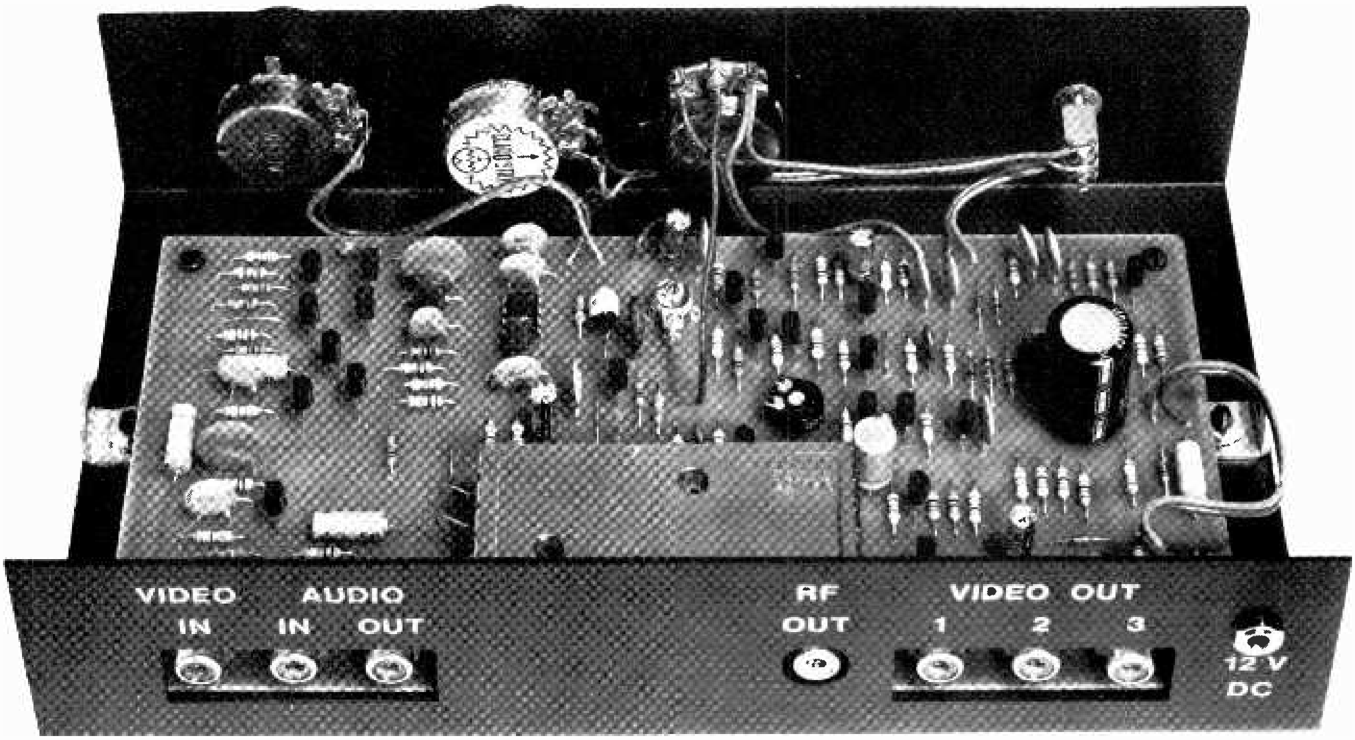

Photo shows interior construction details of the Video Enhancer.

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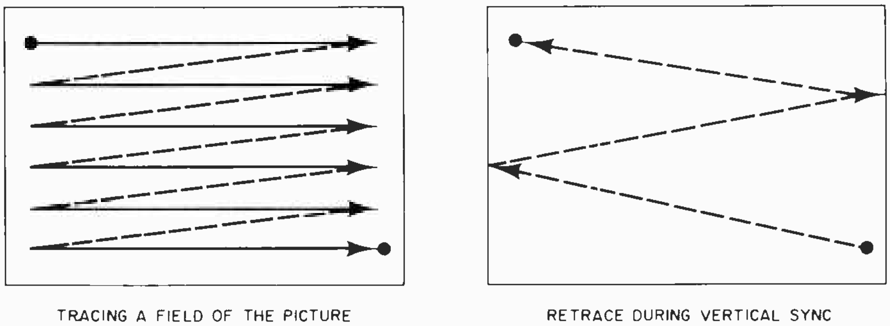
Fig. 1. Shown here is the tracing action of the electron beam on the face of the picture tube during field and retrace.



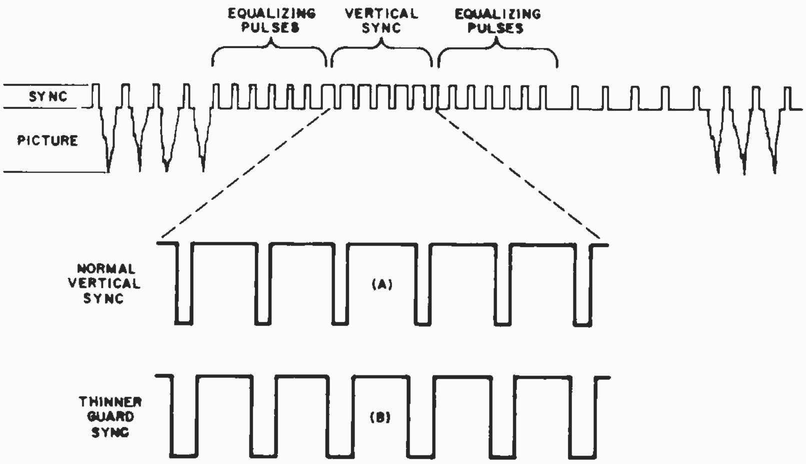
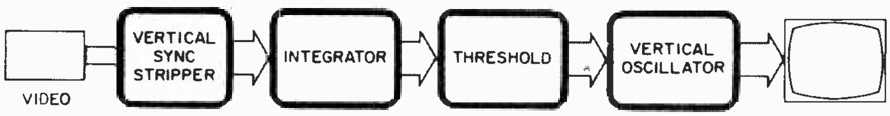


Fig. 2. Top trace shows standard sync pulses and picture signals. Contrast between normal vertical sync and thinner guard sync are illustrated in (A) and (B).

Fig. 3. Path of vertical sync pulses is illustrated in this block diagram.



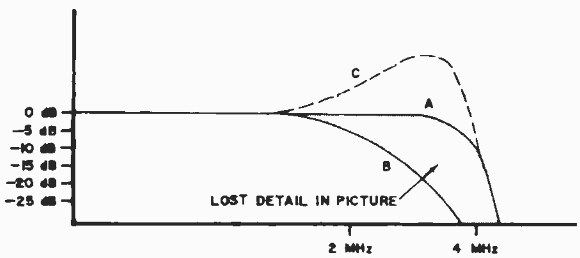
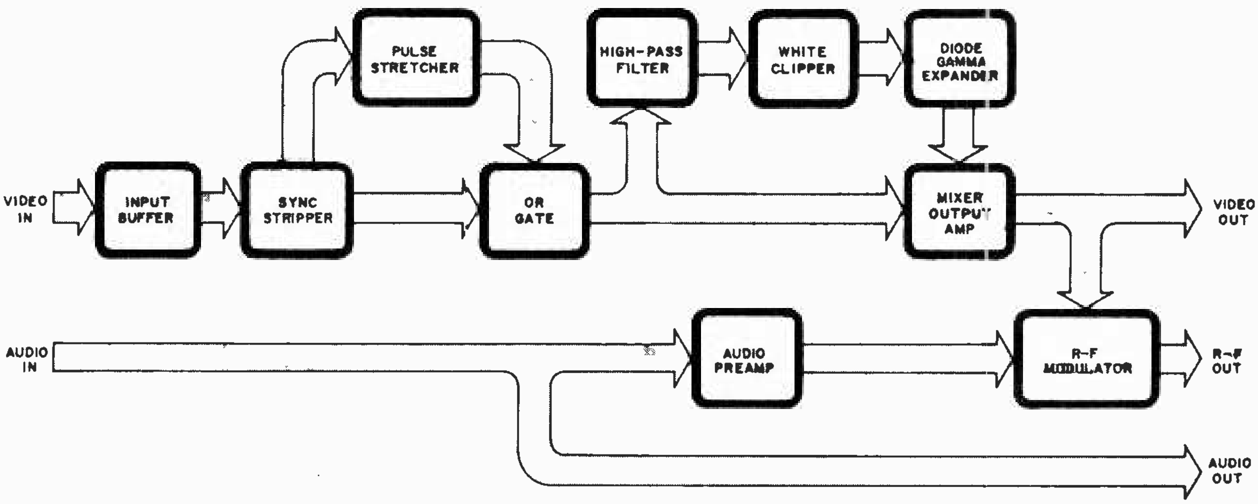


Fig. 4. Plots show (A) frequency bandwidth of standard luminance, (B) bandwidth after recording on tape, and (C) boost in frequency by enhancer.

Fig. 5. This is the block diagram of the enhancer/stabilizer project.



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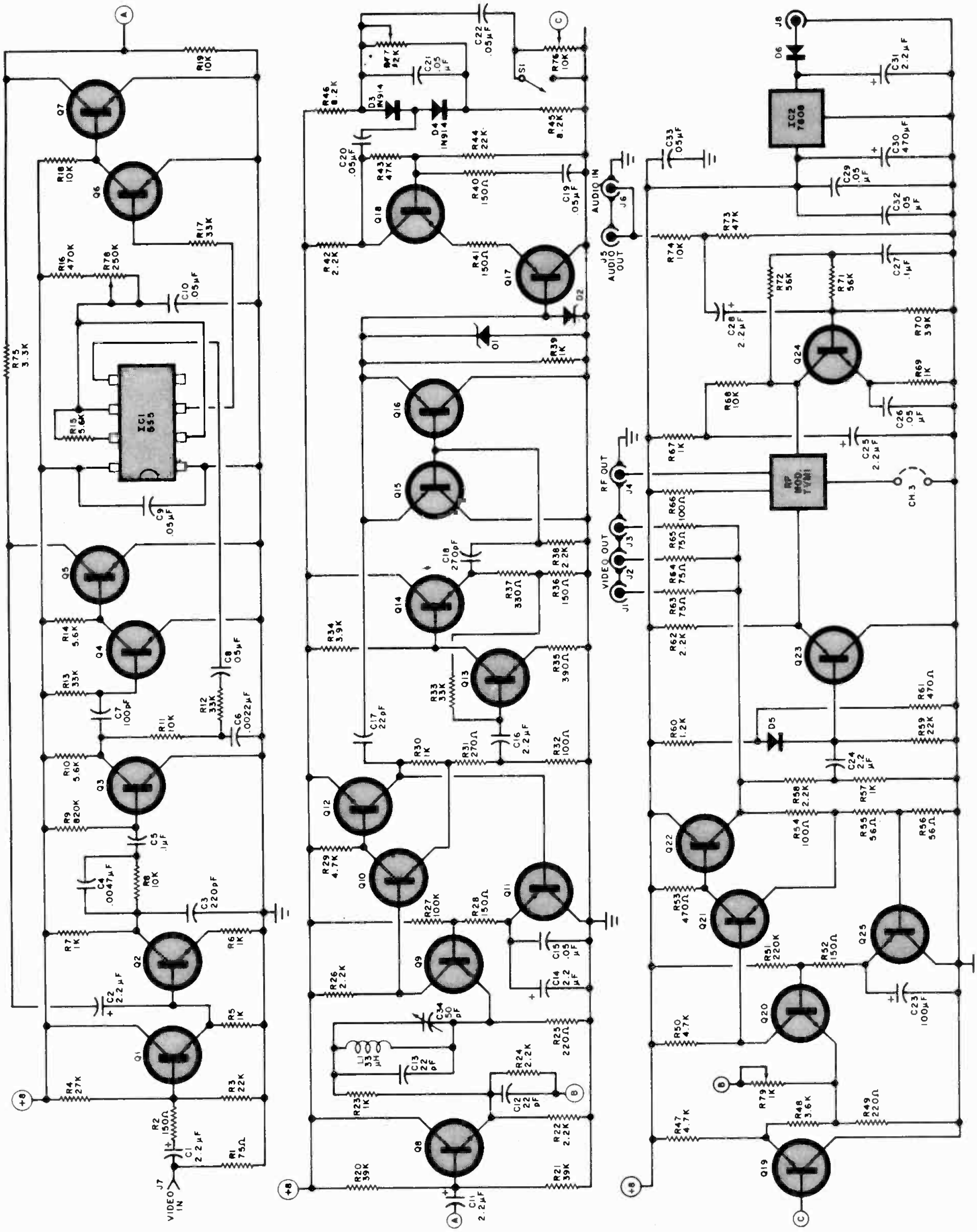

PARTS LIST

C1,C2,C11,C14,C16,C24,C25,C28,C31—2.2-μF, 50-volt electrolytic capacitor
C3—220-pF, 50-volt disc capacitor
C4—0.0047-μF, 100-volt Mylar capacitor
C5,C27—0.1-μF, disc capacitor
C6—0.0022-μF Mylar capacitor
C7—100-pF disc capacitor
C8,C9,C10,C15,C19,C20,C21,C22,C26, C29,C32,C33—0.05-μF disc capacitor
C12,C13,C17—22-pF disc capacitor
C18—270-pF disc capacitor
C23—100-μF, 10-volt electrolytic capacitor

Fig. 6 Schematic diagram of Video Enhancer

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C30—470-μF, 35-volt electrolytic capacitor
C34—5-55-pF trimmer capacitor
D1,D2—1N270 zener diode
D3,D4,D5—1N914 diode
D6—1N4001 rectifier diode
IC1—555 timer IC
IC2—7808 8-volt regulator IC
J1 thru J7—Phono jack
J8—Miniature phone jack
L1—33-uH high-Q inductor coil
Q1 thru Q10,Q13,Q14,Q16,Q18,-Q20,Q21,Q24—Sylvania ECG287 npn transistor
Q11,Q12,Q15,Q17,Q19,Q22,Q23,Q25—Sylvania ECG288 pnp transistor
All resistors 1/4 watt, 5% tolerance:
R1,R63,R64,R65—75 ohms
R2,R28,R36,R40,R41,R52—150 ohms
R3,R44,R59—22,000 ohms
R4—27,000 ohms
R5,R6,R7,R23,R30,R39,R57,R67,R69—1000 ohms
R8,R11,R18,R19,R68,R74—10,000 ohms
R9—820,000 ohms
R10,R14,R15—5600 ohms
R12,R13,R17,R33—33,000 ohms
R16—470,000 ohms
R20,R21,R70—38,000 ohms
R22,R24,R26,R38,R42,R58,R62—2200 ohms
R25,R49—220 ohms
R27—100,000 ohms
R29—4700 ohms
R31—270 ohms
R32,R54,R66—100 ohms
R34—3900 ohms
R35—390 ohms
R37—330 ohms
R43,R73—47,000 ohms
R45,R46—8200 ohms
R47,R50—4700 ohms
R48—3600 ohms
R51—220,000 ohms
R53,R61—470 ohms
R55,R56—56 ohms
R60—1200 ohms
R71,R72—56,000 ohms
R75—3300 ohms
The following are linear-taper potentiometers:
R76—10,000 ohms
R77—2000 ohms
R78—250,000 ohms
R79—1000-ohm trimmer potentiometer
S1—Spdt switch
TVM1—TV r-f modulator and antenna switch (Radio Shack kit Cat. No. 277-122)
Misc.—117-volt ac to 12-volt dc, 300 mA power adaptor; printed-circuit board; control knobs; line cord; aluminum or steel cabinet; machine hardware; hook-up wire; solder; etc.

Note: The following are available from Video Control, 3314 H Street, Vancouver, WA 98663 (tel. 1-206-693-3834): Complete kit containing pc board, power adapter, case, and all parts but excluding r-f modulator/antenna switch for $110; etched and drilled pc board for $15; power adapter for $10. Please add $3.50 for postage and handling.

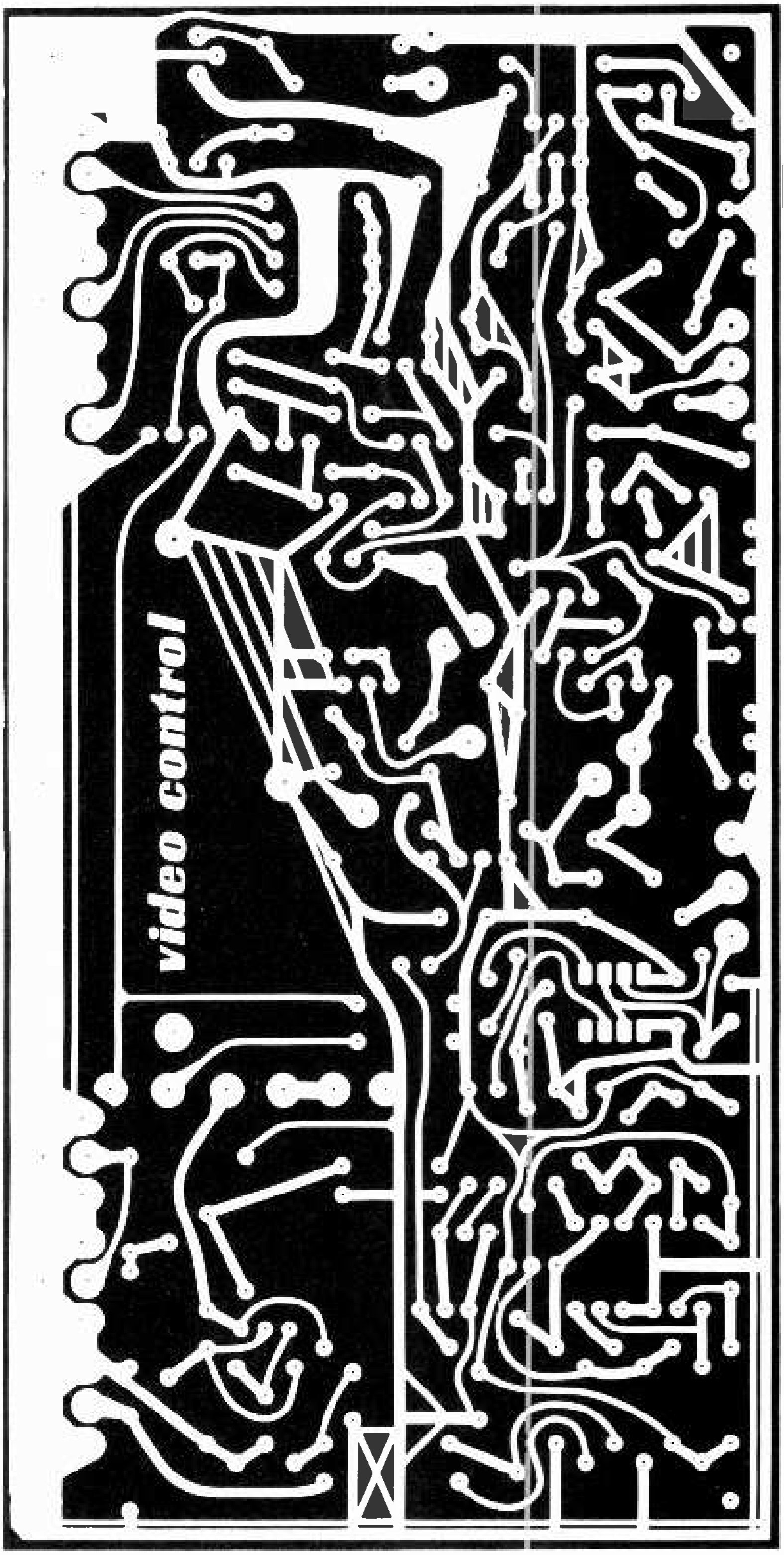

Fig. 7A. Actual-size etching and drilling guide for enhancer project.

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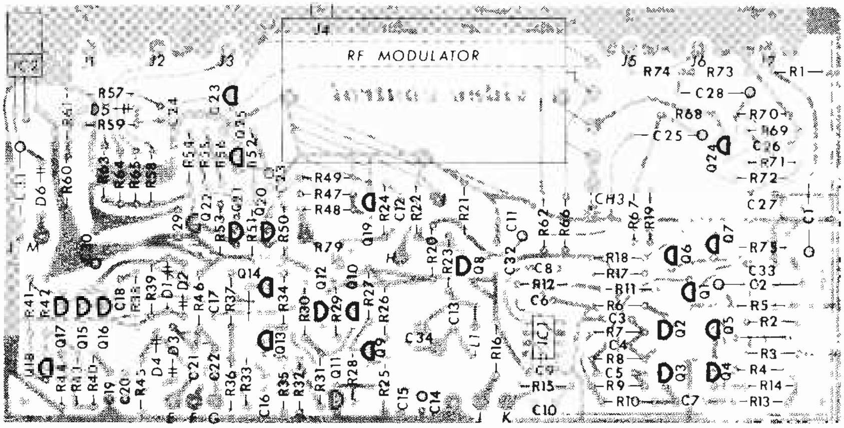

Fig. 7B. Components-placement guide for enhancer project.

Notes: R77 CONNECTS BETWEEN HOLES *E* AND *F*
R76 CONNECTS BETWEEN HOLES *G* AND *I*, WIPER TO HOLE *H*
S1 CONNECTS ACROSS R76 (HOLE *G* TO HOLE *I*)

recording process and limitations of the recording tape (Fig. 4). Every recording (generation) from the original causes more loss in detail. The picture-enhancing portion of this project reequalizes the luminance bandwidth by boosting the high frequencies. When this is done prior to recording, the loss caused by the tape and machine can be canceled out, giving a copy that has as much detail as the original.

Enhancing high-frequency components of the video signal may increase noise, appearing as snow in the picture. This noise is reduced by a logarithmic gamma circuit that acts like an amplitude expander. When properly adjusted, low-level noise is eliminated by the logarithmic gamma circuit.

About The Circuit. This project has three controls. Adjustment of the ENHANCE control increases detail and edge sharpness. Proper adjustment of the GAMMA control complements the enhance adjustment by reducing snow and other low-level luminance noise. The STABILIZE control locks in the picture and cancels the copy-guard signal. (A block diagram of the enhancer stabilizer circuit is shown in Fig. 5.)

As shown in Fig. 6, *Q1* acts as a buffer for video inputs. Transistors *Q2* and *Q3*, capacitors *C3*, *C4*, and *C5*, and resistor *R8* separate sync pulses from the video. Sensing of vertical sync and triggering of *IC1* is accomplished with *C6* and *R11*, while *Q4* and *Q5* clamp the video to ground. The width of the sync pulse is set by *C10*, *R16* and *R78*.

The output of *IC1* drives *Q6* and *Q7*, which mix the new vertical sync pulses in with the video. At this point, any guard signal is eliminated. Buffer *Q8* drives a chroma filter madeves up of *R23*, *C13*, *C34*, and *L1*, which reduces any color shift that may occur as a result of over-enhancing. High-pass filter *C17/R39* is driven by *Q9*, *Q10*, *Q11*, and *Q12*. Clamping transistors *Q15* and *Q16* are driven by inverter *Q13/Q14*. Diodes *D1* and *D2* clamp any signal overshoot, while transistors *Q17* and *Q18* make up a cascode amplifier that drives gamma circuit *D3/D4*. The diodes operate as a nonlinear signal expander whose threshold is controlled by the setting of *R77*.

The gamma circuit reduces any noise introduced by enhancing action, by an amount set by *R76*. Switch *S1* inserts and defeats enhancement. Buffer *Q19* delivers the signal to the output mixer, while *R79* mixes in the original video. The output mixer amplifier is made up of *Q20*, *Q21*, *Q22*, and *Q25*. The video is prepared for r-f modulator *TVM1* by *R60*, *R61*, *D5*, and *Q23*. A modulator designed for reception on a standard TV receiver on Channel 2 or 3 must be used. A typical example of such a modulator is Radio Shack's Catalog No. 277-122, which includes an antenna isolation switch for attachment at the TV receiver's antenna terminals.

Audio preamplifier *Q24* preemphasizes high frequencies for the r-f modulator. System power is regulated at 8 volts by *IC2*, while input power requirements are 12 volts dc at 300 mA, obtained from a standard battery eliminator/charger.

Construction. A printed circuit board is imperative for this project, due to the high-frequency requirements of low stray capacitance. An actual-size etching and drilling guide and a components-placement diagram are shown in Fig. 7.

Proper orientation of parts during assembly is very important. So, take careful note of the directions of transistors, diodes, and electrolytic capacitors. The plus (+) lead holes for the electrolytic capacitors are identified by circles on the board. Since high frequencies are involved, it is a good idea to keep all component leads as short as possible. And, when soldering in the r-f modulator, make certain that the ground pins are fully coated with solder and firmly attached to the copper traces.

Once the project is assembled, install jumper *CH3* if you plan to use the device on TV Channel 3; otherwise, the modulator will transmit on Channel 4. Also, after soldering components to the board traces, clean away the flux residue with alcohol and follow up with a careful inspection to make sure that there are no solder bridges between closely spaced traces.

The project is designed to fit into a custom aluminum case to insure low r-f radiation. Before placing the top on the case, however, connect the project to the VIDEO OUT jack of a VCR and the device's r-f modulator output to a TV receiver's antenna isolation switch. With the enhancer defeated, play a tape and adjust *R79* so that the picture on the screen is as bright as a regular TV program's. Engage the enhancer at full enhancement and adjust *C34* so that the picture is enhanced without altering the color. This done, assemble the case.

Summing Up. The enhancer/stabilizer is an excellent tool for making copies as good as the original and for viewing older video tapes. Furthermore, it will save money spent on tapes by giving comparable viewing quality of the 2-hour mode when recording in the 6-hour mode.

This project is not intended to be used for illegal copying, of course. It is intended solely to correct problems arising when a copy-guarded tape is played on a TV receiver that has a limited-range vertical-hold control. ◇

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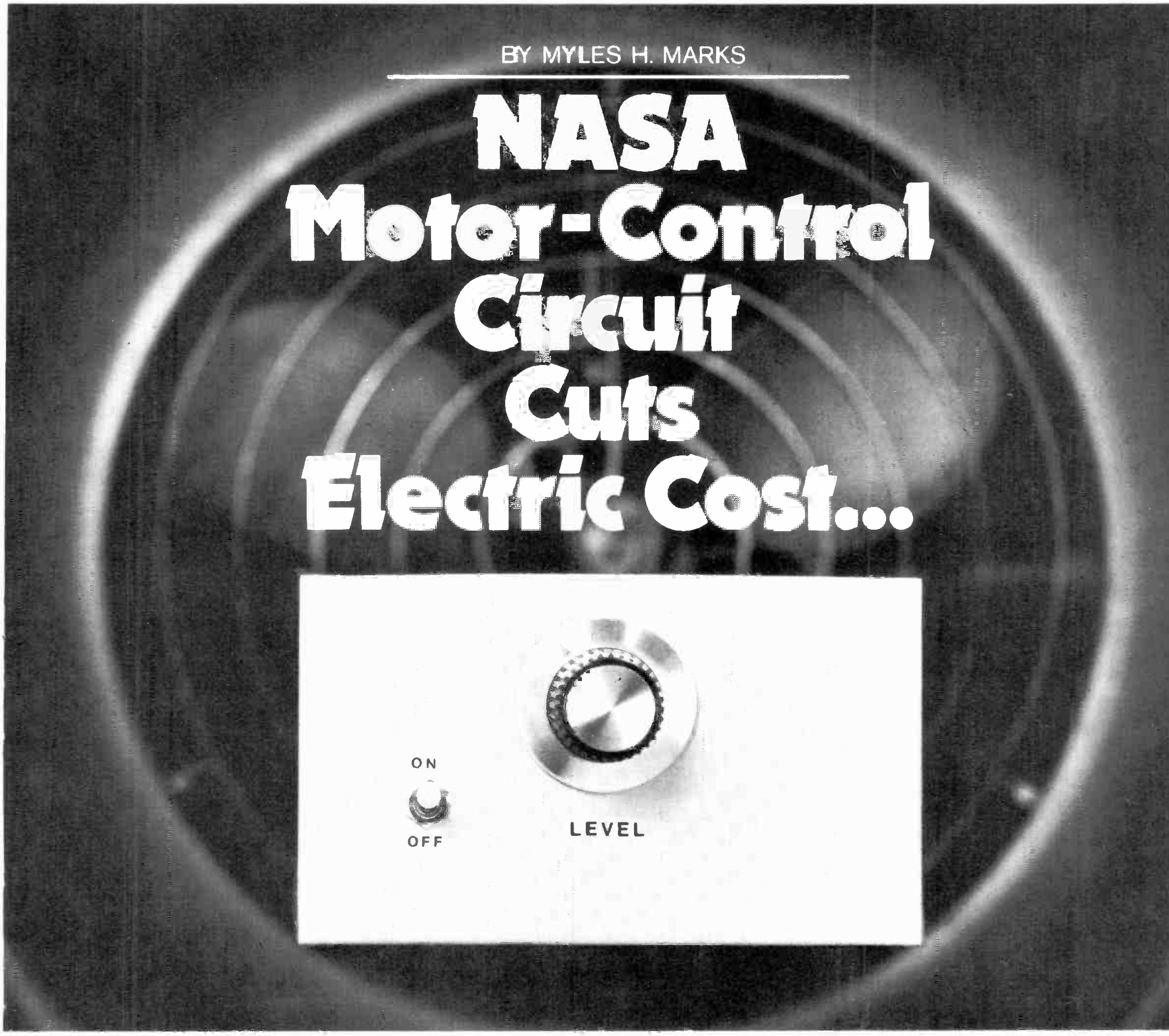
An inexpensive solid-state controller that reduces inefficiencies in electric motors such as those used in refrigerators and dishwashers has been developed at the NASA Marshall Space Flight Center (by Frank J. Nola). Since total electric energy consumed by motors in the U.S. is equivalent to six-million barrels of oil per day and 25% or more of this electricity is pure waste in the form of heat and other factors, the discovery's import is obvious.

BY MYLES H. MARKS

NASA Motor-Control Circuit Cuts Electric Cost...



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Motor-Control Circuit Cuts Electric Cost...

The NASA-developed controller is meant to work with ac induction motors, probably the type most widely used today. They characteristically run at a nearly constant speed that's fixed by power-line frequency and independent of load and supply voltage. When heavily loaded, the motor draws line current that is nearly in phase with the applied voltage, keeping its power factor (cosine of the angle between current and voltage) high and developing a large torque. Under light load conditions, the motor develops less torque by allowing more lag between the voltage and current. This reduces the power factor while leaving the current essentially the same in magnitude.

Though the low power factor means that conversion of electricity to mechanical power is small, the large current causes considerable I^2R losses (heat) in the supply lines and motor windings. This is what reduces efficiency. To minimize this waste, Nola's device monitors the motor's power factor and, when it detects light load conditions, it reduces the supply voltage. This increases "slip" in the motor, which causes a speed reduction of 2% or less so that the motor acts as if it were heavily loaded.

The current, now more nearly in phase with the voltage, therefore does as much useful work as before, but it and the voltage are smaller, resulting in a net saving of electric power.

Power Savings. The device was tested at Marshall Center on over 40 types of motors. Power savings ranged to 60%, depending on the loading. Up to 40-50% power reductions are claimed for motors running lightly or intermittently loaded.

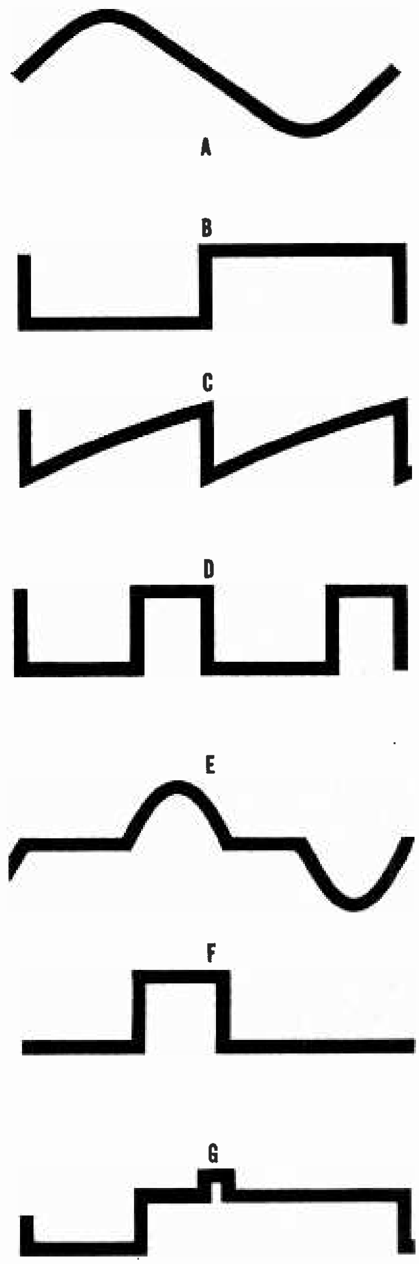
The savings derived by using the controller with motors driving relatively constant loads (refrigeration systems and pumps, for example) are smaller, since the device can then do little more than reduce the 8-10% safety factor allowed for low-voltage conditions. On the other hand, since such motors typically have long duty cycles, significant economies may be realized over a period of time.

Figure 1 was constructed from data averaged from tests made on a ⅓-hp split-phase motor, and ¼- and ¾-hp capacitor-start motors. The top curve shows the typical power required for various loads when no control system is used. The lower curve shows the power consumed when the power-factor controller is used. The controller reduced the no-load power drain by a factor of 5 or 6 and increased the power factor from 0.2 to 0.8. In all three motors, the speed reduction resulting from lower voltage was less than 2%.

Circuit Operation. The circuit shown in Fig. 2, which is a simplified version of the original invention, operates in exactly the same manner. Also shown in Fig. 2, facing the diagram are waveforms for the corresponding letter-in-a-circle points on the schematic.

Typically, current may lag the voltage by 80° in an unloaded motor and only 30° when loaded. The controller continuously monitors phase angle between voltage and current, producing a voltage proportional to that phase angle. This voltage is summed with a preset reference voltage that corresponds to a desired phase angle. The difference between the two produces an error signal that biases a ramp voltage synchronized to the 60-Hz line voltage.

The intersection of the ramp and the error voltages is detected by a squaring amplifier whose output provides proper timing for controlling a triac in series with the motor. The triac is triggered at a point during the cycle, and the circuit switches to "off" as the line current goes through zero. Triggering the triac earlier in each half cycle raises the average voltage to the motor and vice versa.



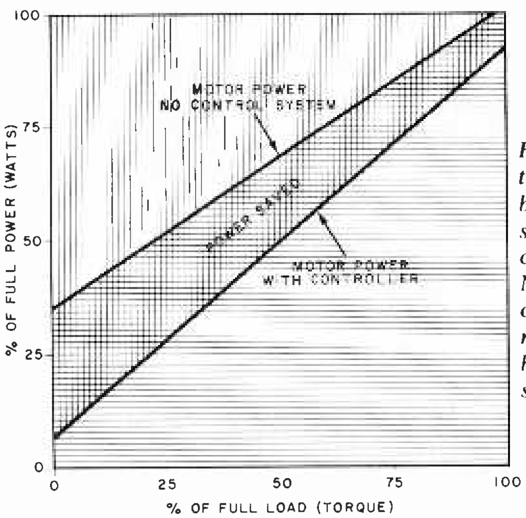


Fig. 1. These curves are the results of tests made by NASA on a ⅓-hp split-phase and ¼- and ¾-hp capacitor-start motors. Note that the power-factor controller reduced the no-load power demand by a factor of 6. Motor slowdown was less than 2%.

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PARTS LIST

C1—1-µF non-polarized capacitor, Mouser Electronics 19NK001 or equivalent
C2—4.7-µF, 20-V electrolytic
C3—6.8-µF, 20-V electrolytic
C4—0.25-µF, 400-V capacitor
C5,C6—470-µF, 35-V electrolytic
C7—2.2-µF, 20-V electrolytic
C8,C9—0.033-µF capacitor
C10—0.33-µF capacitor
D1,D2,D9—1N4148 or 1N914
D3 through D6—1N4001 or similar
D7,D8—1N757, 9.1-V, 400-mW zener
IC1—Quad 741 op amp, LM324N
Q1,Q2,Q3—2N2222 or similar
Q4,Q5—2N2907 or similar
Following are ¼-watt, 5% resistors unless otherwise specified:
R1—0.02 ohm, 5 W (see text)
R2—620,000 ohms (see text)
R3,R18—39,000 ohms (see text for R3)
R4—1800 ohms (see text)
R5—3300 ohms (see text)
R6—1.5 megohms (see text)
R7—100 ohms, 2 W (see text)
R8—51 ohms, 1 W
R9, R13—1000 ohms
R10,R20—3000 ohms
R11,R12,R23,R24,R25—27,000 ohms
R14,R29—9100 ohms
R15—15,000 ohms
R16—68,000 ohms
R17—150,000 ohms
R19—1 megohm
R21—200 ohms
R22—91,000 ohms
R26—36,000 ohms
R27,R28—5600 ohms
R30—20,000-ohm linear taper pot. (see text)
S1—Spst switch
T1—20-V CT, 0.3-A secondary (115/220-volt version is Signal DP-241-4-20 or similar)
Triac—200-V, 15-A (400-V unit for 220-V operation is available. See note below.)
Misc.—Suitable enclosure, heavy-duty ac line cord (male and female connectors), mounting hardware, etc.

Note: The following items are available from M.H. Marks Enterprises, 315 Thornberry Ct., Pittsburgh, PA 15237: Complete kit of all parts for 115-volt system (includes triac, printed circuit board, line cord, cabinet, hardware) for $35.95; complete kit as above for 220-volt system for $41.95; kit of all parts *not* including line cord, cabinet, hardware for $29.95; etched and drilled printed circuit board for $10.95. Please add $3.00 per kit for shipping and handling on domestic orders, appropriate postage for 3 lb for foreign orders. Pennsylvania residents, please add 6% sales tax.

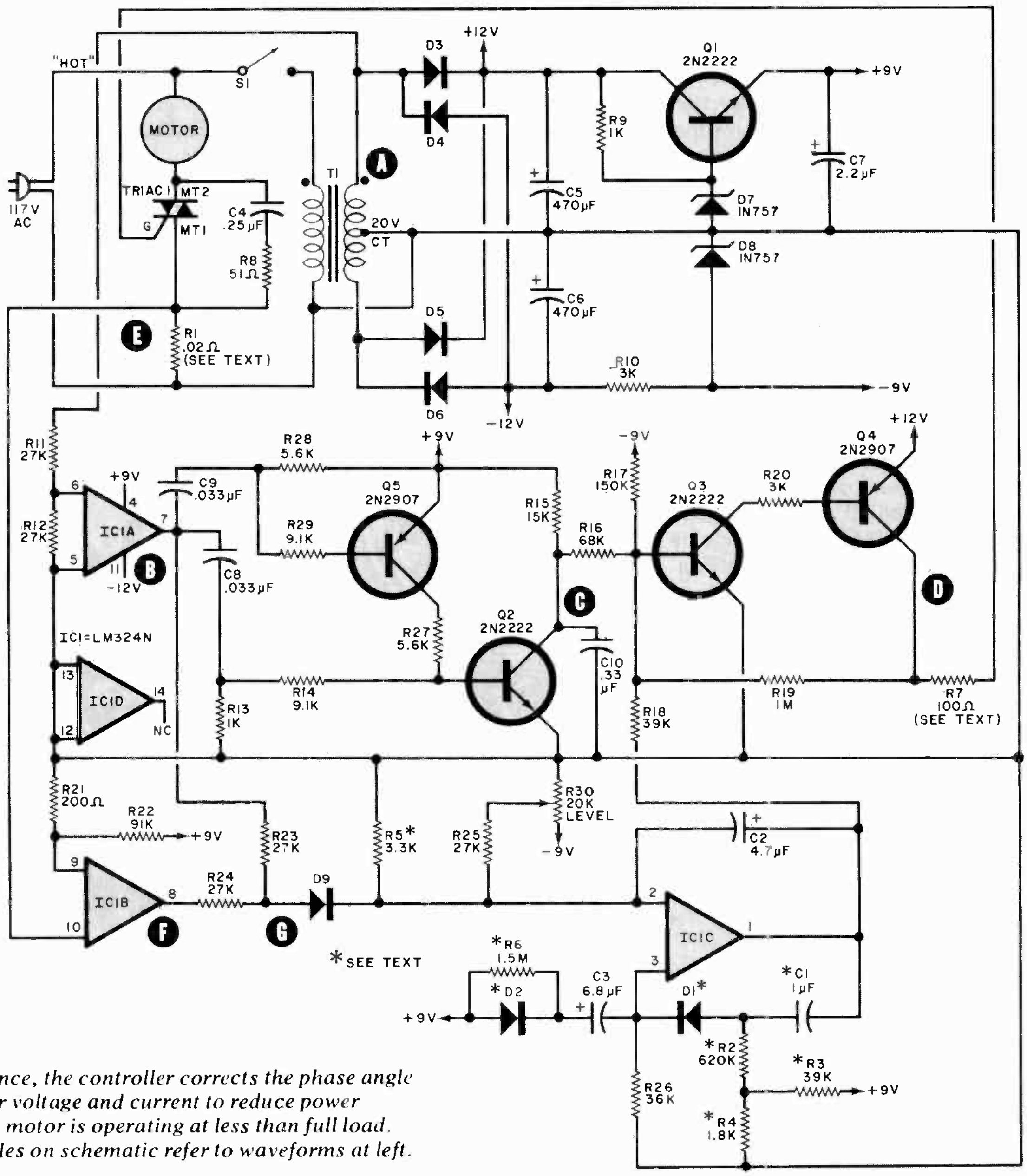

Fig. 2. In essence, the controller corrects the phase angle between motor voltage and current to reduce power required when motor is operating at less than full load. Letters in circles on schematic refer to waveforms at left.

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Motor-Control Circuit Cuts Electric Cost...

The triac's control signal is created by sensing the voltage (A) developed at the top end of transformer *T1*—which also serves as the power source for the conventional dc supply. (Note how the secondary of *T1* is phased with the primary ac power.) The voltage is applied via *R11* to the input of op-amp *IC1A*. Since this op amp is operating at full gain, the output is a square wave at power-line frequency. This IC has two outputs (B). One, via *C8* and *C9*, drives the ramp generator, which consists of *Q5*, *Q2*, and associated components. Capacitor *C10* charges through *R15* to form the ramp. The positive-going step from *IC1A* turns on *Q2*, thus rapidly discharging *C10* to complete the ramp function. The negative-going step from *IC1A* turns on *Q5*, which, in turn, causes *Q2* to saturate, thus discharging *C10*.

Since *IC1A* is triggered at power-line rate, the ramp generated across *C10* is synced to the power line, with each ramp occupying a half power-line cycle. The other output of *IC1A* is coupled through *R23* to diode gate *D9*.

A voltage proportional to the current through the motor (E) develops across sensing resistor *R1*. This voltage is passed to *IC1B*, whose squared-off output (F) is passed through *R24* to diode *D9*, where it combines with the output of *IC1A* to make waveform (G). The summed voltage at the cathode of *D9* is differentiated and fed to integrator *IC1A*, along with a dc control level determined by LEVEL potentiometer *R30*. This control is used to set the motor's optimum phase angle. Time constant network *C3* and *R26* provide a delay to let the motor develop maximum torque when first turned on. Capacitor *C2* provides the high-frequency roll-off necessary for system stability.

Since suddenly applied loads may cause the motor to stall if the system reacts too slowly, the circuit contains some components to prevent this from happening. These parts, which alter the integrator's time constant, are shown with an asterisk in Fig. 2 (*R2*, *R3*, *R4*, *R5*, *R6*, *D1*, *D2*, and *C1*). If you do not need this capability, eliminate these components and tie the positive end of

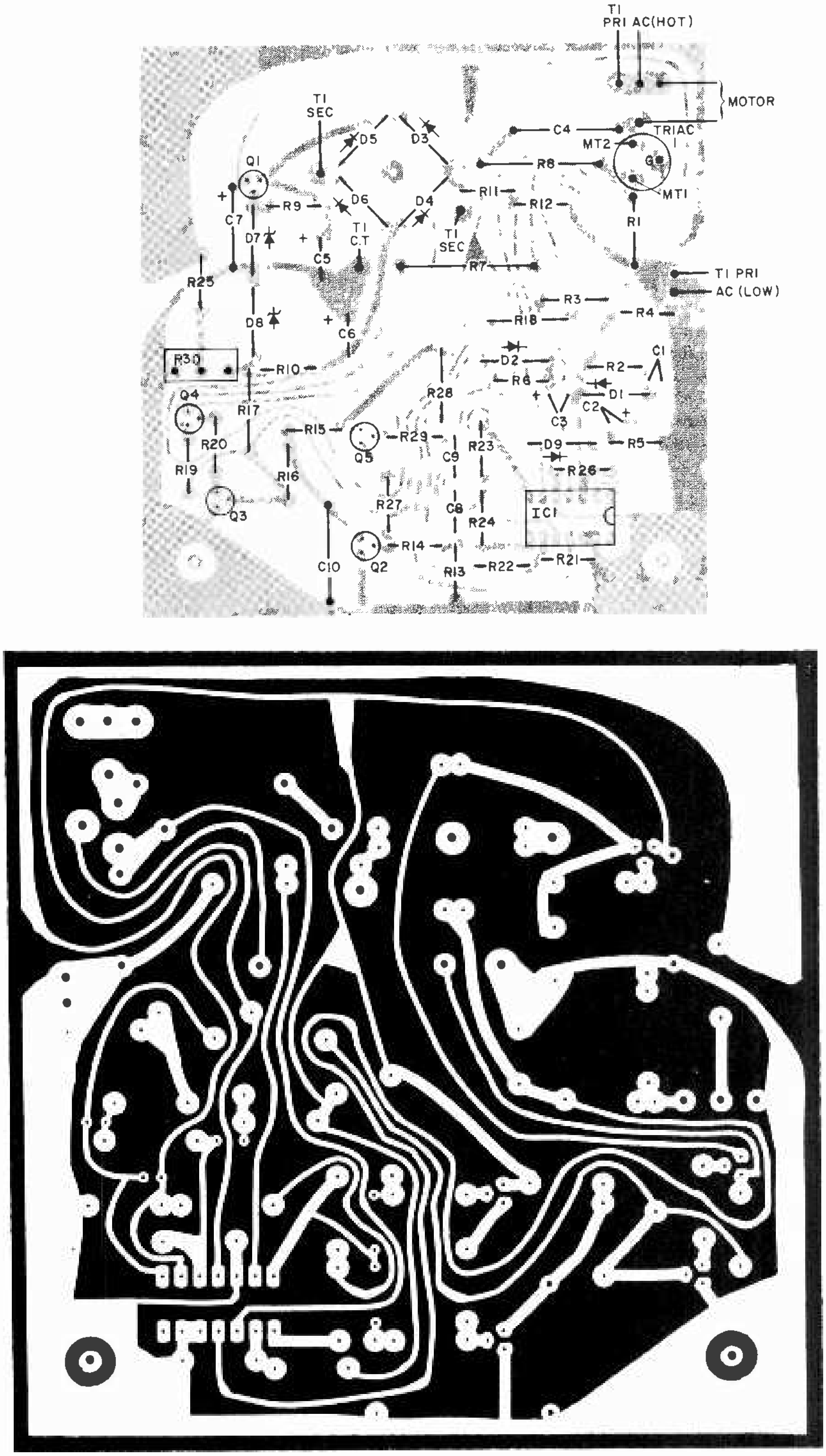

Fig. 3. Actual-size etching and drilling guide for a printed-circuit board for the controller is shown above. Component layout is at top. Note that there are sevral different options regarding components and construction, as outlined in text and Parts List.

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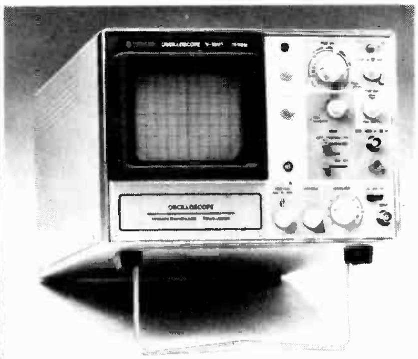

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  - **Display area** 8x10div (1 div = 9.5mm)
  - **Acceleration potential** Approx. 2kV
  - **Intensity modulation** Over 5Vp-p
- **Vertical deflection**
  - **Sensitivity and bandwidth** 5mV/div–5V/div ±5%, DC–15MHz, –3dB; 1mV/div–1V/div ±6%, DC–5MHz Typ., –3dB (Using x5 amplifier)
  - **Rise time** 24ns
  - **Dynamic range** More than 4div at 15MHz
  - **Input R and C** Direct 1M Ohm, approx. 30pF
  - **Maximum input voltage** 600Vp-p or 300V (DC + AC peak)
  - **Display mode** Single-trace
  - **X-Y operation** DC–500 kHz, 200mV/div; Phase difference DC–10kHz 3°
- **Horizontal deflection**
  - **Sweep mode** Auto, NORM, TV (–), TV (+)
  - **TV synchronization** TV sync-separator circuit
  - **Internal** Over 1 div (V sync-signal)
  - **External** Over 1 Vp-p (V sync-signal)
  - **Trigger sensitivity**

Frequency	Internal	External
20Hz–2MHz	0.5div	200mV
2–15MHz	1.5div	800mV

  - **Trigger slope** ±
  - **Sweep time** 0.2µs/div–0.2s/div ±5%, 19 calibrated steps
  - **Sweep-time magnifier** 10 times (±7%)
  - **Max. sweep rate** 100ns/div
- **Amplitude calibrator**
  - **Waveform** 1kHz ±10% Typ. Square wave
  - **Voltage** 0.5V ±3%
- **Power requirements** 100V (120/220/240V) ±10%, 50/60Hz, 40W
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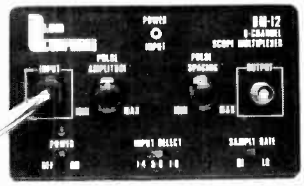

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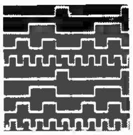

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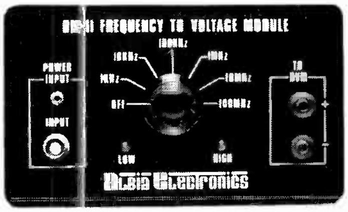

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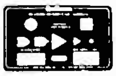

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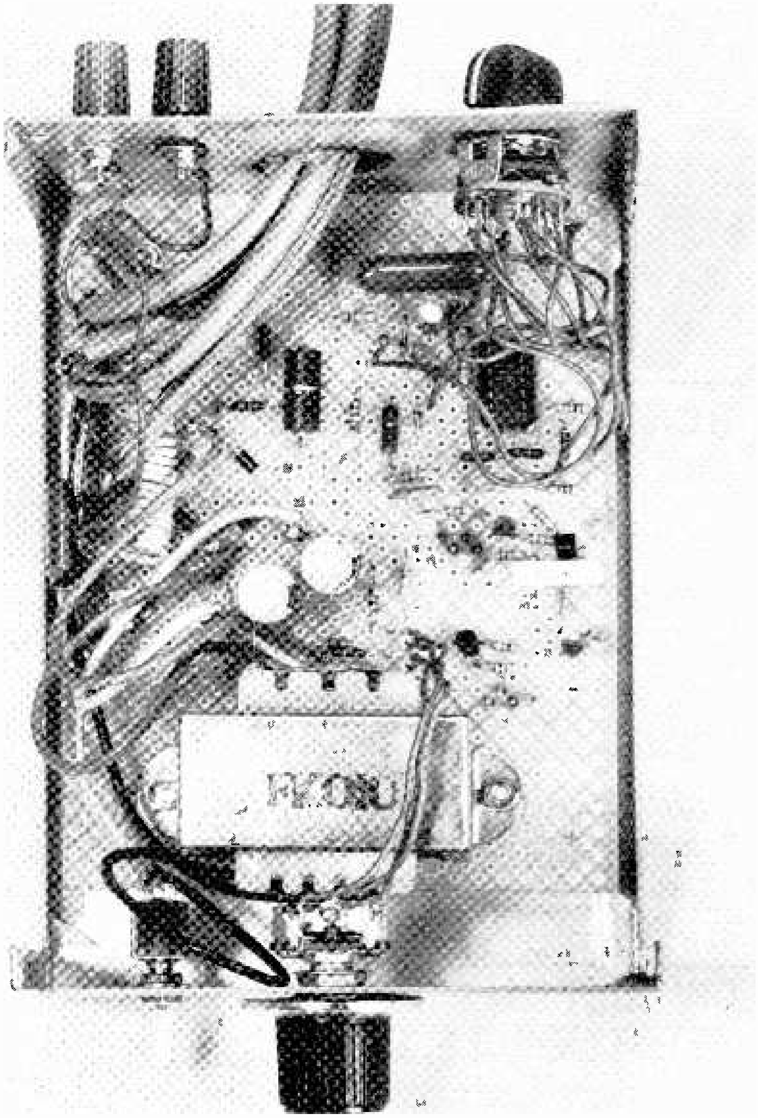
C3 to the +9-volt line by replacing *R6* with a jumper. The *IC1C* output signal is coupled through *R18* to the *Q3* input, in parallel with the ramp from *C10*.

Triac controller *Q3-Q4* is normally biased off by *R17*. When the composite signal (ramp plus pulse) arrives at the base of *Q3*, this transistor will turn on when the peak of the composite signal overcomes the bias. Since the ramp level is fixed, the pulse from *IC1C*, controlled by *R30*, determines when the *Q3-Q4* combination turns on. When turn-on occurs, the waveform shown at (D) triggers the triac, thus applying voltage to the motor.

Construction. The circuit can be most easily assembled on a pc board using the foil pattern and component layout shown in Fig. 3. A bridge rectifier can be used in place of the four rectifier diodes (*D3-D6*). If a 24-volt transformer is employed, increase the value of *R7* to 150 ohms. Resistor *R1* can be fabricated from a 9" length of #22, or a 10" length of #24 solid copper wire that's wound on an insulated support dowel.

At this time, you make the decision about the aforementioned possibility of sudden or clutched-in loads that would require using the asterisked components. Furthermore, if this device is to be used with motors requiring in excess of 300 watts, to prevent damage to the triac or pc board, remove the triac, *R1* and the ac input from the board, mounting a terminal strip in their place. Mount the triac with *R1* to the chassis or optional heat sink (suitably isolated) and wire them into the circuit board, using the terminal strip. Make sure that the "low" side of the ac line is used as the circuit common, and use polarized plugs for all ac-power connections. *Do not use the metal chassis as the common ground!* Failure to observe these precautions may cause a serious shock hazard.

Mount the pc board and transformer in a chassis, securing the board on insulated spacers so that no part of the ac line makes contact with the chassis. If desired, LEVEL potentiometer *R30* can be removed from the pc board and a conventional rotary potentiometer of the same value can be mounted on the chassis. The motor can be plugged into an optional socket mounted on the chassis (wired to the motor-connector pads on the pc board), or use a suitable length of heavy-duty ac line cord having a socket at one end. Do not forget to use ac line cord having sufficient current-carrying capacity to handle the load.



Photograph of the author's prototype which was built on perf board, though a printed circuit board is recommended. The binding posts and switch on the rear were used for testing during design.

Since many of the systems to which the controller can be usefully applied have motors fed from 220-volt ac mains, you may wish to adapt the circuit to work at that voltage. This can be done by exchanging *T1* for a similar transformer with twice as many primary turns and substituting a higher voltage triac (400 PIV minimum). Both "hot" legs of the 220-volt line should be isolated from the chassis, while the center tap should be connected to the ground circuit. An appropriate line plug and receptacle can be used, or the controller can be hard-wired to the load.

Use. Plug the power-factor controller into an ac outlet and connect the motor to be controlled. Turn both on. With the motor operating, slowly adjust LEVEL control *R30* until a slight drop in speed or mechanical power is noticed. Vibration, too, will probably diminish. Slightly back off on *R30* until you feel the point where the speed barely drops off. This should be the optimum setting of the controller. It will probably be necessary to readjust *R30* for each different motor you wish to control.

As noted earlier, the savings effected by using the power factor controller (and the length of time required for the device to pay for itself) depend on the way in which a particular motor is loaded and for what proportion of the time it is in use. Clearly, intermittently used appliances such as power tools are poor candidates. In most households, refrigerators, air conditioners, ventilating fans, swimming-pool pumps, and other machines that run for extended periods will let the power factor controller pay for itself more quickly than smaller and/or intermittently used appliances. Savings will depend on your electric rates, too. In New York City, where one kilowatt-hour costs 11.5 cents in the summer and 9.52 cents in the winter, the controller, used on a 16-cu-ft frostfree freezer, might well pay for itself in about two years. ◇

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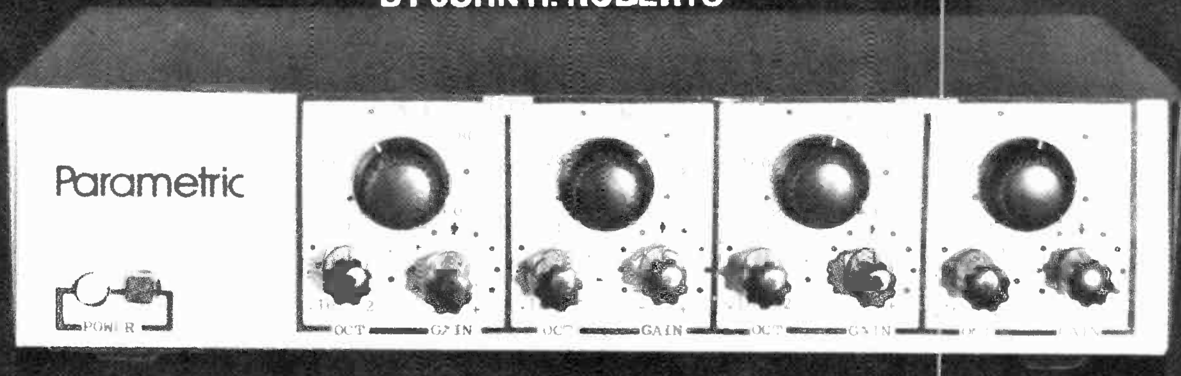
BY JOHN H. ROBERTS



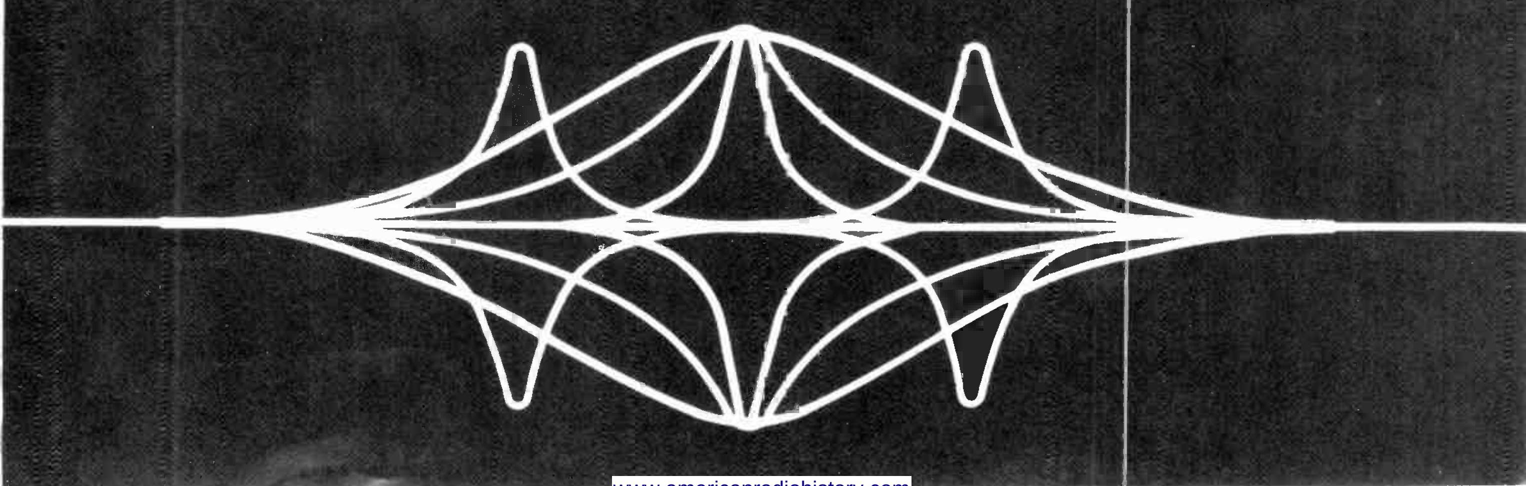
TAILOR THE SOUND OF YOUR AUDIO SYSTEM WITH THIS STEREO PARAMETRIC EQUALIZER

Low-cost, high-performance component employs BIFET operational amplifiers, can be powered by dc or ac sources.

AS THE state of the audio art has matured, whole new families of sophisticated components generically known as *signal processors* have become available for use in sound systems. Among the most popular category of signal processors is the equalizer. And the subcategory that has generated the most excitement among serious audio enthusiasts and sound professionals is the parametric equalizer.

As its name implies, each of the parametric equalizer's key parameters—its center frequency, filter bandwidth or Q, and amount of boost or cut introduced—can be independently adjusted. This provides extraordinary flexibility, allowing the user to tailor equalization to the precise needs for a particular program or room/system combination.

Presented here is a two-band parametric stereo equalizer with several features that commend it to the audiophile. It has been designed so that the Q and BOOST/CUT controls interact to compensate for the perceived change in loudness as filter bandwith increases or decreases. Furthermore, the circuit employs high-performance BIFET op amps, which combine the best of both junction-field-effect and bipolar-junction transistors in each amplifier. It can be powered by either the ac line or a 12-to-30-volt dc supply, making it equally "at home" in fixed, mobile, or portable applications. Finally, the Parametric Equalizer is relatively inexpensive—a line-powered stereo kit costs $99.00.



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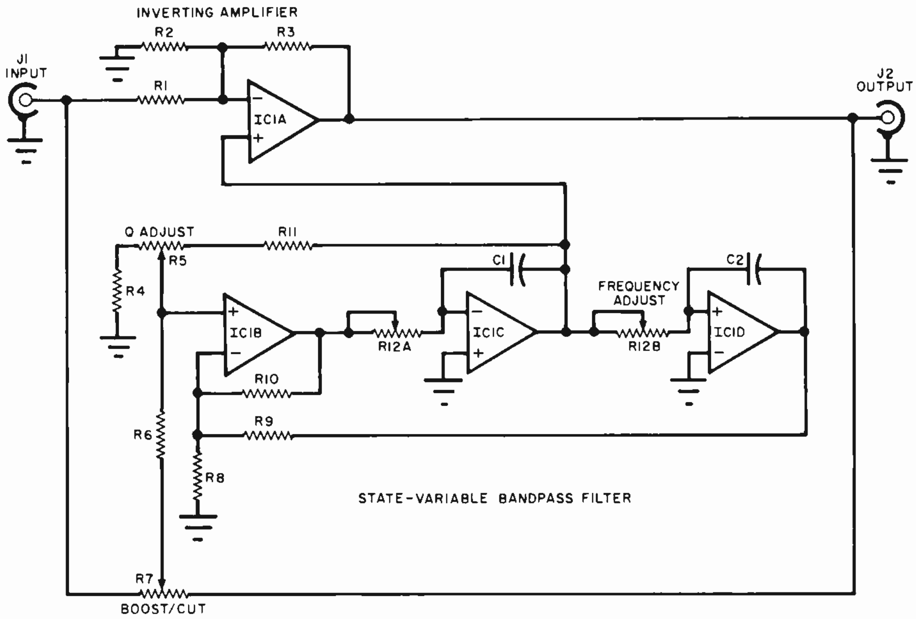
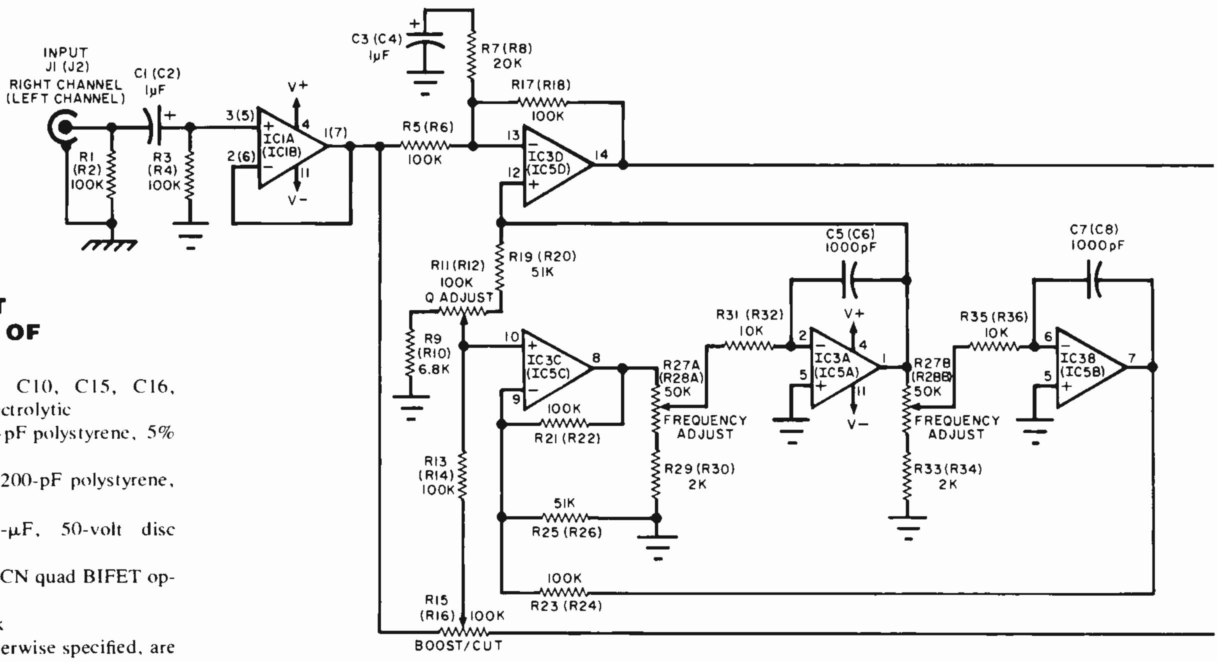


Fig. 1. Simplified schematic of one channel of equalizer shows that an inverting amplifier is interconnected with a modified state-variable active bandpass filter.

A Short Course In Equalization. Here's a brief overview of the subject of audio equalization. The category of signal processors known as equalizer can be broken down into three subcategories: tone control or shelving types; graphic or peaking equalizers; and parametrics. All three are capable of boosting or cutting signal levels, but differ in the manner in which they generate the boost or cut, in the shapes of the frequency-response curves they produce, and in the size of the band of frequencies which they affect.

Tone controls are characterized by a gradual transition between the non-boosted and fully boosted (or unattenuated and maximally attenuated) frequency bands, levelling off to a fixed amount of boost or cut. The resulting frequency-response curve takes on the appearance of a shelf, giving rise to the name *shelving equalizer*.

Graphic equalizers divide the audio spectrum into a given number of bands with individual boost/cut controls for each band. The transition between the unaffected and fully affected regions is determined by the number of bands in the graphic equalizer. An inexpensive five-band or two-octave (so called because each band is two octaves wide) has a lower filter Q and therefore more effect over frequencies somewhat removed from the band of interest than a sophisticated professional equalizer which breaks the audio spectrum down into 30, one-third-octave-wide bands. In most consumer graphic equalizers, the center frequency of each band is fixed, although some more sophisticated units (and most professional graphics) allow the user some leeway in setting the center frequencies. The family of frequency-response curves generated by a graphic



MAIN PARTS LIST (TWO CHANNELS OF EQUALIZATION)

C1, C2, C3, C4, C9, C10, C15, C16, C20—1-μF, 25-volt electrolytic
C5, C6, C7, C8—1000-pF polystyrene, 5% tolerance
C11, C12, C13, C14—8200-pF polystyrene, 5% tolerance
C17**,C18**,C19*—0.1-μF, 50-volt disc ceramic
IC1 through IC5—TL074CN quad BIFET operational amplifier
J1, J2, J3, J4—Phono jack
The following, unless otherwise specified, are ¼-watt, 5% carbon-film fixed resistors.
R1 through R6,R13,R14,R17,R18,R21,R22, R23,R24, R37,R38,R45,R46,R49,R50, R53, R54, R55, R56, R74, R75—100,000 ohms
R7, R8, R39, R40, R63, R64, R67, R68—20,000 ohms
R9,R10,R41,R42—6800 ohms
R11,R12,R15,R16,R43,R44,R47,R48—100,000-ohm, linear-taper potentiometer
R19,R20,R25,R26,R51,R52,R57,R58—51,000 ohms
R27,R28,R59,R60—dual 50,000-ohm linear-taper potentiometer
R29,R30,R33,R34,R61,R62,R65,R66—2000 ohms
R31, R32, R35, R36—10,000 ohms
R69,R70—100 ohms
R71**,R72**,R73*—10 ohms
Misc.—Printed circuit board, pc standoffs, IC sockets or Molex Soldercons, hookup wire, shielded cable, solder, machine hardware, control knobs, suitable enclosure, etc.
*—Dc version only
**—Ac version only

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equalizer resembles a series of peaks and valleys. That's why some audiophiles refer to graphic equalizers as "peaking" types.

The parametric equalizer is a variation on the graphic equalizer theme. In addition to an individual boost/cut control, each band of a parametric equalizer also has center-frequency and bandwidth or filter Q controls. This means that the amount of boost or cut introduced, the center frequency of the band of equalization, and the bandwidth within which the equalization is applied (as well as the transition between the frequencies that are unaffected and those which are boosted or cut the most) are all independently variable. The parametric equalizer thus gives its user the ultimate in control over the sound recorded on tape or reproduced by his speakers.

About the Circuit. A simplified schematic of the Parametric Equalizer is shown in Fig. 1. Only one equalizer section of one channel's circuit is shown, and input buffering and output decoupling details are omitted. Similarly, power supply connections are not shown. It can be seen that the simplified schematic is that of an inverting amplifier (*IC1A*, *R1*, *R2*, and *R3*) interconnected with a modified "state variable" active bandpass filter. Such a filter is composed of two active integrators connected in cascade (*IC1C*, *IC1D*, and associated passive components) and a differential amplifier (*IC1B* and associated passive components).

This circuit was chosen for use in the Parametric Equalizer because its center frequency and Q can be varied independently of each other. The filter's center frequency is selected by adjusting dual potentiometer *R12*. Filter bandwidth and Q are dependent upon the values of *R4* and *R11* and the setting of potentiometer *R5*. For the component values employed in this project, filter bandwidth and Q can be adjusted over a range of 0.16 to 2 octaves at the −3-dB points. (The relationship between bandwidth at the −3-dB points and filter Q is given by the simple equation $BW_{-3\,dB} = 1/Q$.)

To convert a state variable active bandpass filter into the desired all-pass circuit with adjustable boost and cut, a potentiometer (*R7*) is connected between the inverting input and the output of unity-gain amplifier *IC1A*. The wiper of this potentiometer is connected to the input of differential amplifier *IC1B*. Signals appearing at the output of integrator *IC1C*, which are inverted with respect to those appearing at its input, are applied to the noninverting input of *IC1A*.

When the wiper of *R7* is at the *J1* extreme of its travel, the bandpassed signal adds to the input signal, boosting the amplitude of signals within the filter's passband. When the wiper is at the *J2* extreme of its travel, the bandpassed

PERFORMANCE SPECIFICATIONS (Supplied by the Author)

Center frequency range: 40 to 16,000 Hz in two bands—40 to 960 Hz, 500 to 16,000 Hz

Frequency response: 3 to 100,000 Hz, +0 dB, −1 dB with all controls at their flat settings

Input impedance: 50,000 ohms

Input/output gain: 0 dB

Intermodulation distortion (SMPTE): Less than 0.007%

Maximum output: 8 volts rms into a 10,000-ohm load when powered by ±15-volt supply

Maximum boost/cut: ±20 dB at 0.16-octave bandwidth

Output impedance: 100 ohms

Output noise: −70 dBm unweighted, −89 dBm "A" weighted

Range of Q adjustment: 0.16 to 2 octaves (−3-dB bandwidth)

Total harmonic distortion plus noise: below 0.04% from 20 to 20,000 Hz

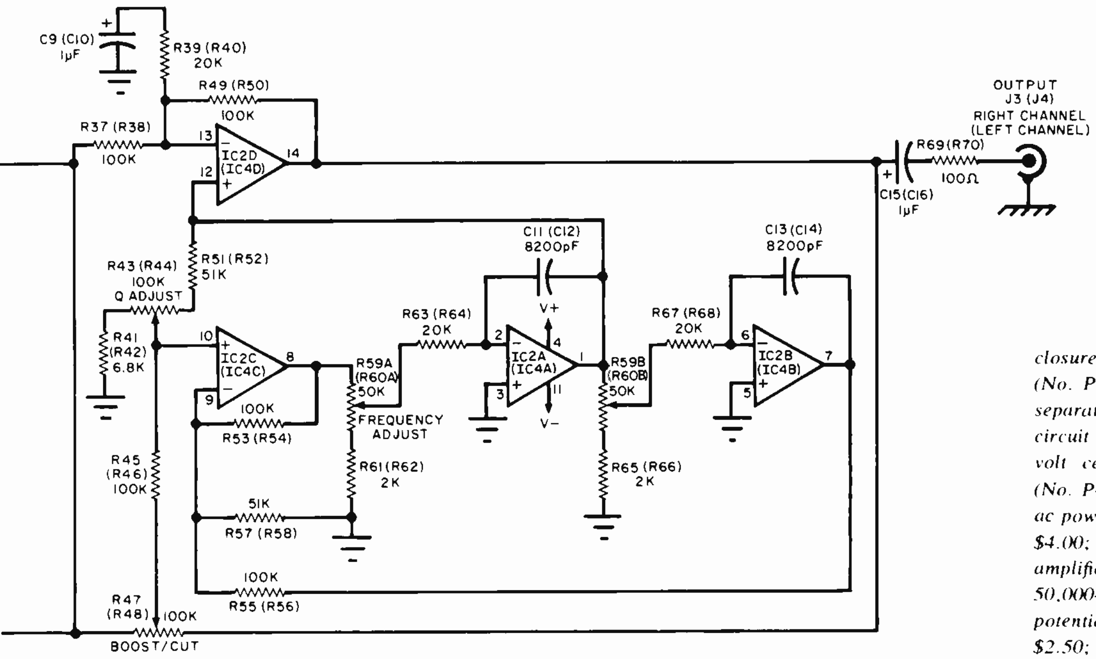

Fig. 2. The complete circuit for a two-channel equalizer. Part numbers not in parentheses are for right channel of a stereo system, others are for left channel. For components with asterisks, see Figs. 3 and 4.

Parts Availability

Note—The following are available from Phoenix Systems, 375 Springhill Road, Monroe, CT 06468 (203-261-4904): Complete kit of parts including enclosure for ac-powered stereo equalizer (No. P-94-S) for $99.00; Complete kit of parts including enclosure for dc-powered stereo equalizer (No. P-94-SC) for $89.00. Also available separately: etched and drilled main printed circuit board (No. P-94-AB) for $8.00; 20-volt center-tapped stepdown transformer (No. P-94-T) for $6.50; etched and drilled ac power supply board (No. P-94-PSB) for $4.00; TL074CN quad BIFET operational amplifier IC (No. P-94-C) for $2.50; dual 50,000-ohm, linear-taper, closely tracking potentiometer (No. P-94-2X50KB) for $2.50; etched and drilled dc power supply board (No. P-94-PSBC) for $4.00; 100,000-ohm, linear-taper potentiometer (No. P-94-100KB) for $1.00; p.c.-mount, push-on/push-off power switch (No. P-94-S1) for $1.00. Add $1.00 handling charge for orders less than $10.00. Add $1.00 for COD orders. Canadians add $2.50 postage. Connecticut residents add state tax.

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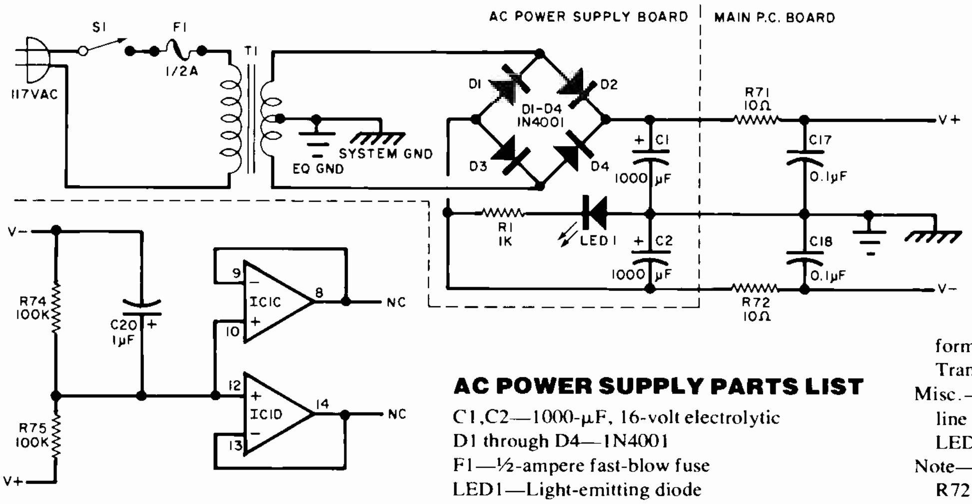

Fig. 3. Schematic of power supply to use with an ac source. It is a conventional full-wave circuit giving plus and minus 15 volts to ground.

AC POWER SUPPLY PARTS LIST

C1,C2—1000-µF, 16-volt electrolytic
D1 through D4—1N4001
F1—½-ampere fast-blow fuse
LED1—Light-emitting diode
R1—1000-ohm, ¼-watt, 5% resistor
S1—Spst switch
T1—20-volt, center-tapped stepdown transformer, secondary rating 100 mA (Signal Transformer No. ST-4-20 or equivalent)
Misc.—Printed circuit board, pc standoffs, line cord, strain relief, hookup wire, solder, LED mounting collar, hardware, etc.
Note— Components C17, C18, C20, IC1, R72, R74 and R75 are mounted on the project's main printed circuit board and are included in the Main Parts List. See Fig. 1 for Parts Availability.

signal subtracts from the input signal, attenuating input signals within the passband of the active filter. Finally, when the wiper of *R7* is at the midpoint of its travel, the output of *IC1A* cancels out that portion of the input signal appearing at the wiper because the two signals are 180° out-of-phase. This means that no signals are routed to the bandpass filter, the filter generates no output, and has no effect on *IC1A*. The result is that inverting amplifier *IC1A* exhibits a flat frequency response.

There are two equalizer sections for each signal channel. (Only one section is shown in Fig. 1.) The center frequency of the low-band equalizer can be adjusted from 40 to 960 Hz, and that of the high-band equalizer from 500 to 16,000 Hz. Both the setting of the BOOST/CUT potentiometer and the value of filter Q determine the amount of boost or cut introduced by each equalizer section. The maximum boost or cut is ±20 dB at a filter bandwidth of 0.16 octave, and ±12 dB at a bandwidth of 2 octaves. This interaction makes the Q control more convenient to use because parametric designs not incorporating it often require readjustment of equalizer gain after the filter Q has been changed.

The master schematic of the main Parametric Equalizer circuit is shown in Fig. 2. The most likely application for this project is in a stereo sound system, so the schematic describes a two-channel equalizer. All components pertaining to the right signal channel have part numbers not shown in parentheses. Those for the left channel, however, have part numbers which are shown in parentheses. The rest of this discussion will refer only to the right signal channel but is equally applicable to the left.

Input signals are applied to jack *J1*, where *R1* and *R3* (which are effectively in parallel) provide a high-impedance load. Capacitor *C1* blocks any dc level that might be accompanying the input signal. Buffering is accomplished by voltage follower *IC1A* which isolates the input from the rest of the circuit. Output signals from the voltage follower are then applied to two cascaded equalizer sections, each of which employs a TL074CN quad BIFET operational amplifier IC.

Each section closely resembles the simplified schematic shown in Fig. 1. That employing *IC3* is the high-band equalizer circuit. Its center frequency is adjustable by means of dual potentiometer *R27* over a range of 500 to 16,000 Hz. Potentiometer *R11* is the filter's Q ADJUST control and potentiometer *R15* (along with the Q of the filter) determine the amount of boost or cut introduced.

The second equalizer circuit (the one

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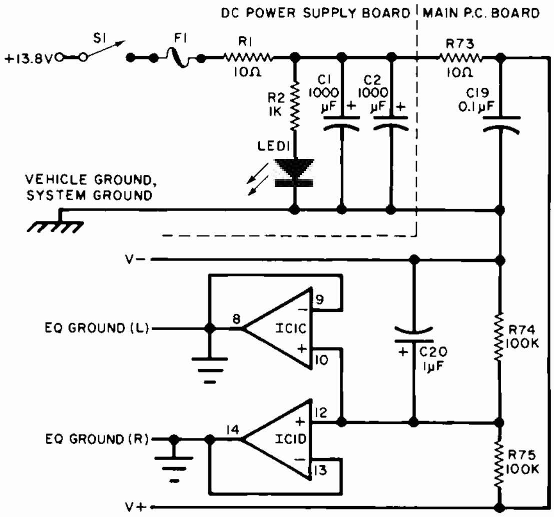

Fig. 4. Use this circuit if a dc supply is to be employed. The IC voltage followers derive an artificial equalizer ground.

DC POWER SUPPLY PARTS LIST

C1,C2—1000-µF, 16-volt electrolytic
F1—½-ampere fast-blow fuse
LED1—Light-emitting diode
R1—10-ohm, ¼-W, 5% resistor
R2—1000-ohm, ¼-W, 5% resistor
S1—Spst switch
Misc.—Printed circuit board, pc standoffs, machine hardware, etc.
Note—Components C19, C20, IC1, R73, R74, and R75 are mounted on the project's main printed circuit board and are included in the Main Parts List. See Fig. 1 for Parts Availability.

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employing *IC2*) is the low-band unit. Dual potentiometer *R59* allows adjustment of its center frequency over a range of 40 to 960 Hz. The filter's Q is adjusted by varying the setting of potentiometer *R43*. Signals within the filter passband can be boosted or cut by means of potentiometer *R47*.

Output signals from *IC2D* are coupled to output jack *J3* via *C15* and *R69*. The electrolytic capacitor blocks any dc offset appearing at the output of the operational amplifier and the resistor provides decoupling. Signals can be routed from the output jack back to the tape monitor loop of a preamplifier or receiver, if that is where drive signals were taken, or to the input of the power amplifier if drive is obtained from the preamplifier output.

Power supply details are omitted from the main schematic for simplicity's sake, but each IC's power supply pins are denoted. The Parametric Equalizer can be powered by either the ac line or a 13.8-volt dc automotive electrical system. Schematic diagrams of the ac and dc supplies are shown in Figs. 3 and 4, respectively. The ac supply is a conventional full-wave circuit employing a 20-volt, center-tapped transformer. Diodes *D1* through *D4* rectify the low-voltage ac into bipolar, pulsating dc which is filtered by *C1* and *C2*. Light-emitting diode *LED1* functions as a pilot light. All components except for decoupling resistors and capacitors *R71*, *R72*, *C17* and *C18* are mounted on a separate power supply circuit board. The output of the supply is ±15 volts dc.

The dc supply employs voltage divider *R74R75* and voltage followers *IC1C* and *IC1D* to derive an artificial equalizer ground at one-half the full voltage delivered by the electrical system powering the circuit. Note, however, that the voltage divider should be connected to the noninverting inputs of the voltage followers even if the ac supply is used to power the circuit. This is done to prevent unwanted oscillation. The outputs of the followers are left uncommitted when the ac power supply is employed.

Light-emitting diode *LED1* acts as a pilot light, and electrolytic capacitors *C1* and *C2* filter any noise present on the dc line. Note that decoupling components *R73* and *C19* as well as the "equalizer ground" deriving circuit are located on the main printed circuit board.

In the dc-powered equalizer, the negative supply voltage pins of the quad operational amplifier IC's are connected to the vehicle and sound system ground (shown in the schematics as "earth

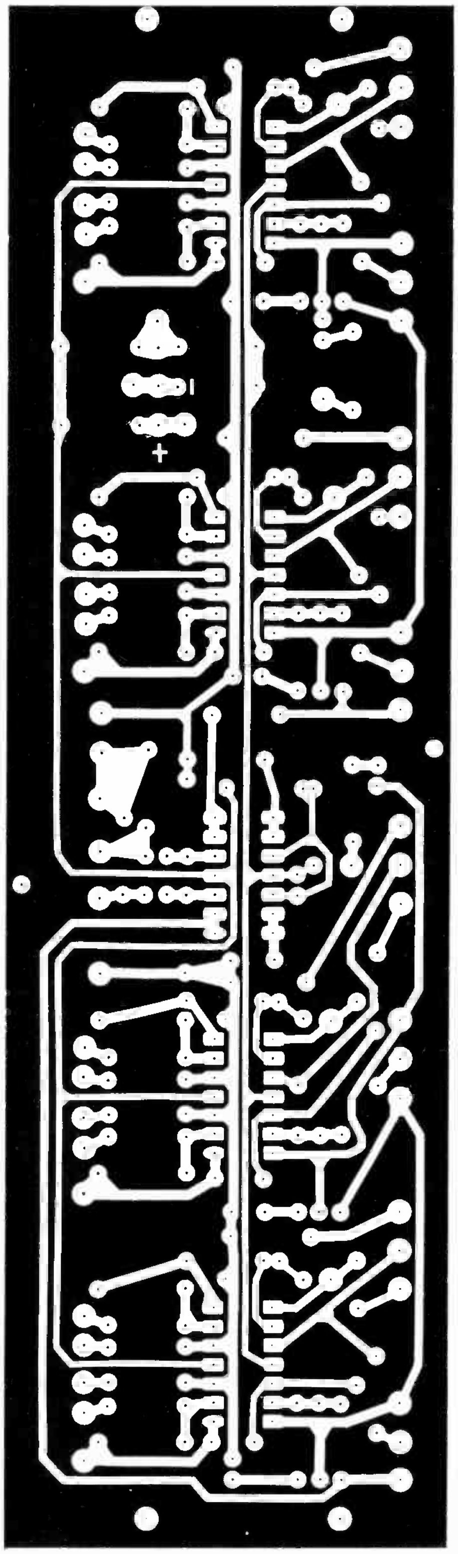

Fig. 5. Actual-size etching and drilling guide for the main printed circuit board.

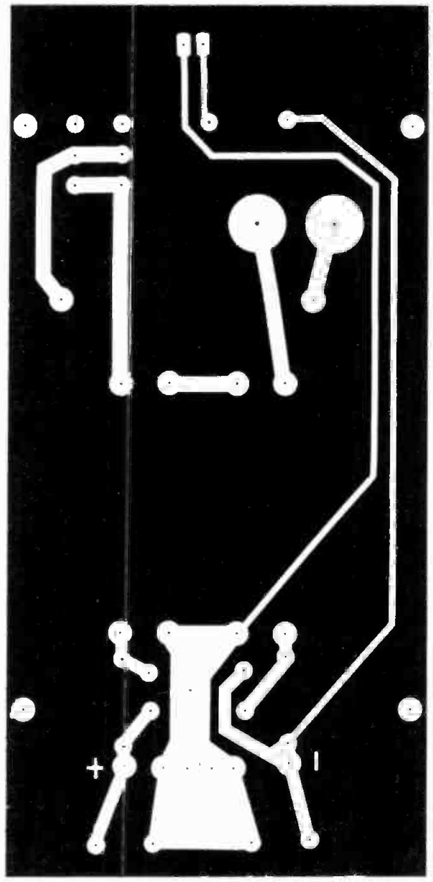

Fig. 6. Use this board for an ac power supply.

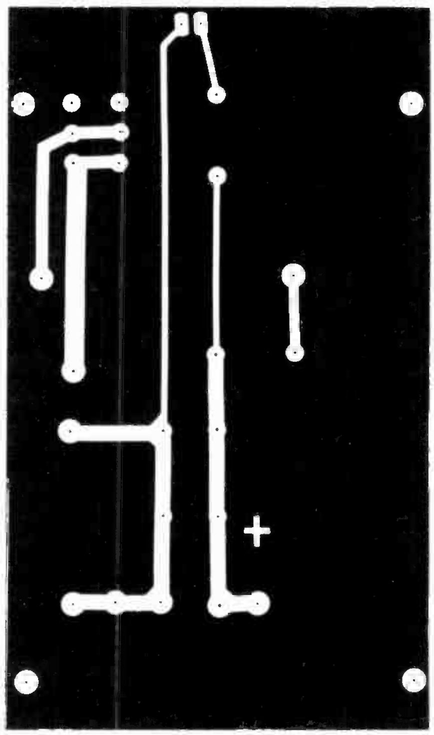

Fig. 7. If a dc supply is available, use this board.

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ground" symbols). The artificial grounds derived by *IC1C* and *IC1D* are shown as conventional "chassis ground" symbols. Note that the grounds within the equalizer sections (for example, the noninverting inputs of the op amp integrators) are artificial grounds above vehicle and system ground.

Capacitive coupling between the input jack and the op amp input buffer and between the output of the high-band equalizer and output jack prevents dc offsets both internal and external to the equalizer from having a deleterious effect on the performance of the entire system. It is because of the dc offsets present in the dc-powered equalizer that the "hot" sides of the input and output jacks are returned to system ground but the signal

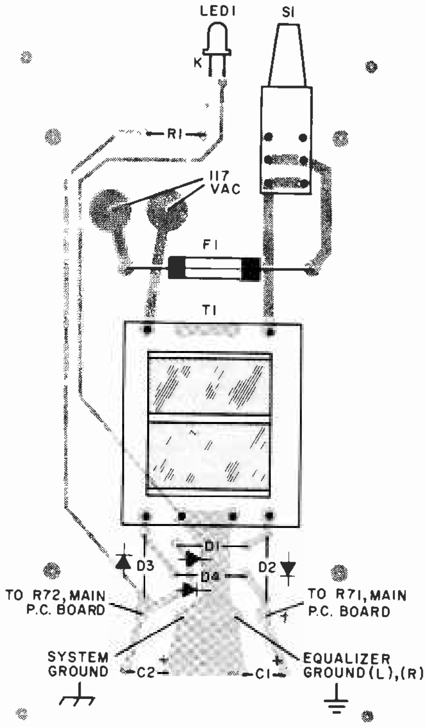
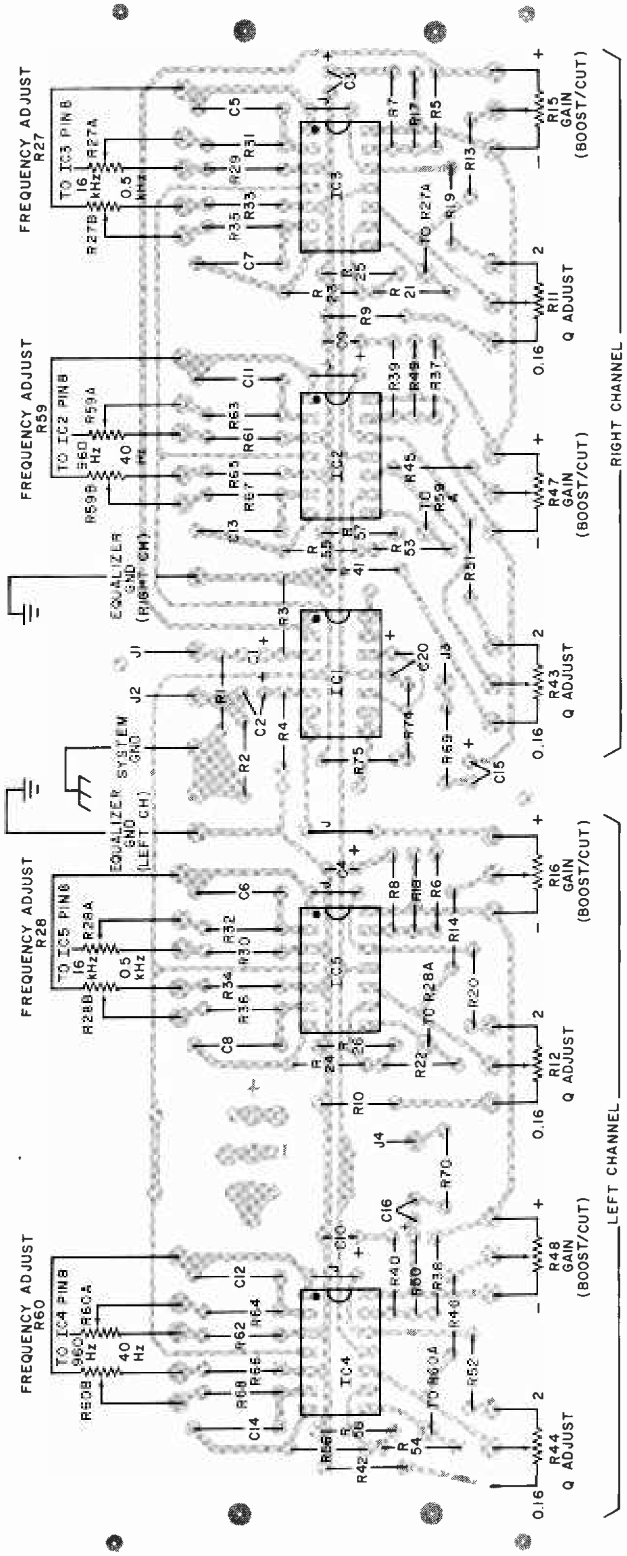


Fig. 9. Component placement for the ac power supply.



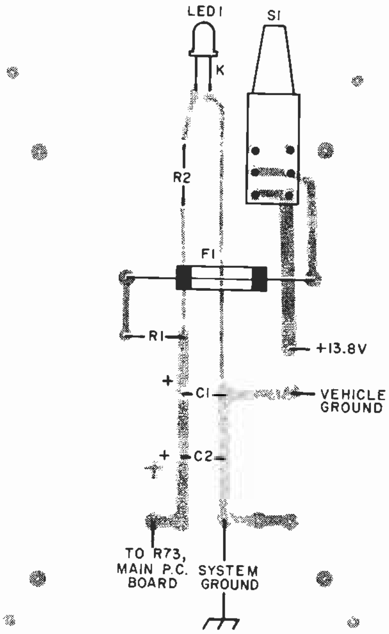


Fig. 10. Component placement for the dc power supply.

Fig. 8. Component placement for the main pc board for the equalizer. Note vacant pads near upper left to make connections to power supplies.

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paths within each equalizer circuit are referenced to the artificial grounds. In the ac-powered equalizer, however, the bipolar dc voltages furnished by the power supply obviate the need for separate system and equalizer grounds. The two are shown connected together in the schematic of Fig. 3.

Results of tests on the prototype performed by the author at his own lab are shown in the box. You will note that all performance specifications but one are identical for both the dc and ac versions of the Parametric Equalizer. The one area in which the two differ is in the maximum voltage swing that can be generated at the output jack. The reason for this is that in the ac-powered equalizer the potential difference between the V+ and V− supply rails is 30 volts, but the potential difference between the supply rails in the dc-powered equalizer is less than half of this value if the dc power source delivers 13.8 volts. However, even in this situation there exists substantial headroom—most (if not all!) autosound power amplifiers require far less drive than 13.8 volts peak-to-peak to develop their maximum levels of output power. Greater output voltage swings can be obtained by increasing the voltage provided by the dc source. The circuit as shown can be used with supplies from +12 to +30 volts.

Construction. The use of printed circuit assembly techniques is recommended. Full-size etching and drilling guides for the main, ac power supply, and dc power supply circuit boards are shown in Figs. 5, 6, and 7, respectively. The corresponding parts placement guides are shown in Figs. 8, 9 and 10.

Mount all components on the circuit boards as shown in the parts placement guides. Begin by installing the jumpers on the main pc board. Then install the fixed resistors and nonpolarized capacitors. Taking care to observe polarities and pin basings, mount the electrolytic capacitors and semiconductors. The use of IC sockets or Molex Soldercons will facilitate replacement of ICs should that become necessary. Interconnection between the main board and the phono jacks and potentiometers can be made using flexible hookup wire. If desired, signal paths between the board and the jacks can be made with shielded cable.

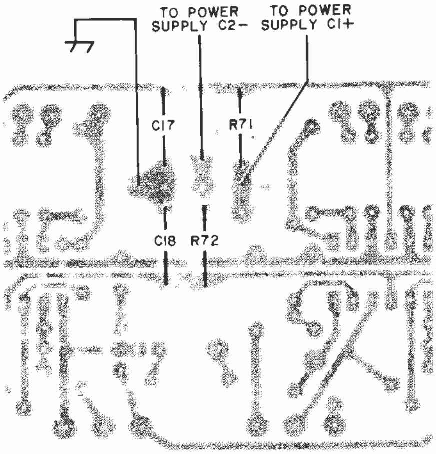

Fig. 11. Special wiring of the main pc board for use with an ac power supply.

This will not be necessary, however, if the project is housed in a grounded metallic enclosure. Special wiring of the main board for ac-powered operation is shown in Fig. 11. Wiring details for dc operation are shown in Fig. 12.

Assemble either the dc or ac power supply to fit the intended application of your Parametric Equalizer. Observe the polarities of electrolytic capacitors and diodes, including the LED pilot light. Fuse *F1* mounts directly on the board and should be soldered to it using pigtail leads. The author designed the power supply boards to accommodate a special push-on/push-off power switch, but any panel-mount switch can be used.

When assembling the circuit boards, be sure to use the minimum amount of heat and solder consistent with the formation of good solder connections. Scrutinize your work after the boards have been completed, paying close attention to polarities, pin basings, power supply wiring and interconnection between the two circuit boards. Make sure that no solder bridges have been created inadvertently.

When all wiring has been completed, mount the circuit boards, jacks and controls in a shielded enclosure. A photograph of the author's ac-powered prototype is shown in Fig. 13. Route power leads out of the enclosure using a protective strain relief. Connect the power leads to a suitable source. Using shielded patch cords, route line-level signals from the tape monitor output of your preamplifier or receiver (or from the preamplifier output) to input jacks *J1* and *J2*. Similarly, patch signals from output jacks *J3* and *J4* back to the tape monitor loop or to the input of the power amplifier. The project is now ready for use.

Using the Parametric Equalizer. Because this project is so flexible, there is no one "correct" way to use it. Its variable Q and center frequency allow the user to boost or attenuate a select group of frequencies. A high Q restricts the boost or cut introduced to a narrow part of the spectrum (less than one octave). A low Q causes broader changes to be introduced.

Adding some sharp boost at the very low and high ends of the audio spectrum allows the user to compensate for speaker rolloff. A broad dip inserted at the midband makes possible the simulation of a loudness contour to enhance low-level listening. The Parametric Equalizer is also adept at compensating for unwanted room resonances. A high-Q cut can reduce audio output at the resonant frequency with little effect on nearby frequencies.

The usual technique for coping with room resonances is as follows. Drive the system with a wideband audio signal

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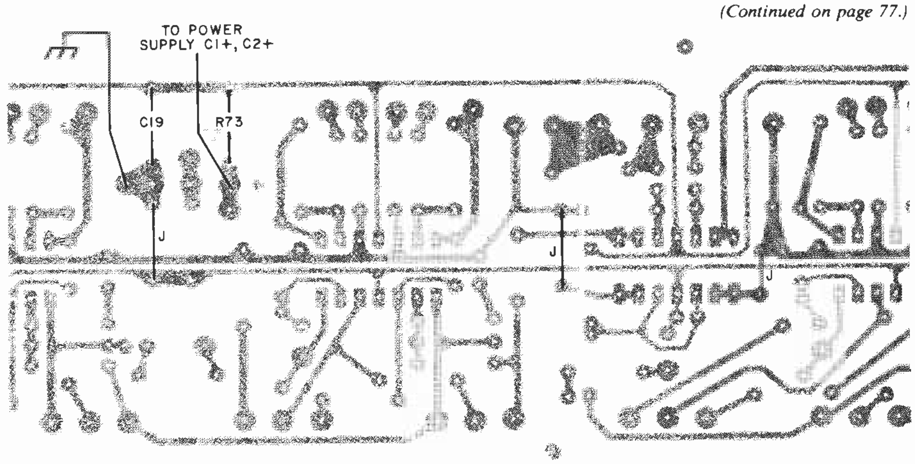

Fig. 12. Special wiring of the main pc board for use with a dc power supply. Note two jumpers on IC1 at right.

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BY WILLIAM VANCURA

Simple circuit triggers electronic system to close garage door after selected time period.

AN AUTOMATIC GARAGE-DOOR CLOSER

THE STANDARD electrically powered radio-controlled garage-door opener has a drawback. It can be falsely triggered by a CB or amateur radio transmitter or other actuating signal, or the user can forget to send a signal command to close the door. In either case, an open garage door could invite thieves to remove valuable equipment—bicycles, lawn mowers, etc. The "Auto Closer" described here overcomes this problem. It automatically commands the system to close the door after a preselected time interval, providing improved security and convenience. The automatic function can be disabled by the user, too, in the event that it is desirable to keep the garage door open.

About the Circuit. The Auto Closer is shown schematically in Fig. 1. Switch *S1* is the door-position sense switch; it remains open as long as the garage door is closed. The open switch prevents the Auto Closer circuit from drawing current from the power supply and keeps it isolated from the rest of the door opener circuit. When the sense switch closes as the door opens, 24 volts ac from the main opener power supply is applied to the Auto Closer. Diode *D5* rectifies the ac into pulsating dc which is filtered by *C1* and *R3*. Zener diode *D1* provides +15 volts regulated for *IC1*, a CMOS 4020 14-stage binary counter.

When power is first applied to the Auto Closer, *R1* and *C2* momentarily keep pin 11 of *IC1* high, ensuring that the counter is reset as the timing cycle begins. A 60-Hz signal from the opener power supply is coupled by *R2* to the counter's CLOCK input. This clocking signal is peak limited by diodes *D2* and *D3*, thereby protecting the counter IC from excessive input levels. The outputs of the twelfth, thirteenth and fourteenth counter stages are available at pins 1, 2, and 3, respectively. When the counter is clocked by a 60-Hz signal, the periods of the square waves at these three outputs are 68 seconds (pin 1), 136 seconds (pin 2) and 272 seconds (pin 3). Each output is high for one-half of its square-wave period.

The time interval that the Auto Closer will hold the door open before automatically closing it is selected by connecting *R4* to one of the output pins of *IC1*. If, for example, *R4* is connected to pin 1, no base current will flow into *Q1* for 34 seconds and the garage door will remain open. At the end of that time, pin 1 will go high and source base current for *Q1* through *R4*. When *Q1* begins to conduct, the coil of reed relay *K1* becomes energized.

This causes the contacts of *K1* to place diode *D4* across the 24-volt ac line. The negative half-cycle of the ac control input is shorted out by the diode. This not only triggers the control circuit to close the door, but also allows the Auto Closer circuit to remain active until the door closes far enough to reopen sense switch *S1*. Capacitor *C3* is connected across the coil of *K1* to keep the relay from chattering and to protect *Q1* from inductive transients. When *S1* reopens, *R5* discharges the Auto Closer capacitors and effectively resets the circuit after a few seconds to ready it for another cycle. Cutout switch *S2* allows you to keep the garage door open for extended periods of time by effectively deactivating the Auto Closer.

Two other time periods are available. If *R4* is connected to pin 2, the garage door will be closed after 68 seconds

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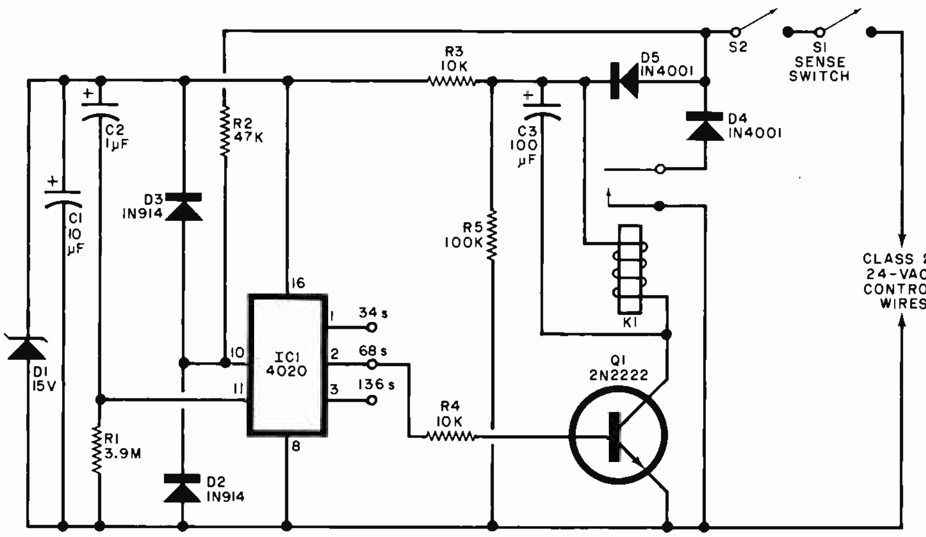

Fig. 1. Schematic of circuit. CMOS counter IC1 develops the delay interval from the 60-Hz line frequency.

PARTS LIST

C1—10-µF, 25-volt electrolytic
C2—1-µF, 25-volt electrolytic
C3—15-100-µF, 25-volt electrolytic
D1—15-volt, 400-mW zener diode
D2, D3—1N914 switching diode
D4, D5—1N4001 rectifier
IC1—CD4020 or MC14020 14-stage CMOS ripple counter
K1—12-volt reed relay (Arrow-M DA-1A or equivalent)
Q1—2N2222 npn silicon switching transistor
The following are ¼-watt, 10% tolerance carbon-composition fixed resistors:
R1—3.9 megohms
R2—47,000 ohms
R3, R4—10,000 ohms
R5—100,000 ohms
S1—SPST spring-loaded, normally open lever switch or other suitable door sense switch (see text)
S2—SPST toggle switch
Misc.—Printed circuit or perforated board, IC socket or Molex Soldercons, suitable enclosure, hookup wire, solder, machine hardware, etc.
Note—The following are available from William Vancura, 4115 35th Avenue, Moline, Illinois 61265: kit of parts, less enclosure and switches $15 plus $1 postage and handling; 12-V reed relay, $5 plus $1 P&H; etched and drilled pc board, $4 with SASE.

have elapsed. Connecting the resistor to pin 3 results in a 136-second delay. The latter interval is enough time for one person to move two cars out of the garage and into the driveway. A delay of 68 seconds is enough time to open the trunk, place a package in it, close the trunk, enter the car, buckle up, start the car once or twice and coax the car into the driveway. A 34-second delay is ideal for an efficient individual who moves quickly but is a cautious driver. Any time delay less than 34 seconds increases the possibility of hitting the door.

The Door Opener. A typical garage door opener employs a "Class 2" wiring system. Basically, this means that the control system comprises a low-voltage supply (usually 24 volts ac derived from a step-down transformer), a control relay which applies power to the motor, and one or more activating switches. The low-voltage power supply cannot cause any serious accidental shocks and permits the use of relatively inexpensive bell wire in connections to the activating switches and relay. A typical system schematic is shown in Fig. 2.

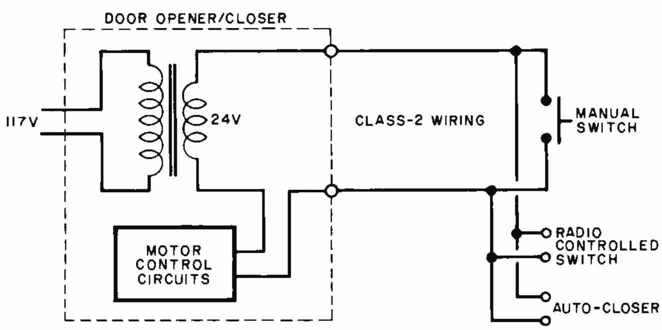

Fig. 2. Class 2 wiring of a typical garage door opener.

Several activating switches can be and usually are wired in parallel across the opener's control input terminals. The system schematic shows a manual pushbutton switch, a radio-controlled switch, and the Auto Closer so connected. Normally, the control circuit is designed to be able to supply power to several low-power switching devices via the Class 2 control wiring itself. This is how both the Auto Closer and radio-controlled switch's receiver are powered.

Construction. Wire-Wrap, point-to-point or printed-circuit techniques can be employed in the construction of the Auto Closer. Parts placement and lead dress are not critical. Although the printed circuit board (see Fig. 3) has been designed to accommodate a reed relay, a perf-board version of the project could use a standard low-power relay such as the Radio Shack No. 275-003.

In any event, use an IC socket or Molex Soldercons with the CMOS counter. Do not insert the IC into its socket at this time. Select the delay period suitable for your application and connect the lead of *R4* to the corresponding pin of the IC socket. Be sure to observe polarities and pin basing of semiconductors and electrolytic capacitors.

The Auto Closer should be housed in a metallic or plastic enclosure approximately 4" x 2¼" x 2¼" (10.2 x 5.7 x 5.7 cm). An on/off switch (*S2*) should be mounted on the enclosure if a remote power switch is not used. The mounting of the sense switch depends on how the door-open condition is to be sensed. One possibility is to install the switch in the Auto Closer enclosure and attach a lever arm to it. The arm can be extended to the door to sense the door-up position. Alternatively, you can replace the travel-limit switch of the motor control circuit with one that has an extra set of contacts. The normally open contacts can be used as sense switch *S1*.

If your garage door opener applies low voltage dc across the control lines instead of 24 volts ac, the Auto Closer circuit should be modified as follows. Replace *Q1*, *C3*, D4 and *K1* with a 1-ampere SCR. The anode of the SCR should be connected to the anode of *D5* and the pole of *S2*. The cathode should be connected to pin 8 of *IC1*, the bottom leg of *R5*, etc. The gate should be connected to *R4*, whose value and that of *R3* should be changed to 1000 ohms. Instead of connecting one end of *R2* to the *D5S2* node, apply 60-Hz CMOS-compatible square waves between it and pin

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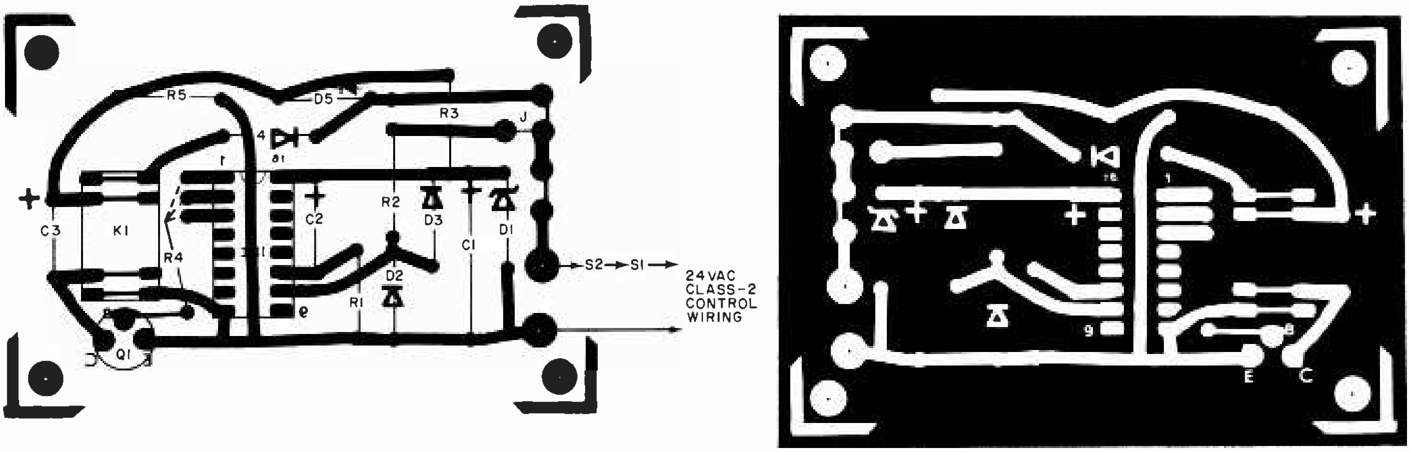

Fig. 3. Etching and drilling and parts placement guides for a suitable pc board.

8 of *IC1*. Zener diode *D1* can be eliminated if the dc control voltage is greater than (or equal to) 12 volts and less than 15 volts.

Checkout. After the Auto Closer has been assembled, but before the CMOS counter has been installed in its socket, temporarily connect one end of a convenient length (about 4 to 6 feet or 1.2 to 1.8 m) of hookup wire to pin 8 of the IC socket. Connect one end of a similar length of hookup wire to the anode of *D5*. Next, attach the two free ends to the Class 2 control wiring of the garage door opener and measure the ac voltage between the anode of *D5* and pin 8 of the IC socket. You should obtain a reading of about 24 volts. Measure the dc voltage between pins 16 and 8 of the IC socket. It should be about +15 volts. Finally, measure the voltage between pins 10 and 8. The meter should read about +15 volts in the dc mode and slightly more in the ac mode. If you have an oscilloscope, look at the signal waveform. You should see a sine wave clipped at 0 and 15 volts.

Momentarily clip a jumper between pins 16 of the IC socket and that to which R4 is connected. The relay coil should become energized and the door opener activated. Removing and replacing the jumper should cause the door opener mechanism to reverse its direction. If the relay chatters while the jumper is connected, the door will jerk back and forth and the Auto Closer will not reliably close the door. This problem can be caused by a defective *C3* or one with insufficient capacitance.

When the Auto Closer is working reliably, it is time to install *IC1* in its socket. The normal precautions should be taken when handling this CMOS device. Disconnect the Auto Closer from the Class 2 wiring and place the circuit board on a 10" x 10" (25.4 x 25.4 cm) sheet of aluminum foil. Also, place the IC (still in its protective foam carrier) and both hands on the foil, which should be grounded. Keeping the heels of both hands on the foil, remove the IC from its protective carrier and insert it into the socket, paying close attention to pin locations. Then permanently install the circuit board in the project enclosure. Reconnect the Auto Closer to the Class 2 control wiring.

If all is well, the door (after having been opened) will begin to close only after the selected delay has elapsed. When the door begins to close, momentarily disconnect the Class 2 control wires from the Auto Closer so that the relay drops out. Each time the Auto Closer is disconnected, the counter will reset itself. Complete the wiring of the sense and power switches and verify the operation of both.

Installation. The Auto Closer is now ready for permanent installation. If a remote sense switch is used, the Auto Closer can be mounted in any convenient location. Just be sure that the control and sense switch wires are positioned so that they do not interfere with the proper operation of the door opener mechanism.

Two methods of mounting an Auto Closer equipped with a built-in lever sense switch are shown in Figs. 4A and 4B. The latter installation is less sensitive to minor variations in the stopping position of a door riding on tracks beside the sense switch lever. A slight bend at the tip of the lever arm prevents the door from snagging and damaging itself. The mounting method shown in Fig. 4B allows the project enclosure to be mounted easily on the door track using a 4" (10.2-cm) hose clamp. The ceiling mount (Fig. 4A) will work equally well with either a single-piece trackless door or a multi-section tracked door.

In Conclusion. You will surely find the Auto Closer to be a great convenience and an effective security device. Keep in mind, however, that you can very easily lock yourself out of the house should you forget your keys, the opener's pocket transmitter, or to disable the Auto Closer! ◇

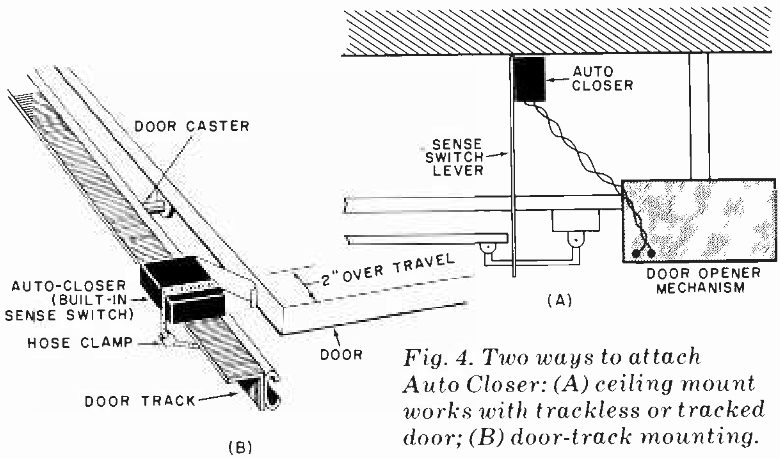

Fig. 4. Two ways to attach Auto Closer: (A) ceiling mount works with trackless or tracked door; (B) door-track mounting.

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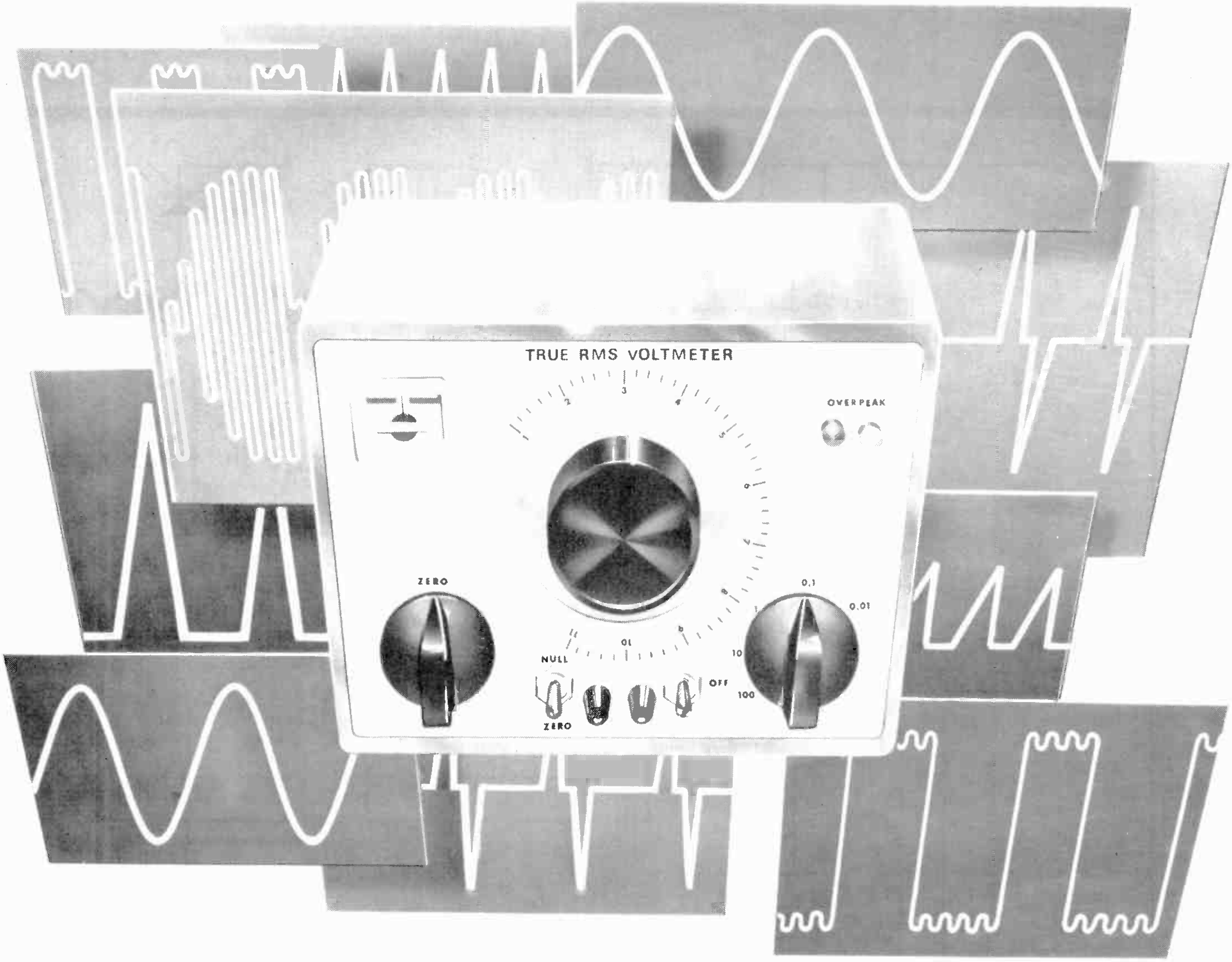

BUILD A TRUE RMS VOLTMETER

Measures effective ac voltage rather than the usual average values

BY DANIEL METZGER

EVENTUALLY, most serious electronics experimenters find a need to measure the power content of an odd-shaped waveform. This might occur in measuring the output of a class-C amplifier, audio, video or noise signals; power supply inverter waveforms; or just plain power supply ripple.

When the time comes, the experimenter will discover that a standard VOM or DVM simply won't provide true rms readings! An rms voltmeter is needed for this purpose. Unfortunately, this is a rather expensive instrument.

With the plans presented here, however, a true rms voltmeter can be built for only about $30. To fully understand how this is accomplished, let's first examine what rms means.

What Is RMS? Rms values allow the expression of the average power content of an ac waveform whose instantaneous power varies from zero (at the zero crossing) to some high value at the voltage peak. In practice, rms means "equivalent to dc." Mathematically, given a voltage and a resistance, power is calculated as $P=E^2/R$.

The "equivalent-to-dc" voltage for a continuously varying ac voltage must be calculated by squaring the voltage at each instant, averaging the infinte number of instantaneous values produced (taking the mean), and then extracting the square root. Hence, the term Root of the Mean of the Square (RMS).

For anything beyond a simple rectangular waveform, this rms technique involves calculus. For sine waves *only*, it produces $E_{RMS} = E_{PK(SINE)}/\sqrt{2} = 0.707E_{PK}$.

For periodic waveforms that cannot be mathematically expressed, the rms technique requires iterative computer analysis. For completely undefinable

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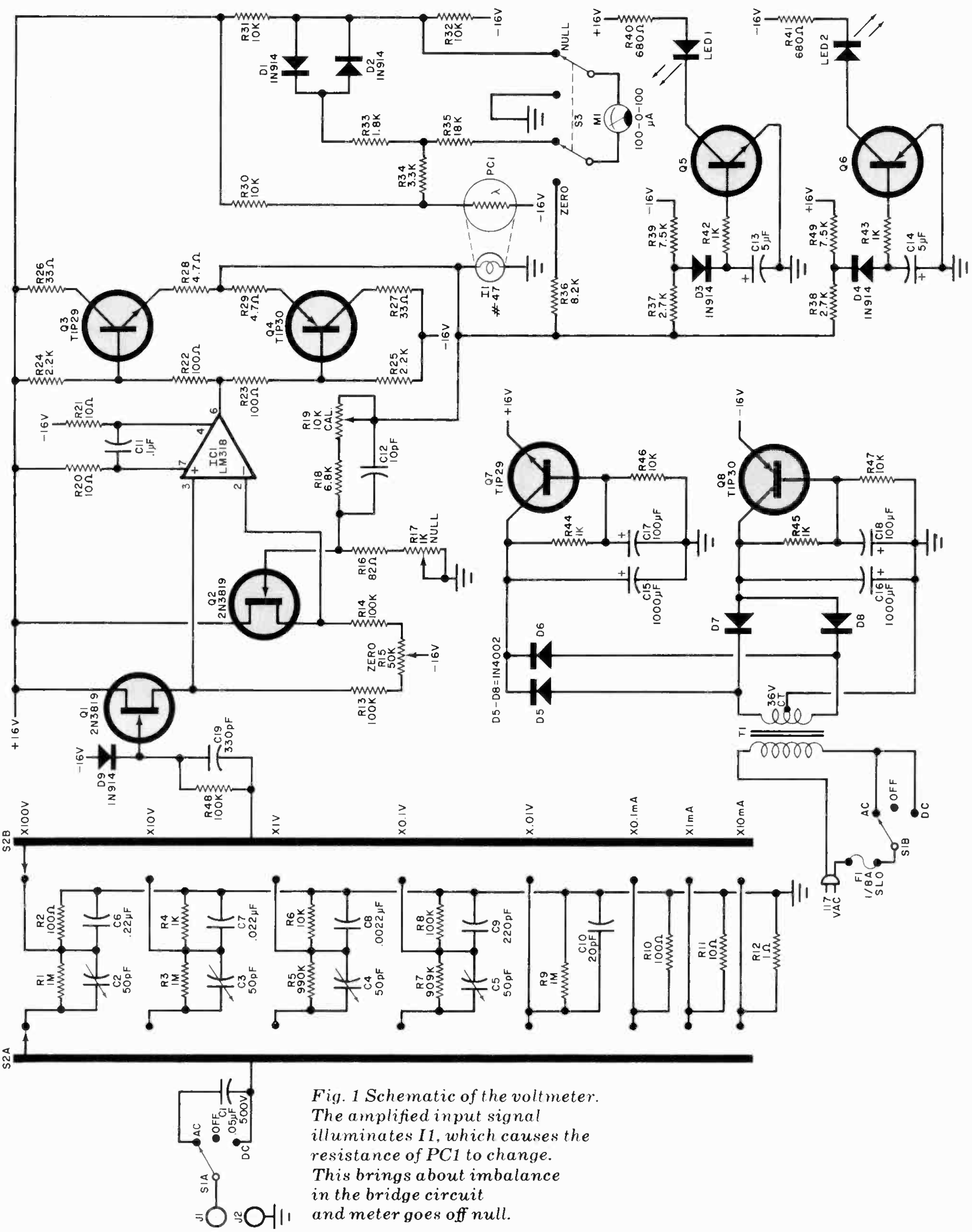

Fig. 1 Schematic of the voltmeter. The amplified input signal illuminates I1, which causes the resistance of PC1 to change. This brings about imbalance in the bridge circuit and meter goes off null.

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PARTS LIST

C1 — 0.05-μF, 500 volt capacitor
C2 through C5 — 50-pF trimmer
C6 — 0.22-μF Mylar capacitor
C7 — 0.022-μF Mylar capacitor
C8 — 0.0022-μF capacitor
C9 — 220-pF capacitor
C10 — 20-pF capacitor
C11 — 0.1-μF capacitor
C12 — 10-pF capacitor
C13,C14 — 5-μF, 15-volt electrolytic
C15,C16 — 1000-μF, 35-volt electrolytic
C17,C18 — 100-μF, 25-volt electrolytic
C19 — 330-pF capacitor
D1 thru D4, D9 — 1N914
D5 thru D8 — 1N4002
F1 — 1/8-ampere slow-blow fuse
I1 — #47 pilot lamp
IC1 — LM318 op amp
J1,J2 — Five-way binding posts, one red (J1), one black (J2)
LED1,LED2 — Light emitting diode
M1 — 100-0-100-μA meter
PC1 — See text
Q1,Q2 — 2N3819,MPF102, or similar
Q3,Q7 — TIP29 or similar
Q4,Q8 — TIP30 or similar
Q5,Q6 — Any small-signal silicon transistor
R1,R3,R9 — 1-megohm, 1%, 1 watt resistor
R2,R10 — 100-ohm, 1% resistor
R4 — 1000-ohm, 1% resistor
R5 — 1-megohm, 1% resistor (try for 990k)
R6 — 10,000-ohm, 1% resistor
R7 — 909,000-ohm, 1% resistor (try for 900k)
R8 — 100,000-ohm, 1% resistor
R11 — 10-ohm, 1% resistor
R12 — 1-ohm, 1% resistor
R13,R14,R48 — 100,000-ohm resistor
R16 — 82-ohm resistor
R18 — 6800-ohm resistor
R20,R21 — 10-ohm resistor
R22,R23 — 100-ohm resistor
R24,R25 — 2200-ohm resistor
R26,R27 — 33-ohm, 1-watt resistor
R28,R29 — 4.7-ohm resistor
R30,R31,R32,R46,R47 — 10,000-ohm resistor
R33 — 1800-ohm resistor
R34 — 3300-ohm resistor
R35 — 18,000-ohm resistor
R36 — 8200-ohm, ½-watt resistor
R37,R38 — 2700-ohm 5% resistor
R39,R49 — 7500-ohm 5% resistor
R40,R41 — 680-ohm resistor
R42,R43,R44,R45 — 1000-ohm resistor
R15 — 50,000-ohm linear-taper potentiometer
R17 — 1000-ohm linear-taper potentiometer (see text)
R19 — 10,000-ohm, trimmer pot
S1 — Dpdt center-off toggle switch
S2 — 8-position, 2-pole rotary switch
S3 — Dpdt switch
T1 — 36-V CT, 0.1-A transformer (Stancor 8611 or similar)
Misc. — Suitable enclosure, dial plate and knob for R17, pointer knobs (2), press-on type, mounting hardware, etc.

waveforms such as noise, however, we have to go back to the rms voltmeter.

This may be a good place to observe that the time-honored VOM and most DVMs actually respond to *average* voltage (not rms) on their ac scales. The average value of one half-cycle of a sine wave is 0.637 of its peak value, or 0.901 of its rms value.

The meter scales are calibrated in rms, but this calibration is valid for sine waves only. VTVMs and FET meters generally respond to peak voltages on their ac ranges. Again, the meter scales are calibrated for rms readout on sine waves only. Consequently, a VOM or VTVM can display different and incorrect conclusions about a 25% duty cycle waveform, as illustrated in boxed section below right. Interestingly, most VOMs err in one direction on non-sine waves, while VTVMs err in the other.

The majority of true rms meters operate by electronically squaring and averaging the input signal, or by taking the log of the signal, doubling it, then taking the antilog. Building such a meter is best avoided because of its circuit complexity. The meter discussed here takes the direct approach of amplifying the input waveform and using its large signal to turn on a conventional incandescent lamp. The brilliance of the lamp is sensed by a photocell so that comparisons between ac and equivalent dc voltages can be made. In operation, the amplifier gain is adjusted until the lamp reaches a predetermined brightness, and the rms voltage is then read off a calibrated "gain" control.

Circuit Description. As shown in Fig. 1, resistors *R1* through *R9* provide voltage divisions by progressive factors of ten while maintaining a 1-megohm input resistance. Capacitors *C2* through *C10* provide identical division ratios at high frequencies while maintaining an input capacitance of 20 pF. Without these capacitors, the reactance of any stray wiring capacitance would completely swamp the resistive dividers above approximately 50 kHz, resulting in false division ratios. Resistors *R10* through *R12* are used in the measurement of current.

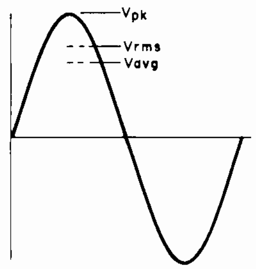
Source follower *Q1* provides an almost infinite input impedance for the amplifier, while FET *Q2* provides compensation for variations in *Q1* source voltage with temperature. Both *Q1* and *Q2* should be well matched—zero volts between their sources with the input grounded and control *R15* (ZERO) near the center of its rotation. Also, the transistor cases should be thermally bonded.

Op amp *IC1* has a gain between 10 and 100, depending on the setting of the CAL *(R19)* and NULL *(R17)* controls. Capacitor *C12* compensates for stray capacitance in the *R16-R17* leg, and is selected for good high-frequency response. Components *R20, R21* and *C11* are required to prevent high-frequency oscillation via power supply coupling. A complementery symmetry voltage follower (*Q3* and *Q4*) provides the high current required by lamp *I1*.

The sensitive surface of photoresistive cell *PC1* is butted to the lamp and secured in place with opaque heat-shrink tubing. The photocell forms one arm of a Wheatstone bridge, with *R30-R32* as the other arms. At one particular

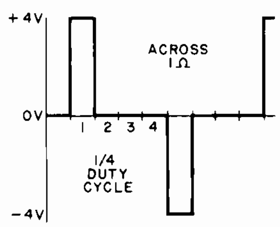
VTVM AND VOM ERRORS

For sine waves:



$V_{PK} = 1.414V_{RMS} = 1.569V_{AVG}$
$V_{RMS} = 0.707V_{PK} = 1.110V_{AVG}$
$V_{AVG} = 0.637V_{PK} = 0.901V_{RMS}$
VOMs read $1.110V_{AVG}$
VTVMs read $0.707V_{PK}$

For this rectangular wave:



$P_{INST} = E^2/R = 4^2/1 = 16\text{ W}$
$P_{AVG} = 16/4 = 4\text{ W}$
$E_{RMS} = \sqrt{PR} = \sqrt{4 \times 1} = 2.00\text{ V}$
VOM reads:
$1.110 \times V_{AVG} = 1.110 \times ¼ \times 4$
$= 1.11\text{ V or } 2.00\ V_{RMS}$
VTVM reads:
$0.707 \times V_{PK} = 0.707 \times 4$
$= 2.83\text{ V or } 2.00\ V_{RMS}$

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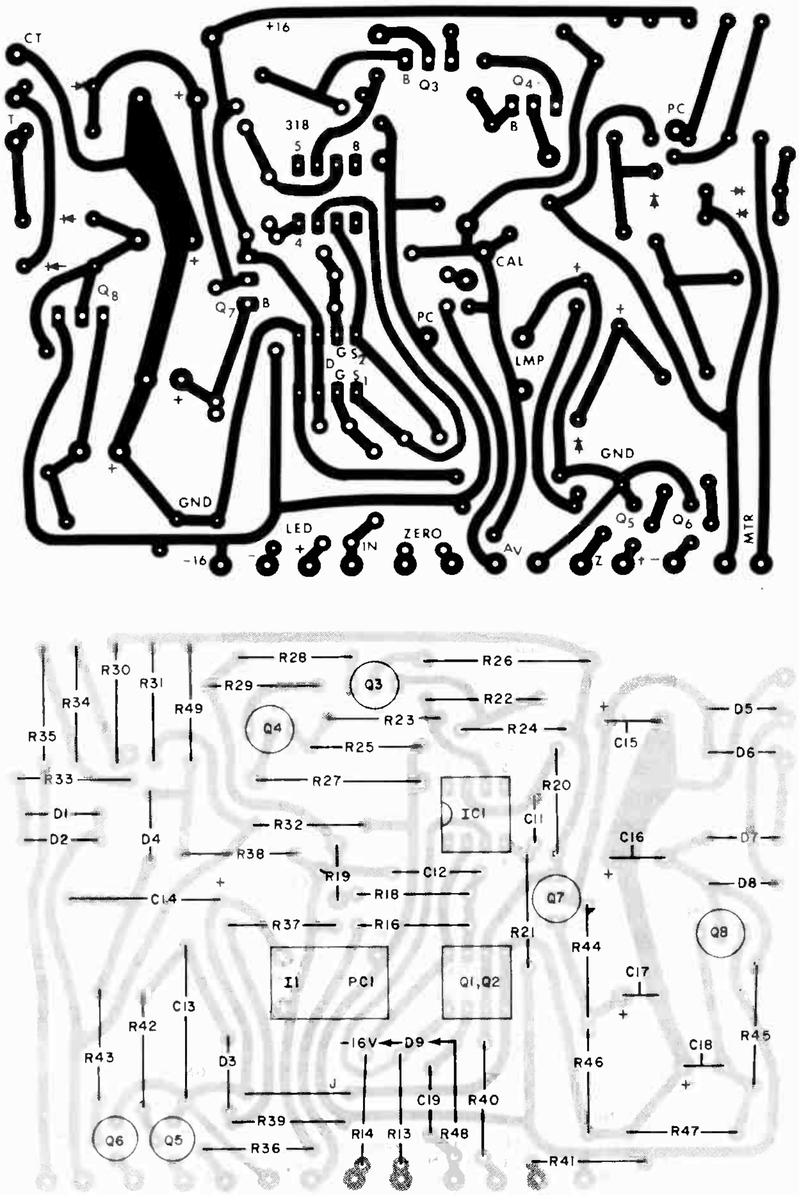

Fig. 2. Actual-size foil pattern for main pc board is at top, component installation immediately above.

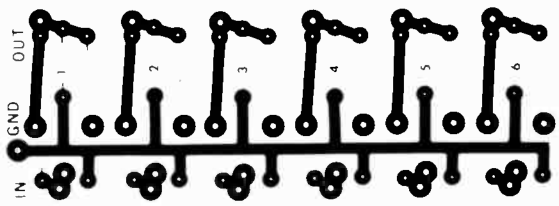

Fig. 3. Components for attenuator can be mounted on board similar to that at left. Only four sections are used in this case.

lamp intensity, the cell resistance will equal 10,000 ohms and meter *M1* will indicate zero. For unbalance currents of about 25μA, *D1* and *D2* will begin to conduct, shunting the meter with *R33*. This makes it easier to find the null by reducing the meter sensitivity for off-null settings. The meter can be switched to zero out dc voltage at the amplifier output when the input is grounded.

In the prototype, balance was achieved with 2 volts rms (or dc) across the lamp. For sharply peaked waveforms, it is possible that the peak voltage might exceed the approximately 10-volt limit of the amplifier resulting in distortion of the waveform. Diodes *D3* and *D4* detect such peaks slightly before the threshold of distortion, causing the front panel positive *(LED1)* or negative *(LED2)* overpeak indicators to glow.

The power supplies are well filtered by the emitter follower action of *Q7* and *Q8*. Regulation is not required because the differential amplifier and bridge circuits remain balanced in spite of power supply voltage variations.

Construction. Few of the components are critical and substitutions are possible. The NULL potentiometer *(R17)* should be of the highest quality since accuracy of the instrument depends on reading its angular position. Resistors *R1* through *R12* directly affect the accuracy and should be held to ±1%. For the voltage dividers, it is the resistance *ratio* that is important. Ceramic trimmer capacitors hold their values better than compression types and are recommended for *C2* through *C5*. Transistors *Q3* and *Q4* may dissipate as much as 2 watts and must have heat sinks.

Using an LM318 for *IC1* results in a 400-kHz bandwidth. A μA741 reduces the bandwidth to about 20 kHz. Any photoresistive cell having a resistance of 10,000 ohms when illuminated by a #47 lamp with 1.5 to 2.5 volts dc applied to it will work for PC1.

The bulk of the circuit can be assembled on a pc board using the foil pattern and component layout shown in Fig. 2. The front-end attenuator can be built up on a small pc board like that shown in Fig. 3. Fixed and trimmer capacitors are mounted on the foil side and the resistors on the other. This board supports the first four attenuator elements. (Two sections on this small board will be unoccupied.) Once the components are mounted on the board, the pc assembly can be wired to the appropriate lugs on *S2* with short lengths of hookup wire.

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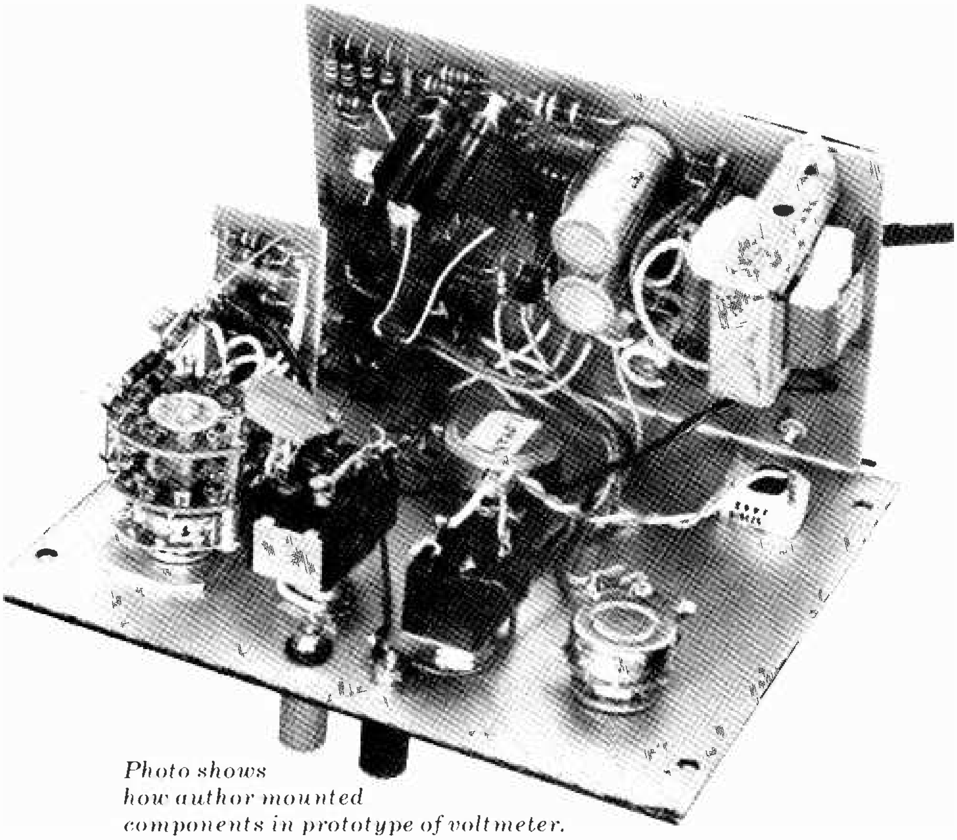

Photo shows how author mounted components in prototype of voltmeter.

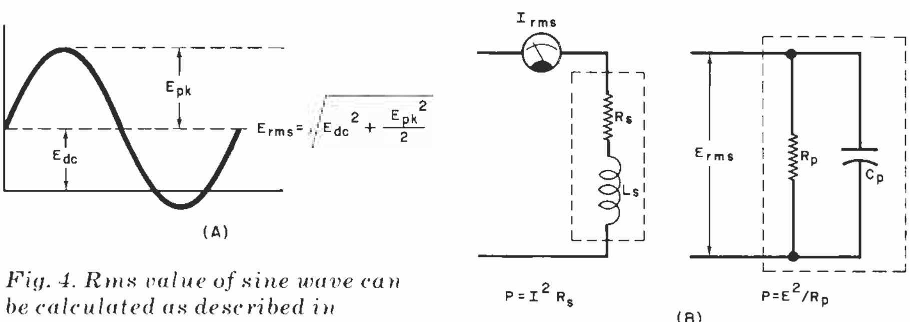

Fig. 4. Rms value of sine wave can be calculated as described in text using diagrams shown here.

The four remaining attenuator sections are resistors that can be mounted directly on *S2*.

As shown in the photograph, on the front panel are the meter, two overpeak LED's, *R17* (NULL), *R15* (ZERO), *S3* (NULL/ZERO), *S1* (AC, DC POWER ON-OFF), *S2* (RANGE) and input jacks *J1* and *J2*.

These components can be mounted as desired on the front panel, leaving enough room around *R17* to accommodate a calibrated dial plate. Press-on type can be used to identify controls.

Calibration. Set the NULL control *(R17)* to its minimum resistance. With *S3* at ZERO, connect *J1* to *J2* and adjust *R15* (ZERO) for no meter deflection (center zero).

Set *S2* to the ×1V position and *S3* to NULL. Apply an accurate 1-volt dc between *J1* and *J2* (positive to *J1*) and adjust the pc board mounted *R19* for a zero meter indication.

Mark the NULL dial plate with a 1, then apply 2, 3, etc., up to 11 volts dc, zeroing the meter with the NULL control and marking the dial as you go. If desired, you can apply a 60-Hz sine wave of known rms value to verify the "equivalent-to-dc" function.

The *S2* trimmer capacitors can be adjusted with the aid of an audio generator. Null the meter to read a 60-Hz sine wave someplace on the ×.1 range.

Without changing the generator output level, raise the frequency to 50 kHz and adjust *C5* for a null at the same spot on the dial. Repeat this for *C4*, *C3* and *C2* on their ranges. Alternately, the trimmer capacitors can be adjusted for the cleanest 10-kHz square-wave response with a scope probe across *I1*.

Use. Here are some guidelines for using the rms meter.

When there is an ac or combined ac and dc voltage across an unvarying (pure) resistance, the rms value of the voltage can be used to compute the average power in the resistance by the equation $P_{AVG} = E_{RMS}^2/R$.

Similarly, if the rms current through an unchanging resistance is known, then $P_{AVG} = I_{RMS}^2 R$.

Note that unlike many rms meters, the meter described in this article can indicate the rms value of a waveform having a dc component. This dc component cannot be simply added to the rms value of the ac component to obtain the rms value of the total waveform. For sine-wave ac components, the rms value can be calculated as illustrated in Fig. 4. For other waveforms, it's either calculus or the dc-coupled rms meter.

An rms measurement cannot be applied to the calculation of power where the load impedance is partly reactive. However, if the numerical value of the resistive part of the load can be determined, true power can be calculated as shown in Fig. B, since the reactive portion consumes no power.

The rms measurement is also not appropriate for calculating the average power delivered to devices having changing ohmic resistance. Examples of such devices are diodes, SCRs, switching transistors and the plate or collector of a class-C amplifier.

For such devices, measure the average voltage and current for the device and calculate: $P_{AVG} = I_{AVG}E_{AVG}$.

If current flows always in one direction, a VOM on the dc range can be used. If ac is involved, you must use a meter that measures the average absolute (without regard to sign) value of voltage and current.

A VOM or average-reading DVM on the ac range fills this function if the readings are multiplied by 0.901 to change the rms calibration to average calibration. Be wary of such meters on the ac ranges since many of them have upper frequency limits below 1 kHz. ◇

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Fun Project

The AUDIO ARTIST Sound-Effects Machine

BY JIM BARBARELLO

You can create any of a number of sounds--from a siren's wail to a clock's tick--to enhance your tape recordings

WHETHER you're an amateur recording engineer, electronic musician, or simply a "sound bug" or chronic knob twiddler, the Audio Artist is sure to appeal to you. It's a special-effects generator which can be used to create such sounds as the wail of a siren, the bubbling splash of a rock falling into a pond, the stock Hollywood sound of a flying saucer, the complex whirring generated by some futuristic machine, and much more. The Audio Artist's five controls interact with each other, resulting in a large variety of possible sound effects.

The project can double as a metronome whose rate is variable from less than 1 Hz to more than 250 Hz. Displaying the output of the Audio Artist on an oscilloscope also creates some interesting effects. The project is easily built, and the total cost of construction is less than $25.

About the Circuit. The Audio Artist employs essentially the same circuit as that of the Cabonga Percussion Synthesizer and its Auto Trigger accessory (POPULAR ELECTRONICS, August and September 1977). It is shown schematically in Fig. 1. A comparison of the two reveals that the Cabonga's manual PITCH control has been replaced with a FET to allow voltage control of the output frequency.

That portion of the circuit built around *IC2B* is the triggering and tone-generating section. Field-effect transistor *Q1* is a voltage-sensitive device whose source-to-drain resistance varies with the magnitude of the voltage applied between its gate and source. The signal applied to the gate of *Q1* is a triangle wave which varies the effective channel resistance of the FET at a rate determined by the setting of potentiometer *R20*. Transistor *Q1*, along with op amp *IC2B*, *R11*, *R12*, and *C4* through *C7*, form a twin-T, active bandpass filter which will generate a damped sinusoidal output each time it is triggered by a positive-going pulse. Damping of the output waveform is determined by the setting of *R10*, and can be varied between the extremes of no output at all and sustained oscillation.

Dual operational amplifiers *IC1* and *IC3* each form oscillators. One (*IC1*) is used to generate trigger pulses which stimulate the active filter into oscillation. The other (*IC3*) produces triangle waves which modulate the channel resistance of *Q1* and hence sweeps the filter. In each oscillator, the noninverting stage (*IC1A* or *IC3A*) acts as a comparator and the inverting stage (*IC1B* or *IC3B*) functions as an integrator. Assuming that the output of the comparator is changing state from V− to V+, the resulting positive voltage step is integrated into a ramp with a positive slope. When the amplitude of the ramp reaches V+/2, the comparator again changes state, generating a negative-going step which is integrated into a ramp with a negative slope. The comparator changes state once more when the amplitude of this ramp reaches V−/2.

This process continues cyclically, producing a square wave at the comparator's output and a triangle wave at the output of the integrator. The slope of the ramp (triangle waveform) determines how quickly the comparator changes state and, consequently, the frequency of oscillation. That slope is determined by the current supplied to *C1* (*C8*) via *R3* and *R4* (*R19* and *R20*). Therefore, the frequency of oscillation is governed by the setting of a single control (*R4* or *R19*) over a range of from 0.5 to more than 250 Hz.

This square-wave output of the tempo generator (*IC1*) is shaped into trigger pulses for active filter *IC2B* by the RC network *R7C2C3* and diodes *D1* and *D2*. Triangle waves generated by *IC3B* are applied to the gate of FET *Q1* via DEPTH control *R18* and *R15*, causing *IC2B* to produce a constantly changing pitch. The two generators (*IC1* and *IC3*) oscillate independently of each other, and can thus be adjusted to beat, to run asynchronously, or to run synchronously for different effects. The project's controls can be adjusted to produce some

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very unusual sounds, in addition to a damped, repetitive sine wave whose frequency varies pseudorandomly.

Signals generated by *IC2B* are buffered by *IC2A*, a unity-gain inverting amplifier, and are presented to output jack *J1* for further amplification or recording. The output signals are of line level and should not be applied to microphone or other weak-signal inputs. The bipolar voltages required by the project's op amps can be furnished by either a line-powered supply or batteries. The author's prototype employs batteries for portability. Total current demand is relatively modest, making the use of a battery supply a practical alternative to a line-powered one.

Construction. The Audio Artist can be assembled using either a perforated or a printed-circuit board (Fig. 2). When assembling the circuit board, be sure to employ the minimum amount of heat and solder consistent with the formation of good solder joints. Take care to observe the polarities of electrolytic capacitors and the pin basings of semiconductors. Mounting the ICs in sockets or Molex Soldercons is recommended.

The project's circuit board can be housed in any suitable enclosure. One measuring 6½" × 3-¾" × 2" (15.9 × 9.5 × 5.1 cm) will provide adequate room for the circuit board, a battery power supply, and the various controls. Mount the board in the enclosure using standoffs and machine hardware. Similarly, install the potentiometers, power switch, and output jack using the hardware supplied with these components. Secure the batteries (if used) to the interior of the enclosure with brackets.

Label the various control positions using dry-transfer lettering. Once the con-

Photo of author's prototype shows pots on front and pc board at rear.

trols, switch and jack have been mounted and identified, interconnect them with the project's circuit board using suitable lengths of flexible hookup wire. Be sure

(continued on page 69)

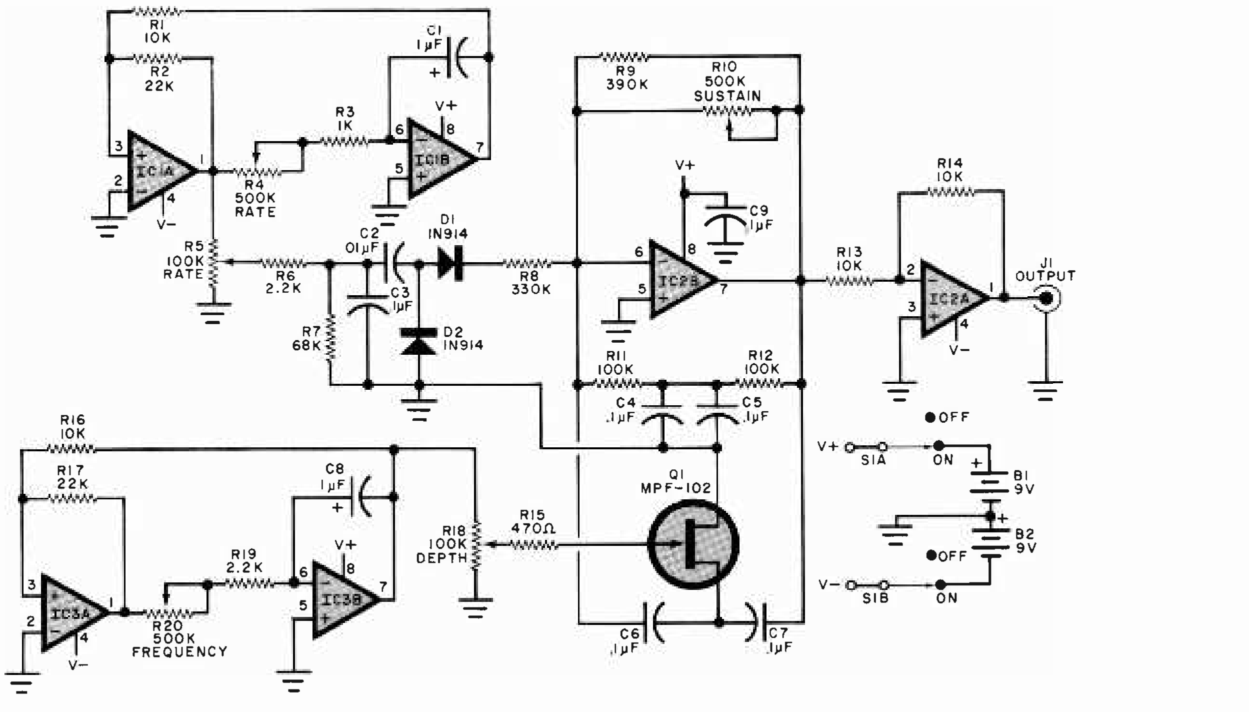

Fig. 1. The circuit around IC2B is the tone-generating section. The five controls react with each other to provide various sound effects.

PARTS LIST

B1, B2—9-volt battery
C1, C8—1-µF, 16-volt upright electrolytic
C2—0.01-µF disc ceramic capacitor
C3 through C7, C9—0.1µF disc ceramic capacitor
D1, D2—1N914 or 1N4148
IC1, IC2, IC3—MC1458N dual op amp
J1—phono jack
Q1—MPF-102 n-channel JFET
The following are ¼-watt, 10% tolerance, carbon-composition resistors unless otherwise noted:
R1, R13, R14, R16—10,000 ohms
R2, R17—22,000 ohms
R3—1000 ohms
R4, R20—500,000-ohm audio-taper pot.
R5, R18—100,000-ohm linear-taper pot.
R6, R19—2200 ohms
R7—68,000 ohms
R8—330,000 ohms
R9—390,000 ohms
R10—500,000-ohm linear-taper pot.
R11, R12—100,000 ohms
R15—470 ohms
S1—Dpdt switch
Misc.—Suitable enclosure, printed circuit or perforated board, IC sockets or Molex Soldercons, battery clips, battery holders, dry-transfer lettering, control knobs, hookup wire, machine hardware, solder, etc.

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Dial your scores into a two-player, double-digit scoreboard

GAMES in which the scores for individual players must be kept are a popular pastime. Not so popular is the usual search for paper and pencil needed for keeping the score. The Electronic Scorekeeper described here eliminates the search so you can get right to the game. As designed, the Scorekeeper can keep score for two players up to a maximum count of 99. However, with a couple of simple modifications, the number of players and the count range can be increased as desired. The circuit uses readily available and inexpensive TTL devices and seven-segment numeric LED displays.

About the Circuit. Since the circuit for each player is identical, only the circuit for player A is shown in Fig. 1. Player B's circuit connects to the pin-6 output of gate *IC1B*. Integrated circuits *IC4* and *IC5* and display *DIS2* make up a conventional 0-to-9 units decade counter whose carry output at pin 8 of *IC4* is fed to a similar tens counter made up of *IC2*, *IC3*, and *DIS1*. Seven-segment displays *DIS1* and *DIS2* are common-anode LED types.

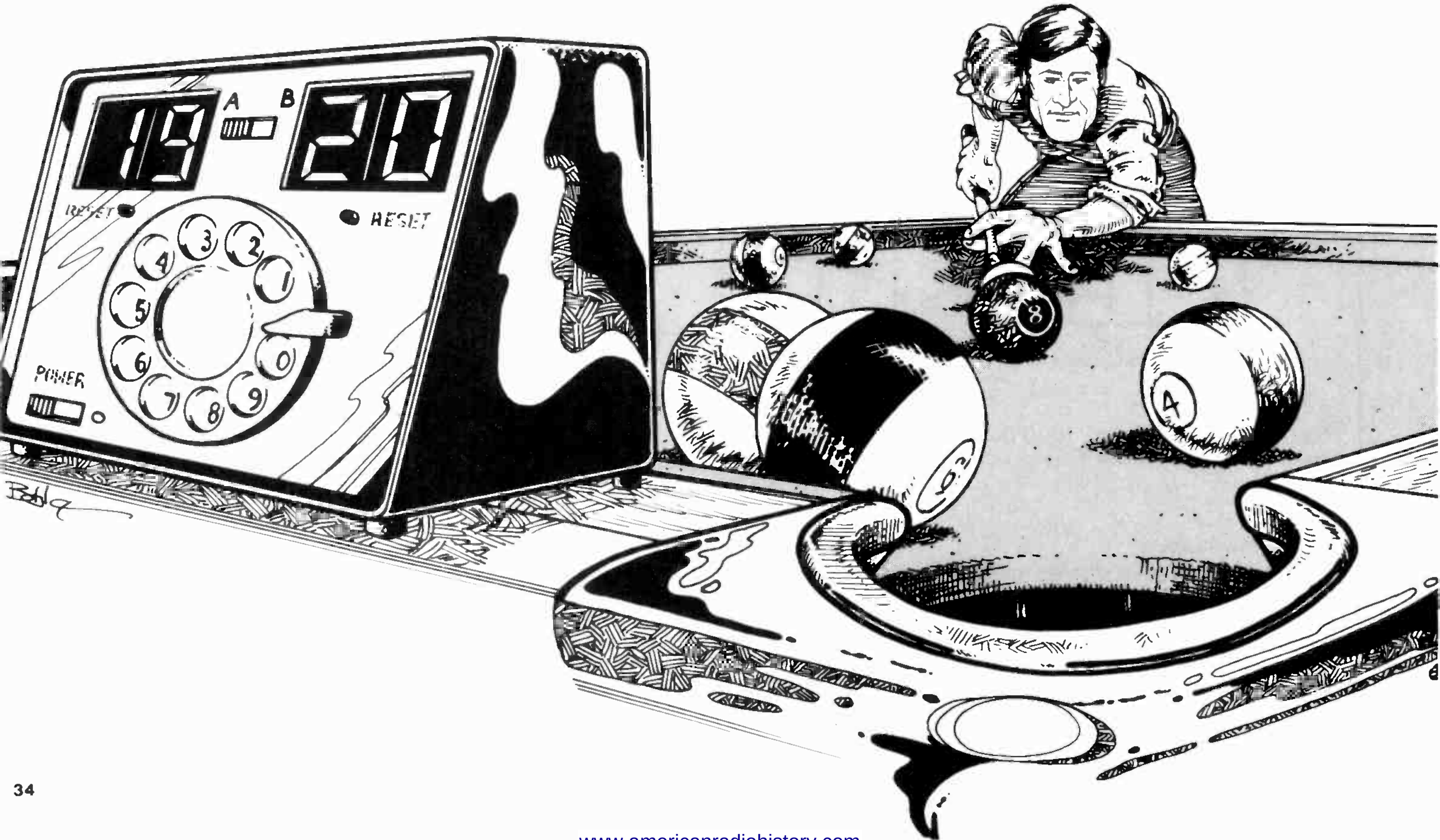
The count for the circuit shown in Fig. 1 can easily be increased as desired simply by adding extra decade counters. When the additional decade counters are used, the input of each successive counter is connected to the carry output of the preceding counter and the RESET lines are connected in common.

Both decade counters shown in Fig. 1 are set to zero by operating CLEAR pushbutton switch *S2* to momentarily raise the reset-to-zero (RST) input at pin 3 of *IC2* and *IC4* to high and then back to ground as the switch is released and pulldown is accomplished by *R2*. When *S2* is pressed and released, both *DIS1* and *DIS2* should display zeros.

SELECTOR switch *S1* permits the person keeping score to choose between player A and player B for score display

Electronic Scorekeeper for Recreation Rooms

BY JOSEPH FORTUNA



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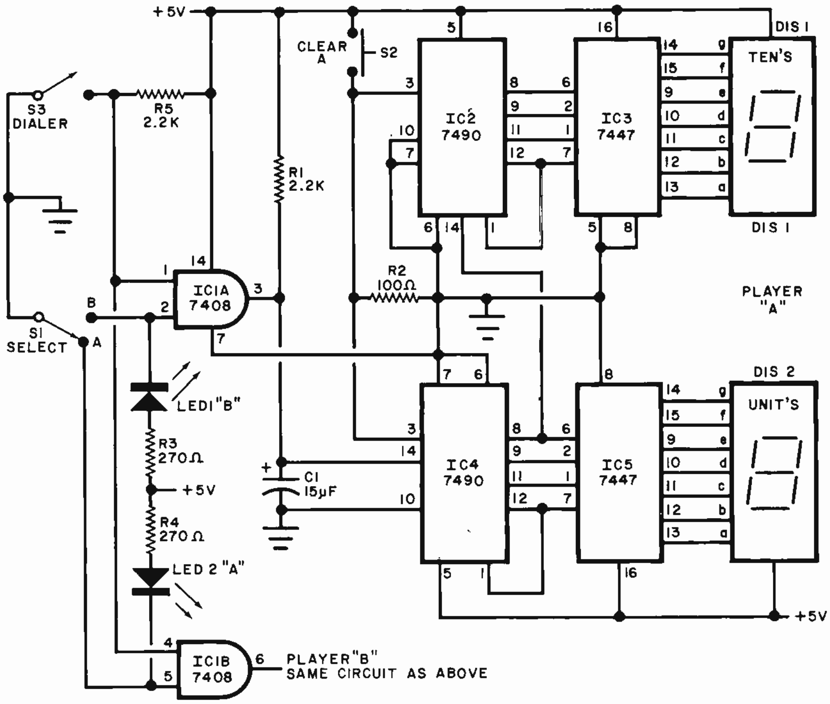


PARTS LIST

C1—15-µF, 15-volt electrolytic
DIS1, DIS2—Common-anode 7-segment LED display
IC1—7408 quad AND gate
IC2, IC4—7490 decade counter
IC3, IC5—7447 BCD-to-7-segment decoder
LED1, LED2—Any discrete red LED
R1, R5—2200-ohm, ½-watt resistor
R2—100-ohm, ½-watt resistor
R3, R4—270-ohm, ½-watt resistor
S1—Dpst switch
S2—Normally open pushbutton switch
S3—Telephone dialer-switch mechanism (see text)
Misc.—Duplicate circuit for player B; regulated 5-volt, 1-ampere dc power supply; perforated or printed-circuit board and hardware; suitable enclosure; sockets for ICs (optional); machine hardware; hookup wire; solder; dry-transfer lettering kit; etc.

Fig. 1. Schematic shows scorekeeping circuit for only one player.

and incrementing. When the player-A position is selected, pin 5 of *IC1B* is grounded and held low, causing *LED2* for player A to come on. At this time, the output of *IC1B* is low and the gate is disabled. Hence, the player-B decade counters will not operate.

Pin 1 of *IC1A* and pin 4 of *IC1B* are made high by pullup resistor *R5*, and mechanical DIALER switch *S3* is connected from ground to this common point. (A surplus mechanical telephone-dial switch assembly can be used for *S3* to allow you to conveniently "dial in" the score updates. Alternatively, you can substitute an ordinary normally open pushbutton switch for this operation, but it will have to be operated for each and every unit increment in the scoring.)

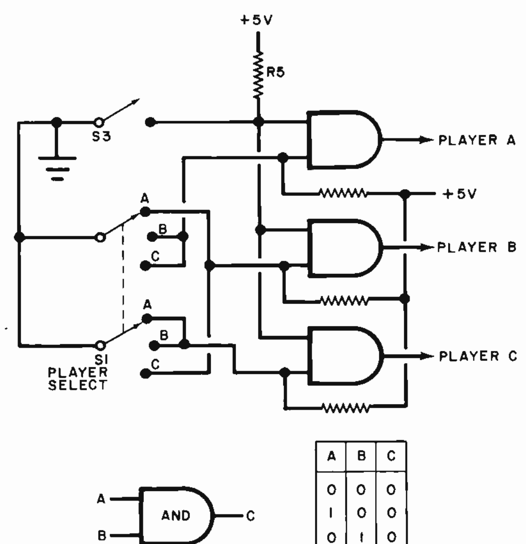
Operating *S3* shorts the common *IC1A* pin-1 *IC1B* pin-4 point to ground the same number of times selected on the DIALER. As the DIALER is operated, IC1A turns on and off with each closure of *S3*. This generates one or more input pulses, depending on the DIALER number selected, for player A's decade counter. (This assumes *S1* is set to A; operation is identical for player B, except that *S1* must be set to B.) Every time the *IC4* units decade overflows at the tenth pulse from *IC1A*, the carry output from *IC4* toggles the *IC2* decade counter.

The circuit in Fig. 1 can be expanded to keep score for more than two players, as shown in Fig. 2. Note here that separate player LEDs are not used. Using the AND gate and truth table shown, you can design further switching to increase the number of players beyond the three shown in Fig. 2.

Construction. Since component layout is not critical, you can use just about any wiring technique that suits you. Perhaps most convenient is a printed-circuit board of your own design, but perforated board and Wire Wrapping is equally suitable. In either case, it is recommended that you use sockets for the ICs.

Once you have assembled and checked the circuit, mount it in an enclosure so that the two pairs of displays can easily be viewed. Mount the LEDs and switches, including the DIALER mechanism, on the top of the enclosure. Finally, use a dry-transfer lettering kit to label the switches and LEDs according to function.

Power for the Scorekeeper can be obtained from any regulated 5-volt dc supply capable of delivering 1 ampere or more of current. ◇



A	B	C
0	0	0
1	0	0
0	1	0
1	1	1

Fig. 2. Circuit for over two players.

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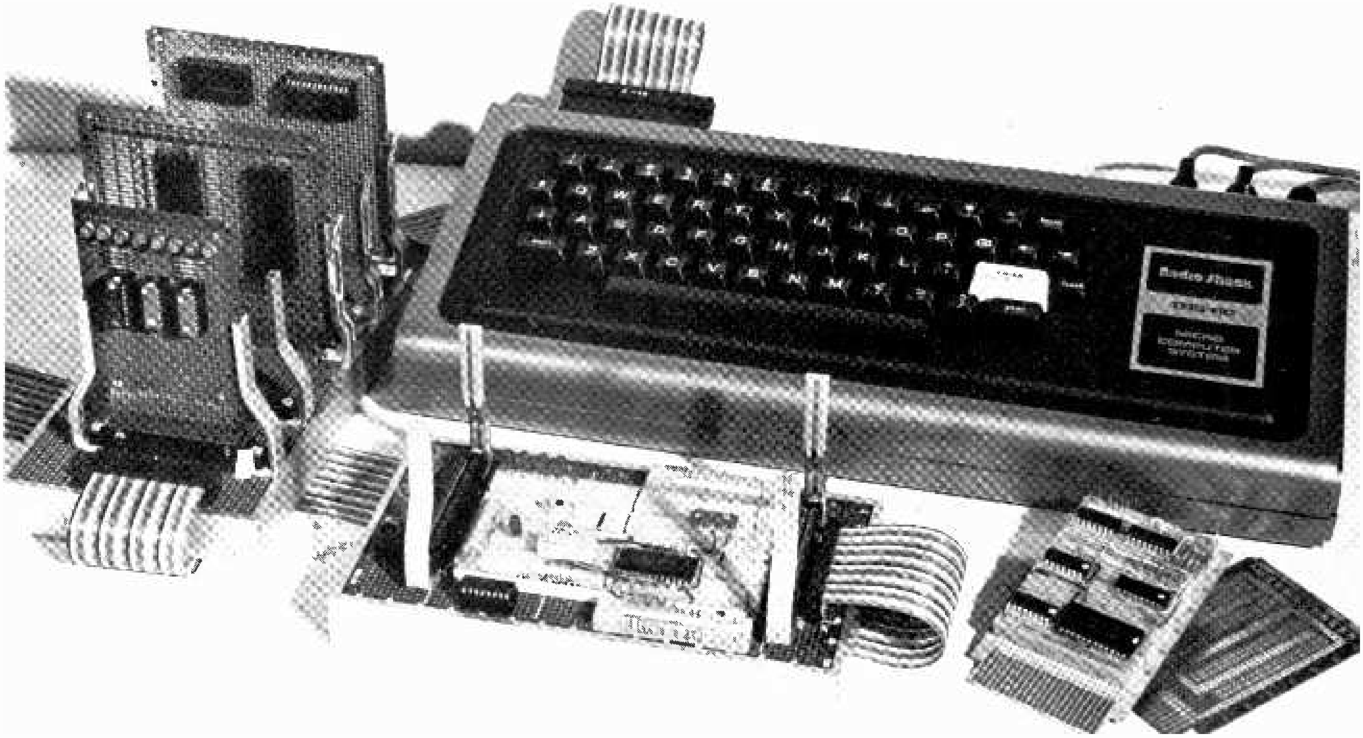

How to Add I/O Ports to Microcomputers

The basics of computer port operation and instructions for using them to expand computer flexibility

BY ADOLPH A. MANGIERI

FOR A microcomputer to "do something" truly useful, it must have input and output ports. The I/O ports make it possible for the computer to "interface" with practical devices—relays for appliance control, switches (or a keyboard) for feeding in desired commands, keyboard and video or hard-copy terminals for communicating with the computer, etc. Though 8080- and Z80-based micros can control up to 256 I/O ports, few are equipped with more than two. In this article, therefore, we will describe how to add I/O facilities to expand a Z80 or 8080 computer's flexibility.

To add the I/O ports described here to any Z80 or 8080 micro, you must have a basic familiarity with port operation and addressing and bus structure. (This information is detailed in manuals that accompany the computers.) A few ICs will get your computer up and running. Port examples presented here are for a Radio Shack TRS-80 Level I computer that uses the T-BUG monitor and a Level II computer with machine code and BASIC. You can use a solderless breadboard to perform experiments and to prototype circuits.

Port Basics. There are a number of different types of I/O ports in use. An elementary port may simply display information on a bank of LEDs, operate relays, or input data from a bank of switches. A complex port, on the other hand, can accommodate such sophisticated devices as an ASCII keyboard, full-graphics CRT monitor, and hard-copy terminal. Although all ports share the common computer bus, each is assigned a specific address and is provided with logic circuitry that enables the port only when it is addressed.

Machine-code instructions define CPU input and output operations. Two-byte instruction D3 XX initiates an output operation to a port. (D3 is the output instruction and the Xs indicate numbers for specific port addresses, such as D3 00, D3 01, D3 02, etc.) When a Z80 CPU fetches and executes this instruction, it generates an $\overline{IOREQ}$ (I/O request) pulse and a $\overline{WR}$ (write) pulse, both active low, as indicated by the lines above them. These are logically added in an external AND gate and delivered as the $\overline{OUT}$ pulse on pin 21 of the TRS-80's bus. The data byte in the CPU accumulator register is placed on data bus lines D0 through D7. Simultaneously, address byte XX is placed on address lines A0 through A7.

Port-select logic constantly examines the $\overline{OUT}$ and address lines, waiting for the simultaneous appearance of the $\overline{OUT}$ pulse and port address. When this occurs, the port is enabled and data on the data bus lines enters the port. Ad-

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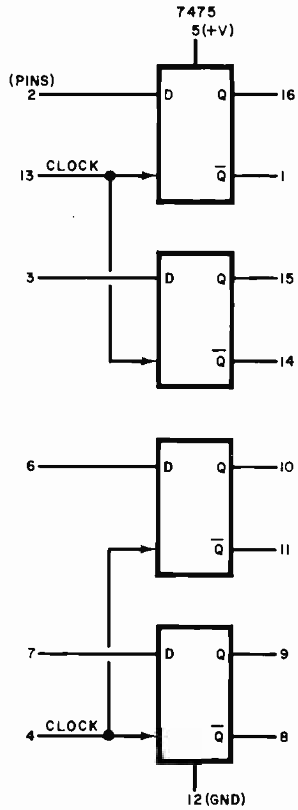

TRUTH TABLE

CLK	D	Q EQUALS
0	X	LATCH
1	0	0
1	1	1

X= DON'T CARE STATE

Fig. 1. Internal logic of 7475 (including pin-out) and truth table for each latch.

dress bytes can be from 00 to FF (hexidecimal) to allow up to 256 output ports to be used with suitable decoding.

Machine-code instructions DB XX (input instruction with port address) initiates an input operation from the selected input port. Here, $\overline{IOREQ}$ and $\overline{RD}$ (read) pulses are added in an external AND gate and delivered as the $\overline{IN}$ pulse on pin 19 of the TRS-80's bus.

Port logic detects the simultaneous appearance of the $\overline{IN}$ pulse and port address and enables the port. At this time, the port connects its output lines to the data bus and the CPU copies any data present on the bus into the accumulator register. After data acceptance, the port frees the data bus for other purposes. The accumulator register is the source and destination of data with the D3 and DB instructions. The Z80 instruction set includes a number of special I/O instructions that effect data transfers to and from other registers and memory, with some instructions allowing movement of data in blocks.

Output and input ports can have the same address, such as output D301 and input DB 01. Port-select logic differentiates between the two by $\overline{OUT}$ and $\overline{IN}$ pulses. "Standard" or "isolated" I/O addressing allows up to 256 input and 256 output ports to be addressed by the computer. This is ample for just about any imaginable home computer system.

An alternative form of port addressing employs memory-mapped I/O. Each port, in effect, is addressed as memory. This method allows thousands of ports to be addressed and affords some programming advantages.

Simple Output Port. Inexpensive 7475 TTL ICs can be used to make 2-, 4-, and 8-bit latching-type output ports. As shown in the truth table in Fig. 1, data latch output Q follows input data D as long as the clock (CLK) line is high. When the CLK line goes low, data D is latched to output Q. The internal logic of the 7475, including pinout, is shown in Fig. 1. Note that each clock line drives two latches.

Two 7475's can be connected as an 8-bit latching port (Fig. 2). LEDs connected to the $\overline{Q}$ outputs turn on when their respective data D input is high.

The port shown in Fig. 2 is addressed by instruction D3 00, which places binary 00000000 on lines A0 through A7. The least-significant bit is on A0. When $\overline{OUT}$ and $\overline{A0}$ are true, the port is enabled by *IC3A* and the data byte held in the CPU accumulator register is displayed in binary on the LEDs. Address line bits A1 through A7 are "don't cares," provided they are not assigned to other ports. Hence, instruction D3 FE also selects this port because bit A0 is low.

Ports that do not require all eight data bits are easily arranged. To set up two 4-bit ports, break the clock line at X and add the second NOR gate (Fig. 2). Port *IC1* is enabled by instruction D3 01, which places binary 00000001 on the address lines. The 0 bit on line $\overline{A1}$ causes selection of port *IC1*. You can also separate the four clock lines and arrange four 2-bit ports, using address lines A2 and A3 for port selection.

Progressive addressing allows up to eight input and output ports to be used. One or more ports can be enabled by one instruction, simplifying programming and hardware requirements. Although it requires additional ICs in each port, full decoding of the address bits allows up to 256 input and output ports. For example, the TRS-80 cassette port is fully decoded and selected by instruction D3 FF. For this and other reasons, the ports described here are assigned active-low address bits for selection. For an elementary example of both fully decoded port and memory-mapped port, refer to the TRS-80 Technical Reference Handbook.

Complex I/O Port. Intel's versatile 8212 I/O chip can be used as either a latching or a nonlatching output port, input port, gated bus driver, or straight-

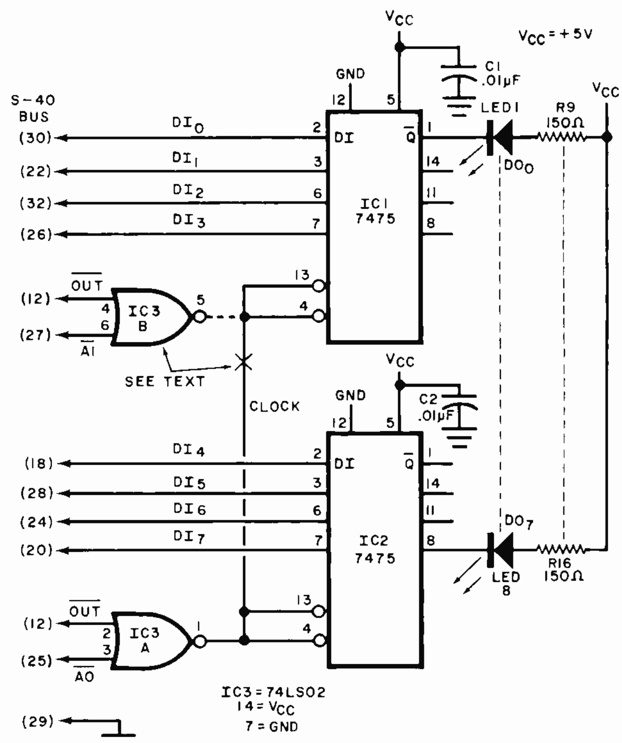

Fig. 2. Two 7475s and a 74LS02 can be used to form an 8-bit port or, with slight modifications, two 4-bit or four 2-bit ports.

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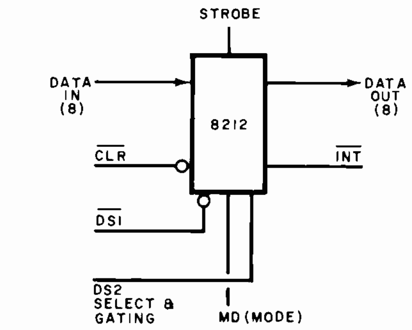

STATE	STB	MD	($\overline{DS1}$ · DS2)	D OUT EQUALS
1	0	0	0	3 STATE
2	1	0	0	3-STATE
3	0	1	0	LATCH
4	1	1	0	LATCH
5	0	0	1	LATCH
6	1	0	1	DATA IN
7	0	1	1	DATA IN
8	1	1	1	DATA IN

Fig. 3. The 8212, an 8-bit I/O port, and its truth table. This IC has tri-state provisions. That is, in one mode it can be electrically isolated from the system bus.

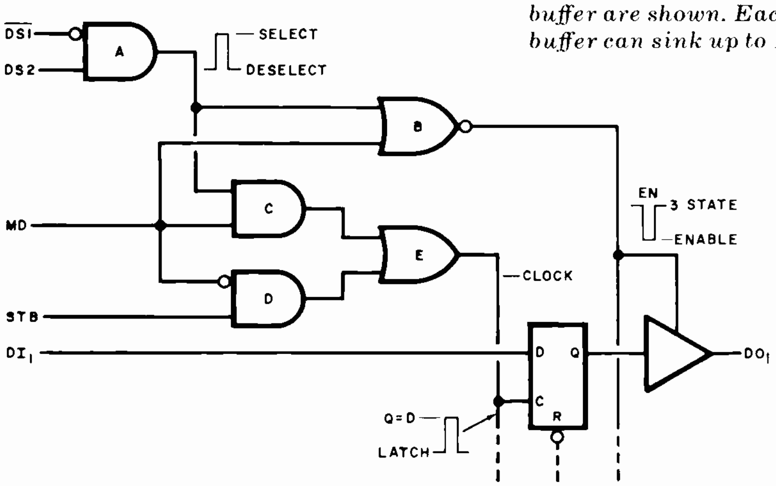

Fig. 4. A portion of the internal control logic of the 8212. Only one of its eight latches and output buffer are shown. Each output buffer can sink up to 15 mA.

through buffer, to name just a few of its applications. This high-speed Schottky TTL device includes eight data latches and output buffers that can be tri-stated (switched to high impedance). Each buffer can sink up to 15 mA.

The function diagram of the 8212 is shown in Fig. 3, which also illustrates chip signals and the IC's truth table. A portion of the internal control logic and one of the latches and its output buffer are illustrated in Fig. 4. For simplicity, CPU-interrupt control logic, which controls interrupt output INT, is omitted.

Familiarity with the control logic simplifies application. Mode control line MD is tied low (logic 0 or ground) for the input-port mode and high (logic 1 or V_{CC}) for the output-port mode. Lines $\overline{DS1}$ and DS2 are the device-select, or gating-control, lines. When $\overline{DS1}$·DS2 is 1, the device is selected by a high at the output of gate A.

The data-latch clock is strobed two ways. When line MD is low, gate C is defeated and strobe line STB passes a pulse through gates D and E to the clock line. When line MD is high, gate D is defeated and gate A passes a pulse through gates C and E to the clock line. Similarly, the output buffers are also operated two ways. When line MD is high, gate B goes low and enables the buffers continuously. This is a necessary requirement for a latching-output port. When line MD is low, the selected pulse from gate A passes through gate B to enable the buffers briefly, after which they return to tri-state. This is a necessary requirement for an input port.

The truth table is simple to use if you keep in mind the port or application requirements. To illustrate, let us implement an input port. In this case, MD must be tied low (grounded). Strike out all rows or states listing MD as 1 in the Fig. 3 truth table. Since MD is assigned, strike out column MD. We know from port basics that the input port must not latch onto the data bus. This eliminates state 5 and all that remain are states 1, 2, and 6, as shown in the reduced truth table in Fig. 5.

FIRST REDUCED TRUTH TABLE

STATE	STB	$\overline{DS1}$ · DS2	D OUT EQUALS
1	0	0	3-STATE
2	1	0	3-STATE
6	1	1	DATA IN

FINAL TRUTH TABLE

STATE	STB	$\overline{DS1}$ · DS2	D OUT EQUALS
1	0	0	3-STATE
6	1	1	DATA IN

Fig. 5. Reduction of the truth table for the 8212.

Clearly, state 6 must be retained for device selection and data transfer. Recalling that the STB line must be used to strobe the latches when line MD is 0, state 2 is deleted so that STB can alternate between 1 and 0. This results in the final truth table shown. Check this truth table to be sure it accomplishes the application's requirements. In this case, state 6 enables the port, placing port data on the data bus. State 1 "deselects" the port and tri-states the output buffers as required.

In the final step, computer pulses are assigned to $\overline{DS1}$, DS2, and STB. Available computer pulses are $\overline{IN}$ and $\overline{A0}$, the latter assigned to this port and active low. Notice that STB and DS2 are active high. With $\overline{DS1}$ active low, connect $\overline{IN}$ to $\overline{DS1}$. Pass $\overline{A0}$ through inverter *IC6A* and then to both STB and DS2 (Fig. 6).

To use the 8212 as a latching output port, tie MD high (to V_{CC}). Port requirements include device select with data in (state 7 or 8) and device deselect with latching (state 3 or 4). STB is a "don't care" line. Connect computer output pulse $\overline{OUT}$ to line $\overline{DS1}$. Address line bit $\overline{A0}$ is inverted by *IC7A* and connected to line DS2 (Fig. 6). The LED is off when D is high. If this is objectionable, add inverting buffers between port outputs and light-emitting diodes.

For the 8212 to serve as a straight-through buffer or line driver, requirements are device select, data out equals data in, and continuously enabled output buffers. State 8 will effect these requirements. Connect line MD and DS2 to V_{CC} and line $\overline{DS1}$ to ground. For use as a bidirectional bus driver, interrupting ports, etc., see the Intel 8080 User's Manual.

Computer Hookup. The TRS-80 accepts a special 40-contact card edge connector. However, you can substitute a standard 44-contact card socket, such as a Vector No. R644-2, after modifying it. To do this, fit a thin piece of hard plastic into the connector slot to cover the two top and bottom contacts at one end of the connector. You now have a 44-contact connector that for all practical purposes has been modified to serve as a 40-contact connector.

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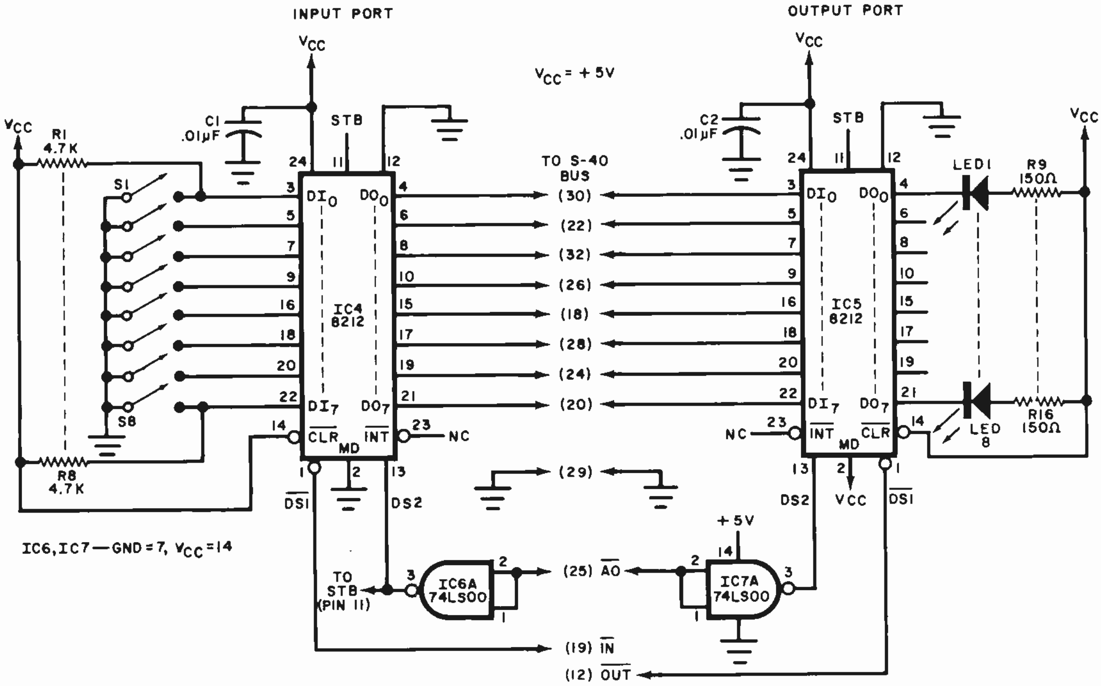

Fig. 6. An experimental I/O port using a pair of 8212s with switch inputs and light-emitting diode outputs.

PARTS LIST

C1,C2—0.01-µF disc capacitor
IC1,IC2—7475
IC3—74LS02 NOR gate
IC4,IC5—8212 8-bit I/O port
IC6,IC7—74LS00 low-power quad 2-input NAND gate
LED1 through LED8—Light emitting diode
R1 through R8—4700-ohm resistor
R9 through R16—150-ohm resistor
S1 through S8—Spst switch

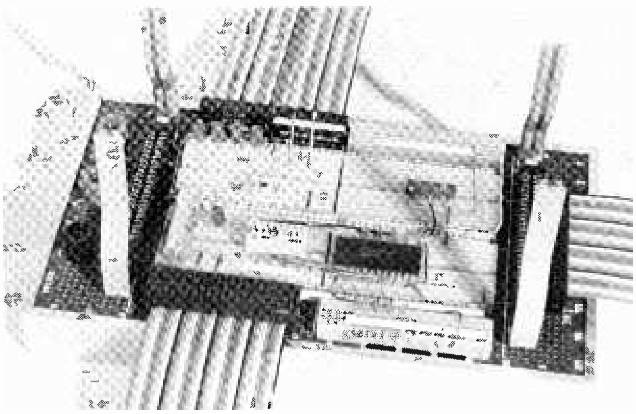

Fig. 7. Three computer links are provided on this patchboard using solderless circuit connections. Wire-Wrapping is used on connectors and under chassis.

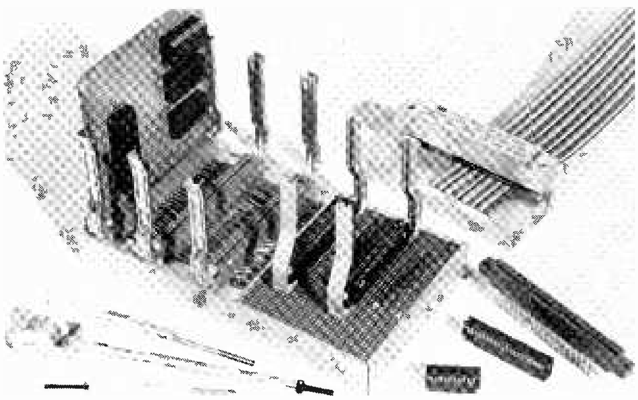

Fig. 8. Typical plug-in card chassis. Three large sockets are Vector Electronics R644-3 with mating BR27-D card guides. Smaller connectors are R644-2 with BR27 card guides. Wire-forming and chiseling tools are also shown.

Install the modified connector in tne TRS-80, making sure that lateral play barely exceeds 1/64" (0.4 mm). Carefully remove the modified connector and cement the small plastic pieces solidly in place. Recheck connector fit before the cement sets. Then use a lettering kit to mark an UP label on the up side of the connector.

Solder a 12" to 18" (30.5 to 45.7 cm) length of color-coded 40-conductor ribbon cable, such as Vector's No. KW2-40, to the connector. Make a record of which conductor connects to and what signals are present on each pin. Refer to the TRS-80 Technical Reference Manual or User's Manual for pin assignments.

At this point, you can choose any of a number of conventional construction approaches. Perhaps the simplest is to use a solderless breadboard on which to experiment with the I/O port. A 40-pin IDC connector, such as a Vector No. KS2-40, can be fitted to the end of the cable, using a Vector No. P187 IDC fixture to make the connection. A mating connector can then be mounted on the solderless breadboard. Shown in Fig. 7 is this author's experimental setup, which includes wiring to a home-built card cage (Fig. 8) to support Wire Wrap circuit cards.

In Conclusion. From the foregoing, you can see that it is relatively simple to interface a computer with external devices to perform useful operations. You could conceivably use all 256 I/O ports to control everything in your home. ◇

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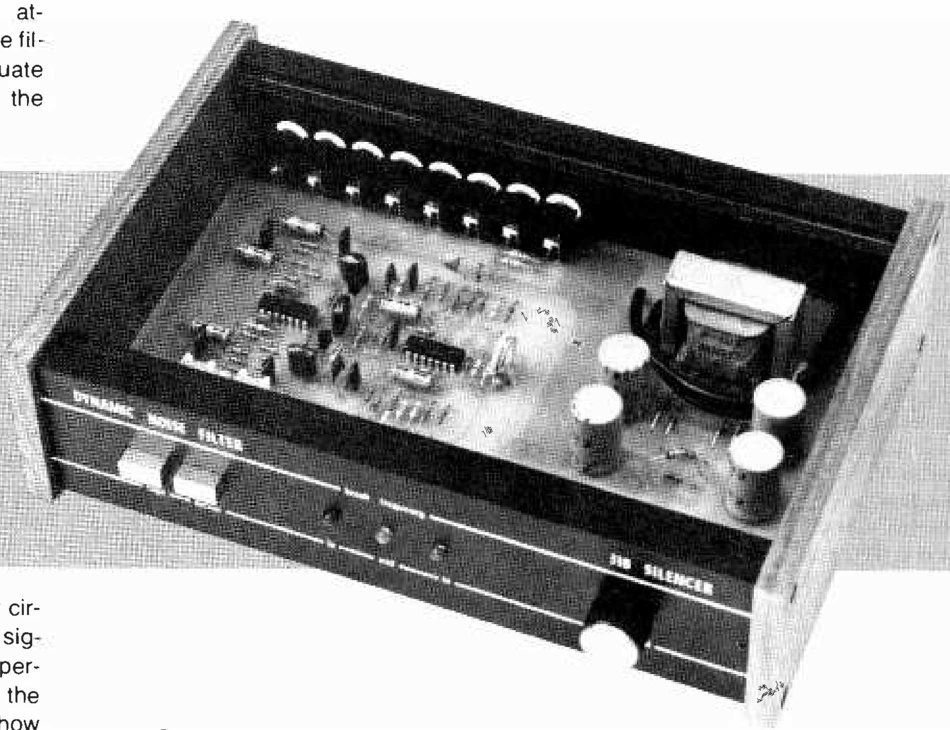
BY COLLEEN McNEICE AND ROGER COTA

Build a DYNAMIC AUDIO NOISE FILTER

Cleans up radio, tape and record signals from any stereo system

THE "SILENCER" dynamic noise filter described here can eliminate tape hiss, record-surface noise, and atmospheric radio noise. Consequently, it is an ideal add-on device for stereo hi-fi systems. Moreover, it does not require encoding and decoding.

The device is essentially a voltage-controlled low-pass filter whose cutoff or break frequency is continually changing to accommodate program material and shut out any detracting noise. It only filters when noise and hiss are audible, when program material is at a low level or absent. The phenomenon of masking is utilized. That is, high-level signals mask noise that would be objectionable if program material level were low. When such masking occurs, the whole signal is passed. When there is no masking by program material, however, the filter extends the bandwidth only as far as required by the music. Beyond this, the high-frequency noise is attenuated. The frequency at which the filter begins rolling off to attenuate high-frequency noise is called the "break frequency."



About The Circuit. The silencer circuit constantly analyzes incoming signals for amplitude, frequency, and persistence. These factors determine the bandwidth at any instant, as well as how quickly the variable low-pass filter changes. Attack and release times vary with the music, thus eliminating a "pump and wheeze" effect of noise modulation.

The device has a continuously variable threshold control, with front-panel LEDs calibrated to indicate "Low," "Mid," and "High" break frequencies. The filter's break frequencies vary between 1.5 and 20 kHz with a roll-off slope of 9 dB/octave (maximum). The Silencer is a single-ended stereo device, making it ideal for use with tapes, records, and tuners for playback and record purposes.

The unit connects either in the auxili-

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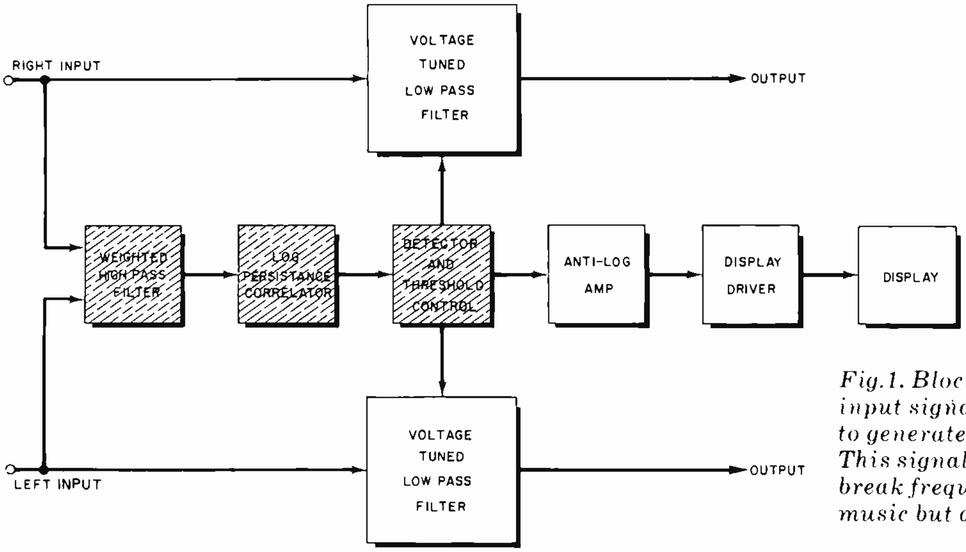

Fig. 1. Block diagram shows how input signals are processed to generate a control signal. This signal governs filter's break frequencies to pass music but attenuate noise.

ary mode or in the tape loop of your audio amplifier. On the back panel are IN and OUT jacks for the tape loop; the front panel also has a TAPE monitor button, and a system DEFEAT.

The block diagram of Fig. 1 shows the functions of the dynamic noise filter. The voltage-controlled low-pass filter is composed of *IC1A* and *IC1B*, as shown in the schematic of Fig. 2. (The components to the left of the dashed line make up one stereo channel; only one is shown in the schematic for clarity.) The gain of op amp *IC1A* is approximately *R3/R5*. At low frequencies, the capacitive reactance of capacitors *C4* and *C5* is very high, making the output of *IC1B* look like a low impedance source. The gain of *IC1A* is then:

$$A = R3/R5 = 10{,}000 \text{ ohms}/1000 \text{ ohms} = 10$$

At higher frequencies, however, the impedance of *C4* and *C5* decreases; *IC1B* generates an output and bootstraps *R5*. This bootstrapping effect causes *R5* to look larger. Therefore, gain A becomes smaller and the filter attenuates the high-frequency energy.

To vary the breakpoint of the filter, FET *Q1* has the ability to shunt the signal at the non-inverting input of *IC1B* to ground. Figure 3A shows the filter with the FET open and the high frequencies attenuated, while 3B illustrates the filter's action with the FET shorting the signal to ground. The control signal applied at the gate of the FET allows the bandwidth of the low-pass filter to be self-adjusting for any frequency. This allows high-frequency signals and subtle harmonics of fundamental bass frequencies to be passed, while unmasked noise is attenuated.

The circuits represented by the shaded blocks of Fig. 1 are the dynamic analytical controls. They automatically judge the program material, adjust the bandwidth to accommodate it, and change the attack and release times to maximize the masking effect and minimize noise modulation. The control signal is applied to the gate of *Q1*. It's determined by the (1) spectral content, (2) amplitude, and (3) persistence of the incoming signal.

The spectral content is sensed by the high pass weighting filter, a network made up by *R8, R29, R30, R31, C6, C17,* and *IC2A*. This network is driven by the output of *IC1B*, which actually determines the quiescent operating point of the low-pass filter. Amplitude is determined by threshold control, *R27*, a 100K-ohm front-panel potentiometer. This pot sets the voltage divider for the positive input to *IC2A*, and the dc level for *IC2A*'s output. The dc output level determines the quiescent operating point of the FET. The dynamic operation of the FET is adjusted by the ac control signal, allowing it to follow the program material. The ac component of *IC2A*'s output is determined by sensing the signal's amplitude on the output of *IC1B*.

The persistence log amp is formed by *R33, D2,* and *C20*. It checks the correlation coefficient of the signal, and adjusts the attack and release time of the low-pass filter to minimize any noise modulation problems. Variable attack and release times allow for the most effective masking of the noise.

The anti-log amplifier *IC2B* also senses the control voltage output of *IC2A*. This signal is then rectified and filtered by *D4* and *C21*, and is then used to drive threshold comparators *IC2C* and *IC2D*. These amps drive the logic network of *D5, D6,* and *D7*, which drives the display. The 10K-ohm trimpot, *R37*, is used to calibrate the LEDs. The red LED indicates a break frequency of 1.5 kHz, the yellow, a break frequency between 1.5 and 10 kHz, and the green that the filter is opening up above 10 kHz.

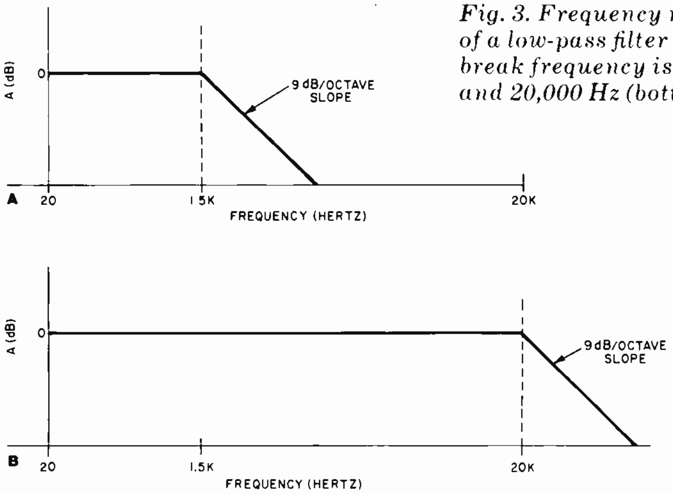

Fig. 3. Frequency response of a low-pass filter when its break frequency is 1500 Hz (top) and 20,000 Hz (bottom).

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PARTS LIST

C1, C3, C9, C11, C21—1-μF 50-volt axial-lead electrolytic capacitors
C2, C10—680-pF disk ceramic capacitors
The following are 100-volt Mylar capacitors:
C4, C5, C12, C13—.022-μF
C6, C14, C17—.001-μF
C7, C15, C20—.1-μF
C8, C16—.01-μF
C18, C22, C23, C24—1000-μF 35-volt radial-lead electrolytic
D1—33-volt Zener diode
D2 through D7—IN914 signal diode
D8 through D11—IN4002 rectifier
F1—½-ampere fuse
IC1, IC2—μA4136 quad op amp (Fairchild)
J1-J8—RCA phono jacks
LED1—Red (Fairchild FLV 110 or equivalent)
LED2—Yellow (Fairchild FLV 410 or equivalent)
LED3—Green (Fairchild FLV 310 or equiv.)
Q1, Q2—Matched pair of 2N5458 JFETs.
The following are ¼-watt, 5% tolerance resistors:
R1, R13, R26—47,000 ohms
R2, R14, R35, R43, R44—4700 ohms
R3, R15, R33, R38—10,000 ohms
R4, R16—100 ohms
R5, R17—1000 ohms
R6, R18—39,000 ohms
R7, R19, R45, R46—2200 ohms
R8, R20, R29—15,000 ohms
R9, R21—11,000 ohms
R10, R22, R36, R39, R41—100,000 ohms
R11, R12, R23, R24, R34—1 megohm
R25—22,000 ohms
R28, R32—470,000 ohms
R30—680,000 ohms
R31—130,000 ohms
R40, R42—27,000 ohms
Other resistors and controls:
R27—100,000-ohm potentiometer with switch (CTS FR-GC-XM 450 or similar)
R37—10,000 ohm thumbwheel trimpot
R47, R48—10 ohms, ½ watt, 5% tolerance resistor
S1, S2—DPDT switches
S3—110-V, 2-A switch (part of R27)
T1—22-volt center-tapped, 50-mA transformer
Misc—Ac line cord, knob for threshold pot, buttons for switches, suitable enclosure, hardware, hookup wire, solder, etc.
Note—The following items are available from Video Control, 3314 "H" St., Vancouver, WA 98663 (Tel. 1-206-693-3834): Complete 318 Silencer kit, including 6063 extruded aluminum chassis and hand-finished black walnut end pieces, $159.00. Also available separately: Etched and drilled circuit board, $16.00; individuality tested and matched 2N5458 FETs. $4.50 Washington state residents please add 5% sales tax.

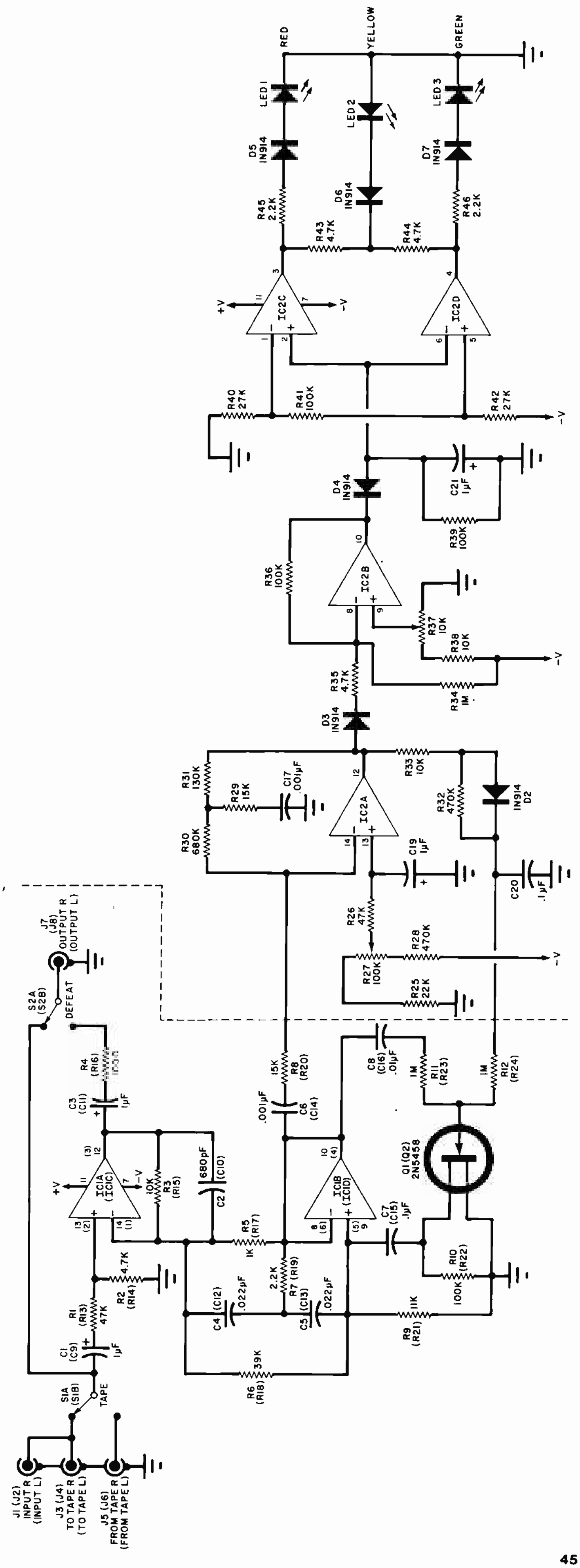

Fig. 2. Schematic diagram. Components of the left-channel filter are numbered in parentheses. The power supply is common to both channels.

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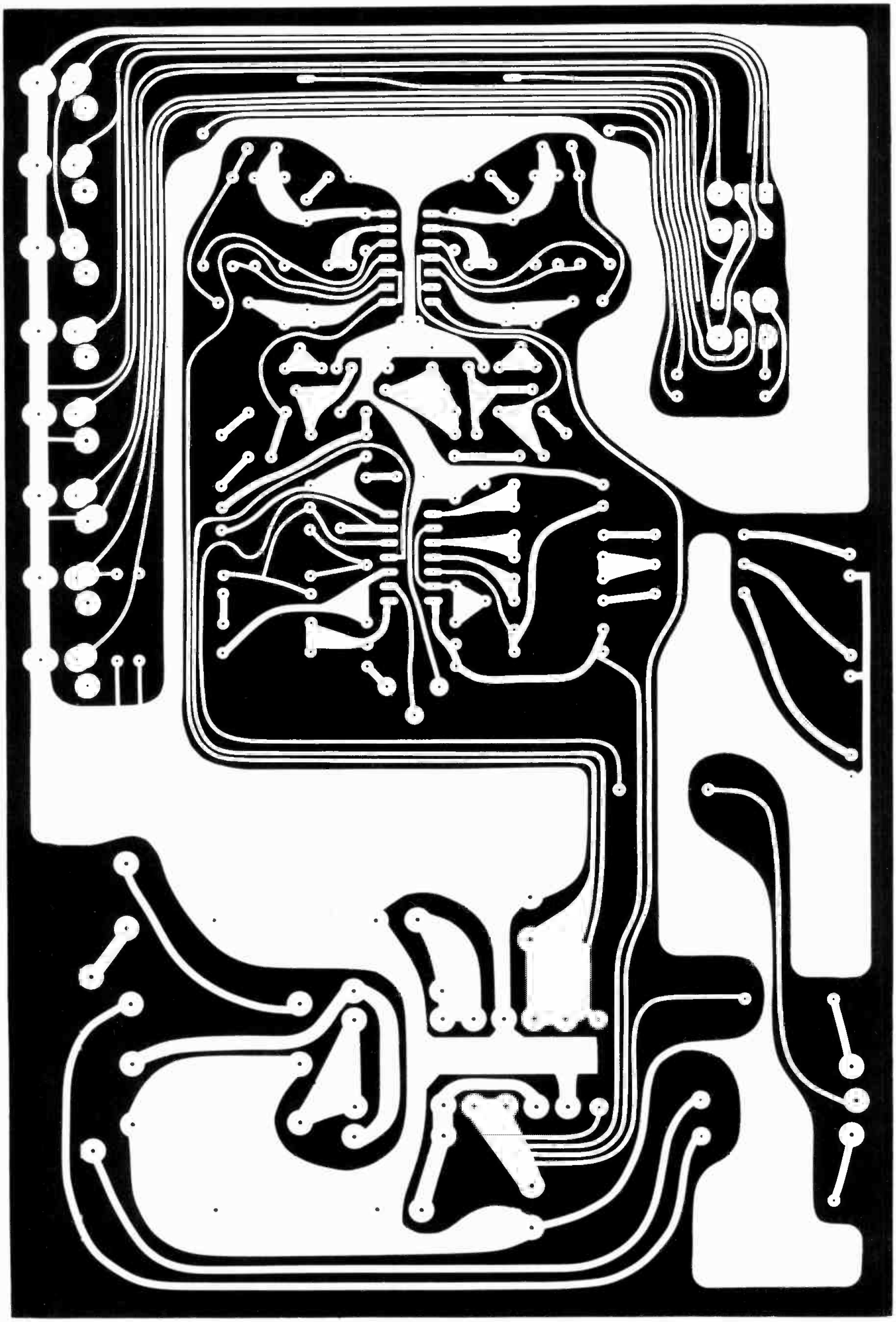

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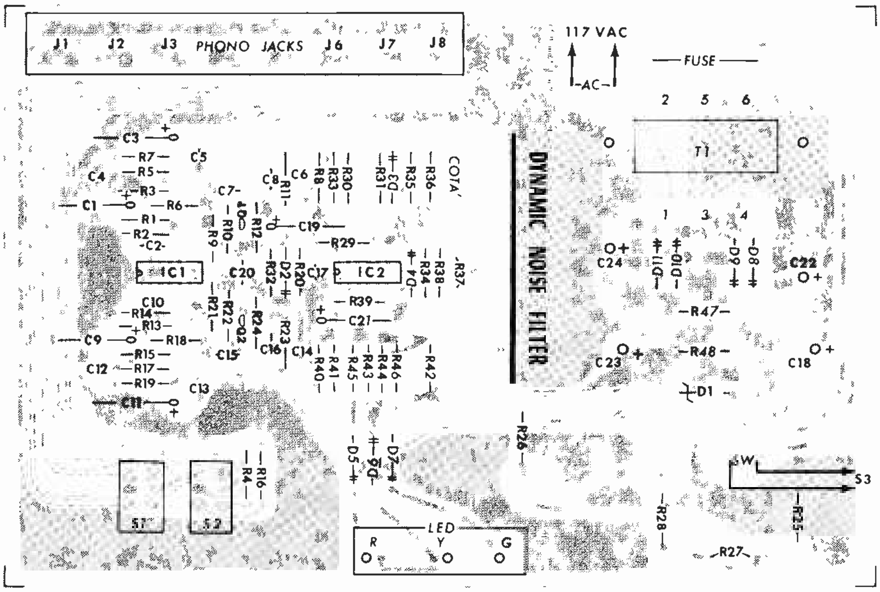

B

Fig. 4. Actual-size etching and drilling guide for the "Silencer." Board is shown at (A); parts placement at (B).

OPERATING SPECIFICATIONS—"SILENCER"

Hiss Reduction:	15 dB at 10,000 Hz
Max. Filter Slope:	9 dB/octave
Frequency Response:	20 to 20,000 Hz ±0.5 dB
Minimum Bandwidth: (Filter Closed)	1500 Hz
Dynamic Range:	Output noise greater than 100 dB below max. output, 20 to 20,000 Hz
S/N Ratio:	Better than 85 dB below 2 V ac output 20 to 20,000 Hz
THD:	Less than 0.1%, at rated output, 20 to 20,000 Hz.
IM Distortion:	Less than 0.01% at rated output 60/7000 Hz mixed 4:1; typically less than 0.005%
Rated Output:	2 V ac into 10,000 ohms
Max. Output:	10 V ac into 10,000 ohms
Input Impedance:	47,000 ohms, single ended
Output Impedance:	100 ohms
Power requirements:	110/120 V, ac 50/60 Hz, 8 W

Note: All measurements made with filter bandwidth open maximum except where specified. (This is the worst-case condition.)

Construction. This unit is most easily constructed using a printed-circuit board. Complete etching and drilling guides are shown in Fig.4A, with the component guide shown in Fig. 4B. Proper orientation of parts is very important. Take careful note of how FETs *Q1* and *Q2* are mounted as well as op amps, diodes, and electrolytic capacitors. Also observe that the dynamic characteristics of the FETs must be matched. Moreover, when choosing op amps, it is important to make sure that the one chosen for the detection circuit, *IC2*, has an open-loop gain of at least 50 dB at 10 kHz. Op amps in the parts section were chosen for their excellent noise figures.

The unit is designed to fit into a custom aluminum extrusion, held by the eight screws in the wood ends. Any suitable enclosure will work, however. The circuit board itself measures 6" × 9". The RCA phone jacks, front-panel switches, and threshold pot are circuit-board mounted for ease of construction and minimum noise. LEDs may be circuit-board mounted or attached to your front panel and then wired. If you choose not to use the furnished printed-circuit guides, make sure that the power supply is as far away as possible from the rest of the circuit to eliminate stray hum.

Calibration. Calibration should be done before you fully enclose the unit. To calibrate, connect the noise filter into your amplifier's or receiver's auxiliary or tape input. Find a low-level noise source—an erased magnetic tape would be ideal. If you don't have tape facilities you may use the inside groove of an LP record. Increase the amplifier's gain so you can hear the noise very well.

Start with the Silencer's threshold pot turned fully counter-clockwise and slowly turn the control knob clockwise. You will hear the noise change character and become more objectionable. Return to the position where the noise-content change just begins (listen several times so you will be able to identify this point). With the threshold knob in this position, adjust *R37*, the thumbwheel trimpot, so that the red LED lights. You should adjust the pot so that it is at the point where only a slight adjustment will cause the yellow LED to light.

In conclusion, this easy-to-build noise-reduction system will be a helpful and versatile addition to any stereo hi-fi system, cleaning up signals from any source. ◇

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BY GEORGE STEBER

The MORSE-A-WORD

PART ONE: Theory and System Operation

LED readout displays words and numbers when Morse code is received

NOW YOU can literally sit back and read messages sent in International Morse even if you don't know the code. The "Morse-A-Word" project presented here automatically converts incoming dits and dahs from a communications receiver or telegraph key into alphanumeric symbols for display on a multicharacter LED readout. The display operates in moving-character fashion to make it easy to read the messages.

With this project, SWLs can listen in on commercial and amateur code traffic. And for beginning as well as veteran radio amateurs, the Morse-A-Word makes an excellent operating and code-training aid. Cost of a complete kit including a prepunched and lettered chassis and two two-character displays is $140. One or two additional displays can be added at moderate cost.

This project is similar to the Morse-A-Letter featured in the January 1977 issue of POPULAR ELECTRONICS. Its display capability has been expanded, however. At the builder's option, the Morse-A-Word can display two, four, six or eight characters simultaneously. All

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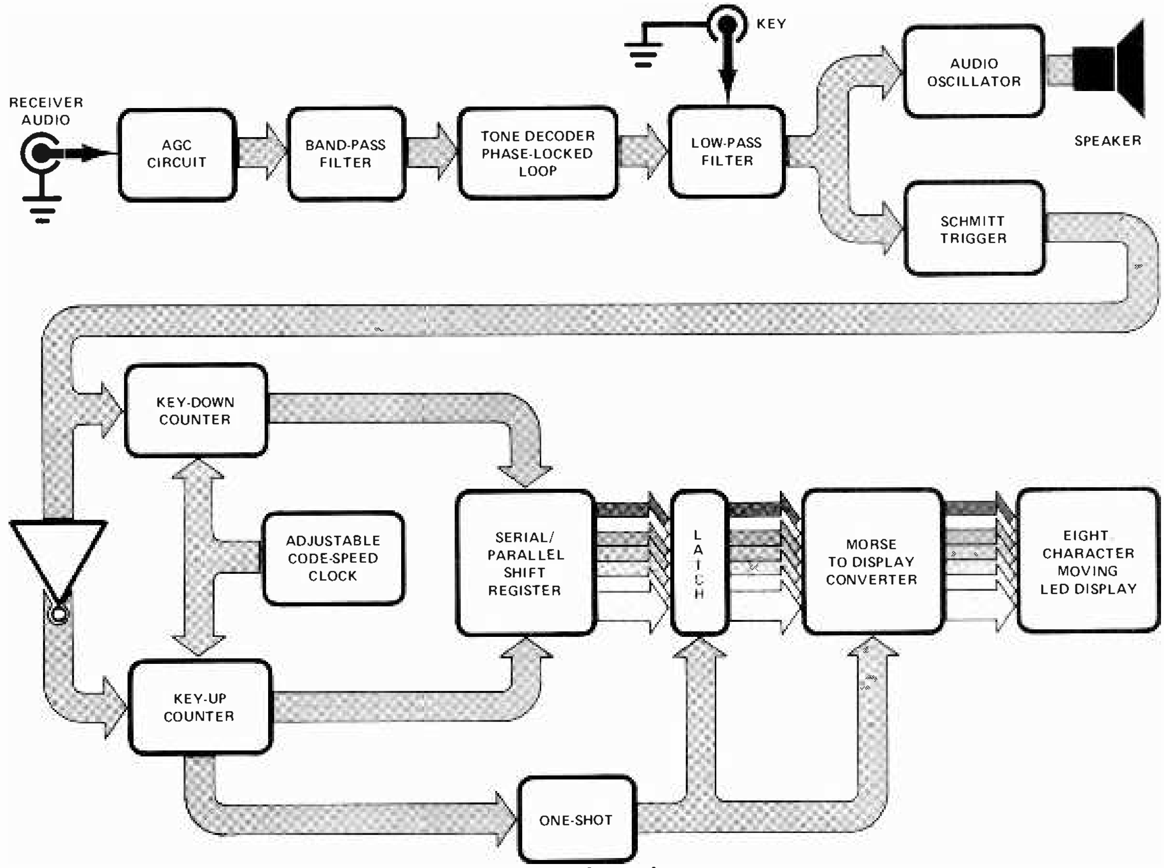

Fig. 1. Block diagram of the Morse-A-Word system shows how the incoming signal in code is processed for alphanumeric display.

characters—letters, numerals, punctuation marks and, if desired, word spaces—are displayed and shifted from right to left as new ones stream in.

Double-sided pc boards hold the LED display and main decoder circuits. A single-sided board accommodates the power supply.

It should be mentioned at the outset that the reliable conversion of Morse code radio signals into alphanumeric characters is not easy. Signal fading, atmospheric and man-made noise, and human errors present major difficulties. Consequently, no device can perfectly decode all received signals all of the time. The highly sophisticated Morse-A-Word circuit has been designed to provide a very high degree of accuracy, however, and will do a very creditable decoding job in far-from-ideal situations.

System Analysis. A block diagram of the Morse-A-Word is shown in Fig. 1. The complete schematic of the main decoding circuit is in Fig. 2, and the display circuit is shown in Fig. 3.

PARTS LIST: MAIN DECODING CIRCUIT

C1,C2,C5,C10,C12,C15,C17,C18 through C21,C23—0.1- or 0.05-µF disc ceramic
C3, C7—22-µF, 10-volt tantalum
C4—0.05-µF disc ceramic
C6,C9,C11—0.01-µF Mylar
C8—1-µF, 10-volt tantalum
C13—0.22-µF Mylar
C14—6.8-µF, 10-volt tantalum
C16—0.47-µF, 10-volt tantalum
C22—27-pF disc ceramic
D1,D2,D3—1N270 germanium diode
IC1, IC2—7495 4-bit shift register
IC3,IC6,IC15,IC17—74161 4-bit counter
IC4,IC8—741 operational amplifier (8-pin mini-DIP)
IC5—74174 hex D flip-flop
IC7—7414 hex inverting Schmitt trigger
IC9,IC10—7489 64-bit RAM
IC11—74121 monostable multivibrator
IC12—555 timer
IC13—567 PLL tone decoder
IC14—1702A PROM
IC16—7402 quad 2 input NOR gate
IC18—7483 4-bit binary adder
IC19—7485 4-bit magnitude comparator
J1,J2—Phono jack
LED1, LED2—Red light-emitting diode
Q1—2N3823 n-channel JFET
The following are ¼-watt, 10% tolerance fixed resistors.
R1,R4,R27—220 ohms
R2—10,000 ohms
R3,R13,R15—470 ohms
R5—15,000 ohms
R6,R17,R21 through R26—1000 ohms
R7—150,000 ohms
R8—330 ohms
R10—680 ohms
R11,R19—6800 ohms
R12—270,000 ohms
R16—47,000 ohms
R18—12,000 ohms
R9,R14—500-ohm pc trimpot
R20—5000-ohm pc trimpot
R28—500-ohm linear-taper potentiometer with ganged spst power switch
S1—Spst slide or toggle switch
SPKR—8-ohm dynamic loudspeaker
Misc.—Printed circuit board, IC sockets or Molex Soldercons, suitable enclosure, LED holders, pc standoff insulators, control knob, machine hardware, hookup wire, solder, etc.
Note—For parts and kit ordering information, refer to the Parts Availability list.

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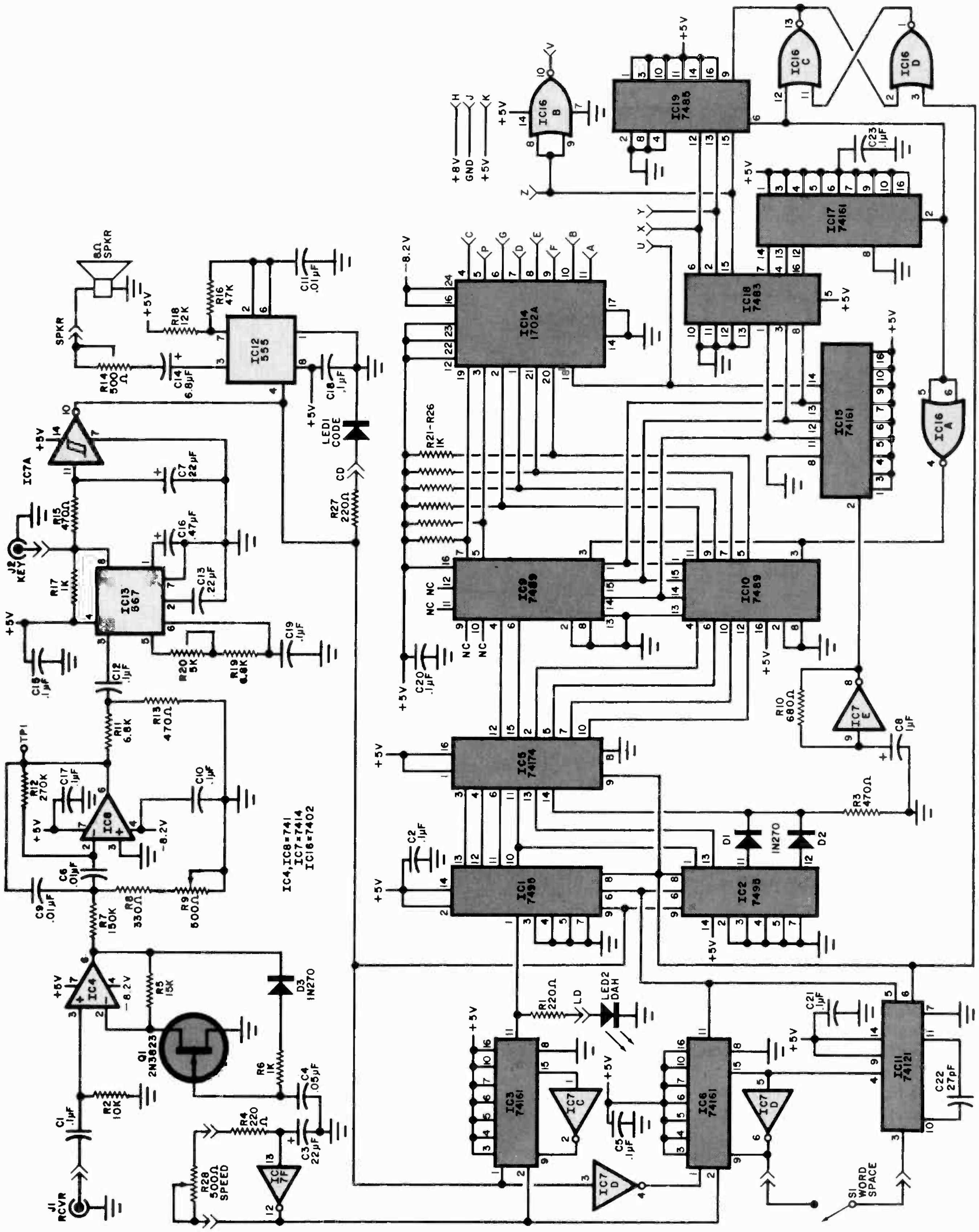

Fig. 2. Schematic diagram of the main decoder circuit. If the audio output of a radio receiver is used, it is applied to J1. An input from a telegraph key is applied to J2. Parts list is on facing page.

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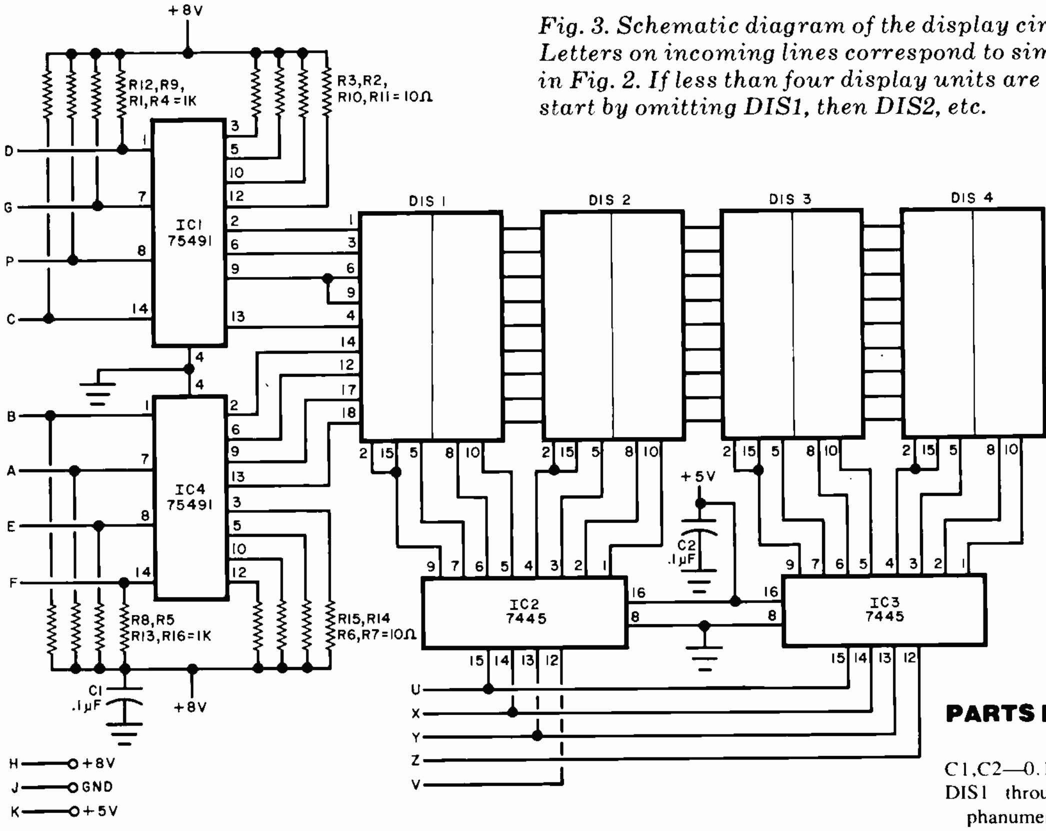

Fig. 3. Schematic diagram of the display circuit. Letters on incoming lines correspond to similar points in Fig. 2. If less than four display units are used, start by omitting DIS1, then DIS2, etc.

PARTS LIST: DISPLAY CIRCUIT

C1,C2—0.1- or 0.05-µF disc ceramic
DIS1 through DIS4—IEE 1785R dual alphanumeric LED display
IC1,IC4—75491 MOS-to-LED display driver
IC2,IC3—7445 or 74145 BCD-to-decimal decoder/driver
The following are ¼-watt, 10% tolerance fixed resistors.
R1,R4,R5,R8,R9,R12,R13,R16—1000 ohms
R2,R3,R6,R7,R10,R11,R14,R15—10 ohms
Misc.—Printed circuit board, Molex Soldercons for displays, Soldercons or IC sockets for driver ICs, red bezel for displays, solid hookup wire, solder, etc.
Note—For parts and kit ordering information, refer to the Parts Availability list.

Referring to Fig. 1, the audio output of a radio receiver is applied to an agc stage which limits the amplitude excursions of the input signal. The output of the agc stage drives an active bandpass filter whose response is centered at 1200 Hz. A tone decoder with a phase-locked loop, whose response is also peaked at 1200 Hz, receives signals from a bandpass filter and demodulates them. This decoder generates a low voltage when the transmitter's telegraph key is down and a high voltage under key-up conditions. A low-pass filter smooths the output of the tone decoder and can accept a telegraph key input for code practice use.

Further signal processing is performed by a Schmitt trigger which "squares up" and inverts the signals applied to it. At the output of the Schmitt trigger, a logic 1 corresponds to a key-down condition, and a logic zero to a key-up condition. Signal processing is now complete, and clean, TTL-compatible Morse signals are available to the di-

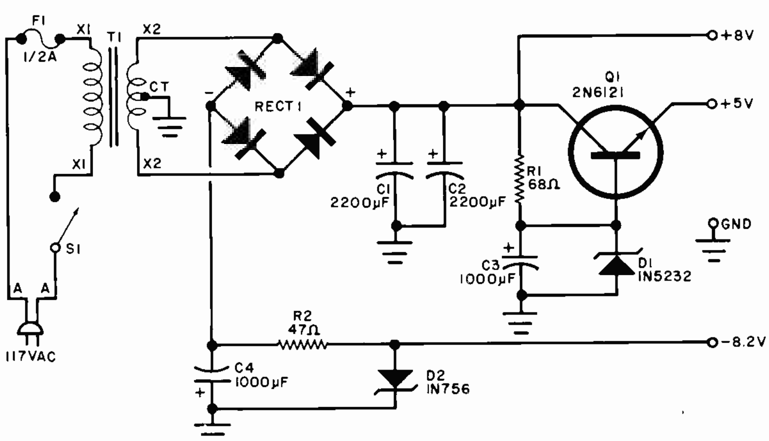

Fig. 4. Schematic diagram of power supply circuit. The main decoder requires 750 mA at 5 volts and 20 mA at −8.2 volts. Display is best with 8-volt supply.

PARTS LIST: POWER SUPPLY

C1,C2—2200-µF, 16-volt upright electrolytic
C3—1000-µF, 10-volt upright electrolytic
C4—1000-µF, 16-volt upright electrolytic
D1—1N5232 5.6-volt zener
D2—1N756 8.2-volt zener
F1—½-ampere fast-blow fuse
Q1—2N6121 npn tab (TO-220) transistor
R1—68-ohm, ½-watt, 10% resistor
R2—47-ohm, ½-watt, 10% resistor
RECT1—1-ampere, 50-PIV modular bridge rectifier
S1—Spst power switch (part of main circuit R28)
T1—12.6-volt, 2-ampere center-tapped transformer (Stancor P8130 or equivalent)
Misc.—Printed circuit board, pc-mount heat sink for *Q1*, silicone thermal compound, fuseholder, pc standoff insulators, line cord and strain relief, hookup wire, machine hardware, solder, etc.
Note—For parts and kit ordering information, refer to the Parts Availability list.

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gital decoding circuits.

The digitized Morse is first applied to two counters. One counter, but not both, will be enabled to count, depending on whether the key is up or down. These circuits count at a rate dependent on the frequency of an adjustable code-speed clock. The clock frequency should be adjusted to match the speed of the incoming code, but this adjustment can be off by as much as ±50% and still result in solid copy.

Whenever the key-up counter detects an element space, a condition that occurs when it counts less than eight clock pulses, it serially transfers a logic 0 or 1 to the next stage, an eight-bit serial/parallel shift register. The latter is always initialized with the binary word 00000001 so that the beginning of each Morse character will be uniquely decodable. Whether a logic 1 or 0 is transferred to the shift register in subsequent steps is determined by the condition of the key-down counter, which distinguishes between dits and dahs. If the key-down counter counts more than seven clock pulses, the code element is a dah and a logic 1 is transferred to the shift register. Otherwise, it is a dit and a logic 0 is transferred to the shift register. The detection scheme is similar to that employed in the Morse-A-Letter, and has been found to be very reliable.

This procedure continues until the key-up counter detects a space longer than an element space (longer than seven clock periods), whereupon the circuit determines that a complete character has been sent. The unique binary code present in the shift register can now be transferred to a latch for decoding and display. However, if the key-up counter continues to count more than 15 clock pulses, this is interpreted as a space between words and a blank character is inserted in the latch after the last character is received. Because many CW stations do not send word spaces, the circuit contains a switch to defeat the word-space feature.

A 16-element RAM (in which only 8 elements are used) stores the Morse characters obtained from the latch. The RAM is synchronized to the eight-character display by an address counter and a ROM which decodes the Morse characters for display. A standard multiplexed circuit is employed for display of stored characters, which appear on IEE 1785R two-character LED displays. The circuit has been designed to provide a moving-character type of display which introduces new characters at the rightmost position and moves each of the existing characters to the left, one position at a time, as characters are received. It takes just a few minutes to accustom yourself to reading this type of presentation. Once you get the hang of it, reading code is a breeze.

The Morse-A-Word's main decoder circuit power requirements are 750 mA at +5 volts and 20 mA at −8.2 volts. The display circuit also calls for 8 volts at approximately 100 mA. Voltages as low as 5 V can be used to power the display, but it will not be as bright. A suggested power supply is shown in Fig. 4.

In Part Two of this article, next month, we will describe how to assemble, align, and use the project. Programming instructions for IC14 will be included. ◇

PARTS AVAILABILITY

The following are available from Microcraft Corp., Box 513, Theinsville, WI 53092:

No. MAWK-1. Complete kit of parts, including prepunched and lettered cabinet and two dual-character IEE 1785R LED displays, $139.95. (One or two additional dual-character displays can be ordered at the builder's option.)

No. EPK-1. Essential parts kit including two (main and display) pc boards, preprogrammed ROM, all ICs, sockets, resistors and capacitors, one dual-character IEE 1785R LED display, but not including power supply, hookup wire, solder, loudspeaker, enclosure, control knob, jacks and miscellaneous hardware, $99.50.

No. PCBK-1. Set of three (main, display and power supply) pc boards, $24.00.

No. MB-1. Etched and drilled, double-sided, glass epoxy main pc board with plated-through holes, $12.50.

No. DB-1. Etched and drilled, double-sided, glass epoxy display pc board with plated-through holes, $7.00.

No. PSB-1. Etched and drilled, glass epoxy power supply pc board, $5.50.

No. PSK-1. Power supply kit, including pc board and all components, $22.00.

No. Rom-1. Preprogrammed 1702A ROM, $10.00.

No. DSP-1. One dual-character IEE 1785R LED display, $9.00.

No. CAB-1. Prepunched and lettered enclosure, $17.00.

No. CT-1. Alignment and code practice cassette tape, $6.00.

Prices include shipping and handling within the continental USA. Wisconsin residents, add 4% sales tax.

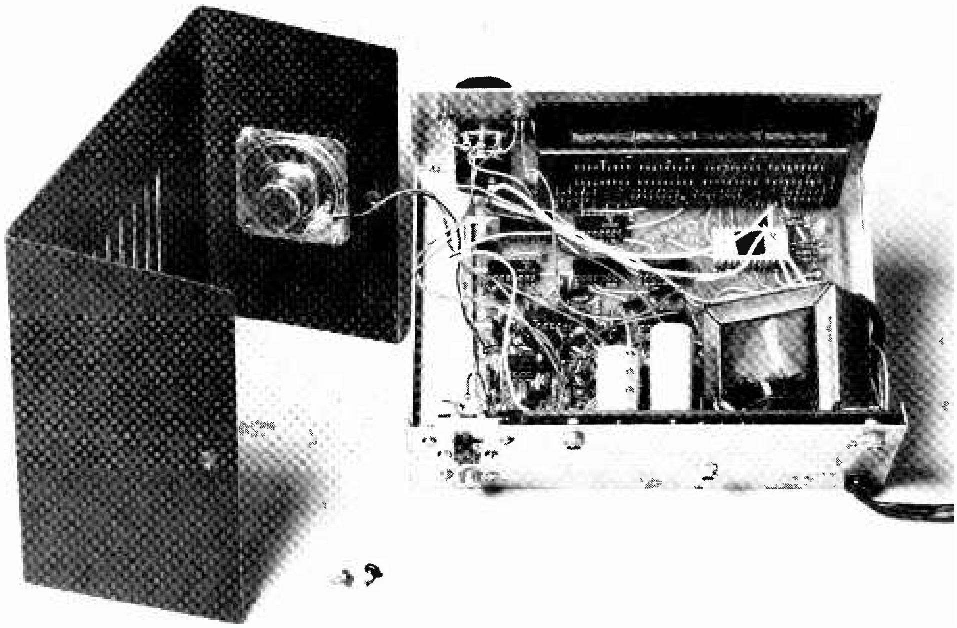

Photo shows internal assembly of the author's prototype. Display board is on front panel, power supply on back.

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The MORSE-A-WORD

BY GEORGE STEBER

PART TWO: Construction, Alignment, and Use

Construction. The Morse-A-Word is most easily assembled using printed circuit techniques. Three pc boards are required—a main circuit board, display board, and a power supply board. The parts placement guide for the main circuit board appears in Fig. 5. Etching and drilling guides are shown in Fig. 6. Similarly, etching and drilling guides for the double-sided display board are shown in Fig. 7. This board's parts placement guide appears in Fig. 8. Finally, the etching and drilling and parts placement guides for the power supply board appear in Figs. 9 and 10.

When soldering components to circuit board foils, use a low-wattage, fine-tipped soldering pencil and fine solder. Be sure to employ the minimum amount of heat and solder consistent with good connections, and take care not to inadvertently create solder bridges between adjacent foils. The use of IC sockets or Molex Soldercons is recommended.

Assemble the main pc board first. Start by inserting and soldering the IC sockets and Molex Soldercons. Install the smallest components next, gradually working up to the larger items. For example, start with the ¼-watt resistors, then install the diodes, the small capacitors and finally the larger capacitors. Be sure to observe the polarities of diodes and tantalum and electrolytic capacitors, and the pin basing of transistors and ICs. The board furnished by the kit supplier has plated-through holes so you need only solder component leads on the bottom side of the board.

Neither the power supply, the display circuits, the sidetone speaker, jacks, CODE and DAH LEDs or the SPEED control are mounted on the main pc board. Insulated wire leads of suitable lengths should be soldered to appropriate points on the pc board now for connection to these components.

Wire the display board next, referring to the parts placement diagram of Fig. 8. Use Molex Soldercons to mount the dual IEE 1785R LED displays. Make sure the Soldercons are properly aligned before soldering them to the board. This will ensure a good fit for the displays. Resistors, capacitors and IC sockets or Soldercons for the driver ICs should be installed and soldered next. The resistors should be mounted in a vertical position. Notice that there are a number of jumper wires to be soldered to this board. These are used to interconnect the display board and the main circuit board and to support the display board. The jumpers should be made of heavy solid wire, about ½" (1.3 cm) long, and bent into "L" shapes so they extend parallel to the board and point downward.

Position the display board perpendicular to the main pc board. Insert the jumper wires connected to the display board through the appropriate holes on the main pc board and push down the display board until it just touches the main board. Check the alignment of the display board and then solder the jump-

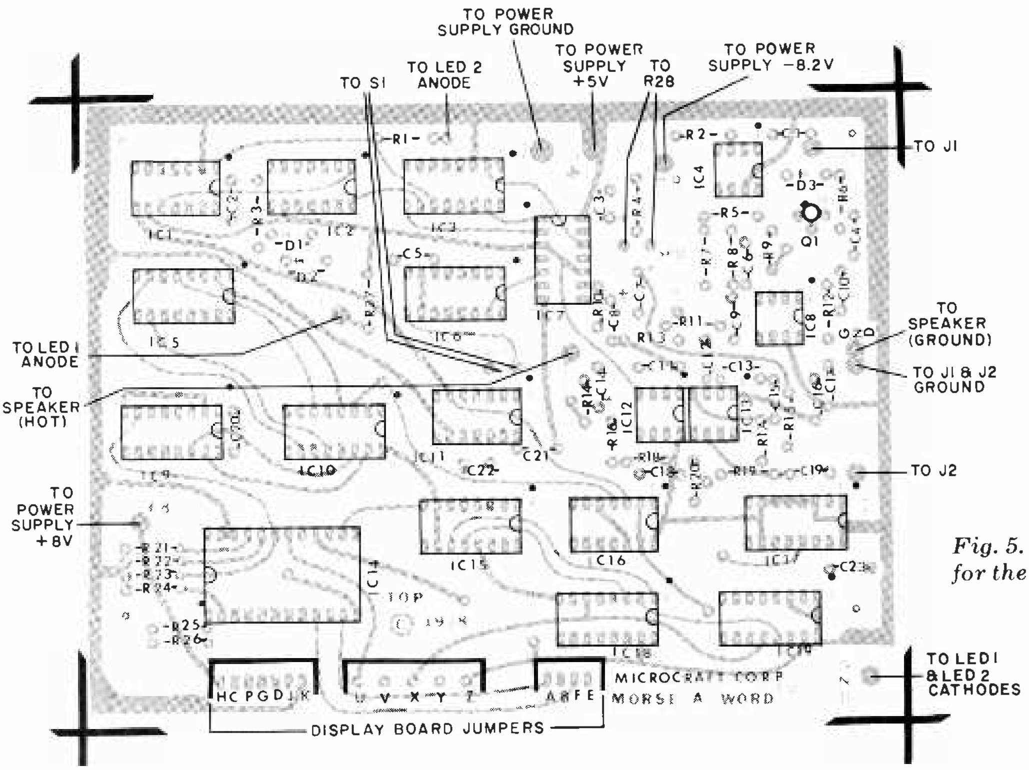

Fig. 5. Parts placement guide for the main circuit board.

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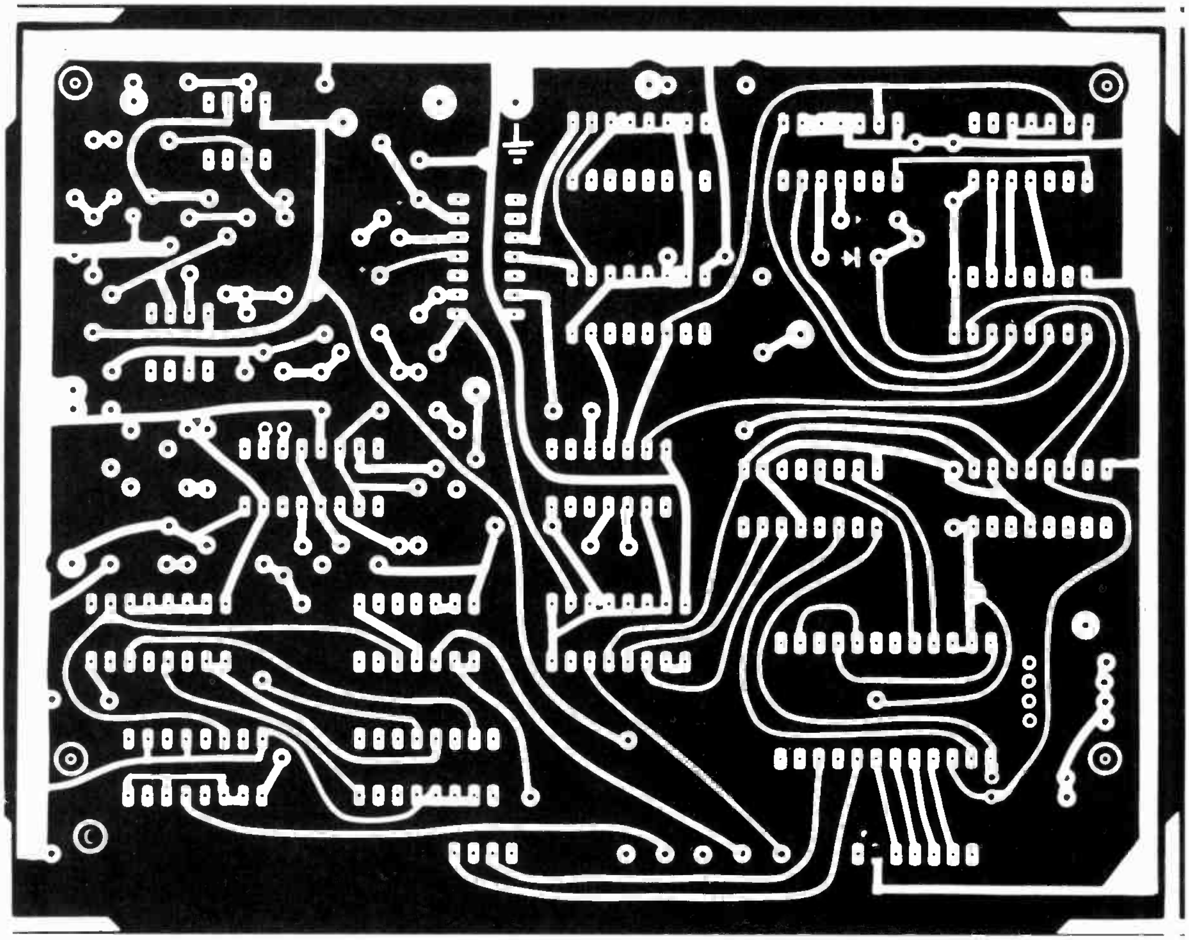

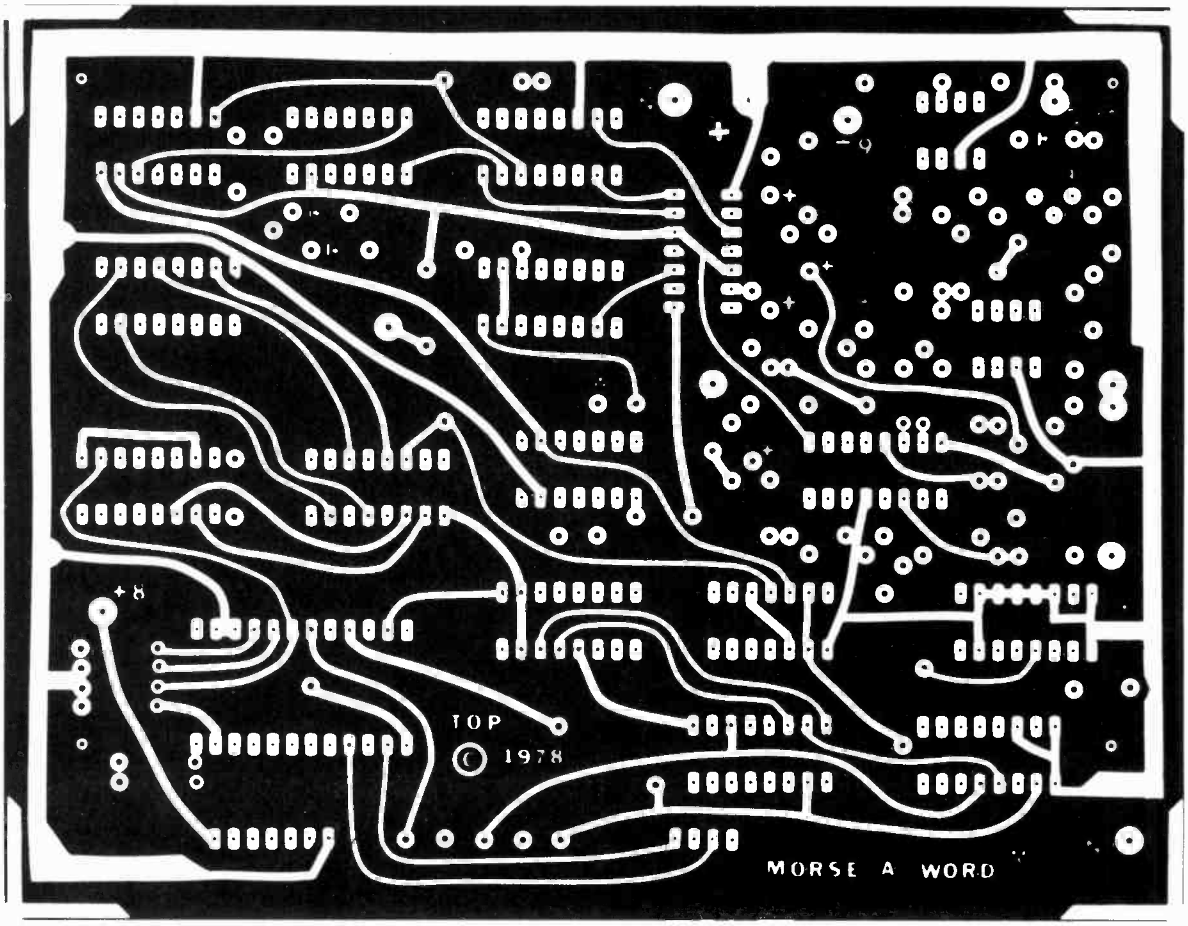

Fig. 6. Actual-size etching and drilling guides for the double-sided main printed circuit board.

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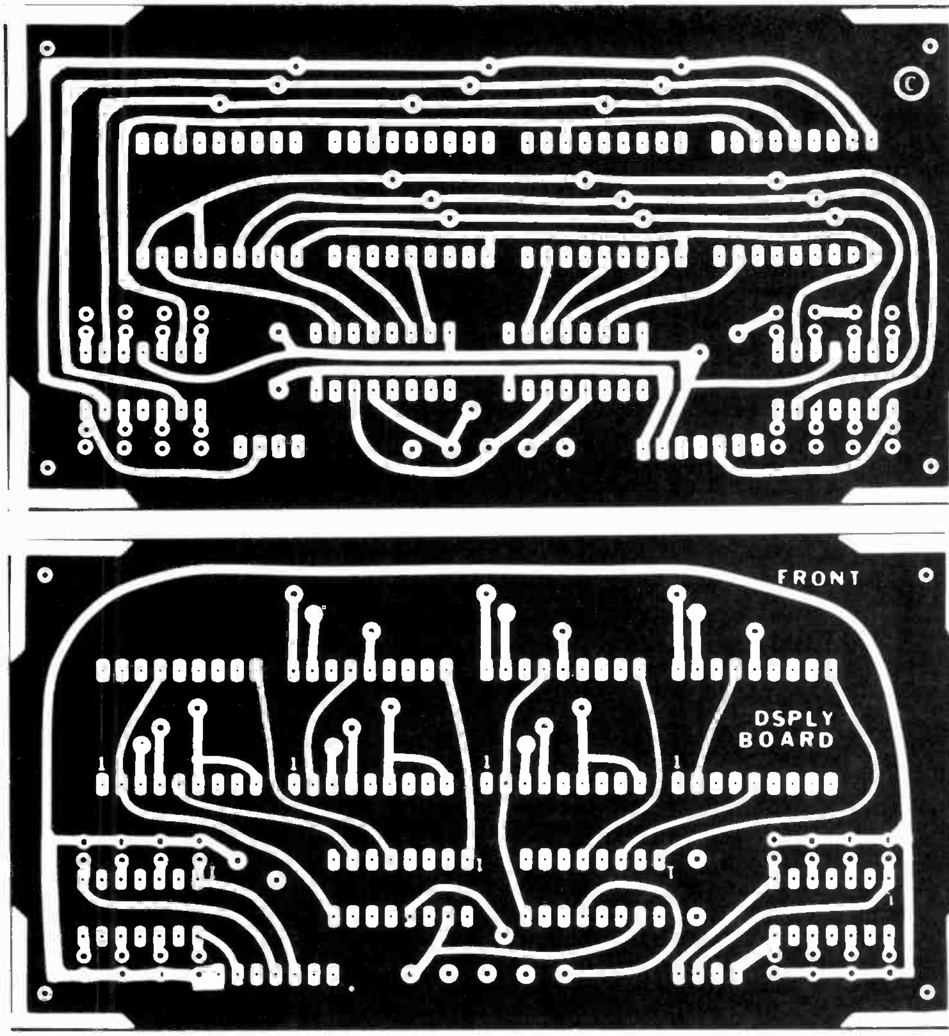

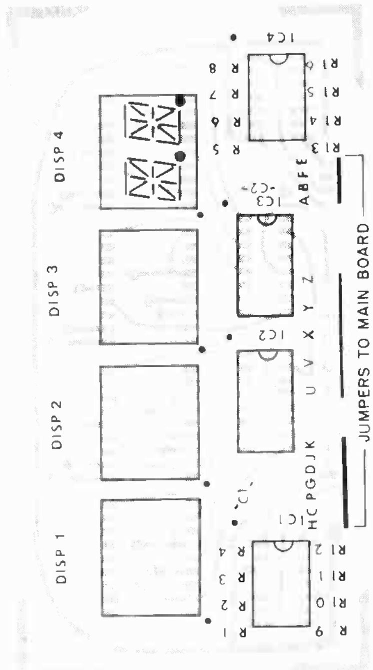

Fig. 8. Parts placement guide for front side of display board is above.

Fig. 7. Actual-size etching and drilling guides for the double-sided display board are at left.

ers to the foils on the bottom of the main pc board. Cut off excess jumper lengths.

For proper Morse decoding, the 1702A ROM must be programmed in accord with Table I. A construction article that appeared in the February 1978 issue of this magazine described a project that allows you to program your own blank ROMs. Some parts distributors will program the 1702A for you if the truth table accompanies your order. The kit supplier for the Morse-A-Word also offers a preprogrammed ROM.

Install the ICs and the dual-character IEE 1784R LED displays in their Soldercons. Make sure they are correctly oriented and take the usual precautions to avoid bending the leads or damaging the MOS ROM. It is not necessary to have a full eight-character display. Those builders with a tight budget, for example, can install only one dual-character IEE 1785R LED readout. However, a minimum of two readouts (four characters) is recommended. If fewer than four readouts are used, make sure they are right-justified (installed at the *DIS4*, *DIS4* and *DIS3*, etc.)

The remaining pc board, that for the power supply, should now be assembled. You will note that the board has space for extra components to be used in another project. These components

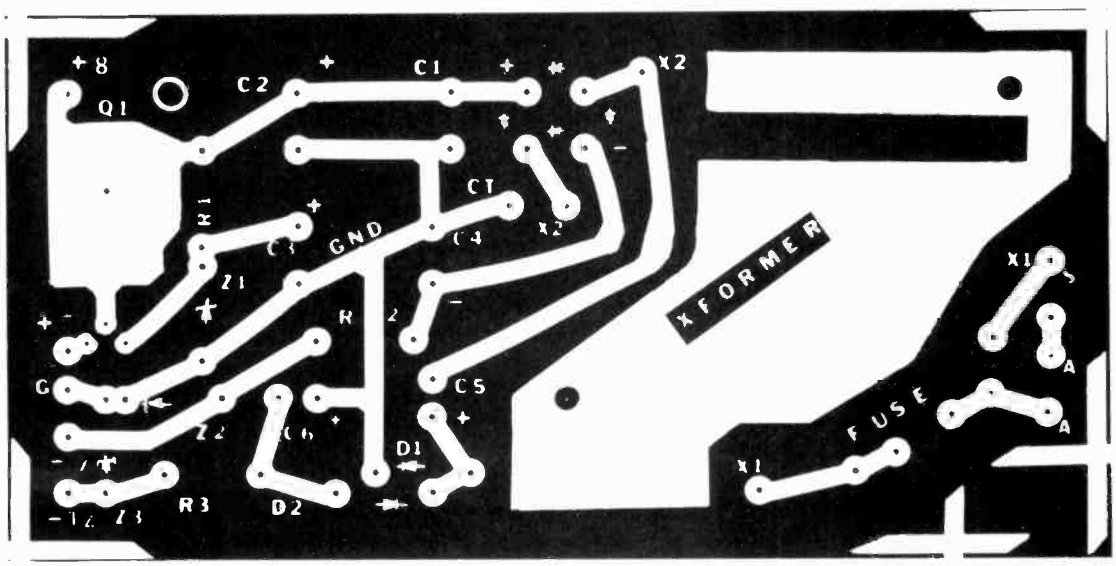

Fig. 9. Actual-size etching and drilling guide for the power supply board.

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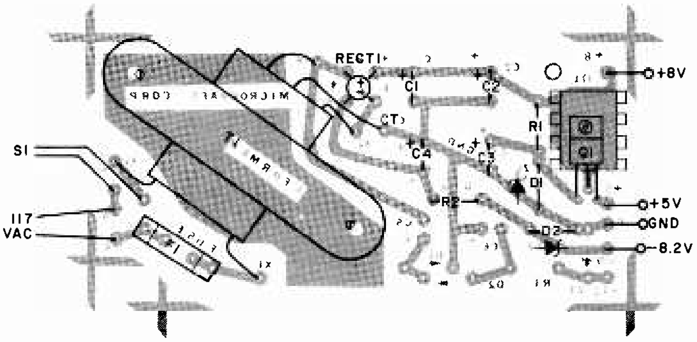

Fig. 10. Component placement for the power supply board shown above.

are not required in the Morse-A-Word and the pc locations for them should be ignored. When you have completed assembly of the board, apply line power to it and verify that the desired voltages are being produced. If the voltages are correct, remove line power and interconnect the supply and main pc boards with suitable lengths of color-coded hookup wire. Then mount the boards in the project enclosure and connect the free ends of the hookup wires already soldered to the main pc board to speaker, jacks, etc. A cutout for the displays must be made on the front panel of the enclosure. This can best be done with a nibbling tool. For those who prefer a prepunched enclosure, one is available from the kit supplier. Display contrast and project appearance will be enhanced by installing a bezel and red filter in the cutout.

Apply power to the project. Several or all of the dual-character displays should start to glow. If they don't, disconnect power and go back and thoroughly check for loose wires, cold solder joints, solder bridges, or incorrect wiring.

Alignment. The center frequency of the bandpass filter and the tone decoder's peak response frequency must be the same if the Morse-A-Word is to function properly. Any frequency between 800 and 2600 Hz is suitable, but the higher frequencies will produce a better circuit response. On the other hand, the higher frequencies tend to be more difficult to tune on a highly selective communications receiver. As a compromise, 1200 Hz was selected as the center frequency of the band-pass filter and the tone decoder.

To align the project, apply a 0.5-volt rms, 1200-Hz signal to the receiver input jack. Connect an ac voltmeter or oscilloscope to the output of the bandpass filter (*TP1*) and adjust trimmer potentiometer *R9* for maximum output. Next, adjust *R20* until a tone is heard in the speaker and the CODE LED lights. Reduce the input signal to as low a level as possible and repeat the procedure. If a 1200-Hz signal is not available and you have a cassette tape recorder, a cassette tape available from the kit supplier has the necessary tone recorded on it. The tape also includes recordings of sample Morse code messages and selections which can be used for code practice.

The only other adjustment is the setting of trimmer potentiometer *R14*, which determines the loudness of the speaker output. A low volume setting is

TABLE I
TRUTH TABLE FOR 1702A PROM IC14

Character	Input								Output							
	A7	A6	A5	A4	A3	A2	A1	A0	D8	D7	D6	D5	D4	D3	D2	D1
A	0	1	1	0	1	0	1	1	1	1	0	0	1	0	0	1
	0	0	1	0	1	0	1	1	0	1	0	0	1	0	0	1
B	0	1	1	1	1	1	0	0	1	0	0	1	0	0	0	0
	0	0	1	1	1	1	0	0	1	1	0	1	1	0	0	1
C	0	1	1	1	0	1	0	0	1	1	0	0	0	0	0	1
	0	0	1	1	0	1	0	0	1	0	0	0	0	0	0	0
D	0	1	1	1	1	0	0	1	1	0	0	1	0	0	0	0
	0	0	1	1	1	0	0	1	1	1	0	1	0	0	0	1
E	0	1	1	1	0	1	1	1	1	1	0	0	1	0	0	1
	0	0	1	1	0	1	1	1	1	0	0	0	0	0	0	0
F	0	1	1	1	0	1	1	0	1	1	0	0	1	0	0	1
	0	0	1	1	0	1	1	0	0	0	0	0	0	0	0	0
G	0	1	1	1	0	0	0	1	1	1	0	0	0	0	0	1
	0	0	1	1	0	0	0	1	1	0	0	0	1	0	0	1
H	0	1	1	1	1	1	1	0	0	1	0	0	1	0	0	1
	0	0	1	1	1	1	1	0	0	1	0	0	1	0	0	1
I	0	1	1	1	1	0	1	1	1	0	0	1	0	0	0	0
	0	0	1	1	1	0	1	1	1	0	0	1	0	0	0	0
J	0	1	1	0	0	0	1	0	0	0	0	0	0	0	0	1
	0	0	1	0	0	0	1	0	1	1	0	0	0	0	0	1
K	0	1	1	0	1	0	0	1	0	1	0	0	1	0	0	1
	0	0	1	0	1	0	0	1	0	0	1	0	0	1	0	0
L	0	1	1	1	1	0	1	0	0	1	0	0	0	0	0	1
	0	0	1	1	1	0	1	0	1	0	0	0	0	0	0	0
M	0	1	1	0	0	0	1	1	0	1	1	0	0	0	0	1
	0	0	1	0	0	0	1	1	0	1	1	0	0	0	0	1
N	0	1	1	1	0	0	1	1	0	1	1	0	0	0	0	1
	0	0	1	1	0	0	1	1	0	1	0	0	0	1	0	1
O	0	1	1	0	0	0	0	1	1	1	0	0	0	0	0	1
	0	0	1	0	0	0	0	1	1	1	0	0	0	0	0	1
P	0	1	1	1	0	0	1	0	1	1	0	0	1	0	0	1
	0	0	1	1	0	0	1	0	0	1	0	0	1	0	0	0
Q	0	1	1	0	1	0	0	0	1	1	0	0	0	0	0	1
	0	0	1	0	1	0	0	0	1	1	0	0	0	1	0	1
R	0	1	1	1	0	1	0	1	1	1	0	0	1	0	0	1
	0	0	1	1	0	1	0	1	0	1	0	0	1	1	0	0
S	0	1	1	1	1	1	0	1	1	0	1	0	0	0	0	1
	0	0	1	1	1	1	0	1	1	1	0	0	0	1	0	0
T	0	1	1	0	0	1	1	1	1	0	0	1	0	0	0	0
	0	0	1	0	0	1	1	1	0	0	0	1	0	0	0	0
U	0	1	1	0	1	1	0	1	0	1	0	0	0	0	0	1
	0	0	1	0	1	1	0	1	1	1	0	0	0	0	0	1

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recommended to avoid confusion when listening to both the receiver and the Morse-A-Word simultaneously.

Use. The Morse-A-Word is easy to operate. The setting of the front-panel SPEED control (*R28*) is the only adjustment that must be made, and only a rough setting is required. Keep in mind that the Morse-A-Word has a sensitive input stage, so don't set the receiver audio gain control higher than is necessary. When the receiver is tuned to the center of the filter passband (1200 Hz), you should hear audio from the project's internal speaker and the code LED should flicker in time with the incoming code. The passband is only about 120 Hz wide, so some care is required when tuning in a signal.

With the signal properly tuned in, adjust the SPEED control so the DAH LED glows only when dahs are sent, and not dits. The alphanumeric readout LED will now display the incoming characters. If word spaces are desired, make sure the WORD SPACE switch is closed. Only a few amateur stations actually send word spaces, so don't expect perfectly spaced copy unless you are tuned to a station such as W1AW which sends machine-perfect code. Invalid Morse characters will be displayed as blanks.

V	0	1	1	0	1	1	1	0	0	1	0	0	0	1	0	1
	0	0	1	0	1	1	1	0	0	0	1	0	0	0	0	0
W	0	1	1	0	0	1	0	1	0	1	0	0	0	1	0	1
	0	0	1	0	0	1	0	1	0	1	0	0	0	1	0	1
X	0	1	1	0	1	1	0	0	0	0	1	0	0	1	0	0
	0	0	1	0	1	1	0	0	0	0	1	0	0	1	0	0
Y	0	1	1	0	0	1	0	0	0	0	1	0	0	0	0	0
	0	0	1	0	0	1	0	0	0	0	1	1	0	0	0	0
Z	0	1	1	1	1	0	0	0	1	0	0	0	0	1	0	0
	0	0	1	1	1	0	0	0	1	0	1	0	0	0	0	0
0	0	1	0	0	0	0	0	0	1	1	0	0	0	1	0	1
	0	0	0	0	0	0	0	0	1	1	1	0	0	0	0	1
1	0	1	0	0	0	0	0	1	0	0	0	1	0	0	0	0
	0	0	0	0	0	0	0	1	0	0	0	1	0	0	0	0
2	0	1	0	0	0	0	1	1	1	0	0	0	1	0	0	1
	0	0	0	0	0	0	1	1	1	1	0	0	1	0	0	0
3	0	1	0	0	0	1	1	1	1	0	0	0	1	0	0	0
	0	0	0	0	0	1	1	1	1	1	0	0	1	0	0	1
4	0	1	0	0	1	1	1	1	0	1	0	0	1	0	0	0
	0	0	0	0	1	1	1	1	0	1	0	0	1	0	0	1
5	0	1	0	1	1	1	1	1	1	1	0	0	1	0	0	0
	0	0	0	1	1	1	1	1	1	0	0	0	1	0	0	1
6	0	1	0	1	1	1	1	0	1	1	0	0	1	0	0	1
	0	0	0	1	1	1	1	0	1	0	0	0	1	0	0	1
7	0	1	0	1	1	1	0	0	1	0	0	0	0	0	0	0
	0	0	0	1	1	1	0	0	0	1	0	0	0	0	0	1
8	0	1	0	1	1	0	0	0	1	1	0	0	1	0	0	1
	0	0	0	1	1	0	0	0	1	1	0	0	1	0	0	1
9	0	1	0	1	0	0	0	0	1	1	0	0	1	0	0	0
	0	0	0	1	0	0	0	0	0	1	0	0	1	0	0	1
.	0	1	0	0	1	0	1	0	0	0	0	0	0	0	0	0
	0	0	0	0	1	0	1	0	0	0	0	0	0	0	1	0
,	0	1	0	0	0	1	1	0	0	0	0	0	0	1	0	0
	0	0	0	0	0	1	1	0	0	0	0	0	0	0	0	0
?	0	1	0	1	1	0	0	1	1	0	0	0	0	0	0	0
	0	0	0	1	1	0	0	1	0	1	0	1	1	0	0	0
/	0	1	0	1	0	1	1	0	0	0	0	0	0	1	0	0
	0	0	0	1	0	1	1	0	0	0	1	0	0	0	0	0
–	0	1	0	0	1	1	1	0	0	0	0	0	1	0	0	0
	0	0	0	0	1	1	1	0	0	0	0	0	1	0	0	0
Space	0	1	1	0	1	1	1	1	0	0	0	0	0	0	0	0
	0	0	1	0	1	1	1	1	0	0	0	0	0	0	0	0
$\overline{AR}$	0	1	0	1	0	1	0	1	1	1	0	1	1	0	0	1
	0	0	0	1	0	1	0	1	0	1	0	1	1	1	0	0
$\overline{SK}$	0	1	0	0	1	0	1	1	1	1	0	1	1	0	0	0
	0	0	0	0	1	0	1	1	1	0	1	1	0	1	0	0
$\overline{KN}$	0	1	0	1	0	0	1	0	0	1	0	0	1	0	0	1
	0	0	0	1	0	0	1	0	0	0	1	1	0	1	0	1
$\overline{AS}$	0	1	0	1	1	1	0	1	1	0	1	1	0	1	0	0
	0	0	0	1	1	1	0	1	1	0	1	1	0	1	0	0

For code practice sessions, connect your telegraph key to the KEY jack and adjust the SPEED control for the approximate sending speed. You can calibrate your SPEED control using the formula: Speed (WPM) = 0.15*f*. That is, the code speed in words per minute equals fifteen hundredths of the clock frequency as set by the SPEED control.

An excellent source of code material is amateur station W1AW, operated by the American Radio Relay League. The station transmits several code practice sessions each day, as well as ham news bulletins, propagation forecasts and OSCAR bulletins, all in Morse code, on 3.58, 7.08, 14.08, 21.08 and 28.08 MHz, as well as vhf frequencies. For a complete W1AW operating schedule, send an SASE to ARRL, 225 Main St., Newington, CT 06111.

Commercial CW stations that transmit ship-maritime, press, and weather messages are also valuable sources of code practice. Table II is a list of eight shortwave stations that you'll want to tune in if you have a general-coverage receiver. Whether you plan to confine your listening to the ham bands or branch out into the other shortwave frequencies, remember that a good antenna and a sensitive, stable communications receiver play key roles in good CW reception.

One final note—international radio regulations prohibit the disclosure to others of *any* information gleaned from press or commercial transmissions. Accordingly, it is illegal to pass information so obtained (except that learned from amateur radio or broadcast transmissions) to a third party. ◇

TABLE II
SOME COMMERCIAL CW STATIONS

Station	Frequency (MHz)	Traffic
NSS	8.090	Weather Bulletins
WSL	8.516	Commercial Press
WAX	8.527	Messages
WCC	8.587	Messages
WCC	8.590	Messages
NBA	8.617	Telegrams
NAM	12.134	Weather Bulletins
NMN	12.720	High Speed CW
NAWS	12.725	Weather Bulletins
WCC	12.926	Messages
WSL	13.026	Messages
NAWS	15.921	Messages
WCC	16.935	Messages

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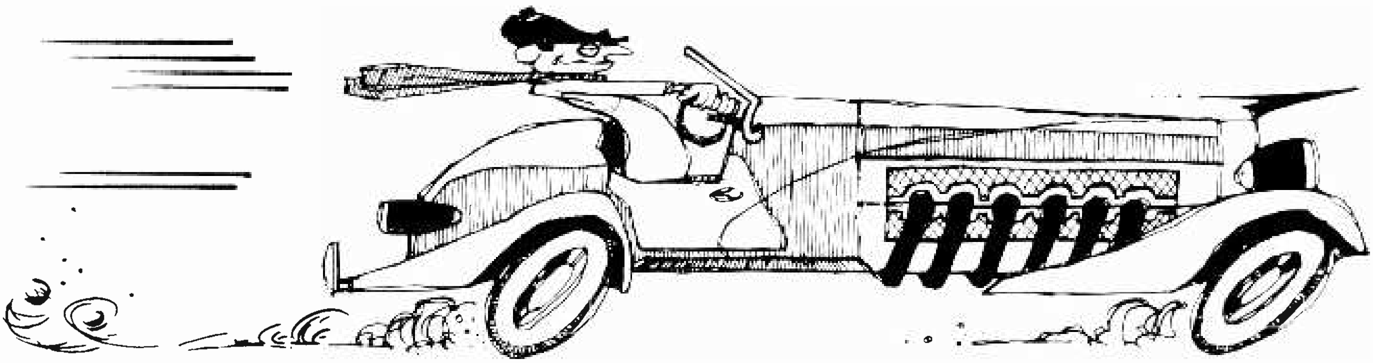
BY ROBERT P. BISEY

BUILD "CRUISEALERT" A 55 MPH SPEED-LIMIT ALARM



Automobile add-on device for highway safety

TO STAY within the 55-mph national highway speed limit, one must keep one eye on the road and the other on the speedometer. This can be a dangerous situation at highway speeds when your whole attention should be fixed on the road. It would be far safer, therefore, if you could keep a constant eye on the road and have some audible means for alerting you when you have exceeded the speed limit. This is exactly what the "Cruisealert" described here was designed to do.

The Cruisealert works on the principal that, with a given vehicle, there is a close relationship between engine rpm and road speed. It constantly monitors engine rpm and is preset to sound an alarm when engine rpm reaches a value that causes your vehicle to travel at 55 mph (or some selected lower speed). When this happens, the Cruisealert sounds a beeper to alert you that you are at the legal speed limit. At no time do you have to take your eyes from the road. And the Cruisealert can be used with 4-, 6-, and 8-cylinder engines.

Circuit Operation. A schematic diagram of the Cruisealert is shown in Fig. 1. Components *R1, R2, C1*, and *D1* both filter and clip the raw signal coming from the engine's distributor contacts. Resistor *R1* and capacitor *C1* form a single-stage low-pass filter that has a time constant of about 1.5 ms, which is long enough to provide smoothing for the transient, oscillatory-like waveforms present at the points. The frequency range is between 40 and 170 Hz, which approximately corresponds to a four-cylinder engine at a road velocity of about 30 mph and an eight-cylinder engine at approximately 70 mph.

Zener diode *D1* clips the input voltage swing to approximately +7 and −0.7 volts, suitable for use by the following circuitry.

Positive-edge retriggerable monostable multivibrator *IC1A* functions as a frequency discriminator, while *IC1B* forms the annunciator section. The filtered and limited signal from the input filter is applied to *IC1A* via input current limiting resistor *R2*. This portion of the dual multivibrator is arranged to deliver an output pulse at pin 6 when triggered by a positive spike. The pertinent waveforms for *IC1A* are shown in Fig. 2.

Resistors *R5* and *R6*, potentiometers *R12* and *R7*, and capacitor *C3* control the on time (T_{on}) of the multivibrator. For the three relationships shown in Fig. 2, the on time of *IC1A* remains constant, regardless of the input frequency, while the off time (T_{off}) changes with the input frequency. As the input frequency increases and approaches the threshold frequency of the multivibrator, T_{on} remains constant while T_{off} diminishes. At the critical threshold frequency, T_{off} diminishes to zero. The resulting output is a constant logic 1 as shown in Fig. 2C.

Diode *D3*, resistor *R13*, and capacitor *C5* form a negative-going integrating pulse detector. As long as the cathode of *D3* (pin 6 of *IC1*) remains at logic 0, *C5* remains fully charged.

For all input frequencies lower than the threshold frequency of the multivibrator, a negative-going T_{off} signal appears at pin 6 of *IC1A*, which forces *C5*'s

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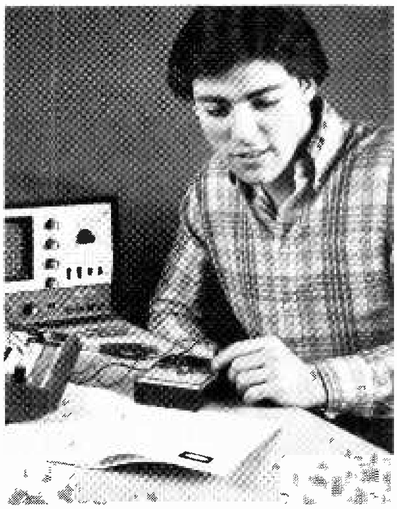

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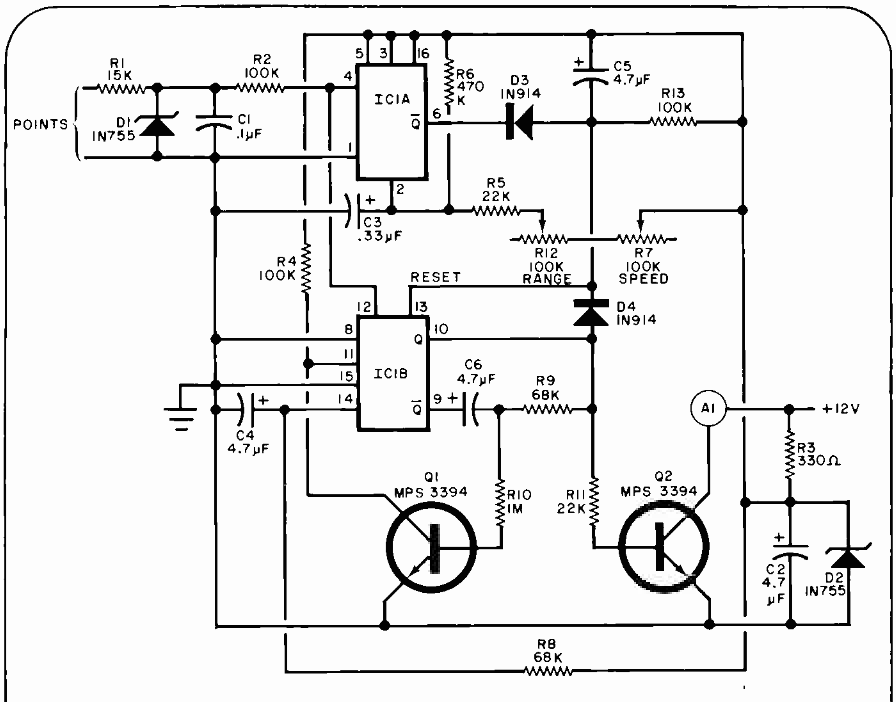

Fig. 1. Frequency discriminator IC1A triggers IC1B to sound alarm when input frequency from distributor points exceeds predetermined limit.

PARTS LIST

A1—SNP Sonalert or similar alarm
C1—0.1-µF, 50-V tantalum
C2, C4, C5, C6—4.7-µF, 50-V tantalum
C3—0.33-µF, 50-V tantalum
D1, D2—1N755 (7.5-V, 400-mW) zener
D3, D4—1N914 switching diode
IC1—MC14528CP dual monostable multivibrator
Q1, Q2—MPS3394 or similar transistor
Following are ½-W, 10% resistors unless otherwise noted:
R1—15,000 ohms
R2, R4, R13—100,000 ohms
R3—330 ohms
R5, R11—22,000 ohms
R6—470,000 ohms
R7—100,000-ohm panel-mount potentiometer
R8, R9—68,000 ohms
R10—1 megohm
R12—100,000-ohm, pc-mount potentiometer
Misc.—4" x 2¼" x 2¼" (10.2 x 5.7 x 5.7 cm) box; control knob; dry-transfer lettering kit; 16-pin IC socket (optional); hookup wire; solder; machine hardware; etc.
Note—A complete kit of parts is available for $29.95 from EALAB Associates, Box 737, Smithtown, NY 11787.

negative terminal to near ground potential. When the input frequency exceeds the critical threshold frequency, the voltage step T_{off} disappears and becomes a logic 1 (Fig. 2C). At this instant, diode *D3* then becomes reverse biased, causing the negative side of *C5* to rise towards the +V through *R13*.

Retriggerable monostable multivibrator *IC1B* and transistors *Q1* and *Q2* form the annunciator section. The main triggering input at pin 12 responds only to voltage transitions, while master reset at pin 13 responds to dc levels. In this circuit, *IC1B* is arranged so that to initiate astable action, a constant ac trigger signal must be present at pin 12. This is accomplished by connecting this input to the filtered and clipped ac signal source generated by the distributor points.

When the input frequency is below the discriminator's threshold frequency, the negative end of *C5* is near ground. Since this point is connected to pin 13, a logic 0 at this input forces *IC1B* to assume a reset condition in which the Q and not-Q outputs are held at logic 0 and logic 1, respectively. At this time, *C6* (connected between the two outputs via resistor *R9*) is fully charged to the voltage difference between the two outputs. The logic 0 level at Q also holds *Q1* in the off condition via *R9* and *R10*. The collector of *Q1* is held at a logic 1 to allow the input pulses at pin 12 to trigger *IC1B*.

When the input frequency rises above the discriminator's threshold frequency, the voltage at the negative end of *C5* assumes a positive (logic 1) potential. The logic 1 at pin 13 causes *IC1B* to be triggered by the input pulse train present on pin *12*. When triggered, the Q and not-Q outputs change state with a logic 1 and logic 0 appearing at the Q and not-Q outputs, respectively.

The logic 1 at the Q output turns *Q2* on via *R11*, which activates alarm *A1*. At this time, the voltage at the junction of *C6* and *R9* instantly drops below ground and then gradually rises above ground due to the charging current through *R9* whose source is the logic 1 at the Q output. When this voltage eventually rises above 0.7 volt above ground (one diode drop), *Q1* switches on and its collector drops to ground level. By virtue of logic-gate action, a logic 0 at pin 11 inhibits the input pulse stream at pin 12 from further triggering the multivibrator. In the absence of triggering pulses, the multivibrator eventually times out as determined by the *C4/R8* time constant.

The subsequent change of state at the Q and not-Q outputs causes *Q2* to switch off, silencing the alarm. Since *R9* now "sees" a logic 0 source at the Q output, the voltage at the *C6-R9* junction eventually drops to ground potential. When this junction reaches 0.7 volt, *Q1* turns off and its collector assumes a logic 1 state via *R4*. This allows the pulse train at pin 12 to once again trigger the multivibrator. It is in this manner that the astable action of *IC1B* is sustained only when the master reset at pin 13 is maintained at logic 1. The waveforms associated with *IC1B* are shown in Fig. 3.

Hysteresis Dead-Band Circuit. The frequency of the mechanical cam-rotor points breaker system used in the majority of engines is inherently unstable. Even if the engine's rpm were to be held absolutely constant, careful examination of the instantaneous frequency of the points would reveal some frequency modulation. This is due to a variety of factors such as a bent distributor shaft, variations in machining tolerances of the cam lobes, and, most of all, badly burned points.

Since the Cruisealert functions solely as a frequency discriminator, frequency modulation of the breaker points can lead to random and erratic triggering. To make the circuit immune to small incremental frequency variations, diode *D4* was added. It's function is to increase and hold the dc voltage at the negative end of *C5* when *IC1B* is operating (Q output is at logic-1). The result of this addition is illustrated in Fig. 4, which depicts the relationship between the point frequency and the alarm state. Examination of this chart shows that the alarm's turn-on frequency is slightly

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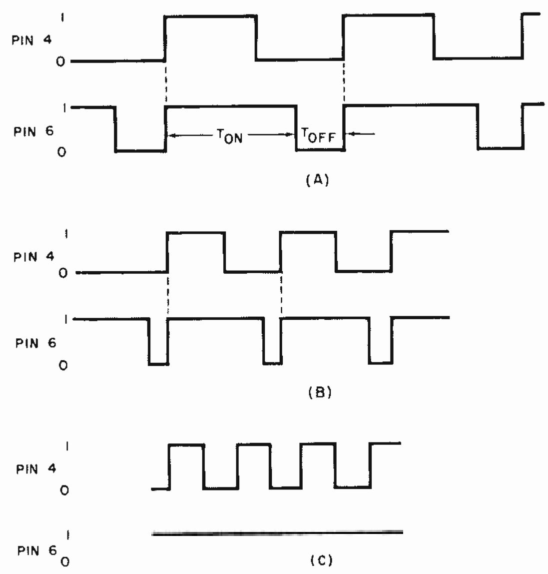

Fig. 2. Waveforms for IC1A show how off time of output (pin 6) varies as frequency of input at pin 4 increases.

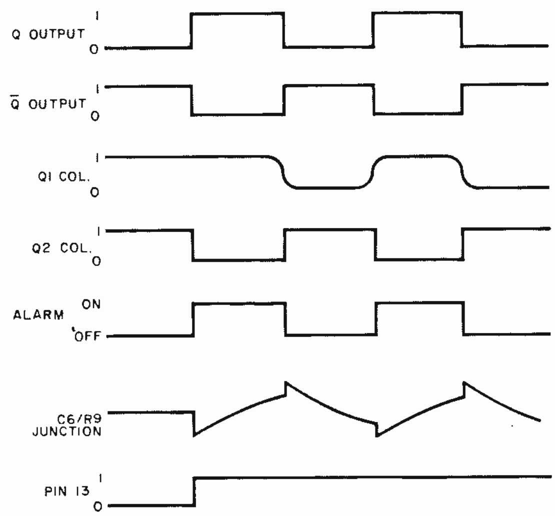

Fig. 3. Timing diagram for IC1B shows waveforms which can be expected at various points in circuit.

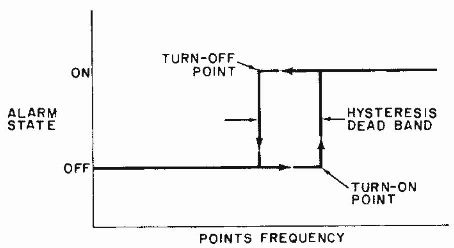

Fig. 4. Hysteresis of input frequency versus alarm state. Difference is less than 2 mph.

greater than its turn-off frequency. The difference between these two frequencies defines the hysteresis deadband, which in terms of vehicle road velocity is less than 2 mph.

Construction. The circuit can be built on a printed circuit board, the etching-and-drilling and components-placement guides for which are shown in Fig. 5. Note that SPEED control potentiometer *R7* and the alarm are both mounted on the box in which the circuit is housed.

After *R7* is mounted, attach a pointer knob to its shaft and provide some kind of marking surface below the knob. Starting at the fully counterclockwise position, mark off 10 equally spaced points to the clockwise limit stop.

Drill a small hole in the Cruisealert's front panel so that trimmer adjust potentiometer *R12* can be reached with a screwdriver after the pc board is in place. Connect the alarm and *R7* to the pc board as shown in Fig. 5. Then connect three long insulated leads to the vehicle's electrical system to provide input.

Select a suitable mounting position in the vehicle. Route the INPUT lead through the firewall and connect it to the screw connector of the ignition coil that goes to the distributor points. Connect the GROUND lead to a convenient metal screw or bolt and the 12-volt lead to a switched +12-volt source, such as the lead that feeds the radio. Insulate all connections.

Photo at right shows the author's prototype with speed control potentiometer R7 at left and loudspeaker at right. The hole in center is to gain access to R12 in making final adjustments with passenger's aid.

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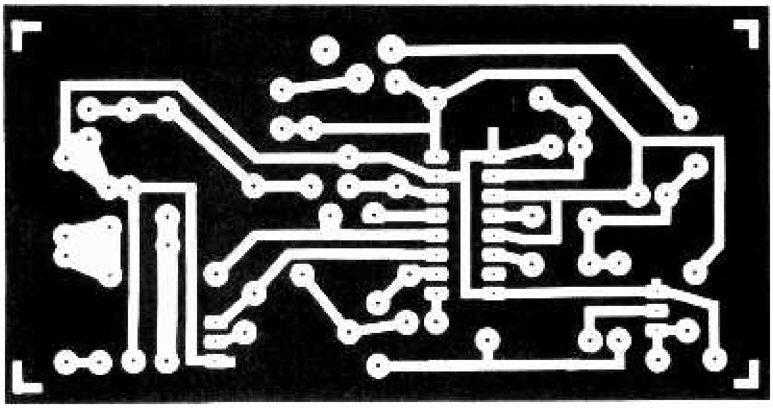
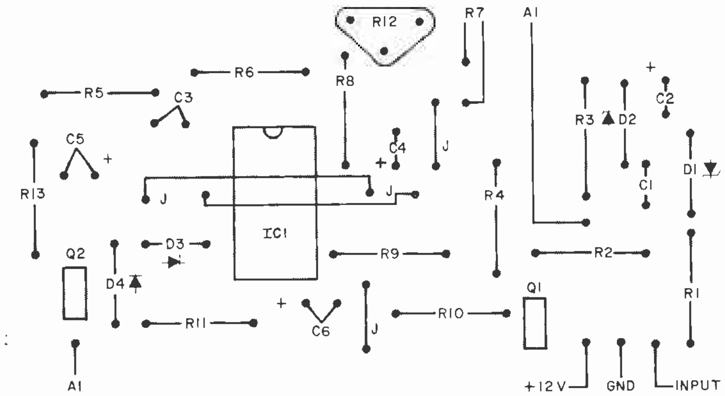


Fig. 5. Actual size etching and drilling guide (above) and components placement (below) for printed circuit board.



Adjustment. The Cruisealert is designed to provide an overspeed alarm indication for selected road speeds between 30 and 70 mph for a four-, six-, or eight-cylinder engine. In the interests of safety, it is recommended that the following adjustments be made by a passenger and not the driver.

Using a small screwdriver, rotate *R12* (RANGE) fully clockwise via the access hole in the front panel and leave the screwdriver engaged in the trimmer slot. Set the front-panel SPEED control (R7) knob to the fifth mark on its scale.

Drive the car until the speedometer indicates 55 mph and try to maintain this speed. Very slowly adjust *R12* until the Cruisealert just starts to beep. Then remove the screwdriver. This completes the range setting, and the SPEED control is set to 55 mph. Note that the SPEED control's scale indications are only relative and do not correspond to vehicle speed.

To preset the Cruisealert to operate at another road speed, rotate the SPEED control fully clockwise, drive the car at the desired speed, and while maintaining this speed, slowly adjust the SPEED knob until the alarm sounds. ◇

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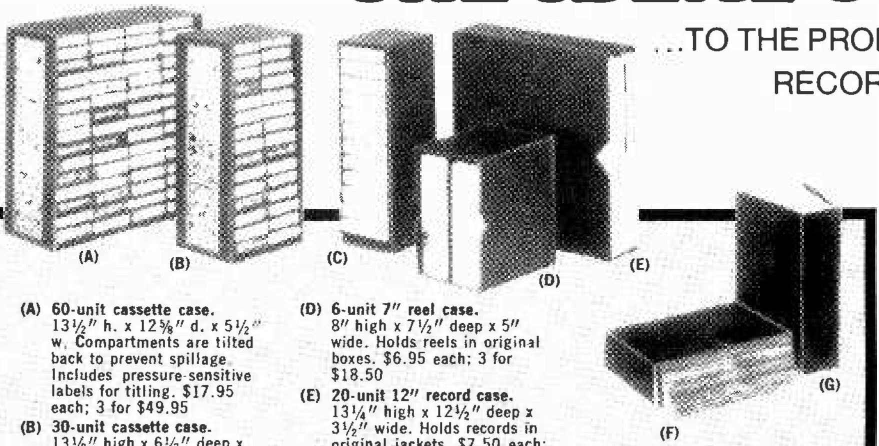

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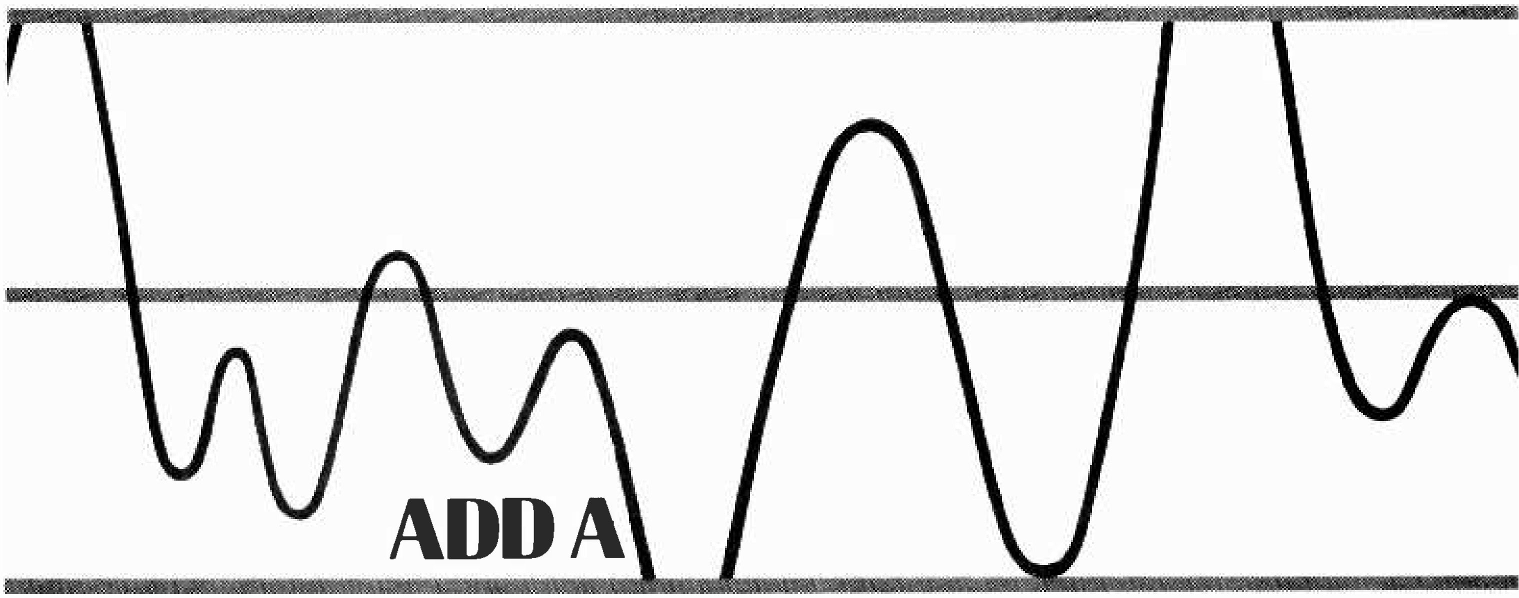

ADD A CLIPPING INDICATOR TO YOUR AUDIO AMPLIFIER

To protect speakers, this simple circuit senses power supply voltages and flashes a warning LED just before the onset of clipping

BY NORMAN PARRON

THE CONSEQUENCES of overdriving an audio power amplifier can range from the unpleasant (ragged, distorted sound) to the catastrophic (burnt, black remains of tweeters and supertweeters). It's obvious, therefore, that the audiophile will want to avoid this condition. The project presented here, an Amplifier Clipping Indicator, will help him do just that. It continually senses both the audio output of the amplifier and the power supply voltages, and flashes a warning LED if the output signal voltage approaches either power supply rail. The user can then reduce the drive level so that the LED stops flashing.

Readily available, inexpensive components comprise the Amplifier Clipping Indicator. Many of them will be found in an experimenter's "junk box." A stereo version can be built in just a few hours, making the Amplifier Clipping Indicator an enjoyable weekend project. The modest amount of power the circuit requires can be tapped from the power amplifier's supply or furnished by a small supply built especially for this purpose.

What Is Clipping? When an audio amplifier is overdriven, it "clips" the input signal. The process is shown graphically in Fig. 1. A power amplifier is driven by a sinusoidal input signal having maximum positive and negative amplitudes of $+V_{IN}$ and $-V_{IN}$, respectively (Fig. 1A). The amplifier generates an output signal that is (ideally) an exact replica of the input except for its increased amplitude.

Because the amplifier must reproduce ac waveforms, it employs a bipolar dc power supply. This means that the most positive voltage it can produce at the output terminals is $+V_{CC}$, and the most negative voltage is $-V_{CC}$. If the amplifier's gain control is adjusted so that the output signal approaches the limits imposed by the power supply, a waveform like that shown in Fig. 1B is generated. It can be seen that the maximum positive and negative swings of the output voltage, $+V_{OUT}$ and $-V_{OUT}$, are somewhat less than the absolute limits of $+V_{CC}$ and $-V_{CC}$.

Adjusting the control for more gain causes the amplifier to attempt to exceed the constraints of the power supply. The result is a clipped waveform like that shown in Fig. 1C. Spectral analysis of such a waveform indicates the presence of high-order harmonic distortion products during the interval that clipping takes place. If the output signal is clipped less than 1% of the time, the effect is usually inaudible. As the duration of clipping approaches 10%, the usual consequence is audible, "raspy" distortion. A severely clipped signal (more than 10% of the time) contains a considerable amount of high-frequency energy. This energy poses a significant threat to midrange and high-frequency drivers because it is directed to them by the crossover network and they are usually capable of dissipating far less power than bass drivers.

Although the example that has been discussed used sinusoidal signals, an audio amplifier usually processes musical signals that are much more complex. It is characteristic of most recorded music that the average signal level is low. However, musical program material does contain a significant number of

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short-lived, high-level transients. An amplifier might be called upon to deliver one watt of output power on an average basis, but accurate reproduction of a bass percussion transient can require fifty to one-hundred times that power level for a brief instant.

All is well if the amplifier has enough voltage and current reserves to pass the transient unclipped. However, if the amplifier cannot do so, the dynamic range of the recording will be compressed and audible distortion products introduced. This, coupled with the fact that perceived loudness is a function of average (as opposed to peak) power, explains the trend toward power output capabilities that were unheard of in audio amplifiers a relatively short time ago. So-called "super-power" amplifiers allow the audiophile to listen to program material at realistic levels without clipping high-level transients, even if inefficient speakers are used.

About the Circuit. The Amplifier Clipping Indicator is shown schematically in Fig. 2. Each channel of amplification in a sound system will require a separate indicator circuit. The most common application for the project is in a stereo system, so component numbers for two channels are shown. Those for the right channel are given in parentheses. The discussion that follows pertains to only one channel, designated the left channel of a stereo pair. Everything that will be said, however, applies equally to as many channels as are needed because the indicator circuit is identical for each.

Output signals from the audio amplifier are applied to an 11:1 voltage attenuator (*R1R3*). Similarly, the positive and negative supply voltages, $+V_{CC}$ and $-V_{CC}$, are applied to attenuators *R5R7* and *R9R11*. The voltage dividers associated with the power-supply outputs, however, employ trimmer potentiometers and have variable attenuation factors. Those portions of the input voltages passed by the attenuators are applied to two 741 operational amplifiers (*IC1A* and *IC1B*) employed as voltage comparators.

Assume that the trimmer potentiometers have been adjusted to attenuate the power supply voltages slightly more than the fixed divider attenuates the audio signal. If the amplifier is being driven by an audio signal, but not to the point of clipping, its output voltage will be smaller in magnitude than either the positive or negative supply voltage. This means that the voltage applied to the noninverting input of *IC1A* is never more positive than that applied to the inverting input, and the output of the comparator remains at −12 volts. Similarly, the voltage applied to the inverting input of *IC1B* remains positive with respect to that present at the noninverting input, keeping the output of *IC1B* at −12 volts.

Diodes *D1* and *D3* form an OR gate whose output goes to +12 volts when either of the comparator outputs does. In the absence of clipping, both *D1* and *D3* are reverse-biased, which keeps transistor *Q1* cut off. Monostable multi vibrator *IC3* remains untriggered and its output (pin 3) is at ground potential. This keeps *D7*, which together with *D5* forms a second diode OR gate, in a nonconducting state. The output of the *D1D3* OR gate is applied to the *D5* input of the second gate. Both inputs are low, so *Q3* receives no base drive and the clipping indicator LED (*LED1*) remains dark.

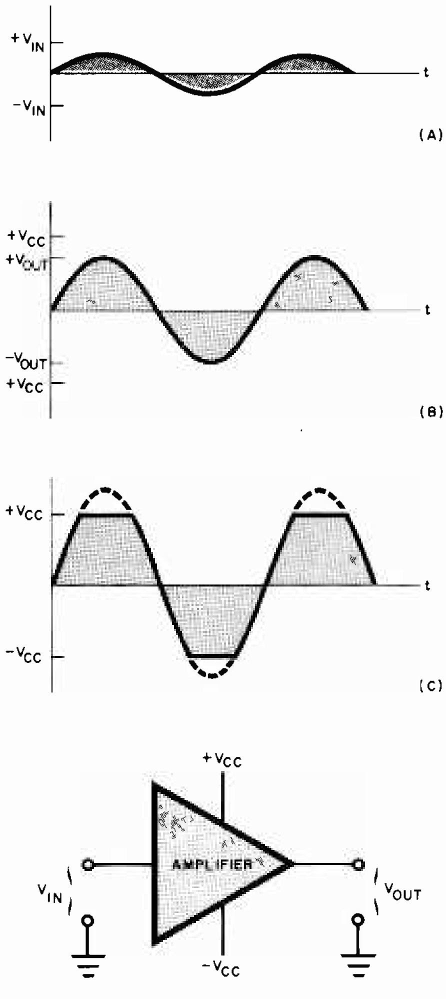

Fig. 1. If input amplitude (A) or gain of amplifier is not excessive, output is not clipped (B). Increasing one or both causes amplifier to clip the output (C).

Now let's assume that the audio amplifier is driven into clipping. The audio output voltage reaches the positive or negative supply voltage (or both) and is clipped like the one shown in Fig. 1C. When the positive portion of the audio waveform applied to the noninverting input of *IC1A* becomes more positive than the voltage at the inverting input, the output of the comparator goes to +12 volts, this forward-biases *D1* and *D5*, and provides base drive for *Q1* and *Q3*. A similar thing happens when the negative portion of the audio waveform is clipped. The voltage applied to the inverting input of *IC1B* becomes more negative than the voltage at the noninverting input, so the output of this comparator switches to a +12-volt level. This forward biases *D3* and *D5*, providing base drive for *Q1* and *Q3*.

When *Q3* is supplied with base current, it turns on and the clipping indicator LED glows. However, the clipping interval can be so short that the eye will not readily detect the brief flash of the LED. That's why *Q1*, *IC3*, and their associated components have been included. Together they function as a pulse-stretching circuit. Here's how.

When the output of either comparator goes high, *Q1* receives base current and its collector drops to ground potential. A negative pulse is passed by *C1* to pin 2 of *IC3*, triggering this monostable multivibrator. The output of the timer IC (pin 3) goes high for an interval determined by the time constant of *R19C5*. For the values given, the width of the output pulse is about 0.25 second. This output pulse is OR'ed with the output of gate *D1D3* and applied to resistor *R21*. Transistor *Q3* receives base drive and sinks current for *LED1*, causing the clipping indicator LED to glow.

The pulse-stretcher turns the LED on for one quarter of a second even if the clipping interval is much shorter. A subsequent trigger pulse received while the monostable is timing will not retrigger it. However, one received immediately after a timing cycle will cause the process to be repeated. If the clipping interval is longer than the width of the output pulse (which can be extended to any desired interval by increasing the value of *R19* or *C5* or both), the OR'ing action of *D5* and *D7* will keep *Q3* in a conducting state. Therefore, the clipping indicator LED will continue to glow even after the output of the monostable has returned to its ground state. It will glow un-

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til the audio amplifier recovers from the clipping condition.

The project requires a bipolar power supply of ±12 volts dc. These operating voltages can usually be tapped from the audio amplifier's power supply. Zener diodes and series current-limiting resistors can be used to drop the amplifier's $+V_{CC}$ and $-V_{CC}$ supply voltages to the desired values. Alternatively, a small line-powered supply can be built into the project's enclosure. Current demand is relatively modest—a few milliamperes for the −12-volt supply and about 50 mA from the positive rail.

Because dynamic voltage comparison is the method employed to sense clipping, this project enjoys a significant advantage over such power-monitoring devices as peak-reading meters and strings of LEDs. A peak-reading meter only indicates that the audio output has reached a given level. It will not necessarily indicate that clipping is taking place. For the sake of illustration, let's consider what happens to an amplifier with an unregulated power supply when it is driven by an audio signal with many high-level transients.

Suppose that our amplifier can deliver 75 watts per channel of continuous power to 8-ohm loads and has an IHF dynamic headroom of 2.04 dB. This means that it can deliver 120 watts of output power into 8 ohms for brief intervals. Consequently, the power supply voltages under full load are +34.6 volts and −34.6 volts. When the demand on the power supply is light, the available voltages are +43.8 and −43.8 volts.

If the supply's filter capacitors have charged up to these higher voltages and a short-lived, high-level transient arrives at the amplifier's audio input, the output stage can momentarily generate an 87.6-volt peak-to-peak waveform without clipping it. However, driving the amplifier this hard causes the voltages across the filter capacitors to decrease. If the amplifier is called upon to reproduce a second high-level transient before the filter capacitors have had an opportunity to recharge sufficiently, clipping will result.

It can thus be seen that a peak-reading audio power meter will not *necessarily* indicate that the amplifier is clipping. In our example, the lowest possible power supply voltages are +34.6 and −34.6 volts, so we can safely say that any audio output signal with a peak power of up to 75 watts as indicated on the peak-reading monitor will not be clipped. Above that power level, however, the meter reading alone will not tell us whether clipping is taking place. By contrast, a flash of the indicator LED in this project warns of the onset of clipping, a warning which takes into account the dynamics of the amplifier's power supply.

Construction. Either printed circuit or perforated board can be used in the assembly of the Amplifier Clipping Indicator. In any event, the use of IC sockets is recommended. Be sure to use the minimum amount of heat and solder consistent with the formation of good solder joints. Also, observe the polarities and pin basings of semiconductors and electrolytic capacitors.

After the project's circuit board has been completed, connect it to *BTS1* and the indicator LED(s) with suitable lengths of hookup wire. Then secure the board to the project enclosure with standoffs and machine hardware. Mount *BTS1* on the rear panel of the enclosure

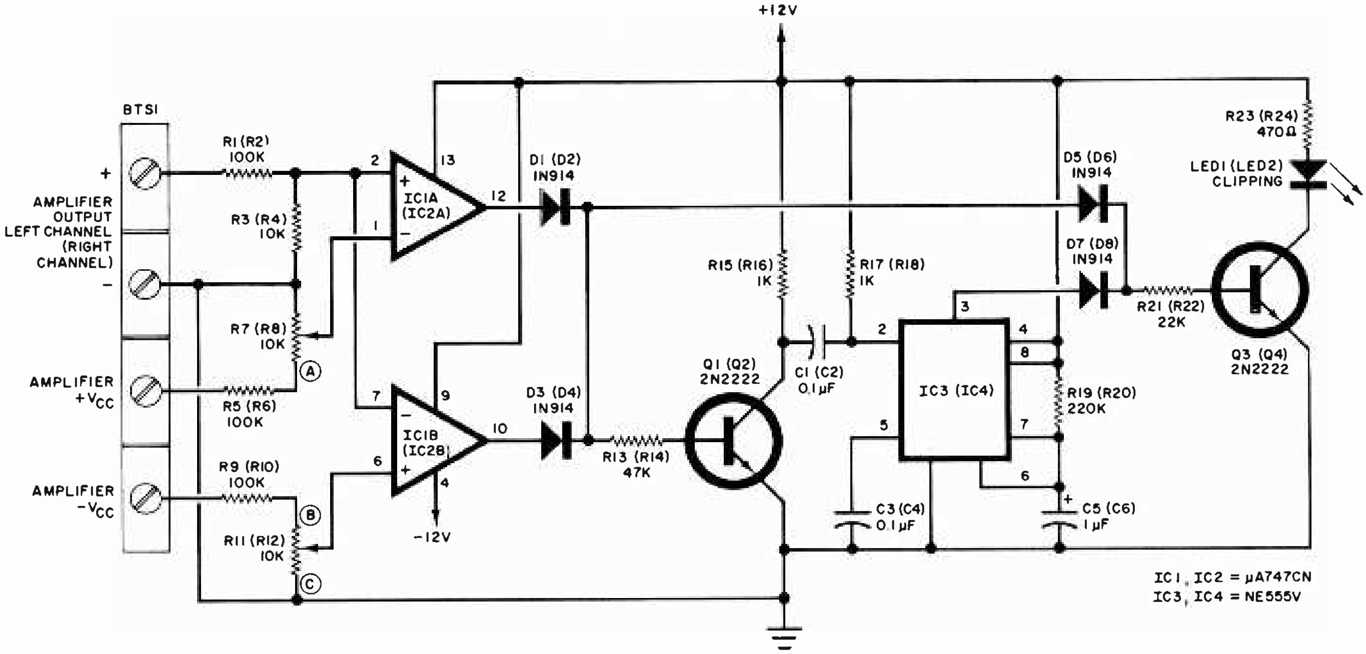

Fig. 2. Schematic of the clipping indicator circuit for one channel.

PARTS LIST

BTS1—Four-position (Six-position) barrier terminal strip
C1 through C4—0.1-µF disc ceramic
C5,C6—1-µF tantalum
D1 through D8—1N914
IC1, IC2—µA747CN dual operational amplifier
IC3,IC4—NE555V timer
LED1,LED2—Light emitting diode
Q1 through Q4—2N2222
The following are ¼-watt, 5% tolerance carbon composition fixed resistors unless otherwise specified.
R1,R2,R5,R6,R9,R10—100,000 ohms
R3,R4—10,000 ohms
R7,R8,R11,R12—10,000-ohm, linear-taper trimmer potentiometer
R13,R14—47,000 ohms
R15,R16,R17,R18—1000 ohms
R19,R20—220,000 ohms
R21,R22—22,000 ohms
R23,R24—470-ohms, ½-watt, 10% tolerance
Misc.—Suitable enclosure, printed circuit or perforated board, bipolar 12-volt power supply, IC sockets or Molex Soldercons, LED mounting collars, machine hardware, hookup wire, solder, etc.

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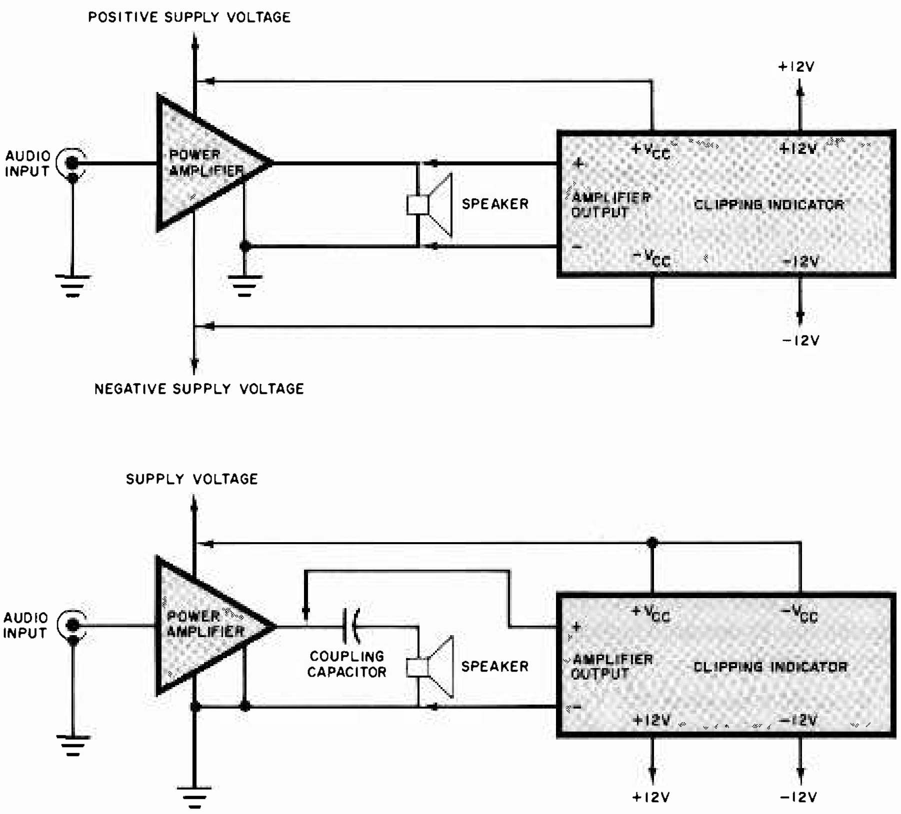

Fig. 3. Diagram showing details of interconnection for amplifiers with bipolar (A) and single-ended (B) power supplies

with machine hardware and the indicator LED(s) on the front panel with rubber grommets or mounting collars made especially for this purpose. As mentioned earlier, operating power for the project can be obtained from a small supply included inside the enclosure or tapped from the amplifier itself if a bipolar dc supply is employed. If the latter approach is taken, the required zener diodes and series resistors will easily fit inside the project enclosure.

Another possible approach, if there is room in the amplifier chassis, is to mount the entire project inside the amplifier and locate the indicator LEDs on the front panel. If this is done, *BTS1* can be eliminated and the connections to the speaker outputs and $+V_{CC}$ and $-V_{CC}$ hard-wired.

Note that the circuit as shown will function properly with audio amplifiers having supply voltages of up to ±60 volts (or +80 or −80 volts in the case of an amplifier with a single-ended supply). That bipolar voltage corresponds to a clipping power of 225 watts into 8 ohms. The project is therefore useable with the vast majority of audio amplifiers commercially available. If you have an amplifier employing greater supply voltages, the circuit can be suitably modified simply by increasing the attenuation factors of the input voltage dividers (increasing the values of *R1*, *R5*, and *R9*).

Interconnection and Adjustment. If your audio amplifier employs a bipolar dc power supply (most do), connect the $+V_{CC}$ and $-V_{CC}$ terminals of *BTS1* to the power supply outputs inside the amplifier. (Note that making these connections will, in most cases void the warranty on your amplifier.) Also, connect the AMPLIFIER OUTPUT terminals of *BTS1* to the amplifier's speaker output terminals in agreement with the polarities indicated in Fig. 2. These connections can be made with standard "zip-cord" or speaker wire. Refer to Fig. 3A for details.

Slightly different connections should be made if your audio amplifier employs a single-ended power supply and a coupling capacitor or transformer between the final amplifying devices and the speaker output terminals. The required connections are as follows: connect the $+V_{CC}$ *and* $-V_{CC}$ terminals of *BTS1* to the "hot" side of the power supply output; and connect the "hot" AMPLIFIER OUTPUT terminal of *BTS1* to the "hot" side of the amplifier output *before* the output coupling (dc blocking) capacitor or transformer. Refer to Fig. 3B.

The circuit's trimmer potentiometers can now be adjusted. Referring to Fig. 2, note the points near *R7* and *R11* designated *A*, *B*, and *C*. If your audio amplifier has a bipolar power supply, adjust the wiper of *R7* so that it is at position *A* and the wiper of *R11* so that it is at position *B*. If your amplifier's power supply is single-ended, adjust the wiper of *R7* so that it is at position *A* and the wiper of *R11* so that it is at position *C*.

Two pieces of test equipment are needed to adjust the trimmer potentiometers properly. The first is a sine-wave generator whose output is of sufficient amplitude to drive the audio amplifier into clipping. (One volt peak-to-peak of drive signal is usually more than adequate.) The second item can be either an oscilloscope or a multimeter, but the former is preferred. We will first describe the procedure to be followed if an oscilloscope is available and then that to be employed if one is not.

Connect a patch cord between the output of the signal generator and the input of the audio amplifier. Then connect the probe running from the oscilloscope's vertical amplifier input to the audio output of the power amplifier. Apply power to the project, signal generator and audio amplifier. Then adjust the amplitude of the generator's output, the gain of the audio amplifier, and the various oscilloscope controls for a stable, sinusoidal trace. The output of the audio amplifier should not be connected to a speaker.

Increase either the gain of the amplifier or the amplitude of the generator output until the oscilloscope trace just begins to reveal clipping of the waveform. Then decrease either the amplifier gain or signal output so that the amplitude of the waveform decreases a few volts below each clipping limit. (This provides a small safety margin so that the indicator LED will start to flash just before clipping actually begins.)

Without disturbing the amplifier, generator, or oscilloscope control settings, adjust trimmer *R7* until the LED starts to flash on positive signal peaks. Make a pencil mark on the circuit board denoting the correct position of the wiper and then return the control to its original setting. Next, adjust *R11* so that the indicator LED starts to flash on negative signal peaks. Once the correct setting of *R11* has been found, don't disturb it. Return to *R7* and adjust its wiper so that it corresponds to the position marked on the circuit board. Decrease the amplitude of

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the generator output or the gain of the amplifier, noting that the indicator LED will be extinguished. If you have built more than one Amplifier Clipping Indicator, say, for use with a stereo or four-channel audio amplifier, repeat the procedure just described for each.

Those who do not have acess to an oscilloscope can use a VTVM, VOM, or similar multimeter to adjust the project. First, the power supply limitations of the amplifier with which the project will be used must be determined. Connect the signal generator to the amplifier as described above and adjust the generator for a 60-Hz output. Connect the amplifier's speaker output to an 8-ohm load (a resistor is best) and apply a moderate amount of drive to the amplifier input. With the power supply loaded, measure its output voltage(s). Increase the gain of the amplifier or the amplitude of the drive signal and note whether the power supply voltages decrease. If they do, measure the *minimum* values.

Having performed these measurements, determine the peak-to-peak voltage swing that the output can generate. For example, if the minimum voltages that a bipolar power supply generates under maximum drive conditions are +30 and −30 volts, the continuous peak-to-peak signal that the amplifier can pass at the onset of clipping is 60 volts p-p. Next, calculate the rms output voltage using the equation $V_{rms} = V_{p\text{-}p}/2.828$. For our example, the rms output voltage is 21.2 volts.

Connect the multimeter probes across the 8-ohm load and adjust the amplifier's gain or the amplitude of the input signal so that the calculated rms voltage is indicated by the meter. Then decrease the gain or the drive signal so that the meter reading is a few volts below the calculated value. (This provides the safety margin previously discussed.) Now adjust the trimmer potentiometers in the same manner described in the procedure employing the oscilloscope. Repeat the procedure for each additional channel of amplification (if any).

Use. The Amplifier Clipping Indicator is now ready for use. With it, you'll be able to adjust drive level and/or amplifier gain so that your amplifier will never go into heavy clipping. Keep in mind that the indicator LED will begin to flash slightly before the the onset of clipping. If the LED starts to blink, back off on the drive level or gain control. Your high-frequency drivers will be glad you did! ◇

Audio Artist . . .

(Continued from page 33)

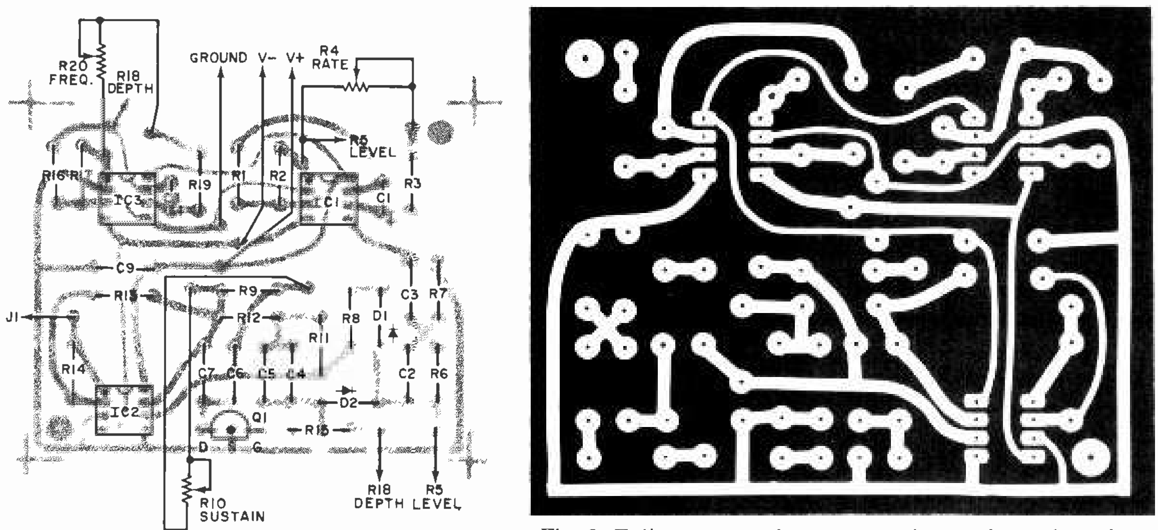

Fig. 2. Foil pattern and component layout for pc board.

to double check your wiring to catch any errors that might have inadvertently been made.

Use. Patch signals from the output jack of the Audio Artist to an audio amplifier which in turn drives a loudspeaker or pair of headphones. Depending on the settings of the Audio Artist's controls, the peak voltage across the output jack can vary from less than one to nine volts. To avoid overloads, apply drive to a line-level input and initially keep the volume low.

Apply power to the Audio Artist and the amplifier and adjust the amplifier's gain control for a comfortable listening level. Setting the SUSTAIN control at its minimum position will reduce the output signal to zero.

Begin to experiment with the Audio Artist by rotating the wiper of the SUSTAIN potentiometer to a maximum of midscale and the wipers of the other controls to their maximum settings. Slowly vary the settings of the RATE and SUSTAIN potentiometers. Vary each control in turn, noting how it affects the sound generated by the project. You will quickly be creating unusual sound effects, and will be surprised to discover how many different sounds the Audio Artist is capable of producing. ◇

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During the heating season, you can feel warmer at 65°F, if you humidify the heated air.

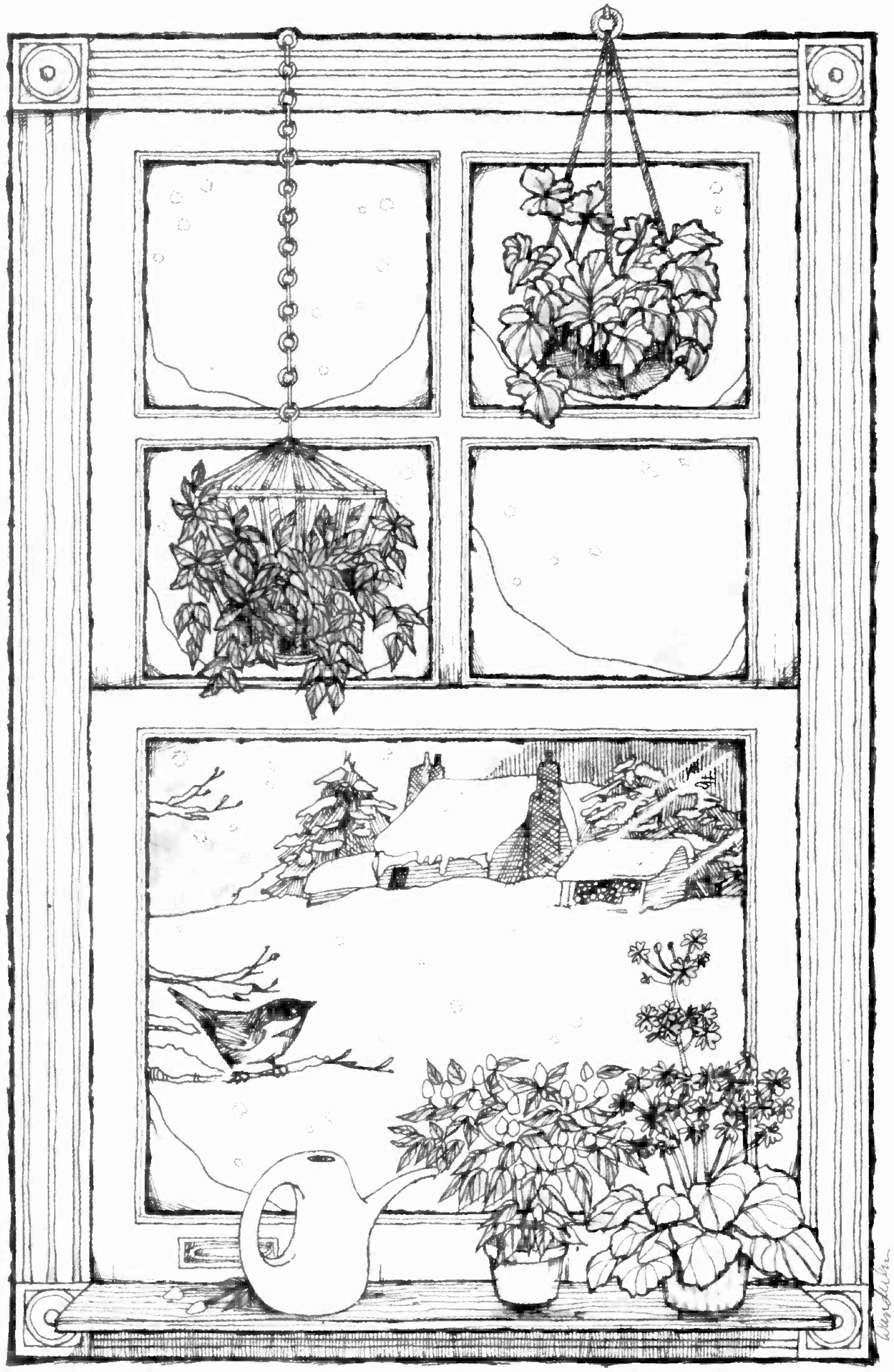
IN THESE days of high energy costs, each of us must pitch in to help reduce energy consumption. In particular, if you are a home owner, you can help simply by turning down the thermostat. At the same time, however, you will want to turn up your "comfort" level. One way to do this is to increase the relative humidity in your home by modernizing the humidification system in your warm-air furnace. The Solid-State Humidity Control described here does just that, economically and reliably.

This automatic humidity control is an easy-to-build, low-cost device that couples to a solenoid-controlled water valve and special humidifier spray nozzle. (Spray nozzles of various water capacities are readily available at many hardware and plumbing-supply outlets at reasonable cost.) To reduce parts cost, you can use a solenoid valve recovered from an old washing machine, if available. The spray nozzle itself mounts anywhere in the direct stream of the heated air so that the fine water mist turns to vapor.

Why Humidify? All heating systems, whether warm-air, hot-water, or steam, should have some form of humidification to promote a healthy environment for you, your family, pets, furnishings, plants, etc. This is especially important during the winter months when the air inside of the home is closed off from the outside air and is heated.

When air is heated, its relative humidity decreases from a healthy 30–50% to as low as an arid 10–20%. The latter is far too low for comfort. It irritates sensitive membranes (which accounts for all those sore throats you get during the winter); is a poor conductor of electricity (hence, the build-up of annoying static-electricity charges on your body when you cross a rug); and makes temperatures in the 50° to 70° F range appear to be colder than they really are.

By raising the humidity level in the air in your home during the winter months, you can alleviate many of these prob-



SOLID-STATE HUMIDITY CONTROL

BY ANTHONY J. CARISTI

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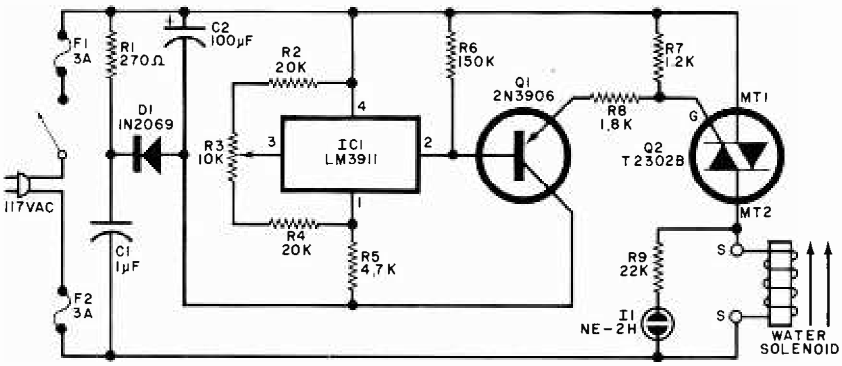

Fig. 1. Temperature sensor IC1, in furnace plenum chamber, determines when solenoid operates to spray fine water mist into chamber.

PARTS LIST

C1—1-μF, 200-V nonpolarized capacitor
C2—100-μF, 25-V tantalum capacitor
D1—1N2069 or similar rectifier
F1,F2—3-ampere slow-blow fuse (MDL-3 or equivalent)
I1—NE-2H neon lamp
IC1—LM3911N temperature-controller (National)
Q1—2N3906 or similar transistor
Q2—T2302B triac (RCA)
The following resistors are ¼-W, 10%, unless otherwise specified:
R1—270-ohm, 2-watt, 10% resistor
R2,R4—20,000 ohms metal film (see text)
R3—10,000-ohm subminiature trimmer
R5—4700 ohms
R6—150,000 ohms
R7—1200 ohms
R8—1800 ohms
R9—22,000 ohms
S1—Spst switch (optional)
Misc.—Printed-circuit board; fuse holders for F1 and F2; 120-volt ac water solenoid (W.W. Grainger No. 6X230 or similar); spray nozzle (see Note below); 4-conductor (color-coded) cable; hookup wire; etc.

Note: The following items are available from A. Caristi, 69 White Pond Rd., Waldwick, NJ 07463: printed circuit board for $3.60; IC1 for $3.00; Q2 for $3.00; spray nozzle (specify A-37 for 0 37 gal/hr or A-50 for 0.50 gal/hr) for $6.75. Add 50¢ for postage per order.

lems. In terms of saving energy (fuel), you can reduce your thermostat setting while maintaining a relative humidity of 30% to 50% and feel warmer at 66° to 70° F than you would at 75° in arid air.

Circuit Operation. The heart of the humidity controller is a special integrated circuit, *IC1*. This four-terminal LM3911 has an entire temperature-control system on a single chip. Included in this IC are a temperature sensor, stable voltage reference, and operational amplifier. As shown in Fig. 1, the internal op amp is wired as a comparator with external resistors so that the pin-2 output switches voltage as the temperature of the IC's case traverses the set-point determined by the reference voltage on pin 3. Since the IC has a built-in 6.8-volt reference supply, the temperature switch-point of the controller is held stable regardless of power-line voltage.

The output current from *IC1* is insufficient to drive the gate of triac *Q2*. Therefore, *Q1* is used to amplify this current to a sufficient level.

When the temperature of *IC1* drops below the set-point, the voltage at pin 3 cuts off *Q1*. Conversely, as the temperature rises above the set-point, *IC1*'s output switches negative and sends *Q1* into conduction. This, in turn, applies sufficient current to the gate of *Q2* to cause the triac to trigger on and energize the water solenoid. Neon lamp *I1*, connected across the solenoid (with dropping resistor *R9*), provides visual indication that the solenoid has been energized and the water is on.

By physically locating *IC1* in the warm-air stream of the furnace, we can ensure that the water solenoid will be energized only when the temperature of the heated air is sufficient to vaporize the mist from the spray head. When the burner shuts off and the air cools, the spray is automatically terminated. Potentiometer *R3* in the input circuit of *IC1* allows you to select warmer or cooler switching temperatures, permitting you to adjust duty cycle of the spray and, thus, the degree of humidification.

Dc power for *IC1* and *Q1* is provided by *R1*, *C1*, *D1*, and *C2*, with *R1* and *C1* forming a voltage-divider network. Since *C1* is purely reactive, it dissipates no power and remains cool. This makes it possible to develop a relatively low voltage across *R1* without need for a power transformer or high-wattage resistor. Current through *D1* charges *C2* to about 20 volts. Fuses *F1* and *F2* protect the circuit against excessive damage in the event of component failure. As either side of the circuit can be connected to the "hot" leg of the power line, both sides are fused.

Construction. The circuit is best assembled on a printed-circuit board, etching-and-drilling and components-placement guides for which are shown in Fig. 2. When laying out the board for etching, be sure to maintain the heavy conductor traces where indicated.

Wire the board as shown, taking care to properly orient *D1*, *C2*, *Q1*, and *Q2*. Be sure to use a nonpolarized capacitor for *C1*.

Metal-film resistors are specified for *R2* and *R4* in the Parts List to maintain greatest temperature set-point stability. If you substitute 10% composition resistors, you can use 18,000 ohms instead of the specified 20,000 ohms.

Note that *IC1* mounts off the board, inside the heated-air chamber (plenum) of your furnace, where it will be able to sense burner operation. Connection between *IC1* and the board is accomplished with a four-conductor, preferably color-coded, cable. Pay strict attention to pin connections when wiring *IC1* to the board via its cable, referring back to Fig. 1 as necessary. (The length of the interconnecting cable is not critical.)

When mounting *IC1* in the plenum, make absolutely certain its terminals are insulated from all metal parts in the furnace, since the IC is electrically connected to a circuit that does not have an isolation transformer. Should the IC touch any metal part of the furnace, the resulting short circuit will destroy the IC. Also, do not locate *IC1* where it will come into contact with the water supply.

The water solenoid valve terminals connect to the pc board at the two points labelled with an S. Use at least 20-gauge wire for the connections. Power input to the circuit, labelled AC on the board, can be made via a conventional line cord and plug. You can install an optional power switch in series with the ac line, if desired.

Typical installation of the system is illustrated in Fig. 3. The water connection between supply and spray nozzle can be made via ¼" (6.4-mm) copper tubing and plumbing fittings as required. Mount the spray nozzle to the wall of the plenum chamber so that it sprays into the heated section of the warm-air chamber. Do *not* place it in the cold-air return. To

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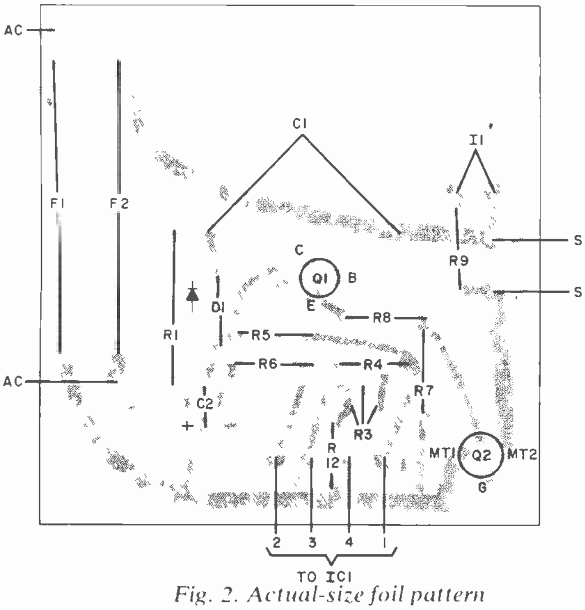

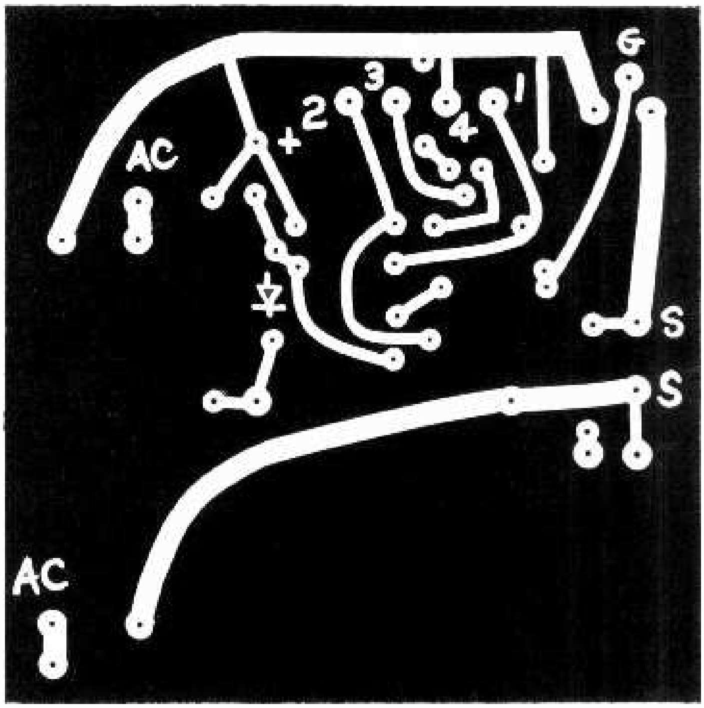

Fig. 2. Actual-size foil pattern for pc board is at right; component installation above. Note that IC1 is not on the board. When mounting board, make sure it is isolated from ground.

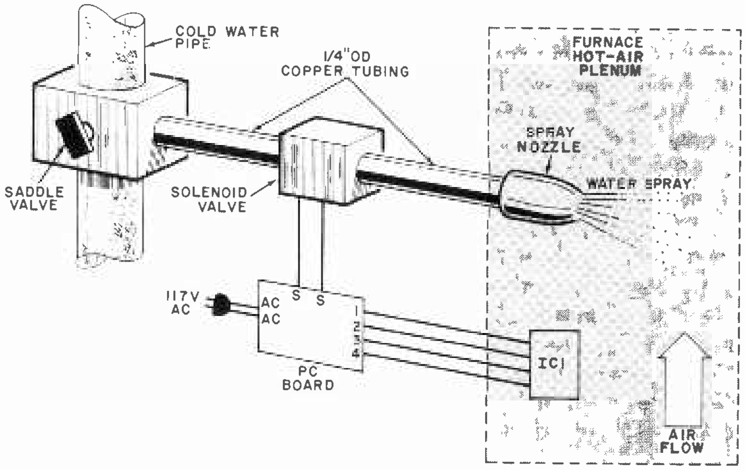

Fig. 3. Typical installation showing plumbing. An optional power switch can be connected to the ac line.

do so will cause improper operation and possible damage to your furnace.

Checkout and Use. Before installing the control system, check for correct circuit operation. You do not need the water-supply connection for this check.

First, make certain that *IC1* is fully insulated from all metal objects and your person. Then connect the solenoid terminals to points S on the board. Rotate *R3* over its range. At some point, you should hear the solenoid click and see the neon lamp light simultaneously. This setting will be close to room temperature. Adjust *R3* so that *I1* just extinguishes to obtain an approximate setting for the potentiometer. This done, disconnect the project from the ac line and install the pc board permanently.

Install the spray nozzle in the plenum. Refer to the Parts List for specifications on two sizes of spray nozzles. Use part No. A-37 for smaller or No. A-50 for larger homes.

Final adjustment of *R3* can be made by placing the furnace in operation and setting the pot to operate the solenoid just after the furnace blower comes on. This will provide maximum humidification for your heating system. If you require less humidification, readjust R3.

In Closing. If you wish to conserve energy and/or reduce your heating costs while maintaining a comfortable living environment, simply modernize your humidifier system with the automatic controller described here. ◇

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Build a Smart Switch

A solid-state wall switch that "remembers" to turn off the lights when you forget!

by Richard Fermoyle

HAVE YOU ever gone into a darkened room "for just a minute," only to return an hour later and find the lights still burning? The "Smart Switch" presented here will correct this most common occurrence.

This useful project, which costs about $17 to build, is a solid-state, 117-volt ac timer switch designed to replace a conventional wall switch. Using the components specified, the Smart Switch can control loads up to 250 watts.

When a pushbutton on the Smart Switch is depressed, power will be supplied to the load (lights) connected to it for approximately one minute. At the end of that interval, power will be automatically removed. An optional bypass switch is provided to override the timer circuit and to power the load continuously. With today's high cost of energy and the need to conserve, this device is a practical and economical addition to your home.

About the Circuit. The Smart Switch is shown schematically in Fig. 1. The heart of the circuit is *IC2*, a 555 timer operating as a monostable multivibrator. When pushbutton switch *S1* is depressed, power from the 117-volt ac line is applied to the timer circuit. Parallel resistors *R3* and *R4* drop approximately 95 volts of the line voltage, resulting in the application of approximately 22 volts ac to the input of modular bridge rectifier *RECT1*. The pulsating dc output generated by *RECT1* is converted into +5 volts regulated by filter capacitor *C7* and IC regulator *IC1*.

When power is initially applied to the timer circuit, pin 3 of *IC2* goes high and forward-biases the infrared-emitting diode within *IC3*, an optically isolated triac driver. This activates the bilateral switch within *IC3* which triggers triac *Q1* into conduction. When the triac turns on, 117 volts ac is applied to the load and to the center contact of switch *S2*. If this switch is placed in position "A", as shown in the schematic, the timer circuit continues to receive line power even though pushbutton switch *S1* is released.

The load and the timer circuit will be powered for a period of time determined by values of components *R6* and *C4*. For the component values shown, this interval is approximately one minute. Once *IC2* has timed out, pin 3 of *IC2* goes low and deactivates *IC3* and triac

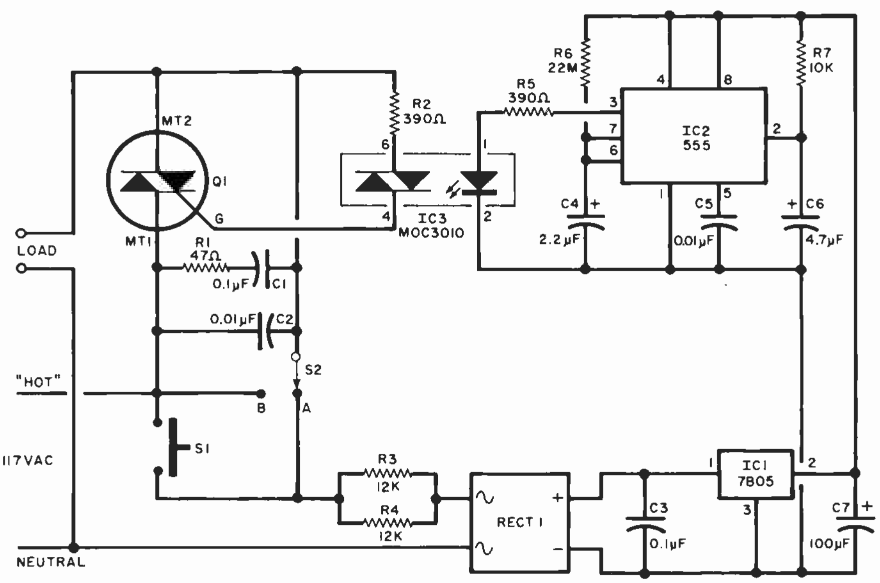

Fig. 1. When power is applied to the circuit by pressing S1, the output of IC2, through IC3, triggers Q1, which supplies power to the load (with S2 on "A") for a time determined by R6 and C4. With S2 on "B", power is supplied directly to the load.

PARTS LIST

C1—0.1-µF, 200-VDC tubular (272-1053)*
C2—0.01-µF, 200-VDC tubular (272-1051)*
C3—0.1-µF disc ceramic (272-1069)*
C4—2.2-µF tantalum (272-1407)*
C5—0.01-µF disc ceramic (272-1065)*
C6—4.7-µF tantalum (272-1409)*
C7—100-µF, 10-volt electrolytic (272-1044)*
IC1—7805 voltage regulator (276-1770)*
IC2—555 timer (276-1723)*
IC3—MOC3010 triac driver ***
Q1—6-A, 200-V Triac (276-1001)*
R1—47-ohm, ¼ -watt resistor
R2—390-ohm, ¼-watt resistor
R3, R4—12,000-ohm, 2-watt resistor
R5—390-ohm, ¼-watt resistor
R6—22-megohm, ¼-watt resistor *
RECT1—1-A, 50-PIV modular bridge rectifier (276-1161)*
S1—Single-pole, normally open pushbutton switch (34-02062V)**
S2—Spdt rocker switch (99-64248V)**
Misc.—Electrical box cover plate, printed circuit board, heat sink, silicone thermal compound, barrier strip (274-657)*, IC sockets (optional), hookup wire, spacers, mounting hardware, etc.

* Radio Shack Part Number
** Lafayette Part Number
*** Motorola Semiconductor component, available from Motorola Distributors

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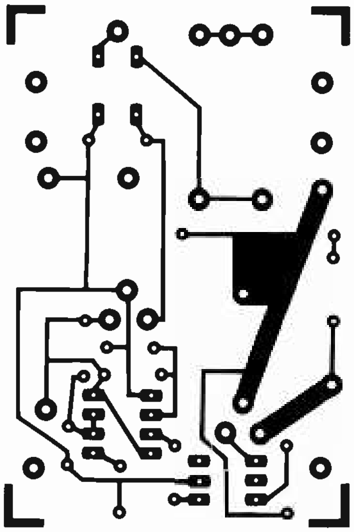

Fig. 2. Actual-size etching and drilling guide for pc board.

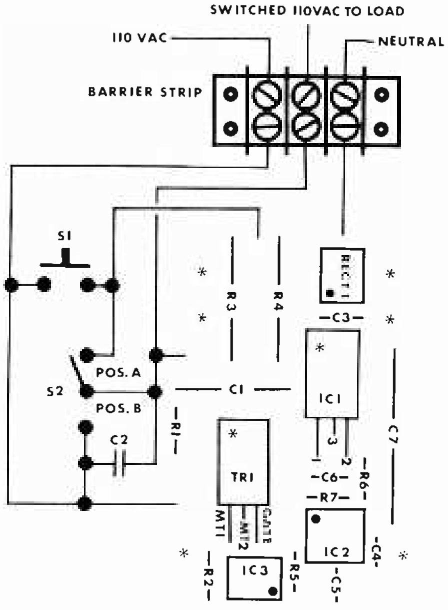

Fig. 3. Parts placement guide for the printed circuit board is shown at right.

Q1. Power is thus removed from the load and the timer circuit.

Placing switch *S2* in position "B" bypasses the triac and applies 117 volts ac directly to the load. This feature has been incorporated into the Smart Switch so that the user can manually keep the load powered for an indefinite period of time. If the bypass feature is not desired, switch *S2* and capacitor *C2* can be eliminated. In that case, however, it will be necessary to connect the *MT2* terminal of triac *Q1* directly to the junction of *S1*, *R3*, and *R4* to ensure proper operation of the timer circuit.

Construction. Most of the circuit can be mounted on a single printed circuit board. The etching and drilling and parts placement guides are shown in Figs. 2 and 3; respectively. Triac *Q1* must be mounted on a heat sink. Also, be sure to use silicone thermal compound to ensure a good heat transfer. A standard plastic electrical wall-box cover plate should be cut out and drilled to accommodate switches *S1* and *S2*. Capacitor *C2* is then mounted directly on the lugs of switch *S2*.

As shown in Figure 4, a three-terminal barrier strip is mounted on standoffs on the component side of the printed circuit board directly above *R3*, *R4*, *C3* and *RECT1*. When soldering capacitor *C3* to the pc board, leave the leads long enough so that the body of the capacitor can be bent back to lay flat on top of *RECT1*. Suitable lengths of hookup wire should be used to interconnect the board and switches *S1* and *S2*.

The completed pc board is then mounted using standoffs on the back side of the plastic cover plate. Be sure to use standoffs that are not too long, as the entire assembly must fit within a standard electrical wall box.

Installation. Before installing the Smart Switch, be sure to turn off the power at the fuse or circuit breaker box. Remove the existing wall switch and cover plate. Then, using the parts placement diagram shown in Figure 3 as a guide, connect the existing wall-switch wiring to the Smart Switch barrier strip. You might find that a neutral wire has not been brought into the wall-switch electrical box. If this is the case, wire the neutral terminal of the barrier strip directly to the metal wall box.

Carefully screw the assembled Smart Switch into place and you're ready to start using it. The finished product will look like the prototype shown in Fig. 5.

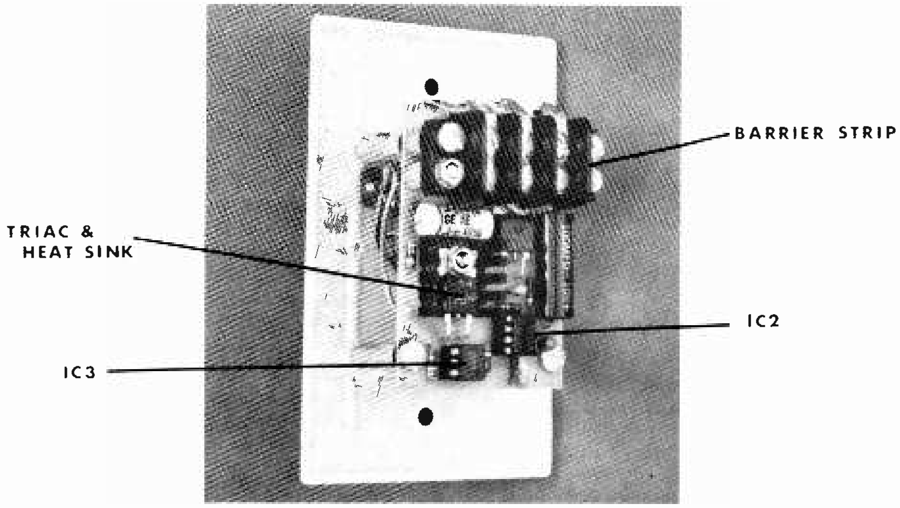

Fig. 4. Photo of the back of the Smart Switch shows how pc board is mounted on a standard plastic cover plate.

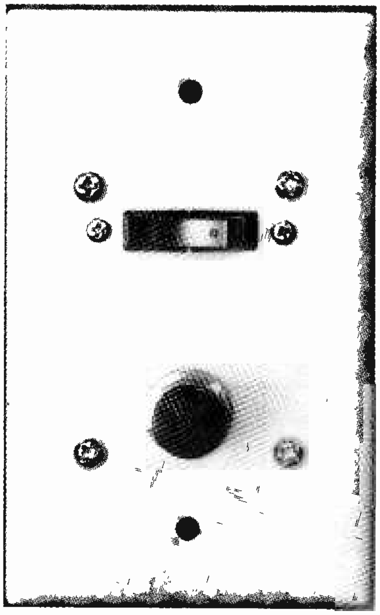

Fig. 5. Completed Smart Switch mounted and ready for use.

Use. If you are only going to remain in the room equipped with the Smart Switch for a short period of time, depress *S1* as you enter. The lamp controlled by the project will remain on for the period determined by the values of the components in the timing circuit, resistor *R6* and capacitor *C4*. If you intend to remain in the room for an extended period of time, place switch *S2* in its "B" position. ◇

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BY WILLIAM P. JOHNSON, JR.

BUILD A STEREO ROTO-BLENDER

Lets you manipulate your stereo to blend or transpose the two channels.

MOST STEREO recordings made in a professional studio begin as a number of "tracks" (usually 16 or more) on tape, which are subsequently mixed down to two channels. During mixing, the apparent location of each instrument and vocalist is fixed in the final left and right channels by its relative loudness. Usually, the listener cannot alter the mix other than by transposing or by blending the two channels to reduce stereo separation. With the "Stereo Roto-Blender," however, he can remix the recording, within certain limitations, to improve the mix and emphasize previously "buried" sounds. It also allows him to blend and transpose the two resulting channels in the conventional manner. The new mix will have roughly the same channel separation as the original program.

The Basic System. The Roto-Blender is made up of two basic parts: a stereo ROTATE control, which is the heart of its remixing capabilities, and a stereo BLEND control (Fig. 1). The ROTATE control "rotates" the performers in a circle around the listener. With the control centered, the mix is unaltered. As it is rotated clockwise, the sounds originating from the left and center shift to the right. The sound originating from the right moves over to the left to complete the rotation.

The above effect is illustrated in Fig. 2. Note that, with the ROTATE control centered (NORMAL), a vocalist is centered between a guitar on the left and a piano on the right. By rotating the control to the left, the vocalist and piano shift one position to the left and the guitar comes over to the right. Exactly the opposite rotation occurs when the control is rotated in the clockwise direction. The control alters both the sonic directions and relative loudnesses of each sound. Normally, when a sound is shifted to the center, it becomes louder, and when it is shifted away from center, it becomes quieter. This allows the listener to emphasize interesting or previously unnoticed sounds.

The BLEND control allows you to reduce channel separation down to monaural as it is turned from fully clockwise to center. Rotating the control counterclockwise causes the separation to increase, this time with the left and right channels transposed. This transposition provides additional flexibility in the remixing process.

About the Circuit. The left- and right-channel inputs to the Roto-Blender in Fig. 1 are buffered by *IC1A* and *IC1B* and passed to differential amplifier *IC1C* whose output is an R − L signal. This signal is similar to the combined left- and right-channel signals minus the center-channel material. NULL ADJ control *R13* permits the center-channel material to be precisely cancelled to achieve optimum results.

The R − L signal is inverted by *IC1D* to produce an L − R signal. The left- and right-channel signals plus the composite signals are applied to ROTATE potentiometer *R14*. Figure 3 illustrates the signals applied to *R14* and indicates how the resulting output signals on each control wiper vary over the range of the potentiometers. An important feature of this arrangement is the cancellation of one channel when the control is at its center of rotation, leaving only the remaining channel, attenuated by one half. In this manner, normal stereo is obtained at center of rotation. The attenuation is counteracted by *IC2A* and *IC2B*, whose boosted outputs are added to the *R14* outputs through *R11* and *R12*. This does not affect the signal at the extreme positions of the ROTATE control, due to the potentiometer's zero source impedance, but increases in effect as the pot is adjusted to its center position. This results in a nearly constant loudness at all positions of the potentiometer for most stereo signals.

After rotation occurs, the signals are applied to buffer amplifiers *IC2C* and *IC2D*. BLEND control *R15* mixes the signals in varying proportions to achieve

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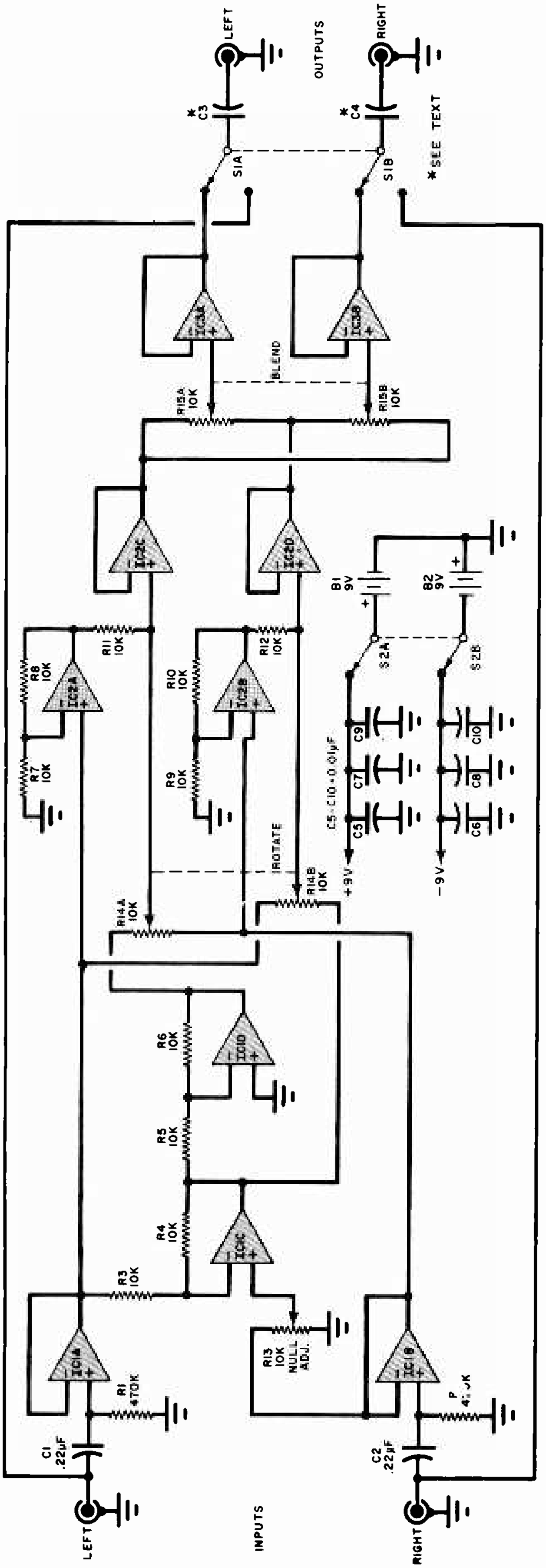

Fig. 1. The left and right stereo signals are buffered in IC1A and IC1B and combined in IC1C. Potentiometer R13 adjusts the null.

PARTS LIST

B1, B2—9-volt battery
C1, C2—0.22-µF disc
C3, C4—1-µF nonpolarized (see text)
C5 through C10—0.01-µF disc (see text)
IC1, IC2, IC3—4136 quad op amp (or any other op amp IC—see text)
J1 through J4—Phono jack
R1, R2—470,000-ohm, ¼-watt, 10% resistor
R3 through R12—10,000-ohm, ¼-watt, 10% resistor
R13—10,000-ohm, linear-taper potentiometer
R14, R15—10,000-ohm, linear-taper dual potentiometer
S1—Dpdt switch
S2—Dpst switch
Misc.—Holders for B1 and B2; printed circuit or perforated board; sockets for IC's; control knobs (3); suitable enclosure; lettering kit; machine hardware; hookup wire; solder; etc.

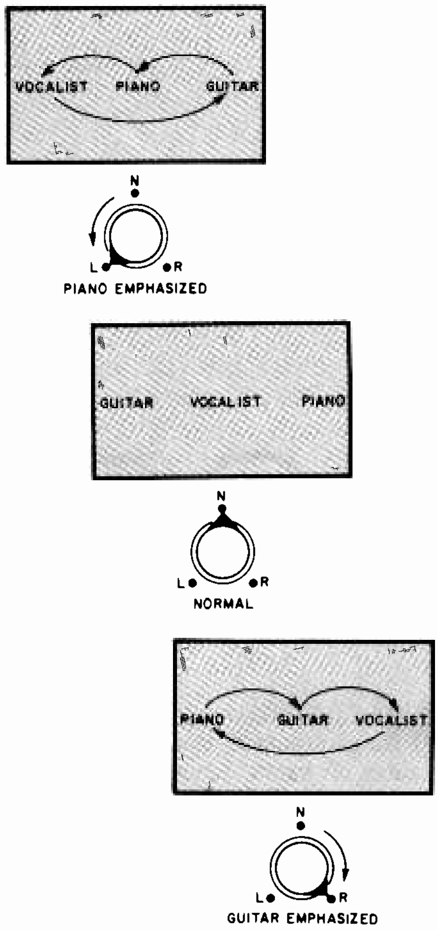

Fig. 2. With R14 centered, as in middle diagram, vocalist is between piano and guitar. With R14 rotated to either extreme, relative positions are changed.

either normal or reversed stereo, mono, or anything in between. These signals are then buffered by *IC3A* and *IC3B*, after which they are delivered to the Roto-Blender's outputs. Capacitors *C3* and *C4* are optional and are required only if the input to the amplifier to which the Roto-Blender is connected does not have similar capacitors. Their values should be chosen to have a low impedance at 20 Hz, compared to the impedance of the amplifier.

The Roto-Blender can be either battery powered as shown in Fig. 1 or driven by a ±6-to-±15-volt ac operated supply, which should be decoupled using capacitors *C5* through *C10* located close to the +V and −V pins of each op amp used. (The op amps used in the author's prototype were 4136 quad types, which required only three IC packages. If you use a different op-amp type, and almost any other type will work in this circuit, you will have to increase the number of 0.01-µF capacitors so that two capacitors are used for each IC package.)

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and boost the bass region using the Parametric. Using a high Q setting, vary the center frequency of the low-band equalizer until you discover the room's fundamental resonant frequency. (That's the one at which the walls start shaking and the furniture moves around the floor.) Now reduce the setting of the BOOST/CUT control for more even-sounding bass. The high-band equalizer can be used to brighten up a room that is too "dead" acoustically or to attenuate treble response in a room that is too "alive."

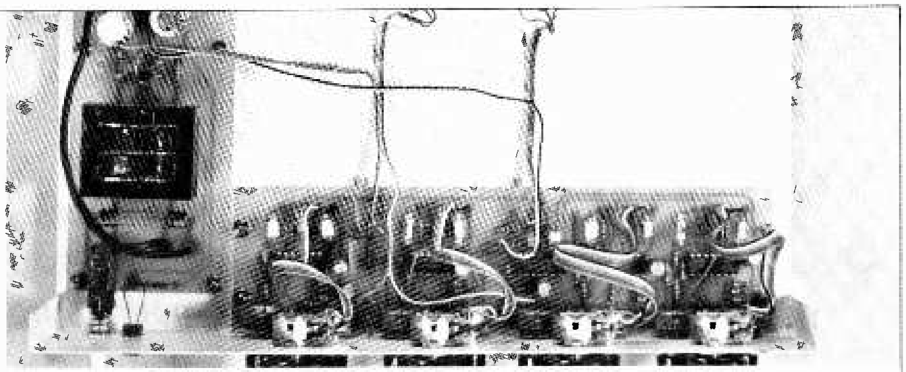

Fig. 13. Interior view of prototype using ac power supply.

You will undoubtedly find other uses for this versatile project. Those who listen to music analytically will appreciate the ability to zero in on one particular instrumental (or human) voice. Amateur recording engineers can employ the Parametric to tailor the sounds of a mix. And, of course, anyone whose speakers have response irregularities will be able to smooth them out.

One word of caution—don't blindly apply large amounts of deep bass and extreme treble boost in an attempt to flatten the response of your system at the upper and lower limits of the audible spectrum. Experience has shown that room/system combinations are best equalized by first employing acoustic methods, followed by electronic equalization. For example, you should first try repositioning the loudspeakers, modifying the absorption coefficients of the room, and adjusting the speakers' crossover level controls (if any).

Most often, a lack of deep bass and extreme highs is due to the limitations of dynamic drivers. Don't try to force flat response out of your speakers by cranking up the BOOST/CUT controls. The results of such attempts frequently include overloaded amplifiers, excessive distortion, and blown voice coils. Remember—equalization should be introduced *intelligently*.

In Conclusion. We have presented a stereo Parametric Equalizer project that is well suited for home, mobile, and portable applications. It provides a high level of performance and the flexibility of control inherent in the parametric design, enough flexibility for most readers. Those who require more bands of equalization per channel can reproduce two or more complete equalizers and connect them in cascade for even greater control over the sounds they record or reproduce. ◇

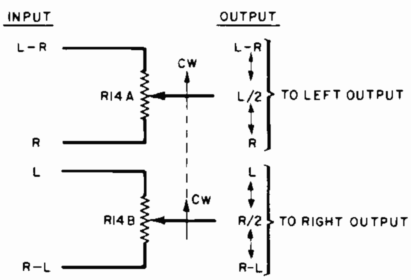

Fig. 3. The left and right signals and the composite are applied to R14 as shown. Note how the resulting output signals on each control wiper vary over potentiometer range.

Construction. The circuit can be assembled on a printed circuit board of your own design or on a perforated board using pencil wiring techniques. In either case, it is a good idea to use sockets for the IC's. Mount the potentiometer controls, input and output jacks, and POWER and IN/OUT switches on the box in which the circuit is housed. Use a dry-transfer lettering kit to label the controls, jacks, and switches according to function and operation.

Application. The Roto-Blender unit should be connected to suitable high-level inputs and outputs for optimum results. You can connect it between a preamplifier and power amplifier or, lacking this facility, into the tape-monitor loop. It is a good idea to hook it up ahead of the headphone amplifier, since the Roto-Blender is best appreciated using headphones.

For proper operation, the Roto-Blender should be nulled to counteract imbalances in the source material and preceding electronics. This can be done by disconnecting the right channel output of the Roto-Blender and, with the ROTATE and BLEND controls fully clockwise, adjusting the NULL ADJ control to exactly cancel the center sounds of the program source. If a mono source is used, adjust for minimum sound. Excessive distortion heard at this time indicates either a worn record or stylus or some other deficiency in the source material or amplifier's electronics.

Cancellation of center sounds with some recordings is not possible when the sounds are reproduced differently in each channel, using reverberation techniques. This case should not be confused with the case where distortion prevents nulling with a raspy sound.

Once nulling is accomplished, the right channel can be reconnected and the ROTATE pot should be centered for normal stereo reproduction. If an instrument on the left—a trombone, for example—is to be emphasized, rotate the sound to the right by turning the ROTATE control clockwise. This moves the trombone to the center, where it will be more dominant. At this point, if the BLEND control is rotated fully counterclockwise, the trombone will remain centered while the left and right channels will be effectively transposed.

The effects achieved by the Roto-Blender are a function of the source material and cannot be fully described here. Perhaps the most fascinating aspect of the Roto-Blender is its ability to bring forth sounds that were never noticed before. ◇

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BY BRIAN DANCE

Listen to a NEW WORLD OF SOUNDS WITH ULTRASONIC DETECTOR

Inexpensive detector converts ultrasonic sounds from insects, compressed gas leaks, etc., to an audio output.

EXPLORING the world of ultrasonic sound—which lies above approximately 20 kHz—can be exciting and educational. Here is a frequency spectrum beyond human hearing where many insects and rodents communicate with each other, where sounds from leaks in pressurized gas lines occur, etc.

The inexpensive circuits presented here convert these ultrasonic sounds to audio frequencies, enabling anyone to hear them. Also included is a simple ultrasonic transmitter circuit that will enhance your ability to probe this interesting electronics area.

An Ultrasonic Receiver. The schematic diagram of a heterodyne-type ultrasonic receiver is shown in Fig. 1. This receiver hyterodynes ultrasonic signals with those from an internal oscillator converting them to audible frequencies for reproduction by a dynamic speaker. Thus, it allows you to "hear" any signals it detects.

Piezoelectric transducer *TR1* converts ultrasonic waves impinging upon it into ac waveforms which are applied to the noninverting input of operational amplifier *IC1A*. Because a single-ended power supply is used, resistors *R1* and *R2* bias the noninverting input to one-half the supply voltage. Resistor *R3*, effectively connected across *TR1* by electrolytic capacitor *C1*, damps the transducer's resonant response and broadens its bandwidth. At dc, *R5* provides 100% negative feedback to stabilize the operating point. At signal frequencies of interest, the gain of *IC1A* is 60 dB for the values given in Fig. 1.

The output of *IC1A* is directly coupled to op amp *IC1B*, a similar amplifier stage. The voltage gain of *IC1B*, about 43.5 dB with the component values specified, is somewhat lower than that of the preceding stage. Signals at the output of *IC1B* are capacitively coupled by *C5* to diodes *D1* and *D2*.

Also applied to the diodes is the output of an ultrasonic oscillator comprising *IC3* and its related components. The frequency of this oscillator is determined by the setting of potentiometer *R12* and the capacitance of *C9*, which is chosen so that the oscillator output corresponds to the resonant frequency of the transducer. (Transducers are readily available from surplus dealers with resonant frequencies ranging from 22 to 44 kHz.)

The two diodes form a nonlinear net-

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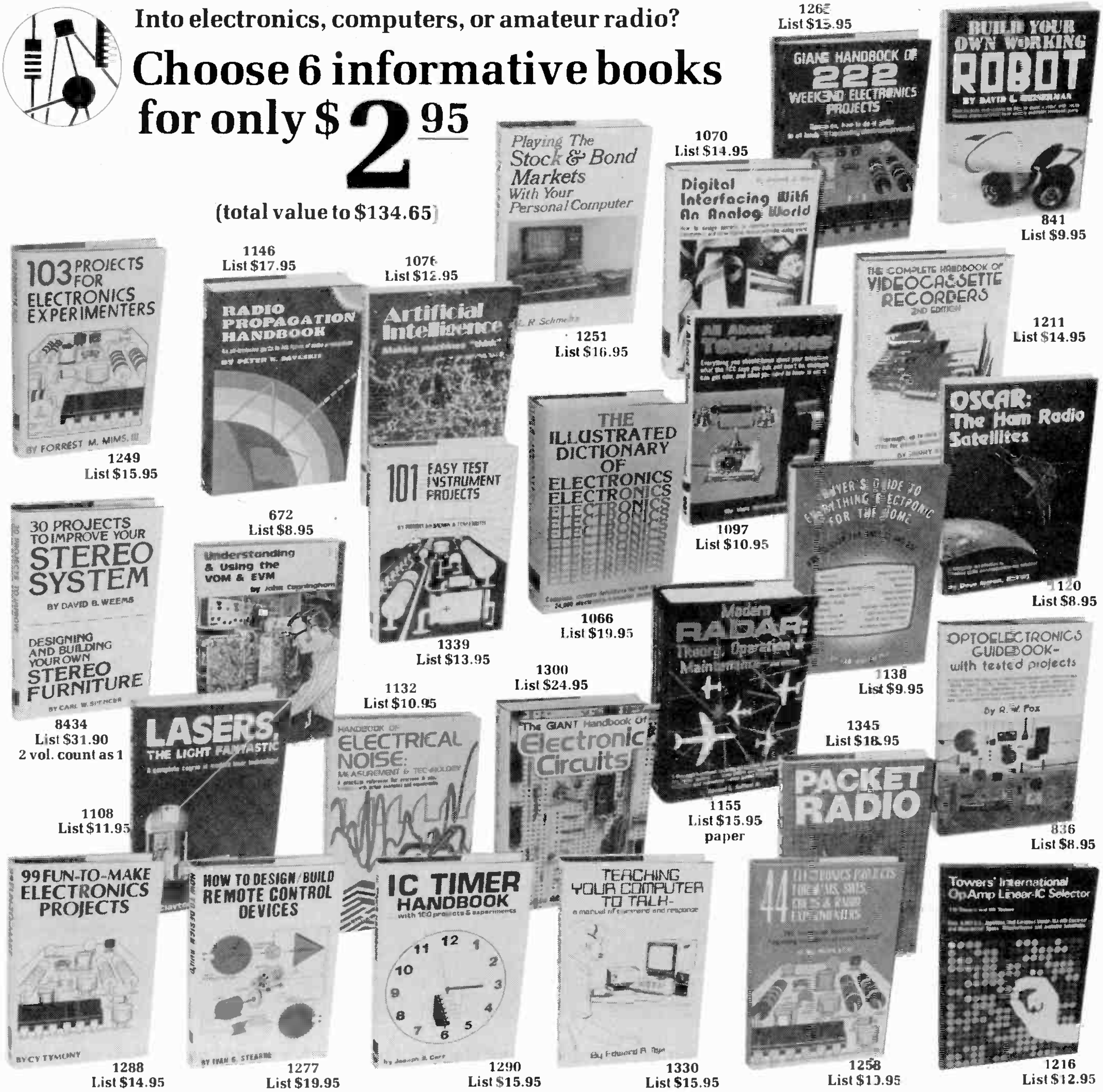

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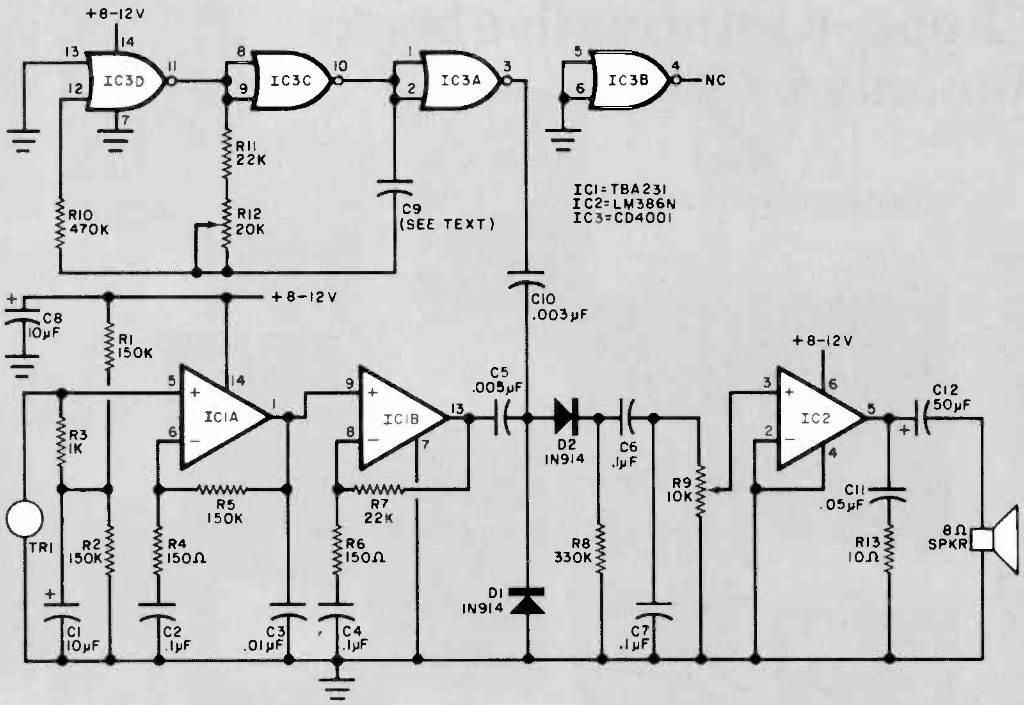

Fig. 1. An ultrasonic receiver, where incoming signals heterodyne with those from the local oscillator to produce an audible output.

PARTS LIST FOR FIG. 1

C1,C8—10-μF, 25-V tantalum
C2,C4,C6,C7—0.1-μF disc ceramic
C3—0.01-μF disc ceramic
C5—0.005-μF disc ceramic
C9—180-pF (or 330-pF) disc ceramic, polystyrene, glass or silver-mica (see text)
C10—0.003-μF disc ceramic
C11—0.05-μF disc ceramic
C12—50-μF, 25-V electrolytic
D1,D2—1N914 signal diode
IC1—TBA231 dual op amp (see note)
IC2—LM386 audio amplifier
IC3—CD4001 quad 2-input NOR gate
The following fixed resistors are ¼-W, 10% carbon composition:
R1,R2,R5—150,000 ohms
R3—1000 ohms
R4,R6—150 ohms
R7,R11—22,000 ohms
R8—330,000 ohms
R10—470,000 ohms
R13—10 ohms
R9—10,000-ohm linear-taper potentiometer
R12—20,000-ohm linear-taper potentiometer
SPKR—8-ohm dynamic speaker
TR1—Piezoelectric ultrasonic transducer
Misc.—Printed circuit or perforated board; suitable enclosure; Hook-up wire; dc power source; machine hardware; etc.
Note—The TBA231 dual op amp is imported from the U.K. by SG-ATES Semiconductor Corp., 435 Newtonville, MA 02160 (Tel: 617-969-1610).

work. Hence, when signals from the oscillator and the op amp are applied, they heterodyne with each other. If *IC3* oscillates at a frequency fairly close to that of an ultrasonic wave detected by *TR1*, an audible beat signal will appear at the cathode of *D2* at a frequency equal to the difference between the two ultrasonic frequencies. The process is similar to that performed in a conventional superheterodyne r-f receiver. The beat note, which can be tuned by adjusting *R12*, is amplified by *IC2*, an audio IC, to a level sufficient to drive the dynamic speaker. Potentiometer *R9* serves as an audio gain control.

An Ultrasonic Transmitter will help you explore the ultrasonic region more fully. A suitable design is shown schematically in Fig. 2. The circuit is similar to the local oscillator stage in the receiver, but the previously unused fourth gate in the 4001 is employed to provide push-pull drive for transducer *TR2*. The output frequency is variable by means of *R3*. The capacitance of *C1* should be chosen so that the nominal oscillating frequency corresponds to the resonance of the transducer. As was the case with *C9* in the receiver, *C1* should be 180 pF if 44-kHz transducers are used, or it should be 330 pF for use with 22-kHz transducers.

Construction. Either printed circuit or perforated board can be used to duplicate the transmitter and receiver circuits. Parts placement is not especially critical. The use of sockets or Molex Soldercons is recommended when mounting the IC's on the boards. Be sure to observe normal precautions when handling the CMOS devices. Install polarized capacitors and semiconductors with due regard for polarity and pin basing. Batteries are well suited to power the transmitter and receiver circuits. Note that, when transmitter switch *S1* is in the OFF position, the output states of *IC1*'s gates are frozen. The quiescent current drain of the circuit is so small that no power switch is necessary. If a battery supply is used with the receiver, however, an spst power switch should be used to disconnect the circuit from the supply when it is not being operated.

Use. Receiver potentiometer *R12* tunes the circuit across a limited portion of the ultrasonic frequency range. Apply power and adjust audio gain control *R9* until some noise is heard through the speak-

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er. Then rub the palms of your hands in front of *TR1*. The receiver will detect the ultrasonic energy from the rubbing.

You will notice that *TR1* has a very directional response. This is due to the fact that ultrasonics have very short wavelengths (compared to those at audio frequencies) and are thus subject to less diffraction at the edges of large objects. Also, ultrasonic waves behave like light waves in that they tend to travel in straight lines.

It's interesting to note that if coupling capacitor *C10* in the receiver is disconnected from the diode mixer, the receiver will still detect ultrasonic signals if more than one frequency is present. The frequencies present at the input will beat against each other to produce an audible output. This can be verified by repeating the palm-rubbing experiment described earlier after the coupling capacitor has been disconnected. The speaker will still generate an audio output even though no local oscillator signal is being injected into the diode mixer.

If an ultrasonic wave generated by transmitter transducer *TR2* now impinges upon *TR1*, the random noise reproduced by the speaker will drop to a low level. No tone will be heard because only one frequency is applied to the mixer. Stray coupling that allows a portion of the local oscillator output to reach the mixer will create an audible beat.

When the receiver and transmitter are operating in the same room, a signal will be heard as *R12* tunes the receiver across its range. The two transducers do not *have* to be directly facing each other if enough hard surfaces in the room reflect the ultrasonic waves, and the room is not so large that it introduces excessive signal attenuation.

The circuits presented have been successfully used with ultrasonic transducers from many different sources, including those used in television receiver remote control accessories. Of course, if you want to tune in several ultrasonic "bands," you can use a multiple-pole rotary switch to select the appropriate transducer and its corresponding oscillator capacitance. Experimentation indicates that the receiver can "hear" the transmitter at distances up to 125 feet if the transducers are aimed at each other. The use of a suitable parabolic reflector in tandem with *TR1* and/or multiple driven transmitter transducers should result in even greater useful range.

Other Suggestions. We have already mentioned the possibility of using these circuits for signalling purposes. Many other practical applications exist. For example, leaks in the rubber sealing of car doors and windows or in the sealing of a freezer door. The transmitter is placed in the car or freezer and fills the interior with ultrasonic waves. The walls of the interior reflect the waves to create a wide dispersion of ultrasonic energy. If the receiver's transducer is moved over the exterior, a tone will be heard whenever it passes any leaks. ◇

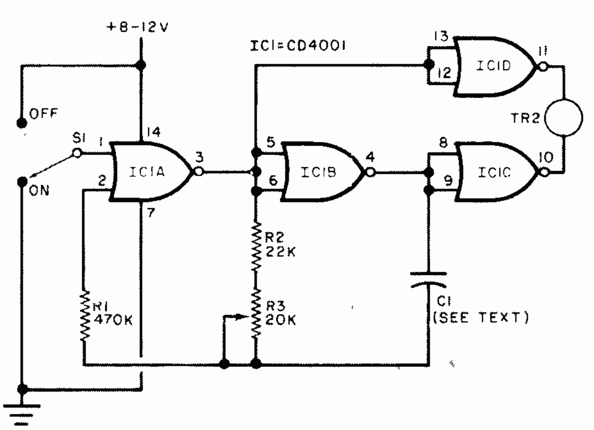

Fig. 2. This ultrasonic transmitter employs four NOR gates.

PARTS LIST FOR FIG. 2

C1—180-pF (or 330-pF) disc ceramic, polystyrene, glass or silver mica capacitor
IC1—CD4001 quad dual-input NOR gate
R1—470,000-ohm 10%, ¼-W resistor
R2—22,000-ohm 10%, ¼-W resistor
R3—20,000-ohm linear-taper potentiometer
S1—Spdt switch
TR2—Piezoelectric ultrasonic transducer
Misc.—Printed circuit or perforated board; suitable enclosure; hook-up wire; dc power source; machine hardware, etc.

SURFER

(Continued from page 116)

good choice is a 4" to 6" (10.2 to 15.2 cm) high-compliance, high-fidelity quality 8- or 16-ohm speaker.

With *Q1* operated in the manner shown, it is susceptible to detecting r-f. Hence, the circuit should be housed in a metal box and the common bus on the pc board electrically connected to the box. The pc board assembly mounts in place with spacers and machine hardware. POWER switch *S2*, VOLUME control *R12*, INTERVAL control *R20*, and SELECTOR switch *S1* mount on the front panel. Transformer *T1* and the holder for fuse *F1* can be mounted on the rear of the box. The line cord should enter the box through a grommet-lined hole.

Adjustment and Operation. Two simple adjustments are required to get the Surfer into proper operating order. Set trimmer pots *R9* and *R16* to their centers of rotation, set the VOLUME control fully clockwise (maximum volume), and set *S1* to the SURF position. Now, apply power and set the INTERVAL control for a period of about 25 seconds. Adjust *R16* so that *I1* extinguishes about a second before the end of the cycle. Wait several cycles and then check *R16* again and readjust it if necessary. Once *R16* is adjusted, place the shield over the lamp/photocell assembly.

Set the VOLUME control to its center of rotation. Then adjust trimmer pot *R9* until the sound is just barely audible at the beginning of the cycle.

During operation, when *S1* is set to SURF, *C5* (see Fig. 1) produces the maximum high-frequency rolloff and *PC1* is connected to the base of *Q3*. This produces a roaring sound that changes in intensity and tone. In the RAIN position of *S1*, maximum high-frequency attenuation occurs with no amplitude modulation. This creates a constant-volume "hiss" whose tone varies. When *S1* is in the NOISE position, *R11* is shorted out, which causes the tone control to lose its effectiveness. In general, a long interval is best for the SURF function, while a short interval is best for RAIN.

The Surfer is not intended to be used as a sound-effects generator to which one consciously listens. Rather, it is meant to provide a nondistracting background of sounds. Best effects are created when the Surfer is positioned 6' (1.8 m) or more from the listener with the VOLUME adjusted so that the sound is barely audible. ◇

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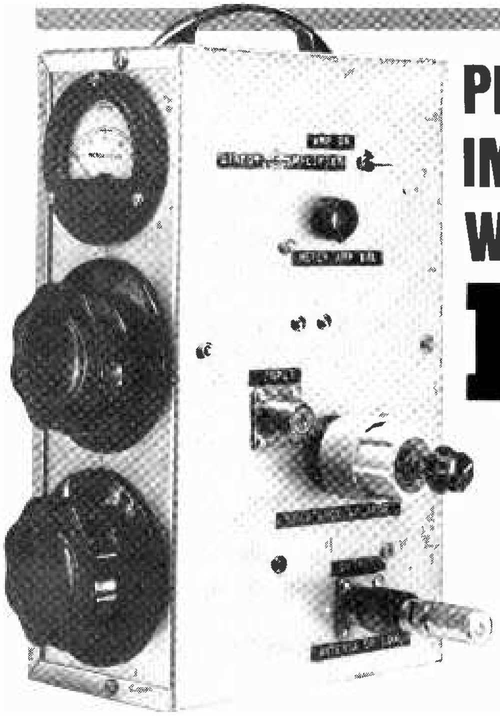

PERFORM COMPLETE IMPEDANCE MEASUREMENTS WITH THIS R-F BRIDGE

BY DON MORAR, W3QVZ

Inexpensive bridge measures *R* and *X* components over a wide frequency range

ONE OF the most useful instruments an experimenter who works with r-f circuits can have is an impedance bridge. The ideal bridge would permit accurate measurement of both the resistive *and* reactive components of an unknown impedance over a wide range of frequencies. Commercial r-f impedance bridges, although they satisfy these requirements, are priced well beyond the means of the average experimenter. On the other hand, those affordable bridges that have appeared as construction projects in amateur radio and hobby electronics magazines only tell half the story—the resistive component.

The bridge described in this article can measure the complex impedance of just about any load at frequencies between 3.5 and 54 MHz with a high degree of accuracy. Moreover, it can be inexpensively built using "junk box" components, and is smaller and lighter than its typical commercial counterpart. The only external items required for calibration and operation are a group of nonreactive resistors, an r-f source such as a signal generator, and, of course, the impedance to be measured.

Among the project's features are a built-in null indicator (a microammeter) and an amplifier which can be switched into the null detector circuit to enhance its sensitivity. The value of the impedance's resistive component is read directly off the bridge's R dial, which is calibrated in ohms. The unit's X (reactance) dial calibration is scaled in terms of frequency. This is done because inductive and capacitive reactance vary with frequency, so an X dial calibrated directly in ohms would be accurate at only one specific frequency. Scaling the X dial's calibration in terms of frequency provides greater operating flexibility.

About the Circuit. The schematic diagram of a basic r-f bridge is shown in Fig. 1. It obviously resembles the classic Wheatstone bridge, which has four resistive arms. Two of these arms are usually derived from a potentiometer. The r-f bridge, however, employs a dual differential variable capacitor (*C1C2*) so that measurements can be performed over a wide range of frequencies. If a potentiometer were used, its frequency-dependent intrinsic reactance would cause the bridge to yield false results.

Some readers might not be familiar with the dual differential capacitor. It is, essentially, two variable capacitors ganged so that when one section (capacitor) exhibits maximum capacitance, the other exhibits minimum capacitance. The use of a dual differential capacitor provides a variable capacitance ratio between the two arms of the bridge that it comprises. This simplifies the calibration and use of the bridge. If the value of *R1* is a constant, the single control knob used to vary the setting of *C1C2* can be calibrated in terms of *R1* or in ohms.

To use the bridge the unknown impedance is connected as R_U and an r-f source is used to energize the network. The dual differential capacitor is then adjusted so that the bridge is balanced. When that happens, no voltage drop will exist across the bridge detector and no current will flow through it. The detector will indicate a null and the value of R_U can be read off the dual differential capacitor's control knob.

The bridge in Fig. 1 will only measure the unknown's resistive component. More complete information about the unknown (including its reactance) can be obtained by measuring it on the bridge shown in Fig. 2. This circuit re-

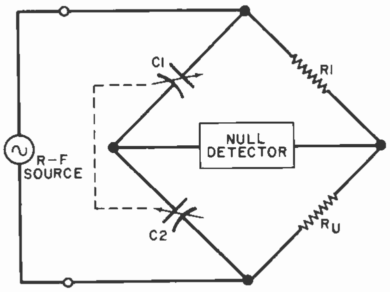

Fig. 1. Basic r-f bridge uses dual differential capacitor but gives only resistive information.

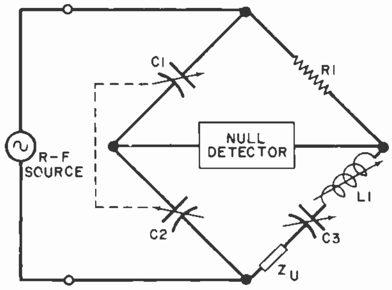

Fig. 2. More sophisticated bridge can measure reactance as well as resistances.

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sembles that of Fig. 1, but two components ($C3$ and $L1$) have been added to the bridge's lower right arms. To underscore the bridge's greater measuring facilities, the unknown is no longer represented as a resistance (R_U), but as a general impedance (Z_U).

As in Fig. 1, dual differential capacitor *C1C2* is used to measure the real (resistive) component of the unknown complex impedance. Variable inductor *L1* and variable capacitor *C3* make possible measurement of both the sign and magnitude of the unknown's imaginary (reactive) component. The bridge thereby provides the user with complete information about the unknown impedance.

The device is initially balanced at the frequency of interest with a purely resistive termination at Z_U. Variable capacitor *C3* is placed at its midrange setting and inductor *L1* is adjusted for resonance. This cancels out any reactance which would otherwise be reflected into the other bridge arms. The nonreactive termination is then replaced with the unknown impedance. Its resistive component is balanced by varying *C1C2* and its magnitude read off the calibrated capacitor control knob scale. The unknown's imaginary component is balanced by shifting *C3* away from its midscale setting in either the clockwise (+, the standard sign for inductive reactance) or counterclockwise (−, the standard sign for capacitive reactance) direction to cancel out any reactance in the unknown.

If the unknown impedance has an inductive component, more capacitive reactance (that is, less capacitance) is required from *C3* to obtain a balance. Conversely, if the load has a capacitive component, more capacitance and less capacitive reactance is required. Once *C3* has been properly adjusted, the bottom right leg of the bridge will look purely resistive, and an excellent null will be obtained on the detector.

The scale of *C3*'s control knob should be calibrated in terms of the magnitude and sign of the reactance present at Z_U. Because inductive reactance varies directly with frequency and capacitive reactance varies inversely with frequency, the calibration of *C3*'s control knob must be scaled in terms of frequency. If it were calibrated directly in ohms, its calibration would hold true at one frequency only. A better approach is to perform the calibration at 1 MHz and frequency-scale it. The exact magnitude of the reactive component can then be determined by a simple arithmetic operation.

The complete schematic of the r-f impedance bridge is shown in Fig. 3. Resistive balancing is performed by dual differential capacitor *C1*. Noninductive resistor *R2* provides the reference against which the resistive component of the unknown impedance is measured. (The unknown is connected to LOAD jack *J3*.) Balancing and measurement of the unknown's reactive component (if any) is the task of *L2* and *C2*.

When the bridge is unbalanced, germanium diode *D1* rectifies r-f into pulsating dc which is filtered by *L1* and *C3*. If *S1* is placed in its DIRECT position, the filtered dc is applied to null indicator *M1*, a 0-to-200-μA meter. For increased reso-

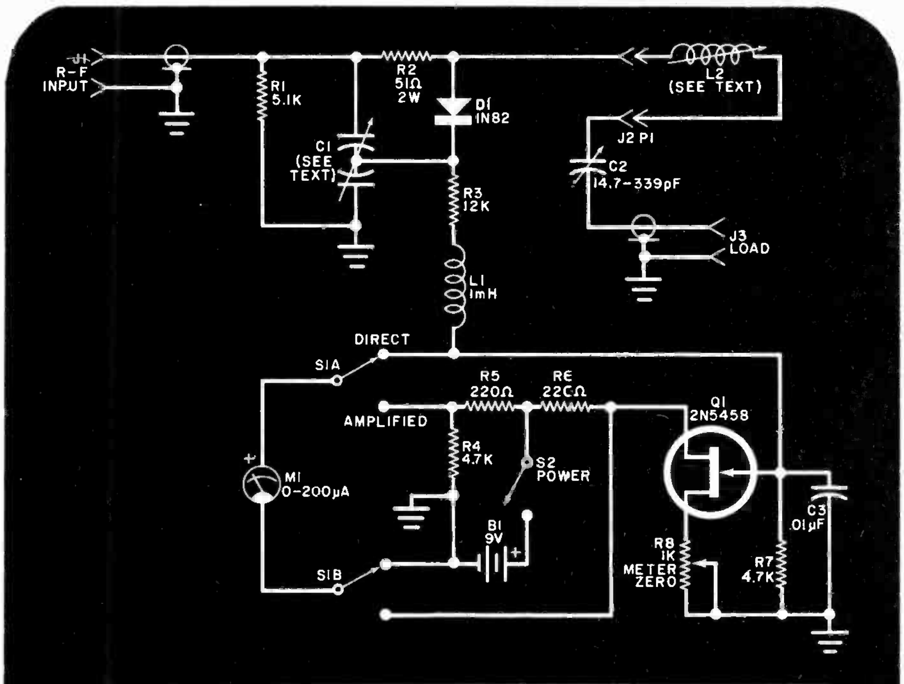

Fig. 3. Complete schematic diagram of the r-f impedance bridge. Amplifier Q1 enhances sensitivity and increases null resolution.

PARTS LIST

B1—9-volt transistor battery
C1—Dual differential variable capacitor, 12-to-150 pF per section, Millen No. 28801 or equivalent (see text)
C2—14.7-to-339-pF variable capacitor (Millen No. 19335 or equivalent)
C3—0.01-μF disc ceramic capacitor
D1—1N82 or equivalent germanium diode
J1,J3—SO-239 coaxial connector
J2—Standard Amphenol 4-prong jack
L1—1-millihenry inductor
L2—See text and Table
M1—0-to-200-μA meter
P1—4-prong plug to match *J2*
Q1—2N5458 n-channel JFET
The following, unless otherwise specified, are ¼-watt, 5% tolerance, fixed carbon-composition resistors.
R1—5100 ohms
R2—51 ohms, 2 watts
R3—12,000 ohms
R4,R7—4700 ohms
R5,R6—220 ohms
R8—1000-ohm potentiometer
S1—Dpdt toggle switch
Misc.—10" × 6" × 3½" (25.4 × 15.2 × 8.9 cm) aluminum utility box (Bud CU3010A or equivalent), 5¼" × 3" × 2⅛" (13.4 × 7.6 × 5.4 cm) aluminum utility box (Bud CU3006A or equivalent), J.W. Miller No. 42000CBI or equivalent slug-tuned coil forms, plastic or metal threaded BX/Romex outdoor electrical-box plugs, 1¼-inch (3.8-cm) PVC pipe, PVC pipe adapters (½-inch or 1.3-cm threads to ½-inch or 1.3-cm pipe), cyanoacrylate cement, control knobs with ⅛-inch (3.3-mm) shaft hole, one small control knob with ¼-inch (6.6-mm) shaft hole, two large Bakelite control knobs with ¼-inch (6.6-mm) shaft hole threaded porcelain standoffs, one noninsulated and two insulated ¼-inch (6.6-mm) shaft couplings, L brackets, battery clip, battery holder, perforated board, several nonreactive resistors whose values have been accurately determined, PL-259 coaxial connectors, convenient lengths of 50-ohm coaxial cable, enamelled copper wire, hookup wire, solder lugs, solder, machine and self-tapping hardware, etc.

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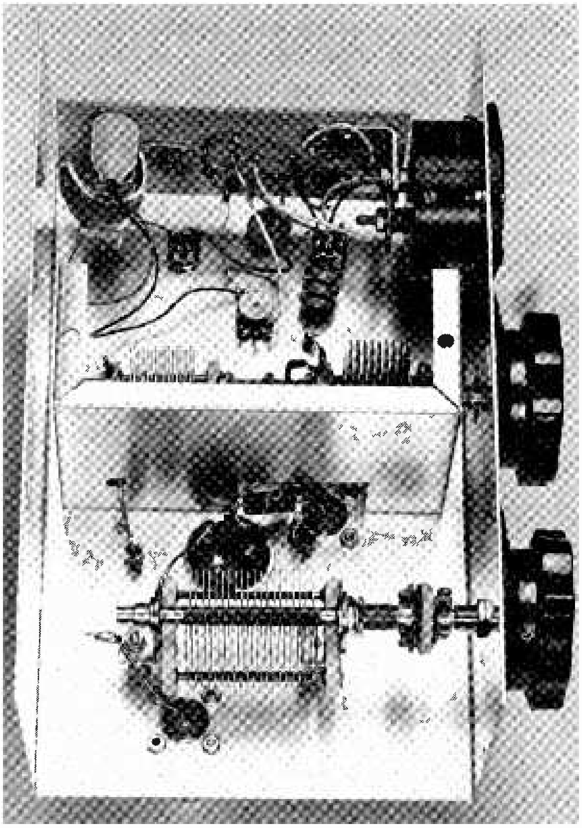

Fig. 4. Photo of author's prototype. Note small shielding box (half of which has been removed) partially obscuring dual differential variable capacitor.

lution of the null, *S1* should be placed in its AMPLIFIED position. The filtered dc is then amplified by *Q1*, which in turn drives the meter movement. Use of the amplifier also increases bridge sensitivity so that the circuit is compatible with low-level signal sources such as solid-state "grid" dippers.

Construction. In any r-f bridge, it is essential that residual and stray reactances be kept to a minimum, and this project is no exception to that rule. Placement of components must be such that lead lengths in the r-f portion of the circuit are absolutely as small as possible. The layout established by the author, which can be seen in the photograph of his prototype (Fig. 4), yielded good results up to 54 MHz.

All of the components were mounted in an aluminum utility box measuring 10" × 6" × 3½" (25.4 × 15.2 × 8.9 cm). The frames and stators of variable capacitors *C1* and *C2* must be insulated from ground (the enclosure), necessitating the use of threaded porcelain spacers or their equivalent. Similarly, insulated couplings should be used with the capacitors' control shafts.

Dual differential capacitor *C1* is partially hidden in the photograph by one half of an aluminum utility box which mounts inside the main enclosure and shields the capacitor from the rest of the bridge. (The other half of the utility box has been removed to expose the capacitor for the photograph.) Dimensions of the box shield used by the author are 5¼" × 3" × 2⅛" (13.4 × 7.6 × 5.4 cm). Totally enclosing the differential capacitor within the grounded utility box helps keep stray reactances small.

To cover 3.5 through 54 MHz with one variable capacitor (*C2*) requires the use of several different inductors. However, band switching of the inductors is not used in this project because it would introduce too much stray reactance and degrade bridge performance. The author's solution to this problem is to use plug-in inductors. J.W. Miller coil forms (No. 42000CBI), ½-inch (1.3-cm) inner-diameter PVC pipe fittings, and 4-prong plugs are used in making the coils.

Details of coil construction are shown in Fig. 5. First, the various coils should be wound on slug-tuned forms. Coil winding data appears in the Table. After the coils are wound, they should be soldered to standard Amphenol four-prong plugs. (Bases removed from discarded four-prong vacuum tubes can be used instead of four-pin plugs.) Take the suggested PVC pipe fittings and modify them as shown in Fig. 5. Then affix each four-pin plug to a modified PVC fitting with cyanoacrylate cement (Eastman 910, "Krazy Glue," or equivalent).

Using a ¼-inch (6.5-mm) bit, drill out the center of a ½-inch (1.3-cm) plastic or metal threaded BX/Romex outdoor electrical-box plug to accommodate the Miller coil form's metal bushing. Mount the threaded plug in the PVC fitting and attach a knob to the coil form's tuning shaft to complete the coil assembly. Repeat this procedure for each inductor.

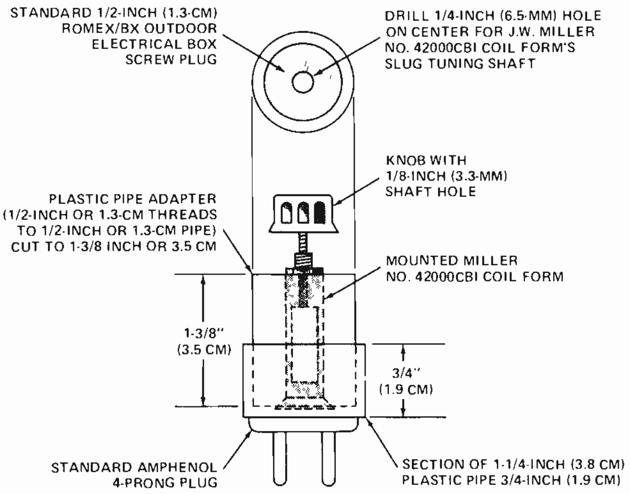

Fig. 5. Assembly details for each plug-in coil (L2).

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Either a commercial dual differential capacitor or a home-brew one can be used for *C1*. The dual differential capacitor should have a capacitance of 12-to-150 pF per section. A Millen No. 28801 dual differential capacitor is suitable, but the author ganged two identical Hammarlund receiving-type variable capacitors rated at 12-to-150 pF each. If two capacitors are used, they should be ganged so that one is at maximum capacitance (plates fully meshed) when the other is at minimum capacitance (plates fully open). The other variable capacitor, *C2*, is rated at 14.7-to-339 pF. A Millen No. 19335 or equivalent component is acceptable.

The frames of both *C1* and *C2* should be mounted on insulated standoffs, and insulated shaft coupling should be used to connect their rotor shafts to the short shafts to which the R and X control knobs are attached. Note that if two capacitors are ganged to form a dual differential unit, a noninsulated coupling should be installed between the two rotor shafts.

Other details of the construction of the author's prototype are apparent in Fig. 4. A portion of the small shield box has been cut away with a nibbling tool to provide room for *J1, R1, R2* and the lead connected to the rotor plates of *C1*. The null detector's amplifier is mounted on a small piece of perforated board which is mechanically supported by L brackets secured to the terminals of *M1*. Because parts placement is critical in the r-f portion of the project, it is best to duplicate the author's layout closely.

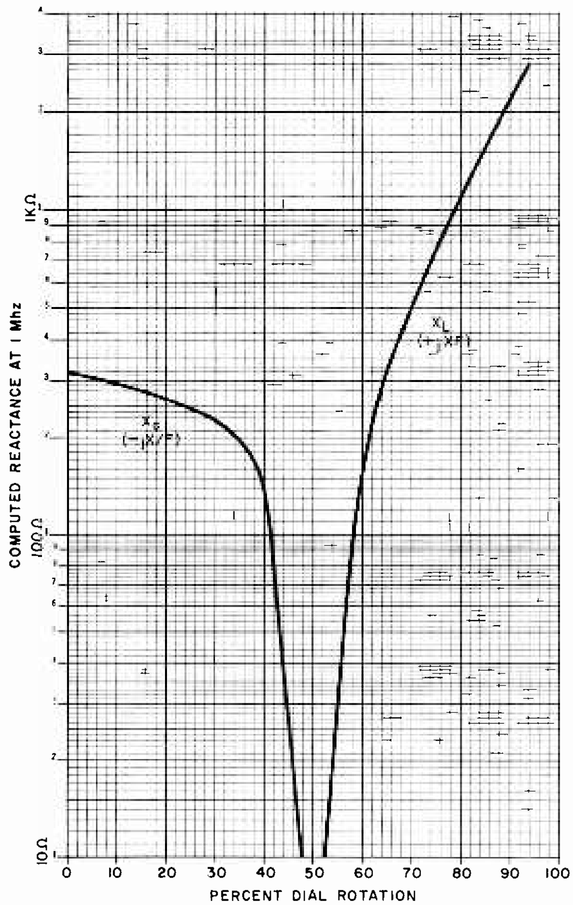

Fig. 6. Calibration curve for the X (reactancescale of C2's control knob.

Calibration. The resistance (R) dial of the impedance bridge can be calibrated by using an assortment of ½-watt carbon composition resistors of various values within the 5-to-200-ohm range. The author measured the exact resistance of each component selected on a General Radio GR-650 bridge to enhance the accuracy of the calibration. If you don't have access to a highly accurate bridge, measure the resistors with a good-quality digital multimeter or use close-tolerance metal film components. Connect the resistors to PL-259 coaxial plugs, keeping lead lengths short.

Calibration should be performed at 3.5 MHz to minimize the effects of reactive strays. Apply the output of a signal generator oscillating at that frequency to *J1* and connect the first load resistor (the one with the lowest resistance) to *J3*. Also, install the 80-meter plug-in coil at *J2*. With *S1* in its DIRECT position and *C2* set at 50% dial rotation (plates half meshed), adjust *C1* and *L1* for the best null possible. Then place *S1* in its AMPLIFIED position and fine-tune for the deepest null you can obtain. Place a notch, tick mark, or other notation of the position on *C1*'s dial. Repeat this procedure for each calibrating resistor.

The author used Bakelite knobs with large skirts as the X and R control knobs. Calibration of the R knob was made by inscribing the appropriate point on the Bakelite skirt with tick marks and numerical values using an electric engraving tool. This technique permits direct calibration of the R knob in ohms. (Suitable Bakelite knobs are available from such surplus electronics dealers as Fair Radio Sales Co., Box 1105, Lima, OH 45802.) Alternatively, a knob with a silver skirt calibrated from 0 to 100 over 180 degrees of dial rotation can be used in conjunction with a graph of dial readings plotted against resistance values.

No direct calibration was performed on the X (reactance) dial. Rather, the following procedure was followed. Using a Southwest Technical Products digital capacitance meter cross-checked against a General Radio GR-650 bridge, the author made a plot of the capacitance of *C2* against dial rotation. Then the standard inductive and capacitive reactance formulas were employed to derive a plot of reactance below and above a resonant frequency of 1 MHz. Assuming that *L2* is adjusted to cancel out bridge reactance (including that of *C2* when its plates are half meshed), the graph shown in Fig. 6 plots the net reactive variation of X_C and X_L below and above the resonance at 1 MHz.

This graph can be used to calibrate the X control knob. For example, at 50% dial rotation, the reactance of the load is 0. At 75% dial rotation, the reactance is +j740 or 740 ohms inductive. Similarly, at 25% dial rotation, the reactance is

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−j250 ohms or 250 ohms capacitive. As was done with the R control knob, a Bakelite knob with a large skirt can be used and the skirt inscribed with an electric engraving tool. Alternatively, a knob with a silver skirt calibrated from 0 to 100 over 180 degrees of dial rotation can be used in conjunction with the graph of Fig. 6 to determine the sign and magnitude of the reactive component.

The accuracy of the X control knob's calibration depends on that of the graph of Fig. 6 and the degree of bridge balance (null sharpness) obtainable. The theoretical curve is apparently very accurate. How much a direct calibration would depart from the curve would depend on stray bridge reactance. The prototype yielded good, sharp nulls and its X calibration was very accurate.

Using the Bridge. Before an unknown impedance can be measured, it is necessary to balance the bridge. Apply an r-f signal to *J1* and connect a nonreactive termination to *J3*. (The author employs a commercial 50-ohm, 5-watt nonreactive termination when performing this step.)

Any signal source producing 1 to 3 volts rms of r-f can be used. A grid-dip oscillator loosely link-coupled to *J1* is satisfactory. The author employs a 4-turn coil of No. 16 enamelled copper wire large enough to accommodate the outer diameter of his grid-dip coil to apply r-f for 80- and 40-meter measurements and 2 turns of the same wire for measurements on 20, 15, 10, and 6 meters. Each coil is connected to a convenient length of 50-ohm coaxial cable, the other end of which is terminated with a PL-259 connector.

Plug the appropriate coil for the frequency at which the measurement is to be performed into jack *J2*. Then set the X control knob to 0 ohms (50% rotation or midscale). Adjust *C1* and *L2* for a good null as indicated by *M1*. After initial adjustments, switch the amplifier into the meter circuit to increase the resolution of the null. If a complete null cannot be obtained, reduce the coupling between the signal source and the bridge.

After the bridge has been balanced, replace the purely resistive load with the unknown impedance. Alternately adjust *C1* and *C2* to obtain the best null and note the readings of the R and X scales. Impedance measurements are in rectangular form. An impedance with an inductive component is of the form $Z = R + jXF$, where R and X are the readings of the R and X scales, respectively. The operator $+j$ denotes inductive reactance, and F is the frequency at which the measurement is performed. An impedance with a capacitive component is of the form $Z = R - jX/F$, where R, X, and F are as defined in the case of a partially inductive impedance. The operator $-j$ denotes capacitive reactance.

As mentioned earlier, the X measurement involves frequency scaling. In the case of an inductive reactance, the exact magnitude is determined by *multiplying* the X scale reading by the frequency at which the measurement is performed. The exact magnitude of a capacitive reactance can be obtained by *dividing* the X scale reading by the frequency at which the measurement is made.

COIL WINDING DATA

Band	Approximate Frequency Range	Coil Data
80 M	3.4 to 4.2 MHz	28 turns of No. 30 enamelled wire, close wound
40 M	6.5 to 7.5 MHz	16 turns of No. 22 enamelled wire, close wound
20 M	13.0 to 15.0 MHz	8 turns of No. 16 enamelled wire, close wound
15 M	19.5 to 22.0 MHz	3½ turns of No. 16 enamelled wire, close wound
10 M	27.0 to 30.0 MHz	2½ turns of No. 16 enamelled wire, close wound
6 M	50.0 to 54.0 MHz	1 turn of No. 16 enamelled wire

All coils are to be wound on a J.W. Miller No. 42000CBI or equivalent slug-tuned form.

In Conclusion. Here are a few hints that you should keep in mind when using this project. Bridge measurements are of course frequency sensitive. The bridge must therefore be rebalanced after a frequency change of 1% or more occurs. Be sure to balance the bridge with a purely resistive test load before performing any measurements. Otherwise, inherent bridge reactances will cause a false reading.

Remember that the bridge requires very little r-f drive. This is no problem when a signal generator or grid-dip oscillator is used as the signal source because the output level of the generator or the coupling between the oscillator and the bridge can be easily reduced. However, if a transmitter is used to provide r-f for the impedance measurement of, say, an antenna or linear amplifier input stage, care must be taken not to overload the bridge. The transmitter's r-f output must be kept at a low level, and the bridge must not be left in the line when more than 0.1 watt of r-f power is flowing.

It is usually very inconvenient to perform impedance measurements directly at an antenna's feed point, so they are commonly performed at the transmitter end of the transmission line. This can result in misleading information if the line is not an integral multiple of an electrical half-wavelength. Note that a transmission line's electrical length is its physical length expressed in free-space wavelengths at the frequency of interest multiplied by the line's *velocity factor*. Solid-dielectric coax (RG-58, RG-59, RG-8, RG-11, etc.) has a velocity factor of approximately 0.66; polyfoam coax has a velocity factor of approximately 0.81.

If it is not convenient to add or subtract enough cable to make the transmission line an integral multiple of an electrical half-wavelength, a Smith chart can be used to transpose the measured impedance at the transmitter end of the line into the actual antenna impedance. To do this, the line length must be accurately determined by physical measurement or by measuring it with a grid-dip oscillator and the far end of the line shorted. Remember that you must employ the electrical length of the line when using the Smith chart.

You are now ready to start using your impedance bridge in r-f work. Its usefulness on your test bench or in your radio shack will be quickly appreciated. ◇

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BY CASS R. LEWART

Make Your Computer Work As a Control Center

Simple circuits enable small-computer owners to perform a variety of external operations.

ONCE YOU tire of playing graphic games on your home computer, have solved all the mathematical problems you care to, and exhausted your list of favorite tunes, you may start thinking about new applications for that wonderful machine. Some of the more attractive uses for a home computer are in the controlling of appliances. In this article, we will present a few simple and proven inexpensive circuits that allow your computer to turn on the coffee pot in the morning, turn lights on and off while you are away to confuse a potential burglar, or control your slide projector and tape recorder in response to various cues.

The great advantage of using a computer to control appliances is its flexibility. No more relays driving relays, where the slightest change in the logic may require redesigning and rewiring your circuit from scratch. A simple change of a few instructions in your program can now accomplish the same objectives relatively painlessly.

Computer Interface. The computer interacts with the outside world by means of I/O (input/output) ports. These ports consist of a connector where specific pins can assume either a high or a low logic status. In most cases, a high corresponds to approximately +5 volts, while a low corresponds essentially to 0 volt (ground). Specific instructions in your program (BASIC or machine language) are used to set voltages to the required values.

As a rule, computer ports can supply only a very small amount of current, usually on the order of 1 mA. Therefore, in order to control any device drawing appreciable power, it is necessary to have interface circuits that translate logic signals from computer ports into relay-contact operations, LED activation, or ac appliance and motor movements.

Because program instructions to control I/O ports differ from one computer to the next, we will not go into details of port programming. Instead, we will assume you are familiar with the programming of your particular computer and know how to set logic signals at its ports low or high.

Some computers use separate ports for input and output, while others use the same ports for both, depending on program instructions. Consult the port operation section in the programming manual for your computer.

In general, when you interface the computer, the program will provide timing and logic for whatever you are doing. Input ports connect to sensors, such as door switches, thermostats, light sensors, etc, while output ports interface to relays, LEDs and solid-state switches. The interface circuits discussed and illustrated in this article deal with computer output ports only.

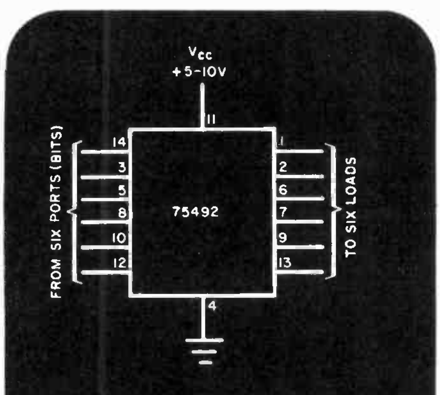

Fig. 1. Six inverters in 75492 can sink up to 200 mA each.

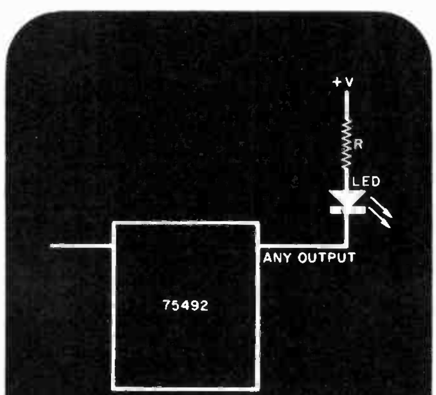

Fig. 2. LED interface circuit used to compute resistance, R.

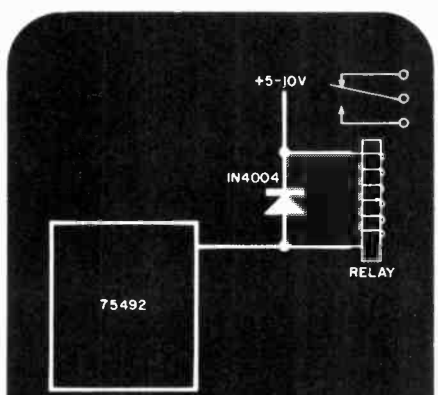

Fig. 3. Low-power relay interface requires protective diode.

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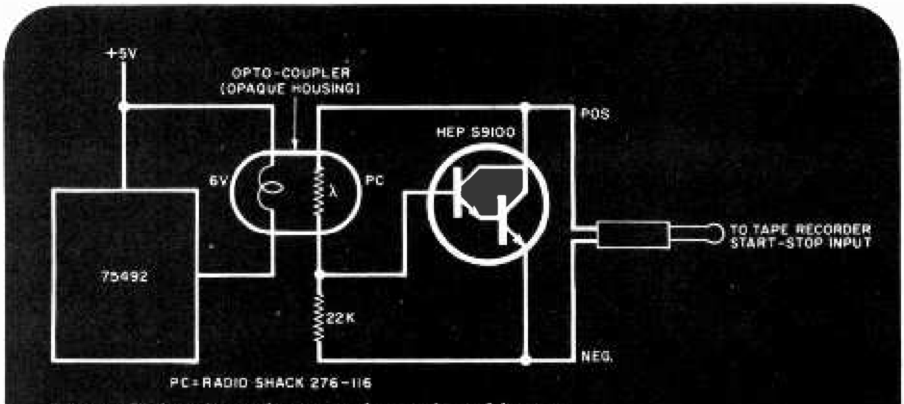

Fig. 4. Relay interface can be replaced by an optoisolator. Here is a homemade version.

Basic Interface. A basic output interface, an inexpensive SN75492 MOS LED-driver IC, is shown in Fig.1. Six computer output-port pins connect directly to the inputs of the device which can sink up to 200 mA on each of its six outputs. This current is sufficient to directly drive a small relay, LED, or optoisolator. All of the interface circuits given in this article employ the SN75492 as the basic building block.

If more than six ports of a computer are being used for control, more than one SN75492 IC can be used. The same port can also drive more than one output (for example, an ac load and a LED to indicate an *on* condition).

LED Interface. Shown in Fig. 2 is a typical LED interface circuit. To compute the values of the dropping resistor in the external circuit, use Ohm's Law: R = E/I, where R is the dropping resistor's value, E is the supply voltage, and I is the current through the LED. Remember to take into account the one-diode voltage drop of the inverter in the IC and the drop across the LED.

As an example of calculating the resistor's value, assume E = 10 volts, I = 20 mA, the voltage dropped across the LED is the typical 1.5 volts, and 0.7 volt is dropped across the internal diode of the inverter. The value of the dropping resistor is R = E/I = (10 - 1.5 - 0.7) /0.02 = 390 ohms. To determine the resistor's power rating, use the formula $P = I^2R$. Plugging in values, we obtain $P = (0.02)^2 \times 390 = 0.156$ watt, which means you can safely use a standard ¼- or ½-watt resistor.

DC Relay Interface. A low-voltage relay whose coil draws less than 200 mA of current can be operated through the output of the IC, as shown in Fig.3. Make sure that the current demand of the relay's coil does not exceed 200 mA, and install a diode as shown to protect the IC from back-emf spikes.

The relay's contacts can be used to turn on and off power for almost any electrical device whose demands are less than the volt-ampere (VA) or current (at the load's operating voltage) rating of the relay's contacts. For heavy loads, the low-power relay can be used to control a power relay with heavy-duty contacts.

Tape-Recorder Interface. Turning on and off a tape recorder under computer control can be very useful for color-slide presentations. Other attractive applications include loading programs from a cassette deck into a computer and storing of programs on tape. The tape deck you wish to control must be equipped with a start/stop control system accessed by way of a jack–usually located near the microphone jack. To turn the tape deck on and off one can connect contacts of a relay (Fig. 3) to a plug inserted in the on/off jack on the tape recorder. If you wish to eliminate the relay, an alternate circuit shown in Fig. 4 uses a Darlington transistor and an optoisolator consisting of a cadmium-sulfide (Cds) photocell and a low-voltage lamp in a light-tight housing. Because this circuit is polarized, it may be necessary to reverse the leads to make the circuit work.

The reason for using an optoisolator in this and the following circuit is to keep the computer and the circuit it controls electrically separate. This is to provide protection for the computer. High insulation resistance between the computer and the ac power line will safeguard low-voltage logic circuits and, not incidentally, the human operator.

Control of AC Appliances. An alternative to a relay or simple light coupler is shown in Fig. 5. The Motorola MOC 3010 is an optoisolator that houses an infrared diode and a small triac. The low power triac, in turn, controls a larger triac, such as the Radio Shack No. 276-1001 that switches the ac power to the load. The rating of the larger triac determines the maximum wattage that can be controlled. For example, the 276-1001 will work with appliances consuming up to 600 watts. Pulsing the appropriate port under program control will result in partial power being delivered to the appliances, allowing the computer to dim lights and run motors at variable speeds.

In Conclusion. The foregoing are just a few possible schemes for interfacing your computer with practical appliances. After you familiarize yourself with these circuits and their capabilities, other schemes may suggest themselves. You may even devise interfaces that you will wish to keep permanently connected. ◇

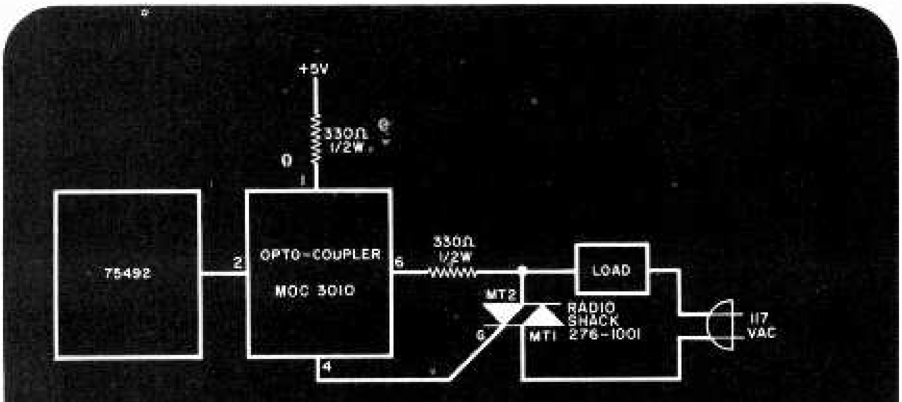

Fig. 5. This circuit, using an infrared optoisolator and a triac, can be used to control loads up to 600 watts.

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A SIMPLE TOUCH CONTROL SWITCH

BY GEORGE PETERKA

Single FET amplifier circuit can be used to control relay or other low-current device

A TOUCH control is an electronic switch that can be activated simply by touching a small conductive plate with a fingertip.

Such controls are easy to build and can be used to enhance many projects. They can also be added to an existing circuit, such as forming an alarm "off" switch for a digital clock.

Circuit Operation. A basic touch control circuit is shown in Fig. 1A. Essentially, it consists of a FET amplifier with a high input impedance (10 megohms) and a conductive touch plate connected to its gate. Operation occurs when the ambient 60-Hz ac field flooding the area is impressed on the touchplate during the finger contact. This signal is amplified and appears at the drain as a 60-Hz square wave, alternating between ground and supply voltage.

Capacitor *C1* shunts any r-f picked up by the "antenna effect" of the touchplate, while capacitor *C2* acts as a transient suppressor.

The drain of *Q1* can be connected to the alarm-off pin of a clock chip, since most of these ICs require that the alarm-off pin be momentarily connected to the supply voltage to silence the alarm.

The circuit of Fig. 1B uses the same FET input stage, but, via *D1*, rectifies the ac waveform at the *Q1* drain and uses the generated positive voltage to turn on transistor *Q2*. The positive voltage developed across *C3* will keep *Q2* turned on until the capacitor is discharged by base current and resistor *Rx*. The value of this latter resistor determines how rapidly the switch will shut off and should be between 10,000 and 100,000 ohms.

The load on *Q2* can be a low-current relay or a resistor (1000 to 5000 ohms) with the signal generated across the resistor used to turn on a high-power transistor. Using the transistor shown for *Q2*, any device that requires 50 mA or less can be powered.

Construction. Any form of construction may be used since the circuit is relatively simple. It should be powered from an ac-line supply for reliable operation.

The touch plate should be relatively small—several square inches are enough. It must be insulated from ground. But it need not be a discrete metal plate; a metal door-knob on a wooden door will suffice. This latter type of touchplate makes an excellent sensor in an alarm project. ◇

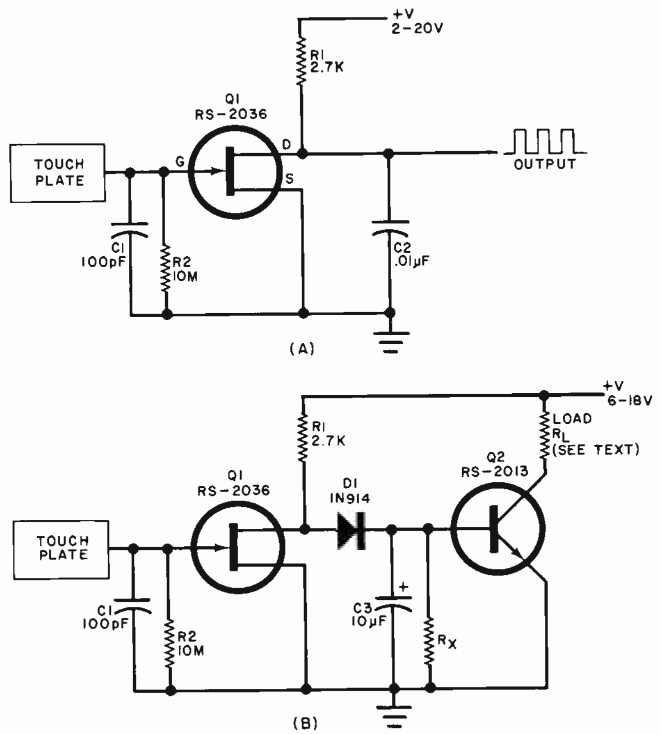

Fig. 1. At (A), high input impedance FET developes a square-wave output when gate is touched by fingertip. Transistor Q2 (B) is added to drive external devices.

PARTS LIST

C1 — 100-pF, disc
C2 — 0.01-µF, disc
C3 — 10-µF, electrolytic
D1 — 1N914 or similar
Q1 — N-channel FET, RS2036 or similar
Q2 — Npn transistor, RS2013 or similar
R1 — 2700 – ohm resistor
R2 — 10 – megohm resistor
Rx — 10,000 to 100,000 ohms (see text)
Touchplate — see text.
Misc. — Perf board, mounting hardware, power supply, etc.

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OPEN-DOOR "FRIDGE ALARM" STOPS FOOD SPOILAGE AND ENERGY WASTE

REFRIGERATORS are among the hungriest of household appliances in terms of electrical power consumption. Every time a refrigerator door is opened, cold air spills out and the warm air that replaces it must be cooled. Needless to say, it pays in dollars and cents to limit the time the door is open to as brief a period as possible. The low-cost Fridge Alarm described here may be just what you need to limit the time you study the contents of your refrigerator or your child forgets to close the door.

The Fridge Alarm is a photoelectric device that is activated as soon as the door opens and the refrigerator's light goes on. It sounds an insistent two-tone signal if the door remains open past a given number of seconds.

About the Circuit. As shown in Fig. 1, when light strikes its photosensitive surface, *Q1* triggers into conduction and causes *Q2* to saturate. This places pin 1 of *IC1* close to ground potential and allows the timer to start operating (Fig. 2). Since the voltage across *C1* is initially zero, *IC1* is triggered into immediate operation. Timing is controlled by *R8*, *R1*, and *C1*.

During the timing sequence, the output of *IC1* at pin 3 remains high (almost at V_{CC}) and keeps *IC2* and *IC3* cut off, since pin 1 of each of these integrated circuits is connected to this line.

Most electrolytic and many aluminum capacitors can have sizable leakage currents. Hence, they should not be used in timing circuits. To avoid this problem, *C1* should be a tantalum capacitor. Using the time constants shown, *R8* can be set for periods of from 4 to 17 seconds. (This range was selected because 8 seconds is about the mean time for access to a refrigerator.) Because *C1* discharges through *D1* and the 15,000-ohm internal resistance of *IC1*, pin 7 is left unconnected.

If the light striking *Q1* is interrupted during the timing cycle, both *Q1* and *Q2* turn off and timing capacitor *C1* rapidly discharges through *D1* and *IC1*, resetting the timer. In darkness, *Q1* has a very high collector-emitter resistance. With *Q2* in cutoff, standby current is extremely low.

Should the light striking *Q1* be constant, the timing cycle will run its course and the output at pin 3 of *IC1* goes low. This effectively grounds pin 1 of both *IC2* and *IC3*, activating these ICs.

Integrated circuits *IC2* and *IC3* are wired to operate as astable multivibrators. The oscillating frequency of *IC2* is about 4 Hz. This 4-Hz signal "modulates" *IC3*, and the output of *IC3* directly drives a small loudspeaker.

The two-tone sound is created by alternately shunting the *IC2* end of *R4* between V_{CC} and ground at a 4-Hz rate. When pin 3 of *IC2* is high, the parallel combination of *R4* and *R5* produces about a 500-Hz tone. When pin 3 is low, *R4* is effectively shunted to ground. This reduces the voltage at pin 7 of *IC3*. Since *C6* must now charge to 80% and then discharge to 40% of this new value to activate the comparators inside *IC3*, about a 330-Hz tone is generated. The two tones alternate at a 4-Hz rate as long as the circuit is activated.

Construction. All components, except *B1* (and its optional battery holder) and the small loudspeaker can be mounted on a printed circuit board. The actual-size etching-and-drilling guide and components-placement guide for the pc board are shown in Fig. 3.

The leads of *Q1* can be identified with the aid of an ohmmeter and light source if an unmarked phototransistor is used.

The project can be mounted inside a small translucent box that permits sufficient light to pass through and trigger *Q1* into conduction. Any of the various polyethylene refrigerator-type storage containers on the market will suffice as long as they are large enough to accommodate the circuit. The loudspeaker is best secured to the bottom of the container (after drilling a number of small holes for

Sounds an alarm after preset time when refrigerator door is left open

BY ELLIOT K. RAND

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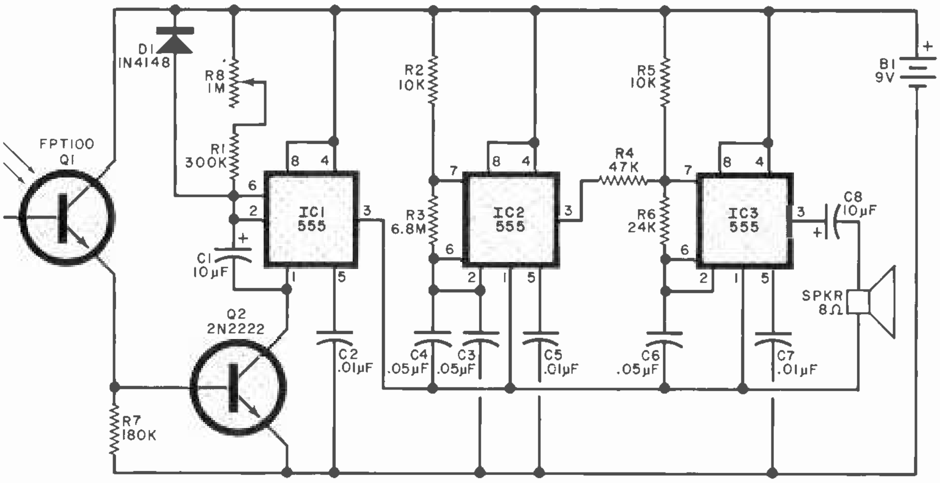

Fig. 1. Timing action of circuit is initiated by light striking Q1.

PARTS LIST

B1—9-volt battery
C1—10-µF, 25-V tantalum capacitor
C2,C5,C7—0.01-µF disc capacitor
C3,C4,C6—0.05-µF disc capacitor
C8—10-µF, 25-V aluminum capacitor
D1—1N4148 or similar diode
IC1, IC2, IC3—555 timer
Q1—FPT100 or equivalent
Q2—2N2222 or similar transistor
All resistors ¼-watt, 10% tolerance:
R1—300,000 ohms
R2, R5—10,000 ohms
R3—6.8 megohms
R4—47,000 ohms
R6—24,000 ohms
R7—180,000 ohms
R8—1-megohm trimmer potentiometer
SPKR—Miniature 8-ohm loudspeaker
Misc.—Battery holder; translucent plastic refrigerator container (about 3" square); silicone-rubber cement; hookup wire; etc.

Note: The following items are available from Rand Laboratories, P.O. Box 468, Cape Canaveral, FL 32920: complete kit of parts including drilled case for $9.95 postpaid. Also available: pc board only, $4.25 postpaid. Florida residents, please add sales tax.

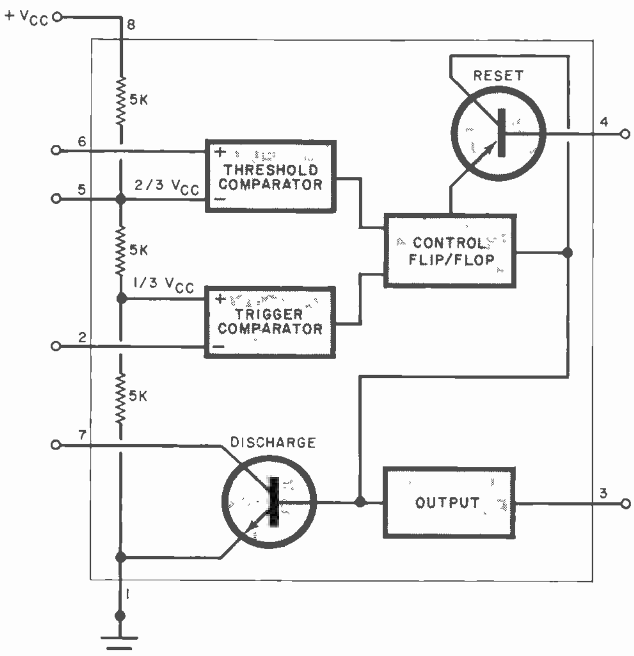

Fig. 2. Block diagram of principal circuits in the 555 IC. In this case, one 555 is used as timer, and two as astable multivibrators.

the sound to escape down through the shelf) with silicone-rubber cement. The speaker and pc board are interconnected with #20 wire so that the board can be positioned to allow maximum exposure of Q1 to the lamp.

The assembled alarm can be tested by placing it in a darkened location and shining a light on it. After a several-second delay, the alarm should sound. Count the number of seconds between the time the light goes on and the alarm sounds. Adjust *R8* as needed for the desired delay between the two events.

Place the Fridge Alarm inside your refrigerator in a location where it will receive the maximum amount of light from the refrigerator's lamp. Make sure it is in a location where there will be no possibility of liquid spills on it. Equally important, make sure that the selected location will obviate any possibility of obstructing the light. ◇

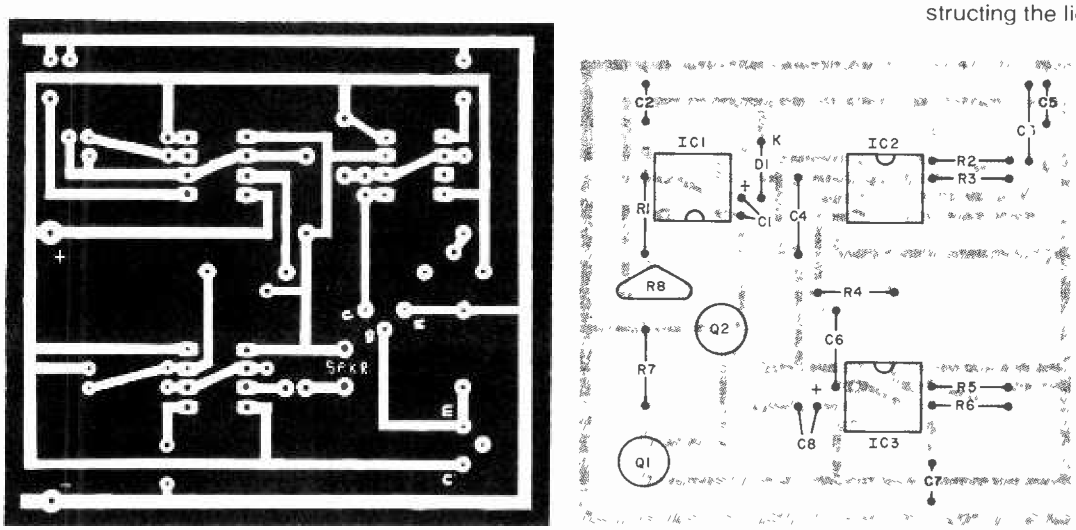

Fig. 3. Components are mounted on board as shown at left and enclosed in a translucent box.

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BY DOUGLAS KOHL

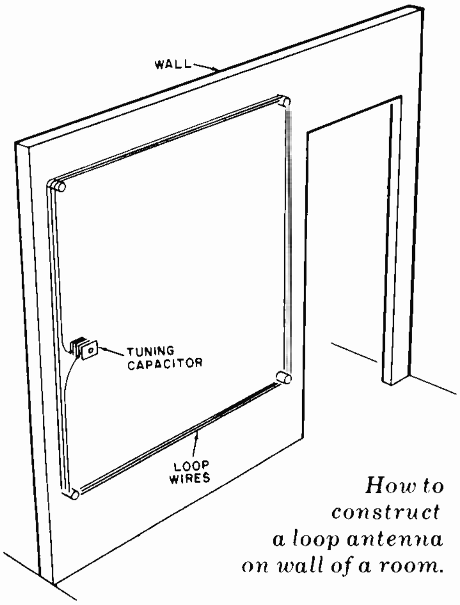
Low-cost LOOP ANTENNA EXTENDS AM RADIO RECEPTION

Easy-to-build air-core loop helps pull in distant stations on inexpensive radios.

EVEN IF you're vacationing too far from home for normal AM reception, you can still pick up home-town broadcasts with an ordinary AM radio. Alternatively, if you're staying at home, you can receive out-of-town sports broadcasts to keep tabs on your favorite team. Using an inexpensive external loop antenna will do the trick. Here's why it works and how to build one.

Because portable and desk-top AM receivers employ relatively small, internally mounted ferrite-core loop antennas, they can deliver only enough signal for good reception of local stations. However, if an external loop with a larger effective cross-sectional area is substituted for the internal one, or used in tandem with it by mutual coupling, the working sensitivity of the receiver is increased in direct proportion to the ratio of the loop areas.

If the loops are used in tandem, no connections or modifications to the receiver are necessary. Signals will be coupled to the small internal loop inductively when the two loops are placed in proximity to each other. If your home is of wood-frame construction and the walls do not have metal lath, you can mount the large loop on a wall or even conceal the loop behind it. The loop can then be a source of fun as a mystery spot where your neighbor's $5 transistor radio will work better than ever before!



How to construct a loop antenna on wall of a room.

Constructing a Loop. A typical large loop antenna is shown in the figure. It is made simply by winding a series of turns of wire on some supporting structure. The loop is tuned by a variable capacitor connected across it. The antenna can be supported by wooden pegs inserted into the wall or by a free-standing wooden cross frame. Insulated copper wire, No. 20 or larger, should be used. Bell wire or even No. 14 house wire will yield excellent results. Such a loop can be concealed if other members of the family consider it unsightly.

Plan to make your loop square, or at most, slightly rectangular. This makes it easy to compute the area inside the loop. Construct your loop so that it is as large as possible. A 7-ft × 9-ft (2.1-× 2.7-m) loop, for example, is suitable if you have 8-ft (2.4-m) walls. If possible, mount the loop on a wall which is in line with the distant radio stations you want to receive. The antenna is most sensitive to signals parallel to the plane of the loop, and least sensitive to signals propagating in directions perpendicular to it (striking the antenna broadsides).

To calculate how many turns of wire are needed, compute the area of the proposed loop and use the following formula: $N = 242.3/\sqrt{A}$ where N is the number of turns and A is the area in square inches (1 square inch = 6.45 cm^2). For example, suppose the planned loop will measure 6′ 9″ (2.1 m) on each side. Its area will be 6,561 square inches (4.2 m^2 or 42,330 cm^2) and the number of turns required will be three. For your convenience, here are the loop sizes corresponding to an integral number of turns:

N	Length of each side
3	80-11/16″ (2.05 m)
4	60-5/8″ (1.60 m)
5	48½″ (1.23 m)
6	40-3/8″ (1.03 m)
7	34-5/8″ (88 cm)
8	30-5/16″ (77 cm)
9	26-7/8″ (68.3 cm)
10	24-3/16″ (61.4 cm)

Incidentally, you can make a small, portable loop on a wooden frame to take along on picnics, or on a boat. A loop two feet square (0.3716 m^2 or 3716 cm^2) will provide good results with a "pocket portable" receiver.

Connect the loop ends to each side of an ordinary air dielectric variable tuning capacitor (one loop end to the rotor plates and the other to the stator plates). The capacitor, which can be removed from a junked AM receiver or purchased new (or surplus), should have a maximum capacitance of at least 360 pF. Multisection capacitors can be wired in parallel to extend the loop's tuning range. Be sure to solder all connections using rosin core solder.

Using the Loop. A loop antenna will provide some improvement in reception of all stations, not just the one at the frequency to which it is tuned. However, for best results the loop should be resonated. Tune the receiver to the desired station's frequency and place it in the vicinity of the large loop. Orient the receiver so that its internal ferrite bar is perpendicular to the plane of the loop. Then rotate the shaft of the antenna's tuning capacitor until the signal peaks.

Enhanced reception will be experienced when the receiver is placed up to approximately one side dimension of the loop in front of, or behind the wall on which the loop is mounted. Experiment with the placement of the receiver to determine the location that gives best results. The closer the receiver is to the loop, the more signal coupled to the internal ferrite antenna. For casual listening, as opposed to chasing weak DX signals, the degree of coupling between the loop and the receiver will not be critical, thanks to the large measure of improvement the loop provides. ◇

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A SIMPLE PRECISION POWER SUPPLY for your work bench

BY FRAN HOFFART

Adjustable (1¼ to 33 volts) supply delivers up to 1½ amperes with excellent regulation

ONE ITEM that belongs on every experimenter's work bench is a source of clean, regulated dc. The ideal hobbyist supply would be relatively inexpensive, easily built, adjustable over a fairly wide range of output voltage, and capable of sourcing an ampere or more of dc. In addition, it would have a high degree of line and load regulation and contain such protection as automatic current limiting, maximum power limiting, and thermal shutdown.

The project presented here satisfies these requirements handily. It is built around an LM317, a monolithic variable-voltage regulator IC that can provide an output voltage from 1.25 to 40 volts. The supply can generate output currents up to 1.5 amperes at 1.25 to 33 volts. It has a low parts count and is rugged enough to withstand the abuse to which most bench supplies are subjected at one time or another.

Before we examine the power supply circuit, let's first look at the LM317 variable-voltage regulator IC. Because this chip is the essence of the supply, a prior understanding of its operation will simplify our later discussion of the power supply as a whole.

The LM317 Regulator IC. Shown in Fig. 1 is a simple schematic which illustrates the basic operation of the LM317. The integrated circuit keeps the voltage drop between the output terminal (the case of the regulator, which is housed in a TO-3 package) and the adjustment terminal (Pin 1) a constant 1.25 volts. In practice, resistor *R1* is connected between these two terminals, thereby setting up a constant adjustment current. The magnitude of this current and the setting of potentiometer *R2* determine the output voltage of the regulator.

If the adjustment current is sufficiently large, the output terminal is always 1.25 volts more positive than the adjustment terminal. Accordingly, setting the wiper of *R2* so that the adjustment terminal of the IC is grounded causes the LM317 to act as a 1.25-volt regulator. Rotating the control shaft of *R2* elevates the adjustment terminal above ground, simultaneously increasing the voltage at the output terminal.

Any voltage greater than 1.25 volts can be obtained at the output terminal simply by increasing the resistance between the adjustment terminal and ground. Although the manufacturer (National Semiconductor) rates the IC's maximum differential input-to-output voltage at 40 volts, the LM317 can be used to provide higher regulated voltages. However, such operation necessitates the inclusion of additional components to protect the regulator from excessive differential voltages.

Note that the LM317 has no ground terminal. This means that all quiescent operating current for the IC must flow through its output terminal, and necessitates that a minimum load current be established if the regulator is to function properly. Shown in Fig. 2 is a plot of the minimum operating current required against the differential input-to-output voltage. A convenient way to satisfy this minimum-load-current requirement is to select a value for the adjustment current that is suitably greater than the quiescent operating current of the IC.

The internal design of the LM317 makes possible a wide variety of applications other than that of a series-pass voltage regulator. These include tracking preregulators, switching regulators, ac voltage regulators, two-terminal current regulators, and power regulators, to name just a few!

Even more important than the inherent simplicity and flexibility of the three-terminal regulator IC is its ability to protect itself from practically every type of overload condition, thereby greatly in-

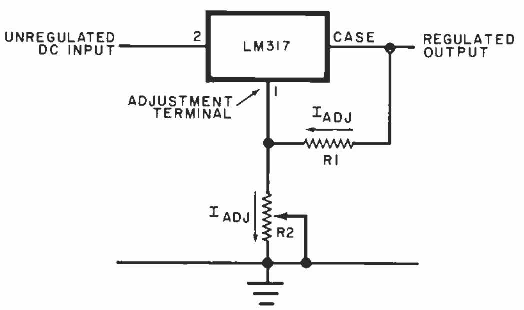

Fig. 1. Basic circuit showing the operation of the LM317 regulator integrated circuit.

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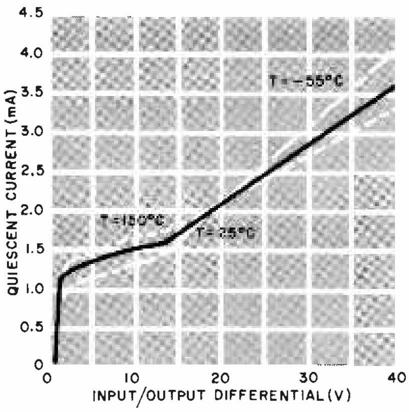

Figure. 2. Plots of minimum load currents vs. differential input-to-out voltage.

creasing reliability. Output current is limited to 2.2 A to protect the IC as well as the power transformer and rectifier. Safe-area protection limits the maximum power dissipated by the regulator to approximately 20 W, thus guarding the series-pass transistor located on the chip against a destructive secondary breakdown. The safe-area protection circuit decreases the maximum possible output current as the differential input-to-output voltage increases, thereby limiting the power dissipated to a safe value. A plot of output current limiting versus differential input-to-output voltage is shown in Fig. 3.

Thermal protection built into the LM 317 limits the maximum chip temperature to approximately 170°C. This protects the regulator from overheating, regardless of the type of overload or the amount of heat sinking provided. The temperature is sensed on the chip at a point near the series-pass transistor, enabling the regulator to shut down quickly if a potentially destructive overload condition occurs. Once the overload has been corrected and the chip cools down, the regulator turns back on and resumes normal operation.

All these protective circuits remain functional as long as an input-to-output differential of at least 2 volts exists, even if the adjustment terminal is accidentally disconnected from the rest of the circuit.

About the Circuit. The complete schematic of a 32-volt, 1.5-ampere bench power supply is shown in Fig. 4. Using the LM317 voltage regulator greatly simplifies the design and construction of the supply but keeps performance and reliability at high levels.

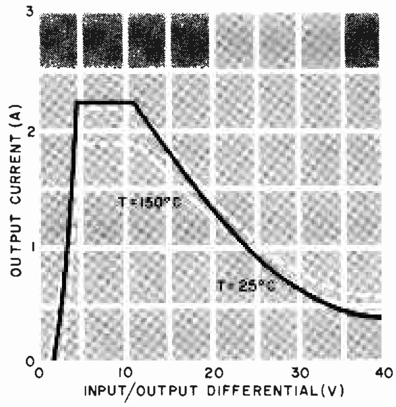

Fig. 3. The 20-watt safe-area curves for the LM317 regulator IC at three operating temperatures.

Power transformer *T1* is a "universal" multiwinding unit. Switch *S2* selects one of two primary winding configurations, causing the ac input to the full-wave bridge rectifier to vary from 18 volts in the LOW position to 32 volts in the HIGH position. This minimizes power dissipation by the LM317 regulator. It also allows full output current to be generated at low voltages by reducing the input voltage to the regulator when the supply is being used at low output-voltage lev-

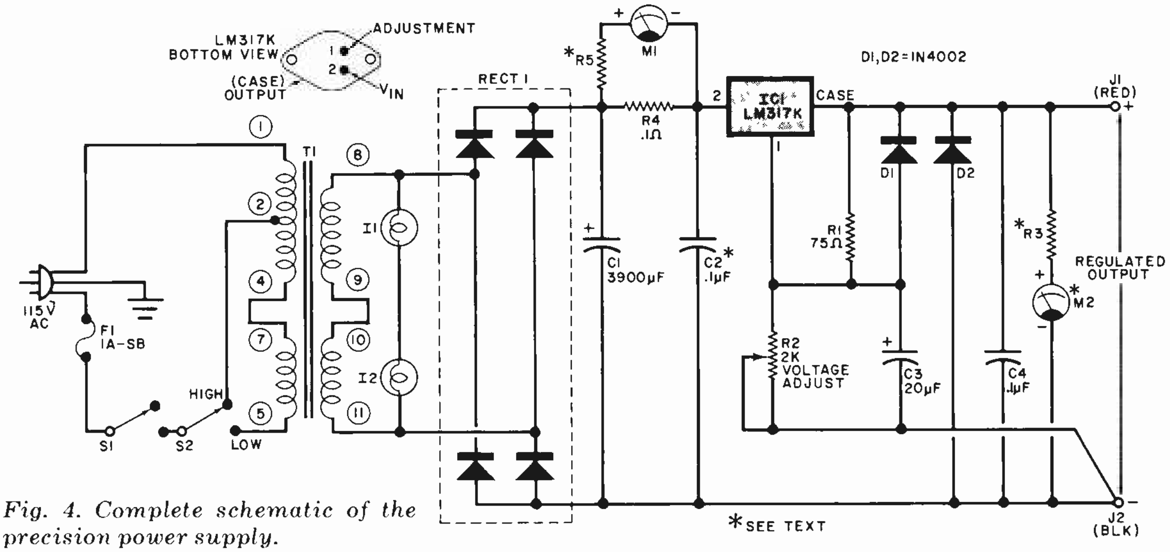

Fig. 4. Complete schematic of the precision power supply.

PARTS LIST

C1—3900-μF, 50-V electrolytic
C2,C4—0.1-μF disc ceramic (C2 optional)
C3—20-μF, 50-V electrolytic
D1,D2—1N4002
F1—1-A slow-blow fuse
IC1—LM317K TO-3 voltage regulator (National Semiconductor)
I1,I2—15-V pilot lamp
J1,J2—Color-coded 5-way binding post
M1—1-mA, 2-inch (5.1-cm) panel meter, relabeled to read 0 to 1.5 A.
M2—1-mA, 2-inch (5.1-cm) panel meter, relabeled to read 0 to 30 V.
R1—75-ohm, 1%, ¼-W metal-film resistor
R2—2000-ohm, 10-turn potentiometer
R3—Select for individual meter used (approximately 30,000 ohms for 1-mA meter movement)
R4—0.1-ohm, 5%, ½-W resistor
R5—Select for individual meter used (typically 10 to 100 ohms for 1-mA meter movement)
RECT1—2-A, 100-PIV modular bridge rectifier
S1,S2—Spdt miniature toggle switch
T1—30-V, 2-A secondary "universal" power transformer (Stancor RT-202 or equivalent)
Misc.—Suitable aluminum enclosure (LMB-564 or similar), heat sink, TO-3 socket and mica washer, silicone thermal compound, line cord, fuse holder, control knob, rubber feet, dry-transfer lettering, hookup wire, solder, hardware, etc.

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els. An added benefit is a cooler-running heat sink.

Filter capacitor *C1* keeps the peak-to-peak ripple voltage under 2 volts at the input of the regulator, resulting in less than 300 microvolts rms of ripple at the output. Ceramic disc *C2* should be located close to the IC regulator and, although it is designated in the Parts List as optional, it is required if filter capacitor *C1* is located more than 4 inches (10.2 cm) from the IC. It is good practice, however, to include *C2* even if *C1* is close to the LM317 regulator.

Resistors *R4* and *R5* are the current shunt and calibrating resistors, respectively, for milliammeter *M1*. Note that *R4* is located on the input side of the regulator rather than the output side, so that it will not degrade load regulation. The exact value of *R5* will depend on the characteristics of the particular meter used for *M1*. It will usually be between 10 and 100 ohms if a 0-to-1-mA meter movement is employed.

Precision metal-film resistor *R1* establishes an adjustment current of 16 mA which flows through VOLTAGE ADJUST potentiometer *R2*. Adjusting *R2* for maximum resistance places the adjustment terminal of the IC at 32 volts above ground. This sets the power supply output voltage at 33.25 volts required. For a high degree of resolution in adjusting the output voltage, a ten-turn potentiometer is specified for *R2*. However, a lower-cost, single-turn potentiometer could be substituted if your budget won't accommodate a precision component or if you have difficulty procuring one.

Capacitor *C3* filters out any ripple voltage appearing at the adjustment terminal, increasing the ripple rejection at high output voltages. Transient response and stability of the power supply are improved by the addition of *C4*. Diode *D1* provides a discharge path for *C3* in the event of a short circuit at the supply output. The IC regulator is protected by *D2* against reverse voltages that might be accidentally applied to the output of the supply.

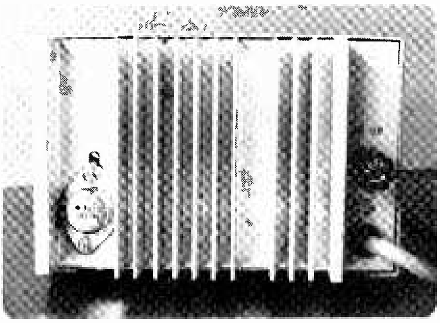

Fig. 6. Rear view of heat sink on prototype. A mica washer provides good thermal contact between regulator and heat sink.

Resistor *R3* is the calibration component for the output voltmeter *M2*. Both this meter and *M1* are standard 1-mA meter movements with the meter faces relabeled. The exact value of *R3* will depend on the characteristics of the individual meter used for *M2*. It will be approximately 30,000 ohms if a 0-to-1-mA meter movement is employed. Incandescent lamps *I1* and *I2* illuminate the supply's meters and act as pilot lights.

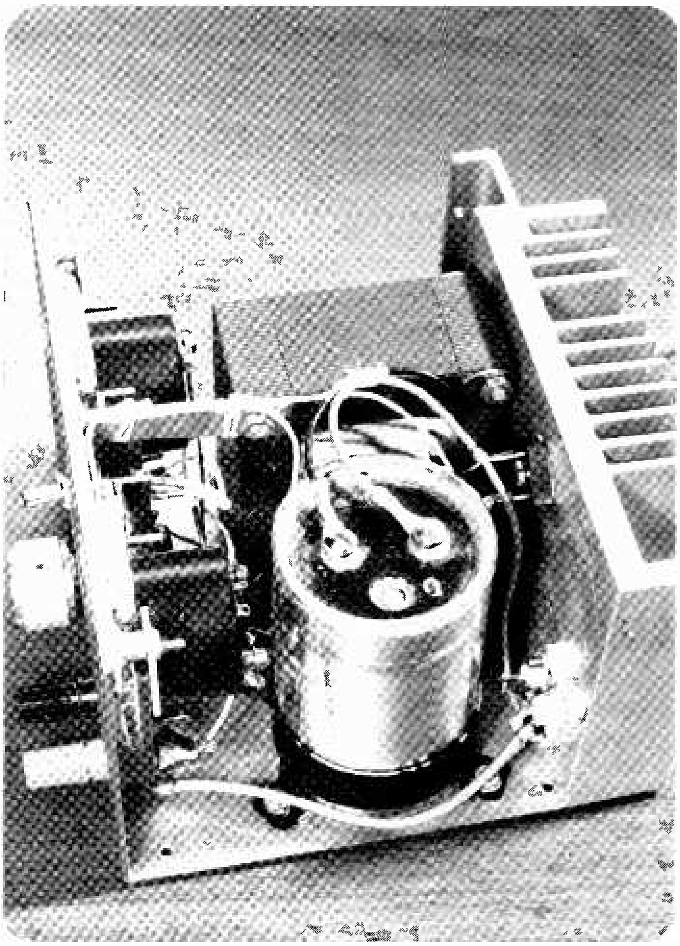

Fig. 5. Interior view of the author's prototype. Heat sink must limit IC temperature to 75°C above ambient.

Construction. The 1.5-ampere bench supply was constructed in an aluminum enclosure measuring 4"H by 6"W by 5"D (10.2 x 15.2 x 12.7 cm). The back of the case was removed and replaced with an aluminum heat sink containing thirteen 1-inch (2.5-cm) fins. An interior view of the author's prototype is shown in Fig. 5. The heat sink selected must be of sufficient size to limit the regulator temperature to no more than approximately 75°C above ambient when dissipating a maximum of 25 watts.

A mica washer will provide good thermal conductivity between the case of the LM317 regulator and the heat sink while maintaining electrical isolation. Be sure to apply a layer of silicone thermal compound on each side of the mica washer. Also bolted to the heat sink of the author's prototype is the bridge rectifier, *RECT1*. The rectifier, however, is mounted on the inside of the modular heat sink and is not visible in the rear view of Fig. 6.

After drilling holes for the various components, and cutting the front panel for the meters, the cabinet was painted and labeled with dry-transfer lettering. Next, using suitable hardware, all components except the filter capacitor and the meters were mounted. This provided

(Continued on page 126)

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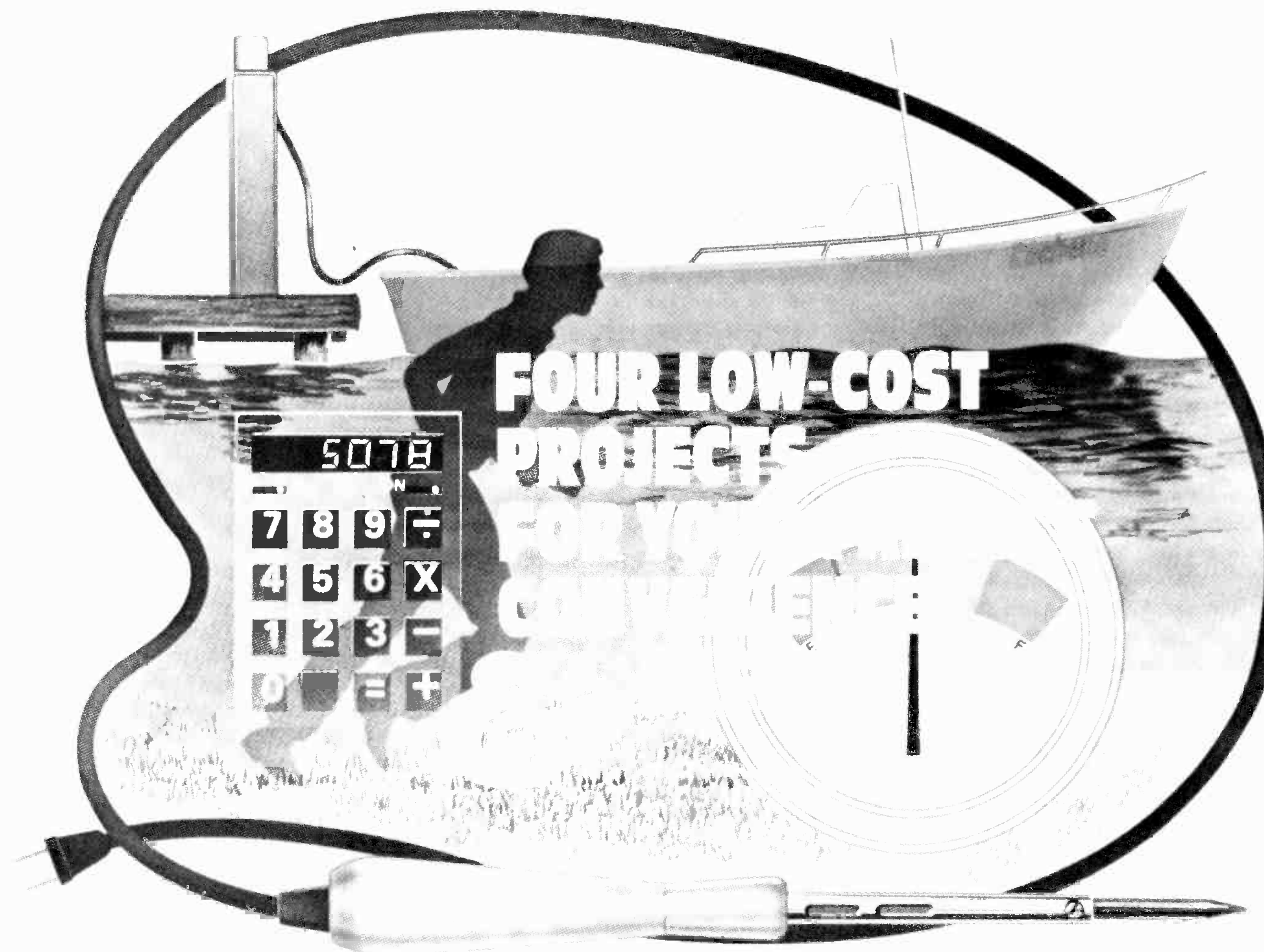

1. Solid-state level-sensing switch for sump pumps

By Phillip Windolph

A FLOODED basement is a minor household disaster. That's why most homeowners whose basements are prone to flooding install sump pumps. Some, however, have discovered to their chagrin that the pump somehow fails to operate when it is most needed. In many instances of failure, the pump itself is actually in perfect working order. Rather, it is the water-detecting actuator switch that's the culprit, never sending a turn-on command to the pump.

Here's a simple, dependable circuit to replace the often-unreliable (usually mechanical) switch supplied as part of the pump assembly. It will automatically activate the pump when the water level reaches the level of a pump trigger probe. Once activated, the pump will remain energized until the water level falls below a keep-alive probe. If the pump fails or cannot keep the water in check, an optional alarm will sound as the water level reaches a trigger probe specifically for that purpose. The project can be powered either by batteries or the ac line. Inexpensive components are employed, most of which will be found in any well-stocked junk box.

About the Circuit. The Electronic Sump Pump Switch is shown schematically in the figure. Positive voltage from the power supply is applied to the COMMON probe via resistors *R1* and *R2*. (This and all other probes are stiff wires or metal rods suspended above and extending to different levels in the sump.)

Replaces often-unreliable pump switch and sounds alert if water level continues to rise or pump isn't working

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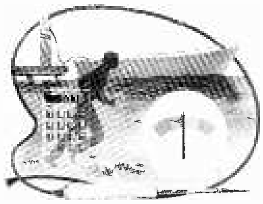


Low-cost Projects continued...

As can be seen in the figure, the COMMON probe extends almost to the bottom of the sump. Any water entering the sump comes into contact with this probe, but as yet nothing which would cause activation of the pump happens.

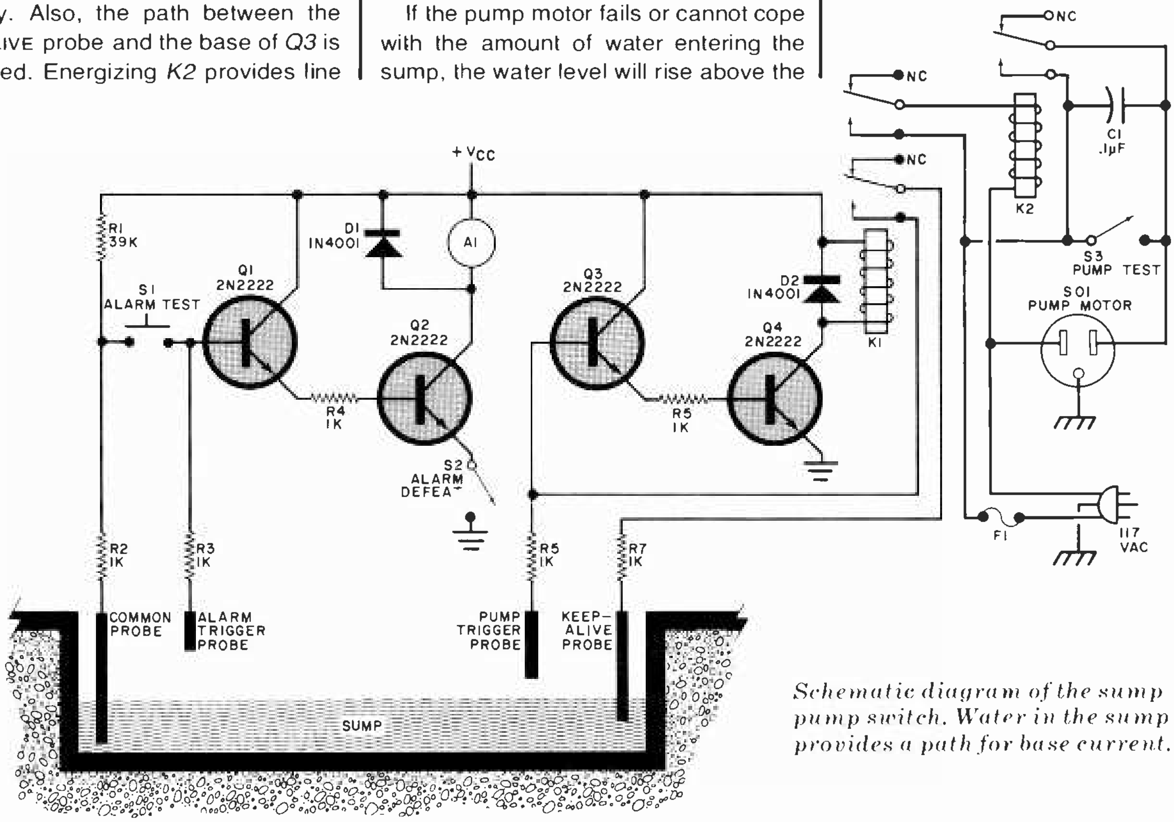
As the water level in the sump rises, the KEEP-ALIVE probe touches the water, but this still does not activate the pump. If the water reaches the level of the PUMP TRIGGER probe, current can flow from the positive supply voltage terminal through *R1*, *R2*, the water in the sump, *R5* and finally into the base of *Q3*. This transistor then turns on and provides base current for *Q4*. When *Q4* conducts, it energizes the coil of relay *K1*.

Once this relay is energized, the normally open contacts are closed and two things happen. Line current is able to flow through the coil of *K2*, a heavy-duty ac relay. Also, the path between the KEEP-ALIVE probe and the base of *Q3* is completed. Energizing *K2* provides line voltage across *S01* for the pump. If the sump pump is connected to the socket, it will be activated and will start to pump the water out of the sump.

As the water level drops, the conductive path provided by the water in the sump between the COMMON and PUMP TRIGGER probes will be interrupted. However, current will continue to reach the base of *Q3* via the KEEP-ALIVE probe, *R7*, and one set of contacts of relay *K1*. Because this probe extends almost to the bottom of the sump, relays *K1* and *K2* remain energized (as does the pump motor) until practically all of the water has been evacuated. When the water level drops below the free end of the KEEP-ALIVE probe, *Q3* is deprived of base current and is cut off. This causes *Q4* to stop conducting, deenergizing *K1*, *K2* and the pump motor.

If the pump motor fails or cannot cope with the amount of water entering the sump, the water level will rise above the PUMP TRIGGER and KEEP ALIVE probes and eventually reach the ALARM TRIGGER probe. This probe is part of the optional alarm circuit and should be mounted near the top of the sump. Although the alarm circuit is independent of the pump controller, it is a valuable addition to the project.

The alarm circuit closely resembles that of the pump controller and operates in a similar manner. Water reaching the ALARM TRIGGER probe provides a path for current to reach the base of *Q1*. This transistor begins to conduct and provides base drive for *Q2*. Transistor *Q2* then conducts and completes the circuit for audible alarm *A1*, which alerts the homeowner to the fact that water in the sump has risen to a critically high level. He can then try to get the pump working



Schematic diagram of the sump pump switch. Water in the sump provides a path for base current.

PARTS LIST

A1—Dc-energized buzzer, bell, Sonalert™ or similar audible alarm*
C1—0.1-µF, 1000-volt disc ceramic
D1,D2—1N4001 rectifier
F1—Fast-blow fuse*
K1—Dc-energized relay*
K2—117-volt relay*
Q1 through Q4—2N2222 or similar npn switching transistor*

The following are ¼-watt, 10% tolerance carbon-composition resistors:
R1—39,000 ohms*
R2 through R7—1000 ohms*
S1—Normally open pushbutton switch
S2—Miniature spst toggle switch
S3—Spst toggle switch*

SO1—Ac power socket
Misc—Line-powered, regulated or battery dc supply*; suitable enclosure; barrier terminal strip; perforated board; fuseholder; line cord; metal rods or stiff, solid-conductor wire; hookup wire; solder; self-tapping and machine hardware, etc.

*See text.

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or, if necessary, bail the water out of the sump manually.

Two switches are associated with the alarm circuit and one switch is included in the pump controller. These switches provide test facilities for the alarm and pump (*S1* and *S3*, respectively) and the ability to silence the alarm (*S2*). The currents handled by *S1* and *S2* are relatively small, so miniature components can be used in these locations. Switch *S3*, however, as well as the contacts of *K2* must be capable of handling the current demanded by the pump motor, so use heavy-duty components.

The author employed a solenoid/spring-type buzzer as his prototype's audible alarm. Diode *D1* is connected across the buzzer to protect *Q2* from inductive spikes generated by the buzzer. Other types of alarms can be used, some of which will not require the inclusion of *D1*. A Sonalert™ or similar audio oscillator will not necessitate diode protection for *Q2*, but an alarm bell will.

Which type of audible alarm you choose is largely a matter of personal preference and parts availability. Similarly, there is a great deal of leeway in the choice of components *Q1* through *Q4* and *R1* through *R7*. General-purpose 2N2222 transistors are suggested in the parts list. Just about any low-power npn transistor is suitable for use as *Q1* and *Q3*. Exactly which transistor types are acceptable for use as *Q2* and *Q4* depends on the audible alarm and relay (*K1*) used. If the current demand of either load is fairly low, say, 300 mA or less, a general-purpose component such as type 2N2222 can be used as a relay or alarm driver.

However, if a load draws more than 300 mA, a higher-power driver will have to be used. A good rule of thumb is to use a transistor with a collector current rating that is double the current required by the alarm or relay coil. The author employed a sensitive 6-volt relay for *K1* (Sigma No. 70R4T-6DC), which permitted the use of a low-power npn driver. Diode *D2* was included to protect the relay driver from inductive spikes.

The values specified for the fixed resistors (*R1* through *R7*) are nominal ones. Substitutions can be made freely if you want to use components you have on hand. However, do not make the fixed resistances so low that they tax the base current ratings of the transistors employed in the project.

Either a line-powered or battery supply can be used for the project. The exact value of supply voltage is not critical and can be chosen to accommodate a particular dc relay (*K1*). Practical supply voltages range from 6 to 15 volts. Although it is not necessary, voltage regulation is desirable in a line-powered supply. The widespread availability of voltage regulator ICs makes the inclusion of regulation simple and inexpensive.

If the alarm circuit is included in the project, battery power enjoys a significant advantage over a line-powered supply—it will still provide power to the project if the commercial power line is blacked out. Of course, if line power is not available, the pump motor will not be activated, even though *K1* will be energized. The alarm circuit, however, will be activated if the water in the sump rises to the level of the ALARM TRIGGER probe. This will alert the homeowner that water is accumulating and had best be bailed out before any damage occurs. Also, when neither the alarm nor pump controller circuit is triggered, practically no current is drawn from the battery supply. If nonrechargeable batteries are used to power the project, long operational life can be expected.

Construction. The circuit is relatively simple, which suggests the use of perforated board and point-to-point wiring techniques. Remote mounting of the alarm and pump controller circuits will simplify any future servicing. If this is done, the circuit board, relays, switches and power supply can be housed in a suitable enclosure which can be installed at some convenient location.

A four-terminal barrier strip can be mounted on the control box for the leads running to the sump probes. These probes can be fashioned from either metal rods or lengths of solid No. 12 or No. 14 copper wire and should be mounted rigidly above the sump. The probes are of varying length, with the COMMON probe extending almost to the bottom of the sump, the KEEP-ALIVE probe extending almost as deeply, the PUMP TRIGGER probe reaching about halfway down, and the ALARM TRIGGER probe extending only a short distance into the sump. Suitable lengths of hookup wire should be soldered to the probes and routed to the barrier terminal strip on the control box.

When constructing the control box, be sure to observe the polarities of all semiconductors and, if a line-powered supply is built in, of electrolytic capacitors. Use the minimum amount of solder and heat consistent with making good connections. Take special care in wiring the 117-volt ac portions of the project so that no shock hazard is present.

Checkout and Installation. After the control box has been wired, connect short lengths of hookup wire to the barrier terminal strip. Remove a portion of the insulation from the free end of each wire. Next, fill a drinking glass or measuring cup with water and place the wire connected to the COMMON probe terminal into the water. Place the wire connected to the KEEP-ALIVE probe terminal into the water. (Keep these and all probes from touching each other to realistically simulate actual operation. No damage will occur, however, if the probes accidentally come into contact.) Activation of the pump controller, indicated by a click as the relays are energized, should not yet happen.

Now insert the wire connected to the PUMP TRIGGER probe terminal into the water. You should hear a click as the relays are energized. If desired, a lamp can be connected to power socket *SO1* and the line cord connected to the power line (assuming this has not yet been done). The lamp will then indicate that the relays are energized and that line power is reaching socket *SO1*.

Remove the PUMP TRIGGER wire from the water. The relays should remain energized and no click should be heard. Then remove the KEEP ALIVE wire from the water. At this time, the relays should drop out and a click heard. Finally, insert the ALARM TRIGGER wire into the water. The alarm should sound and remain on until the wire is removed from the water.

Press the ALARM TEST pushbutton and keep it depressed. The alarm should sound and remain activated until the ALARM DEFEAT switch is opened. The operation of the PUMP TEST switch can be checked by closing it and observing whether the load connected to socket *SO1* receives line power.

Once it has been determined that the control box is functioning properly, a permanent installation can be made. Mount the control box at some convenient point and interconnect it with the sump probes and pump motor. Be sure to bypass the stock pump-activating switch as it is no longer needed. As a final check, you can quickly fill the sump with water. The alarm should sound until the pump has lowered the water level beyond the reach of the ALARM TRIGGER probe. The pump should remain on until the KEEP ALIVE probe is no longer immersed, at which point nearly all of the water will have been taken out. ◇

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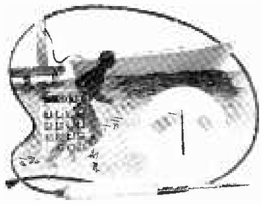

Low-cost Projects continued...

2. Vehicle low-fuel indicator

Alarm sounds when level in vehicle fuel tank drops to a predetermined level

By Bradley Albing

RUNNING out of gas can be an exasperating experience. The low-fuel indicator described here can help you avoid this situation. It will sound an alarm when the fuel level in your gas tank reaches a predetermined minimum. This level can be preset by a simple potentiometer adjustment.

Circuit Operation. In most vehicles, the fuel-level sensor is a float-controlled potentiometer (sender) wired in series with the dashboard-mounted fuel gauge (meter) and connected between the chassis and + 12-volt line as shown in Fig. 1. As the fuel level changes, the resistance changes, making the meter indication change.

The voltage level thus generated across the fuel-level sensor can be tapped off (at the meter) and, as shown in Fig. 2, applied through a low-pass filter *R8-C4* so that the voltage across *C4* is the average across the sender. This low-pass filter also eliminates any rapid voltage fluctuations due to gasoline sloshing and a bouncing sensor float, or

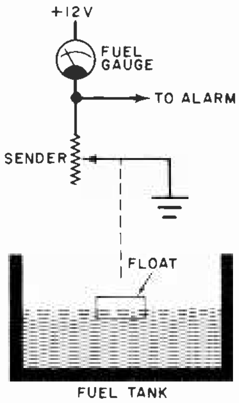

Fig. 1. Typical fuel-gauge circuit.

voltage transients generated by the switching voltage regulator as used in some vehicles.

The *C4* voltage is applied to the non-inverting (+) input of comparator *IC1*, and rises with decreasing fuel in the tank. When this voltage exceeds the *R4* preset voltage on the inverting (-) input, the output of *IC1* (pin 6) goes high.

This voltage (approximately 9 volts) is high enough to cause zener diode *D6* to conduct and turn on transistor *Q1*. When turned on, this transistor draws current through audible alarm *A1*, and turns on optional indicator *LED1*.

As long as the fuel level is low, the output of *IC1* remains high. To silence the alarm until the tank is filled, CANCEL switch *S1* is depressed to trigger *SCR1*. When triggered, the SCR brings the junction of *R5-D6* (the input to *Q1*) down to approximately 2.2 volts, which is not high enough to cause *D6* to conduct and activate the alarm circuit. Since the SCR is powered by dc, it will remain turned on as long as the *IC1* output is high (the fue. level is low).

As long as *SCR1* is conducting, there will be about 1.2 volts (two diode drops) across *D7* and *D8*, enough to turn on *Q2* and cause *LED2* to operate. This LED is a special type that incorporates a built-in flasher circuit that makes the LED flash at a 2.5-Hz rate as long as the LED is

Cable on author's prototype has connector for +12 volts, ground and tank sender unit.

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PARTS LIST

A1—Sonalert, buzzer or other 12-volt alarm (Radio Shack 273-060 or similar)
C1,C2—100-μF, 25-V aluminum electrolytic
C3,C5—0.1-μF, 25-V disc or Mylar
C4—300-μF, 15-V tantalum electrolytic
D1,D7,D8—1N914
D2—1N5742, 18-V, 400-mW zener
D3,D4,D9—1N751A, 5.1-V, 400-mW zener
D5—1N4001
D6—1N5732, 6.8-V, 400-mW zener
IC1—3140E op amp
LED1—red LED
LED2—Litronix FRL-4403 flashing LED (Radio Shack 276-036)
Q1—2N3053 or similar
Q2—2N3904 or similar
The following are ¼-watt, 10% tol. resistors.
R1,R11—100 ohms
R2—33 ohms
R3,R5,R12—470 ohms
R6—10 megohms
R7—470,000 ohms
R8—33,000 ohms
R9—330 ohms
R10—10,000 ohms
R13—820 ohms
R14—200 ohms
R4—25,000 ohm potentiometer
SCR1—2N5062
S1—normally open pushbutton switch
Misc.—Suitable enclosure (Radio Shack 270-285 or similar), interconnecting leads, mounting hardware

Note: The following are available from BFA Electronics, P.O. Box 212, Northfield, OH 44067: No. LF-2 pc board for $4.50 + $1 p&h; No. LF-2-KIT all parts in Fig. 2 except case and connector for $18 + $2 p&h. Ohio residents, please add sales tax.

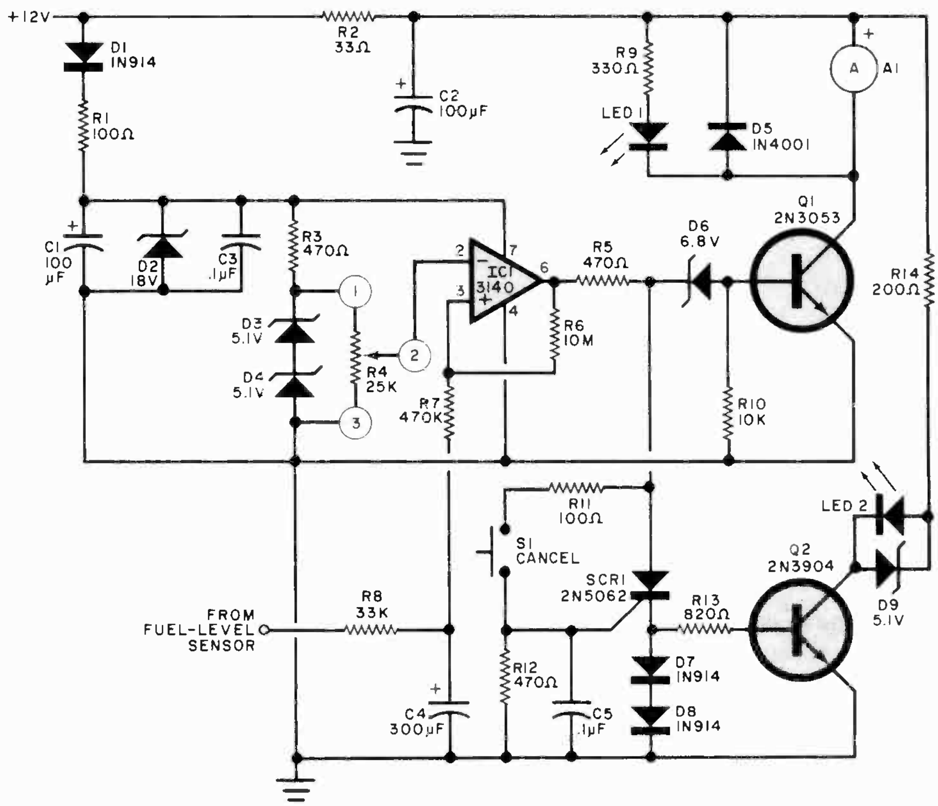
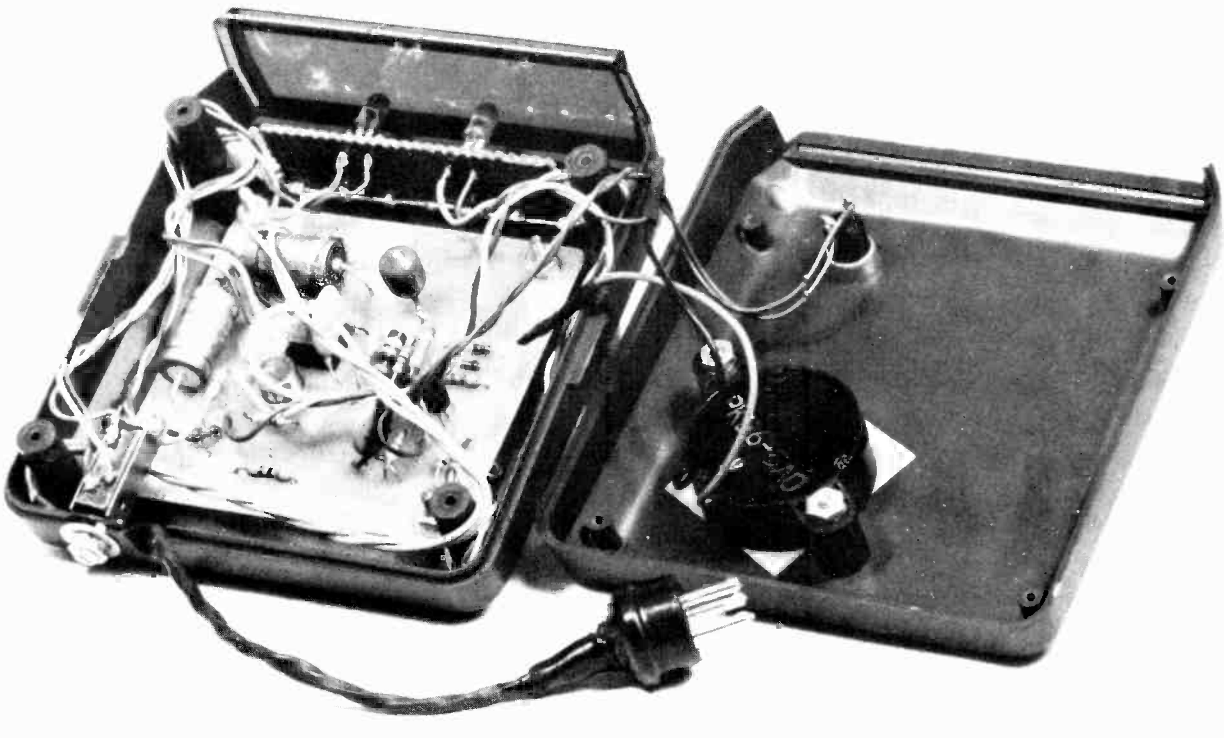


Fig. 2. Comparator IC1 turns on when fuel drops below some predetermined level, and sounds the alarm. The SCR circuit energizes a flashing LED during the Cancel mode.

PC board mounted in prototype with alarm and CANCEL switch on top.



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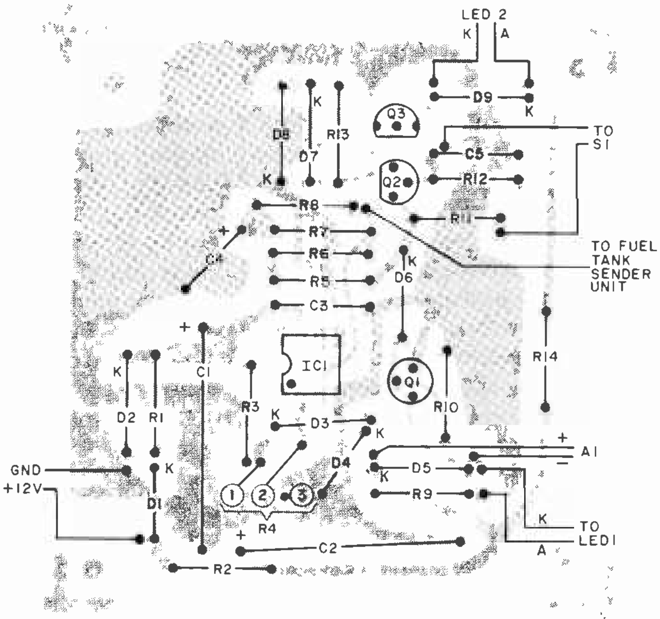

Fig. 3. Actual-size etching and drilling guide is shown at left. Component placement guide is above.

powered. The maximum voltage permitted across this special LED is 6 volts, hence the presence of 5.1-volt zener diode *D9*.

The incoming dc power line is noise decoupled by *R1*, *C1* and *C3*. Zener diode *D2* clamps any transients to a maximum of 18 volts while diode *D1* makes sure that the correct polarity is supplied to *IC1*. Filter *R2-C2* decouples the power line to the alarm and indicator circuit. Diode *D5* clamps any voltage spikes that may occur if an inductive load, such as a buzzer, is used as the alarm. Resistor *R6*, connected between the output of *IC1* (pin 6) and the noninverting (+) input, adds a small amount of positive feedback to give the comparator a little hysteresis and speed up the transition from low to high. This also reduces the likelihood of comparator oscillation.

Construction. The circuit may be constructed on perf board, Wire-Wrapped, or on a pc board such as that shown in Fig. 3 along with the component installation.

The two LED indicators, CANCEL switch *S1*, level-select potentiometer *R4*, and the selected audible alarm are not mounted on the pc board.

The finished pc board can be mounted within a selected enclosure that will also mount the off-board components. Power can be derived from any +12-volt source that becomes active when the vehicle ignition key is operated. The ground can be any convenient metal element that is solidly connected to the vehicle chassis.

You will have to locate the dashboard end of the fuel sensor lead. Test this lead by measuring the voltage across it with various levels of fuel. Usually, the lower the fuel level, the higher the voltage. It is possible for this voltage to vary due to the action of the vehicle switching voltage regulator (if your vehicle uses one) so this must be considered.

If you have any doubt as to the type and wiring of the fuel-level sensor in your vehicle, consult the vehicle repair manual.

Calibration. There are two ways to calibrate the system. The first is to wait until the fuel level is down to the selected low level, then adjust *R4* until the alarm sounds off.

The second approach is to disconnect the fuel gauge from its feed line to the fuel sender but leave the lead connected to the low-fuel alarm, then connect a resistor-substitution box between the fuel gauge and ground (as a substitute for the fuel sender). Adjust the value of the resistor until the fuel gauge indicates the desired level. Adjust *R4* to sound the alarm at that point. Disconnect the resistor box and replace the fuel sender line.

Once the fuel-level turn-on point has been determined, depress *S1* to silence the alarm. After the tank is filled, the alarm will be reset until the fuel level drops below the predetermined point. ◇

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By Cass Lewart

3. Portable gas leak meter

Ultra-sensitive instrument gives quantitative indication of natural gas, propane, fuel vapors, etc.

TOXIC and explosive gases are an ever-present danger in our modern society. They include natural gas, propane, fuel vapors, and invisible and odorless carbon monoxide.

The ultra-sensitive gas-leak detector presented here indicates the quantitative presence of these gases and enables one to track down and pinpoint the *source* of a gas leak by observing the unit's meter indication. Moreover, it is a portable, battery-powered model for use in boats, cars at campsites, etc.

Circuit Operation. The gas sensor, *GS1* in Fig. 1, consists of an electrically heated tin-oxide pellet that changes resistance when exposed to carbon monoxide, hydrogen, propane, alcohol, gasoline vapor, and other oxygen-reducing gases. Power for the circuit can be obtained from either six D cells, preferably rechargeable, connected in series or from an optional 9-volt battery eliminator. Regulator *IC1* reduces the available 9-volt level to the 5 volts required by the circuit. Optional *LED1* is a

Current from the regulator heats gas sensor *GS1*'s semiconductor pellet. The sensor, *R4*, *R7*, and *R8* are arranged in a bridge configuration. The null indicator consists of *M1* and *R6*, while *D1* and *D2* serve as protection for *M1*. Overall circuit sensitivity is determined by the value of resistor *R5*, while *S2* provides a BATT. TEST function.

Once the bridge is balanced, by NULL potentiometer *R8*, any change in the resistance of *GS1* will create an unbalanced condition. When this occurs, the meter's pointer swings up-scale, by an amount proportional to the change in resistance of *GS1*.

Construction. With the exception of *GS1*, *J1*, *M1*, *B1*, *R8*, *S1*, and *S2*, the circuit can be assembled on a piece of perforated board. Select an enclosure large enough to accommodate the board and all off-board components, including *B1* and its holder.

PARTS LIST

B1—Six D cells in series
D1,D2—Germanium diode (1N34A or similar)
GS1—Model 812 gas sensor
IC1—5-volt regulator (Radio Shack No. 276–1770 or similar)
J1—Normally closed miniature phone jack (Radio Shack No. 274–281 or similar)
LED1—Red light emitting diode
M1—50-µA meter (Radio Shack No. 22–051. No substitute)
R1—1000-ohm, ½-W, 10% resistor
R2—22,000-ohm, ½-W, 10% resistor
R3,R7—4700-ohm, ½-W, 10% resistor
R4—15,000-ohm, ½-W, 10% resistor
R5—2700-ohm, ½-W, 10% resistor
R6—2200-ohm, ½-W, 10% resistor
R8—10,000-ohm linear potentiometer
S1—Spst switch
S2—Dpdt switch
Misc.—7-pin miniature tube socket; battery holder; enclosure; 9-volt dc calculator-type ac adapter (optional); machine hardware; hookup wire; solder; etc.
*Available for $10.00 postpaid from Southwest Technical Products Dept., PE-2, 219 W. Rhapsody, San Antonio, TX 78216.

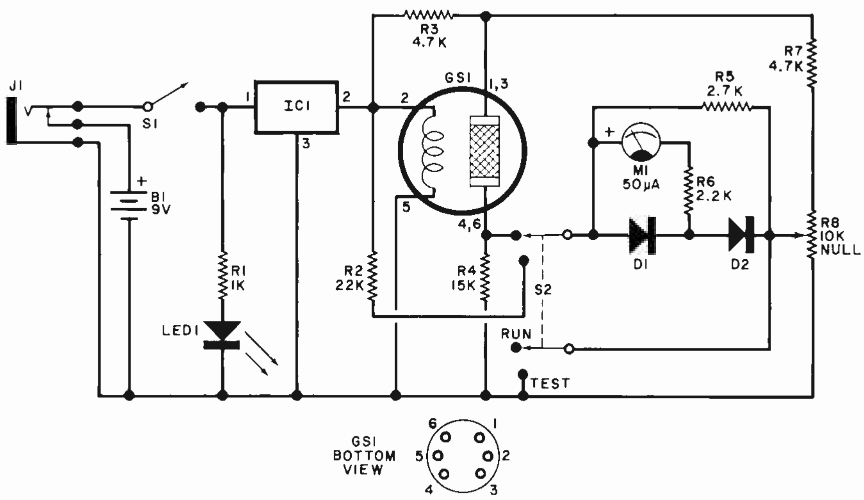

Fig. 1. The gas sensor forms one arm of a Wheatstone bridge. Pins 1, 2 and 3 can be interchanged with pins 4, 5 and 6. Once bridge is balanced by R8, a change in resistance of GS1 will cause meter pointer to swing upscale.

Mount the meter movement on one side of the enclosure's front panel, the remaining off-board components (except *B1* and *J1*) on the other side of the panel. The battery holder and optional battery-eliminator/charger jack *J1* are best mounted on the rear wall of the enclosure, set *S1* to ON and *S2* to BATT. TEST, and make a note of the point on the meter's scale at which the pointer comes to rest. Turn off the power and carefully remove the cover from the meter's face. Use a felt marker to identify the battery-test point on the meter's scale. *S2* to RUN and, in a neutral atmosphere, adjust NULL control *R8* until the meter indicates zero. Now, place a drop of alcohol or gasoline on a finger and approach the sensor. The meter pointer should swing up-scale. Move the finger away from the sensor; it will take a min-

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closure. If desired, *GS1* can be mounted either directly on the front panel or in a separate housing, the latter fitted with a cable to connect it to the main enclosure. The sensor itself takes a miniature 7-pin tube socket.

After the project is assembled, install a fresh set of D cells in the battery holder.

Operation. Set *S1* to ON and allow the sensor to heat up for about two minutes. Set *S2* to BATT. TEST and check that sufficient voltage is available from the battery. (A set of fresh D cells will last about 20 hours. An external 9-volt battery-eliminator/charger can be used.)

After the sensor has warmed up, set ute or so for the sensor to settle back for the next measurement. Readjustment of *R8* may be necessary occasionally. If setting time is too long, change *R7* to 1000 ohms.

When looking for a gas leak, note locations where the meter swings upscale to narrow down the location. ◇

4. Electronic pedometer for joggers

By Andrew A. Modla

How to convert a calculator into a pedometer to record distance covered while walking or jogging.

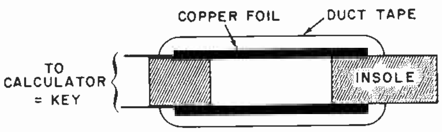
AN INEXPENSIVE pocket calculator can be converted to operate as an electronic pedometer to keep an ongoing tally of the number of steps taken while walking and jogging. Then, with a simple conversion, you can use the calculator to determine the number of yards, meters, miles, or kilometers travelled. Although the conversion described here is "hard wired" into the calculator, you sacrifice none of the calculator's basic built-in capability.

Calculator Conversion. The first thing you must do is determine whether or not your calculator has a built-in constant function. To do this, press CLEAR, 1, +, 1, =, =. If your calculator has the necessary constant function, the display should read 3 and should increment by 1 for each additional operation of the = key. Having established the fact that your calculator does indeed have the constant function, you can proceed with the conversion.

Conversion of the calculator consists in simply wiring a foot-operated switch across the = key. First, carefully open the calculator's case and locate the contacts for the = key. Then solder a 5' (1.5-meter) or so length of 26-gauge flexible stranded wire to each = switch contact. Insulate the soldered connections with a layer of electrical tape.

Now, test your hookups in the following manner. Turn on the calculator and key in 1, +, 1. Touch together and separate the free ends of the wires two times. With the first touch, the display should read 2 and with the second, 3. If the test checks out properly, turn off the calculator and reassemble it, routing the wires out through the side of the case. If necessary, use a sharp knife to cut a slot to allow the wires to exit the case. No other modification is necessary.

Footswitch Fabrication. As shown in the drawing, the footswitch is fabricated from a commercially available "airpillow" foam insole. Begin by cutting a 1" (25.4-cm) square away from the center of the heel area of the insole. Cement a square of copper-coated Mylar or any other flexible conductive material over the cutout on both sides of the insole, conductive surfaces face-to-face.



Place copper foil on each side of insole hole and insulate with tape.

Solder the free ends of the flexible wires from the calculator to the conductive material. Then cover the "switch" assembly with duct or other durable tape to keep out dirt and moisture.

Slide the insole into your shoe and put on the shoe. Turn on the calculator and key in 1, +, 1. Now, as you walk around, the display should read 2, then 3, then 4, etc., as you successively put weight on the switch shoe with each step. If you do not obtain these results, turn off the calculator and carefully check out the switch arrangement.

Determining Distance. Every time you use the pedometer, you must first key in 1, +, 1. Thereafter, the calculator increments the display by 1 for each step taken by the shoe in which the switch is installed. To determine how far you have run or walked, you must find out how many steps you take in a given measured distance (mile, kilometer, etc.). You must, therefore, measure off the "control" distance and walk or run it to determine how many steps are required to cover the course.

Let us assume you wish to know how many miles you have walked and have previously determined that it takes you 1056 steps to walk a mile. (Note that a step is two strides. If the switch is in your right shoe, a step is completed every time you set down your right foot.) Now, subtract 1 from the total displayed by the calculator. This is necessary because the first step you take will register 2. If we assume you stopped at 7200 steps, simply divide this number by 1056, your "control" number, using the calculator to obtain the number of miles walked. Therefore, 7200/1056 = 6.82 miles. ◇

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BUILD AN

IN-CIRCUIT TRANSISTOR TESTER FOR $15

By Jules Gilder

Indicates transistor quality and type without unsoldering from a circuit

LOCATING a bad transistor on a circuit board crowded with components all soldered in place can be a vexing problem. With an in-circuit transistor tester, however, you can determine the component's general quality and also avoid damaging components and/or the foil pattern due to excessive soldering-iron heat.

The simple, low-cost (under $10) tester described here will indicate when a suspect transistor is good or bad and, as a bonus, tell you the component's type (pnp or npn). Indication is through a pair of flashing LEDs. One LED flashes if the device is a good pnp transistor, while the other LED flashes if the device is a good npn type. If it is not good, either both LEDs will flash or neither will flash, depending on the type of transistor failure.

Circuit Operation. The circuit, shown in Fig. 1, is based on a 555 (*IC1*) timer operating as a 12-Hz multivibrator. The output at pin 3 drives one flip-flop of *IC2*. This flip-flop divides the input frequency by two, but more important, delivers complementary voltage outputs at pins 15 (Q) and 14 (not-Q).

These complementary outputs are connected to indicators *LED1* and *LED2* via current-limiting resistor *R3*. The LED's are arranged so that when the polarity across the circuit is one way, only one LED will glow, and when the polarity reverses, the other LED glows. Thus, with no transistor being tested, the LED's flash alternately.

The *IC2* complementary outputs are also connected to resistor network *R4* and *R5*, with the junction of these two resistors connected to the base of the transistor under test.

With a good transistor connected to the B, C and E terminations, when the correct voltage is applied to the three connectors, the transistor will turn on. This produces a short circuit across the LED pair. For example, when a pnp transistor is under test, during the interval when the Q output is low and the not-Q output is high, the pnp device will turn on. In this mode, *LED1* is shorted, *LED2* is reverse biased and, for that half cycle, neither LED will glow. On the next half cycle, the conditions of Q and not-Q are reversed with Q high and not-Q low. Under these conditions, *LED1* is off because it is reverse biased, and since the pnp transistor is cut off, it does not prevent *LED2* from glowing. Thus, when testing a good pnp device, *LED2* will flash, and when testing a good npn type, *LED1* will flash.

If the transistor under test is open, both LEDs will flash. If the transistor has an internal collector-to-emitter short, neither LED will flash.

To compensate for low-valued resistors that may be present in the circuit being tested, *R4* was selected to supply a large amount of base current to the transistor under test. This makes it possible to overcome in-circuit resistances across the collector-base or base-emitter junctions of as little as 40 ohms.

Diodes *D1* through *D4* become important if the transistor being tested has an internal short between its collector-base or base-emitter junctions. In such a case, half of the transistor acts like a diode and would normally conduct and indicate a good transistor. To overcome the possibility of this type of problem's occurring, diodes *D1* through *D4* are added in series with the collector.

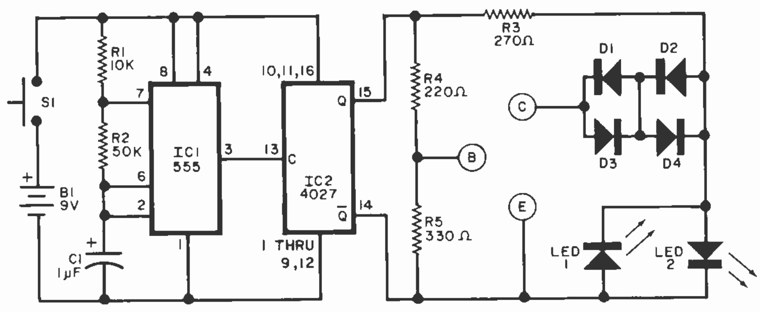

Fig. 1. As shown above, circuit is based on 555 timer operating as a 12-Hz multivibrator.

PARTS LIST

B1—9-volt battery with holder
C1—1-µF, 16-volt electrolytic
D1 through D4—1N4148 or similar
IC1—555 timer
IC2—4027 dual flip-flop
LED1, LED2—Light emitting diode
R1—10,000-ohms, ½-watt, 10% resistor
R2—50,000-ohms, ½-watt, 10% resistor
R3—270-ohms, ½-watt, 10% resistor
R4—220 ohms, ½-watt, 10% resistor
R5—330-ohms, ½-watt, 10% resistor
S1—Normally open spst pushbutton switch
Misc.—Suitable enclosure, mounting hardware, et al.

Note: The following is available from Redlig Systems Inc., 2068 79 St., Brooklyn, NY 11214: kit of parts (no battery or case) for $15, including postage. NY residents, please add 8% sales tax.

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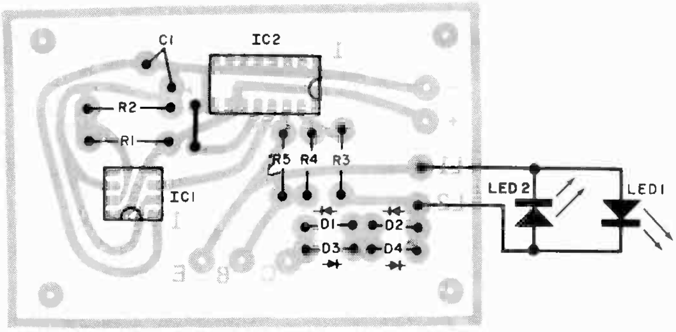

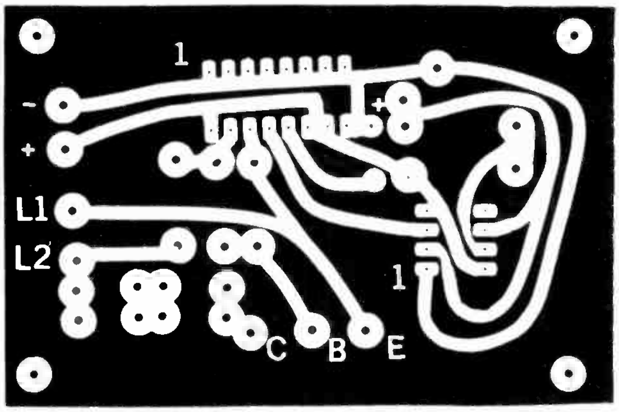

Fig. 2. Component placement guide (above) and actual-size foil pattern (at right).

When *D1* and *D2* or *D3* and *D4* are conducting, they create a voltage drop of about 1.2 volts across the operating pair. This voltage adds to the drop across the transistor being tested, and if the transistor is good, the drop across it will be about 0.1-volt, and the total drop across the LED's will be 1.3 volt for the half cycle that the transistor is turned on. This is not enough voltage to turn on the appropriate LED. If, on the other hand, the transistor has a base-emitter or base-collector short, the 1.2 volts of diode drop is added to another 0.6-volt drop to produce a total drop of 1.8 volts— enough to turn on the LED. Therefore, internal shorts will cause both LED's to flash alternately.

Construction. The circuit is not critical with regard to parts placement and can be built up on a small piece of perforated board or the pc board whose foil pattern is shown in Fig. 2. Sockets for the ICs are optional, and be sure to observe the polarity of *D1* through *D4* and *C1*. The three leads to be connected to the transistor under test can be terminated at a transistor socket, or used as three color-coded leads terminated with small alligator clips or some form of needle tip to make in-circuit transistor pad connections.

The completed pc board can be mounted within a small enclosure that will also support the battery (and holder) and on/off switch *S1*. The two LEDs can be mounted on the cover using rubber grommets. To check the tester, depress *S1* and note that the two LEDs alternately flash. If they flash together, one of them is improperly connected. ◇

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PLAY "SPACE BATTLE" ON YOUR VIDEO MONITOR

Here's an exciting visual game program for any computer using an RCA COSMAC 1802 CPU

BY DONALD R. SCHROYER

THIS program for 1802-based systems equipped with an 1861 video display, puts you and your opponent in command of a pair of space vehicles. The ships move in accordance with commands entered via the hex keypad, and fire their weapons via a pair of pushbutton switches tied to input lines EF3 and EF4. When a player has scored eight hits against his opponent, the losing ship is displayed as destroyed. At the end of the game, the score is displayed. About 2K of memory is required.

The Program. This consists of an initialization routine, six subroutines, the main program, six data pointers and six data locations. These are shown in Tables I through IX.

The Input and Movement subroutine shown in Table I inputs data via the hex keypad. Depressing keys 0, 4, 8, or C results in movement of the left ship. Keys 0 and 4 move the ship downward with 0 causing motion at twice the speed of key 4. Keys 8 and C move the left ship up with the rate of movement doubled when key C is depressed. Keys 1, 5, 9, or D freeze the left ship where it is.

Keys 3, 7, B, or F operate the right

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ship in a like manner. Keys 2, 6, A, or E freeze the right ship. If your system has a hex display, the last two movements will be displayed.

In operation, movement of the two ships is accomplished by changing R(7) the left ship position pointer, and R(8) the right ship position pointer. The ships are moved up by subtracting 08 from the pertinent position pointer. Downward motion is performed by adding 08 to the pointers. To double the speed, 10 is added or subtracted. The subroutine tests the last position of each ship to assure that neither moves out of the screen's display area.

The Load Counter R(5) subroutine of Table II replicates the data pointed to by R(3) at the display location pointed to by R(7). The data pointed to by R(3) forms the silhouette of the left ship. The silhouette of the right ship, pointed to by R(4) is loaded into the display area pointed to by R(8).

The Left Fire Counter R(A) subroutine of Table III tests the switch wired to $\overline{EF4}$. If this line is low, the data flag is set and R(9) is called, otherwise the main program continues. The Right Fire Counter R(B) subroutine of Table IV tests $\overline{EF3}$ in the same manner.

The Weapons Counter R(9) subroutine of Table V is brought into play only when called by subroutines R(A) or R(B). When entered, the Weapons subroutine tests the data flag and fires either the left or right ship's weapon as appropriate. The weapons of both ships cannot be fired simultaneously, and this is compensated for in the main program. The Weapons subroutine also tests each firing to see if a hit has been scored. If so, the score stored in R(E).1 is incremented by adding 10 for a left hit and 01 for a right hit. Register R(E).1 stores the left score in the high four bits and the right score in the low four bits. The Weapons subroutine tests to see if either score is equal to eight. If so, the game ends with the score displayed, otherwise the program continues.

The Interrupt Counter R(1) shown in Table VI displays a two-page segment of memory in a 64 x 64 format. This display area is from 0400 to 05FF.

Table VII, Main Counter R(F), is a sequence of calls to the various subroutines. Since the Weapons subroutine cannot simultaneously fire both ship's weapons, the left and right fire subroutines alternate. Once the program is running, the alternation gives each player roughly an even chance of firing first if both fire switches are operated simultaneously. The main program inserts a delay routine between each call to R(C) and R(5) to slow the ship's movements for easier control.

TABLE I—INPUT AND MOVEMENT COUNTER R(C)

```
0200 DF
0201 F8 06 B6 A6 E6
0206 6C FA F0 AE
020A FA 20 32 0F 16 16
0210 8E F6 F6 F6 F6 56
0216 F8 06 A6
0219 72 FA 0F AD
021D FA 02 32 22 16
0222 8D 56 F8 06 A6 8D 56
0229 8E F4 56 64
022D F0 32 55
0230 F0 FB 03 32 69
0235 F0 FB 0C 32 7D
023A F0 FB 0F 32 91
023F F0 FB 04 32 A5
0244 F0 FB 07 32 B9
0249 F0 FB 08 32 CD
024E F0 FB 0B 32 E1
0253 30 F3
0255 97 FB 05 3A 5F
025A 87 FC 58 33 F3
025F 87 FC 10 A7
0263 97 7C 00 B7
0267 30 F3
0269 98 FB 05 3A 73
026E 88 FC 51 33 F3
0273 88 FC 10 A8
0277 98 7C 00 B8 30 F3
027D 97 FB 05 32 87
0282 97 FB 03 32 F3
0287 87 FF 10 A7
028B 97 7F 00 B7 30 F3
0291 98 FB 05 32 9B
0296 98 FB 03 32 F3
029B 88 FF 10 A8
029F 98 7F 00 B8 30 F3
02A5 97 FB 05 3A AF
02AA 87 FC 50 33 F3
02AF 87 FC 08 A7
02B3 97 7C 00 B7 30 F3
02B9 98 FB 05 3A C3
02BE 88 FC 49 33 F3
02C3 88 FC 08 A8
02C7 98 7C 00 B8 30 F3
02CD 97 FB 05 32 D7
02D2 97 FB 03 32 F3
02D7 87 FF 08 A7
02DB 97 7F 00 B7 30 F3
02E1 98 FB 05 32 EB
02E6 98 FB 03 32 F3
02EB 88 FF 08 A8
02EF 98 7F 00 B8
02F3 86 FB 07 3A FB
02F8 16 30 2D
02FB E2 30 00
```

TABLE II—LOAD COUNTER R(5)

```
014F DF
0150 83 A6 93 B6
0154 F8 00 AE
0157 97 73 87 52
015B 46 57 1E
015E 8E FB 0A
0161 32 6D
0163 87 FC 08 A7
0167 97 7C 00 B7
016B 30 5B
016D 72 A7 F0 B7
0171 84 A6 94 B6
0175 F8 00 AE
0178 98 73 88 52
017C 46 58 1E
017F 8E FB 0A
0182 32 8E
0184 88 FC 08 A8
0188 98 7C 00 B8
018C 30 7C
018E 72 A8 F0 B8
0192 30 4F
```

TABLE III—LEFT FIRE COUNTER R(A)

```
0194 DF
0195 3F 94 F8 01 F6
019A D9 30 F1 00 00
```

TABLE IV—RIGHT FIRE COUNTER R(B)

```
019F DF
01A0 3E 9F F8 00 F6
01A5 D9 30 F6 00 00
01F0 DF 3F 94 30 F0
01F5 DF 3E 9F 30 F5
```

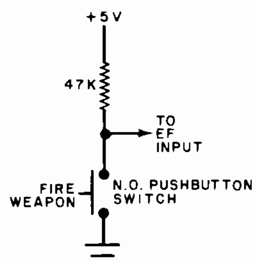

Fig. 1. To implement the space battle program, the circuit shown here must be duplicated, with one attached to $\overline{EF3}$ and the other connected to $\overline{EF4}$.

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TABLE V—WEAPONS COUNTER R(9)

```
0300 DF
0301 7A 7B
0303 33 07
0305 30 79
0307 87 FC 19 A6 97 7C 00 B6
030F 96 73 86 52 F8 00 AE E6
0317 F8 FF 56 16 1E
031C 8E FB 06 3A 17
0321 F8 FF F2 E2 3A 2C
0327 F8 FF 56 30 5C 22
032D 94 73 84 73
0331 F8 10 A4 F8 06 B4
0337 F8 00 AE 98 73 88 52
033E 44 58 1E 8E FB 0A
0344 32 50
0346 88 FC 08 A8 98 7C 00 B8
034E 30 3E 9E FC 10 BE
0354 72 A8 72 B8 72 A4 72 B4
035C 72 A6 F0 B6 F8 00 AD BD
0364 1D 9D FB 01 3A 64
036A F8 00 AE F8 00 56 16 1E
0372 8E FB 07 3A 6D 30 E8
0379 88 FC 27 A6 98 7C 00 B6
0381 96 73 86 52 F8 00 AE E6
0389 F8 FF 73 1E 8E FB 06
0390 3A 89
0392 F8 FF F2 E2 3A 9D
0398 F8 FF 56 30 CD 22
039E 93 73 83 73
03A2 F8 10 A3 F8 06 B3
03A8 F8 00 AE 97 73 87 52
03AF 43 57 1E 8E FB 0A 32 C1
03B7 87 FC 08 A7 97 7C 00 B7
03BF 30 AF 9E FC 01 BE
03C5 72 A7 72 B7
03C9 72 A3 72 B3
03CD 72 A6 F0 B6
03D1 F8 00 AD BD 1D
03D6 9D FB 01 3A D5
03DB F8 00 AE
03DE F8 00 56 26 1E
03E3 8E FB 07 3A DE
03E8 9E FA 88 32 00
03ED C0 01 C0
```

TABLE VI—INTERRUPT COUNTER R(1)

```
01CE 72 70
01D0 C4 22 78 22 52
01D5 F8 04 B0 F8 00 A0
01DB C4 C4 E2
01DE 80 E2 20 A0 E2
01E3 3C DE
01E5 80 E2 20 A0 E2
01EA 34 E5 30 CE
```

Note: *A cassette tape of the program for Space Battle is available from Donald R. Schroyer, 209 Brinker St., Latrobe, PA 15650, for $10.00.*

TABLE VII—MAIN COUNTER R(F)

```
0650 3F 50 37 52 D5 DC
0656 F8 00 AD BD 1D
065B 9D FB 02 3A 5A
0660 D5 DA DB DC
0664 F8 00 AD BD 1D
0669 9D FB 02 3A 68
066E D5 DB DA
0671 30 55
```

TABLE VIII—INITIALIZATION

```
0100 71 00
0102 F8 06 B2 BD BF
0107 F8 05 B7
010A F8 04 B8
010D F8 03 B9
0110 F8 02 BC
0113 F8 01 B1 B3 B4 B5 B6
011A BA BB A9 AC
011E F8 D0 A1 F8 3F A2
0124 F8 AA A3 F8 B4 A4
012A F8 50 A5 AF
012E F8 CC A6
0131 F8 C0 A7
0134 F8 07 A8 AD
0138 F8 95 AA F8 A0 AB
013E F8 00 BE
0141 F8 01 5D 1D
0145 F8 02 5D
0148 E6 69 E0
014B 70 2F
```

TABLE IX—DATA POINTERS

Left Ship
```
01AA 00 00 F2 F7 24
01AF 28 78 38 00 00
```
Alternate (see text)
```
01AA 00 00 18 FF 3C
01AF 18 3C FF 00 00
```
Right Ship
```
01B4 00 00 06 FF CE
01B9 86 0F 0F 00 00
```
Alternate (see text)
```
01B4 00 00 03 3F 7F
01B9 D4 3E 3E 00 00
```
Blast
```
0610 28 04 50 99 4E
0615 39 88 34 45 A2
```
End of Program
```
01C0 F8 06 A6 B6 E6
01C5 9E 56 64 26 30 C9
```
Display Area
```
0400 - 05FF; 00 loaded in display
             ; memory locations
```

In addition to the mentioned pointers, register R(6) serves as a temporary pointer to save and increment other pointers. The stack pointer is R(2) with the stack starting at 063F.

Memory location 0607 contains the current move instruction for the left ship, while 0608 contains the same data for the right ship. At Initialization (Table VIII), the contents of these two locations are 01 and 02, respectively. As stated earlier, R(E).1 holds the current score and is loaded with 00 on initialization. R(E).0 is an index counter used in loading the ship silhouettes and firing the weapons. R(D) is used to load 0607 and 0608 and for delays encountered in the main program.

Table IX contains the remainder of the program. Both the left and right ship are illustrated. Location 0610 to 0619 contains the blast that occurs after every hit and the debris of the destroyed ship. If the ship is moving fast enough at the time of a hit, debris is shown scattered in its wake.

Hardware. The only hardware connection required is a pair of circuits similar to that shown in Fig. 1, with one connected to $\overline{EF3}$ and the other to $\overline{EF4}$. These are the weapon firing switches.

Playing the Game. Load the program from Tables I through IX, making sure that the correct code is placed at the correct addresses. Since the 1802 starts running at address 0000, key in C0 01 00 starting at address 0000. Save the program on cassette if you have this provision.

To begin the game, reset and run the computer, then depress and release the INPUT switch on an ELF II, or bring the $\overline{EF4}$ line low then high. The two ships will appear in opposite diagonal corners.

The left ship cannot move until the right player has made a move. Once the right player has moved, the left ship will move according to the keypad data last entered. To fire the weapon, depress the appropriate FIRE WEAPON pushbutton switch. When the program is restarted after eight hits, each ship can be moved across the screen to remove "ghosts".

Program Changes. If you desire another look for the spaceships, try the alternate versions shown in Table IX. You can even design your own ships using graph paper. Make the first and last two bytes 00 or the ships will leave a trail as they move.

The winning score can be 1,2,4 or 8 as determined by address 03EA. To change ship speeds, locations 065D and 066A can be loaded with higher numbers to slow the ships down. If you desire more speed, use 01 or NOP's. ◊

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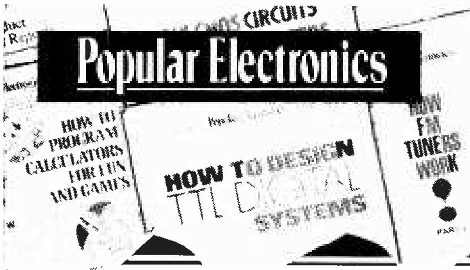

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BY RICHARD GOODPASTURE

How to Add TRIGGERED SWEEP TO AN OSCILLOSCOPE

Increase the performance capabilities of your scope by permitting expansion of waveforms.

WORKING with an oscilloscope that uses recurrent sweep can be frustrating when it comes to getting the sync locked in—and keeping it there. The situation is particularly touchy when one is trying to observe fast pulses that have low repetition rates. A much more practical approach to the problem is to use triggered sweep, where the sweep is synchronized by the actual signal that is being observed.

If you have a scope that does not have built-in triggered sweep, there is no need to trade up to a new scope. Instead, you can adapt it for triggered-sweep observation of waveforms, using the circuit shown in the schematic. This add-on circuit can convert virtually any recurrent-sweep scope into a modern triggered-sweep instrument.

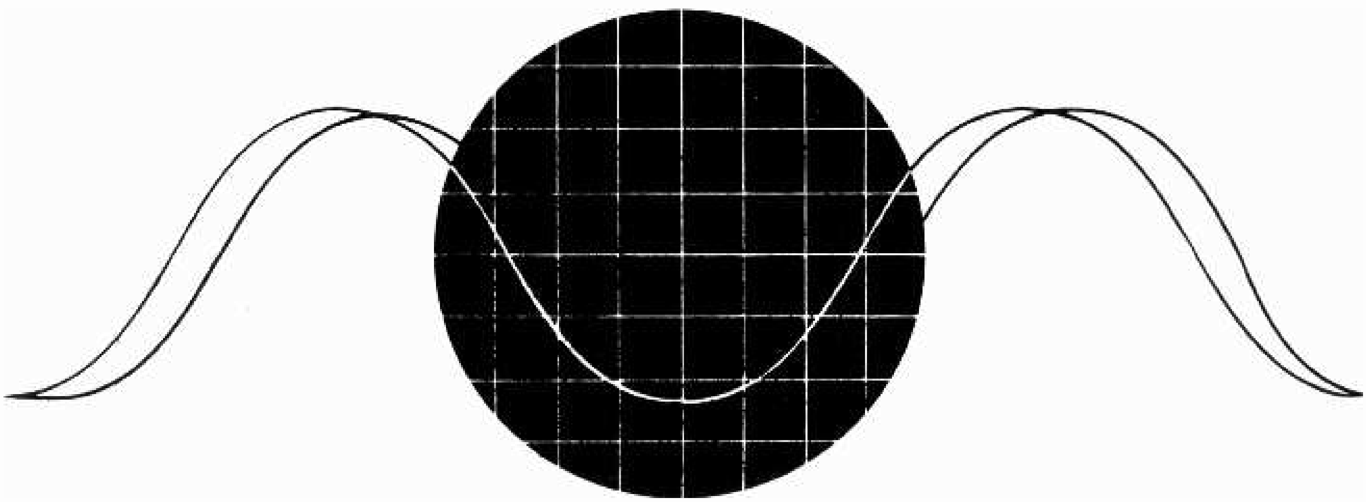
About the Circuit. Transistor *Q1* and resistors *R3*, *R4*, and *R5* make up a constant-current source for charging the sweep-range capacitor selected by switch *S1B*. Resistor *R1* determines the voltage to which the selected capacitor is to charge. The value of *R1* plus the charging current selected by *S1A* determine the sweep frequency. Resistor *R1* also determines the triggering sensitivity; and, with the 3300-ohm value specified, the sweep amplitude is 5 volts peak-to-peak. Omitting *R1* increases the sweep to 10 volts but decreases triggering sensitivity. (The value of *R1* can be changed without affecting the scope's sweep calibration.)

The unblanking pulse from pin 3 of *IC1* is coupled through an isolation capacitor with a typical value of 0.01 μF at 1.5 kV to the control grid of the CRT to intensify the trace during the sweep. Adjusting the scope's brightness control keeps beam intensity low while waiting for the next sweep.

The actual trigger signal can be taken from any point in the vertical amplifier channel where there is sufficient signal amplitude to trigger the sweep circuit.

The new triggered-sweep circuit is substituted for the existing sweep system that now drives the horizontal amplifier in your oscilloscope.

Construction. Just about any method of construction, from fabricating a printed circuit board to Wire Wrapping, can be used to assemble the circuit. Resistors *R8* and *R9* and capacitors *C5* through *C9* mount directly on switch *S1*.



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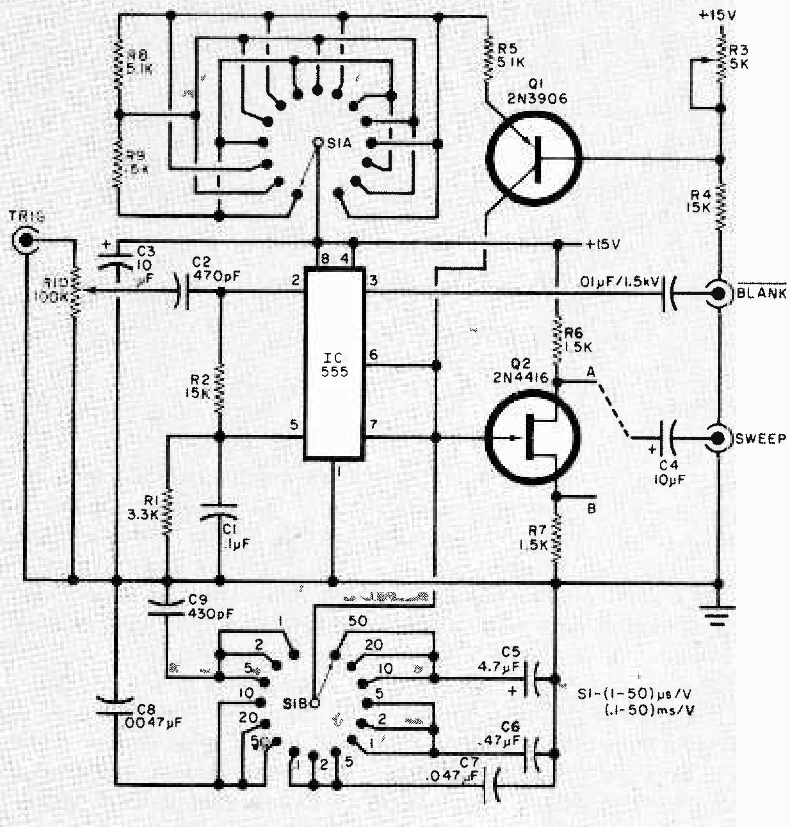

Constant-current source Q1 and timer IC1 combine to form a low-cost triggered sweep circuit.

PARTS LIST

C1—0.1-μF, 50-volt ceramic capacitor
C2—470-pF ceramic capacitor
C3,C4—10-μF, 25-volt electrolytic
C5—4.7-μF, 16-volt tantalum capacitor
C6—0.47-μF, 16-volt tantalum capacitor
C7—0.047-μF 50-volt polystyrene or Mylar capacitor
C8—0.0047-μF, 50-volt polystyrene or Mylar capacitor
C9—430-pF polystyrene, Mylar, or silver mica capacitor
IC1—555 or 1455 timer
Q1—2N3906 or similar pnp transistor
Q2—2N4416 or similar n-channel FET
The following are ¼-watt, 10% tolerance resistors:
R1—3300 ohms
R2,R4,R9—15,000 ohms
R5,R8—5100 ohms
R6,R7—1500 ohms
R3—5000-ohm trimmer potentiometer
R10—100,000-ohm potentiometer
S1—2-pole, 15-position nonshorting rotary switch
Misc.—Printed circuit board and socket (optional) for IC1; connectors (optional) for input and outputs; machine hardware; hookup wire; solder; etc.

The 15 volts required to drive the triggered-sweep circuit can be obtained from the power supply in the scope.

Calibration and Use. Potentiometer *R3* must be adjusted to provide a voltage drop of 2.35 volts across *R5* when *S1* is set to the 1-μs/V position and the collector of *Q1* is grounded. A more accurate setting for *R3* can be had by using a signal whose frequency is accurately known once the sweep has been coupled to the scope.

The best place to inject the new sweep signal is via the scope's EXT HORIZ input, after first determining which polarity of the sweep output produces a left-to-right deflection on the CRT. If a negative voltage produces the appropriate deflection, couple the sweep output to jumper point A. On the other hand, if your scope requires a positive voltage, connect to point B.

The TRIG input requires a negative-going pulse of about 2.5 volts to trigger the circuit. The width of the pulse can be as short in duration as 100 ns if a greater amplitude pulse is available. Use potentiometer *R10* to adjust the trigger amplitude to obtain a stable waveform. ◇

5000 ohms. (If you were making only one circuit, you could hand-select the components to make it work, but this is not a safe approach to use in a construction article.) Now note t_{WQ} when C_{ext} is 1000 pF and R_{ext} is 10,000 ohms. The width of the pulse can be between 2.76 and 3.37 μs. Hence, the value can range from +8.9% to −11.2% of the typical specified value for the given R and C values. Note also that the spec sheet does not tell you that this error is linear throughout the t_{WQ} range. For all we know, this may be the best point on the curve. So, when designing such a circuit, make certain that your design can accommodate this type of tolerance.

Note the column in Table II headed Nom (nominal). This value is the one for which you should strive, but you may find that it is not possible to obtain or hold it through the design.

It should be understood that one parameter may affect another. For example, consider the effect of varying the power supply voltage on the output sinking current (I_{OL}). The output sinking current is a linear function of the power supply voltage, as shown in Fig. 2. When the supply potential is 4.75 volts, the output can sink 15 mA. A similar condition can be observed in Fig. 3, where the maximum input forward current (I_F) is shown as a function of input voltage. Here again, the variation of one parameter can cause a variation in another.

All of the parameters thus far discussed have been of the type that can cause circuit failure, not failure of a component. Most IC data sheets carry a set of catastrophic characteristics, such as those listed in Table III. With resistors and capacitors, characteristics like maximum power dissipation and breakdown voltage should never be exceeded. Never come close to these specifications in your circuit designs.

Summing Up. If you use the techniques detailed in this article, or keep them in mind, your circuits will work and so will other circuits built from your design. If you build projects from magazines, steer clear of broad-tolerance components, especially in critical components. Do not be afraid to test semiconductors and passive components before using them. ◇

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Build THE SURFER

BY MAYNARD GRADEN

Psycho-acoustic device simulates sound of surf, gentle rain, or white noise to provide tranquil background for sleep, study or concentration.

IT IS WELL known that certain naturally occurring sounds can help us to relax, to study, and to concentrate. Among the more familiar of these are the sounds of the surf, rain, and white noise (as when you cup a seashell or a drinking glass to your ear). The "Surfer" is a simple electronic device that can generate all three of these pleasant, tranquil sounds at the flip of a switch. You can use the Surfer to set just the right mood for falling off to sleep, studying, or concentrating simply by setting the selector switch to the appropriate sound position.

About the Circuit. Transistor *Q1* in Fig. 1 serves as a white-noise source by virtue of its being operated in the reverse-bias breakdown mode. The "white-noise" signal is generated across *R2* and is amplified by *Q2*, whose high-frequency rolloff is determined by *C3* (and *C5* when *S1* is in the SURF position). With *S1* set to SURF, *Q3* has two inputs. One is the white-noise signal from *Q2* and the other is determined by the light intensity from *I1* falling on photocell *PC1*. As the intensity of light from *I1* varies, the bias on *Q3* is shifted which, in effect, amplitude modulates the white-noise signal. Trimmer potentiometer *R9* can be adjusted to prevent the white noise from being cut off.

When *S1* is set to RAIN, *PC1* is out of the *Q3* input circuit. This causes *Q3* to function as a conventional emitter-follower stage. The white noise at the emitter of *Q3* is passed through *R11* and VOLUME control *R12* for amplification by *IC1*. The circuit and its operation remain the same when *S1* is set to NOISE, except that, in this mode, limiting resistor *R11* is bypassed.

Photocell *PC2*, which is also illuminated by *I1*, and capacitors *C8*, *C9*, and *C10* form a nonlinear high-cut tone control circuit whose high-frequency attenuation is proportional to the light intensity that illuminates *PC2*.

The period of relaxation oscillator *Q6* is determined by the values of *C16* and the preset value of potentiometer *R20*. The interval controlled by *R20* can be adjusted between 7 and 35 seconds.

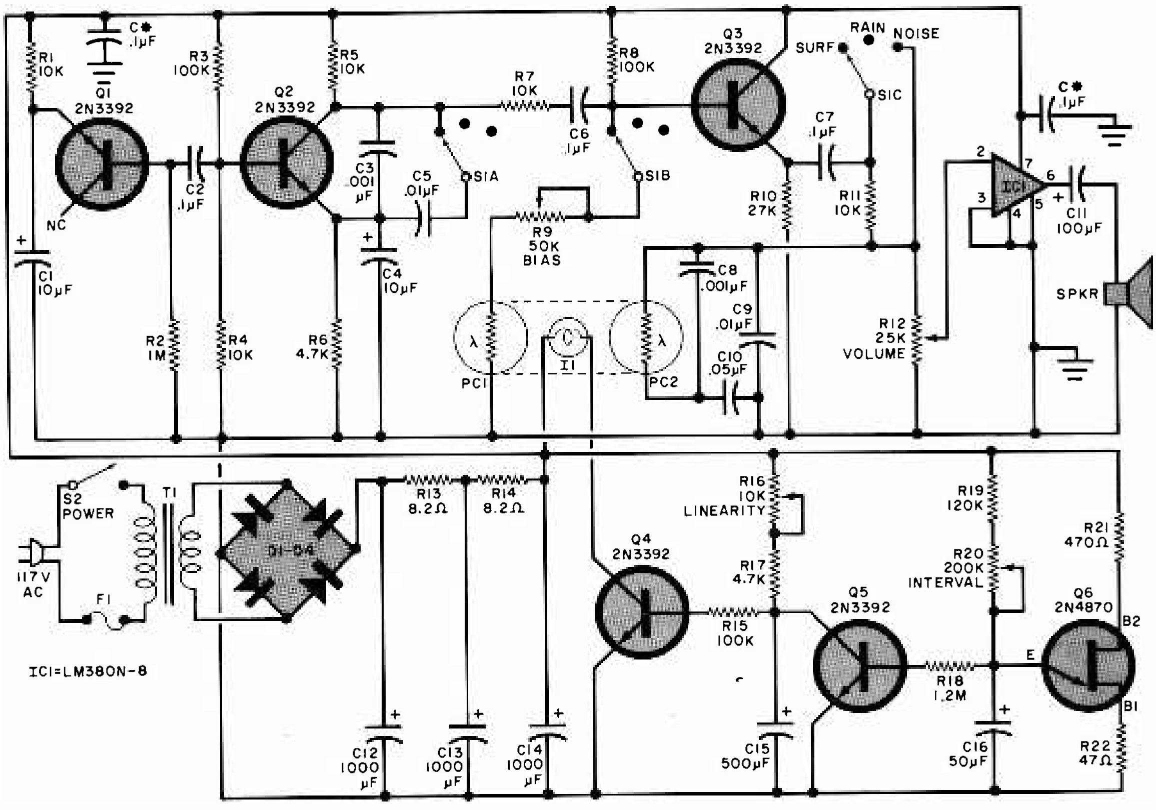

Fig. 1. White noise generated in Q1 is amplitude-modulated and filtered to produce different sounds.

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PARTS LIST FOR FIG. 1

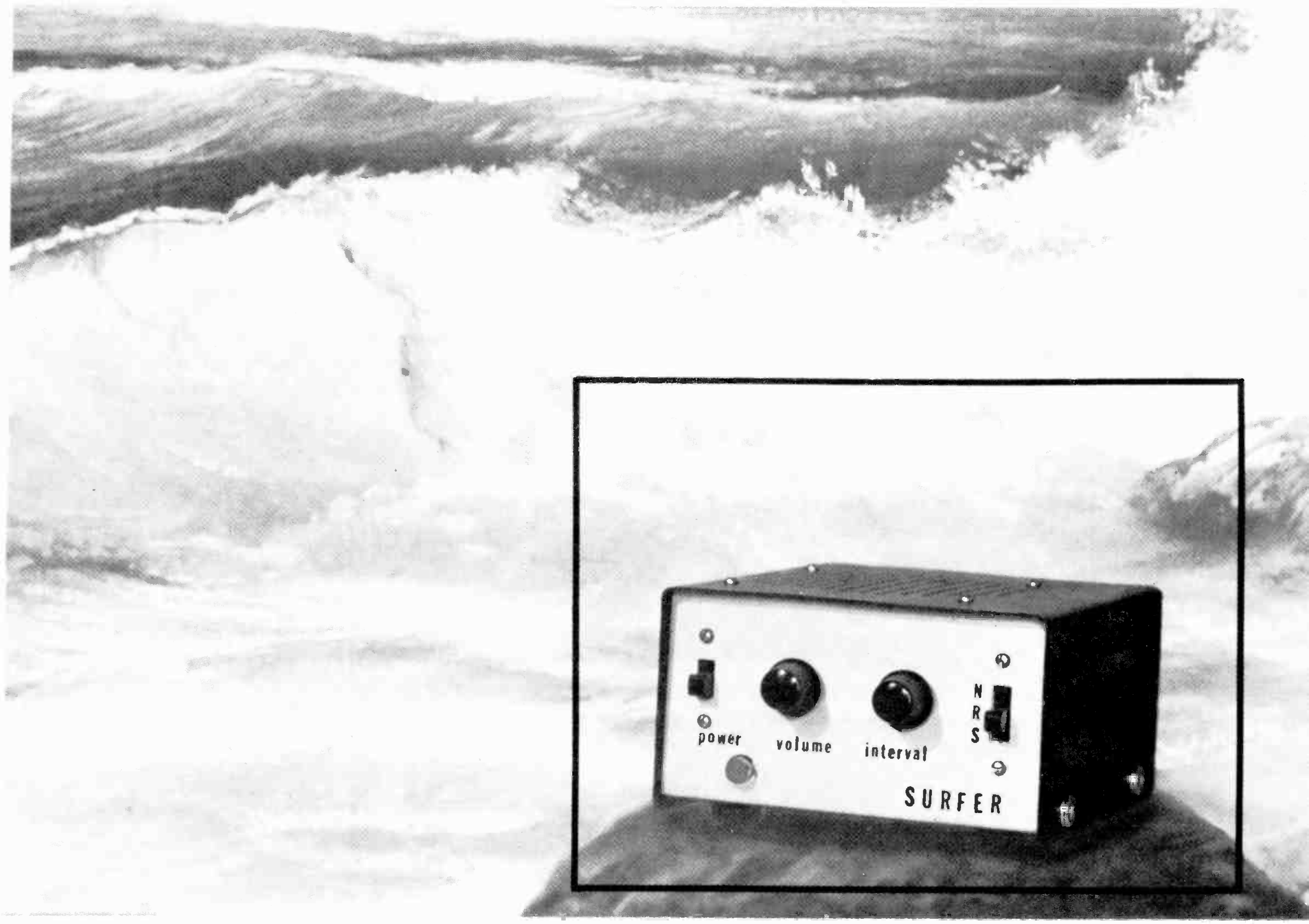
C1,C4—10-μF, 25-volt electrolytic
C2,C6,C7—0.1-μF disc capacitor
C3,C8—0.001-μF disc capacitor
C5,C9—0.01-μF disc capacitor
C10—0.05-μF disc capacitor
C11—100-μF, 25-volt electrolytic
C12,C13,C14—1000-μF, 25-V electrolytic
C15—500-μF, 25-V electrolytic
C16—50-μF, 25-V electrolytic
C*—0.1-μF capacitor (see text)
D1 through D4—1N4001 rectifier diode
F1—¼-ampere fuse and holder
I1—#1869 miniature lamp (10 V, 0.014 A, with wire leads)
IC1—LM380CN (National) or similar audio amplifier
PC1,PC2—Photoresistive cell with 5-megohm dark and 15,000-ohm light resistances (Clairex No. CL702L or Vactec No. VT322L)
Q1 through Q5—2N3392 transistor
Q6—2N4870 unijunction transistor
The following resistors are ¼-watt, 10%:
R1,R4,R5,R7,R11—10,000 ohms
R2—1 megohm
R3,R8,R15—100,000 ohms
R6,R17—4700 ohms
R10—27,000 ohms
R13,R14—8.2 ohms
R18—1.2 megohms
R19—120,000 ohms
R21—470 ohms
R22—47 ohms
R9—50,000-ohm trimmer potentiometer
R12—25,000-ohm audio-taper potentiometer
R16—10,000-ohm trimmer potentiometer
R20—200,000-ohm linear-taper potentiometer
S1—3-pole, 3-position nonshorting slide or rotary switch
S2—Spst switch
SPKR—8- or 16-ohm loudspeaker (see text)
T1—12.6-volt, 300-mA transformer
Misc.—Metal chassis box; line cord; materials for light shield (see text); rubber grommet; spacers; machine hardware; hookup wire; solder; etc.

The waveform present at the emitter of *Q6* is directly coupled to buffer/amplifier *Q5*, whose *C15/R16/R17* output circuit linearizes the signal before it is applied to lamp driver *Q4*.

Typical waveforms for the circuit are shown in Fig. 2. The upper waveform is present at the emitter of *Q6* when *R20* is set for a period of 17 seconds. The next waveform down illustrates the voltage at the collector of *Q4* after waveshaping by the *Q5* buffer stage. The inverse of this waveform is the voltage across *I1*. The third and fourth traces illustrate the outputs of the *Q1* white-noise source and *IC1*, respectively. Note that the output of *Q1* is a constant 10 mV, while the output of *IC1* varies between 100 and 400 mV. This amplitude variation is the result of the varying bias on *Q3* caused by the *I1/PC1* system.

Construction. There is nothing critical about circuit layout and you can use any wiring scheme you prefer. If you want printed-circuit board construction, you can make your own pc board using the etching-and-drilling guide shown in Fig. 3, which also contains component-installation instructions.

Lamp *I1* and photocells *PC1* and *PC2* must be mounted close together and shielded from outside light. Mount *I1* vertically with *PC1* and *PC2* on opposite sides of it so that they receive equal illumination. (Once the circuit is adjusted,



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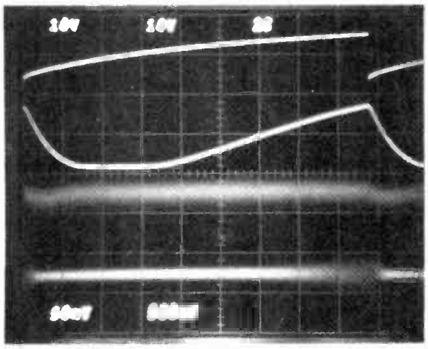
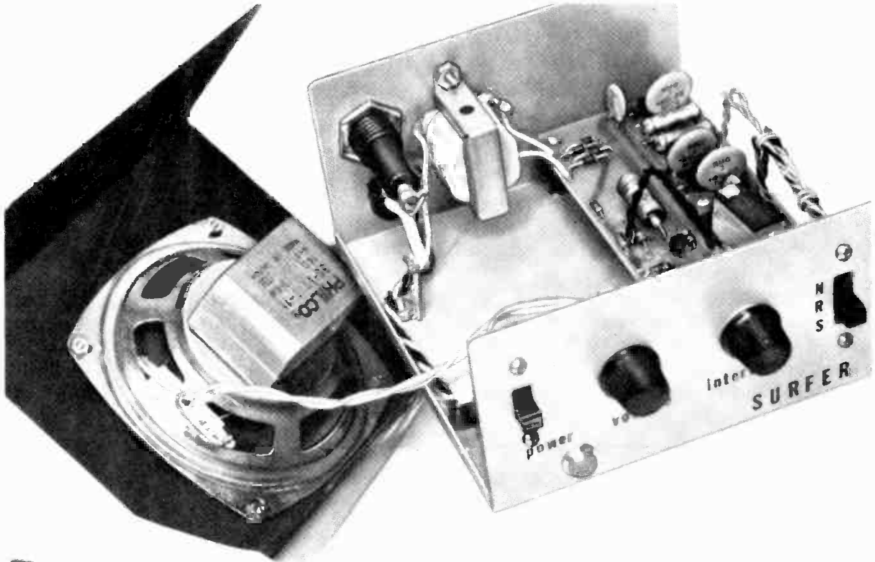
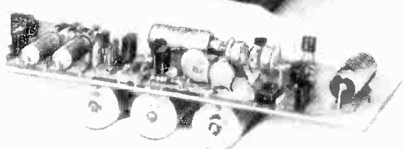
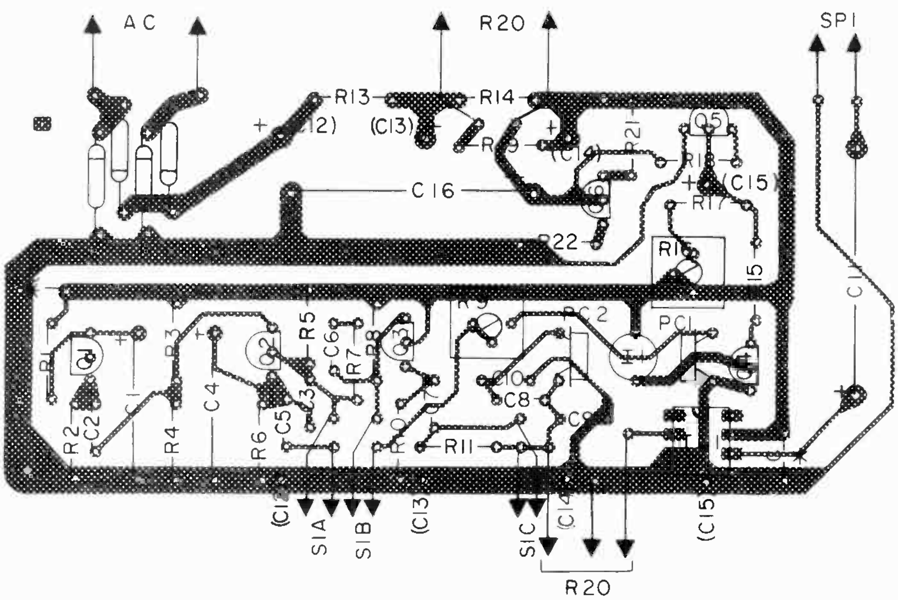


Fig. 2. Circuit waveforms show, at top, output of UJT; and, below that, collector of Q4 after waveshaping. Next two traces are white noise. One from Q1 at 10 mV and, at bottom, final audio output, with swing from 100 to 400 millivolts.





Printed circuit board is mounted on spacers in metal enclosure with T1, fuseholder and line cord on rear apron.



opaque tape or cardboard tubing can be used to make a light shield to exclude external light.

Note that there are two capacitors identified by asterisks (*). One of these decoupling capacitors must be mounted as close as possible to pin 7 of *IC1*, while the other should be mounted as close as possible to the positive end of *R1*. If you use the pc board pattern shown in Fig. 3, *C12*, *C13*, *C14*, and *C15* go on the *foil* side of the board and should be the last components installed.

The quality of the loudspeaker used in the circuit is a significant factor in performance. The maximum output power from *IC1* is about 0.5 watt into 8 ohms. Needless to say, do not use a 4-ohm speaker because it can damage *IC1*. A

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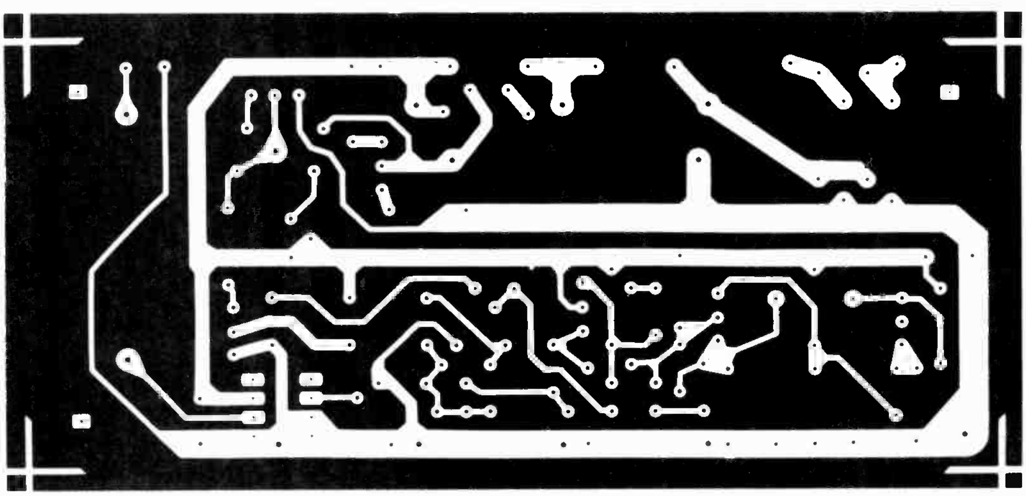

Fig. 3. Actual-size foil pattern is at left, with component placement above. Capacitors C12, C13, C14 and C15 are mounted on foil side.

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BUILD A Speaker Protection Circuit

BY MIKE ROGALSKI

AFTER LONG periods of listening to reproduced music played at a high volume level, it's not uncommon for one's hearing to become insensitive to average loud sounds. As a result, the listener often turns up the gain to compensate for this diminished sensory perception.

The best way to protect our hearing ability—and do a good turn for our speaker systems—is to put an upper limit on the decibel level our sound systems can generate. This is precisely what the automatic audio-overload/speaker-protection circuit described here does.

There are, of course, many circuits that use zener diodes and SCRs to shunt power to dummy loads. Most act too fast, however. This causes important dynamics such as drum rolls, cymbal crashes, and trumpet blasts to get "crunched." A slow-acting threshold sensor that has built-in hysteresis and a comparator circuit would be excellent for providing automatic level limiting, but it requires a power supply. The speaker-protection system here, on the other hand, is far simpler in circuitry, self-powered, automatic in action, and connects directly between the power amplifier and the speaker system it is to protect. It is also inexpensive to build.

About the Circuit. The output from the power amplifier to the speaker-protection circuit is shown in Fig. 1. (The rectifier diodes should have a forward resistance of approximately 600 ohms to introduce minimal signal distortion.) The signal then goes to the normally-closed relay contacts and out to the speaker system.

At high signal levels, the charging circuit consisting of *R1* and *C1* generates sufficient voltage levels to energize *K1* and open its contacts. When *K1* energizes, *R2*, is connected in series with the speaker system to drop the sound level. Then, when the input signal level drops, *K1* de-energizes and its contacts automatically close, removing *R2* from the circuit.

Construction. The simplicity of the protection circuit lends itself to just about any method of construction desired. For those who wish to use printed-circuit construction, an actual-size etching-and-drilling guide and components-placement diagram are given in Fig. 2. Once wired, this compact pc assembly can be permanently mounted inside the speaker system's enclosure or connected directly to the speaker terminals.

Relay *K1* should have a dc coil resistance of about 100 ohms and a dc pull-in rating of at least 2 volts *less than* the required rms voltage cutout point of the speaker system. This allows for the voltage drop across the rectifier circuit. The diodes and capacitors should have twice the peak voltage rating of the signal passing through them. The components specified in Fig. 1 are for a 4- and an 8-watt unit and will protect a speaker system rated at 5 to 10 watts with a 20% derating factor for safety.

Resistor *R1* can be bypassed to move the operating point of *K1* down to 4 watts.

Adjustment. Make certain that the common of each amplifier output circuit is connected to the common of the speaker protector and observe proper speaker phasing when connecting the device into your audio system. With a relay whose coil resistance is about 100 ohms, the circuit shown in Fig. 1 will cut out at 4, 8, or 12 watts if the value of *R1* is 0, 50, or 100 ohms, respectively. Since the circuit is basically a voltage divider, doubling the value of *R1*, shifts the rms point 50% higher. You can also experiment with the value of *R2* to obtain the low level desired. ◇

PARTS LIST

C1—100-µF, 50-volt electrolytic
D1 thru D4--Silicon rectifier diode (see text)
K1—Spst relay with 100-ohm dc-resistance coil (American Zetler No. A 535-11-2 or similar) (see text)
R1—Value depends on power protection level: 0 ohms for 4 watts; 50 ohms for 8 watts; 100 ohms for 12 watts
R2—50-ohm, ½-watt resistor

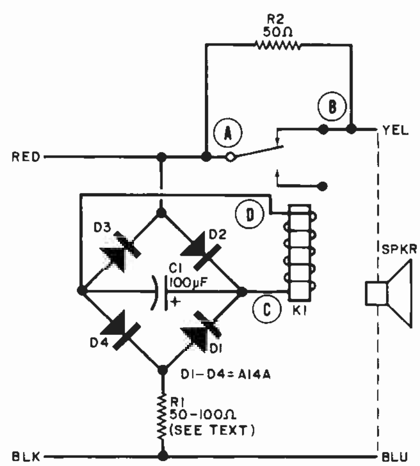
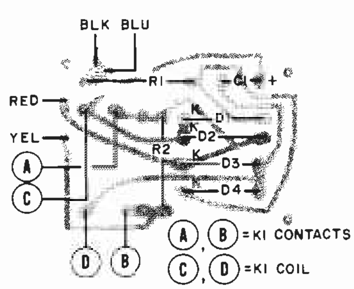


Fig. 1. The self-powered circuit, left, automatically reduces speaker level when peaks occur.



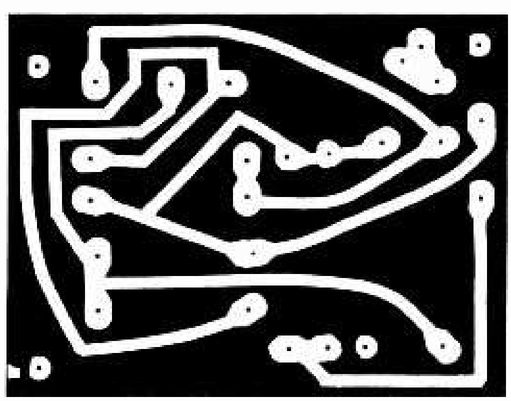


Fig. 2. Actual-size etching and drilling guide, right; component placement above.

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Designing Circuits for Worst-Case Operation

BY STEVEN L. CHEAIRS

How to choose components with tolerances to insure that circuits work properly.

ONE CONSTANTLY recurring problem for many hobbyists is that some circuits in the projects they build fail to work properly. Other than improper assembly and bad components, the most probable cause of this problem is that a "typical" circuit design was used. A typical circuit design might be sound on paper, but unless component characteristic variations are taken into account, the design may not produce a working circuit. And the cause is *normal* component parameter variations. It is important, therefore, that when you design or build a project, you take into account the possible variation range of the components you will be using to ensure that the project works properly.

In this article, we will discuss why component characteristics vary and what can be done to circumvent possible problems. Stated differently, we will discuss how to design for worst-case conditions.

Why They Vary. Component characteristics can vary for any number of reasons. For example, IC's are manufactured in "batch" lots, wherein a number of identical chips are fabricated simultaneously on a single silicon wafer. This approach results in significant manufacturing savings and a very low cost per circuit element. Unfortunately, the parameters of the individual components can vary greatly from one wafer to the next, even though component characteristics on a single wafer will "track" very closely.

It is not uncommon to find a circuit that contains components whose parameters fall anywhere between their worst-case limits. If the circuit was designed around devices that have typical parameters, there is the possibility that it will not function because it contains a device that operates at an extreme end of its parameters. Here is an example.

Assume a circuit has 50 components. Of these, 80% have typical parameters and 10% are sensitive to parameter variations. That means 20%, or 10 components, have atypical characteristics and 5 components are parameter sensitive.

The probability of an event occurring can be defined by the equation P = M/N, where P is the probability, M is the number of times the event is expected to occur, and N is the number of trials. Hence, the probability of a sensitive component occurring per circuit is 1/10, while the probability of a component having atypical performance is 1/5.

By the Law of Multiplication Probability (compound probability), when an event is regarded as occurring if a number of subevents independently occurs, the compound probability of occurrence of the event is equal to the product of the individual probabilities of the subevents. This can be expressed in a mathematical way by the equation P = P1 × P2 or P = (M1/N1) (M2/N2). Therefore, 1/5 times 1/10 or one out of every 50 circuits may not function due to the typical design technique used in our example.

In all likelihood, the figures used in the example are applicable to many hobbyist-built projects and account for the occasional project that fails to work even when all the wiring is correct. It should also be noted that this condition is worsened when "surplus" components are used, since the probability of using a component that is just barely within its specifications increases. Using such components, it becomes possible for a designer to produce a working design prototype that when duplicated by others will fail to operate.

Given the above conditions, it becomes mandatory for all circuit designs to be subjected to worst-case design analysis if the circuit is to be duplicated by others. Any circuit can be so analyzed. The most convenient method is to use worst-case parameter values during the initial design phase to insure proper operation from the start.

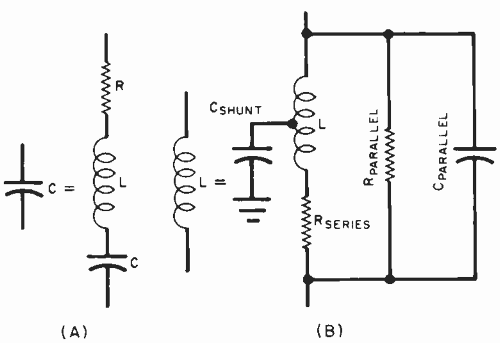

Fig. 1. A theoretical capacitor (A) actually contains parasitic R and L. Theoretical equivalent of an inductor (B) is even more complex.

Defining the Problem. It is essential to recognize which parameter or combination of parameters create the worst case for a particular circuit. Unfortunately, these conditions and how they affect circuit performance vary from circuit to circuit. Also, there may be different performance specs for any given circuit.

Most modern circuits contain both IC's and discrete components. When an IC and a number of discrete components can be combined to make a subcircuit, it is acceptable and may even be desirable to consider the subcircuit thus formed as a self-contained entity. This is

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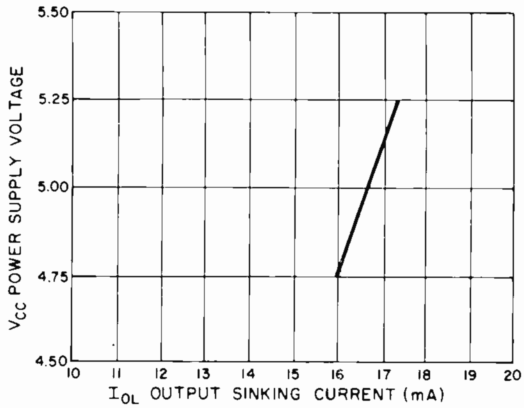

Fig. 2. Maximum sinking current as a function of power supply voltage variation.

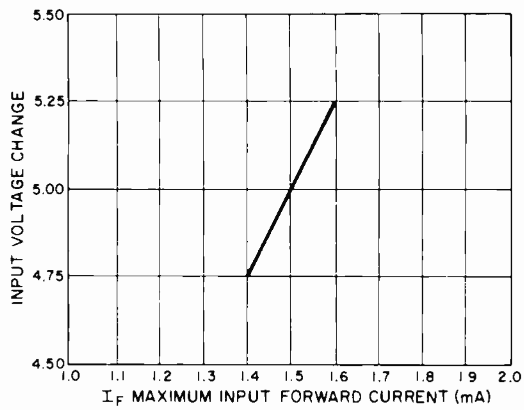

Fig. 3. Maximum input forward current as a function of the input voltage.

also true for combining gates and other elements of IC's. When circuit elements are so combined, a block diagram is created. The self-contained entities can then be individually analyzed and the results combined to analyze total circuit performance. This approach also allows system partitioning and interconnection methods to be considered, as well as such problems as impedance matching, level shifting, and fan-out.

The entire circuit's specifications can be divided down to the individual blocks that are sufficiently detailed to be treated on a stand-alone basis. All characteristics must be considered. If the circuit block does not satisfy the detailed requirements (input and output impedance, temperature range, threshold levels, propagation delay, hold times, etc.), the circuit must be modified.

Every component in a circuit must be allowed to vary over its full range of values, as specified by its tolerance, and still allow satisfactory circuit operation. It is the tolerance range that specifies the worst-case parameter range.

Every component contains parasitic components, such as capacitance, inductance, and resistance. In many circuits, the parasitic components are observed only during worst-case conditions. For example, consider a capacitor. A capacitor cannot simply be added to a high-frequency circuit with the expectation that the circuit will behave as if a theoretically pure capacitor were added. This simple component is actually quite complex, as can be seen in Fig. 1A. An inductor is even more complex, as shown in Fig. 1B. Therefore, for proper worst-case operation, these parasitic effects must be considered when designing and building circuits.

Fixed resistors also have broad tolerance specifications that can range up to ±10% (±20% in older resistors) of their specified nominal values.

The Spec Sheet. Manufacturer specification sheets for a particular IC should be consulted for pinout and to gain a working knowledge of the device itself. A typical spec sheet, this one for a 74123 dual retriggerable monostable multivibrator IC, is shown in Table I.

Assume you require a 50-ns pulse and decide to use the 74123 to generate it. Note in the table that $t_{wQ(min)}$ (minimum output pulse width) has a typical value of 45 ns and a worst-case value of 65 ns when external capacitance C_{ext} is zero and external resistance R_{ext} is

(Continued on page 126)

TABLE I—SWITCHING CHARACTERISTICS V_{CC}=5 V, T_A=25°C

Parameter*	From input	To input	Test	Min.	Typ.	Max.	Units
t_{PLH}	A	Q	C_{ext}=0		22	23	ns
	B		R_{ext}=5k		19	28	ns
t_{PHL}	A	Q	C_L=15 pF		30	40	ns
	B		R_L=400		27	36	ns
t_{PHL}	Clear	Q			18	27	ns
t_{PLH}		Q			30	40	ns
$t_{wQ(min)}$	A or B	Q			45	65	ns
t_{wQ}	A or B	Q	C_{ext}= 1000 pF, R_{ext}=10k, C_L=15 pF, R_L=400	2.76	3.03	3.37	µs

*t_{PLH}= propagation delay time, low- to high-level output
t_{PHL}= propagation delay time, high- to low-level output
t_{wQ}= width of pulse at output Q

TABLE II—RECOMMENDED OPERATING CONDITIONS

Parameter	Min.	Nom.	Max.	Units
Supply voltage, V_{CC}	4.75	5	5.25	V
High-level output current, I_{OH}			−800	µA
Low-level output current, I_{OL}			16	mA
Operating free-air temp., T_A	0		70	°C

TABLE III—ABSOLUTE MAXIMUM RATING OVER FREE-AIR TEMPERATURE RATING

Supply voltage, V_{CC}*	7 V
Input voltage	5.5 V
Operating free-air temperature range	0-70°C
Storage temperature range	−65 to +150°C

*Voltage values are with respect to network ground terminal.

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BY PAUL LUTUS

BUILD THE "SUPER MARKER"

Inexpensive marker generator with selectable 100-, 50-, 20-, or 10-kHz output allows precise tuning of shortwave receivers.

A STABLE source of marker frequencies is one accessory that belongs in every shortwave listener's shack. Many receivers contain built-in 100-kHz calibrators, but for exact tuning, smaller marker increments are required. The Shortwave Super Marker described in this article is an inexpensive, easily built frequency standard that will provide precise markers at selectable increments of 100, 50, 20, or 10 kHz. Built around a quartz crystal, two npn transistors, and a CMOS divider IC, the project can be assembled in two hours or less. Total parts cost is about $15.

About the Circuit. Transistor *Q1*, the quartz *XTAL* and their associated components comprise a stable 100-kHz oscillator. Trimmer capacitor *C1* allows the user to zero-beat the oscillator against a frequency source of known accuracy such as radio station WWV or WWVH. The 100-kHz output of the oscillator is applied to pin 14 of *IC1*, the CLOCK input of a CD4017 CMOS decade counter/divider with ten decimal outputs.

Depending on the position of *S1*, the RESET terminal of *IC1* is either grounded or connected to one of three decoded decimal outputs. When the RESET terminal (pin 15) is grounded, *IC1* functions as a ÷10 counter and a 10-kHz pulse train appears at pin 2. If pin 15 is connected to pin 1, the counter resets itself every 5 clock pulses and a 20-kHz pulse train is developed. Connecting pin 15 to pin 4 causes the counter to reset after every second clock pulse. The counter than acts as a ÷2 stage and produces a 50-kHz output. When pin 15 is connected to pin 2, the counter resets itself on the negative edge of each clock pulse, acting like a ÷1 stage and producing a 100-kHz pulse train at pin 2.

Transistor *Q2* and its associated components comprise an amplifier which is driven by the programmable counter's output pulses. This stage amplifies the harmonics of the fundamental pulse train frequency so that they are of usable strength up to 30 MHz. Accordingly, if *S1* is placed in the 100-kHz position, the user will hear marker signals every 100 kHz as he tunes across the dial of his general-coverage receiver. Successively higher-order harmonics will be increasingly weaker, but usable markers will be found to at least 30 MHz, the upper limit of most receivers.

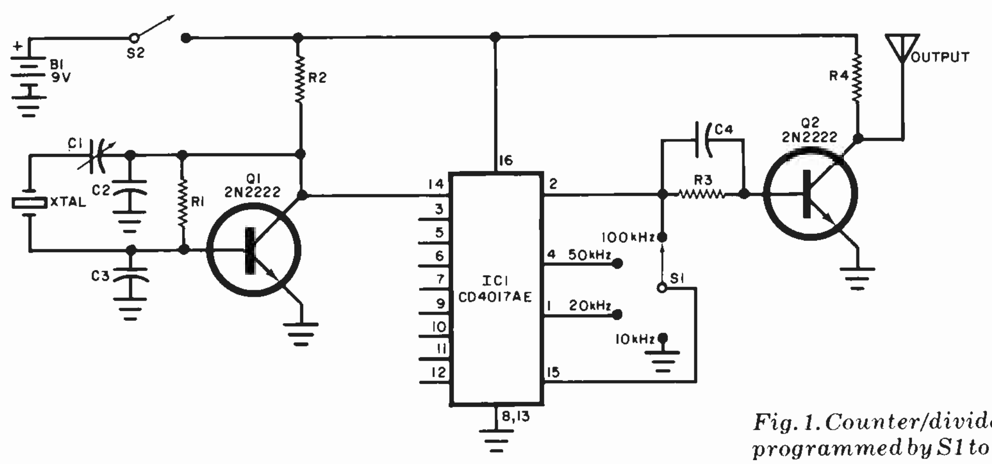

Fig. 1. Counter/divider IC1 can be programmed by S1 to provide marker frequencies at selected intervals.

PARTS LIST

B1—9-volt transistor battery
C1—7-to-45-pF trimmer
C2,C3—0.001-μF disc ceramic
C4—50-pF disc ceramic
IC1—CD4017AE CMOS decade counter/divider with decoded decimal outputs
Q1,Q2—2N2222 npn silicon transistor

The following are ¼-watt, 10% tolerance fixed carbon-composition resistors:
R1—150,000 ohms
R2,R3—8200 ohms
R4—5600 ohms
S1—1-pole, 4-position nonshorting rotary switch

S2—Spst toggle switch
XTAL—100-kHz quartz crystal
Misc.—Molex Soldercons or IC socket, printed circuit or perforated board, suitable enclosure, battery holder and clip, hookup wire, machine hardware, circuit board standoffs, solder, etc.

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Construction. Parts placement is not critical, so printed circuit or point-to-point perforated board techniques can be employed. The use of an IC socket or Molex Soldercons is recommended for mounting the CMOS device. Carefully observe the standard precautions when handling the CMOS device and pay attention to the pin basing of both the IC and transistors. Any four-position rotary switch can be used for *S1*. If you already have a switch with more than four positions, you can use it in the circuit if the extra positions are grounded.

The output antenna shown in the schematic is simply a length of hookup wire that can either be wrapped around the antenna lead-in (if a single wire feed is used) or physically placed close to the r-f input stage. No direct connection between the Super Marker and receiver is required. The project can be housed in any small enclosure or even mounted inside the receiver if space is available. For simplicity, a 9-volt battery is used as the power source. However, a small well-filtered, line-operated supply can be used instead. A third alternative is to tap the receiver's dc supply or, if the project is to be used with an older tube-type receiver, the ac filament voltage can be rectified, filtered and zener regulated.

Calibration. Tune your receiver to WWV or WWVH at 2.5, 5, 10 or 15 MHz. With the Super Marker's antenna coupled to the input of the receiver and switch *S1* in the 100-kHz position, close power switch *S2*. You should hear both the NBS transmission and an audio tone whose pitch will vary as trimmer capacitor *C1* is adjusted. If you don't hear the audio tone, increase the coupling between the Super Marker's antenna and the receiver input.

Carefully adjust *C1* so that the audio tone decreases in pitch and becomes a "flutter" on the NBS transmission. Ideally, *C1* should be set for a zero beat. That is, the marker and r-f carrier are at exactly the same frequency and no beat note is created. Adjust the trimmer capacitor during the portions of the WWV or WWVH transmission when only second ticks and no continuous audio tone superimposed on the ticks are heard. Otherwise, you may zero beat the marker to the modulating tone instead of the r-f carrier.

A nonmetallic screwdriver, alignment, or neutralization tool should be used when making these adjustments. Even so, you might find that the presence of the tool and/or your hand will affect the oscillator's frequency. Withdraw the tool and your hand between adjustments to ensure that a true zero beat has been obtained.

It's a good idea to drill a hole in the project enclosure so that *C1* can be adjusted after the enclosure has been "buttoned up." This will minimize the detuning effects of hand, tool and even enclosure capacitance. Also, this hole will enable you to periodically touch up the adjustment of *C1* without having to remove the top of the enclosure.

Use. Turn on the receiver's bfo and tune up the band until another marker is encountered. (Don't confuse a broadcaster's carrier with a marker. Open and close power switch *S2*. The marker tone should appear and disappear as power is applied to and removed from the circuit.) Note the frequency indicated on the dial and tune back to WWV or WWVH. Next, place *S1* in the 50-kHz position and tune up the band until you encounter a marker. The dial frequency should be midway between that of WWV or WWVH and the previously noted marker frequency. With *S1* in the 20-kHz position, you should detect five markers—one at the NBS station's frequency, one at the previously noted marker frequency, and three spaced evenly between the two. In the 10-kHz mode, the Super Marker should generate ten evenly spaced markers across this 100-kHz band segment.

The Super Marker will allow you to tune your receiver very precisely even if its tuning mechanism and dial are less than optimum. Let's assume that a weak DX station you've been chasing is listed as transmitting on 15.370 MHz. Tune your receiver to WWV or WWVH and turn on the Super Marker. Place *S1* in the 100-kHz position, turn on the receiver's bfo and tune up three markers to 15.300 MHz. Place *S1* in the 50-kHz position and tune up one marker to 15.350 MHz. Next, place *S1* in the 10-kHz position and tune up two markers and open *S2*. Your receiver is now tuned to exactly 15.370 MHz. If propagation conditions are favorable, you'll hear the station with no need for further tuning.

If the desired station is transmitting at, say, 15.380 MHz, the procedure is less complicated. After tuning to 15.300 MHz, place *S1* in the 20-kHz position and tune up four markers to exactly 15.380 MHz. You can develop your own tuning procedures after you have logged some time practicing with the Super Marker. ◇

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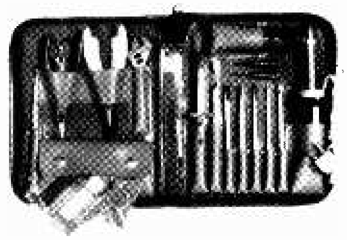

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CIRCLE NO. 5 ON FREE INFORMATION CARD

BY JOSEPH J. CARR

Practical, low-cost references for checking accuracy of test instruments

Precision References for Current & Voltage

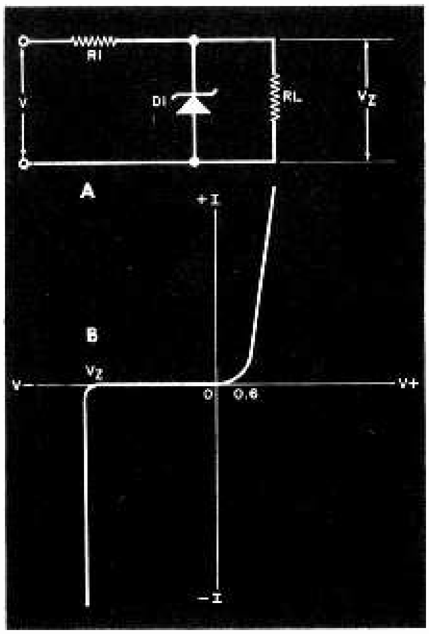

Fig. 1. Zener-diode reference supply. The circuit is shown at A, while the characteristic curve is below at B. The fact that the zener voltage is dependent on temperature must be kept in mind in circuit design.

PRECISION voltage and current references are routinely used in a variety of applications. These references are useful for calibrating vertical deflection factors of oscilloscopes and indications of voltmeters and ammeters, for example. They also find use in data conversion, where almost all analog-to-digital (A/D) and digital-to-analog (D/A) converters employ a reference voltage or current. Regulated dc power supplies that employ error amplifiers also require precision voltage references. (The output of an error amplifier controls regulator action, which is, in turn, controlled by the difference between actual output-supply and reference voltages.)

Precision references were quite costly only a few years ago. Now, owing to low-cost ICs, good precision references can be built at low cost.

Before we examine IC voltage and current references in detail, let us review some of the older methods for obtaining references so that principles will be more readily understood.

Zener Diode. A simple circuit in which a zener diode is the regulator element is shown in Fig. 1. Also shown is a typical zener-diode characteristic curve. In the V+ forward-bias region, the zener diode behaves much like any other silicon diode that conducts a +I forward-current when V+ is greater than about 0.6 volt. In the V− reverse-bias direction, however, there is a distinct difference between zener and conventional silicon-diode behavior.

Normally, a conventional silicon diode does not conduct current when reverse biased (except when applied reverse-bias voltage exceeds the diode's rated PIV). A zener diode acts quite conventionally between 0 and some V− value called the "zener potential," or V_Z. When V− reaches or exceeds V_Z, the diode breaks over and begins to conduct a reverse current.

As long as ambient temperature is held constant, V_Z will also be constant. Bear in mind that the V_Z value is different for different types of zener diodes and that even then there is a "nominal" voltage. This means that a large number of identical zener diodes will have values that cluster closely around V_Z.

When you build a precision power supply or reference source, you must keep either of two considerations in mind. You must provide a constant-temperature environment or use temperature-compensated circuitry. Then either hand-select the zener diode or provide a means for adjusting the output of the circuit so you can compensate for incorrect zener potentials.

Unfortunately, temperature cannot always be maintained constant in practical circuits, especially where cost is a factor. An attempt at solving temperature dependence by using several zener diodes to produce the desired V_O output voltage is shown in Fig. 2. The actual V_Z value for the different diodes will vary with changes in temperature, but if all diodes used are in the same thermal environment, any temperature change affects all diodes equally. The output voltage, which is the differential voltage between the two points shown, remains constant regardless of temperature changes. Output potential $V_O = (V5 + V6) - V3$.

A problem with circuits such as that shown in Fig. 2 is that the output voltage is not ground referenced. If a ground-referenced potential is required, V_O should be applied across the differential inputs of an operational amplifier. The output potential from the op amp would then be the product of op-amp gain A_V

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and output voltage V_O, or $V_O \times A_V$.

Op amps are frequently used to buffer zener-diode regulators and allow more precise setting of actual output voltage. The circuit usually used for reference sources is shown in Fig. 3. You should recognize this as a special case of a noninverting gain-follower circuit in which zener diode *D1* is used to set the potential applied to the noninverting input. Voltage gain for this circuit is $1 + R_a/R1$. Therefore, $V_O = V_Z(1 + R_a/R1)$. Feedback resistor R_a is actually fixed resistor *R2* and potentiometer *R3*. If *R3* is a 10-turn (or more) trimmer pot and has a total resistance of only 10% to 20% of R_A, V_O can be set precisely.

The circuit in Fig. 3 is used for moderate-precision applications. It still requires a constant-temperature environment to maintain V_O stability. Both zener diode and op amp may tend to drift with changes in temperature. Several commercial voltage-reference standards are available in which circuits such as that shown in Fig. 3 are installed in proportional control temperature ovens.

Precision IC References. IC voltage references allow you to build simple voltage and current references that perform as well as all but the best discrete references previously available.

National's LM199 (and companion LM299 and LM399), shown in Fig. 4, contains a zener diode, whose V_Z is nominally 6.95 volts, inside an IC that also has a built-in conductor heater. (The zener diode is buried in the same semiconductor die as the heater circuitry. This yields lower-noise operation and provides thermal stability.) Ordinarily, zener potentials can change as much as 5 mV/°C, but in temperature-controlled LM199s, drift is limited to microvolts.

Although the LM199 is rated to have an initial stability of ±2% of rated voltage, its stability is very good. Long-term stability is rated at 20 ppm and short-term, as low as 1 ppm. Use of the thermal insulator cap supplied with the LM199 is highly recommended.

In normal operation, the LM199 is used in the same manner as any other zener diode, provided the heater terminals are connected across a 9-to-40-volt dc power supply. Pins 2 and 4 are usually grounded to keep the internal substrate reverse biased. Pin 3 connects to the dc power supply, while zener cathode pin 1 connects into the circuit. Circuits like that shown in Fig. 3 are often used with the LM199.

Another precision device that can be

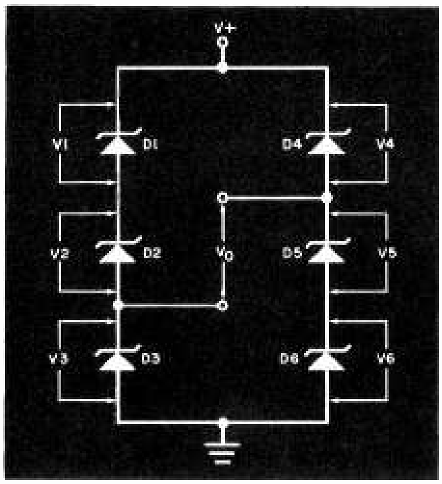

Fig. 2. Several zener diodes may be used in an attempt to stabilize variations in circuit operation due to changes in temperature.

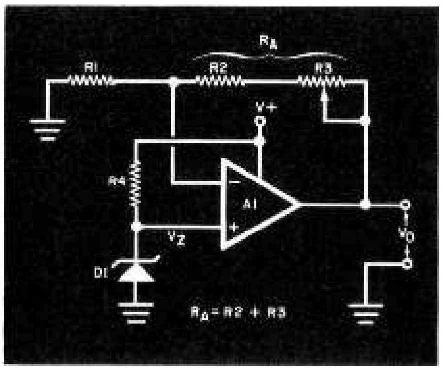

Fig. 3. By adjusting the gain of the op amp (by varying R3), the output voltage can be trimmed to a precise value.

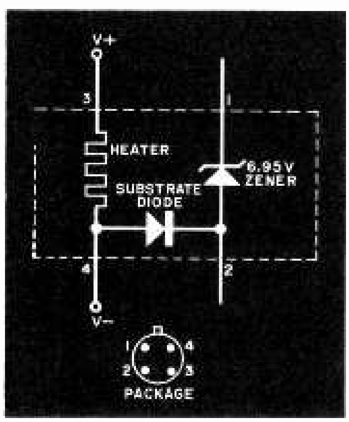

Fig. 4. The LM199 temperature-stabilized zener diode uses an on-chip heater.

treated as if it were a zener diode is the band-gap regulator from Ferranti Electric, Inc. (Semiconductor Products Div., 87 Modular Ave., Commack, NY 11725). Versions in either 2.45 or 1.26 volts are available. Construction of these regulators is shown in Fig. 5. The internal circuit is basically a current-boosted op amp with a band-gap reference diode at the noninverting input. These are two-terminal devices that can be used as if they were ordinary zener diodes.

The Ferranti ZN404 and ZN458A/B are 2.45-volt regulators. They differ as to initial calibration tolerance and thermal drift. While all are better than ordinary zener diodes, the ZN458B is the best in the line. (The ZN423 is a 1.26-volt version of the ZN458B.) All three Ferranti devices sink up to 120 mA and have a 2-to-120-mA operating range.

Long-term stability of the Ferranti regulators can approach 10 ppm/1000 hours, while temperature coefficient is rated at 0.003%. Current-limiting resistor R's value is $[V+ - V_{REF}]/I_{REF}$ (R = resistance in kilohms; V+ = supply potential; V_{REF} = nominal rated reference potential of the device; and I_{REF} = device current).

As in all precision voltage reference supplies, the resistor should have a low temperature coefficient. In most cases, this means use of a wirewound or metal-film precision resistor. Fortunately, actual resistance value is not too critical. So, if *R*'s calculated value is an odd number, the nearest standard value can be used. This changes the I_{REF} value but not the output voltage. As an example, if a ZN458B had to operate from a +12-volt supply and pass 10 mA, *R*'s calculated value would be 955 ohms. Since this is a difficult value to obtain, you can substitute a 1000-ohm resistor, which changes the current from 10 to 9.6 mA.

Precision Monolithics' REF-01 and -02 shown in Fig. 6 produce output potentials of 10.00 and 5.00 volts, respectively. The untrimmed output from the REF-01, with pin 5 open, is in the 9.9-to-10.1-volt range. With trimming, it can be set to 10.00 volts ±300 mV, which is a 3% range. In most applications, the output can be trimmed to exactly 10.000 volts. Trimmed, the temperature coefficient is 0.7 ppm/°C.

The REF-01 can supply up to 20 mA of current and operates from supplies in the +12- to +40-volt range. A large $V_{IN} - V_O$ difference in any regulator is undesirable, however, because it increases device power dissipation and, hence, temperature.

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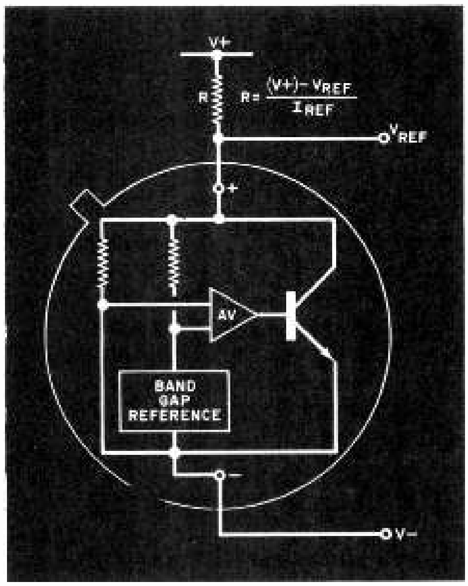


Fig. 5. The Ferranti Electric voltage stabilizers ZN404, ZN423T, and ZN458A/B are used as two-terminal devices.

The 5-volt REF-02 has a 4.975-to-5.025-volt output. It can be trimmed to 5.00 volts ±150 mV (3% range). Input range is 7 to 40 volts, but, once again, avoid high input/output voltage differentials. The REF-02 differs from the REF-01 in that it has a temperature output terminal. Use of an internal band-gap regulator produces an output at pin 3 of approximately 2.1 mV/°C. This allows the REF-02 to be used as a sensor in a simple electronic thermometer project.

The standard thermometer circuit in which the REF-02 is used is a differential operational amplifier. The 5.00-volt output from the REF-02 goes to the inverting input, while the pin-3 output connects to the noninverting (–) input of the op amp. Op-amp gain can be set so the output voltage is numerically the same as the temperature. This type of thermometer is easy to build and can be used to drive a digital voltmeter to obtain a temperature display. Measurement is linear over the –55° to +125° C range of the REF-02, which makes the REF-02 an excellent replacement for nonlinear thermistors.

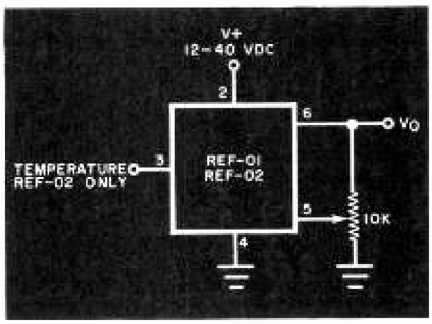


Fig. 6. The Precision Monolithics, Inc. REF-02 reference has a temperature/voltage output on pin 3, for use as an electronic thermometer sensor.

A similar reference source IC is offered by Motorola as the MC1404X and MC1504X devices. These use the same pin-outs as the REF-02 but are housed in 8-pin miniDIP packages instead of the 8-pin metal can of the REF-02. These devices are available in three different standard output voltages: 10 volts (MC1404U10), 6.25 (MC1404J6), and 5 volts (MC1404U5).

Current References. A reference current, rather than a reference voltage, is required in some cases. If load impedance remains constant, a reference current can be generated by applying a known reference voltage to a fixed low-temperature-coefficient resistor. In cases where load impedance may vary, a dynamic regulator circuit that compensates for these changes must be used. Several such circuits are shown below in Fig. 7.

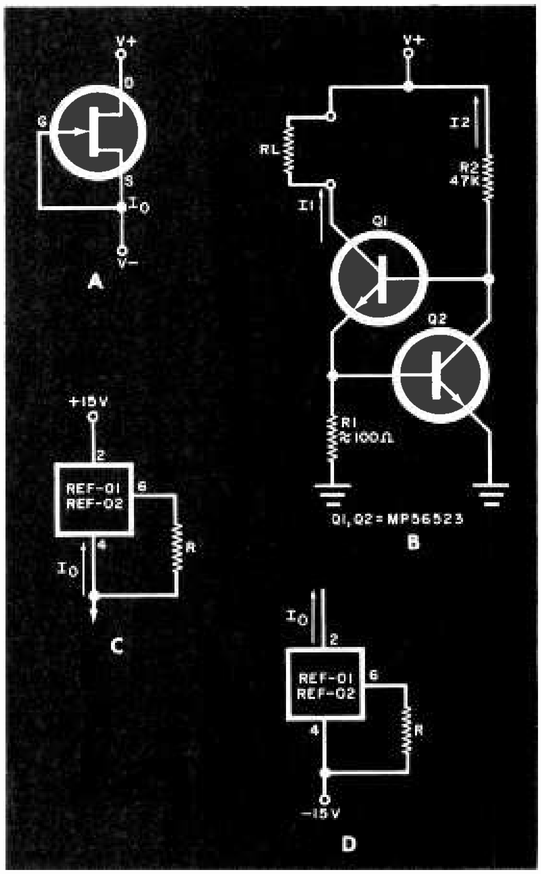
The simplest constant-current source, shown in Fig. 7A, consists of a junction-FET with its gate and source leads connected together. Several manufacturers offer constant-current "diodes" that are little more than a two-terminal package containing a JFET connected as in Fig. 7A. The problem with this type of arrangement, however, is that the current value is determined by the JFET's characteristics. Attempts at varying current by placing a potentiometer in series with the source usually reduce regulation.

A circuit in which two bipolar transistors are used to sink a constant current is shown in Fig. 7B. This circuit was designed around Motorola MPS6523 transistors, but thermal tracking might be improved if a dual npn transistor, such as the MAT-01, is used. Output current I1 is approximately 0.6/*R1*. Current I2 is variable but is best set to about 0.1 × I1.

The REF-01 and -02 voltage regulators can also be used to produce a constant-current source and are capable of either sinking or sourcing current, depending on circuit configuration. Both circuits are the same, except for polarity (Fig. 7C and D, respectively). In the case where a REF-01 is used, output current is (10/*R*) + 1. Since current is usually in milliamperes, it is simplest to express *R* in kilohms to avoid having to convert the decimal number obtained from the equation.

Conclusion. Our aim in this article was to introduce you to the various modern devices that can be used to obtain precision current and voltage references and regulators at low cost. Precision reference/regulators need no longer be confined to the laboratory. ◇

Fig. 7. Constant current sources: (A): using a JFET, (B) bipolar transistors, (C) REF-01 as a current source, and (D) REF-01 as a current sink.



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Build a PINK NOISE GENERATOR for AUDIO TESTING

BY DENNIS BOHN

AN INCREASING number of audiophiles are incorporating graphic equalizers into their hi-fi music systems. The new component is most often used as a "super" tone control that offers a degree of frequency response compensation beyond the capabilities of bass and treble controls. However, adjusting 10 to 30 controls to compensate for acoustic deficiencies in the listening room can be challenging. This project—a pink noise generator—makes the job a little easier. It provides a reference signal for performing equalizer adjustments, and uses just one IC and a few passive components.

The IC, National's MM5837 or AMI's S2688, is a digital pseudo-random sequence generator which will produce a broadband white noise signal for audio applications that's converted to pink noise by a passive filter. Unlike traditional semiconductor junction noise sources, these ICs provide uniform noise quality and output amplitude.

White vs. Pink Noise. The output of the MM5837 is broadband white noise. Since pink noise is used in most audio work, it is helpful to understand the difference between the two.

White noise is a composite signal with contributions from all frequencies and a spectral density substantially independent of frequency (equal energy per constant bandwidth). It is characterized by a 3-dB increase in amplitude per octave of frequency change. In comparison, pink noise has a flat amplitude response per octave of frequency (equal energy per octave). Pink noise allows correlation between successive octave equalizer stages by insuring that the same amplitude of input signal is used for each as a reference.

The network required to convert white noise to pink noise is simply a −3-dB/octave low-pass filter; but it presents an interesting problem in circuit design. If capacitive reactance (and thus the response of a simple RC or first-order filter) varies at a rate of −6 dB/octave, how can a slope of less than −6 dB/octave be obtained? The solution lies in cascading several stages of lag compensation so that the zeros of one stage partially cancel the poles of the next stage. Such a network, shown in Fig. 1, has a −3-dB/octave characteristic (±¼ dB) from 10 to 40,000 Hz.

The complete pink noise generator in Fig. 2 gives a flat spectral distribution (per octave) over the audio band from 20 to 20,000 Hz. An 11.5-V p-p random pulse train appears at pin 3 of the IC, and is attenuated by the filter. The actual output across *C5* is about 1 V p-p ac of pink noise riding on an 8.5-V dc level.

Construction. Since the circuit is fairly simple, it can be constructed on a small circuit board using printed circuit, point-to-point wiring, or Wire-Wrap techniques. Resistors in the filter network should have close tolerances. Premium-grade tantalum and polystyrene, ceramic, and film capacitors are recommended. Observe standard precautions in handling the MOS device, and use an IC socket or Molex Soldercons. ◇

KIT AVAILABILITY

A complete kit of parts, including an MM5837 or S2688 IC and plated etched and drilled pc board but less miscellaneous items is available as Kit PN-1 for $14.95 from TOLECO Systems, P.O. Box 401, Kingston, WA 98346. Washington State residents, please add 5.3% sales tax.

PARTS LIST

C1—1-µF, 35-V metallized polyester
C2—0.27-µF, 35-V metallized polyester
C3—0.047-µF metallized polyester
C4—0.033-µF metallized polyester
C5—10µF, 16-V electrolytic
C6—100-µF, 35-V electrolytic capacitor
IC4—MM5837 noise generator IC
R1—6800-ohm, ¼-W, 5% resistor
R2—3000-ohm, ¼-W, 5% resistor
R3—1000-ohm, ¼-W, 5% resistor
R4—300-ohm, ¼-W, 5% resistor
Misc.—Circuit board, 15-volt regulated supply, output jack, output connector, IC socket or Molex Soldercons, hookup wire, solder, suitable enclosure, etc.

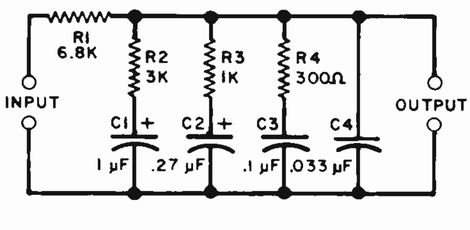

Fig. 1. Low-pass filter with −3-dB/octave response.

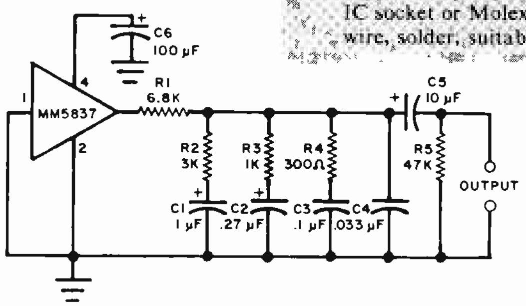

Fig. 2. Schematic diagram of the pink noise generator.

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POWER SUPPLY . . .

(Continued from page 98.)

enough room to wire the switches, potentiometers, etc.

To relabel the meter faces, first remove the meter covers. Then use a pencil eraser to carefully remove the existing numbers and a dry-transfer lettering kit to relabel the meter faces.

Although the wiring of the supply is simple and straightforward, several precautions should be taken to ensure the best possible load regulation. One lead of *R1* should be connected directly to the case of the TO-3 package, the output terminal of the regulator IC. A separate heavy wire is connected from the case of the regulator directly to the positive output jack. Finally, the wiper of the VOLTAGE ADJUST control and the potentiometer lug connected to it should be wired directly to the negative output jack. This improves load regulation by providing remote ground sensing directly at the output of the supply.

In Conclusion. The finished supply will prove to be a worthwhile investment for any home lab. Although the goal of simplicity guided the design of this power supply (less than 25 components were used), its performance is more than adequate for most bench work. With the VOLTAGE ADJUST control set for 15 volts, a 1-ampere load current will typically result in less than a 15-mV drop at the output and less than 1 mV peak-to-peak of ripple. Varying the ac from 90 to 125 volts causes less than a 10-mV output change. Output voltage drifts of 0.01%°C can be obtained if stable components are used for *R1* and *R2*.

With a few additional parts, the output can be made to go down to zero. This is accomplished by connecting the wiper (and the lug connected to it) of the VOLTAGE ADJUST potentiometer to a stable −1.25-volt reference. Output current can be increased by replacing the LM317 with either an LM350 (3 amperes) or an LM338 (5 amperes). If this is done, the power transformer and rectifier must be replaced with components that have increased current-handling capabilities. If a bipolar dc power supply is needed, you can build two supplies and connect the positive output post of one to the negative output post of the other. Alternatively, you can build a modified version of this project incorporating both an LM317 and a negative-voltage LM337. ◇

WORST CASE . . .

(Continued from page 119.)

5000 ohms. (If you were making only one circuit, you could hand-select the components to make it work, but this is not a safe approach to use in a construction article.) Now note t_{wQ} when C_{ext} is 1000_pF and R_{ext} is 10,000 ohms. The width of the pulse can be between 2.76 and 3.37μs. Hence, the value can range from +8.9% to −11.2% of the typical specified value for the given R and C values. Note also that the spec sheet does not tell you that this error is linear throughout the t_{wQ} range. For all we know, this may be the best point on the curve. So, when designing such a circuit, make certain that your design can accommodate this type of tolerance.

Note the column in Table II headed Nom (nominal). This value is the one for which you should strive, but you may find that it is not possible to obtain or hold it through the design.

It should be understood that one parameter may affect another. For example, consider the effect of varying the power supply voltage on the output sinking current (I_{OL}). The output sinking current is a linear function of the power supply voltage, as shown in Fig. 2. When the supply potential is 4.75 volts, the output can sink 15 mA. A similar condition can be observed in Fig 3, where the maximum input forward current (I_F) is shown as a function of input voltage. Here again, the variation of one parameter can affect another.

At this point, you should realize that you must know which characteristics are important so that you can design with a knowledge of their probable variations. To do this, you must know just what will affect a given parameter.

All of the parameters thus far discussed have been of the type that can cause circuit failure, not failure of a component. Most IC data sheets carry a set of catastrophic characteristics, such as those listed in Table III. With resistors and capacitors, characteristics like maximum power dissipation and breakdown voltage should never be exceeded.

Summing Up. If you use the techniques detailed in this article, or keep them in mind, your circuits will work and so will other circuits built from your design. If you build projects from magazines, steer clear of broad-tolerance components, especially in critical components. ◇

NOTICE TO READERS

We consider it a valuable service to our readers to continue, as we have in previous editions of this guide, to print the price set by the manufacturer or distributor for each item described as available at presstime. However, almost all manufacturers and distributors provide that prices are subject to change without notice.

We would like to call our readers attention to the fact that during recent years the Federal Trade Commission of the U.S. Government has conducted investigation of the practices of certain industries, in fixing and advertising list prices. It is the position of the Federal Treade Commission that it is deceptive to the public, and against the law, for list prices of any product to be specified or advertised in a trade area, if the majority of sales of that product in that trade are made at less than the list prices.

It is obvious that our publication cannot quote the sales price applicable to each trading area in the United States. Accordingly, prices are listed as furnished to us by the manufacturer or distributor. It may be possible to purchase some items in your trading area at a price that differs from the price that is reported in this edition.

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ELECTRONIC EXPERIMENTER'S HANDBOOK 1982

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Input Impedance:	1 megohm in parallel with 50 pF
Time Base:	0.05 μSec/div to 0.2 Sec/div, 21 calibrated ranges
Horizontal Bandwidth:	200 kHz
Trigger Modes:	Automatic, Internal, External and Line
Power Sources:	
Internal:	Rechargeable lead acid batteries
External:	115 VAC or 230 VAC, 50-60 Hz via plug-in transformer
Size:	2.9" H x 6.4" W x 8.6" D (74 mm x 163 mm x 218 mm)
Weight:	3 lbs. 10 oz. (1.65 Kg)

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