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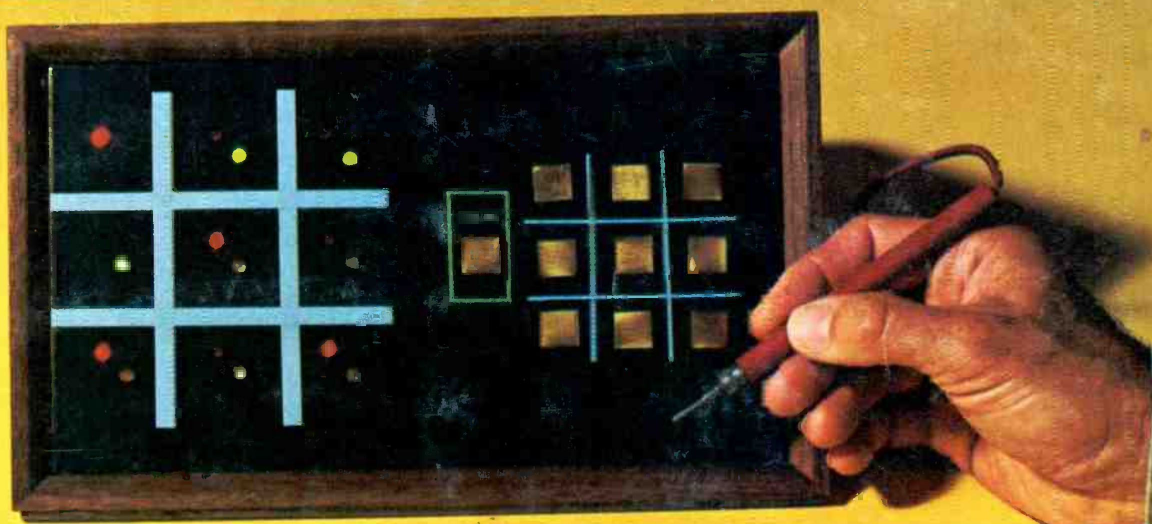

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MATERIALS

Anodized aluminum bases (also serve as ground plane); acetal copolymer dielectric; non-corrosive nickel/silver tie-point terminals; rubber bench feet.

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COVER PHOTO: Conrad Studios
For details on how to build the "Tic-Tac-Toe Logic Machine," see page 9

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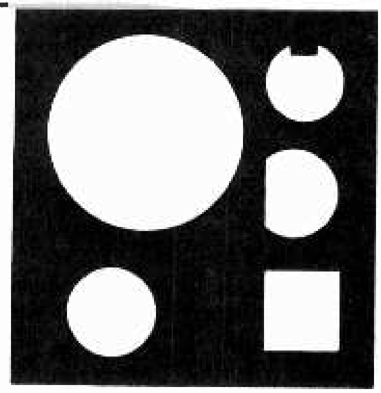

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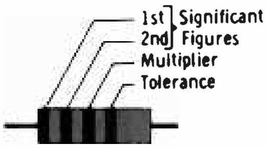
COLOR CODE CHARTS

COLOR BAND SYSTEM



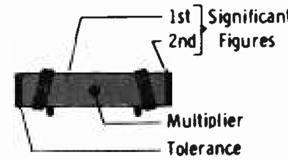
Resistors With Black Body Color Are Composition, Non-Insulated.
Resistors With Colored Bodies Are Composition, Insulated.
Wire-Wound Resistors Have The 1st Digit Color Band Double Width.

RESISTOR CODES (RESISTANCE GIVEN IN OHMS)

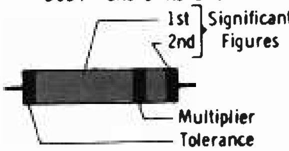
COLOR	DIGIT	MULTIPLIER	TOLERANCE
BLACK	0	1	±20%
BROWN	1	10	±1%
RED	2	100	±2%
ORANGE	3	1000	±3% *
YELLOW	4	10000	GMV*
GREEN	5	100000	±5% (EIA Alternate)
BLUE	6	1000000	±6%*
VIOLET	7	10000000	±12 1/2%*
GRAY	8	.01 (EIA Alternate)	±30% *
WHITE	9	.1 (EIA Alternate)	±10% (EIA Alternate)
GOLD		.1 (JAN and EIA Preferred)	±5% (JAN and EIA Preferred)
SILVER		.01 (JAN and EIA Preferred)	±10% (JAN and EIA Preferred)
NO COLOR			±20%

*GMV = guaranteed minimum value, or -0 + 100% tolerance.
±3, 6, 12 1/2, and 30% are ASA 40, 20, 10, and 5 step tolerances.

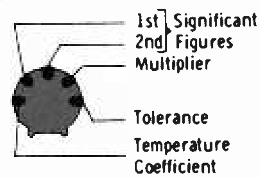
BODY-END-DOT SYSTEM



BODY - END BAND SYSTEM



DISC CERAMICS (5-DOT SYSTEM)



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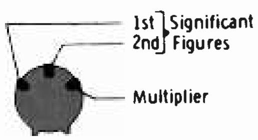
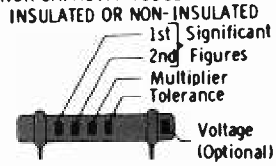


CERAMIC CAPACITOR CODES (CAPACITY GIVEN IN pF)

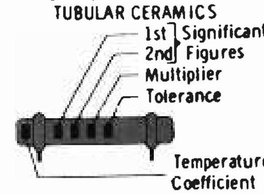
COLOR	DIGIT	MULTI-PLIER	TOLERANCE 10 pF or LESS	TOLERANCE OVER 10MMf	TEMPERATURE COEFFICIENT PPM/°C	EXTENDED RANGE TEMP. COEFF. SIGNIFICANT FIGURE	EXTENDED RANGE TEMP. COEFF. MULTIPLIER
BLACK	0	1	±2.0 pF	±20%	0(NPO)	0.0	-1
BROWN	1	10	±0.1 pF	±1%	-33(N033)		-10
RED	2	100		±2%	-75(N075)	1.0	-100
ORANGE	3	1000		±2.5%	-150(N150)	1.5	-1000
YELLOW	4	10000			-220(N220)	2.2	-10000
GREEN	5		±0.5 pF	±5%	-330(N330)	3.3	+1
BLUE	6				-470(N470)	4.7	+10
VIOLET	7				-750(N750)	7.5	+100
GRAY	8	.01	±0.25 pF		+30(P030)		+1000
WHITE	9	.1	±1.0 pF	±10%	General Purpose		+10000
SILVER					Bypass & Coupling		
GOLD					+100 (P100, JAN)		

Voltage ratings are standard 500 volts for some manufacturers, but 1000 volts for other companies.

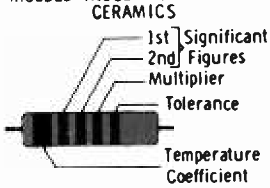
HIGH CAPACITY TUBULAR CERAMIC INSULATED OR NON-INSULATED



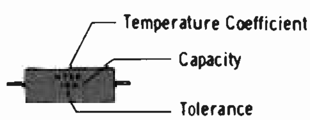
TEMPERATURE COMPENSATING TUBULAR CERAMICS



MOLDED-INSULATED AXIAL LEAD CERAMICS

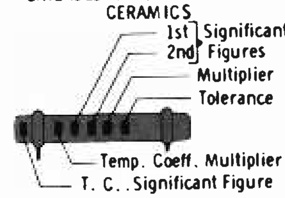


TYPOGRAPHICALLY MARKED CERAMICS

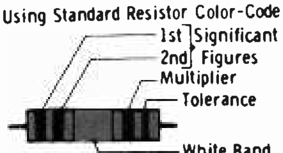


JAN LETTER	TOLERANCE 10 pF or LESS	TOLERANCE OVER 10 pF
C	±0.2 pF	
D	±0.5 pF	
F	±1.0 pF	±1%
G	±2.0 pF	±2%
J		±5%
K		±10%
M		±20%

ENTENDED RANGE T.C. TUBULAR CERAMICS

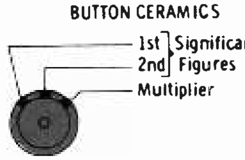


MOLDED CERAMICS
Using Standard Resistor Color-Code



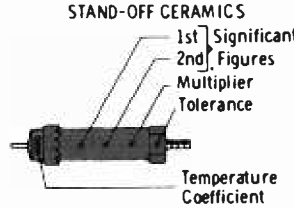
Distinguishes Capacitor From Resistor

BUTTON CERAMICS

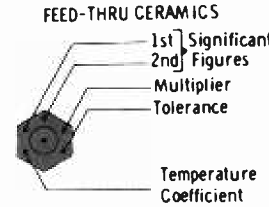


Viewed From Soldered Surface

STAND-OFF CERAMICS



FEED-THRU CERAMICS



MOLDED MICA CAPACITOR CODES
(Capacity Given In pF)

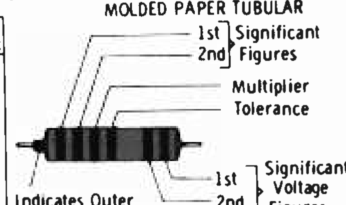
COLOR	DIGIT	MULTI-PLIER	TOLERANCE •	CLASS OR CHARACTERISTIC
BLACK	0	1	20%	A
BROWN	1	10	1%	B
RED	2	100	2%	C
ORANGE	3	1000	3%	D
YELLOW	4	10000		E
GREEN	5		5% (EIA)	F(JAN)
BLUE	6			G(JAN)
VIOLET	7			
GRAY	8			I(EIA)
WHITE	9			J(EIA)
GOLD		.1	5%(JAN)	
SILVER		.01	10%	

Class or characteristic denotes specifications of design involving Q factors, temperature coefficients, and production test requirements.
All axial lead mica capacitors have a voltage rating of 300, 500, or 1000 volts.
•or ±1.0 pF whichever is greater.

MOLDED PAPER CAPACITOR CODES
(Capacity Given In pF)

COLOR	DIGIT	MULTI-PLIER	TOLERANCE
BLACK	0	1	20%
BROWN	1	10	
RED	2	100	
ORANGE	3	1000	
YELLOW	4	10000	
GREEN	5	100000	5%
BLUE	6	1000000	
VIOLET	7		
GRAY	8		
WHITE	9		10%
GOLD			5%
SILVER			10%
NO COLOR			20%

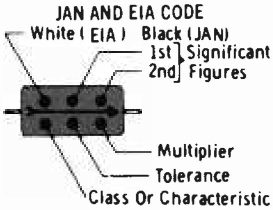
MOLDED PAPER TUBULAR



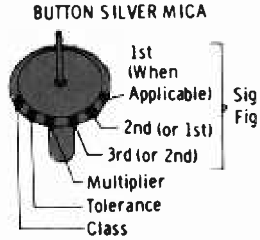
Indicates Outer Foil. May Be On Either End. May Also Be Indicated By Other Methods Such As Typographical Marking Or Black Stripe.

Add Two Zeros To Significant Voltage Figures. One Band Indicates Voltage Ratings Under 1000 Volts.

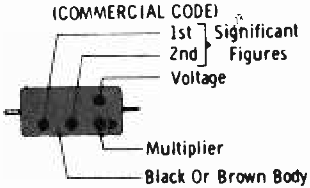
CURRENT STANDARD JAN AND EIA CODE



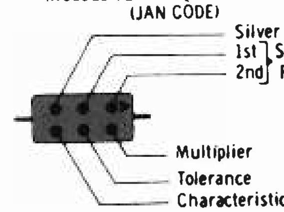
BUTTON SILVER MICA



MOLDED FLAT PAPER CAPACITORS (COMMERCIAL CODE)



MOLDED FLAT PAPER CAPACITORS (JAN CODE)



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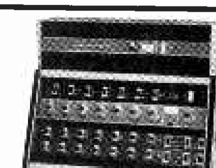

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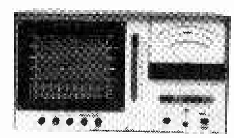

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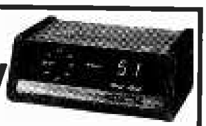

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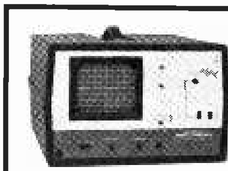

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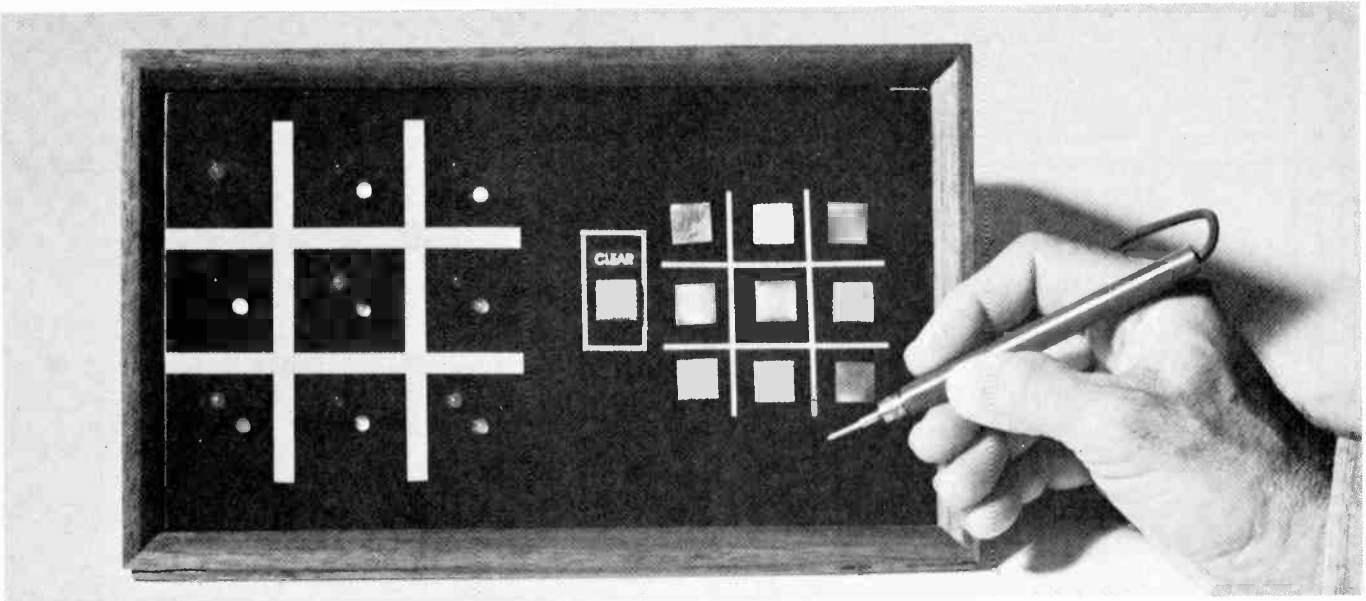


BUILD THE "TIC-TAC-TOE" LOGIC MACHINE

Features low-cost diode construction and electronic "win" secret

BY HERB COHEN

THERE'S hardly a person in the Western World who hasn't played the universally popular game, "tic-tac-toe." Here's the same game, but with a new twist—playing against an electronic logic machine that one cannot beat. (The best that the human challenger can do is tie this logic machine.)

Previous electronic tic-tac-toe projects required a substantial investment in sophisticated integrated circuits, whereas the circuit presented here consists largely of inexpensive diode logic. Furthermore, assembly is easier. And, as an extra fillip, there's a concealed "cheat" switch to force the machine's logic circuits to make a mistake, enabling the owner to beat the "unbeatable" machine while other players cannot.

The game is played in true computer style, using an electronic "pencil" and a grid of touchplates. As shown in Fig. 1, readout is via red LED's to indicate machine moves and green LED's to show where the human player goes. Nine of the touch squares form the conventional tic-tac-toe pattern while the tenth square is used to reset the game.

The machine always makes the first move, as indicated by red *LED9* within square-1 (upper left corner) coming on as power is applied (or after reset). Using the electronic "pencil," the player can then touch any other square, which will cause a green LED to illuminate at that position. The machine makes an instantaneous response by causing a red LED to come on as the machine has already decided on a game plan based on the player's first choice. The machine will try for a winning move on its second countermove. If that square is already taken by the player, the machine will try for a double win in the last move.

An example of this is the following game: machine has square-1 to start,

(Text continues on page 11)

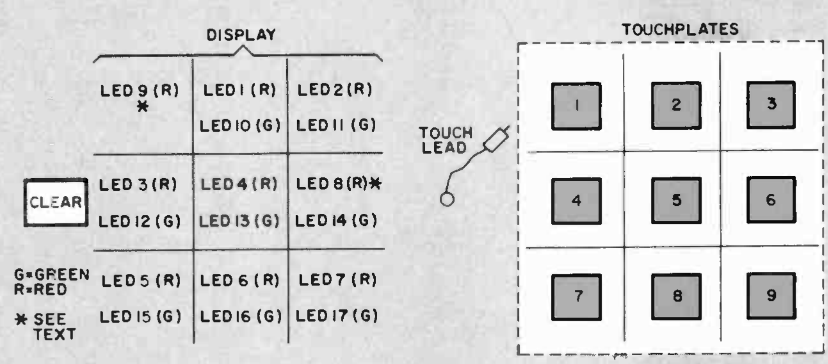
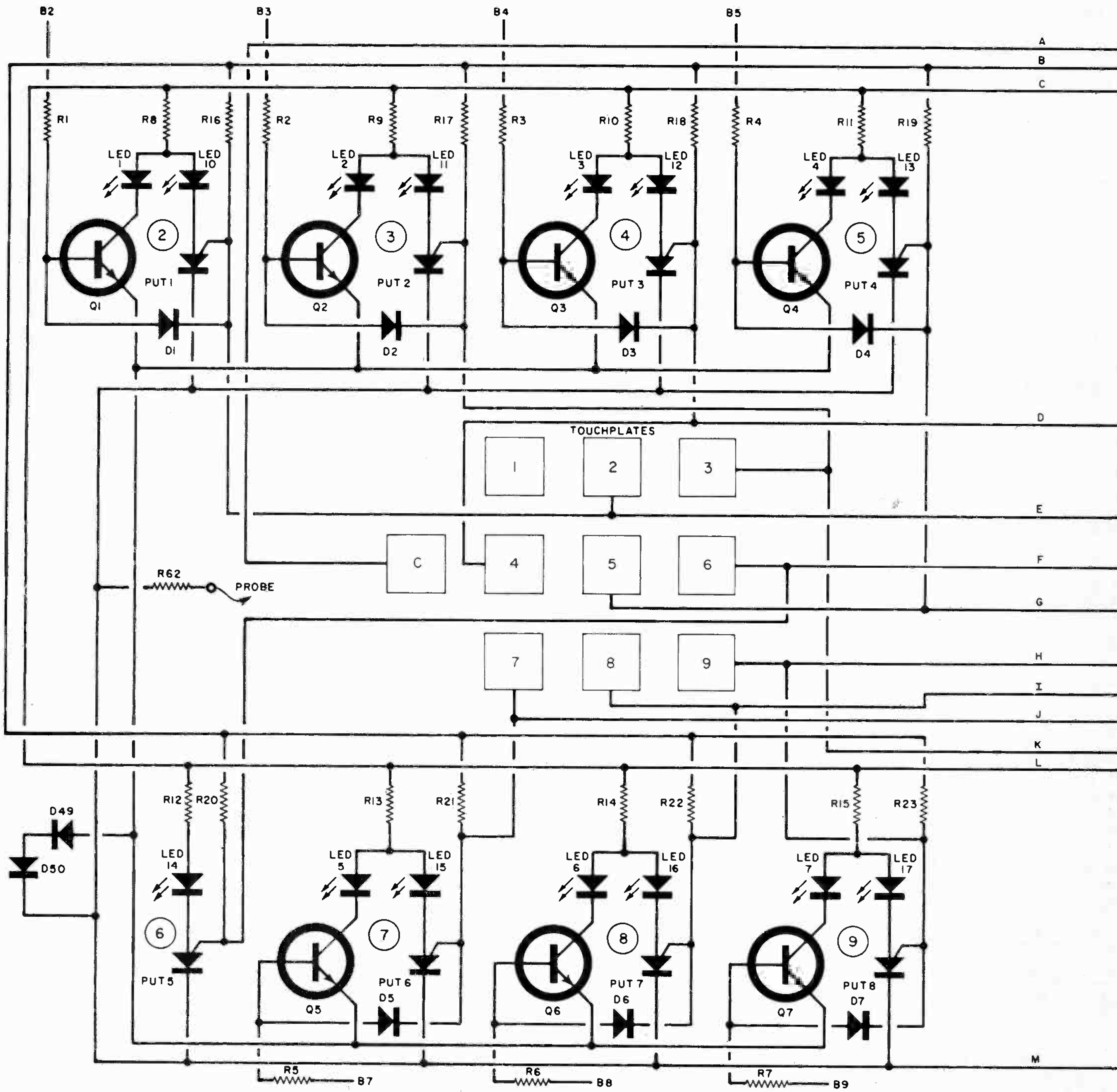


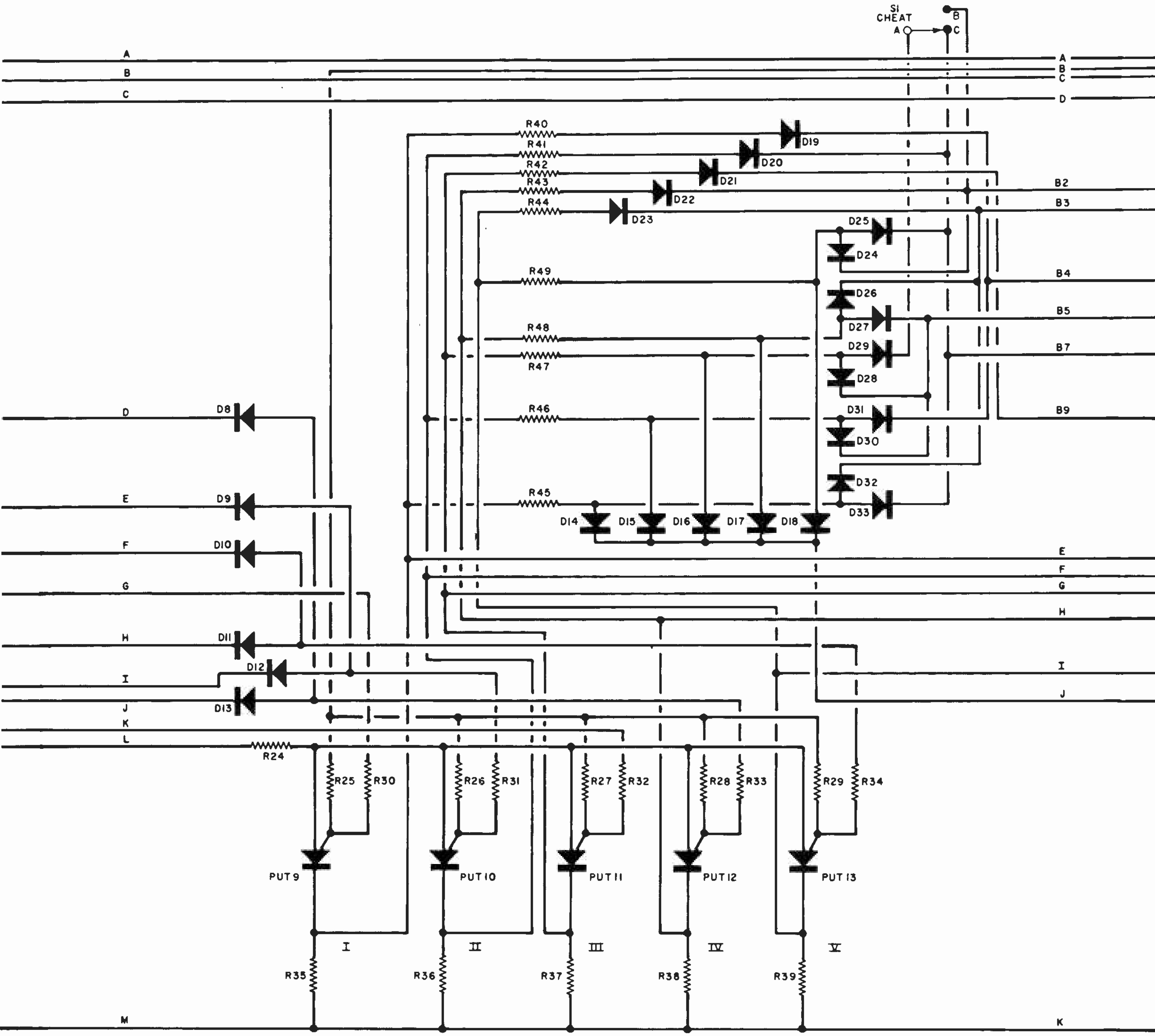
Fig. 1. Layout of the touch squares and operation of the LED's.



PARTS LIST

C1—0.05-μF disc capacitor
C2—1000-μF, 25-volt electrolytic capacitor
D1—D48—1N914
D49—D52—1N4003
D53—D56—50-volt, 1-A rectifier
D57—4.7-volt, 1-W zener diode (HEP Z0405)
LED1-LED9—Any red LED (Note: LED 8 is a dummy LED to complete the matrix)
LED10—LED17—Any green LED
PUT1—PUT13—Programmable unijunction transistor (Motorola MPU133)
Q1—Q9—2N4946, HEP S3019
Q10,Q12—2N4917, HEP S0013
Q11—40406 (RCA or HEPS5013)
R1—R7—3900-ohm, ¼-watt resistor
R8-R15—470-ohm, ¼-watt resistor
R16—R23, R25-R34—10,000 ohm, ¼-watt resistor
R24,R35—R39,R45—R54—1000-ohm, ¼-watt resistor
R40—R44,R63—2200-ohm, ¼-watt resistor
R55,R56—22,000-ohm, ¼-watt resistor.
R57—5600-ohm, ¼-watt resistor
R58,R59—3300-ohm, ¼-watt resistor
R60—560-ohm, ¼-watt resistor
R61—1500-ohm, ¼-watt resistor
R62—6800-ohm, ¼-watt resistor
R64—5000-ohm trimmer potentiometer
S1—Spdt switch
T1—Filament transformer: 12-V, 1-A
Misc.: Test lead with tip; metal for touchplates; tape for marking; epoxy; line cord; mounting hardware; perf board and mounting clips (if used); suitable cabinet; etc.
Note: A PC board is available from: The Lynkeus Corporation, P.O. Box 1512, East Hampton, N.Y. 11937. (Model TTT-1) $16.00, postpaid. New York residents add applicable sales tax.

Fig.2. Complete circuit (third section on next page) and parts list for "Tic-Tac-Toe."

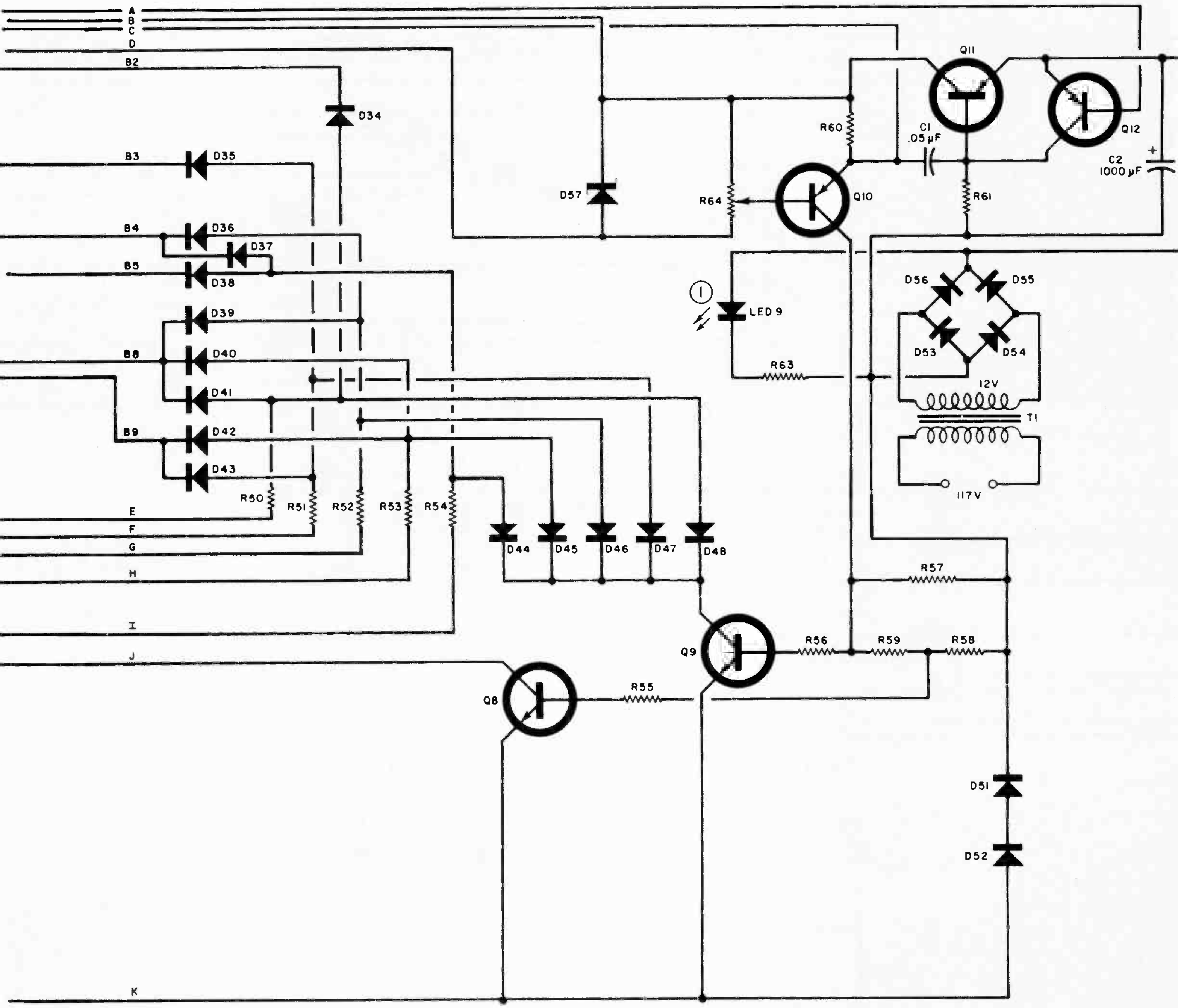


the player selects square-3, the machine plays square-9, the player goes to square-5, then the machine plays square-7—catching the player in a fork. On his next move, the player would have to block square-4 and square-8 simultaneously to prevent a machine win—a move that is illegal (and impossible). If nobody has a clear-cut win by the fourth move, the game is a tie.

Circuit Operation. The complete circuit is shown in the three parts of Fig. 2. The probe consists of a test lead connected to ground via *R62*. Now assume you touch plate-9. This action loads the gate of *PUT8*, firing it and causing green *LED17* to glow. It also holds *Q7* off by clamping *D7* to ground through the PUT. The gate also clamps *D11* to ground, thereby firing game plan-1 *PUT13*.

Once a game-plan PUT is fired, it latches out the remaining game-plan PUT's to assure only one game plan per game. As *PUT13* fires, six volts appear across *R39*. Current then passes through *R44, D33,* interconnects point *R3*, to the base of *Q2* which, in turn, causes *LED2* (red) to come on at square-3. This is the machine's response to the player touching square-9.

The voltage across *R39* also feeds *R49*, which is the machine's second countermove, and *R54*, the machine's last countermove. Both resistors are shorted to ground via diodes *D18* and *D44* and transistors *Q8* and *Q9*. Both transistors are biased on by *Q10* acting as a constant-current source.



The current from all the gates of the player's PUT's must pass through *R60*. As each player's LED (green) goes on, the current through *R60* would normally increase. However, since *Q10* is a constant-current source, this current is held at a constant value.

On the first player move, transistors *Q8* and *Q9* are both in saturation. On the second player move, *Q8* is turned off. Current now flows through diodes *D24* and *D25*, through *B2* and *B7*, to drive *Q1* and *Q5* on. Transistor *Q2* controls square-2 while *Q5* controls square-7, and the game continues.

The player's third move adds more gate current to that already flowing through *R60*, making it enough to turn *Q10* off. In response, *Q9* turns off, thus initiating the third machine response.

The game is "cleared" by touching the clear square with the probe. This action turns on *Q12*, which turns off *Q11* (the power pass transistor). The PUT's lose power and are automatically reset.

It's frustrating not being able to beat the machine logic (the best you can do is tie). Because of this, a "cheat" switch *(S1)* has been added. Diode *D29* is switched from *B2* to *B7*. Then in the *B7* position, the game runs as follows: player takes square-3, the machine responds with square-9. Then, if the player takes square-5, the machine, which would normally respond by taking square-7 thus blocking the player's 3,5 diagonal and setting the machine for a win on the third move, instead will select square-2, allowing the player to complete his diagonal line by taking square-7 on his third move and winning the game. Of course, this "cheat" switch should be hidden from view (preferably on the bottom of the chassis where it can be operated secretly when the player

(Text continues on page 14)

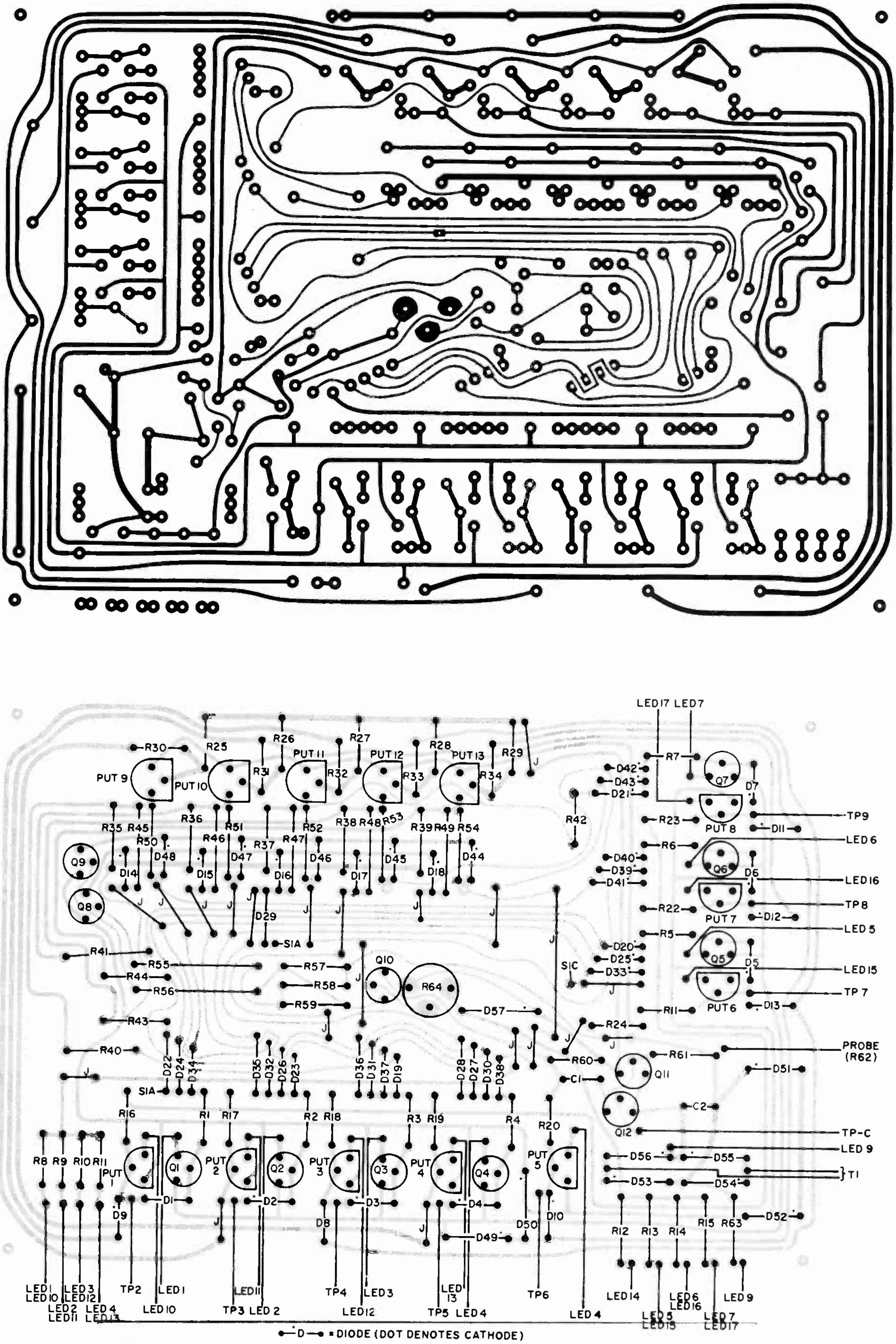


Fig.3. (Top) Full-size foil pattern for PC board; (bottom) component parts placement.

physically moves the chassis to face him).

Construction. The circuit, other than the LED's and 10 touchplates, can be built up on a PC board, like that shown in Fig. 3. If desired, point-to-point wiring on a conventional perf board may be employed. The wiring is not critical, but there is a lot of it. So take care during construction.

The easiest way to make the touchplates is to cut small squares from a sheet of PC board. Drill a small hole through the cabinet top at the center of each touchplate position, then a mating hole through each touchplate. Feed an insulated lead through the non-foil side of the PC touchplates, and solder on the foil side. Feed the wire through the mating hole on the cabinet top and wire it to the circuit board. The author's prototype used a plastic cabinet cover. Each touchplate was made from a small square of brass sheet, soldered to thick pins which mount on the plastic cover. Wiring was from the bottom of the cover. Drill a hole for the lead and mount lead.

The LED's are mounted in a tic-tac-toe configuration, as shown in Fig. 1 and the photograph. Note that square-1 (the machine's first move) has only one red LED (*LED9*). The other squares are formed by two LED's, one green (player) and one red (machine). Drill mounting holes just large enough for the LED's to make a tight fit. Install each pair of LED's near the center of each tic-tac-toe square and epoxy them in place. Narrow lengths of white tape can be used to create the criss-cross pattern. Note also, that the red LED in square-6 (*LED8*) is not actually used and a dummy LED is used to complete the matrix.

Once the wiring has been completed, recheck to eliminate possible shorts, wrong connections, etc.

Operation. Once the circuit is built, turn the power on and adjust potentiometer *R64* until *Q10* is cut off and no voltage appears across *R57*.

Using the test probe (clear lead), contact square-5 (the center square). Red LED's 4,3,7,2, and 8 may come on. Adjust potentiometer *R64* until only *LED4* (center row, first column) remains lit. Next touch square-7 and note that *LED3* comes on. If not, adjust *R64* until it does. Last, touch square-2 and note that red *LED10* lights. Cement the rotor of *R64* so that it will not move and you are in business. ◈

UNIVERSAL DIGITAL CLOCK ALARM FUNCTION

BY EDWARD FRIEDMAN

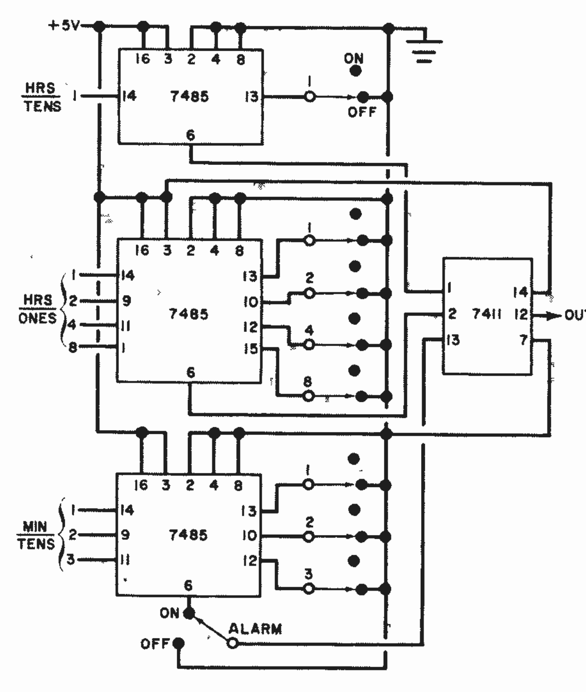
THE popularity and availability of electronic digital clocks suggests some interesting applications and modifications of these devices. The most obvious change is to add an alarm function, and this is the purpose of the circuit shown below. Although the circuit is general in nature, it shows one approach to creating the alarm function.

The BCD data (1, 2, 4, 8) from the tens and unit hours and the tens of minutes are extracted from the appropriate digital counters. The data from each digit goes to a 4-bit comparator (7485) which is wired to produce a positive-going output (at pin 6) when the counter input data exactly matches the switch-selected values of the desired alarm time.

ception of the minute-tens digit, the standard 1-2-4-8 code is used. The minute-tens digit must rest to 0 after a count of 5, and the output of this modulo-6 counter is coded 1-2-3.

If the switches are set as just described, at exactly 07:30, all three 7485's will have positive outputs. If the "alarm" switch is turned "on," this data will appear at the three-input AND gate (7411) which then produces an output signal for as long as the switch settings are appropriate. In this case, the alarm will remain on for 10 minutes, or until the "alarm" switch is turned off.

The output of the 7411 can be used to drive an npn transistor using a couple of hundred ohms in series with the base. The load for this transistor can



Four comparators combine clock digit counters with switch-selected inputs and, when combination is correct, output is developed to sound alarm.

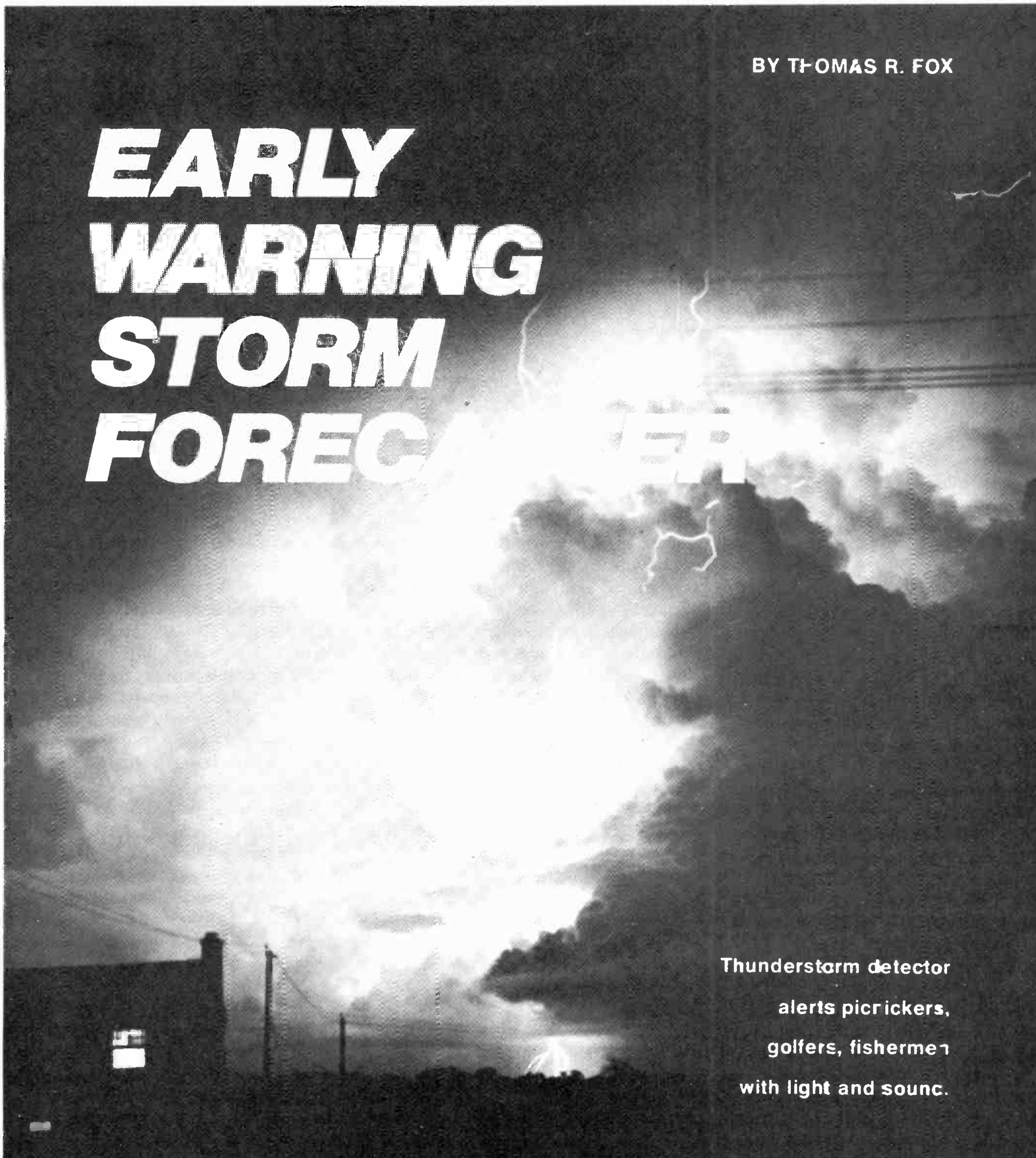
Let's assume you wish the alarm to sound at 07:30. The hours-tens switch is left closed (off); the hours-ones switches are all opened (on) except "8" (1 + 2 + 4 = 7); and the minutes-tens switch 3 is opened.

It should be noted that with the ex-

be a lamp, LED, buzzer, or other audible alarm device. A relay can be used as the collector load, with this relay driving almost any type of alarm. You can also use this output to drive an SCR or triac if higher driving power is required. ◈

BY THOMAS R. FOX

EARLY WARNING STORM FORECASTER



Thunderstorm detector alerts picnickers, golfers, fishermen with light and sound.

SUDDEN violent storms cause many millions of dollars worth of damage in the United States every year. Even worse, they sometimes cause the loss of human lives. In most cases, such storms are detected in advance by authorities and emergency announcements are made over local radio and TV stations.

However, many times people are not listening to such broadcasts. They may be sleeping, boating, camping out, golfing, or just sunning on the beach. If they can get sufficient warning, they can take steps to protect themselves and their property from the elements. That is the purpose of the Storm Forecaster.

The Forecaster uses the sferics generated by lightning—lightning flashes that often announce the approach of a storm—to turn on a warning light or provide an audible alert signal. Since it is battery powered, the

PARTS LIST

A1—Audible signal device (Mallory Sonalert or equiv.)
B1,B2—9-volt battery
C1—3000-μF electrolytic capacitor (see text)
D1—Silicon rectifier diode
I1—#48 pilot lamp
Q1—Transistor (RCA 40399 or HEP721)
Q2—Transistor (2N307 or HEP230)
R1—1-megohm potentiometer
R2—100,000-ohm potentiometer
R3—25,000-ohm potentiometer
R4,R5—100-ohm, ½-watt resistor
R6—15-ohm, ½-watt resistor
S1—Dpst switch
S2—Spst normally closed pushbutton switch
SCR1—Silicon controlled rectifier (C106AL or GEX-5)
Misc.: Miniature transistor radio; suitable plastic case with plastic cover (6" × 5" × 2"); battery connectors; knob; lamp socket; mounting hardware; etc.

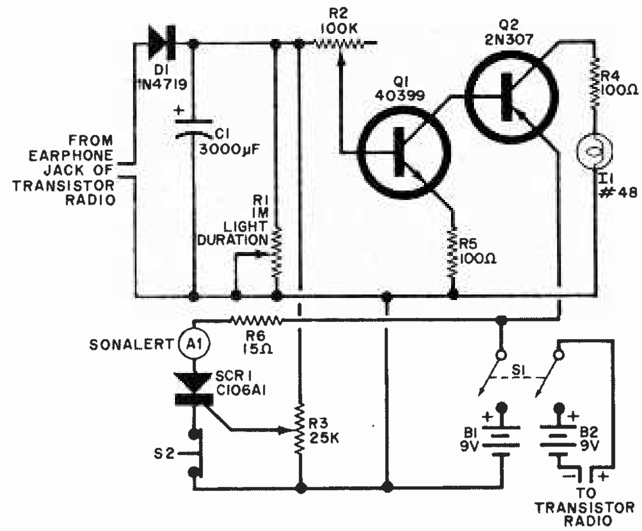


Fig. 1. Static "noises" from the radio charge up C1 which turns on lamp I1. SCR circuit can also be set to provide audible as well as visual warning.

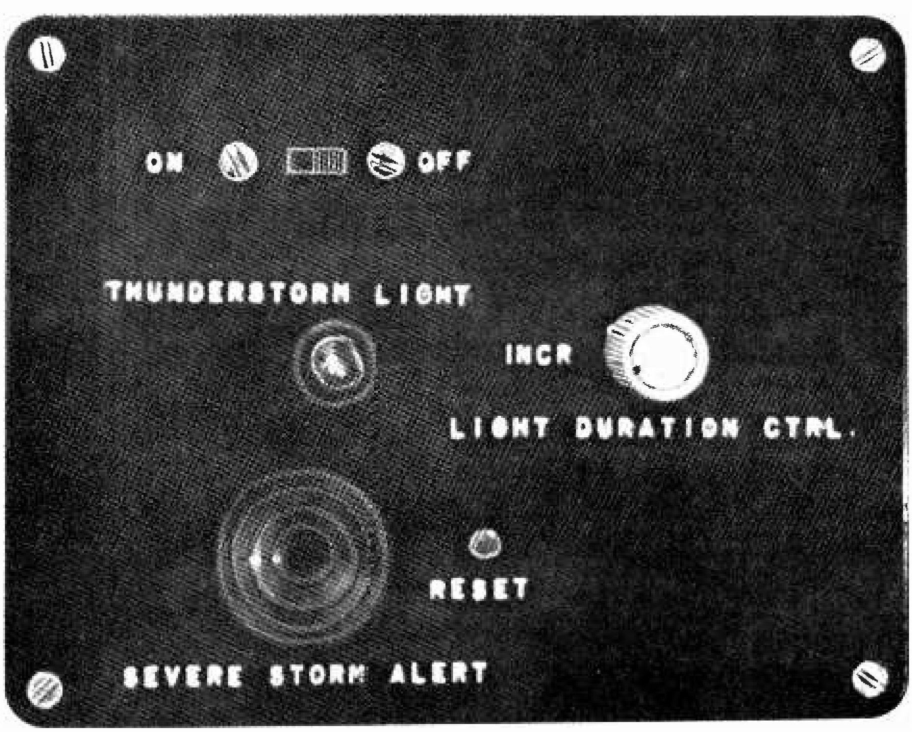
device can be used far from the commercial power line. The detection range can be pre-set to a value of a few miles to several hundred miles and, when a storm comes within the detection range, the warning is given.

Lightning flashes create radio waves that can best be detected in the frequency range between 100 kHz and 1MHz (long waves at the lower end of the conventional broadcast band). If you tune a conventional radio to the 550-kHz end of the broadcast band, weak lightning crashes as much as 100 miles away will cause static on the radio. The louder the static, the closer the storm. Because of the avc (automatic volume control) action in most radios, and if you are listening to a station on the low end of the dial, the static will be louder on a radio tuned to a weak station than one tuned to a strong station.

In addition to signaling the approach of a possible dangerous thunderstorm, the Forecaster is useful indoors as a tornado warning device. There is a strong relationship between severe thunderstorms and the formation of a tornado. In fact, tornados are formed in thunderstorms, but the exact mechanism is not known. The U.S. Weather Service issues Tornado Watches to inform the public that weather conditions may be ripe for the formation of tornados. Researchers at Iowa State University believe that frequencies between 1 and 53 MHz appear to be the most useful in detecting tornados.

When there is a Tornado Watch called for in your area, just turn on the Forecaster and go about your business. This is especially helpful if the warning is called late in the evening and you want to get as much sleep as possible. The loud acoustic warning signal generated by the Forecaster will wake you up.

Front panel of Forecaster holds warning light and severe storm alert.



Construction. The main element of the Forecaster is a small (shirt-pocket) solid-state radio. These can be purchased almost anywhere at very low cost and most are surprisingly sensitive. There are no changes to be made in the radio, except for the use of the earphone jack and cable. When not used in the Forecaster, the radio can operate quite conventionally.

The circuit shown in Fig. 1 can be assembled on a small piece of perf board or a PC board. In either case, the smaller the better. To avoid electromagnetic shielding, the outer case must be either plastic or Bakelite. Do not use a metal case. Pick a size that can accommodate the radio, the small board, and two 9-volt batteries. Light duration control *R1*, the normally closed pushbutton reset switch *S1*, the thunderstorm light, the Sonalert, and the power switch are all mounted on the front (plastic) cover of the case.

Although a Sonalert is specified for the severe storm alert, you can substitute a low-current buzzer; or if you live in a bad storm area, you can use a relay instead of the Sonalert. The relay can actuate a burglar-alarm-type of siren. It is also possible to substitute a sensitive relay for *R4* and *I1*, and use this relay to turn on either another audible warning device or power a much brighter (and hence more visible) light. The various modifications can be made to suit almost any condition.

The input to the circuit board is through a conventional headphone jack plugged into the transistor radio. Remove the earphone and connect the two leads as shown in Fig. 1. This should also silence the radio.

How it Works. The input from the radio earphone jack is rectified by *D1* and places a charge on *C1*, which has a large capacitance value. When the charge on *C1* builds up to a sufficient value, it energizes two separate circuits. One circuit is a two-stage dc transistor amplifier (*Q1* and *Q2*) which is easily driven into saturation and whose output turns on *I1*. The second circuit consists of a sensitive SCR whose output drives the audible alert. Reset switch *S2*, when depressed, opens the cathode circuit of the SCR and turns off the alarm if *C1* is sufficiently discharged.

Calibration. The radio used should be tuned to the lowest frequency that is free of broadcast stations. It should be left in this position unless the Forecaster is moved to an area in which a station happens to be broadcasting at, or near, the frequency to which the radio is tuned.

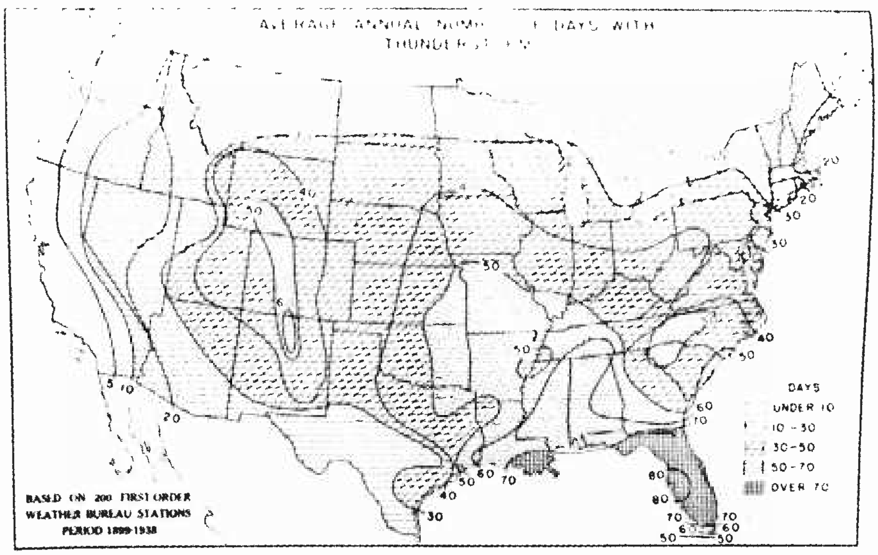
Light-duration control *R1* does not have to be set at a specific point for the Forecaster to work. The purpose of the potentiometer is to provide an easy way of discharging *C1* (when desired) and a way of judging the intensity of an approaching storm by setting the potentiometer close to its minimum value. In that case, if the storm is weak, *I1* will go out periodically.

Controls *R2* and *R3* must be properly adjusted before the device can be used. Potentiometer *R2* is basically a sensitivity control for the warning light, while *R3* controls the input current to *SCR1*. The former should be adjusted so that the thunderstorm warning light goes on at the desired amount of static. For example, you may want to be alerted if moderate thunderstorms are 100 miles (almost 3 hours) away. By listening to the radar weather reports put out by the FAA on the longwave band or to a TV weather program, you can adjust *R2* so that the light just barely goes on when storms are that far away.

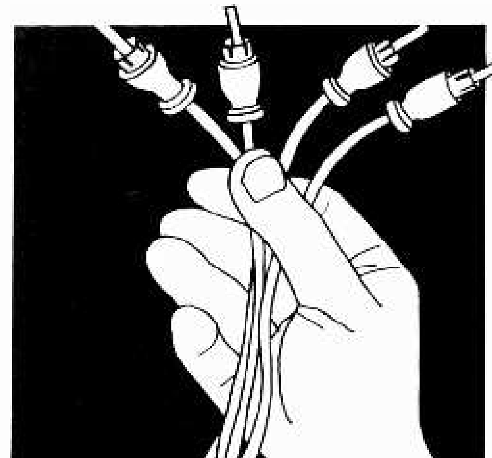
Potentiometer *R3* should be set so that the severe storm alert goes off when a bad storm is quite close (25 miles, for example). The setting of this potentiometer is rather critical. Once you have adjusted *R3*, re-adjust *R2*.

Since thunderstorms appear when you least expect them, you can give the Forecaster an initial test and adjustment by creating your own static. A soldering gun, when turned on and off rapidly, sends out a signal similar to that of a lightning flash. Corrections in this initial adjustment should, of course, be made when an actual thunderstorm appears. If you want to be warned of the presence of a weak storm (with little lightning), use a smaller capacitance value for *C1* (2000-µF or less). Since less current is required to charge a smaller capacitor, fewer pulses from the radio will be required to charge *C1*. With this modification, however, the thunderstorm warning light won't be on constantly during most storms and the increased sensitivity may be troublesome.

After it is checked out and adjusted, the Forecaster may falsely signal that a thunderstorm is near. In this case, some nearby electrical equipment is probably emitting pulses or electrical "hash." A faulty fluorescent light or motor may be to blame. The interference must be corrected or removed before the Forecaster can function properly. ◈



Map of U.S. showing average number of days in which thunderstorms occur in various areas of country.



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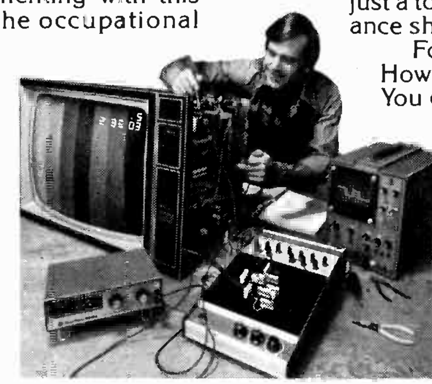

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Build the Super Audio Sweep Generator

PROVIDES LOG AND LINEAR SWEEP. DELIVERS 1 Hz TO 100 kHz SINE, TRIANGLE, SQUARE WAVEFORMS AT A BREAKTHROUGH LOW COST.

BY GEORGE deLUCENAY LEON, JON D. PAUL, AND LUIS E. RICO

ONE of the most valuable pieces of audio test equipment is the audio sweep generator. Unfortunately, it is not widely used because heretofore only costly professional instruments contained the desirable functions and accuracy: a logarithmic function which eliminates tedious point-by-point frequency plots, a frequency range extending virtually flat to beyond 20 kHz, calibrated attenuators, and high stability. Now, you can build the "Super Sweeper" audio sweep generator and have a laboratory-quality instrument for about what you would expect to pay for a common audio signal generator *without* the sweep function.

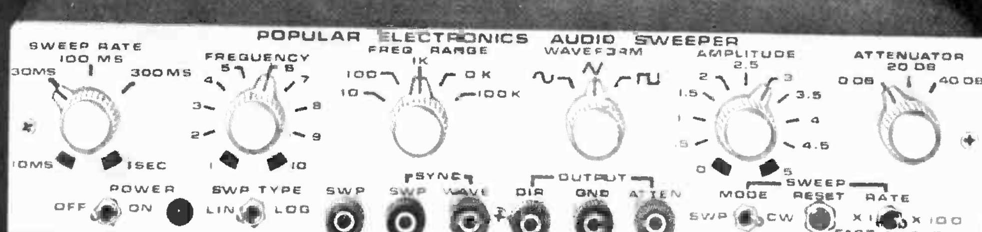
Also, the Super Sweeper sacrifices nothing as a conventional wide-range audio signal generator. Its controls can be set to provide any frequency- and amplitude-adjustable square, sine, or triangle waveform for signal tracing or what have you.

The overall block diagram for the Sweeper is shown in Fig. 1, with the various blocks referring to the schematics of Figs. 2 through 6.

Rather than go through a lengthy stage-by-stage discussion on how the sweeper works, we will be concentrating on calibration procedures and how to use the instrument.

Construction. For ease of assembly, we recommend the use of a PC board. Due to the complexity of the board, the foil pattern is not shown. A foil pattern and component layout (included in the step-by-step instructions) or a complete board are available from the source in the Parts List. Mount the following switches and control potentiometers on the front panel: SWEEP RATE (*R4*), FREQUENCY (*R33*), FREQUENCY RANGE (*S4*), WAVEFORM (*S5*), AMPLITUDE (*R60*), ATTENUATOR (*S7*), SWEEP MODE (*S2*), RESET (*S8*), SWP TYPE (*S3*), and SWEEP RATE (*S1*). Also locate on the front panel the pilot lamp and the six binding posts for the outputs. All other controls are to be mounted on the PC board.

You can use any type of chassis that suits your fancy. A pre-drilled and screened chassis/cabinet is also available, if desired.



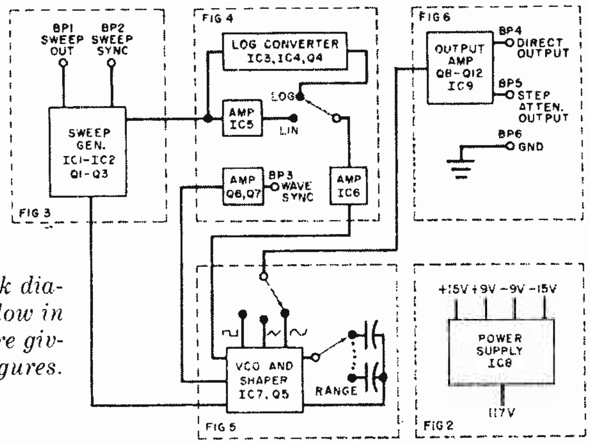


Fig. 1. Overall block diagram showing signal flow in the Sweeper. Blocks are given in detail in other figures.

PARTS LIST

BP1—BP6—Five-way binding post
C1,C12—1-μF, 50-volt Mylar capacitor, 10%
C2—100-μF, 15-volt tantalum capacitor, 10%
C3,C23,C24—0.01-μF, 50-volt disc capacitor
C4,C14—100-pF, 50-volt disc capacitor
C5,C17,C18—0.001-μF, 50-volt disc capacitor
C6,C11—0.1-μF, 50-volt Mylar capacitor, 10%
C7,C8,C19,C20—15-μF, 20-volt tantalum capacitor
C9—910-pF, 100-volt mica capacitor, 10%
C10—0.01-μF, 50-volt Mylar capacitor, 10%
C13—10-μF, 25-volt tantalum capacitor, 10%
C15,C16—1000-μF, 25-volt electrolytic capacitor
C21—150-pF, 50-volt disc capacitor
C22—3.3-pF; 50-volt disc capacitor, 10%
C25—5-pF, 50-volt disc capacitor, 10%
D1—D6,D9—D12—1N914 or 1N4148 signal diode
D7,D8—1N4001 50-volt rectifier diode
F1—¼-A 3AG fuse and holder
IC1, IC3—IC6—741 op-amp IC
IC2, IC9—301A op-amp IC
IC7—Function generator IC (Intersil ICL8038CC)
IC8—Voltage regulator IC (Raytheon RC4194TK)
LED1—Light-emitting diode (Monsanto MV5023)
Q1,Q10—2N4250 transistor
Q2,Q3,Q6,Q7,Q11—2N3642 transistor
Q4—2N2916 dual npn transistor
Q5—MPF-111 n-channel FET (Motorola)
Q8,Q9—2N5210 transistor
Q12—2N3645 transistor
R1—2400-ohm, ¼-watt resistor, 5%
R2,R36—1000-ohm trimmer
R3,R16—5600-ohm, ¼-watt resistor, 5%
R4—500,000-ohm log-taper potentiometer
R5,R53—3900-ohm, ¼-watt resistor, 5%
R6,R41—2000-ohm trimmer
R7—1800-ohm, ¼-watt resistor, 5%
R8,R17,R26,R29,R52—4700-ohm, ¼-watt resistor, 5%
R9,R20,R42—4020-ohm resistor, 1%
R10,R25,R32—5010-ohm resistor, 1%
R11,R12,R13,R61—10,000-ohm, ¼-watt resistor, 5%
R14—3300-ohm, ¼-watt resistor, 5%
R15—8600-ohm, ¼-watt resistor, 5%
R18,R27,R31—5000-ohm trimmer
R19,R56—22,600-ohm resistor, 1%
R21—100-ohm resistor, 1%
R22,R51—82,000-ohm, ¼-watt resistor, 5%
R23—10,000-ohm trimmer
R24—330,000-ohm, ¼-watt resistor, 5%
R28,R30,R35,R37—10,000-ohm resistor, 1%
R33—10,000-ohm linear-taper potentiometer
R34—1100-ohm resistor, 1%
R38,R39—100,000-ohm trimmer
R40—4750-ohm resistor, 1%
R43,R62—15,000-ohm, ¼-watt resistor, 5%
R44,R58—100,000-ohm, ¼-watt resistor, 5%
R45,R49—22,000-ohm, ¼-watt resistor, 5%
R46—33,000-ohm resistor, 1%
R47—120,000-ohm, ¼-watt resistor, 5%
R48—30,000-ohm, ¼-watt resistor, 5%
R50—6800-ohm, ¼-watt resistor, 5%
R54—680-ohm, ½-watt resistor, 5%
R55—71,500-ohm resistor, 1%
R57—68,000-ohm, ¼-watt resistor, 5%
R59—2200-ohm, ¼-watt resistor, 5%
R60—1000-ohm linear-taper potentiometer
R63—39,000-ohm, ¼-watt resistor, 5%
R64—100-ohm, ¼-watt resistor, 5%
R65,R66—1500-ohm, ¼-watt resistor, 5%
R67,R68—51-ohm, ¼-watt resistor, 5%
R69—6200-ohm, ¼-watt resistor, 5%
R70,R72,R73—620-ohm, ¼-watt resistor, 5%
R71—62-ohm, ¼-watt resistor, 5%
R74—560-ohm, ¼-watt resistor, 5%
R75—20,000-ohm resistor, 1%
S1,S2—Dpdt minature toggle switch
S3—Spdt miniature toggle switch
S4—Sp 5-pos. shorting rotary switch
S5,S7—Sp 3-pos. shorting rotary switch
S6—Spst miniature toggle switch
S8—Spst normally open pushbutton switch
T1—Transformer: secondary 12 V at 150 mA
Misc.: Suitable chassis; mounting hardware; wire; solder; etc.

Note: The following are available from MITS, 6328 Linn, N.E., Albuquerque, New Mexico 87108. PC step-by-step instructions including foil pattern and component layouts (free, send self-addressed, stamped #10 envelope); PC board $14.00; complete kit including board, instructions, and enclosure $119.95; complete unit assembled $149.95.

Calibration. Using a VTVM and an oscilloscope, calibrate the Super Sweeper as outlined in the Table. The procedure given will yield a frequency calibration accurate to within 10 percent. If greater accuracy is desired, the range capacitors can be selected or padded to be within 1 percent of the specified value.

An alternate method of calibration is to measure the frequency of the sine-wave CW output of the Sweeper by comparing it with the output of a well-calibrated audio oscillator, using Lissajous figures. Set the Sweeper's FREQUENCY control (*R33*) to 10, its fully CW position, and measure the output frequency for each range. Set the FREQUENCY RANGE switch (*S4*) to the setting that had the greatest error *below* the intended frequency. Then set LIN CAL (*R31*) so that the frequency of this range is accurate to within 1 percent. Now, all of the other ranges will be high in frequency; pad each of these ranges with a capacitor until all outputs are accurate to within 1 percent.

Operation. For use as a fixed-frequency oscillator, set the SWEEP MODE switch to CW and SWP TYPE switch to LIN. The sweep rate controls will now have no effect on the Sweeper's output. When the WAVEFORM switch is set to sine, *BP4* (OUTPUT DIR) will supply a 0 to 5-volt rms sine wave at 600 ohms impedance. In the triangle and square positions of the WAVEFORM switch, the output at *BP4* will have a peak voltage variable between 0 to 7 volts. The *BP5*

SPECIFICATIONS

Modes: CW, linear sweep, log sweep
Waveforms: sine, square, triangle
Range (5 steps): 1 to 100,000 Hz
Response: ±0.1%, to 20,000 Hz; ±0.15%, to 100,000 Hz
Distortion (sine): 1.5%, 10 to 20,000 Hz
Rise time (square): 2 μs
Output voltage: 0-5 V rms sine; 0-7 V peak sine, square, triangle
Attenuator: 0, 20, 40 dB ±1 dB; 600 ohms constant impedance
Sweep time (2 steps): 10 ms to 100 s
Sweep output (ramp): 0 to 5 V, 5000 ohms output impedance
Sync Outputs: 4-V positive pulses, 5000 ohms impedance. Sweep sync pulse starts at end of sweep, returns to zero start of next sweep. Can be used for blanking; 5 μs rise time. Wave sync is square wave, amplitude independent of control settings; 0.5 μs (maximum) rise time.

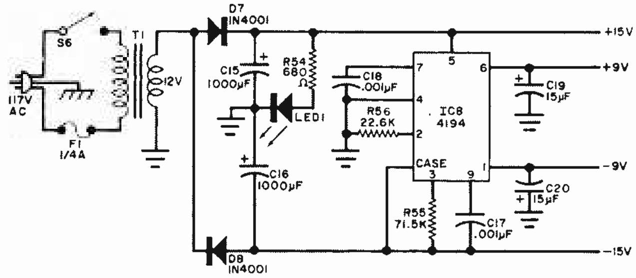


Fig. 2. The power supply has a single IC and a conventional rectifier filter approach.

(OUTPUT ATTEN) binding point will have the same open-circuit output as the direct output when the ATTENUATOR switch is set for 0 dB. In the 20-dB position, the output will be 1/10th of the direct output, while in the 40-dB position it will be down to 1/100th. Loads (or short circuits) on either of these outputs will have no effect on the other.

The output frequency can be set at any point from 1 Hz to 100 kHz by adjusting the FREQ RANGE switch and FREQUENCY control. The SYNC WAVE output (*BP3*) provides a 4-volt, positive-going square wave for sync or counter use.

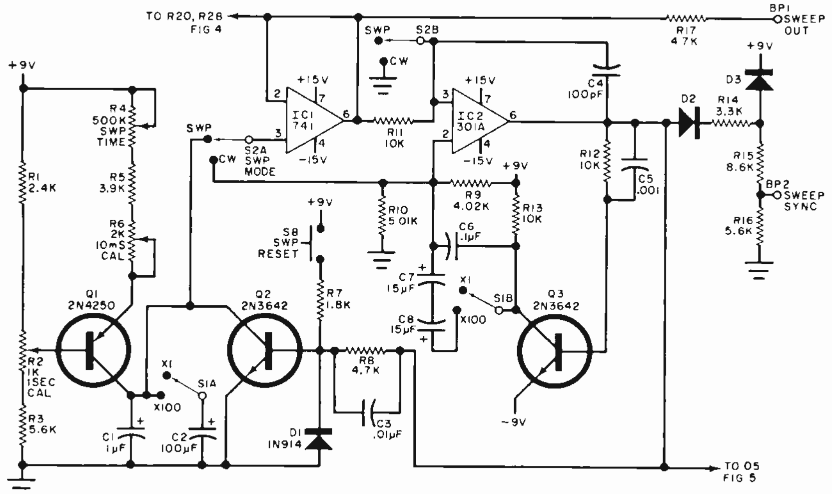


Fig. 3. Linear sweep voltage is generated in Q1 and Q2, available at BP1. Output of IC2 is a pulse which is attenuated and brought out to jack BP2.

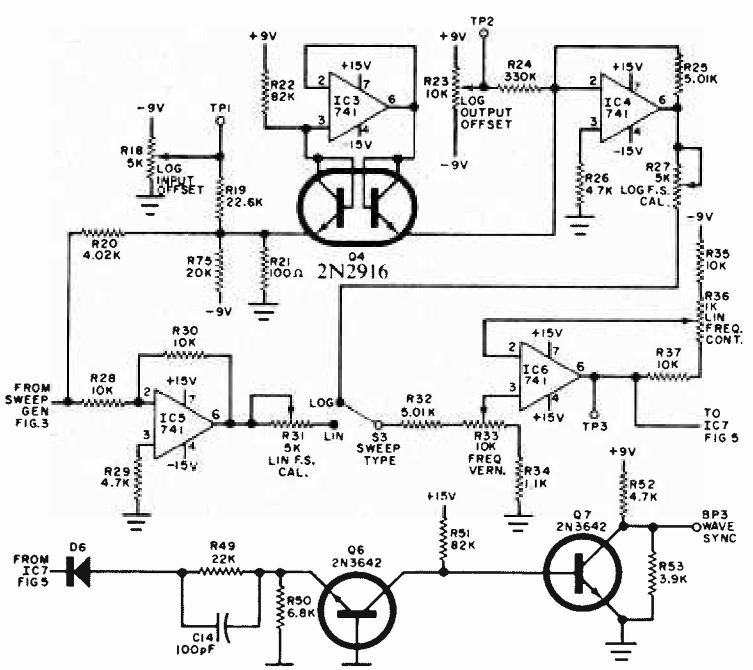


Fig. 4. Converter circuit shapes the linear sweep into logarithmic waveform, since second half of Q4 has current flow proportional to exponent of applied voltage. IC4 converts current to voltage. S3 selects either the linear or logarithmic output.

Refer to Fig. 7 for setting up a sweep display. In the sweep mode, the output voltages and waveforms are the same as they are in the CW mode, but the frequency is swept from nearly zero up to the frequency set by the FREQ RANGE switch and FREQUENCY control. The SWEEP RATE control varies the time for a full sweep from 10 ms to 1 s in the FAST and 1 s to 100 s in the SLOW position. The sweep rate is adjusted to the correct value by observing the output display with an oscilloscope. Sweeping too fast causes the display to smear. Sweeping too slowly will cause a flicker in the display and make observation difficult. The point at which the display smearing occurs depends on the bandwidth of the unit being tested and on the sweep width setting.

The linear sweep allows the frequency of any point on the plot to be read directly, since the frequency starts at zero and changes at a constant rate. The one disadvantage of the linear sweep is that only a narrow region of the audio band is shown in detail. Thus, for a 10-kHz sweep, the bass and midrange (20-1000 Hz) are compressed into the first 10 percent of the sweep, with the remaining 90 percent covering the treble response in detail.

The logarithmic sweep solves this problem by devoting equal area to each band of audio. The rate of frequency change with time increases at a constant pace. The log sweep covers two decades, or about six octaves. Notice that the starting point is not dc (zero frequency), but is offset, since the logarithm of zero is minus infinity.

The ability of the sweeper to go as slowly as 100 seconds/sweep permits plots of systems with narrow bandwidths. A graphic chart plotter or an oscilloscope camera will provide a permanent record of the response curve.

The audiophile can use the 100-second sweep to detect resonant objects in a room by "playing" the

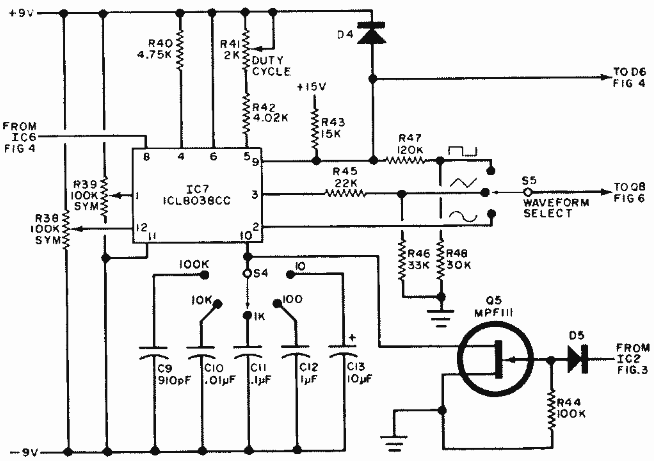


Fig. 5. The vco uses one LSI IC to generate sine, triangle, or square wave output. Chip is reset from the sweep generator. Switch S4 selects the range.

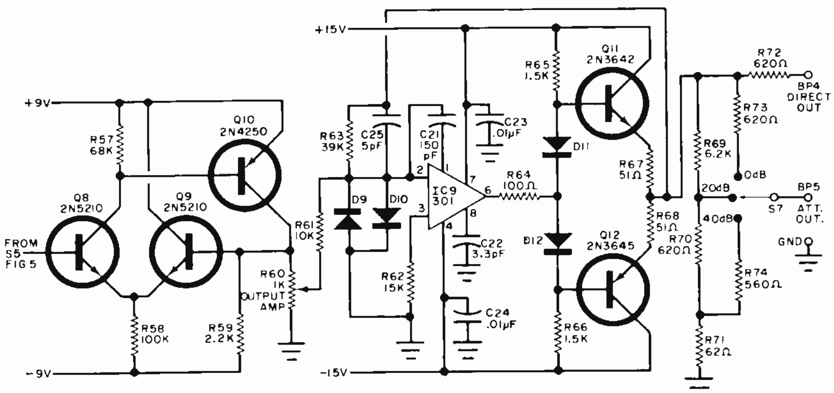


Fig. 6. In output stage, Q8, Q9, and Q10 form unity-gain buffer, whose output is applied to amplitude control R60 and is then amplified by IC9. Transistors Q11 and Q12 provide high output current. A three-step attenuator is made up of R69 to R74 with amount of attenuation selected by S7. Output impedance is 600 ohms.

PROJECT EVALUATION
(Hirsch-Houck Lab Report)

In our tests, the Super Sweeper easily met or surpassed all of its performance specifications. In fact, the rated accuracy and response uniformity of our laboratory-grade test equipment was not adequate to verify the "flatness" of the Sweeper's output. However, it left us with no doubts as to the capabilities of this versatile instrument.

The uniformity of output in the CW mode was within ±0.1 dB from 5 Hz to 100 kHz. Using an oscilloscope, we judge that the output rose to +0.8 dB at 1 Hz, although this involved some guesswork. In the sweep mode, we recorded the output on a General Radio 1521-B Graphic Level Recorder, whose rated flatness is comparable to that of the Sweeper. The chart calibration was not synchronized in frequency with the swept signal, but it showed a total variation of less than 0.25 dB over the full 100-kHz sweep in the linear mode. The logarithmic sweep was almost perfectly flat up to and beyond 60 kHz (on the 100-kHz range); and the output then dropped about 0.5 dB as the sweep continued to 100 kHz. Using less than the maximum sweep capability of the unit, the flatness was generally within 0.1 dB.

The maximum output voltage (sine) was 5.3 volts into an open circuit, dropping to 2.6 volts with a 600-ohm load. The attenuator error at 1000 Hz was −0.1 dB at the 20-dB setting and −1 dB at the 40 dB setting.

The harmonic distortion (sine-wave output) varied with frequency. The lowest reading, 1.0%, was measured at 20 Hz, and the highest, 1.9% to 2.0%, was in the range between 100 and 1000 Hz. Over the audio range, the average distortion was about 1.5%. Of course, one would not use this instrument for making distortion measurements on high-quality amplifiers; it is principally a tool for frequency-response measurements.

The frequency calibrations proved to be surprisingly accurate. The error was typically less than 3% and at only one point did it even approach the 10% figure specified.

The square and triangle waves appeared to be good, judging from a visual examination on an oscilloscope. The square-wave rise time was approximately 2 microseconds.

The logarithmic sweep, a virtually necessity for meaningful audio measurements, is rarely found—even on generators costing many times the price of this one. The lack of logarithmic frequency calibration is a minor inconvenience, but an external marker generator can be mixed with the output of the Sweeper; and once the calibration has been made for a commonly used frequency range (such as up to 20 kHz), it should remain valid for a long time without further checks.

ALIGNMENT PROCEDURE

CONTROL/SETTING	OBSERVE	ADJUST	DESIRED
POWER/on SWP/lin	IC8-5 IC8-case IC8-6 IC8-1	— — — —	+15 ±3 V −15 ±3 V +9 ±0.1 V −9 ±0.1 V
SWP MODE/SWP SWP RATE/X1 SWP RATE/1 sec	SWP out	R2	1-s sweep time (1 Hz)
As above with SWP RATE/10 ms	SWP out	R6	10-ms sweep time (100 Hz)
FREQ RANGE/10K FREQUENCY/10 SWEEP RESET/in	OUTPUT DIR	R36	30-90-Hz output
As above with SWEEP MODE/CW SWEEP RESET/out	OUTPUT DIR	R31	10-kHz output
As above with WAVEFORM/square SWEEP MODE/CW	OUTPUT DIR	R41	Equal time for both states of square wave
As above with WAVEFORM/sine FREQ RANGE/1K FREQUENCY/5	As above	R38 R39 R41	Minimum distortion (on sine wave)
SWP TYPE/LOG SWEEP MODE/SWP SWEEP RATE/X1	TP1	R18	−5 V
SWEEP RATE/1 SEC	TP2	R23	−5 V
As above	TP3	R18	Sweep voltage should drop 10% (middle of trace)
As above with FREQ RANGE/10K FREQUENCY/10 SWEEP RESET/in	OUTPUT DIR	R23	30-70-Hz output
As above with SWEEP MODE/CW SWEEP RESET/out	OUTPUT DIR	R27	10-kHz output

Note: All observation points (Column 2) referenced to ground.

sweeper through his audio system and moving around the room listening for resonances. When using slow sweeps, the SWEEP RESET pushbutton is handy for restarting the sweep before it is completed. Hold down the button to synchronize the sweep manually.

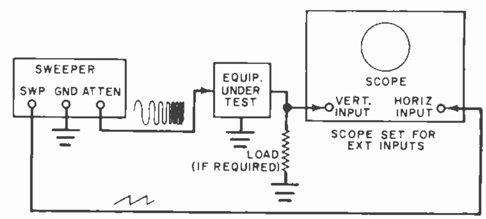


Fig. 7. Setup for hooking sweeper to gear under test and scope.

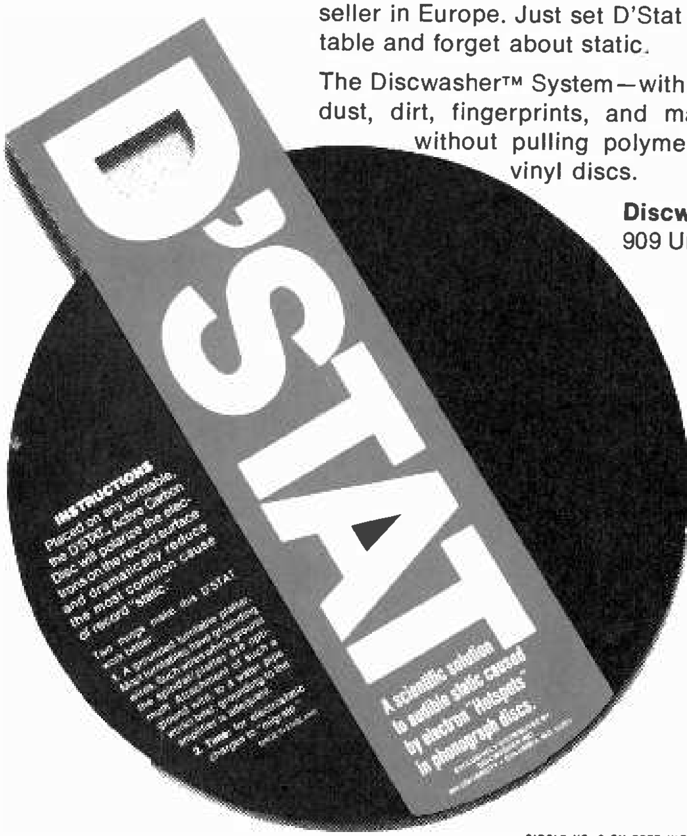
Applications. An audio sweep generator has many applications. For example, it simplifies setting a tape recorder's bias and aligning head azimuth. Line and load regulation and output impedance vs frequency for power supplies can be checked. It can also be used to test room and speaker-enclosure resonance, microphone-element sensitivity, SSB filters, and telecommunications systems, not to mention a host of all-audio applications. ◈

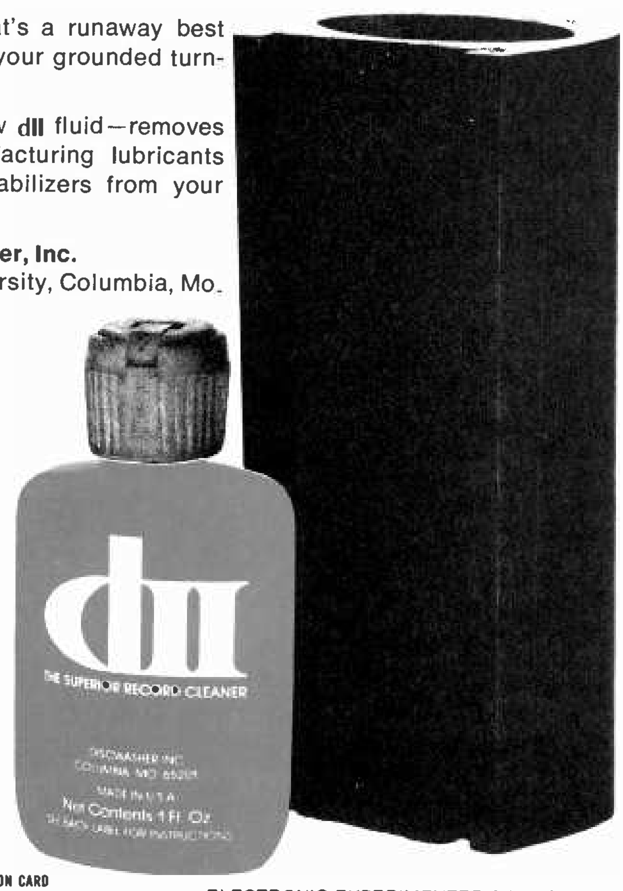
From DISCWASHER

D'STAT. An active carbon disc that's a runaway best seller in Europe. Just set D'Stat on your grounded turntable and forget about static.

The Discwasher™ System—with new **dII** fluid—removes dust, dirt, fingerprints, and manufacturing lubricants without pulling polymer stabilizers from your vinyl discs.

Discwasher, Inc.
909 University, Columbia, Mo.





CIRCLE NO. 9 ON FREE INFORMATION CARD

Build the Low-cost TRANSISTOR OSCI -TESTER

A-f & r-f tests, checks junctions, identifies type

BY JOHN F. HOLLABAUGH

THE Transistor Osci-Tester is more than just a simple go/no-go checker. Of course, it does indicate whether the junctions of a transistor are good, but it also determines if the transistor will oscillate at about 5 kHz for audio functions and whether it will provide gain at about 3 MHz for r-f applications. The latter test eliminates the measurement of gain, junction capacitance, and leakage. If the transistor will oscillate at r-f, it must be in good shape. The tester also shows transistor type.

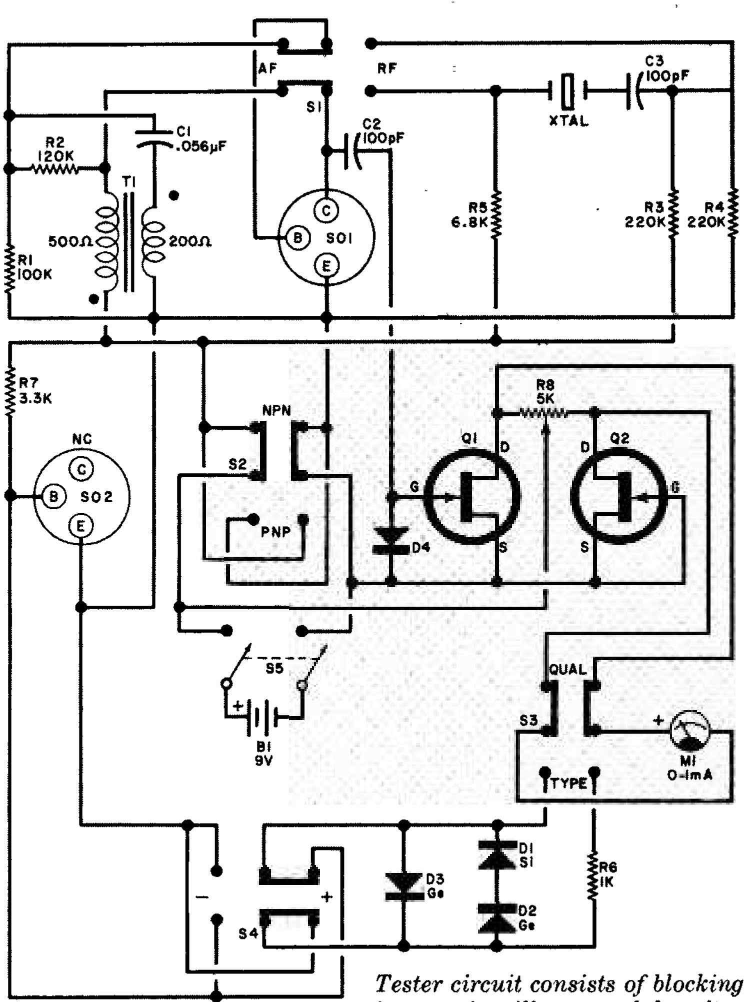
Circuit Operation. The a-f test is made by including the unknown transistor in a blocking oscillator circuit consisting of *T1*, *C1*, *R1*, and *R2*. Resistors *R1* and *R2* determine the operating bias of the unknown to give a partial indication of the operating frequency.

The oscillator output is passed through *C2* to drive a dc voltmeter consisting of *D4*, *Q1*, *Q2*, and *M1*. The quiescent (zero) current of *Q1* is balanced by the channel resistance of *Q2*, which has zero bias and matches the zero bias of *Q1*. Diode *D4* rectifies the oscillator output, producing the negative voltage required to drive *Q1*. The setting of potentiometer *R5* balances the relative quiescent voltage drops across the channels of *Q1* and *Q2*.

The r-f test is made by connecting the unknown transistor in a Pierce oscillator consisting of a crystal (channel 5 used in prototype), *R5*, *R3*, *R4*, and *C3*. In this circuit, the base of the unknown transistor is driven by the collector output through the crystal. This produces positive feedback at the crystal frequency. If you want to check the harmonics of the crystal frequency, loosely couple the r-f oscillator output to the receiver antenna.

Increasing *R3* to 560,000 ohms may improve r-f oscillation.

The third portion of the tester converts it into a one-volt dc meter, which is used to measure the forward voltage drop (barrier voltage) across the forward-biased base-emitter junction of the unknown transistor. Voltage of the correct polarity is determined by the setting of *S2* which applies this current through *R7*. The meter then indicates the approximate 0.3-volt drop of a germanium junction or the approximately 0.7-volt drop of a silicon junction. Diodes *D1* and *D2* are a silicon and a germanium connected in series to limit the open-circuit voltage to about 1 volt when the transistor under test is disconnected.



Tester circuit consists of blocking oscillator, r-f oscillator, and dc voltmeter.

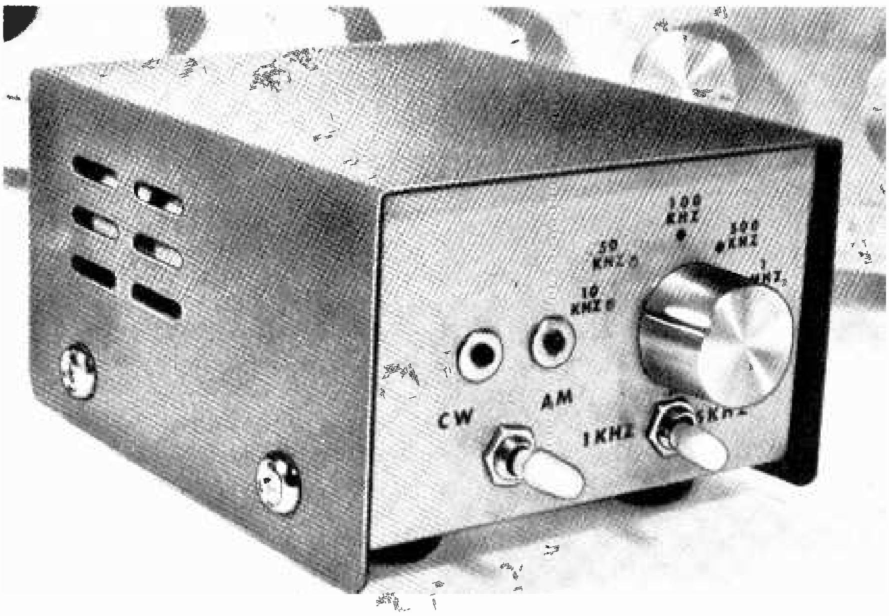
PARTS LIST

B1—9-volt battery
C1—0.056-µF capacitor
C2,C3—100-pF capacitor
D1—Silicon diode
D2—D4—Germanium diode
M1—0-1-mA meter
Q1,Q2—HEP801 transistor
R1—100,000-ohm resistor
R2—120,000-ohm resistor
R3,R4—220,000-ohm resistor
R5—6800-ohm resistor
R6—1000-ohm resistor
R7—3300-ohm resistor
R8—5000-ohm potentiometer
S1—S4—Dpdt switch
S5—Dpst switch
SO1, SO2—Transistor socket
T1—500:200-ohm transistor driver transformer (Radio Shack 273-1581 or equiv.)
Xtal.—CB crystal (channel 5 used in prototype)
Misc.: Battery holder and connector; mounting hardware; etc.

Construction. The circuit can be assembled in any way. The prototype was built on a piece of perf board and put in a small plastic case. The meter and necessary switches were installed on the front panel.

Operation. Connect the unknown transistor to *SO1*. Place *S1* in the AF position, *S3* in the QUAL position, and turn on the power. Operate *S2* for a meter indication. The position of *S2* will indicate the transistor type. If you get a meter indication, the transistor will oscillate at audio. Switching *S1* to RF will show whether the transistor will operate at r-f.

With the transistor in *SO2* and *S3* in the TYPE position, operating *S4* will show either a low-scale indication for germanium types or a high-scale indication for silicon types. ◈



BUILD A

DELUXE FREQUENCY STANDARD

One crystal and four TTL devices produce seven crystal-controlled calibration frequencies.

BY JOE A. ROLF, K5JOK

FOR the shortwave listener, a precision frequency standard is an important accessory for locating the exact frequency of a hard-to-get station. Hams need a calibration oscillator to take to the guesswork out of determining where they are in the band. And the electronics experimenter or technician finds uses for a frequency standard, from calibrating a receiver to using it as a precision timing source. A useful instrument!

The deluxe frequency standard described here can be used for all of these purposes since it provides highly accurate AM or CW signals at 1, 5, 10, 50, 100, and 500 kHz and at 1-MHz intervals to well above 50 MHz. It is small in size, battery operated, and easy to build.

The compactness, versatility, and accuracy of the standard are a result of the use of readily available, economy priced TTL integrated circuits. The standard (Fig. 1) uses one 7404 hex-inverter and three 7490 decade dividers as a master 1-MHz oscillator that can be calibrated to WWV. A divider chain generates six sub-frequencies.

The master-oscillator, digital-divider approach to a frequency standard provides extreme accuracy on all

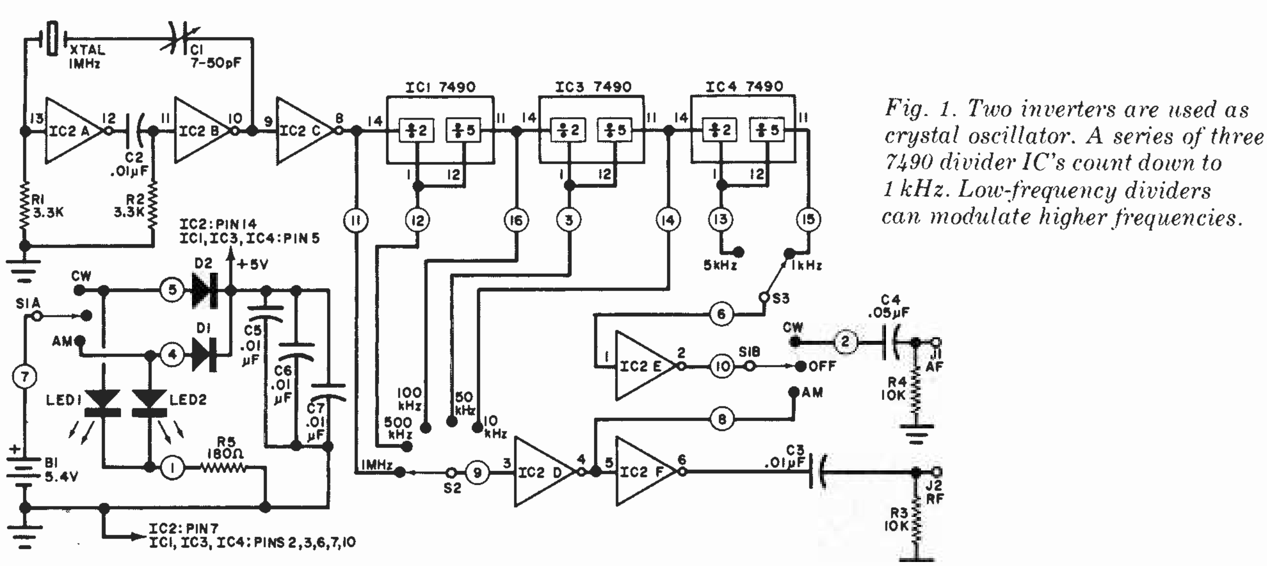


Fig. 1. Two inverters are used as crystal oscillator. A series of three 7490 divider IC's count down to 1 kHz. Low-frequency dividers can modulate higher frequencies.

PARTS LIST

B1—5.4-volt mercury battery (Mallory TR-134 or equiv.)
C1—7-50 pF midget trimmer (Arco #403, Calectro A1-246, or equiv.)
C2,C3,C5,C6,C7—0.01-µF, 25-volt disc capacitor
C4—0.05-µF, 12-volt disc capacitor
D1,D2—0.2-A, 25-PRV diode (1N4444, 1N4450, 1N914, or equiv.)
IC1,IC3,IC4—7490 decade divider
IC2—7404 hex inverter
J1,J2—Miniature pin jack (GC 33-216, Smith 223, or equiv.)
LED1,LED2—Miniature light-emitting diode
R1,R2—3300-ohm, ¼-watt resistor
R3,R4—10,000-ohm, ¼-watt resistor
R5—180-ohm, ¼-watt resistor
S1—Miniature dpdt center-off toggle switch (Alco MST-205N, Calectro E2-129, or equiv.)
S2—Miniature 5-pos. non-shorting rotary switch (Calectro E2-163 or equiv.)
S3—Miniature spdt toggle switch (Alco MTS-115D, Calectro E2-122, or equiv.)
XTAL—1-MHz crystal (PR Type Z-9 HC6/U holder or equiv.)
Misc: Suitable cabinet; rubber feet (4); spacers; mounting hardware; etc.

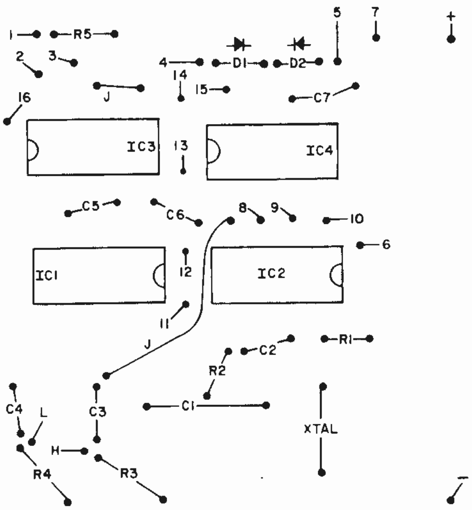
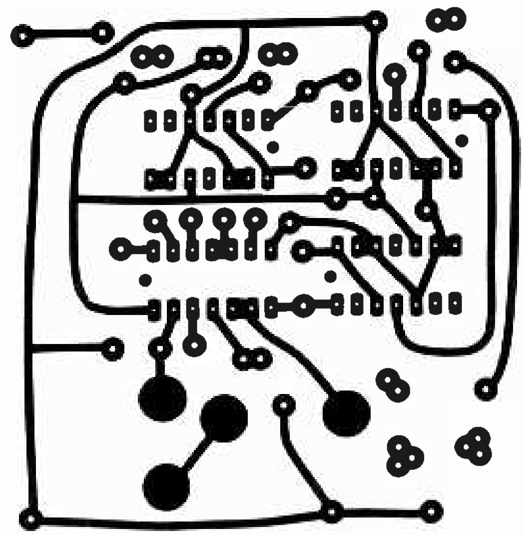


Fig. 2. Actual-size etching guide (right). On component-placement guide (left) numbered points connect to panel controls.

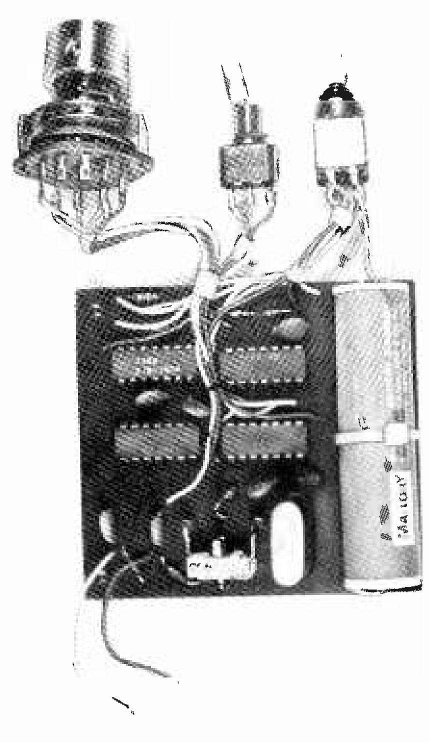


frequencies since digital dividers always divide by a fixed number. For intance, if the 1-MHz master oscillator is tuned to within 10 Hz of WWV at 20 MHz, it will be within 0.5 Hz at 1 MHz and 0.0005 Hz at 1 kHz. This represents an accuracy of 0.00005%.

Three of the inverters in *IC2* form the master oscillator; *IC2A* and *IC2B* generate a crystal-controlled 1-MHz signal that can be precisely tuned by *C1*, and *IC2C* provides isolation. Three 7490 IC's (*IC1, IC3,* and *IC4*) divide by 2 and then by 5 to provide the subfrequencies. Switch *S2* selects the proper output points in the divider chain and feeds the signal to two inverters for shaping and output.

All outputs above 10 kHz can be modulated at 5 kHz or 1 kHz as selected by *S1* and *S2*. Switch *S3* selects the 1- or 5-kHz outputs, while *S1* selects either *J1* for an audio output or the input of *IC2F* to modulate the high-frequency output.

Switch *S1* is a center-off toggle switch that selects either the AM or CW mode. Mode indication is given by two LED indicators located on the front panel. Diodes *D1* and *D2* serve to isolate the LED's from one another in the selector switch circuit.



Assembled board module with front-panel switches wired in; two wires at bottom go to input and output jacks on rear panel. Battery held by cable tie.

Construction. Printed-circuit board construction should be used for ease of wiring and compactness. Details on the board are shown in Fig. 2. The photos show the board assembly and the completed unit.

Trimmer *C1* is mounted by soldering two small perf-board terminals to the board and then soldering the capacitor terminals to these so that the adjusting screw is located just behind the rear edge of the board. A hole with grommet in the rear panel will provide access to this control.

Carefully solder the crystal in place and trim the pins protruding below the board. The battery is secured with a small cable tie and pad and carefully soldered to leads from the board. Since the calibrator will normally be used only for short intervals, battery replacement should be infrequent. Use ⅝" spacers to mount the board to the case.

Put the controls and LED's on the front panel and *J1* and *J2* on the rear apron. Use plastic cement to hold the LED's in place. Drill a hole for access to *C1*. Be sure to clip all unused terminals of *S2* flush with the switch wafer. For appearance and neatness, bundle wires from the PC board with small wraps of masking tape.

Operation. Put *S1* in the AM and then the CW positions, and note that the appropriate LED illuminates. Set the mode switch to CW and the output selector switch to 1 MHz. Using a shortwave receiver and a short piece of wire connected to *J2*, determine that there is a signal from the calibrator every 1 MHz on the receiver. Place *S3* to either 1 or 5 kHz and set the mode switch to AM. The signal should now be tone modulated at 1 or 5 kHz.

To calibrate, tune to WWV at either 10, 15, or 20 MHz and set the calibrator for a 1-MHz CW output. Carefuly adjust *C1* with an insulated screwdriver to zero beat with WWV. As you approach the zero beat, tune very carefully to get the best possible calibration. By listening carefully or observing the receiver's S meter, you can set the standard within a few hertz.

A short piece of wire connected to the output terminal is sufficient to provide strong marker signals through 50 MHz. You can locate 1-MHz points by first turning the selector switch to that position and, by stepping down successively, identify 500-, 100-, 50-, or 10-kHz points as either a carrier or a tone-modulated signal. For a 1- or 5-kHz output, connect a short piece of wire to *J1* and turn the mode switch to CW. Major calibration points can be taken from *J2* simultaneously by setting the selector switch to the desired position. ◈

The Easy Way to Make PC Boards

THE PHOTOPOSITIVE METHOD

SIMPLIFIES PC CONSTRUCTION FOR HOBBYISTS AND EXPERIMENTERS

BY WILLIAM T. ROUBAL

FABRICATING printed-circuit boards, believe it or not, is not a fine art that can be mastered by only a talented few. Anyone who isn't "all thumbs" and has patience can master the techniques needed to turn out commercial-quality PC boards. The trick lies in how you go about it and the medium in which you choose to work.

For all but the most basic, least-detailed PC layouts, the photosensitive process is best to use. There are basically two types of photosensitive resists available. The most commonly used—because it was developed first and received most of the attention—is the negative-type photoresist. The other type, much more convenient for the hobbyist and experimenter to use, is called positive photoresist.

All negative-type photoresists suffer from one inherent drawback. Before you can use them, you must first prepare a negative from your positive artwork. Hence, you are faced with double the work—unless you elect to use a reversing film—because the positive artwork must be converted into a negative if it is to be usable. Additionally, because of the nature of the process, any alterations or corrections must normally be made on your positive, which means that you must make a new negative.

Alternatively, you can use positive-type photoresists and cut your work in half by using the positive artwork directly. What you see in the original artwork is what you'll get when you etch away the copper. Realistically, then, this is the easier way to work.

Positive photoresists differ chemically from negative types. With positive resists, the portions exposed to light are dissolved away during the development process. Another difference lies with the developer itself. Volatile hydrocarbon solvents used with negative resists are expensive and often hard to keep uncontaminated. By contrast, a dilute solution of ordinary household lye, or caustic soda, in water is all you'll need as a developer for the positive resist.

The Artwork. The best way (actually the only realistic way for multi-IC and other finely detailed PC layouts) of preparing your positive artwork is to use sheet Mylar or acetate film and any of the various dry-transfer and/or stick-on patterns available. The film used should be between 0.002 and 0.004 inch thick.

It is imperative that all drafting aids be opaque to ultraviolet (UV) light. While electronics-type materials invariably meet this requirement, not all brands of dry-transfer materials will pass the test. Prestape, Chart Pak, and

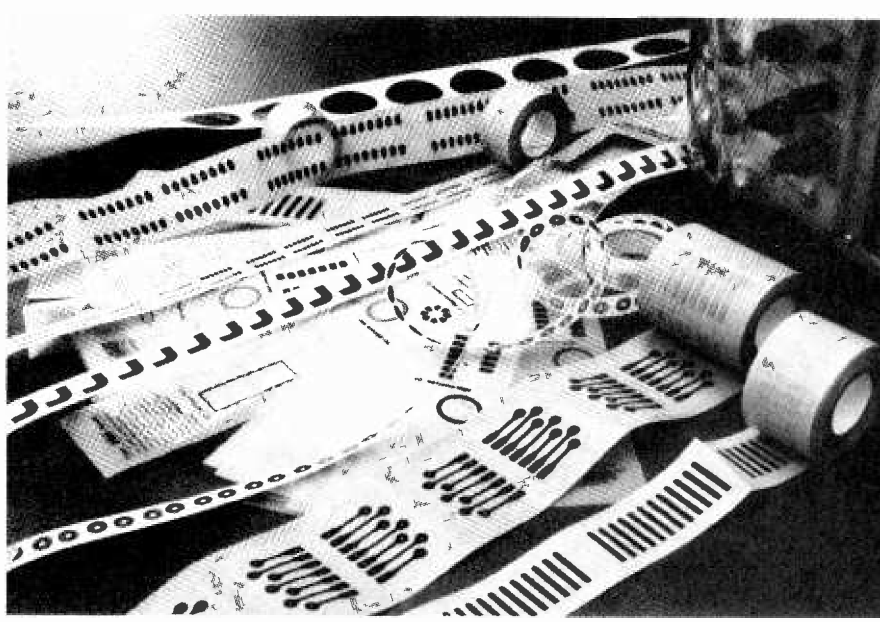


Fig. 1. Opaque transfers and stick-ons are used for making photopositive artwork.

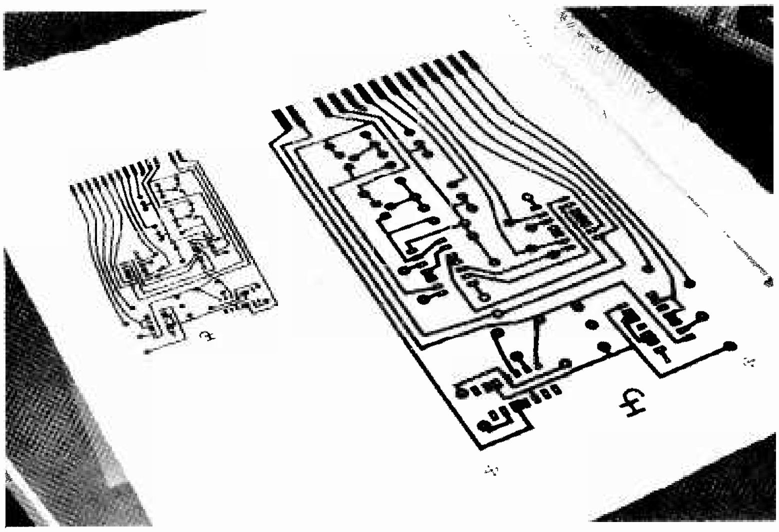


Fig. 2. Twice-size artwork and final photopositive. The final of photopositive is made from lithographic film. Procedure requires use of darkroom and enlarger.

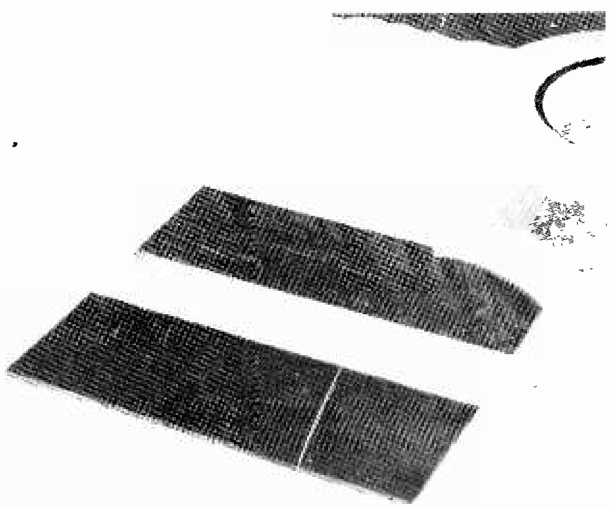


Fig. 3. First step in coating copper by spinning is to adhere masking tape on back.

Para Type are quite satisfactory, while Cello-Tak is not opaque enough.

Representative examples of the drafting aids currently on the market are shown in Fig. 1. Commonly available patterns include various-size solder pad "donuts" and "tear drops," edge connectors, right-angle L's and T's, circles and dots, three- and four-lead transistor pads, and all IC pad configurations. Some companies produce dry-transfer patterns, while most make pressure-sensitive stick-ons. The only thing not shown in Fig. 1 is the opaque crepe tape, also available in various widths, that you'll need for interconnecting solder pads.

You can obtain the drafting aids in various scales, the most common being 1:1 and 2:1. For most jobs, 1:1 will meet your requirements. If you are working on a very detailed and crowded layout, you might be better off working twice up (2:1 scale) to simplify the job and give you better control over the work. Working twice up means that you'll have to reduce, by photographic means, your artwork to the proper size; but the extra step is worth it if there's any chance of introducing errors when using the 1:1 aids. Of course, when you work in any scale larger than 1:1, select interconnecting crepe tape to suit the scale. An example of 2:1 artwork is shown in Fig. 2.

Preparing the PC Blank. The copper-clad board that you will use for making a PC board is called a "blank." The copper must be perfectly clean

STEP-AND-REPEAT ARTWORK

When you must make two or more of the same PC board, you can save considerable time if you make a multiple positive for exposing several blanks at once. The least time-consuming method is a "step-and-repeat" process by which your original artwork is duplicated two or more times with the aid of "Trans-O-Paque" (TOP) film, Type G-2.

The TOP film is sensitive to UV light; so, you can safely work in a dimly (incandescent) lighted area. A 50-sheet package of 8" × 10" TOP film costs $22.50 ($33.75 for 10" × 12" size). Order it from one of the following Dynachem Corp. outlets:

Far West: P.O. Box 12047, Santa Ana, California 92711

Midwest: 449 Fullerton Avenue, Elmhurst, Illinois 60126

New England: 22 B Street, Burlington, Massachusetts 01803

Southeast (NYC to Fla.): 234 Dominion Rd., Vienna, Virginia 22180

You will need a large vacuum-type exposure frame. Carefully measure your original artwork, add about one-eighth inch to the length and width, and use the figures just obtained to cut an opening of the same size in the center of a sheet of aluminum foil. The foil must be large enough to permit only the portion of the film that shows through the cutout to be seen, no matter where on the TOP sheet the mask is placed.

Tape the mask to the inside glass cover of the exposure frame and position and tape your original artwork into the opening. Start at one corner with a UV lamp 8 inches away and expose the film through the artwork for about 10 minutes or until all visible yellow areas become transparently clear. Turn off the lamp and slide the film to an adjacent unexposed portion and expose. Repeat until all areas of the film have been used up.

Place the exposed film in a large container along with an open jar of 28 percent ammonia (get from a drug store). Put a cover on the container and develop the film in the ammonia fumes for 20 to 30 minutes in the dark.

When fully developed, the duplicated positive will be UV-opaque amber color. Allow the ammonia fumes to completely dissipate before taking the film into an area where you have your stock of unexposed TOP film.

Your step-and-repeat positive can be used as one large sheet if you have a PC blank large enough to accommodate it. Cut up, it can be used with several medium- and small-size blanks.

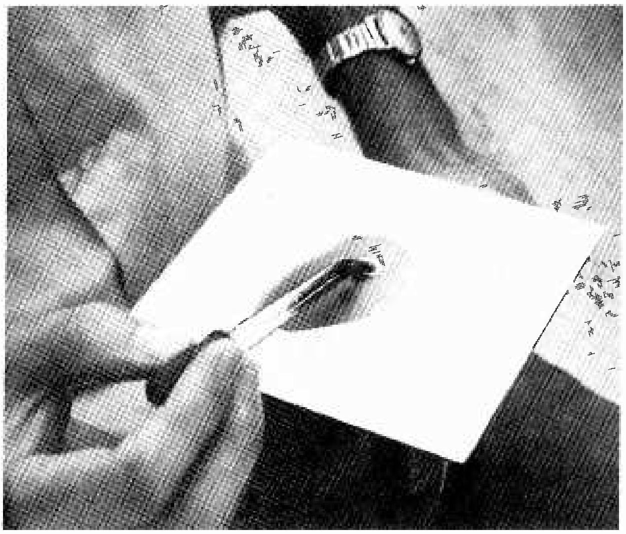


Fig. 4. Once board has been attached to the spinner, pour on AZ-111 positive photoresist.

before you attempt to apply the photoresist. Any dirt or oil will prevent the resist from adhering to the copper.

First, cut the blank to shape, allowing about a quarter-inch extra in length and width. Do *not* make any cutouts that will appear in your finished board at this time. Deburr all cut edges with a medium or fine file. Then clean the copper by light scrubbing with scouring powder and a wet cloth. Thoroughly rinse the blank under running water and remove all traces of grit and immediately blot dry with absorbent paper towels or a lint-free cloth. If the blank is allowed to air-dry, the copper will quickly tarnish. From now on, handle the board only by its edges.

Type AZ-111 positive photoresist is easiest to apply by the spinning method. Place the blank, copper side down, on a lint-free cloth and affix to its unclad surface a wide strip of masking tape (adhesive on both sides or lay one-side tape back on itself) as shown in Fig. 3. Now, centrally position the rubber disc of an electric drill sanding attachment over the blank and press it home. Check to see that the tape is firmly bonded to both the blank and the rubber disc. Then chuck the disc/blank assembly in a variable-speed electric hand drill.

Working in a dimly lighted (make certain that the light is from an incandescent lamp—*not* a fluorescent fixture), up-end the drill assembly and place a few drops or a small puddle of photoresist onto the center of the copper surface (Fig. 4). The amount of resist to use will be governed by the size of the blank. With a little experience, you will quickly learn how much to use for any size blank. Now, quickly brush the resist out toward the blank's edges with a clean artist's brush and up-end the drill assembly over a newspaper-lined wastebasket. Be sure that the blank is several inches below the open top of the wastebasket; then turn on the drill for 5-8 seconds at a speed of 300-400 rpm (Fig. 5).

Working very carefully so as not to touch or disturb the resist coating, remove the blank from the rubber disc, while still in the darkened area. Place the treated blank in a lightproof, dustproof box or cupboard to dry. Since the resist has a strong odor of solvent, it is best to set the box under a kitchen hood with the exhaust fan on until most of the solvent has dissipated. The blank can also be force-dried by placing it in an oven set at 160° F with the door open—don't forget to keep the room in darkness—for 10 to 15 minutes.

Exposing & Developing Blank. When the PC blank is dry and as solvent-free as possible, it's ready to be developed. One test you can make to determine if the blank is ready is to sniff it at close range; if you detect only a faint odor of solvent, it's ready. If you mistrust your olfactory sense, lightly press the tip of a finger against the coating at one corner; any tackiness at all indicates that the board isn't ready.

Once you have satisfied yourself that the blank is ready, you can proceed to expose it as follows: Position the positive over the resist-coated surface of the blank. Next, sandwich the

PC BOARD CONSTRUCTION

Step

A1. Cut and clean PC board
A2. Coat copper with photoresist (photosensitize)
A3. Dry
B1. Prepare artwork (positive)
B2. Prepare multiple positive (step-and-repeat, optional)
C. Place sensitized board and positive in exposure frame and expose to ultra-violet light
D. Develop photoresist
E. Etch copper
F. Drill holes
G. Plate copper (optional)

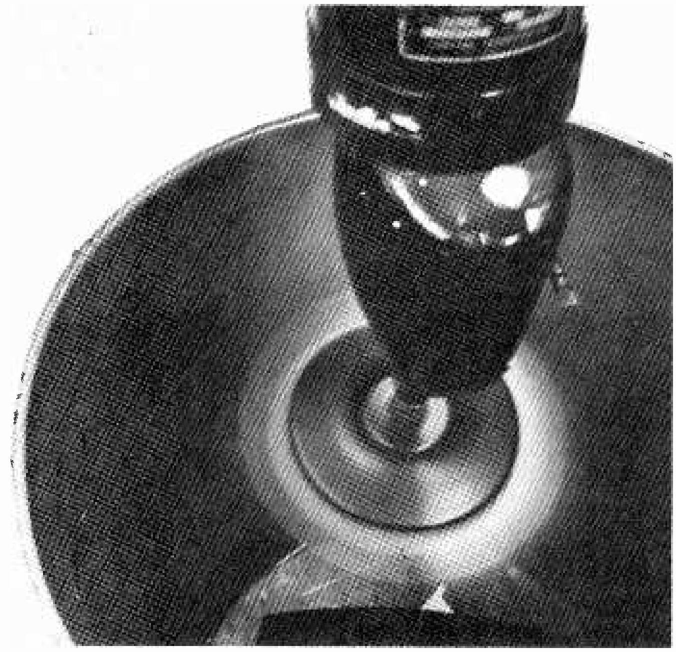


Fig. 5. Thin coat of photoresist is deposited on copper by spinning at moderate speed.

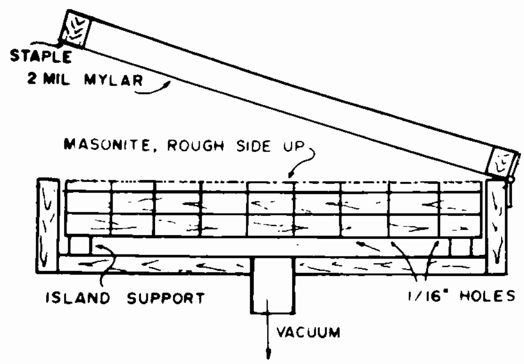


Fig. 6. Building a vacuum exposure frame. It is easy to construct from readily available materials.

assembly between the glass and pressure plate of the exposure frame. (See Fig. 6 for details on how to make a professional-quality vacuum-type frame.) Then expose the positive-masked resist to a 275-watt UV sunlamp for 8 to 10 minutes at a distance of 14 to 18 inches. If you already have a fluorescent UV lamp, feel free to use it. In any case, you'll probably have to experiment a little to determine the optimum exposure time and distance for your setup. Although it is difficult to over-expose the resist, too short an exposure will prevent the resist from dissolving away when you attempt to develop the exposed blank.

Since we've introduced the vacuum-type exposure frame in Fig. 6, let's go into a little more detail. The frame itself is made from ordinary pine lumber. The lumber need not be fancily painted; a single coat of flat black paint will do nicely. Nor are any fancy construction techniques required during assembly. You can make the frame as small or as large as you want (a practical size is 12 inches long, 8 inches wide, and about 2 inches, or less, deep). Don't forget to drill the 1/16-inch holes through the platform as indicated; space them on grid centers about 2 inches apart. And don't forget the island supports. The vacuum tube can be seamless plastic tubing of a size that will provide a friction—not binding—fit for your vacuum cleaner. A foam rubber gasket around the drop lid will increase vacuum efficiency.

The main function of the exposure frame in PC work is to provide a rock-steady system for holding the exposure positive and sensitized blank together and properly registered during the entire exposing time. The vacuum feature is a convenience which comes in handy when you're working with a warped PC blank or an exposure positive that insists upon curling up.

While your blank is being exposed is a good time to mix up a batch of developer. Do this by dissolving about three teaspoonfuls of lye in a quart of water in a shallow Pyrex or enameled tray. If you use an enameled tray, make certain that there are no cracked or chipped areas of the enamel coating. (Warning. *Use only household lye that is free from metal particles.*)

As soon as the blank is completely exposed, immerse it in the developer solution, resist side up. Rock the tray back and forth to agitate the developer and speed up the developing process. The resist will turn purple and the exposed portions will slowly wash away. (Note: If the caustic solution is too weak, development time will be prolonged, or the development will be impossible, indicating that you must increase the concentration of caustic in the bath. Again, experiment until you know the right proportions of lye to water you will need for any given job.)

When all exposed areas on the PC blank are free of resist, remove the blank from the developing bath. Don't just reach in to retrieve the blank; use rubber gloves or plastic tongs. Rinse the blank under gently running water. Then pat the exposed blank dry with absorbent paper towels or a soft cloth.

Etching the PC Board. Submerge the board, copper side up, in the etching solution of your choice. Most people use syrupy ferric chloride for etching. This chemical is fast-acting but requires that the board be removed from the bath several times during etching to check the progress of the chemical action, mainly because ferric chloride is very dark and so dense that it is opaque. You might consider using ammonium persulfate crystals and water, with just a "pinch" of mercuric chloride as a catalyst. (*WARNING: Mercuric chloride is highly poisonous; handle it with extreme care.*)

During the etching process the bath is aerated vigorously by forcing air up through holes drilled in a false bottom of a Plexiglas etching tank; etching is complete in 2-5 minutes. If you don't have a source of compressed air, an aquarium aerator does a good job.

The chemical action is speeded up if the etchant starts off warm. To warm ferric chloride, simply immerse its stoppered container in hot water. The amonium persulfate solution is even easier to warm; just add the crystals and catalyst to hot water (about 150° F).

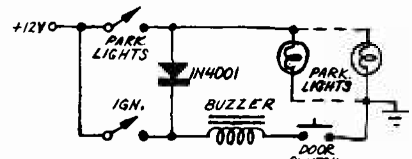
When the etchant has done its work, use rubber gloves or plastic tongs to remove the PC board from the tray in which it was etched and rinse it off under running water. Dry the board. Then remove the remaining resist with a soft cloth dipped into acetone or with very fine steel wool. Trim the board to its exact finished size and drill the component mounting holes.

You might consider plating the copper foil after the PC board is trimmed and drilled, using electroless tin solution that requires only one step and does not contain cyanide. Plated copper resists tarnish and corrosion and, more important, serves as a "wetting" agent that greatly facilitates soldering.

Once you've worked with positive-resist techniques, it's almost guaranteed that you'll never go back to the negative-resist technique that requires almost double the work to obtain the same results. You'll also save money by using inexpensive household chemicals that keep for a long time and do not require special storage. ◈

AUTO LIGHTS WARNING BUZZER

Leaving your parking lights on all night can be almost as hard on the car battery as the headlamps. This simple warning system uses the ignition key/door buzzer. Only a diode is required. Install the 50-PIV diode as shown. If the lights are left on when the ignition switch is open, the diode turns the



warning buzzer on. Since most headlight/parking light switches are ganged, you will be alerted when the headlights and/or parking lights are left on.

—Paul Reckling

• SMALL SIZE • LOW POWER • LOW DISTORTION
AUDIO POWER AMPLIFIER

BY MICHAEL S. ROBBINS

OUTPUT-transformer-less (OTL) audio amplifiers are almost as old as as audio power transistors. But if an audio amplifier is also output-capacitor-less (OCL), several advantages are gained. Presented here is an excellent OTL/OCL hi-fi amplifier designed to deliver 3 to 5 watts rms output power into an 8-ohm load. Its frequency response is a flat ±0.5 dB from 10 Hz to 20,000 Hz, and its total harmonic distortion, measured at a 1-watt output level, is less than 0.2 percent. Input sensitivity is 150 mV rms for full output.

State-of-the-art IC's are employed in the amplifier, reducing the outboard components required to a bare minimum. The whole amplifier, minus the power transformer, fits neatly on a 3″ × 2″ PC board.

Theory of Operation. A typical OTL stage is shown in Fig. 1A. The transistors and current-limiting resistors form a voltage divider across the power-supply voltage. This results in a dc potential that is equal to half the power-supply voltage that appears between the output and ground. To prevent this voltage from appearing across the speaker, a very large coupling capacitor is usually used as a dc blocker. Detrimental side effects of this capacitor include decreased low-frequency response, phase shift, etc.

One way to eliminate the output capacitor is shown in Fig. 1B. This method is called "split-supply" and is often used in high-power amplifiers. A big disadvantage of the circuits shown in Figs. 1A and 1B is that the audio output potential (peak) across the speaker is limited to half the supply voltage. Therefore, the output power of these circuits can be no greater than a fourth of the output power of the circuit shown in Fig. 1C.

The Fig. 1C circuit illustrates a simplified bridge-type OCL amplifier. Transistor *Q1* is a phase splitter, used to drive each half of the bridge amplifier. Since *Q2* and *Q5* conduct only when *Q3* and *Q4* are cut off, and vice versa, the maximum speaker output voltage is practically equal to the supply voltage.

The circuits shown in Fig. 1 are provided for illustration purposes only and should not be used for actual amplifiers. A number of components have been omitted, including those required for proper biasing. A practical amplifier schematic diagram—our hi-fi OTL/OCL amplifier—is shown in Fig. 2. The transistors and most of their allied components have been replaced by a pair of power op amps. The differential input stage of each IC provides both an inverting and a non-inverting input. By connecting the inverting input of one IC to the non-inverting input of the other, no phase splitter is needed.

Capacitor *C2* limits the high-frequency response of the amplifier to the audio range. Without *C2*, the amplifier is flat to about 200,000 Hz and could draw excessive current when

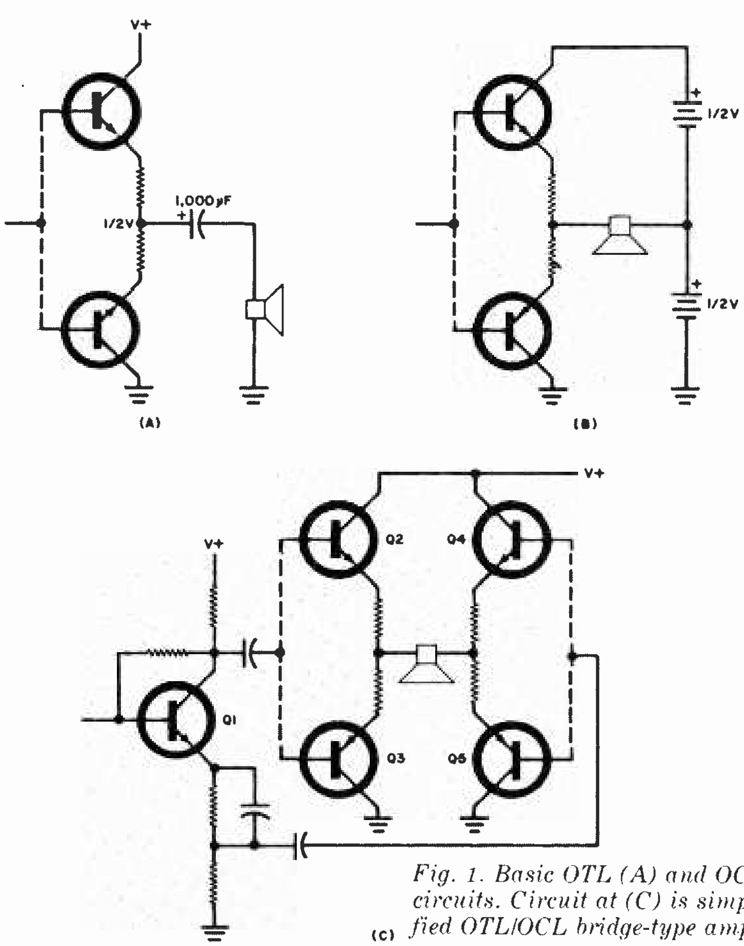


Fig. 1. Basic OTL (A) and OCL (B) circuits. Circuit at (C) is simplified OTL/OCL bridge-type amplifier.

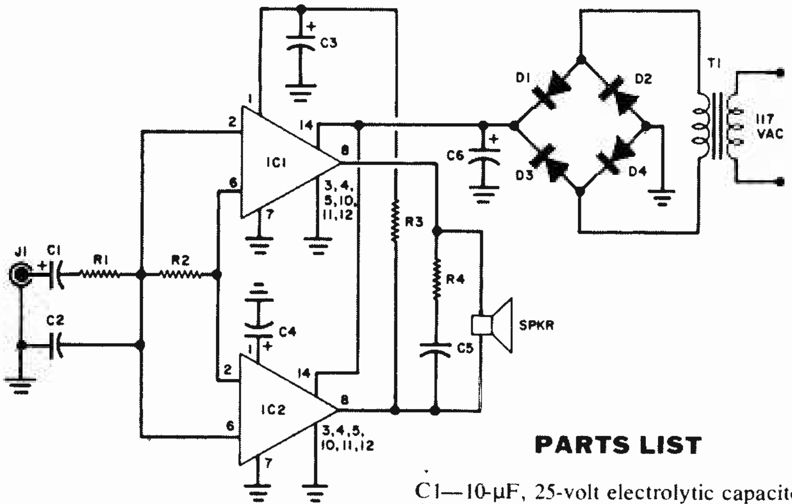
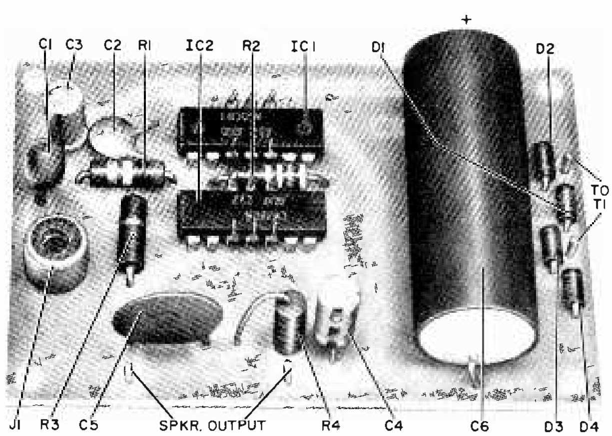


Fig 2. Complete schematic diagram of a practical OTL/OCL hi-fi amplifier using IC's.

PARTS LIST

C1—10-μF, 25-volt electrolytic capacitor
C2—100-pF ceramic capacitor
C3, C4—10-μF, 16-volt electrolytic capacitor
C5—0.1-μF, 25-volt ceramic capacitor
C6—500-μF, 25-volt electroytic capacitor
D1—D4—1N4001 rectifier diode
IC1, IC2—LM380N integrated circuit (National)
J1—Phono jack
R1—100,000-ohm, ½-watt resistor
R2—470,000-ohm, ½-watt resistor
R3—2200-ohm, ½-watt resistor (usually not required)
R4—2.7-ohm, ½-watt resistor
T1—12-volt, 1-A filament transformer
Misc.: Printed-circuit board; heat sink made from 0.04-inch-thick copper (see text); etc.



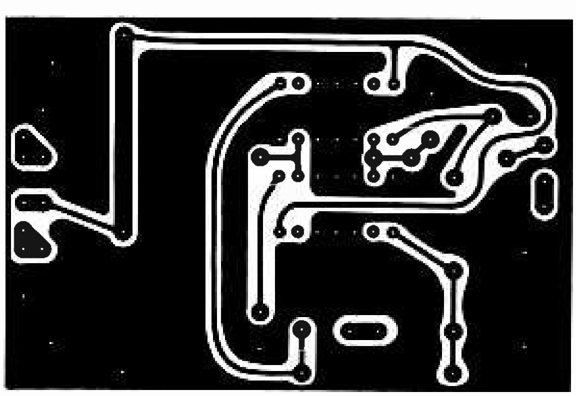


Fig. 3. PC foil pattern and photo showing parts layout.

amplifying high-frequency noise.

The gain of the amplifier is determined by the value of *R2*. With the value shown, about 150 mV will drive the amplifier to full output. By substituting a small 500,000-ohm potentiometer for *R2*, you can make your own sensitivity adjustments. However, if a potentiometer is used, shield its leads.

As with most direct-coupled amplifiers, a small dc offset voltage appears across the output. Resistor *R3* (usually not required) is used to adjust the bias on *IC1* and maintain the offset at a minimum.

The power supply shown in Fig. 2 will provide about 17.5 volts dc at 1 ampere. At this voltage, the amplifier will deliver about 3 watts of rms power into an 8-ohm speaker—more than enough to drive an efficient speaker in a bass-reflex enclosure. By increasing the supply potential to 22 volts, however, the amplifier will deliver about 5 watts rms to the load. (Note: Do not go beyond 22 volts; this is the absolute maximum for the IC's.) A 12-volt battery or power supply will provide about a 1-watt output from the amplifier.

Construction. The complete amplifier, minus power transformer *T1*, can be assembled on a printed-circuit board, the etching and drilling guide and component placement for which are shown in Fig. 3. Note the large ground areas used for shielding and heat sinking. If the amplifier is to be used with a 12-volt dc power supply, additional heat sinking is not required, and pins 3, 4, 5, 10, 11 and 12 of both IC's should be soldered to the PC board.

Assuming that you will be operating the amplifier at 17.5 or 22 volts dc, pins 3, 4, 5, 10, 11, and 12 should be carefully bent upward and fitted into the holes in the heat sink (see Fig. 4 for details on how to make the heat sink) and solder in place. Be sure to solder a short bare copper wire from the hole next to pin 11 of *IC2* on the heat radiator to the hole in the PC board. ◈

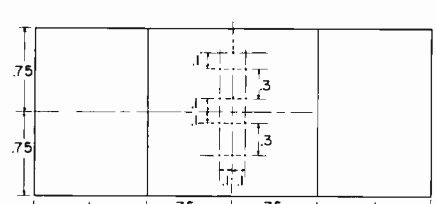


Fig. 4. Mechanical details of IC heatsink.

7400N TTL

SN7400N	.16
SN7401N	.16
SN7402N	21
SN7403N	.16
SN7404N	.21
SN7405N	24
SN7406N	45
SN7407N	45
SN7408N	.25
SN7409N	25
SN7410N	20
SN7411N	.30
SN7412N	42
SN7413N	85
SN7414N	70
SN7416N	43
SN7417N	43
SN7418N	25
SN7420N	21
SN7421N	39
SN7423N	37
SN7425N	43
SN7426N	31
SN7427N	37
SN7429N	42
SN7430N	26
SN7432N	31
SN7437N	47
SN7438N	.40
SN7439A	.25
SN7440N	21
SN7441N	1 10
SN7442N	1 08
SN7443N	1 05
SN7444N	1 10
SN7445N	1.10
SN7446N	1 15
SN7447N	.89
SN7448N	.99
SN7450N	26

SN7451N	27
SN7453N	27
SN7454N	41
SN7459A	25
SN7460N	22
SN7470N	45
SN7472N	39
SN7473N	45
SN7474N	45
SN7475N	80
SN7476N	47
SN7480N	50
SN7482N	1 75
SN7483N	1 15
SN7485N	1 12
SN7486N	45
SN7488N	3 50
SN7489N	3 00
SN7490N	59
SN7491N	1 20
SN7492N	82
SN7493N	.82
SN7494N	91
SN7495N	91
SN7496N	91
SN74100N	1 25
SN74107N	49
SN74121N	55
SN74122N	49
SN74123N	1 05
SN74125N	60
SN74126N	81
SN74132N	3 00
SN74141N	1 15
SN74142N	6 50
SN74143N	7 00
SN74144N	7.00
SN74145N	1 15
SN74148N	2.50
SN74150N	1 10

SN74151N	1 25
SN74153N	1 35
SN74154N	1 25
SN74155N	1 21
SN74156N	1 30
SN74157N	1 30
SN74160N	1 75
SN74161N	1 45
SN74163N	1 65
SN74164N	1 65
SN74165N	1 65
SN74166N	1 70
SN74167N	5 50
SN74170N	3 00
SN74172N	18.00
SN74173N	1 70
SN74174N	1 95
SN74175N	1.95
SN74176N	90
SN74177N	90
SN74180N	1 05
SN74181N	3 55
SN74182N	95
SN74184N	2 30
SN74185N	2 20
SN74187N	6 00
SN74190N	1 50
SN74191N	1 50
SN74192N	1 50
SN74193N	1 40
SN74194N	1 45
SN74195N	1 00
SN74196N	1 25
SN74197N	1 00
SN74198N	2 25
SN74199N	2.25
SN74200N	7 00
SN74251N	2 50
SN74284N	6 00
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CD4019	.69
CD4020	1 69
CD4022	1.25
CD4023	.29
CD4024	1.50
CD4025	34
CD4027	.85
CD4028	1 65
CD4029	2.90

CD4030	.65
CD4035	1 85
CD4040	2.45
CD4042	1.90
CD4044	1 50
CD4046	2.51
CD4047	2.75
CD4049	79
CD4050	79
CD4051	2.98
CD4053	2 98
CD4060	3 25
CD4066	1.75
CD4069	.45
CD4071	45
CD4081	.45
74C00N	39
74C02N	55
74C04N	75

74C10N	65
74C20N	65
74C30N	65
74C42N	2 15
74C73N	1 50
74C74	1 15
74C90N	3 00
74C95N	2 00
74C107N	1 25
74C151	2 90
74C154	3 00
74C157	2 15
74C160	3 25
74C161	3 25
74C163	3 00
74C164	3 25
74C173	2 60
74C193	2 75
74C195	2 75
80C97	1 50

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LM308H	1 00
LM308CN	1.00
LM309H	1 10
LM309K	1 25
LM310CN	1 15
LM311H	90
LM311N	90
LM318CN	1 50
LM319N	1 30
LM319D	9 00
LM320K 5 2	1 35
LM320K 12	1 35
LM320K 15	1 35
LM323K-5	14 00
LM324N	1 80
LM339N	1.70
LM340K-5	1 95
LM340K 12	1 95
LM340K 15	1 95
LM340K 24	1 95
LM340To 5	1 75
LM340To 6	1 75
LM340To 12	1 75
LM340To 15	1 75
LM340To 24	1 75
LM350N	1 00
LM351CN	65
LM370N	1 15
LM370H	1 15

LM373N	3 25
LM377N	4 00
LM380N	1 39
LM380CN	1 05
LM381N	1.79
LM382N	1 79
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NE536T	6 00
NE540L	6 00
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NE555V	75
NE565H	1.25
NE565N	1.95
NE566CN	1.95
NE567H	1.25
NE567V	1.95
LM703CN	45
LM709H	29
LM709N	29
LM710N	.79
LM711N	39
LM723N	55
LM723H	55
LM733N	1 00
LM739N	1 29
LM741CH	3/1 00
LM741CN	3/1 00
LM741 14N	39
LM747N	79
LM748H	39
LM748N	39
LM1303N	90
LM1304N	1 19
LM1305N	1 40
LM1307N	85

LM1310N	2.95
LM1351N	1 65
LM1414N	1 75
LM1458C	65
LM1496N	95
LM1556V	1 85
LM2111N	1 95
LM2901N	2 95
LM3065N	69
LM3900N	55
LM3905N	60
LM5556N	1 85
MC5558V	1 00
LM7525N	.90
LM7528N	2 20
LM7534N	2 20
LM7535N	1 25
8038B	4 95
LM75450	49
75451CN	39
75452CN	39
75453CN	39
75454CN	39
75491CN	79
75492CN	89
75494CN	89
RCA LINEAR	
CA3013	1 70
CA3023	2 15
CA3035	2 25
CA3039	1 35
CA3046	1 15
CA3059	2.46
CA3060	2 80
CA3080	85
CA3083	1 60
CA3086	.59
CA3089	3 25
CA3091	8 25
CA3123	1 85
CA3600	1 75

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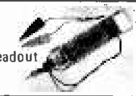

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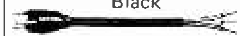

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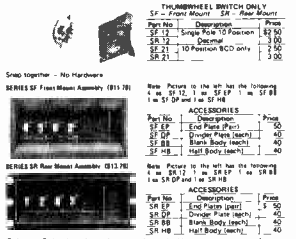

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Part No	Description	Price
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	84P44 062	4 50	8 50	2 56	2 31
	169P44 062	4 50	17 00	5 04	4 53
	469P44 062	8 50	17 00	9.23	8 28
EPOXY GLASS COPPER CLAD	169P44-062C1	4 50	17 00	6 80	6 12

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22 pin	65	59	53				

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18 pin	75	68	.62	40 pin	1 75	1 55	1 40

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ASST. 2	5 ea:	68 OHM 180 OHM	82 OHM 220 OHM	100 OHM 270 OHM	120 OHM 330 OHM	150 OHM 390 OHM	1/4 WATT 5% = 50 PCS.
ASST. 3	5 ea.	470 OHM 1 2K	560 OHM 1 5K	680 OHM 1 8K	820 OHM 2 2K	1K 2 7K	1/4 WATT 5% = 50 PCS.
ASST. 4	5 ea:	3 3K 8 2K	3 9K 10K	4 7K 12K	5 6K 15K	6 8K 18K	1/4 WATT 5% = 50 PCS.
ASST 5	5 ea:	22K 56K	27K 68K	33K 82K	39K 100K	47K 120K	1/4 WATT 5% = 50 PCS.
ASST. 6	5 ea:	150K 390K	180K 470K	220K 560K	270K 680K	330K 820K	1/4 WATT 5% = 50 PCS.
ASST. 7	5 ea.	1M 2 7M	1 2M 3 3M	1 5M 3.9M	1 8M 4.7M	2 2M 5.6M	1/4 WATT 5% = 50 PCS.

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ASST 10	2 ea	CD4001 CD4016	4002 4017	4011 4019	4012 4023	4013 4030	CMOS	$9 95 ASST
ASST 11	2 ea	LM301T LM311T	301N 565T	302T 567T	307T 741T	309K 741N	LINEAR	$10 95 ASST

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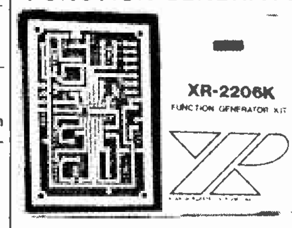
FUNCTION GENERATOR KIT



features sine triangle and square wave, THD 0 5% typ AM/FM capability

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Includes monolithic function generator IC, PC board, and assembly instruction manual

XR-2206KB $29.95
Same as XR 2206KA above and includes external components for PC board.

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IN746	3 3	400m	4/1 00	IN4003	200 PIV	1 AMP	10
IN751A	5 1	400m	4/1 00	IN4004	400 PIV	1 AMP	10
IN752	5 6	400m	4/1 00	IN3600	50	200m	6/1 00
IN753	6 2	400m	4/1 00	IN4148	75	10m	15/1 00
IN754	6.8	400m	4/1 00	IN4154	35	10m	12 1 00
IN965B	15	400m	4/1 00	IN4734	5 6	1w	28
IN5232	5 6	500m	28	IN4735	6 2	1w	28
IN5234	6.2	500m	28	IN4736	6 8	1w	28
IN5235	6 8	500m	28	IN4738	8 2	1w	28
IN5236	7 5	500m	28	IN4742	12	1w	28
IN456	25	40m	6/1 00	IN4744	15	1w	28
IN458	150	7m	6/1 00	IN1183	50 PIV	35 AMP	1 60
IN485A	180	10m	5/1 00	IN1184	100 PIV	35 AMP	1 70
IN4001	50 PIV	1 AMP	09	IN1186	200 PIV	35 AMP	1 80
IN4002	100 PIV	1 AMP	10	IN1188	400 PIV	35 AMP	3 00

TRANSISTORS

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50 VOLT CERAMIC DISC CAPACITORS

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10pf	05	04	03	001	05	04	035
22 pf	05	04	03	0047	05	04	035
47 pf	05	04	03	01	05	04	035
100 pf	05	04	03	022	06	05	04
220 pf	05	04	03	047	06	05	04
470 pf	05	.04	.035	1	12	09	075

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.001mf	12	10	.07	022mf	.13	11	08
.0022	.12	.10	.07	.047mf	21	.17	.13
.0047mf	12	.10	.07	.1mf	.27	.23	17
.01mf	12	10	07	22mf	.33	.27	.22

± 20% DIPPED TANTALUMS (SOLID) CAPACITORS

1	35V	28	23	17	1 5 35V	30	26	21
15	35V	28	23	17	2 2 25V	31	27	22
22	35V	28	23	17	3 3 25V	31	27	.22
33	35V	28	23	17	4 7 25V	32	28	23
47	35V	28	23	17	6 8 25V	36	31	25
68	35V	28	23	17	10 25V	40	35	29
1 0	35V	28	23	17	15 25V	63	50	40

MINIATURE ALUMINUM ELECTROLYTIC CAPACITORS

Axial Lead					Radial Lead				
47	50	15	13	10	47	25	15	13	10
1	50	16	14	11	47	50	16	14	11
3 3	50	15	13	10	1	16	15	13	10
4 7	25	16	14	12	1	25	16	14	11
10	25	15	13	10	1	50	16	14	11
10	50	16	14	12	4 7	16	15	13	10
22	25	17	15	12	4 7	25	15	13	10
22	50	24	20	18	4 7	50	16	14	11
47	25	19	17	15	10	16	14	12	09
47	50	25	21	19	10	25	15	13	10
100	25	24	20	18	10	50	16	14	12
100	50	35	30	28	47	50	24	21	19
220	25	32	28	25	100	16	19	15	14
220	50	45	41	38	100	25	24	20	18
470	25	33	29	27	100	50	35	30	28
1000	16	55	50	45	220	16	23	17	16
2200	16	70	62	55	470	25	31	28	26

CIRCLE NO. 15 ON FREE INFORMATION CARD

the IC "time machine"

Once you understand how the 555 IC timer works, there are many fascinating projects you can build.

BY WALTER G. JUNG

ONE might think that the number of "standard" IC building blocks would be limited since, by definition, a standard device is one which is usable in a wide variety of applications. However, just in the past few years, a new type of IC has shown signs of becoming a "standard." Interestingly enough, this category of device was not represented previously by an IC specifically designed to fulfill its function. This chip is the "555" IC timer, a versatile self-contained timing control circuit with the capability of a stable (free-running) or monostable (one-shot) operation over a wide range of pulse widths from microseconds to minutes. Furthermore, it operates from a single wide range power supply (+4.5 to +16 volts), and, as another bonus, has an output current of 200 mA.

Timing functions can, of course, be realized by other IC techniques, such as digital or op-amp multivibrators. However, when high-current loads are to be driven or single-supply operation is a must, both of these methods can be unattractive due to the number of components required. The picture changed, though, when Signetics introduced the first IC timer, the NE555, an 8-pin commercial device with a price tag in the range of $1. The 555 has quickly established itself and is available from a number of sources There are also dual and quad versions on the market.

The usefulness of the 555 is enhanced by its impressive performance specifications. Consider, for instance, its initial monostable timing accuracy, which is typically within 1% of the calculated value. This degree of accuracy is good for supply voltages of +5 or +15V, since the 555 by design provides an output pulse width that is independent of the supply voltage. This means you needn't be concerned with regulated supplies to maintain stability. In addition, once it is set up, a 555 will hold its pulse width. For instance, pulse-width variation is typically only 0.005% per degree C of temperature change—which is quite stable. In fact, the 555 can be considered to be temperature-independent over the modest temperature environments of experimental projects. (This is true, of course, *if* the *R* and *C* timing components are also temperature stable.)

What Makes the Timer Tick. Knowing the basic principle of what it does, a look inside a 555 is helpful in determinig how it works and how to use it most effectively. The block diagram of Fig. 1A shows the 555's functional components and its basic mode of operation—as a triggered one-shot timer. The internal circuit, while fairly complex, has a minimum of external connections (8 pins).

The circuit provides the functions of control, triggering, level sensing, and discharge, with a power output stage which delivers a high-level gate (near the V+ level) for the duration of the timing interval. Yet the complete timing operation is determined by only two external components, resistor R_A and capacitor *C*.

Monostable Model. In the standby state, the control flip-flop holds *Q1* on, clamping timing capacitor *C* to ground. In this state, the output (pin 3) is at ground level. The internal bias divider composed of three 5000-ohm resistors, provides bias voltages of ⅔ V+ and ⅓V+ to the upper and lower limit comparators, respectively. These two levels determine the voltage thresholds which, in turn, determine the timing interval.

Since the lower comparator is biased at ⅓V+, it stays in its standby state as long as the trigger input (pin 2) is held high (greater than ⅓V+) by R_P. When pin 2 goes low, the lower comparator sets the flip-flop, turning

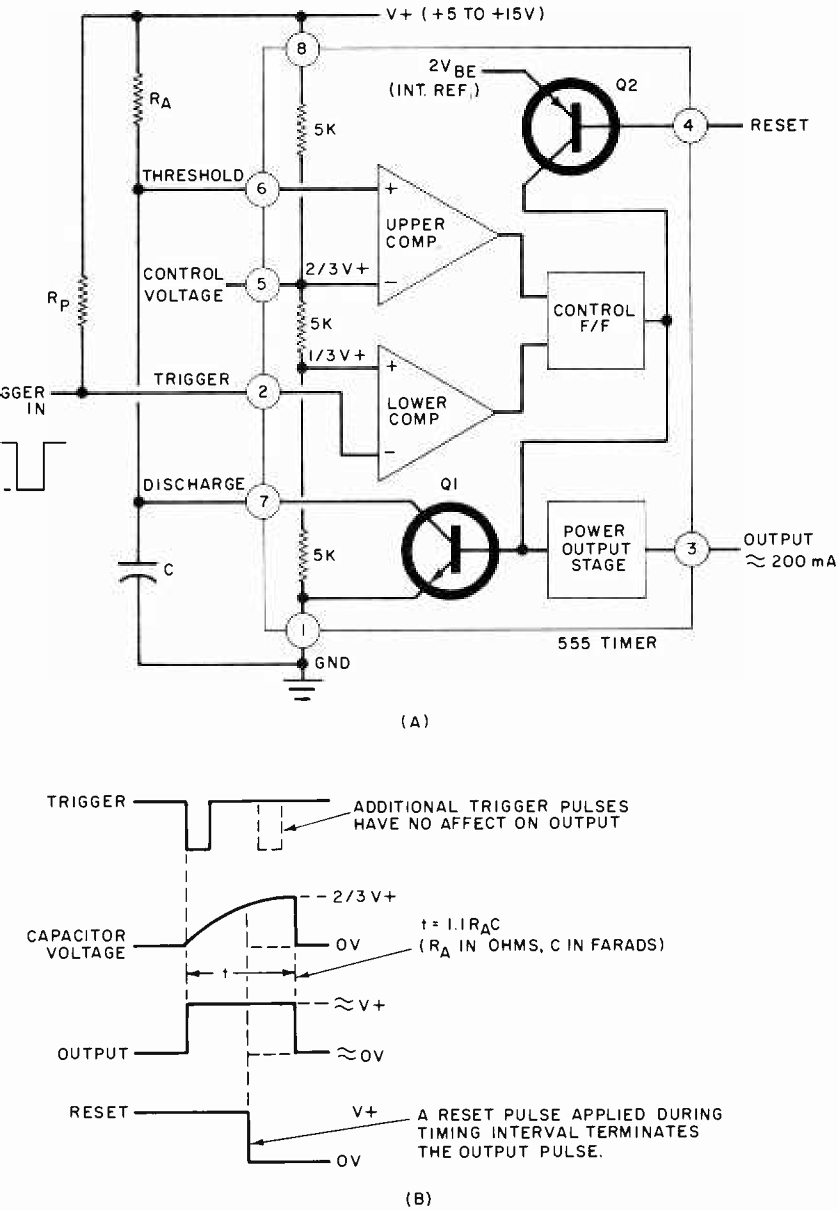


Fig. 1. Internal logic (A) of the 555 timer and wave-forms (B) during triggered operation.

off $Q1$, and the output goes to its high state (near V+). Since capacitor C is now unclamped, it charges exponentially (through R_A) toward V+. After a period of time equal to $1.1R_AC$, the voltage across C reaches ⅔V+, which is the threshold of the upper comparator (pin 6). At this time, the upper comparator resets the flip-flop, which turns on $Q1$, discharging C to zero and returning the output to the low (standby) stage.

The 555 monostable timing sequence is shown in Fig. 1B. In addition to the basic operation just described, there are two other points of interest. One is that any additional input triggers (shown dotted in Fig. 1B) during the timing interval will not affect the output. That is once triggered, the cycle will time out regardless of a subsequent trigger. Trigger pulse duration should be less than the output pulse width. This can be accomplished by differentiation, which also improves noise immunity.

A second point is that the reset function, when activated by a low-level input at pin 4, turns on $Q1$ and terminates the output pulse. The output is held low as long as pin 4 is low. The use of the reset input is optional. If not used, pin 4 should be tied to V+ to avoid possible triggering from noise.

The 555's interesting feature of an output pulse width that is independent of supply voltage comes about from the fact that the timing voltage reference (⅔V+) and the charging rate of C are both proportional to the supply voltage. Consequently, variations in the supply affect both in a manner that cancels changes in the time interval.

Note also that the upper threshold voltage is made available at pin 5. This allows external control of pulse width if desired. If this feature is not used, it is recommended that pin 5 be bypassed to ground with a small (0.01-µF) capacitor to prevent noise problems.

Triggered Monostable. A triggered monostable circuit is shown in Fig. 2A. It includes the $R1C1$ network, which prevents any possibility of mistriggering on positive edges. Values for *R1* and *C1* are not critical.

Values for R_A and C are selected from the timing chart shown in Fig. 2B. For best performance, there are several guidelines that should be followed concerning R_A and C. Stay within the range of resistances shown, and avoid the use of large-value electrolytics if possible since they tend to be leaky Leakage is, of course, more of a problem with long timing periods (large values of C), a "fact of life" which limits the upper range of timing. If electrolytic capacitors are necessary, tantalums should be used because of their low leakage. Voltage derating will also help minimize leakage current. With timing components of good quality, the 555 will provide accurate, stable pulses.

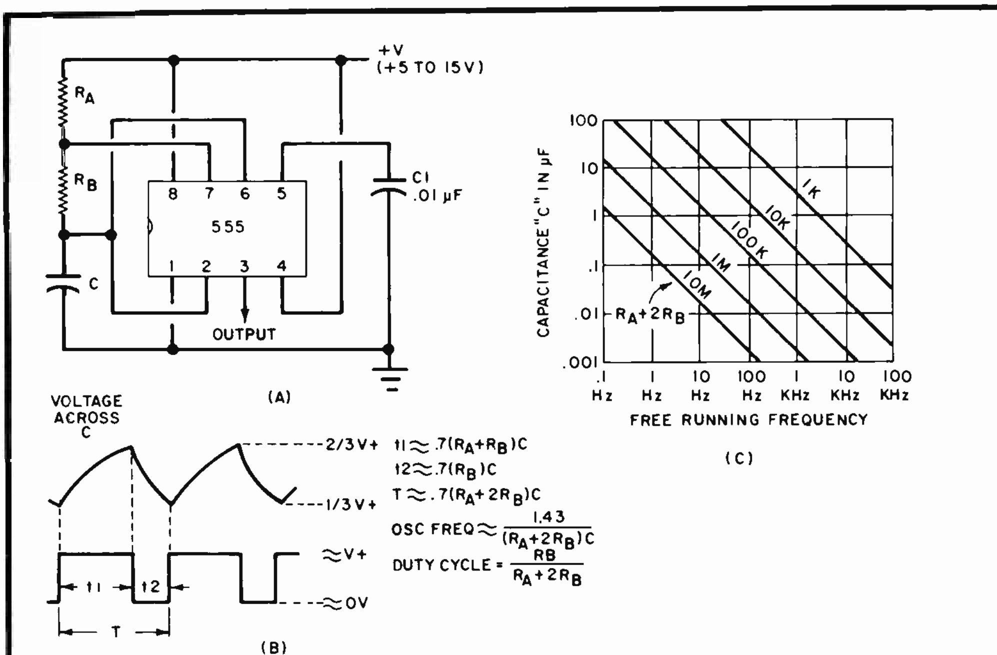


Fig. 3. As an astable multivibrator (A), the 555 has the waveforms shown in (B). Frequency is determined from (C).

Astable Mode. The second basic operating mode of the 555 is as an astable multivibrator (Fig. 3A). Here the timing resistance is split into two sections, R_A and R_B, with the discharge transistor (pin 7) connected to the junction. Upon start up, C charges toward V+ through R_A and R_B until the charge reaches ⅔V+, which triggers the upper comparator. The capacitor then starts to discharge toward ground through R_B until the charge reaches ⅓V+, when the lower comparator triggers. This starts a new charge cycle.

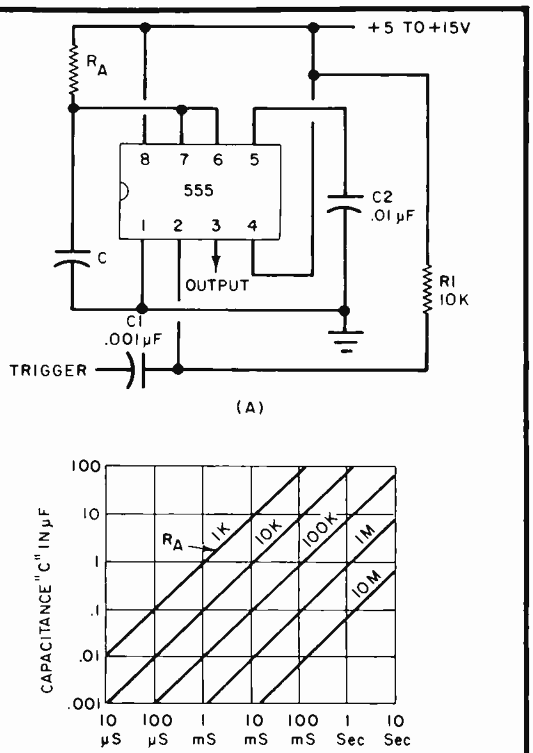


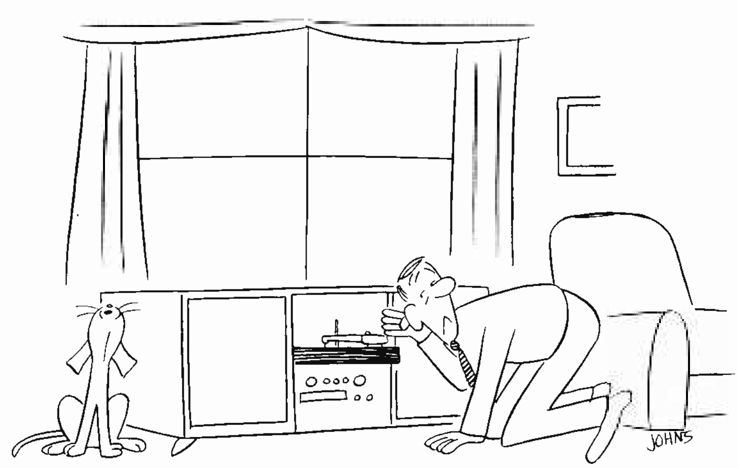
Fig. 2. (A) Use of 555 as triggered monostable. Chart (B) is used to select capacitor and resistor values.

The capacitor is charged and discharged between the limits of ⅔ V+ and ⅓ V+, as shown in Fig. 3B. The output state is, as before, high during the charge cycle and low during discharge. Timing equations for this mode are somewhat more complex (Fig. 3B). However, values for the resistances and capacitances can be chosen by using Fig. 3C. Since the capacitor is charged by two timing resistors and discharged by only one, the output waveform is asymmetrical, not square.

Times *t1* and *t2* (and thus the frequency) are independent of V+, as in the monostable circuit.

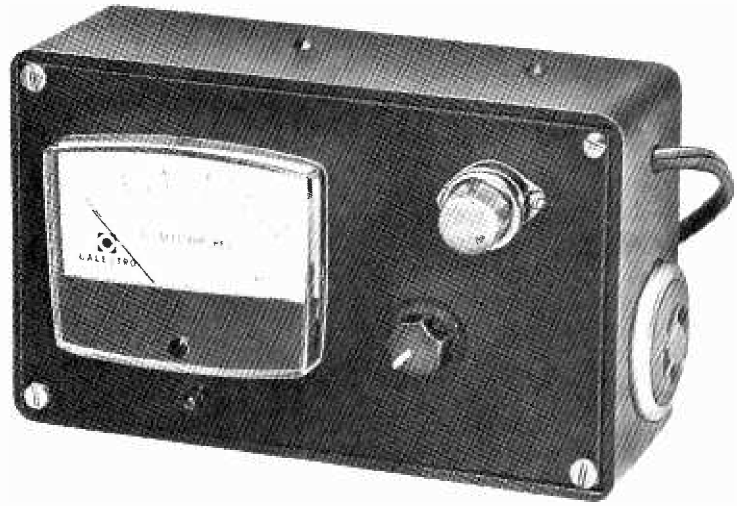
Types and Sources. Type numbers for 555 timers from various manufacturers are: Signetics, NE555V; National LM555CN; Motorola MC1455P1; Lithic Systems LS555; Fairchild NE555; Intersil NE555V; Silicon General SG555C; and Raytheon RC555DN. These are all single, 8-pin minidip devices.

Dual 555 units are: Signetics NE556A; Exar XR-556CP; Raytheon RC556DP; Fairchild NE556A; Silicon General SG556CN; Motorola MC3556CP; and Lithic Systems LS555-2. The Signetics and similar devices are 14-pin dual inline units, while the Lithic Systems unit is a 16-pin, with pins arranged indentically to two "vertically stacked" 555's. Signetics also manufactures 555 type quad timers, NE553 and NE554. ◈



BY HERB COHEN

Low-cost sensor sounds an alarm with as little as 50 parts per million of most toxic gases



POISONED AIR DECTECTOR

IF YOU live in a large industrial area, the air you breathe may contain varying amounts of carbon monoxide, hydrocarbons, soot, smoke, cooking gas, and many other potentially dangerous contaminants. These toxic elements may be odorless; most of them are combustible, and some are dangerously explosive.

Even a home in the "clean" suburbs could contain excessive amounts of cooking gas, cleaning chemicals, paint fumes, carbon monoxide seepage from garages, or potentially dangerous smoke. In a closed camper or boat, you can get carbon monoxide or gasoline fume leakage due to faulty engine exhaust and chassis or deck leakage.

What do you do about all this? You can't do much unless you know that it is there. Thanks to a simple gas detector semiconductor you can now find out whether the contaminants in your air are dangerous or not. This semiconductor is used in the construction of a low-cost (about $28) sensor system that can detect a number of potentially dangerous gases at levels of less than 50 parts per million—well below the government safety standards for industrial hygiene. When the detector senses a sufficient amount of gas, a buzzer sounds off or, if a relay is added, an external audible alarm can then be powered.

How it Works. The actual detector (DET) is an n-type semiconductor of tin dioxide, heated by a platinum wire (Fig. 1). In the presence of a gas, the difference in electron energy levels between the molecules of gas and the semiconductor causes electrons to move from the gas to the semiconductor, decreasing its bulk resistance. For example, with a propane gas level of only 1000 parts per million, the sensor resistance will decrease to 5% of its resistance in clean air.

The semiconductor's internal heater operates at 1.5 volts and 500 mA ac, supplied by half of the secondary of *T1* through dropping resistor *R1*. Resistor *R2* is the output load for the detector and is connected in series with meter *M1*, which is used as a readout for sensor current—hence gas presence. Diode *D3* provides a constant 0.5-volt offset to allow the transistor amplifier to work at low levels of sensor current. Potentiome-

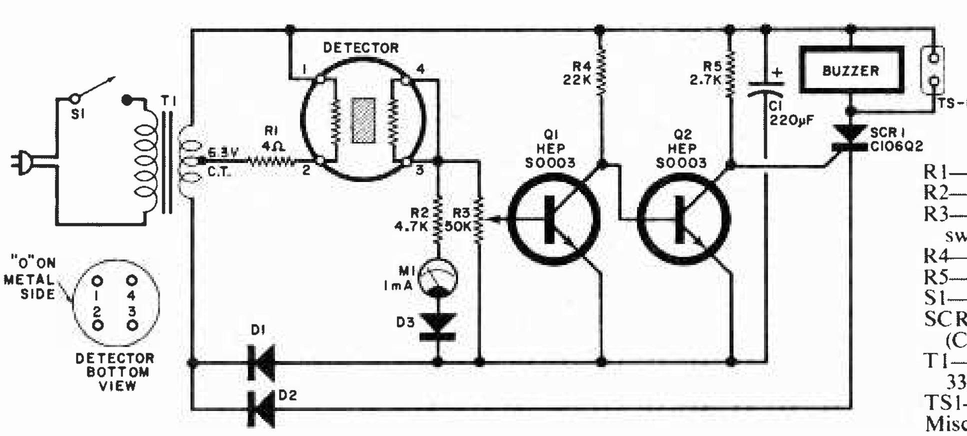


Fig. 1. When detector senses gas, current flow in meter indicates relative level.

PARTS LIST

C1—220-μF, 12-volt electrolytic capacitor
D1, D2, D3—1-A, 50-V PIV silicon diode
DET—Gas sensor (see note below)
M1—1-mA meter (Calectro D1-912 or equiv.)
Q1, Q2—HEPS0003 transistor
R1—1-ohm, 5-watt resistor
R2—4700-ohm, ½-watt resistor
R3—50,000 ohm linear taper pot (with switch S1)
R4—22,000-ohm resistor
R5—2700-ohm resistor
S1—Spst switch (on R3)
SCR1—Silicon controlled rectifier (C106Q2 or equiv.)
T1—6.3-VCT transformer (Lafayette 33E80490 or equiv.)
TS1—Two-post terminal strip
Misc.: 7-pin vacuum-tube socket, suitable chassis, 6-volt buzzer, perf board, knob, etc.
Note: The gas sensor can be ordered as DET-2 from Detectron Inc., P.O. Box 313A, Sag Harbor, NY 11963, for $8.95 postpaid. New York residents add applicable sales tax.

ter *R3* determines the alarm amplifier's operating point.

When a gas is present, the voltage across *R2* increases (meter indicating up-scale) and, depending on the setting of *R3*, transistor *Q1* turns on. Transistor *Q2*, which was in saturation while *Q1* was off, comes out of saturation, causing its collector to go positive. This turns on the gate of *SCR1* causing current to flow through the alarm buzzer (or the external circuit connected to terminal strip *TS1*).

The SCR is isolated from the amplifier power supply by rectifier *D2*. This gives the SCR a source of half-wave ac which allows it to turn off at the next zero crossing, after the gas level drops and the transistors return to their normal operating states. An optional LED and associated 470-ohm series resistor can be connected across the buzzer for visual indication.

Construction. Other than the detector (*DET*), *T1*, *M1*, and the buzzer and *TS1*, the circuit can be assembled on a piece of perf board, using the meter terminals as the mounting. Almost any chassis can be used as long as it will hold the transformer and buzzer and has a front panel large enough to accommodate the meter, *R3*, and the socket for the detector.

A conventional 7-pin vacuum-tube socket can be used to mount the detector. The socket should be attached to the exterior of the front panel. Note that the detector has a small circle stamped on its side between pins 1 and 2.

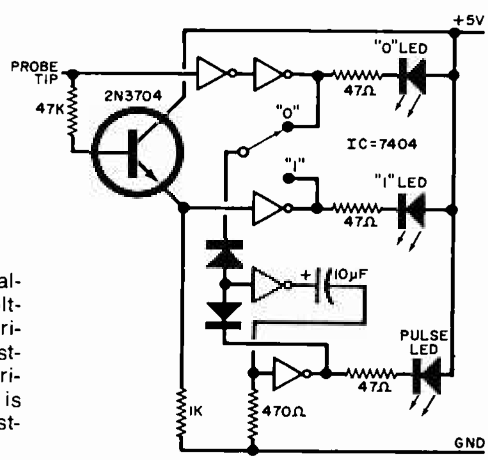
Operation. Before applying power, set *R3* to its minimum position to keep the alarm from sounding off immediately. Apply power and note that the meter needle rises to full scale and remains there for some time. This large sensor current is due to the "burning off" of impurities collected on the detector's surface while it was not in use. In cases of severe contamination, it may take quite a while for the sensor to clean itself, during which time the meter indication will gradually drop to some minimum value. Once the meter has dropped to its minimum, advance *R3* (sensitivity control) until the buzzer starts to sound off. Back *R3* off slightly until the buzzer stops and mark this point on the knob scale. This will be your local "normal." If the air is very clean, the buzzer may not sound, even at full sensitivity.

The unit is now ready for testing. You can blow cigarette smoke at the sensor or open a bottle of ammonia, perfume, etc. and blow the fumes toward the detector. The meter should suddenly jump up-scale and the buzzer should sound off.

For relative measurements, such as hunting for gas leaks, the buzzer can be silenced by rotating *R3* to its minimum with the meter indications used for probing for maximum gas concentrations. ◈

LOW-COST LOGIC PROBE

BY RANDALL GLISSMAN



DIGITAL circuit experimenters usually use a dc scope or some type of voltmeter to determine the logic states at various points of a circuit. Now, for an investment of only a few dollars, they can fabricate a logic probe with its own readout. It is the equal of many commercial probes costing much more.

As shown, the circuit uses a low-cost TTL chip, two silicon and three LED diodes. The presence of a logic 0 or 1 is indicated by their respective LED's, while the presence of a high-speed pulse, which may be far too fast for the logic circuits to catch, turns on the third LED. The latter will also light when a pulse train is probed.

With sufficient care, the complete circuit can be built into an ordinary metal cigar tube, with the insulated probe tip protruding from one end and the two power leads (one for the 5-volt supply and the other for ground) coming out the other end.

A small PC board can be fabricated to fit in the tube. The smallest available resistors should be used. The three LED's are mounted so that they can be viewed through holes cut in the appropriate places on the tube wall.

Power for the probe is taken from the TTL board being tested. Connect the +5-volt line and note that the proper LED is lit. Touching the tip to ground should cause the 0 LED to glow. The pulse-catching LED can be switched. ◈

Section III of this GE Guide

Written for the experimenter/ hobbyist

For GE-MOV varistors, tubes and transistors as well as your copy of the latest edition of the GE Replacement semiconductor Guide, see your General Electric distributor of entertainment receiving tubes and semiconductors.

Tube Products Department
General Electric Company
Owensboro, Kentucky 42301

GENERAL ELECTRIC

CIRCLE NO. 14 ON FREE INFORMATION CARD

HOW TO MAKE CUSTOM METERS FROM SALVAGED PARTS

Surplus d'Arsonval movements are easily converted to special-purpose voltmeters and ammeters.

BY PROF. ROBERT KOVAL

WITH the switch to digital logic and numeric readout devices in modern test equipment, the surplus market is becoming glutted with d'Arsonval meter movements. Actually, the availability of these parts is a boon to the electronics experimenter because the going prices for the movements are often only a small fraction of what he would have to pay if purchased from an industrial supply house.

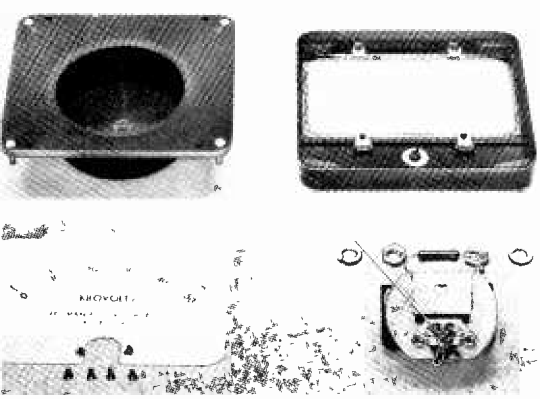


Fig. 1. First step is to disassemble and clean the surplus meter.

Most surplus meter movements can be refurbished and custom designed to suit just about any metering need imaginable. The process is relatively simple.

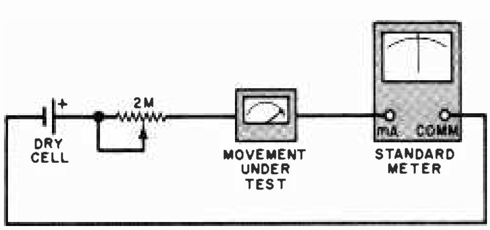


Fig. 2. Use this setup (with VOM and 1.5-V cell) to check full-scale value.

Preliminary Steps. Because the meter movement is from a surplus parts store, the first task is to clean away all dirt and other foreign matter from the case. This can be done with warm water and soap. For tough, greasy build-ups, try using some rubbing alcohol.

Once cleaned, carefully disassemble the movement (Fig. 1). Then inspect the movement to determine whether or not any resistors have been installed. Since you need only the basic movement for the next step, any resistors you find can be discarded.

Now, get out your VOM, a 2-megohm potentiometer, and a 1.5-volt dry cell with holder. Wire up the circuit shown in Fig. 2, but do not install the battery in its holder until you have adjusted the pot for maximum resistance. Connect the battery and slowly adjust the setting of the pot to obtain exactly full-scale pointer deflection on the meter movement. (Note: Temporarily replace the old meter scale to locate the full-scale position.) Since the meter under test is in series with the VOM, both units carry the same magnitude of current. Hence, the VOM's reading is the full-scale current sensitivity of the meter movement.

At this point, the resistance of the meter movement (*Rm*) must be determined. Do not use an ohmmeter to measure the movement's resistance; the current supplied by the ohmmeter could easily damage the movement beyond repair. A method has been developed for calculating *Rm* using only the basic movement, two resistors of known value, and a 1.5-volt dry cell. The circuit hookup is shown in Fig. 3. The resistor *Rser* should have a value large enough to permit *I1* to fall within the upper third of the scale. As a guide for choosing *Rser*, use Ohm's law. Assume the dry cell will be delivering 1.5 volts, and work this against the basic movement's full-scale current sensitivity. A fixed precision resistor would be ideal for *Rser*. The value of *Rsh* should be 1/10 or 1/20 the value of *Rser*. You can determine *I1* and *I2* from the meter's scales. Calculate *Rm* as follows:

$$Rm = \frac{Rser \times Rsh \times (I1 - I2)}{Rser \times I2 + Rsh\ (I2 - I1)}$$

You now have enough information to custom-design a voltmeter or ammeter.

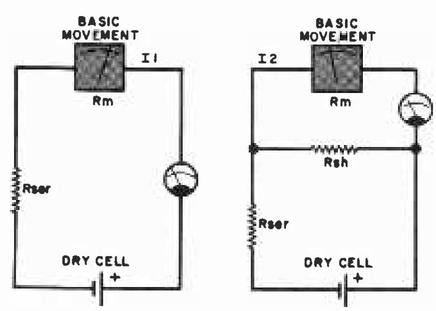


Fig. 3. Circuits for determining resistance of original meter movement.

The Custom Voltmeter. It is usually convenient to customize a meter movement in such a manner that it retains the same numeric sequence on the original meter scales to obviate the necessity of relabeling the scales. However, this is not absolutely necessary if you do not mind the task of removing the old and applying new legends.

Since the meter movement shown in Fig. 1 has a numeral 50 at its full-scale index, let us design a voltmeter with a

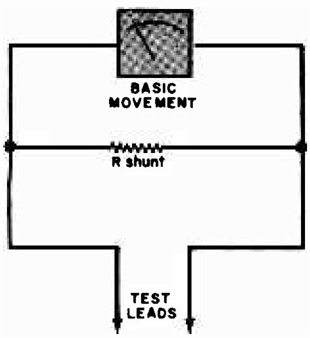


Fig. 4. Basic setup for determining the shunt resistor for an ammeter.

0 — 5-volt range. Assume that 50 μA is needed to deflect the pointer to full scale and that *Rm* is 2090 ohms. To calculate the value of the multiplier resistor (*Rmult*) for any given voltage range (*Vr*), use the following equation.

RESISTANCE PER UNIT LENGTH OF COPPER WIRE AT 25° C			
Gauge	Ohms per 1000 ft.	Gauge	Ohms per 1000 ft.
18	6.510	30	105.2
20	10.35	32	167.3
22	16.46	34	266.0
24	26.17	36	423.0
26	41.62	38	672.6
28	66.17	40	1069.0

$$Rmult = (Vr \times 1/Im) - Rm$$

In the equation, *Rm* is the basic movement's resistance (2090 ohms in our example), *Vr* is the voltage range desired (0 — 5 V full-scale), and 1/*Im* is the reciprocal of the current needed to obtain full-scale pointer deflection (1/0.000050). Hence, *Rmult* = (5 × 1/0.00005) − 2090 = 97,910 ohms.

As illustrated in the example, a 97,910-ohm resistor will yield a 0—5 volt range when connected in series with the basic meter movement. To change ranges, simply substitute the desired full-scale figure for *Vr* in the equation. If you want multi-range capability, calculate *Rmult* for each range desired and use a rotary switch for range selection.

Very likely, the value calculated for *Rmult* will not be readily available from the commercial selections listed. Do not let this deter you. It is a simple matter to arrange two or more resistors in series/parallel hookups to yield the required ohmic value. Alternatively, you can "trim" an ordinary carbon resistor to the proper resistance with the aid of a file (see Fig. 5). Select a fixed resistor of slightly lower value than required. For example, if you need 97,910 ohms, a standard 91,000-ohm carbon resistor can be used. Use an ohmmeter to verify that it is indeed less than 97,910 ohms; a 10-percent tolerance resistor can go as high as 100,100 ohms, a useless figure for the trimming procedure.

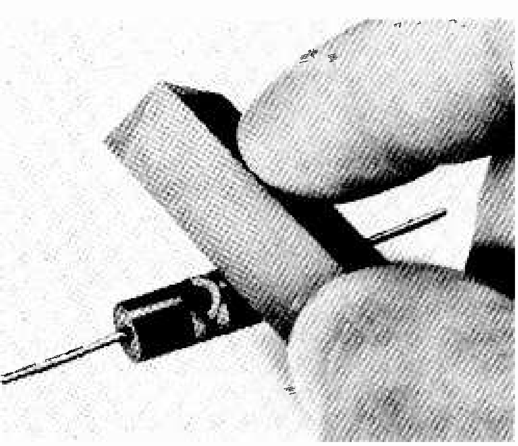


Fig. 5. Resistance of ordinary carbon resistor can be trimmed with a file.

Use a resistance bridge or an ohmmeter to monitor your progress as you cut into the resistor with the corner of a triangular file. Work very carefully so as not to trim away too much of the composition resistance material and end up with a value too high for your needs. When the resistor is trimmed to the proper value, liberally coat the notch with coil dope to seal out moisture. This will assure a constant resistance under changing humidity conditions.

The multiplier resistor can be mounted inside or outside the meter's case. A tag indicating the range and units can then be affixed to the meter face. Make it large enough to completely cover the original legend.

The Custom Ammeter. A custom ammeter can be designed around the basic meter movement with much the same ease encountered when making the voltmeter. The basic hookup is shown in Fig. 4. The equation to use for determining the resistance of the shunt resistor is:

$$Rshunt = \frac{Rm \times Im}{Imax - Im}$$

Maximum current *Imax* is the desired full-scale current the meter is to indicate, *Im* is the current required to deflect the meter's pointer to full-scale, and *Rm* is the resistance of the basic movement.

Assume that you want a range of 0-50 mA and that *Rm* and *Im* remain the same as in the voltmeter example. Then, *Rshunt* should be equal to (2090 × 0.00005/ 0.05 − 0.00005), or 2.092 ohms. Again, if a different range or ranges are desired, the maximum current wanted would be inserted into the equation as *Imax*. A switching arrangement would be used to provide several ranges.

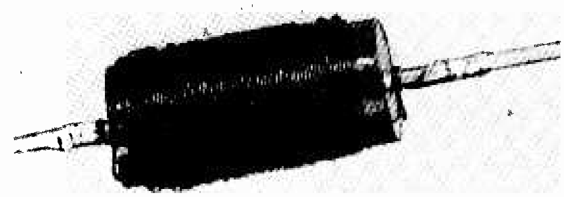
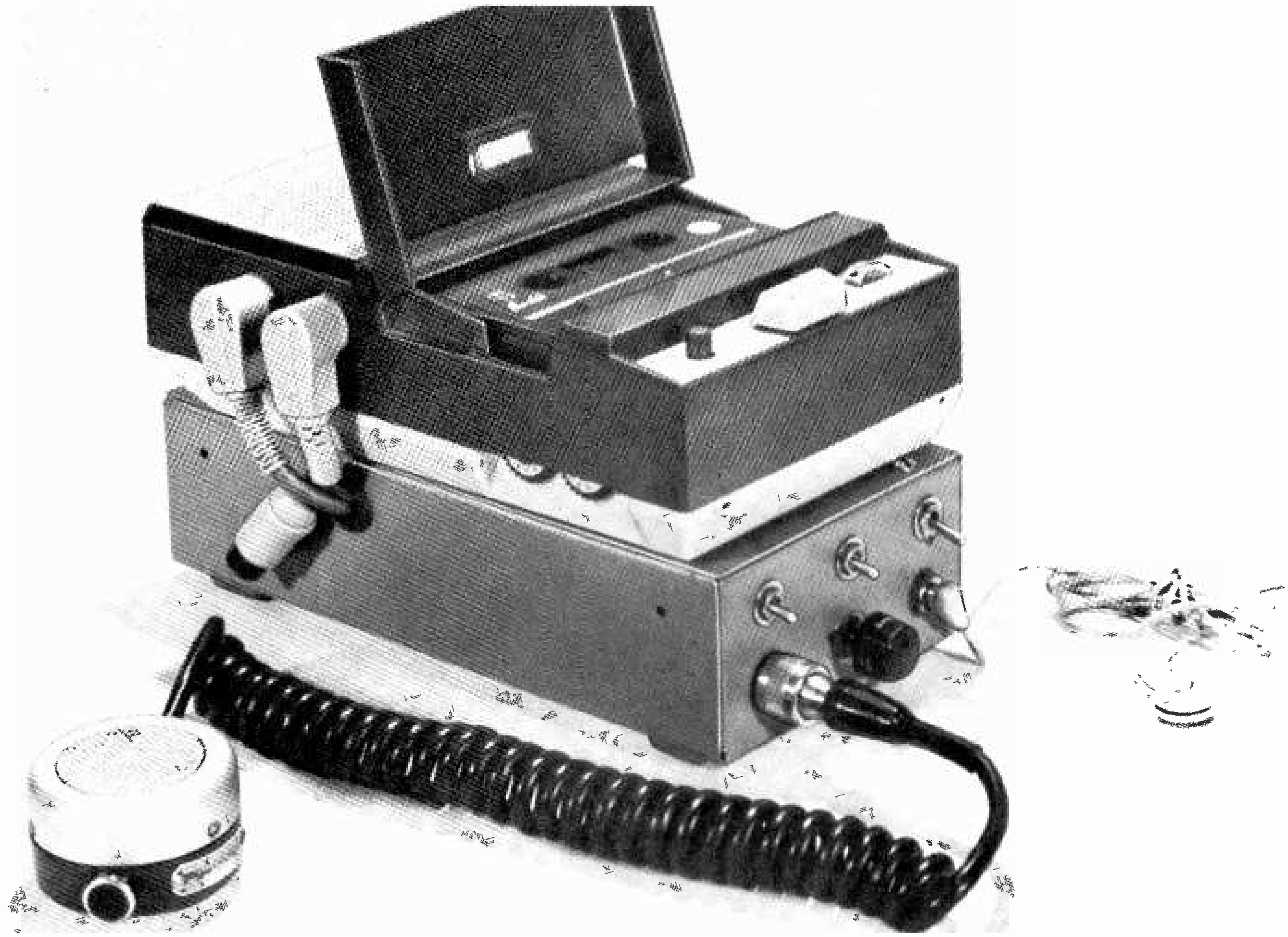


Fig. 6. Hand-wound shunt resistor. Next, the assembly is protected with a good coil dope.

The value of *Rshunt* will normally be very low, sometimes on the order of only a fraction of an ohm. In cases where its value would be too low to be conveniently trimmed with a file, you will have to wind your own shunt resistors. Enamel-coated copper wire can be used as the resistive element, while the resistor form can be any high-value resistor (1 megohm will do). Wire gauges and the resistance they yield are given in the Table. A hand-wound shunt resistor assembly is shown in Fig. 6. After winding the wire onto the resistor body and soldering the wire's ends to the resistor's leads, coat the assembly with coil dope.

As with the voltmeter, the ammeter's shunt resistor can be mounted inside or outside of the meter's case. Also, be sure to label the meter face with the range and unit for which it is designed. To check out your ammeter, connect it in series with a VOM and current source; both meters should indicate the same magnitude of current. ◈



VERSATILE TAPE RECORDER CONTROL

BY MARSHALL LINCOLN

Adds audio compression, squelch-activated start, and earphone amplifier.

THERE are many fine tape recorders available and they have all sorts of features to provide better recordings and, at the same time, make the job easier. But, as with most everything, there is usually room for improvement. Here are three "for instances." Sometimes audio-input level settings are so critical that satisfactory adjustment is difficult. This can be helped by the addition of an audio compressor to regulate the input. Improvement number 2 is the need for a way to turn the recorder on automatically when a transmission is picked up on a monitor receiver. A squelch-operated relay does this trick. Finally, suppose you want to use earphones to monitor recorded material or listen privately on a recorder that has only a low-level output jack. For this, all that is needed is an additional amplifier to drive the earphones.

Described here are circuits for solving all three of these problems, simply and economically. By using perfboard construction, you can build any one or all three of the circuits (shown in Fig. 1) to upgrade your recorder.

Compressor and Mike Preamp. Transistor *Q1* and its associated components comprise a conventional audio compressor which prevents overdriving of the recorder input by loud sounds such as talking too close to the microphone. Transistors *Q2* and *Q3* can be added to the compressor to provide adequate input level in case your particular recorder requires additional input sensitivity. Either one or both of these additional stages can be omitted.

Receiver Squelch Relay. Transistors *Q6, Q7*, and *Q8* drive a 1000-ohm relay (*K1*) to start the recorder whenever a signal is sensed in the squelch or limiter circuit of an FM monitor receiver. Switch *S3* activates this relay circuit and the setting of *R20* determines its sensitivity. Resistor *R19* is connected to the receiver squelch or limiter circuit at a point which goes negative upon receipt of a signal. The exact point will vary among different receivers, but it can

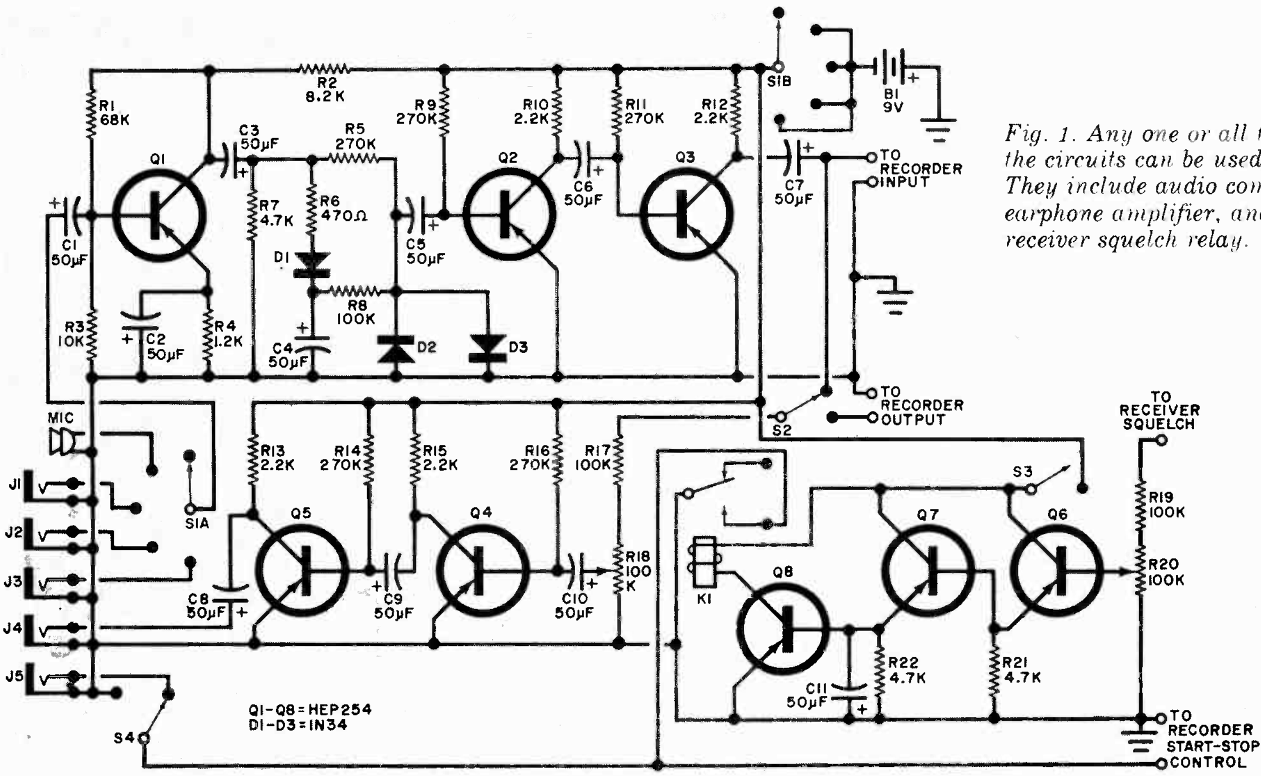


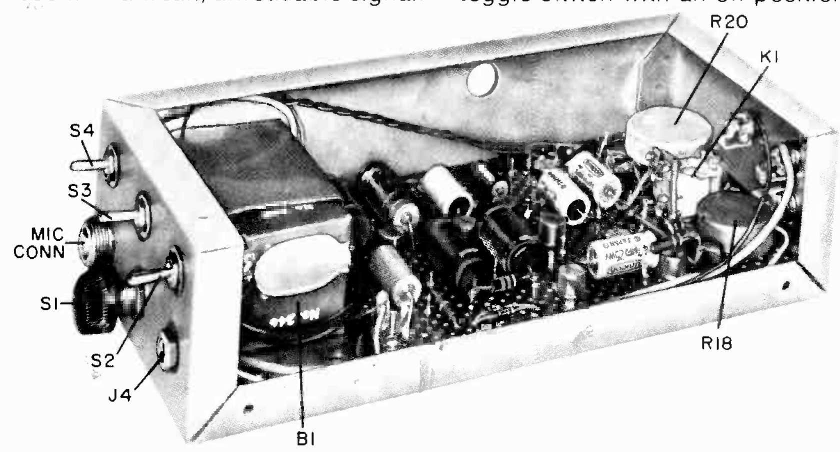
Fig. 1. Any one or all three of the circuits can be used. They include audio compressor, earphone amplifier, and receiver squelch relay.

PARTS LIST

B1—9-volt battery
C1—C11—50-μF, 25-volt electrolytic capacitor
D1—D3—1N34 diode
K1—1000-ohm, 3.5-mA relay (Calectro D1-962 or equiv.)
Q1—Q3—HEP254 transistor
R1—68,000-ohm resistor
R2—8200-ohm resistor
R3—10,000-ohm resistor
R4—1200-ohm resistor
R5,R9,R11,R14,R16—270,000-ohm resistor
R6—470-ohm resistor
R7,R21,R22—4700-ohm resistor
R8,R17,R19—100,000-ohm resistor
R10,R12,R13,R15—2200-ohm resistor
R18,R20—100,000-ohm potentiometer
S1—Dp 5-pos. rotary switch
S2,S4—Spdt switch
S3—Spst switch
Misc.: Suitable chassis; perf board with clips; battery clip; knobs; mounting hardware; etc.

be found by switching the monitor receiver to an active channel and checking voltage swings at various points in the limiter and squelch circuits (with a VTVM) to find one with noticeable voltage swing when a signal appears, without degrading receiver performance. When the point is found and the circuit is connected, adjust *R20* until the relay closes with a readable signal in the receiver, but does not close with a weak, unreadable signal.

Earphone Monitor Amplifier. Two low-level audio amplifier stages (*Q4* and *Q5*) will easily drive earphones from a low-level signal—such as that from the output connector on some recorders intended for feeding the playback signal to an external amplifier. This amplifier also drives the earphones with the output from the compressor and mike preamp (through *S2*) if desired. By using a toggle switch with an off position for *S2*, you can remove the earphone amplifier from the circuit. Potentiometer *R18* is set to produce the desired audio level.



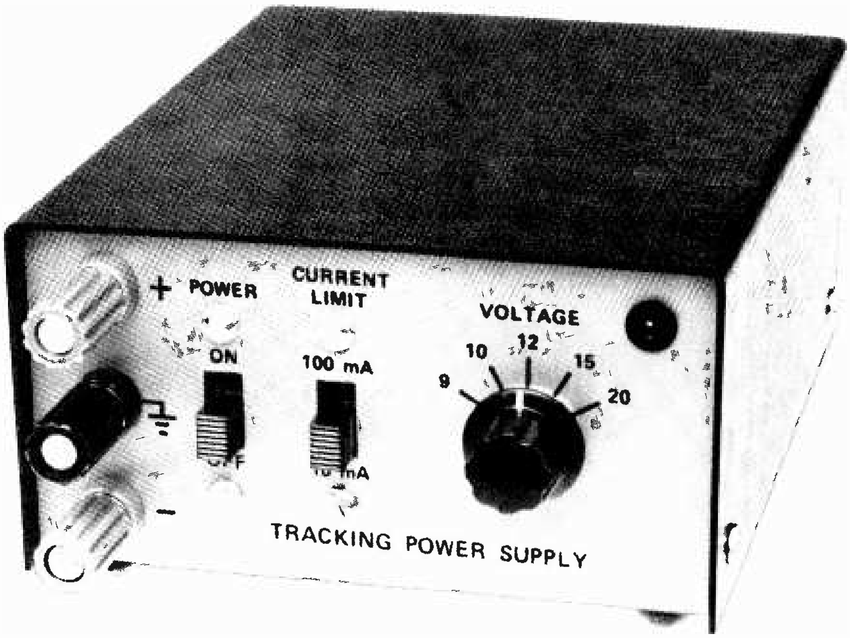
Layout of the prototype, although any type of chassis can be used.

Switching Circuits. Switch *S4*, when in the position shown, allows the recorder to be keyed on by either mike push-to-talk button (through *J5*) or the squelch relay.

Switch *S1* turns on the battery power and selects the desired input. With the switch in position 1 the power is off. Position 2 is for the mike, and positions 3, 4, and 5 are connected to miniature jacks on the rear panel for receivers, telephone pickup coil, or any other convenient device. Jack *J4* should be on the front panel for earphone monitoring.

Construction. The construction shown in the photos illustrates one of the many possible ways to assemble such a unit. The enclosure used in the prototype was 4" × 8" × 2", but the size will ultimately be determined by the recorder with which the add-on is to be used. ◈

HIGH-QUALITY BENCH POWER SUPPLY



BY MICHAEL S. ROBBINS

Single positive/negative supply has regulation better than 0.06%.

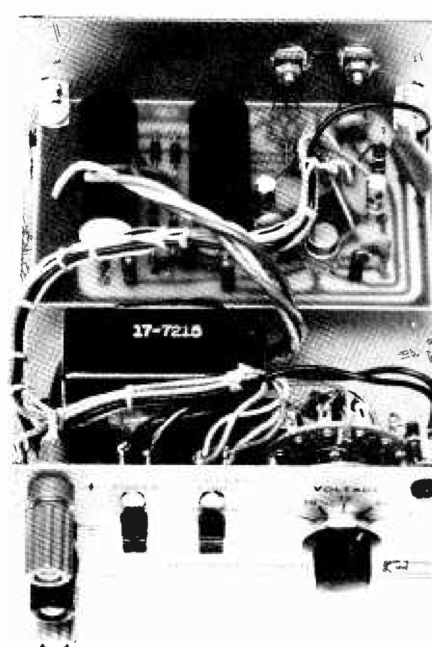
IF YOU are convinced that the op amp is here to stay and that two power supplies (positive and negative) are one too many, you need the compact single power supply described here. It uses a sophisticated IC to provide both positive and negative outputs which remain within 300 millivolts of each other; and it has line and load regulation of better than 0.06%. To keep the supply compact and easy to use, five pairs of switch-selected output voltages (+9, +10, +12, +15, and +20) and two current limits (10 mA and 100 mA) are provided instead of a control and a meter.

Circuit Operation. The IC used here is unique in that it contains two voltage regulators—one for positive and one for negative output. The portion that is the negative regulator is the key to the provision for variable-voltage outputs from both supplies. By varying the value of a single external resistor, the output can be changed over a wide range. Since the positive regulator "tracks" the negative regulator, the two outputs are the same—with opposite polarities.

Since the small pass transistors in the IC can dissipate only 0.68 watt, their outputs are used to drive external high-power pass transistors, *Q1* and *Q2*, as shown in Fig. 1. Current-limiting circuits in both sides of the IC regulator sense the voltage developed across *R4*, *R5*, *R6*, and *R7*. If this voltage exceeds 0.6 V, the output voltage drops.

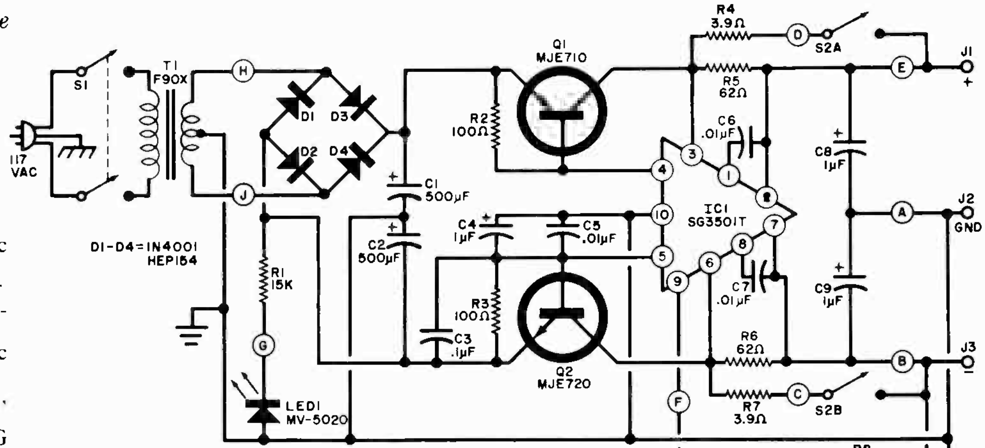
Construction. Layout of the supply is straightforward and many variations are possible. The use of the printed-circuit board shown in Fig. 2 is suggested, to avoid oscillations. Leads between the panel and the circuit board can be bundled, for neatness, as shown in the photograph.

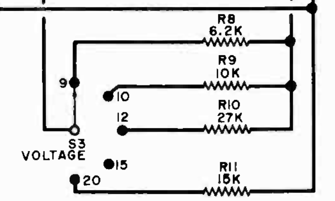
The cabinet can be fabricated from two pieces of 0.050-inch-thick aluminum, although a standard utility box can be used. The two pass transistors (*Q1* and *Q2*) must be heat sinked to the cabinet, insulated with a mica washer coated with heat-sink compound, and fastened with screw, nut, and lock washer.



Prototype photo shows how two transistors are mounted, using back wall of cabinet as heatsink.

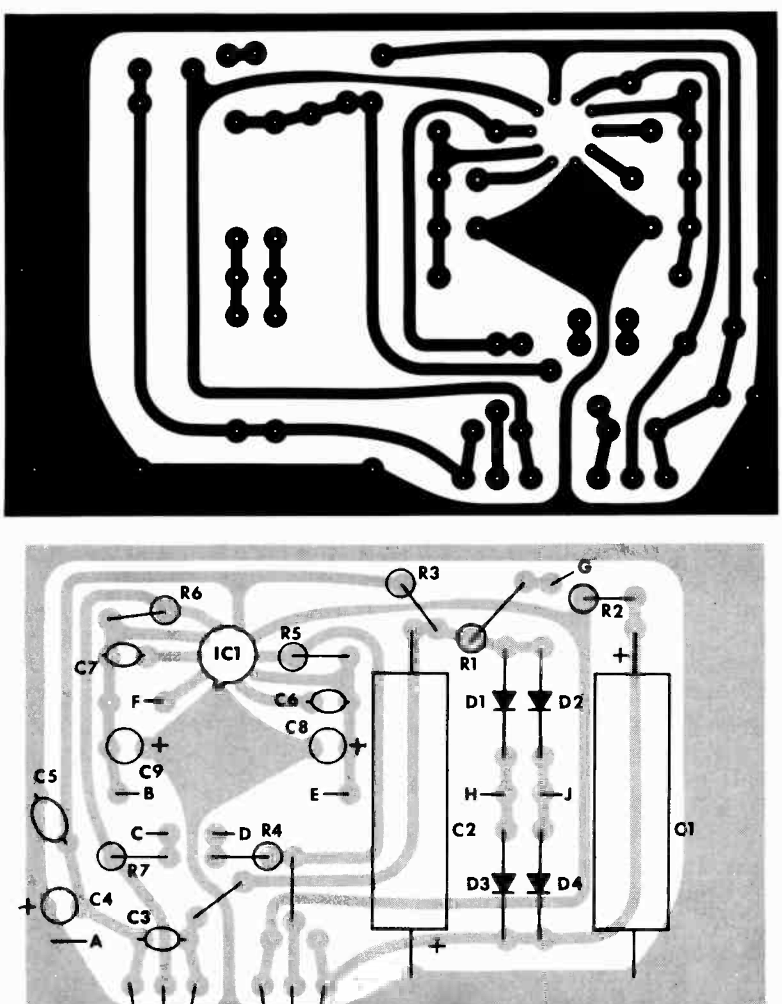
Fig. 1. One IC contains both positive and negative regulators, each feeding its own pass transistor. Voltage output is preset by S3, while S2 determines 10- or 100-mA current limit. Power-on is shown by LED.





PARTS LIST

C1, C2—500-μF, 25-V electrolytic capacitor
C3—0.1-μF, 25-V disc ceramic capacitor
C4, C8, C9—1-μF, 25-V PC-type electrolytic capacitor
C5, C7—0.01-μF, 25-V disc ceramic capacitor
D1—D4—Silicon rectifier (1N4001, HEP154, or equiv.)
IC1—Dual regulator (Silicon General SG 3501T)
J1—J3—Binding post (red, black, and blue)
LED1—Light-emitting diode with mounting clip (Monsanto MV-5020)
Q1—Pnp transistor (Motorola MJE710)
Q2—Npn transistor (Motorola MJE720)
R1—15,000-ohm, ½-watt resistor, 10%
R2, R3—100-ohm, ½-watt resistor, 10%
R4, R7—3.9-ohm, ½-watt resistor, 5%
R5, R6—62-ohm, ½-watt resistor, 5%
R8—6200-ohm, ½-watt resistor, 5%
R9—10,000-ohm, ½-watt resistor, 5%
R10—27,000-ohm, ½-watt resistor, 5%
R11—15,000-ohm, ½-watt resistor, 5%
S1, S2—Dpdt switch
S3—Sp 5-pos. shorting-type rotary switch (Oak 399-418K or Centralab PA-1000)
T1—Power transformer (40 VCT at 0.1A) (Triad F-90X or equiv.)
Misc.: Suitable cabinet, ac cord, mounting hardware, wire, solder, etc.



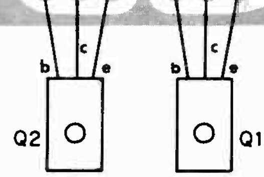


Fig. 2. Actual-size foil pattern and component installation for the power supply.

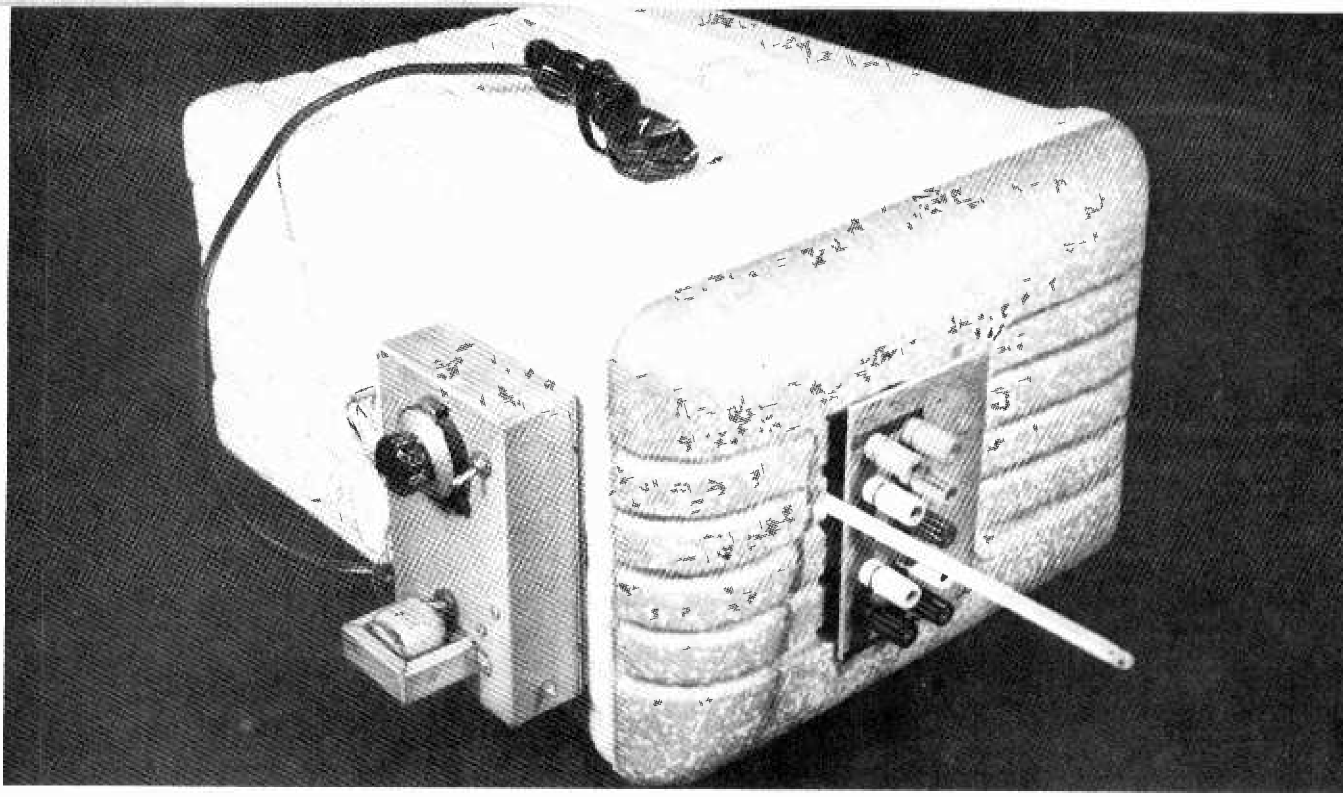
If the Triad F-90X transformer is used, cut off or tape the blue, black, black/white, and black/green leads. The red and green are the ends of the secondary, while the yellow is the center tap. The black/red and black/yellow leads are the primary.

Care should be taken when mounting the electrolytic capacitors, diodes, transistors, and IC. The leads on *Q1* and *Q2* should be left full length so that the transistor body can be mounted on the metal chassis.

Output voltages other than the five provided are made possible by changing the values of *R8, R9, R10,* and *R11*. Note that *R8, R9,* and *R10* connect between *S3* and the negative supply, while the 15-volt position does not have a resistor, and *R11* connects the 20-volt position to ground. A resistor is not needed at the 15-volt "crossover" point, while outputs above 15 volts require a resistor to ground. As the maximum output of this supply is approximately 20 volts, the minimum value of resistance is used for *R11*.

The PC board is laid out so that it can be used independently of the switches. It can be used as the internal power supply in any piece of equipment and can regulate currents up to about 1 ampere with suitable resistors.

Operation. The supply is ideal for use with IC's and hybrid circuits requiring regulated positive and negative voltages. It can also be used single-ended, since balanced loads are not required. ◈



BUILD THE TORTURE BOX

Miniature environmental test chamber can be set from 14°F to 158°F with 1-degree accuracy.

BY RALPH TENNY

WE ALL know how strict the temperature tolerance specifications are on components and systems for military and space applications; but do we ever stop to think whether the projects we build in our workshops will operate satisfactorily "in the field?" A fire detector, for example, that works fine in the controlled conditions of the workshop can go haywire in an attic in the summer when the temperature can reach 140° F. A metal locator may operate quite differently in the coolness of the forest in the fall and in the heat of summer on the beach.

Maybe it's time to take the guesswork out of building for unusual temperature ranges and install your own temperature test chamber, simply by building the Torture Box described here. It can be used to test circuits at temperatures from below −10° C (14° F) to +70° C (158° F). Of course, this range is probably more than you will need since it exceeds the range of many commercial components.

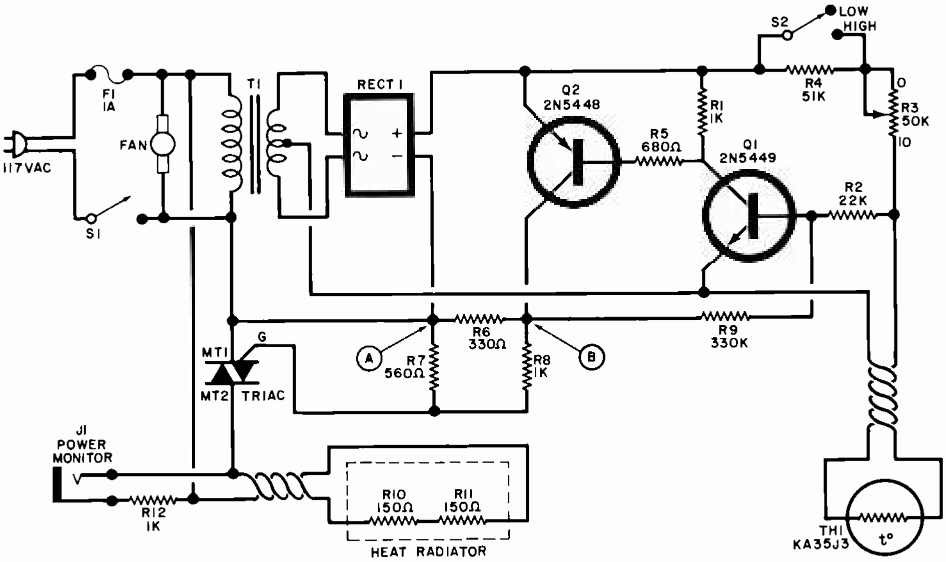
The Torture Box is a low-cost project that provides a change of pace for experimenters. The electronic circuits are fairly simple, but the project uses a combination of materials and techniques that is a little different. The basic box is an ordinary molded plastic picnic-type cooler. All subassemblies in the Torture Box are fastened to thin pieces of plywood or wall-panel material, which are fastened to the plastic using either white furniture glue or aliphatic (fatty, acrylic) resin. Do not use an aromatic glue or cement!

The operating range of the Torture Box can be extended, but temperatures higher than 80° C (176° F) should not be attempted because they may soften the plastic. A large quantity of dry ice will lower the temperature below −28° C (−18° F) but the nonlinearity of the control thermistor may hamper control below about −10° C.

Construction. Select a picnic cooler of sufficient internal volume. The one shown in the photos of the prototype is 12″ × 9″ × 12″ and has an internal volume of about 700 cubic inches.

The assembly of the small mechanical units that are attached to the chamber is described in the following paragraphs. Plan the location of these units in your particular cooler so that the weight distribution will not cause the finished chamber to tip. (Remember that the basic cooler is very light compared to the weight of the mechanical subassemblies.) As shown in the photos, the cooler was placed on its wide side, and four small pieces of similar plastic were glued to the bottom to serve as feet. Use a sharp instrument to make the required openings and holes. Keep the hot soldering iron away from the plastic. The cover should be tight fitting. If necessary, some type of locking device can be used.

Fan Motor. Any small motor is suitable. In the prototype, a shaded-pole motor/fan combination originally intended for electronic chassis ventilation was used. After drilling a hole for the motor shaft in the cooler wall, bend a mounting bracket or 1/16″ aluminum to secure the motor to the wooden mounting plate. Extend the motor shaft (using tubing) so that the fan will be located about ¾″ inside the cooler. Attach the motor mounting to the cooler as shown in Fig. 2.



PARTS LIST

F1—1-ampere fuse and holder
J1—Open-circuit jack (Calectro F2-842) or neon lamp (Radio Shack 272-1105)
Q1—2N5449 transistor
Q2—2N5448 transistor
R1, R8, R12—1000-ohm, ¼-watt resistor
R2—22,000-ohm, ¼-watt resistor
R3—50,000-ohm potentiometer
R4—51,000-ohm, ¼-watt resistor
R5—680-ohm, ¼-watt resistor
R6—330-ohm, ¼-watt resistor
R7—560-ohm, ¼-watt resistor
R9—330,000-ohm, ¼-watt resistor
R10, R11—150-ohm, 50-watt resistor (Dale RH-50 or equiv., with heatsink mounting)
Rect. 1—50-volt, 2-A rectifier (Radio Shack 276-1151)
S1, S2—Spst slide or toggle switch
T1—Transformer: 12.6-VCT, 0.1-A secondary (Calectro D1-750)
TH1—Thermistor (Gulton 35 TF1, Fenwal KA35J3, YSI44007)
Triac—RCA 40529
Misc.: Small shaded-pole motor and fan (see text); plastic container; white glue; ¼" × ¼" pine stock; 4" brass tube; sheet metal; wire screen; thermometer; perf board; mounting clips; sockets; control dial (Radio Shack 274-605); 5-way binding posts; plywood; mounting hardware; etc.

Fig. 1. Thermistor TH1 senses heat radiated by power resistors R10 and R11.

Control Circuit. The control circuit is mounted in a suitable chassis, the bottom plate of which is affixed to the cooler on the side opposite the fan as shown in the photo of Fig. 3.

With the exception of the thermistor (*TH1*), the triac, *T1, R10, R11*, and potentiometer *R3*, the circuit can be assembled on a small perf board, which is mounted in the upper portion of the control chassis. Potentiometer *R3* is mounted on the front panel and provided with a vernier dial drive.

The thermistor is connected to the end of a length of twisted-pair wire which is fed through a narrow tube 3" or 4" long. The tube is then inserted through the Styrofoam so that the thermistor is located within the box and the twisted pair can be connected to the perf board. The triac is mounted on a small heatsink isolated from the metal chassis. Range switch *S2* and power-monitor connector *J1* are mounted on the front panel. The transformer is mounted on the outside of the control chassis.

Power resistors *R10* and *R11* are mounted on a three-piece heat radiator whose configuration is shown in Fig. 4. The radiator consists of three pieces of thin brass sheet at least 2" wide and 4" long. Use heatsink grease between the pieces of the radiator and between each power resistor and heatsink.

When the electronic assembly is complete, temporarily disconnect the triac and connect a 10-volt dc voltmeter between points A and B of Fig. 1. With *R3* set to a low resistance, no dc voltage should be indicated between the test points. As the resistance of *R3* is increased, a 10-volt signal will appear. Make a check for both positions of range switch *S2* and note that the dc voltage appears at a much higher resistance on *R3* when *S2* is in the low range. If everything is OK, disconnect the unit from the power line and replace the triac.

Air Baffle. The baffle covers the fan and directs the air to the rear and thus counterclockwise around the interior of the chamber. The layout is shown in Fig. 5. The baffle is made of thin metal

HOW IT WORKS

The environmental chamber creates hot or cold temperatures by balancing a heater against the cooling effect of dry ice. A fan continuously circulates the air in the chamber, while a thermistor-controlled regulator circuit (Fig. 1) adjusts the temperature to the desired value, which is set on a dial. Transistors *Q1* and *Q2* form a complementary Schmitt trigger which normally has about 1.5 volts of lag (hysteresis). Since the trigger is powered by full-wave rectified dc with no filtering, the circuit voltage sweeps from zero through about 17 volts at a rate of 120 times per second. This varying power reduces the hysteresis to a few millivolts and thus provides control to ±1 degree.

If the thermistor resistance is below the set point (temperature dial setting), both *Q1* and *Q2* are cut off and *R7* keeps the triac cut off. As the chamber cools, the thermistor resistance increases until *Q1* starts to turn on. Shortly after that *Q2* turns on and feedback through *R9* increases the turn-on signal for *Q1*, causing the trigger to snap full on. A pulse of current through *R8* turns on the triac until the end of that half cycle of ac power. As the power passes through zero, the triac turns off and the cycle starts again. If the thermistor resistance is greatly out of balance, the triac will be turned on early in each cycle; a small unbalance will delay the triac turn-on until late in the cycle. Consequently, heating power (triac current in *R10* and *R11*) is applied in proportion to the difference between the actual temperature measured by the thermistor and the temperature set by the control dial.

Range switch *S2* and potentiometer *R4* extend the control range to low temperatures, without losing the resolution on *R3*. Consequently, the set point resolution approaches 1 degree F per division on the specified control dial.

TABLE 1

Control Dial Settings (Major Div.)	Temperature (° F) (S2 Position) (Low)	(High)
0	7	43
1	10	49
2	13	55
3	18	60
4	21	66
5	24	74
6	29	86
7	32	94
8	38	108
9	—	129
10	—	161

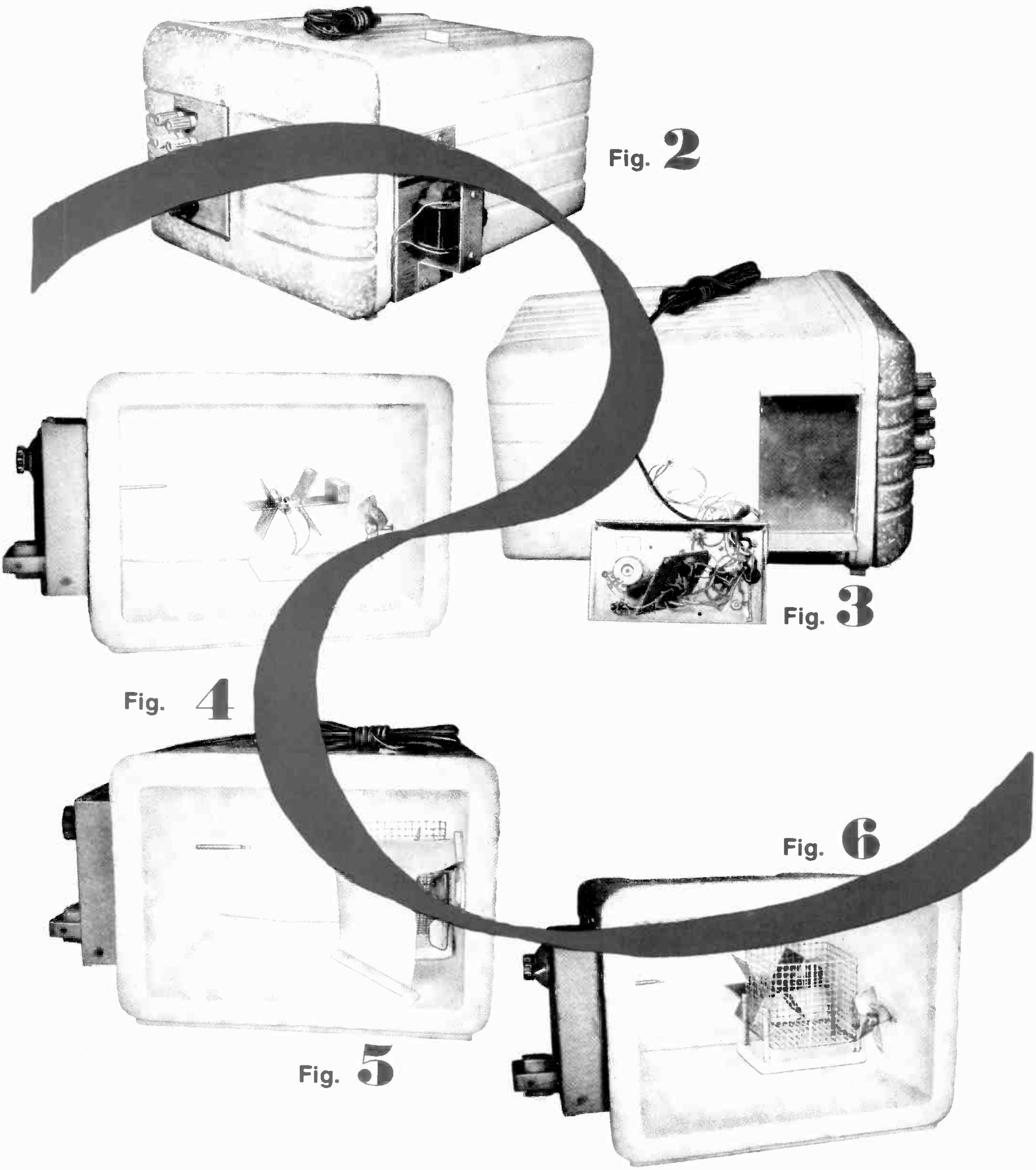


Fig. 2

Fig. 3

Fig. 4

Fig. 5

Fig. 6

stock but you should make a pattern using a piece of paper first to get the proper size and configuration. The baffle will be fixed to the side wall and bottom of the box using ¼" square pine blocks. Once the shape has been determined, cut the metal stock and install.

Ice Basket. The basket is an open-topped cube, about 3" on each edge, made of wire screen. Four ⅛" round dowels are glued to the corners with epoxy and the dowels are used to secure the basket to a plywood or plastic plate which is secured to the base of the chamber as shown in Fig. 6. When the basket is in place, cut a small hatch directly over it as shown in Fig. 7. Note that the hatch is cut with sloping sides so that the cover cannot drop into the cooler. Any small handle can be used on the cover.

Input Terminal Block. A minimum of ten 5-way color-coded binding posts should be provided for input, output, and power-supply connections to the equipment being tested. The terminals are affixed to a piece of plywood as shown in Fig. 2, with their leads protruding through the cover of the cooler.

Internal Circuit Board. As shown in Fig. 8, the internal terminal block is

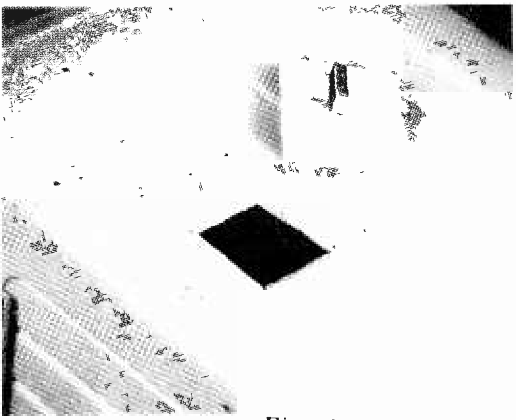


Fig. 7.

made from a 4½" × 6" glass-epoxy laminated board mounted in a frame of ¼" pine strips so that the board is far enough from the cover to be well within the chamber. Make sure that the wooden frame is waterproofed with varnish. The various input binding posts can be connected to color-coded perf-board pins on one edge of the board. Various combinations of sockets and perf-board pins can be attached to the board for testing different types of circuits.

Note also, in Fig. 8, that a conventional laboratory-type immersion thermometer is inserted through the cover to check the internal temperature. The thermometer must have an appropriate temperature range so that it can be read from the outside of the chamber.

Test and Calibration. Recheck the mechanical assembly of all cooler-

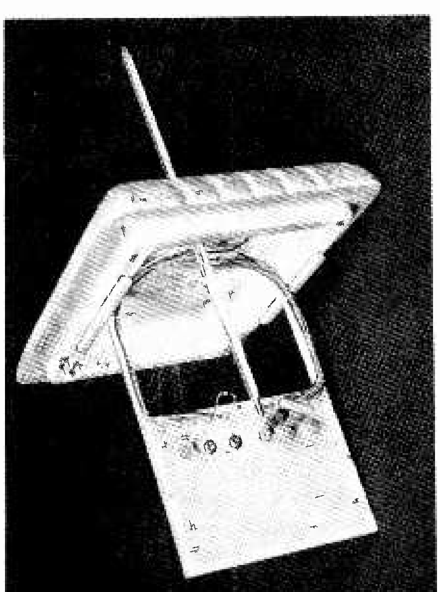


Fig. 8.

mounted components, making sure that all elements are firmly secured and that all glued joints are hard and dry. Recheck all the wiring in accordance with Fig. 1. Keep in mind that power-line ac is present on some leads and be very careful to avoid the possibility of an electrical shock.

Set the vernier dial on *R3* to 10 and slip the shaft of *R3* until the in-circuit resistance is about 3000 ohms. Set the range switch to high and set the control dial to zero. Connect a 150-volt ac meter to *J1* and with a thermometer inserted into the chamber, turn on the power. The fan should start to run and the voltmeter should indicate zero.

Advance the temperature-control dial toward 10 until the voltmeter indicates up-scale and note the dial indication. Advance the control toward the next major dial graduation and wait until the voltmeter shows that the heater power is cycling on and off every four or five minutes. Record the dial indication and the thermometer temperature. Continue this process until the control dial has reached 10 or the temperature reaches 70° C (158° F). Slip the shaft on *R3* until the 10 on the temperature-control dial causes the temperature to stabilize at 70° C.

Set the range switch to low and the temperature dial to 5. Put approximately 3 cubic inches of dry ice into the ice basket (through the small hatch on the top) and operate the system until the voltmeter shows that the heater circuit is cycling. Note the temperature and try new settings until the dial setting for 0° C (32° F) is found. At this point, the operation has been checked and the end points of the operating range have been found and calibrated. You can now fill in a calibration chart by recording temperatures at other major dial settings on both ranges. A typical calibration chart is shown in Table 1.

One-half pound of dry ice (usually available from ice cream stores) is sufficient for most tests. *Do not* handle dry ice with the bare hands as severe frostbite can result. A wide-mouth Thermos bottle can be used to store dry ice for as long as 8 hours, but *do not* close the lid tightly. Long-term storage of dry ice is essentially not possible for the home experimenter, but between 25% and 50% of a given amount will remain after 24 hours if stored in a good Thermos. To break dry ice into chunks, wrap it in a heavy cloth and pound with a hammer.

The power-monitor jack (*J1*) can be replaced with a neon lamp if desired since, once the monitor is calibrated, there is no further need for the jack—unless recalibration becomes necessary.

Using the Chamber. To test a circuit, you can assemble the circuit on the chamber's internal perf board or attach a finished board to the internal board mounts. Connect the power leads, inputs, and outputs to the cover binding posts and check for normal operation of the circuit with the chamber at room temperature.

Then supply power to the chamber, set the desired elevated temperature and see how your circuit works. If it passes this test, cool the chamber, checking circuit operation along the way. If the circuit doesn't pass the temperature test or (more commonly) if its operation drifts with temperature, the circuit must be temperature-compensated to limit drift to allowable levels. This means selecting components whose temperature coefficients

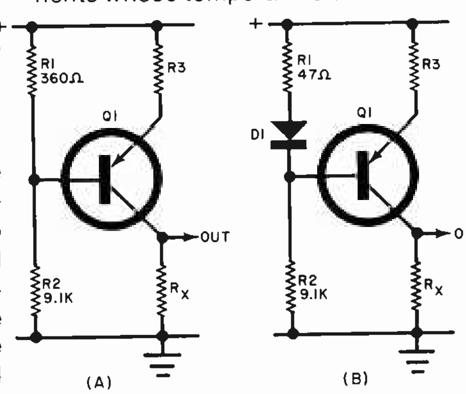


Fig. 9 Temperature compensation circuits.

compensate for temperature change or adding components that drift in the opposite direction.

The term "temperature coefficient" simply means how much a component will change in value with changes in temperature. This is usually expressed as % per degree C. For example, a fixed resistor of 1000 ohms having a 0.1%/°C temperature coefficient will change 1 ohm for each 1 degree C change in temperature. A +0.1%/°C coefficient indicates that the resistor will increase 1 ohm for each 1° C change in temperature. If the 1000 ohms is measured at 25° C, the resistor will measure 1050 ohms at 75° C and 975 ohms at 0° C.

There are capacitors with either positive or negative temperature coefficients. Most thermistors are resistors with negative temperature coefficients, although some companies also make thermistors with positive temperature coefficients. Also, silicon or germanium diodes can be added to a circuit to compensate for temperature drifts in transistors of the same material.

As an example of temperature compensation, consider the circuit of Fig. 9A, where *Q1* is a current source feeding a load, *Rx*. Resistors *R1* and *R2* set the reference level, while *R3* determines the amount of current flowing through the load. As the circuit elements heat up, the current through *Q1* will start to increase, thus increasing the load current. One way of compensating for this increase is shown in Fig. 9B, where a diode has been added in series with *R1*. If *Q1* is a silicon type, the diode must also be silicon. The modified circuit acts exactly the same as before except that the reference voltage is now the voltage across *R1* and *D1*. Resistor *R2* helps to control the current through the diode, but has less effect than it did in Fig. 9A.

To make a complete and proper compensation of load current with temperature, it is now necessary to vary *R2* and *R3* to get the desired current level and good stability with changes in temperature. You will see this method of temperature compensation used in many commercial units. ◈

BUILD A LOW-COST SQUELCH CIRCUIT

Useful addition for receivers without built-in squelch.

BY JOHN G. RAMSEY

MOST modern vhf monitors include an adjustable squelch to quiet the annoying hiss that is usually present when a signal is not being received. However, many of us have either older (non-squelch) sets or homemade versions that do not include this ear-saving circuit. Now, if you build the adjustable squelch shown here—at a cost of about $2.50—you can add this feature to any solid-state vhf/FM (police, fire, etc.) receiver.

Although the circuit shown is for a set using pnp transistors, simply by changing the type of transistor used for *Q1* and the connection to the ratio detector, you can use this circuit on a set with npn types.

Theory of Operation. As shown in the diagram, the emitter resistor of the set's audio driver is cut out of the circuit and replaced by *R1*. The potential at the top of *R1* is controlled by *Q1*. When a signal is not being received, the voltage across *Cx* (ratio detector capacitor in the set) is very low so that *Q1* is turned on. In this case, the audio-driver emitter is reverse-biased; and that stage will not be in operation. When a signal is received, *Cx* is charged up, which turns *Q1* off, allowing *R1* to complete the audio-driver emitter circuit and turn on the stage.

Incidentally, using a squelch will lengthen battery life because the current-consuming audio-output stage is not operating when there is no signal.

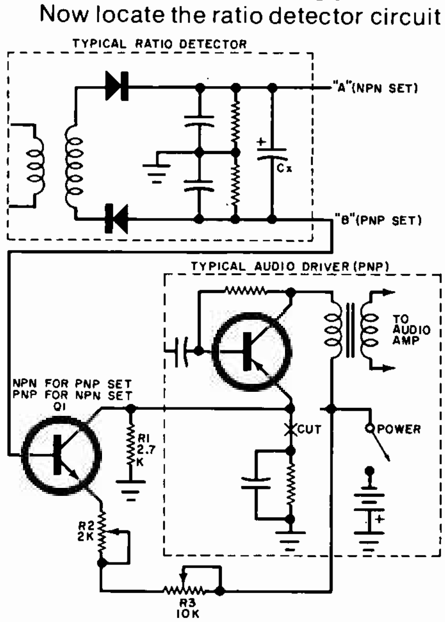
Construction. First, determine whether your receiver uses npn or pnp transistors. If the majority of them are black epoxy, the receiver is npn. If most of the transistors are in metal cases, the receiver is pnp. Select the transistor for *Q1* accordingly.

Now locate the ratio detector circuit in the receiver. The ratio detector consists of two i-f transformers inside the same case or just located very close to each other. Next to these are two diodes and two resistors, the latter having values between 220 and 1500 ohms. To one side of these resistors you will find an electrolytic capacitor with a value usually about 10 μF. The positive side of the capacitor is point A for npn receivers; the negative side is point B for pnp sets.

Next locate the audio driver stage and the resistor-capacitor combination in the emitter circuit. Cut this lead and connect the proper side of *Q1* and *R1* to the emitter side of the cut connection. Connect the slider of *R3* to the negative of your receiver, but remember to connect it after the switch.

There is usually sufficient room in most receivers to mount the additional transistor and resistor. Ideally, the controls (at least *R3*) should be mounted on the front panel. Although two potentiometers are shown for the squelch adjust, it is possible to get away with using only *R3*.

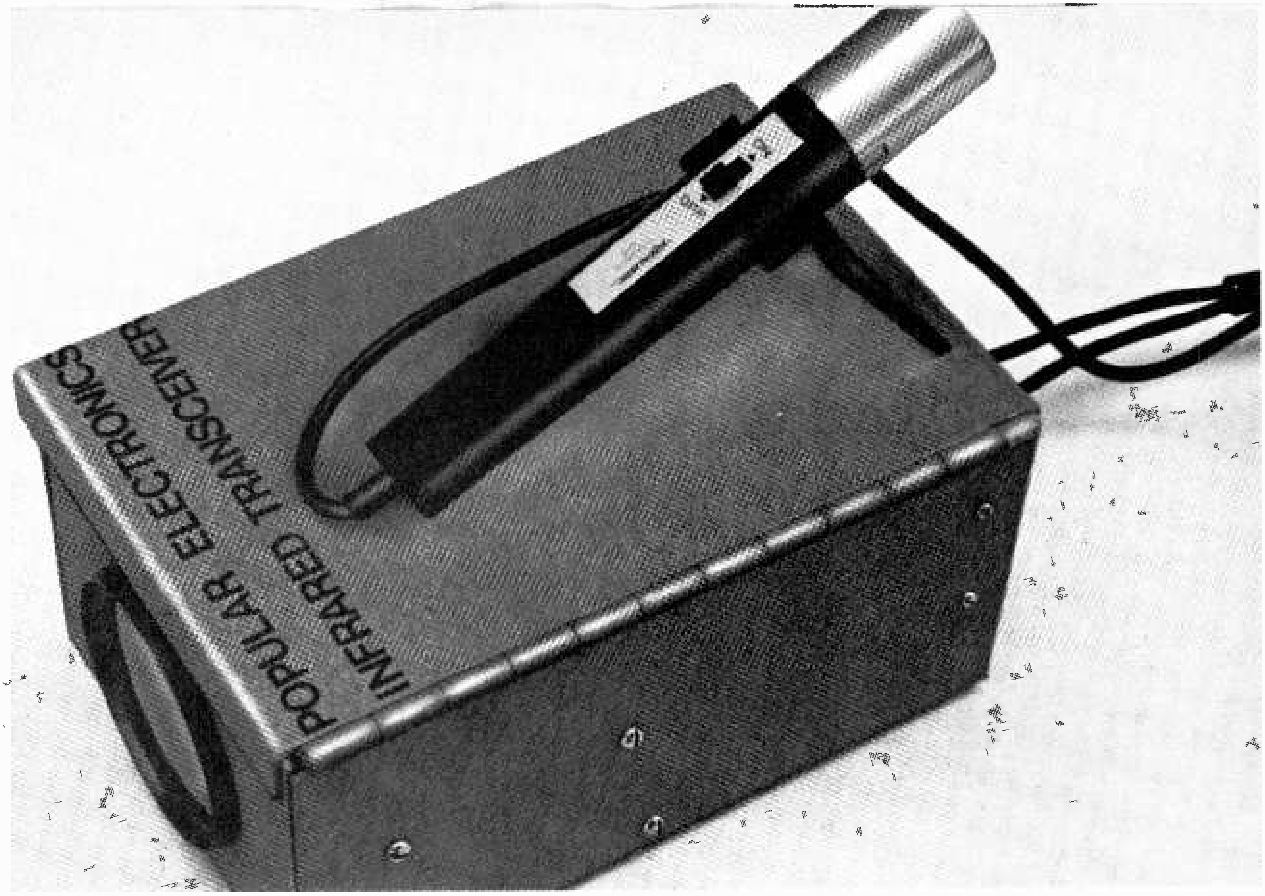
Adjustment. To adjust the squelch circuit, set *R2* so that its rotor is nearest *R3*. Then adjust *R3* until you hear noise from the receiver. Now, set *R2* to its half-way point and adjust *R3* until you hear noise from the receiver. Now, set *R2* to its half-way point and adjust *R3* until the noise is just barely audible. After this adjustment, *R2* becomes a fine squelch adjust and can be set until the noise disappears; and when the signal comes in, the receiver will not be squelched. The spot where the squelch is most sensitive is where background noise just disappears. ◈



PARTS LIST

Q1—For pnp receivers; most any npn switching transistor (2N5129, 2N3904, 2N4123) For npn receivers; most any pnp switching transistor (2N5139, 2N3906, 2N4125)
R1—2700-ohm, ½-watt resistor
R2—2000-ohm linear-taper potentiometer
R3—10,000-ohm linear-taper potentiometer

Circuit, with minimum of parts, can be added to existing receiver. Emitter resistor of set's audio driver is cut out of circuit and replaced by R1. Potential at top of R1 is controlled by Q1. With no signal, Q1 is on and audio stage is off. When signal is received, Q1 is turned off by charge on Cx and R1 completes driver stage.



COMMUNICATE OVER LIGHT BEAMS with the FIRST SINGLE-LED Transceiver

Reduces cost and simplifies construction

BY FORREST M. MIMS

THE EVOLUTION of the new light-beam communicators has opened a whole new vista in modern optoelectronics. So far, all previous light-beam communicators have required separate light sources and detectors for proper operation. Now, for the first time, it is possible to build an optical communicator that uses a single semiconductor diode as both source *and* detector of near infrared radiation.

The semiconductor source/detector is an ordinary light-emitting diode (LED), a semiconductor device designed for the efficient generation of visible or infrared light. What has not been bruited about is that LED's, just as most semiconductor diodes, can be made to detect as well as generate light; so, LED's can be used as detector elements, too.

Using a single LED as both source and detector provides certain important advantages. As can be seen in the accompanying photos, the POPULAR ELECTRONICS Infrared Transceiver employs only one lens, a feature not found in any other present light-beam communicator. Besides reducing the cost and simplifying the construction procedure, a single lens simplifies optical alignment between two infrared transceivers. Of even more significance is the fact that the entire front of the transceiver can be taken up by the

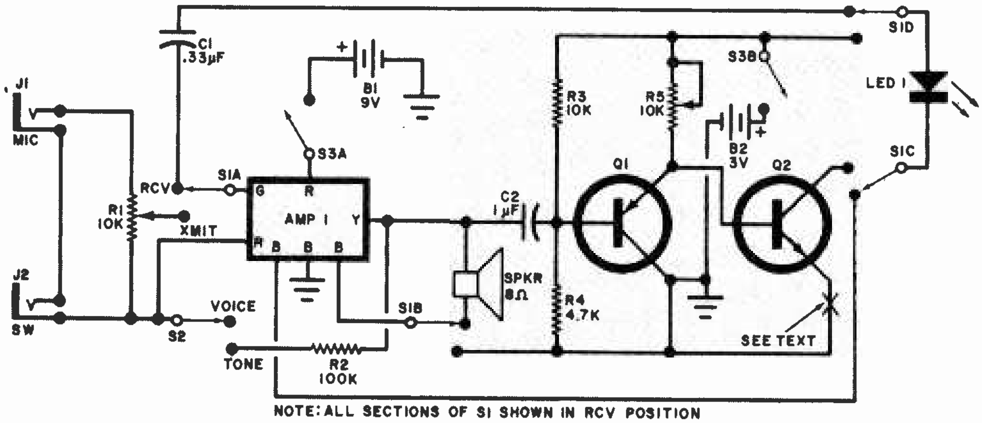


Fig. 1. A single LED is used for both transmitting and receiving. A two-transistor current modulator and a commercial audio amplifier complete circuit.

PARTS LIST

AMP1—Audio amplifier module (see text)
B1—9-volt battery
B2—Two 1.5-volt AA cells
C1—0.33-µF, 10-volt capacitor
C2—1-µF, 10-volt unpolarized capacitor
J1—⅛" miniature phone jack
J2—3/32" subminiature phone jack
LED1—Light-emitting diode (see text)
Q1—2N2907 transistor
Q2—TI P33 power transistor
R1, R5—10,000-ohm miniature trimmer potentiometer
R2—100,000-ohm, ¼-watt resistor
R3—10,000-ohm, ¼-watt resistor
R4—4700-ohm, ¼-watt resistor
S1—4pdt rotary switch
S2—Spdt switch
S3—Dpdt switch
Misc: Chassis box (see text); lens (see text); battery holders; speaker (miniature 8-ohm); low-impedance microphone with built-in switch; perforated phenolic board; grommets; hookup wire; solder; etc.

lens. This results in narrower beamwidths and much higher light-collection efficiency than is obtained with conventional dual-lens systems installed in an identical amount of space.

Construction. Assembling the infrared transceiver is a straightforward job. A 6" × 4" × 3" hinged steel chassis box is ideal for the project, but other boxes of similar size will do. As shown in Fig. 2, begin fabrication by drilling the holes to accommodate the panel switches and jacks. The amplifier, modulator, speaker, and battery holder come next. (Note: Don't forget to drill holes to permit the sound from the speaker to escape.)

The hole for the lens is best cut with a 2" chassis punch. But if such a large punch is not available, you can drill a 2" circle of small holes, knock out the center, and use a file to smooth the edges of the opening.

When you mount the switches and jacks, as shown in Fig. 3, make them only finger tight. It may be necessary to remove some of these parts during soldering to facilitate easy connection of hookup wire. When you install *R1*, be sure its lugs are easily accessible since connections will be made to all three.

Mount the modular amplifier with four 4-40 × ½" machine screws and nuts, sandwiching, between the board and chassis box, small rubber grommets at all four locations to serve as spacers. A Radio Shack No. 277-1240 four-transistor modular amplifier was used in the prototype, but any general-purpose audio amplifier can be used as long as impedance matching between microphone and speaker is observed.

Use two sets of 4-40 × ¼" machine hardware to mount the dual AA-cell holder (see Fig. 2). The 9-volt battery that supplies power for the amplifier can be mounted between two 8-32 × 1" screws, being held in place by a metal or plastic retainer and two 8-32 nuts.

The LED modulator is so simple that a perforated phenolic board can be used as the assembly medium, using Fig. 4 as a guide to parts layout. Note how potentiometer *R5* is mounted with its adjustment screw facing upward for easy access. Mount the LED to one side of the board, at the midpoint of the cabinet's vertical dimension. Then use a pair of L brackets and some 4-40 machine hardware to attach the modulator assembly. Be sure the mounting holes orient the LED at the horizontal midpoint of the cabinet. (Before the mounting holes are

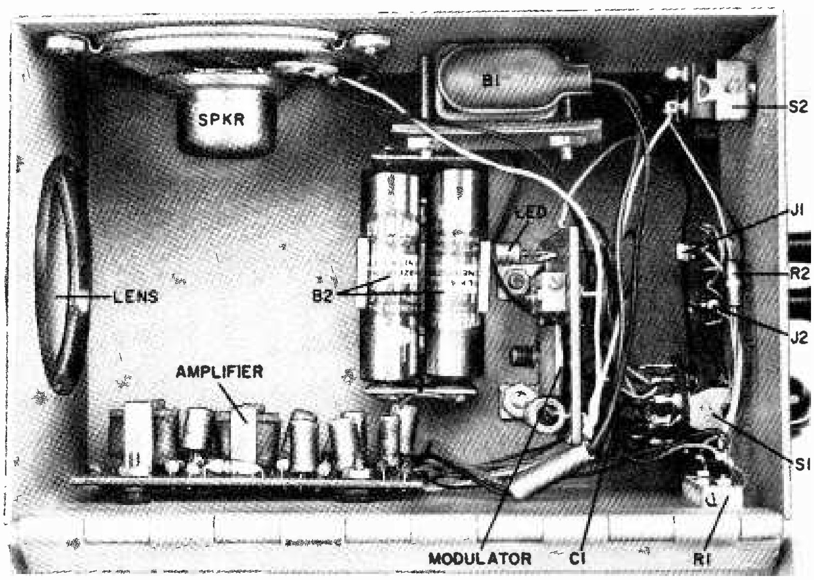


Fig. 2. Make sure that none of the components gets in way of light path between the LED and lens. Photo is of prototype.

drilled, measure the focal length of the lens so that the LED can be placed approximately at the focal point.)

The lens used in the prototype is made of red plastic to filter out unwanted light when the devices are in the RECEIVE mode. Its 2" diameter and 4" focal length give an *f*-number of 2, which is very inefficient in the TRANSMIT mode since only about 20 percent of the infrared radiation from the LED is collected. Somewhat more radiation can be collected by a lens with a focal length similar to its diameter. (A great variety of lenses is available from Edmund Scientific Co., 300 Edscorp Bldg., Barrington, NJ 08007. You might write for a catalog to find out what they have.) Do not mount the lens at this time.

Mount the speaker with three sets of 4-40 × ¼" screws and nuts. You will have to make three small tabs from 1/16"-thick aluminum stock.

Complete internal assembly by soldering all connections, referring to Fig. 1 as you proceed. Use particular care when soldering to *S1* since an error will cause the transceiver to malfunction. Note that *R2* is soldered directly between *S2* and *C2*, and *C1* is soldered directly to *S1*. It may be necessary to extend some of the leads from the amplifier module.

Assembly of the project is completed with the mounting of the lens and labeling of the controls. You can use GE Silastic silicone cement to mount the lens. Ideally, the lens should be cemented to the inside of the chassis box for best external appearance. Use dry-transfer letters to label the controls.

LED Selection. Any LED will operate as both a source and a detector in the finished transceiver, even visible red, yellow, and green units. For best results, however, use a silicon compensated near-infrared LED made from gallium arsenide (GaAs). These LED's are by far the most efficient available. Note that all LED's emitting 930 to 940 nanometers (9300-9400 angstroms) are silicon compensated. For the prototype, a General Electric SSL-55C was used. This is one of the most efficient LED's commercially available.

Current compensation may be required for the LED. As designed and shown in Fig. 1, the circuit will operate with LED's capable of handling 100 mA continuously without a heat sink. Most metal-glass-packaged LED's are rated at this current level. If a low-current LED is used, a current-limiting resistor must be installed at point X in Fig. 1. Determine the value needed for the resistor by temporarily installing a 100- or 500-ohm potentiometer at X and a 0-100-mA meter movement in series with the pot. Set *R5* at about midpoint to allow for circuit adjustment and adjust the potentiometer until the milliammeter indicates the maximum allowable LED current. Without disturbing its setting, disconnect the potentiometer and measure its resistance. Solder an equivalent fixed resistor into the circuit at point X.

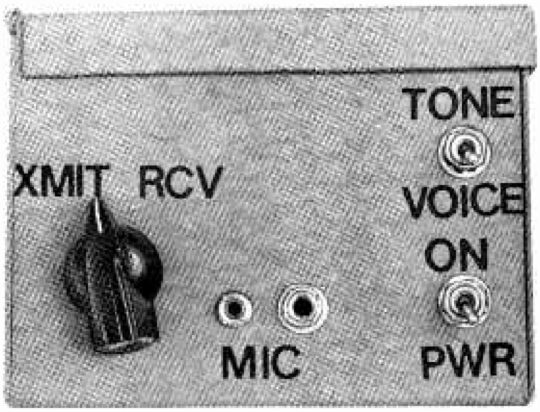


Fig. 3. Suggested front-panel arrangement.

Transceiver Operation. Unless you have a conventional amplitude-modulated LED communicator, it will be necessary to build two transceivers to test the circuit. Plug in a low-impedance microphone and set *S1* to XMIT and *S2* to TONE. An audio tone should be heard from a second transceiver pointed toward the first when *S1* is set to RCV. If no tone is heard, check all battery connections and the batteries themselves to insure that they are fresh. Then check the wiring, paying particular attention to the connections made to *S1*. When the transceivers are operating properly, reverse both *S1*'s and check RCV/XMIT operation.

When two-way operation has been verified, check voice operation by repeating the above procedure with *S2* in the VOICE position. If the receiver seems to be overmodulated, try slightly misaligning the two units to reduce the amount of IR radiation falling on the LED in the detector mode. If volume is too low, try adjusting *R1* in the transceiver set to the XMIT mode. Reverse *S1* in both units and repeat the test.

The initial tests should be followed by a check of LED current to avoid possible overheating of the LED in the XMIT mode. This is easily done by tem-

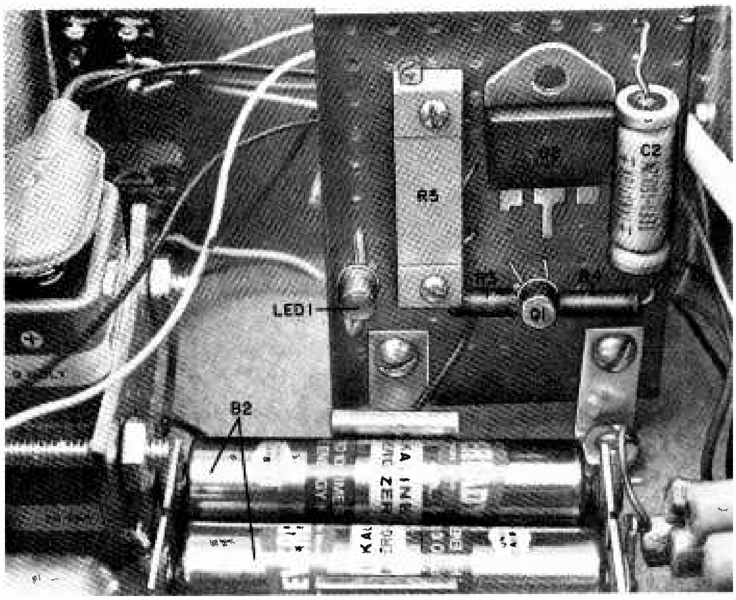


Fig. 4. Current modulator is built on perf board. LED must align with lens and adjustment R5 must be available after mounting the board.

HOW IT WORKS

Despite its novel detection scheme, operation of the infrared transceiver (Fig. 1) is straightforward. In the XMIT mode, a commercial solid-state amplifier (*AMP1*) is connected to a two-transistor (*Q1* and *Q2*) current modulator via XMIT/RCV switch *S1*. Audio signals from the amplifier are fed into the modulator via *C2*, and *Q1* and *Q2* provide linear modulation over a range greater than 75 percent.

LED's are current-sensitive devices. The peak current through *LED1* is normally determined by the setting of *R5*. Since a variety of LED's can be used in the circuit, additional current control may be necessary to prevent exceeding device specifications.

To simplify the alignment of two transceivers, the transmitter circuitry is provided with *R2*, which causes feedback oscillation when connected from the output to the input of the amplifier via TONE/VOICE switch *S2*. With *S2* set to TONE, the transmitter generates a tone whose frequency can be changed from a low to a high pitch by disconnecting the microphone at *J1* from the circuit with its self-contained switch plugged into *J2*.

In the RCV mode, the same LED used to transmit the optical signal is switched to the input of the modular amplifier via *S1*. Capacitor *C1* blocks undesirable dc signals from *LED1* from getting to *AMP1*. In the RCV mode, the modulator circuit is disconnected from the power source to conserve battery power.

Incoming optical radiation striking the sensitive surface of *LED1* generates a photo-current that is proportional to the amplitude of the signal modulations. The photo-current is amplified by *AMP1* and passed to a miniature 8-ohm speaker.

porarily inserting a 0-150-mA meter in series with the LED at point X. Alternatively, connect the meter in series with one of the batteries by removing one cell and using clip leads to connect cell and meter to the holder; be careful to avoid a short circuit.

The current reading should not exceed the peak allowable current through the LED if a heat sink is not used. If the current is too high, adjust *R5* to reduce it; if well below the peak allowable value, again adjust *R5* to bring it up. A quick test for excess current can be made by touching the LED. If it is hot, turn off the power immediately and adjust *R5* to reduce the current. It may be necessary to insert a permanent limiting resistor at point X as described.

Range Testing. Place one transceiver on a steady support and point it along a path unimpeded by obstacles for several hundred feet. Set *S1* of this transceiver to XMIT and *S2* to TONE. Set *S1* in a second transceiver to RCV. Now, walk about 15 feet away from the first transceiver, pointing the second one toward the first until a tone is heard. Due to the very narrow field of view of the receiver and the tight beam of the transmitter, alignment will be difficult at first. This highly directional nature of reception illustrates the significance of optical communications —totally private jamproof transmissions.

Complete the testing by walking away from the transceiver with the receiver while listening to the tone. Daylight range will be shorter than night range due to the increase in detector noise caused by ambient light.

Modifications. Although the Infrared Transceiver can be used as is, it lends itself to several interesting modifications. First, for permanent field installations, mount each unit on a tripod. This will greatly ease optical alignment and make possible continuous transmissions with few realignment problems.

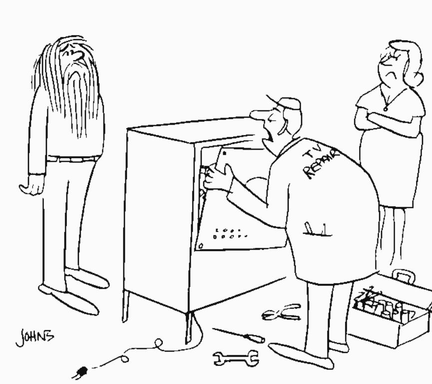
For more range, increase the size of the lens. The light-collecting area of a lens is proportional to the square of its diameter; so a small increase in diameter yields a significant increase in receiving area. A lens 3" in diameter has more than twice the collecting area of a 2" lens. A diverging beam of light follows the inverse square law. Therefore, doubling the lens collection area will, in theory, double the range. But due to variations in atmospheric absorption amd ambient light, doubling lens area will not necessarily double the range.

The most interesting modification of all is to connect two transceivers together with a single fiber-optic link. A transceiver can be converted for both atmospheric and fiber-optic operation by mounting the LED to the modulator board with a miniature phone plug and jack. For fiber-optic operation only, the LED's become an integral part of a single fiber-optic assembly.

A hole bored through a vacant corner of the front of the transceiver will facilitate installation of the optical fiber link. Use a rubber grommet to line the hole to protect the fiber from damage. For best results, choose a length of large-diameter (40-mil) fiber. Remove the caps from the LED's and place a layer of optically clear epoxy over the chip and cement the fiber as close as possible to the chips with the epoxy. Secure the assembly in a fixed position until the epoxy has fully set. Exercise care during the epoxying operation to avoid damaging the delicate LED chip and electrodes. Solder the leads of each LED to a miniature phone plug and pack the connections with more epoxy to make a rigid, durable assembly.

The fiber-optic mode of operation is a precursor of what telephone systems of the future are likely to resemble. For this reason, the Infrared Transceiver is an entertaining, educational, and highly functional project. ◈

(Editor's Note: The author is pursuing patent protection for concepts described in this article. However, readers may build the project for personal use.)



"I found your trouble—there was a hair in the gears."

APPLICATIONS FOR THE 555 IC

Some interesting circuits using the 555 timer-on-a-chip

BY WALTER G. JUNG

NOW that you understand the operating principles of the 555 (see "The IC Time Machine" elsewhere in this Handbook), let's see how it can be put to work in five practical circuits.

These circuits in no way exhaust the applications in which this versatile timer-on-a-chip can be used. We have barely scratched the surface with the circuits given here, but hope that we have suggested some new ideas that will be useful in designing other projects.

In a case like this, the best approach to new circuit design is to understand fully the internal workings of the IC itself and know just what the inputs and outputs are at each pin. Then let your imagination go to work. The best way to do this is to make a "breadboard" and play with the IC, using different connections and varying the external components.

A WARBLE ALARM CIRCUIT

The warble alarm circuit shown in Fig. 1 used two 555 IC's as an audible attention getter. The first 555, *IC1*, oscillates at a frequency slightly below 10 Hz. Its rectangular output is filtered by *R1C1* to produce a triangle wave, which in turn is used to frequency-modulate *IC2*. The latter is operated at approximately 1 kHz and is modulated at a 5-Hz rate.

The output current of the 555 IC is sufficient to drive a small speaker and *R2* is used to prevent excessive loading, but the audible level of the tone produced is still quite noticeable. The exact frequency, rate, and deviation of the circuit can be easily modified to produce almost any type of warble sound desired. The "on-off" control is most efficiently utilized by interrupting the supply line so as to minimize standby power. The switch can be relay contacts or some other means of applying power when an alarm condition is sensed.

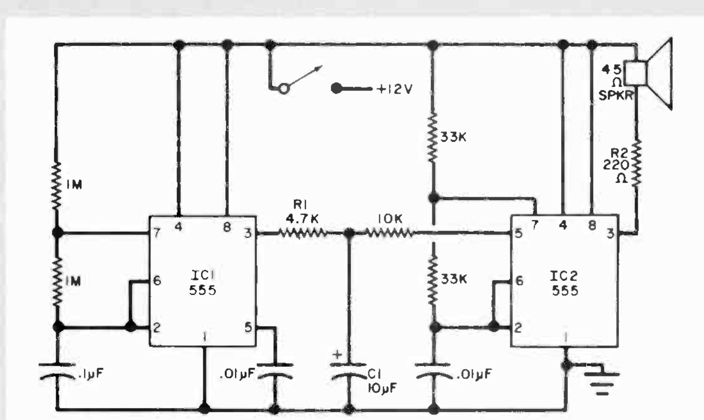


Fig. 1

SCHMITT TRIGGER OR BISTABLE BUFFER

Aside from its basic use in timing functions, the 555 IC can be applied to advantage to other switching circuits. One example is the Schmitt trigger circuit shown in Fig. 2. In this circuit, the two comparator inputs (pins 2 and 6) are tied together and biased at half of the applied dc voltage through the voltage divider made up of *R1* and *R2*. Since the upper comparator (pin 6) will trip at ⅔ of the applied dc and the lower one at ⅓ of the applied voltage, the bias provided by resistors *R1* and *R2* is centered within the comparator's trip limits.

A sine-wave input of sufficient amplitude to exceed the reference levels causes the internal flip-flop to be set and reset. In this way, it creates a square wave at the output. As long as *R1* is equal in value to *R2*, the 555 will be automatically biased correctly for almost any supply voltage. Note that the output waveform as shown in the diagram is 180 degrees out-of-phase with the applied input sine wave. Because of the 555's high output current capability, the circuit can be used to good purpose as a signal shaper/buffer circuit.

Such a circuit can also find application if you have a sine-wave-only audio generator and you would also like to have a simultaneous square-wave output. The major advantage of this circuit is that, unlike a conventional multivibrator type of squarer, which divides the incoming frequency in half to square it, the Schmitt trigger simply squares the input frequency without changing the frequency. A circuit of this type can easily be installed within almost any audio generator.

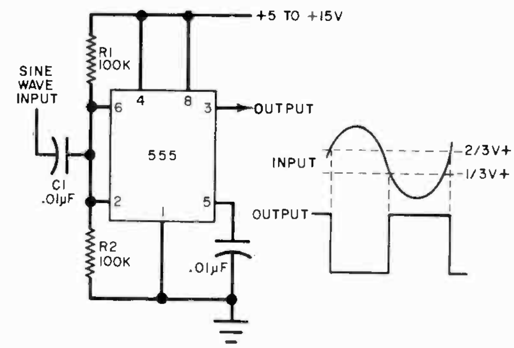


Fig. 2

Inverting Bistable Buffer. By modifying the input time constant of the circuit shown in Fig. 2 (reducing the value of input capacitor *C1* to 0.001 μF, for example) so that input pulses will be differentiated, the arrangement can also be used either as a bistable device or to invert pulse waveforms. In the latter case, the fast time-constant of the combination of *C1* with *R1* and *R2* causes only the edges of the input pulse or rectangular waveform to be passed. These pulses set and reset the flip-flop; and a high-level, inverted output is the result.

SQUARE-WAVE OSCILLATOR

A conventional astable circuit using a 555 IC does not normally produce a symmetrical output waveform. However, square waves can be obtained from a 555 by using the simple circuit shown in Fig. 3.

The asymmetry of a conventional astable circuit is the result of the fact that the charging and discharging time constants are not equal. If the timing capacitor can be charged and discharged through the same (or equivalent) resistance value, the symmetry can be restored.

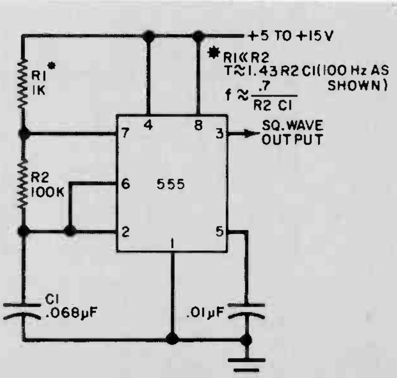


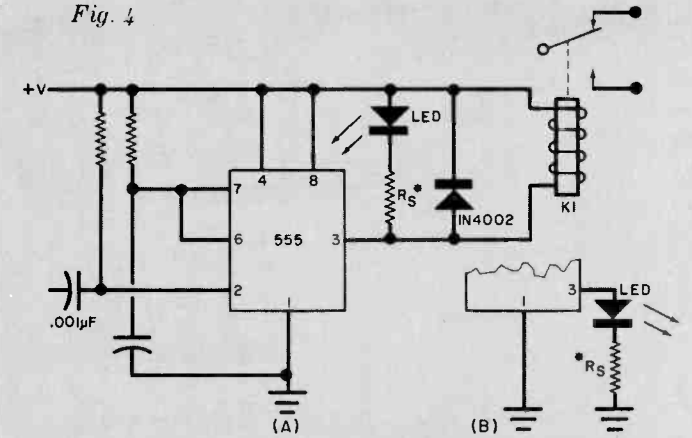
Fig. 3

In the circuit shown, capacitor *C1* is charged through *R1* and *R2* and it is discharged through *R2*. If *R1* is made very small in resistance compared to *R2*, then both time constants will be reduced so that they depend essentially on *R2* and *C1*.

The frequency of operation (*f*) of this circuit is approximately equal to 0.7 divided by the product of *R2* and *C1*. The frequency is, of course, independent of the supply voltage.

OUTPUT DRIVE CONSIDERATIONS

Fig. 4



*SELECT FOR DESIRED LED CURRENT. +5V USE 150Ω, +15V USE 680Ω

The 555 timer IC can provide up to 200 mA of output current in either its high or low state. However, this value should not be considered too strictly since some types of loads have a voltage limitation. If, for example, the 555 is used with a 5-volt supply to drive TTL logic, the output current is limited to much less than 200 mA because of the required input voltage for the following TTL stage. Since TTL output stages are normally specified for 0.4 volt at rated current, a more realistic maximum output current for the 555 is 5 mA, which is far less than the 200 mA specified.

Other types of loads, such as incandescent lamps, relays, or light-emitting diodes are not as critical in terms of voltage and they can be driven by using the circuit shown in Fig. 4. Depending on the logic involved in the application, these types of loads can be connected from pin 3 to either +V or ground. In a timer such as that shown in (A), the output (pin 3) is normally at the ground potential and goes high during the timing interval. Therefore, a LED connected as shown at left will be on when pin 3 is low, and it will go off when pin 3 is high (during the timing cycle).

Since a 555 can operate over a wide dc supply range and a light-emitting diode requires about 1.6 volts, a series resistor (*Rs*) is used to drop the excess voltage and limit the LED current.

Relays can be driven as shown in this circuit by selecting a relay that is compatible with the applied dc. Of course, it will have to have the contact arrangement desired. Since the 555 has a healthy current output, the relay selected need not be particularly sensitive.

This permits relays rated at 12 volts and 100 mA to be used. The diode across the relay coil is used to prevent the back-emf from damaging the IC chip. If the current demand is not too high, both an LED and a relay can be used at the same time.

The connections shown in (B) are for the opposite type of logic where the LED is normally off and is pulsed on.

WIDE-RANGE PULSE GENERATOR

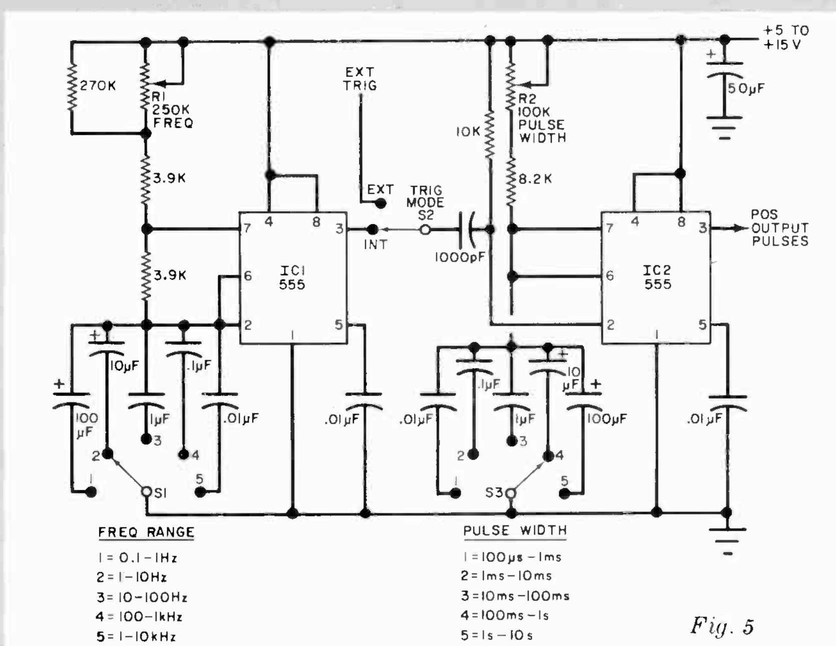


Fig. 5

The most sophisticated of the 555 applications described here is the wide-range pulse generator, whose circuit is shown in Fig. 5.

The general-purpose pulse generator consists of an astable oscillator (*IC1*) whose output frequency can be varied over a 10:1 range by potentiometer *R1* (frequency control). Range selection is made by *S1*, with five ranges from 0.1 Hz to 10 kHz. Tantalum capacitors are used for the two lower ranges, while Mylar capacitors should be used for the upper ranges. The output of *IC1* feeds *S2*, which can be used to select either internal or external signals for *IC2*, a monostable circuit.

Integrated circuit *IC2* is a monostable generator whose output is a pulse with a width that can be varied over a range of 10 to 1 by changing *R2*. Switch *S3* provides five ranges from 100 microseconds to 10 seconds. The output of the latter stage consists of positive-going pulses whose frequency (rate) and width can be set to almost any desired values. If the external mode of triggering *IC2* is selected, almost any negative-going pulse can be applied to the external trigger input. ◈

HOW TO BUILD "FREE-POWER" RADIOS

Successors to crystal radios use single high-gain transistor amplifier.

BY TERRY L. LYON

EXPERIMENTERS and hams have liked to fool around with battery-less radios since wireless communications were first considered. Although notable improvements have increased the sensitivity and selectivity of the devices, their performance is limited unless the newest design techniques are used. Described here are three battery-less receivers which have improved gain as a result of using a simple transistor amplifier powered by random electrical fields which are everywhere. These circuits, which are relatively inexpensive to build, have higher volume and better reception than a crystal radio.

The first circuit (Fig.1A) is a broadcast-band receiver and requires the fewest number of components. The circuit of Fig. 1B also tunes the broadcast band but it has increased gain due to a more efficient design. The circuit of Fig. 1C has improved selectivity and sensitivity due to regeneration, and it is designed to receive shortwave as well as conventional broadcast transmissions.

In the construction, although circuit layout is not critical, it is wise to keep component leads short and neat. The antenna and ground leads from the receiver could have various lengths of stranded insulated wire with alligator clips attached for connecting the receiver to large metallic objects.

If some components prove difficult to find, substitute others with similar characteristics. For example, the tantalum capacitor (*C2*) can be replaced by an electrolytic with the same specifications. The 1N459 diode can be replaced by another low-power silicon unit with small reverse-current characteristics. Likewise, another small-signal, high-gain silicon unit can be used instead of the 2N3391 npn transistor. A 4700-ohm resistor can be used for *RFC1*. Finally, the crystal earphones can be interchanged with high-impedance magnetic phones, using a suitable series capacitor.

Operation. Once the receiver is completed, a tuner dial can be added. Calibration of the dial is accom-

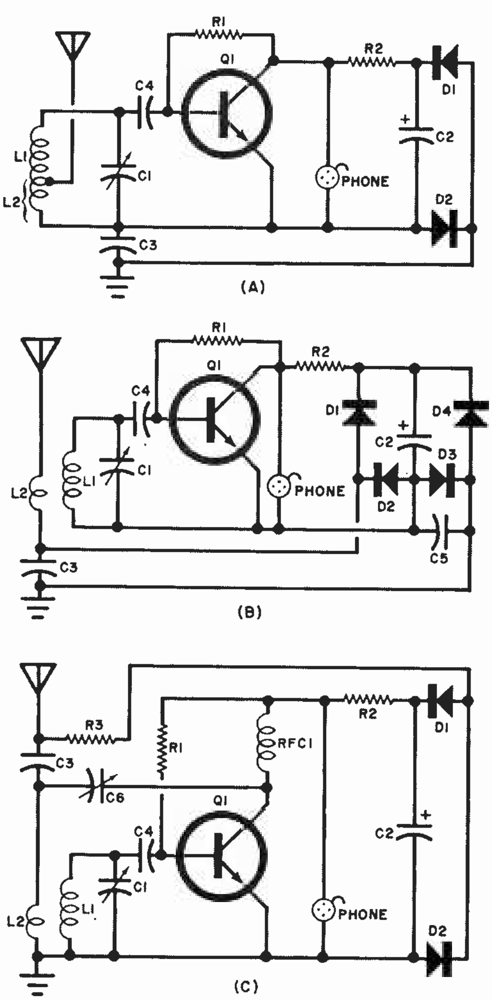


Fig. 1. Three versions of simple single-transistor radios that derive their operating power from the random electrical noise that is usually found in the atmosphere

PARTS LIST

C1, C6—365-pF variable capacitor
C2—5-µF, 50-volt tantalum capacitor
C3—0.002-µF ceramic disc capacitor
C4, C5—0.005-µF ceramic disc capacitor
D1—D4—1N459 silicon diode
L1—Fig. 1A; tapped transistor antenna coil
 Fig. 1B; transistor antenna coil
 Fig. 1C; see Fig. 2
L2—Fig. 1B; 15 to 20 turns of #24 enameled wire wound directly over antenna coil.
 Adjust turns or reverse leads for optimum performance.
 Fig. 1C; see Fig. 2
Q1—2N3391 transistor
R1—10-megohm resistor
R2—470,000-ohm resistor
R3—10,000-ohm resistor
RFC1—2.5-mH r-f choke
Phone—Crystal earphone

plished by listening to stations which have a known transmitting frequency or by coupling a variable r-f signal generator to the receiver through a suitable antenna. If the receiver is not operating in the specified range, adjust the core of *L1* in the first two circuits or add or remove a few turns from *L1* in the third circuit.

To operate the third circuit (Fig. 1C) advance (counterclockwise) the regeneration control (*C6*) until a slight hiss is heard. The proper position of *C6* depends on the length of the antenna, the receiver coil, and the position of the tuning capacitor (*C1*). However, the receiver may not operate with regeneration at high frequencies, but *C6* will serve to boost the receiver's performance. Shortwave reception is obtained by changing coils in accordance with Fig. 2.

For optimum performance, these receivers require a good earth ground and a large metallic antenna. Water pipes and other low-lying metallic objects make good grounds. The antenna lead can be clipped to a window screen, roof gutter, refrigerator, or similar item. Sometimes just touching the antenna lead with the hand is sufficient to power the receiver. To increase reception, attach a 9-volt battery across *C2*, observing the correct polarity.

For listening to weak signals, connect two earphones in parallel to form a headset. Local stations in the broadcast band may interfere with distant transmissions. If so, a series *LC* circuit may be constructed to remove the unwanted station. This circuit connects between the receiver's antenna and ground and is built using a standard antenna coil connected in series with a 365-pF variable capacitor. When this circuit is tuned to the interfering frequency, the latter will be effectively removed. However, the antenna coil must be kept away from *L1* and the chassis of the capacitor should be connected to ground. ◈

HOW IT WORKS

The noise and signal are separated by coupling the series *L2C3* resonant circuit to the parallel *L1C1* resonant circuit. This arrangement functions as a bandpass filter, allowing broadcast information to appear across *L1C1* while leaving the noise across *L2C3*. When *L1C1* is adjusted to a standard broadcast frequency, an amplitude-modulated carrier is produced across the tuned circuit. This r-f signal is sent through dc blocking capacitor *C4* to the base-emitter junction of transistor *Q1*, a common-emitter amplifier.

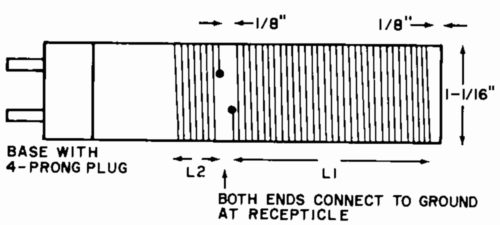
The transistor is biased by a large value of shunt feedback (*R1*) and its load resistance (*R2*) also has a large value. This arrangement performs several functions. First, the voltage drop across the base-emitter junction is quite small. This allows the junction to detect the incoming signal by changing it to modulated dc. Although the shunt feedback biasing arrangement lowers *Q1*'s input impedance, its emitter current is so small that the input impedance is still very large and does not appreciably load the tuned circuit.

Second, the transistor is biased in a region of extremely high gain and some non-linearity. The latter acts to a small degree as an agc. When signals get larger, the amplifier's gain is reduced whereas, on weak signals, the gain is large.

The power supply for the transistor derives its energy from the noise obtained across *L2C3*. This noise derives primarily from a 60-Hz field radiated from household wiring, lights, and appliances. The noise is rectified by *D1* through *D4* and the resulting dc is filtered by *C2*. Limiting resistor *R2* connects the supply to the transistor circuit.

Although the three receivers operate in basically the same manner, there are several differences among them. The first two rectify voltage fluctuations (low-frequency noise) appearing across *C3*. The first circuit has a voltage-doubling diode arrangement to reduce the number of components. On the other hand, the second circuit utilizes a full-wave bridge rectifier with improved efficiency; but it requires the addition of *C5*, *L2*, and two diodes. Capacitor *C5* is used to reference the *L1C1* circuit to ground, which increases the signal and minimizes hum.

The third receiver uses a voltage doubler; however, it is connected across the *L2C3* circuit through *R3*. This arrangement allows high-frequency noise as well as low-frequency noise to be rectified with high efficiency. This receiver also has exchangeable coils so that several bands can be received. Some of the amplified signal in this circuit is returned to the input of *Q1* by *C6*, *L2*, and *RFC1*. This adds positive feedback and further increases the receiver's gain.



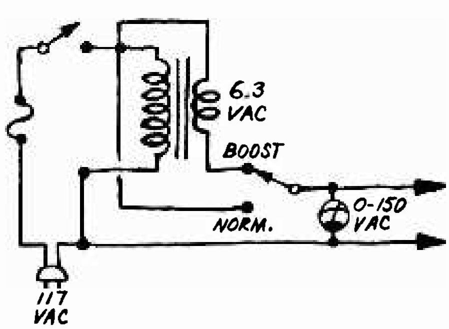
Range	Turns	Wire
540-1500 kHz	L1: 149.6 closewound	#28
	L2: 41.3 closewound	#28
1.5-4.0 MHz	L1: 49.2 even for 2"	#24
	L2: 11.2 even for 7/16"	#24
4.0-11.0 MHz	L1: 18.4 even for 2"	#22
	L2: 4.2 even for 7/16"	#22

All wire is enamel coated, wound on low-loss 1 1/16" diameter forms at least 3½ in. Use plastic pill containers or thin-wall cardboard tubing. Coat with clear lacquer, if desired, to keep wire in place.

Fig. 2. Winding details for the coils used in the circuit of Fig. 1C.

INEXPENSIVE VOLTAGE BOOSTER

At times of peak demand, power companies drop their voltage 5% to stay within safety limits of their equipment. This small voltage drop can be hard on certain appliances—television receivers, small motors, etc. Using a 6.3-V filament transformer in the circuit below allows you to compensate for low line voltage. In the Boost position, voltage is stepped up about 5.4%. Any device which draws less than the rated current of the transformer may be used. For example, a 3-A transformer can handle 330 W, enough for most color TV receivers. The fuse should have the same current rating as the transformer.



–T.R. Fox

DO YOU KNOW YOUR BIPOLAR TRANSISTORS?

TRANSISTORS BIPOLAR YOUR KNOW YOU DO

Check your knowledge of this important basic transistor theory.

BY LOTHAR STERN
Motorola Semiconductor Products Inc.

THE TRANSISTOR is the most versatile component of the semiconductor family. Its most important characteristic is current and/or voltage amplification. Because of this capability, it is the heart of most solid-state electronic circuits involving signal amplification and switching.

Transistor Types. There are two varieties of transistors: unipolar transistors, usually referred to as field-effect transistors (FET's), and bipolar transistors. The FET was invented circa 1930 but was not exploited commercially at the time. The bipolar transistor was invented around 1948 and rapidly became a practical product. The FET was finally developed as a commercial product by the early 1960's, many years after the bipolar transistor had already established itself. FET's, especially MOSFET's, are now used widely in low-noise front ends for TV and FM and are gaining importance in digital elements. However, in this article we will confine ourselves to a discussion of bipolar transistors.

Transistor Basics. The basic element of any bipolar transistor is the pn junction, or diode, formed by chemically uniting a layer of p-doped semiconductor material with a layer of n-doped material. When the junction is forward-biased (p layer positive, n

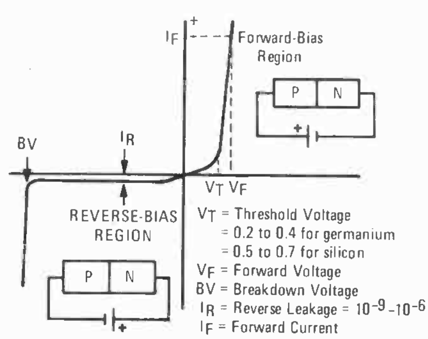


Fig. 1. Forward and reverse characteristics of typical junction diode.

layer negative), the current through the device increases rapidly as the voltage is increased. When the junction is reverse-biased, an increase in voltage causes only a very small amount of leakage current, until the reverse voltage becomes high enough to cause "breakdown" of the junction. See Fig. 1.

A transistor is formed by sandwiching a very thin layer of n-doped material between two layers of p-doped material (pnp transistor) or a thin layer of p-doped material between two layers of n-doped material (npn transistor). The characteristics of this

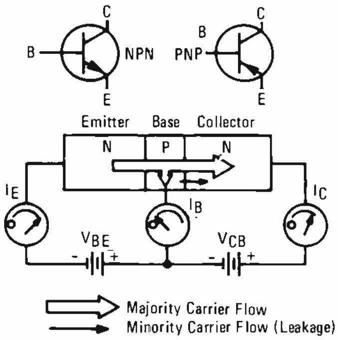


Fig. 2. Current flow in a transistor.

3-layer, 2-junction structure are achieved by varying the thickness and the resistivities and geometries of the three layers.

Theoretically, an npn transistor should operate at higher frequencies than does a pnp device because electrons (the principal current carriers of an npn unit) are more mobile than holes (the principal current carriers of pnp devices). In practice, however, both types have similar characteristics. The main difference is that pnp transistors operate with a negative voltage on the collector element while npn transistors operate with a positive collector voltage. This makes it possible to have complementary circuits that often provide improved performance over the single-polarity circuits that can be implemented with only one type of transistor.

The three layers of a bipolar transistor (Fig. 2) are the emitter, the base, and the collector. The emitter represents the current source, where the current carriers originate. The base is the control element, and the collector is the element through which the current carriers are transferred to an external circuit.

In operation, the collector-base junction is always reverse-biased. If the emitter-base junction is forward-biased, the emitter-base current is high, as in any forward-biased diode junction. Instead of flowing out of the base terminal, however, most of this current diffuses through the very thin base region and crosses the collector-base junction, with only a small fraction flowing out the base lead. This action gives rise to the two commonly used expressions for current gain in a transistor. The first, α (alpha), is the ratio of I_C/I_E when the base-emitter junction is forward-biased and is normally called the common-base current gain. It is always slightly less than unity (usually 0.95 to 0.998). The second gain expression is β (beta), the ratio of I_C/I_B called the common-emitter current gain. It can be quite high, depending on the value of α, since $\beta = \alpha/(1-\alpha)$. Thus, if $\alpha = 0.95$, $\beta \approx 20$; and if $\alpha = 0.99$, $\beta \approx 100$.

Circuit Configurations. There are three basic transistor circuit configurations: common-emitter, common-base, and common-collector. They differ principally in the manner in which the signal is applied to the transistor and where the load is attached. Fig. 3 shows these basic circuits. Since the common-emitter circuit is by far the most prevalent, most data sheets characterize the transistor in terms of this circuit.

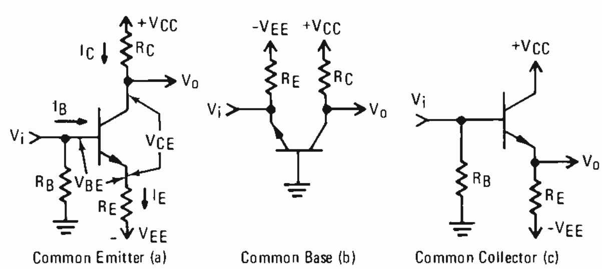
Fig. 4 shows the characteristics of a typical transistor in a common-emitter circuit. The input circuit curve in quadrant III shows that a base-emitter voltage (V_{BE}) of less than approximately 0.5 volt (for silicon; for germanium, about 0.2 volt) causes virtually no base or collector current to flow, thereby keeping the transistor cut off. Above 0.5 V, I_B rises sharply, limited only by the ohmic resistance of the base region. Since the latter is very small, a very small rise in V_{BE} (beyond the threshold voltage) causes a large rise in I_B and I_C, as seen in quadrant II. From such a plot, the transistor current gain, I_C/I_B, can be established. It is also evident that considerable nonlinearity exists in the I_C/I_B curves. This indicates that distortion for large-output signals can be quite high.

The collector curves in quadrant I are the most significant. They show, for example, that the maximum reverse-bias collector voltage that can be applied, before breakdown of the collector-base junction, is BV_{CEO}. The minimum collector voltage needed to keep the collector junction reverse-biased, thereby sustaining transistor action, is $V_{CE(sat)}$. The maximum voltage excursion, therefore, is $BV_{CEO} - V_{CE(sat)}$.

When the transistor is cut off ($I_B = 0$), a residual current, I_{CEO} flows in the collector circuit. This leakage current can be reduced somewhat (but not to zero) by applying reverse-bias to the emitter junction. The limiting value, with reverse-bias applied to the emitter-base junction, is I_{CBO}. This is equivalent to the collector-base leakage current when the emitter is open-circuited.

Collector current increases rapidly as the base is energized. The maximum I_C is that which would damage the internal transistor structure. This value of collector current is given as a maximum rating on data sheets. Thus, the output current could range from $I_{C(max)}$ to I_{CEO} (or I_{CBO} in the event of reverse-bias), but usually I_C is limited to a value far less than $I_{C(max)}$.

Power dissipation, P_D is the product of V_{CE} and I_C and it causes the collector junction to heat up. Beyond a critical junction temperature, $T_{J(max)}$, the device could be damaged. Thus, the $P_{D(max)}$ rating of a transistor limits the maximum V_{CE} and I_C that can be applied simultaneously. The locus of a $P_{D(max)}$ rating is a parabolic curve on the V_{CE}/I_C plot. The load line, XQY,



$V_{CE} \approx V_{CC} - I_C R_C \quad V_{BE} \approx 0.7$ V (for silicon)

Circuit	Current Gain	Voltage Gain	Input Resistance	Output Resistance
(a)	High	High	Low	High
(b)	< 1	High	Very Low	Very High
(c)	High	< 1	High	Low

Fig. 3. Three basic transistor circuits and their characteristic parameters.

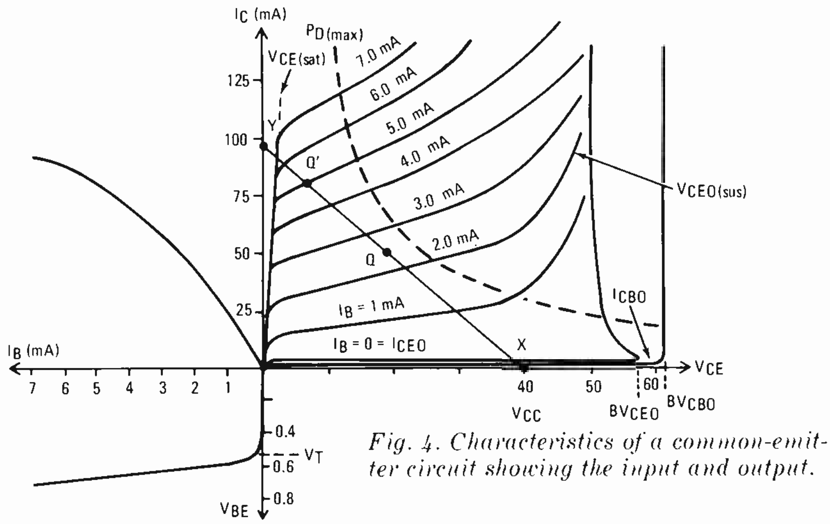


Fig. 4. Characteristics of a common-emitter circuit showing the input and output.

must be chosen to prevent steady-state operation to the right of the locus.

From the above it is evident that, when the voltage applied between base and emitter is less than approximately 0.5 V, there is only a very small current flow through the transistor. Between collector and emitter, therefore, the transistor represents a very high resistance, or open switch. When V_{BE} is raised above the threshold voltage, the internal resistance of the transistor drops to a very small value so that the device approximates a closed switch. As shown in Fig. 4, it takes a change of only 7 mA of I_B along loadline XQY to cause V_{CE} to change from V_{CC} to $V_{CE(sat)}$, that is, from an open switch to a closed switch. This switch action is accomplished by a V_{BE} of only a few tenths of a volt.

Biasing. When operated as an amplifier, the transistor must first be biased to some quiescent value of collector current, so that both positive- and negative-going input voltage excursions will cause corresponding changes in output voltage and current. The ideal bias point is represented by Q on the loadline in Fig. 4 because this permits approximately equal excursions in I_C and V_{CE} in both directions along the loadline without signal clipping. The bias point is established by a quiescent base current that results in a dc collector current of approximately $I_{C(sat)}/2$.

Several circuits are used for establishing the bias point. Among the most familiar are those in Fig. 5. The basic performance difference is in the bias-point stability. At point Q on the loadline, the transistor has a beta of approximately 20. If a transistor with a beta of 40 were substituted (simulated by dividing all I_B values by 2), and if I_B were held by the bias circuit to 2.5 mA, as before, the operating point would move up the loadline to point Q', a much higher value of I_C. As a result, considerable distortion would occur for high-value input signals.

The bias point stability factor (S) is defined as the percent-change in I_C for a percent-change in β, or $\Delta I_C/I_C = S\Delta\beta/\beta$. If a percent-change of β causes a corresponding percent-change in I_C, the least desirable condition, then $S = 1$. If I_C is independent of β (corresponding to a zero change in I_C when β is varied), then S = 0. The formulas accompanying Fig. 5 give I_C and S as functions of β and assign values for S under specific operating conditions. The bias arrangements in Fig. 5c and 5d using emitter degeneration are preferred because by proper choice of resistor values, the effect of β on I_C can be made almost negligible. This prescribes a large value of R_E, so that the voltage, I_ER_E, at the emitter is much larger than V_{BE} or I_BR_B. To prevent degenerative ac feedback, R_E is normally bypassed by a large-value capacitor. (Fig. 5c is used when a positive and negative power supply is available. For single-supply operation, Fig. 5d is preferred.)

In practical transistor amplifiers (RC-coupled amplifier, for example), the operating point is influenced by both dc and ac conditions. Fig. 6 shows a typical RC-coupled amplifier and its representative loadline plot. Note that there are two loadlines—a dc loadline whose slope is affected only by the value of R_C and an ac loadline whose slope is determined by r_L, the equivalent resistance of R_C and R_L.

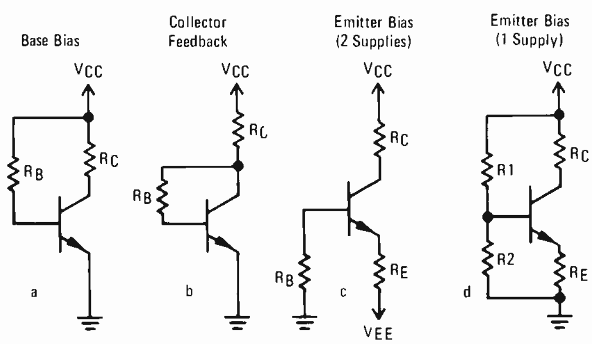


Fig. 5. Conventional common-emitter bias circuits. Table gives approximate characteristic expressions.

	a	b	c	d
I_C →	$\beta \frac{V_{CC}}{R_B}$	$\frac{V_{CC}}{R_C + R_B/\beta}$	$\frac{V_{EE} - V_{BE}}{R_E + R_B/\beta}$	$\frac{R2}{R1 + R2} \times \frac{V_{CC}}{R_C}$
S →	1	$\frac{1}{1 + \beta R_C/R_B}$ for $R_B = \beta R_C$, $S \approx 0.5$	$\frac{1}{1 + \beta R_E/R_B}$ for $R_B = R_C$, $S \approx 1/\beta$	$\frac{1}{1 + \beta R_E \frac{R1R2}{R1 + R2}}$
To Make $I_C = \frac{I_{C(sat)}}{2}$	$R_B = 2\beta R_C$	$R_B = \beta R_C$	$R_E = 2 \frac{V_{EE}}{V_{CC}} R_C$	$\frac{R1}{R2} = 1 + 2 \frac{R_C}{R_E}$

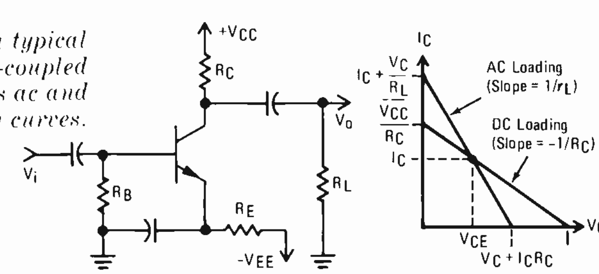


Fig. 6. Circuit of a typical common-emitter RC-coupled amplifier and its ac and dc loading curves.

MID-FREQ. EQUATIONS

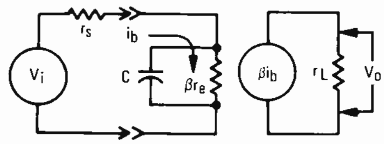
$$i_b = \frac{V_i}{r_s + \beta r_e}$$

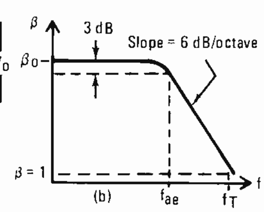
$$V_o = \beta i_b r_L$$

$$\text{Voltage Gain} = \frac{V_o}{V_i} = \frac{\beta r_L}{(r_s + \beta r_e)}$$

$$\text{Input Resistance} \approx \beta r_e$$

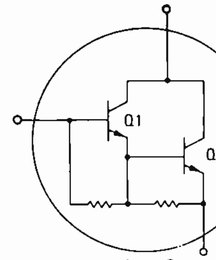
$$\frac{25\ \text{mV}}{I_E} < r_e < \frac{50\ \text{mV}}{I_E}$$





β_o = Low-Frequency β
f_{ae} = β Cutoff Frequency
f_T = Current Gain – Bandwidth Product
= Frequency where $\beta = 1$

Fig. 7. Equivalent high-frequency common-emitter circuit (a) and its response (b).



If $\beta_{Q1} = \beta_{Q2}$:
$r_i = \beta^2 r_e$
$\beta_{total} = \beta^2$

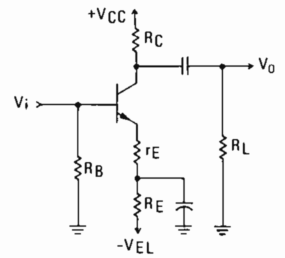
Fig. 9. Darlington transistor pair.

in parallel. The dc loadline represents the path along which the operating point can be established. The ac loadline intersects the dc loadline at the operating point, and the actual signal varies along the ac loadline, which sets the V and I output limits.

The ac performance of the circuit in Fig. 6 can be established from the high-frequency equivalent circuit in Fig. 7a. (For this approximation, it is assumed that the signal frequencies are high enough that all capacitive reactances of Fig. 6 are negligibly small.)

Each transistor junction has an associated junction capacitance. These are quite small (on the order of a few picofarads), but they do affect transistor action at high frequencies. A typical transistor frequency response plot is shown in Fig. 7b. At the frequency where the reactance of the parasitic input capacitance equals the input resistance βr_e, the current to the input resistance is bypassed through the capacitance to the point where the effective β is down 3 dB from its low-frequency value. This is called the β-cutoff frequency, f_{ae}. If the frequency is further increased, β continues to decrease at a rate of 6 dB per octave. The frequency at which β equals unity is specified on data sheets as f_T, the current-gain/ bandwidth product. Given f_T, it is possible to determine transistor beta for any frequency between f_{ae} and f_T from the relation $\beta = f_T/f$.

Negative Feedback. While the dc degenerative feedback associated with R_E of Fig. 6 stabilizes the operating point, making it independent of changes in beta and other temperature-dependent parameters, the bypass capacitor keeps it from compensating for the deleterious effects of these changes on the ac signal. Moreover, while proper placement of the operating point can reduce non-symmetrical signal clipping, it cannot reduce the distortion for large signal swings caused by nonlinearity of the I_C/I_B characteristics (Fig. 4). These characteristics can be greatly improved by means of negative



Voltage Gain with $r_E = 0$

$$A = r_L/r_e \quad r_L = \frac{R_C R_L}{R_C + R_L} \quad r_e \approx \frac{25\ \text{mV}}{I_E}$$

$$\text{Feedback Factor} = F = \frac{r_E}{r_L}$$

Effect of negative feedback on gain:

$$A_{(f)} = \frac{A}{1 + AF} = \frac{1}{F} \quad (\text{when } AF \gg 1)$$

Effect of negative feedback on amplitude distortion:

$$D_{(f)} = \frac{D}{1 + AF}$$

Effect of negative feedback on frequency response:

$$f_{ae(f)} = f_{ae}\ (1 + AF)$$

Effect of negative feedback on input resistance:

$$R_{i(f)} = R_i\ (1 + AF)$$

Fig. 8. One-stage amplifier and equations for feedback effects.

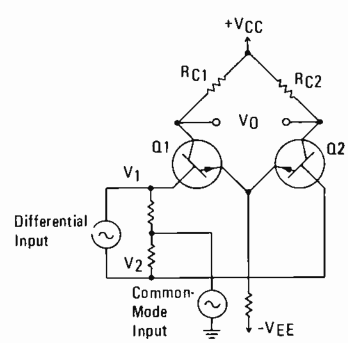
signal feedback, which requires a small unbypassed resistor, r_E, in series with R_E as shown in Fig. 8. (This is only one of many possible feedback arrangements.) In addition, negative feedback improves frequency response and compensates for changes in output voltage (and gain) due to variations in temperature-sensitive parameters such as r_e and β_{ac}.

The equations accompanying Fig. 8 describe the basic advantages achieved through negative feedback, as well as the price paid for them in terms of voltage gain. However, since feedback increases input resistance, the loss of gain can partly be recovered because of an increase in the gain of a previous stage caused by the increase in input resistance.

Darlington Transistors. Modern semiconductor technology not only has led to complete circuits on a single chip of silicon (integrated circuits) but also to compound-connected transistors. For the circuit designer, the latter provides some cost and space savings, while still permitting unrestricted circuit design freedom. One of these devices is the Darlington pair shown in Fig. 9.

Although consisting of two interconnected transistors, the device can actually be treated as a single transistor with extremely high current gain and input resistance. Normally, Darlington pairs are employed in the grounded-collector configuration. Commercially, they are available as small-signal and power devices, in both npn and pnp polarities and with betas ranging from several hundred to several thousand.

Differential Amplifiers. With the advent of integrated circuits, the circuit in Fig. 10 has become increasingly important. Being dc-coupled through a common-emitter resistor, it has no low-frequency limit; but, unlike other types of dc-coupled amplifiers, it exhibits excellent stability and drift-free operation without requiring elaborate compensating circuitry.



Characteristics, if Q1 and Q2 are perfectly matched and $R_{C1} = R_{C2}$:
$V_i = V_1 - V_2$
$$V_O = \frac{R_C}{r_e}(V_1 - V_2)$$
$r_{in} = 2\beta r_e$

For differential input:
$V_1 = -V_2$
For common-mode input:
$V_1 = V_2$

Fig. 10. Basic differential amplifier.

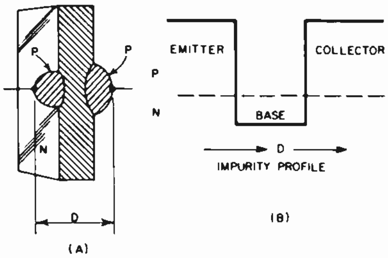


Fig. 11. Typical alloy transistor.

This is its most important characteristic. Operated in the differential mode, as shown, the output voltage responds only to "difference inputs" to the two bases. If a common-mode

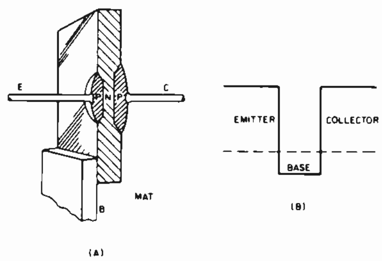


Fig. 12. The microalloy structure.

signal were applied (as in the case of ground line or power-supply noise) or if the characteristics of the transistors were to change in response to a change in temperature, the collector

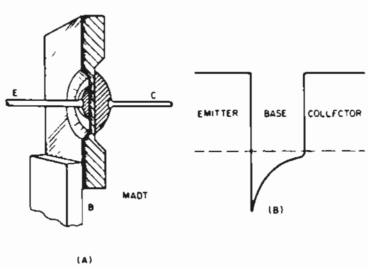


Fig. 13. Microalloy diffused type.

current of both transistors would be affected equally. As a result, the output voltage between the collectors would remain constant.

Transistor Fabrication Processes. Over the years, many processes and structures have been used in transistor fabrication. Most of them are still being used, although the older processes no longer offer the best obtainable performance. The major sequential developments in the processing of the bipolar transistor are shown in Figs. 11 through 15.

In Fig. 11A is a typical alloy transistor, while Fig. 11B shows its impurity profile. It is simple and inexpensive to build. It provides excellent low-frequency beta and can operate at high currents and power levels, but not at high frequencies or high voltages.

Fig. 12 shows the construction detail and impurity profile of a typical microalloy (MAT) structure. It is similar to the technique shown in Fig. 11 except that shallow pits are etched into the base substrate prior to collector and emitter alloying. The thinner base improves the frequency response but results in a fragile structure and further reduces breakdown voltage.

The process shown in Fig. 13 uses diffusion of impurities into a thin base membrane prior to alloying to permit a closely controlled, graded impurity profile. This technique offers frequency responses up to 100 MHz.

The process shown in Fig. 14, with extremely thin collector and base regions and unrestricted use of different material resistivities, provides high-frequency performance up to a gigahertz. It also provides high gain and high breakdown voltage. However, sensitive pn junctions are ex-

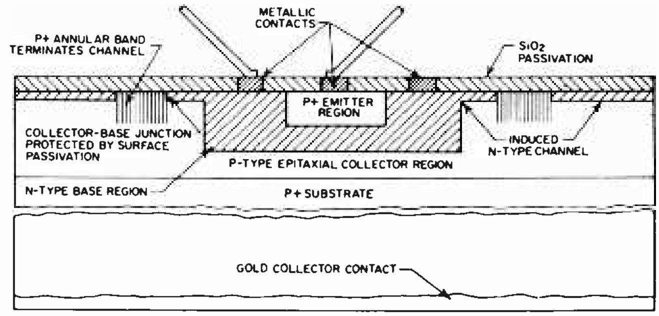


Fig. 15. Latest process is epitaxial planar with annular ring.

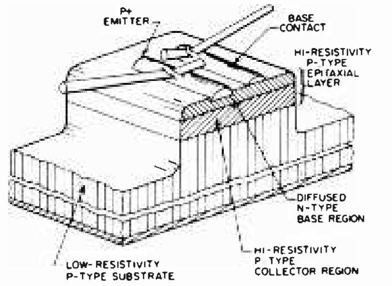


Fig. 14. Epitaxial mesa structure.

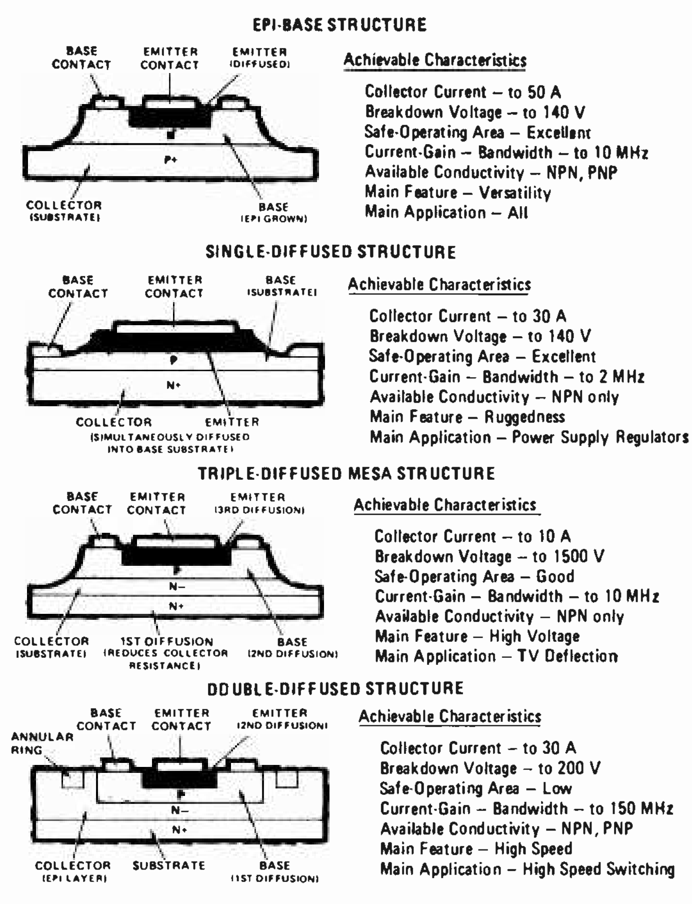
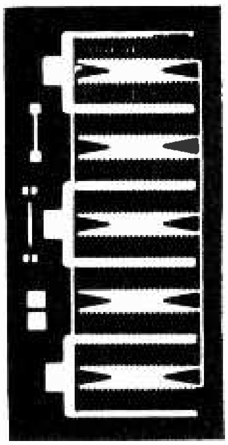
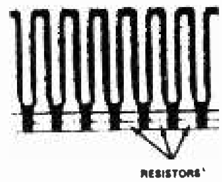


Fig. 16. Cross sections of the processes employed in making transistors for power applications, with some characteristics.





Typical high-frequency transistor structure has INTERDIGITATED geometry. Complex structure is due to large number of separate emitters all interconnected to form a single transistor. Resistor depositions equalize current to the individual emitter areas.

RELATIVE CHARACTERISTICS OF RF STRUCTURES		
GEOMETRY	**ADVANTAGES**	**LIMITATIONS**
Interdigitated	High emitter periphery to base-area ratio. (≈7.0). Low r_b'. Adequate manufacturing experience. Easily balanced with emitter resistors.	Processing critical due to narrow interdigitated fingers. Limited to reletively low currents due to narrow metal stripes.
Overlay	Wider fingers for higher current capability. Reduced manufacturing difficulty.	Higher r_b' and considerable emitter-base parasitic capacitance limits high-frequency response. Relatively low P/A ratio, (≈4.0). Emitter balancing difficult.
Network Emitter	Wide metal fingers. Highest P/A ratio, (7.0 to 8.0).	High contact resistance. Emitter balancing difficult. Processing critical.

Fig. 17. At very high frequencies, power transistors take on some very complex geometries to compensate for current crowding. Characteristics of structures at right.

posed in the atmosphere, resulting in high leakage current.

Newer Processes. Among new techniques, the latest process (Fig. 15) diffuses controlled-geometry base and emitter regions into the collector layer and covers the entire device with a protective coating of silicon dioxide to eliminate impurity contamination. It permits operation at extremely high frequency, high voltage and high current, and provides good reliability at low cost.

For small-signal, low-frequency transistors, the epitaxial planar structure with annular ring is by far the most widely used. For high-power applications, however, other processes are often employed to optimize characteristics needed for special requirements. Fig. 16 shows the most common power-transistor processes.

At very high frequencies, power transistors take on very complex geometries to compensate for current crowding. The latter restricts the emission of charge carriers to the edges of the emitter at high current levels. With rectangular or round emitters, therefore, the center portion of the emitter does not contribute to current emission, but it does add parasitic capacitance which reduces high-frequency response. High-frequency structures, therefore, have very long, thin emitters which may be folded many times to fit within a given base area (Fig. 17). This greatly increases the ratio of emitter periphery to base area, thereby maximizing the emission-to-capacitance ratio. These extremely complex structures stress present processing technology to its limits, accounting for the rapid increases in prices as frequency capabilities of power transistors go beyond 100 MHz.

Transistor Capabilities. For small-signal applications, today's transistors cover virtually every conceivable requirement. Darlington (compound-connected) transistors offer high input impedances and betas up to 75,000 at audio frequencies. Amplifiers for r-f oscillator and mixer applications run well into the GHz band. For special applications, low-noise transistors with noise figures around 2 dB are common, and for high-voltage applications, devices with ratings up to several hundred volts are no longer unique. And, due to plastic packaging, prices are so low that there are few device capabilities that can't be purchased for under $1.00 even in unit quantities. See Table 1.

Amplifiers	Switches	Special Purpose
f_T to 1200 MHz		Low Noise Amplifiers NF < 2 dB
I_C to 500 mA, BV_{CEO} to 100 V	Speeds to 0.5 ns	High Voltage Transistors BV_{CEO} to 300 V
		Darlington Transistors $\beta_{(min)}$ to 50,000

Table. 1. Small-signal transistor capabilities.

Typical Plastic Package		Typical Metal Package (TO-3)	
3-15 A, 30 to 80 V	$0.94 to $2.80	4-15 A, 60-80 V	$1.00 to $4.00
0.5 A, 225-350 V	$1.10 to $1.50	16-50 A, 60-80 V	$5.00 to $15.00
		10-16 A, 100-325 V	$4.00 to $8.00

Table 2. Low-frequency power device capabilities.

Low-Frequency Power. Power transistors for the lower frequencies are also plentiful, but do not yet fill the desired applications as completely as do the small signal devices. See Table 2. Prices are quite low for plastic packaged devices rated up to 15 A and about 100 W. At currents below 5 A, even some of the metal packaged devices are inexpensive. But prices rise rapidly at higher current and voltage levels, particularly if higher frequency operation is required. Considerable room for further development still exists.

High-Frequency Power. At very high frequencies, the power picture is still more limited. Power outputs of one to five watts are available up to 1 GHz, with up to 50 watts at 500 MHz and 100 watts in the 150-MHz region. Prices remain high at the upper limits of power and frequency.

R-F Power. Fig. 18 shows the range of transistor power commonly available for a given frequency and power-supply voltage. Specific device types have been included to suggest possible choices for a particular application. The power-supply voltages given are those most usually encountered in practice. Devices are tailored for best operation at these voltages. Using a transistor with a higher than necessary voltage rating can result in performance degradation. ◈

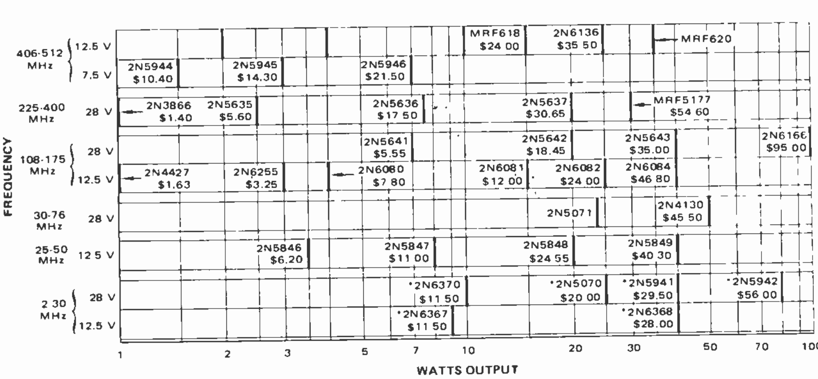
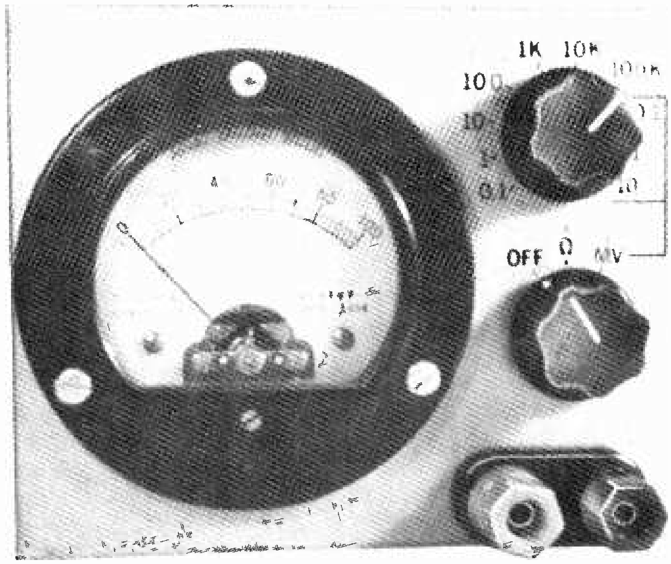


Fig. 18. Range of transistor power available for given frequency and supply voltage.



LINEAR-SCALE OHMMETER for ACCURATE READINGS

Zero to 10 megohms in seven ranges

plus three ranges of dc millivolts.

BY DALE HILEMAN

HANDY as it is, the ohmmeter portion of a typical VOM has four major drawbacks: the nonlinear scale makes accurate readings difficult; it can apply too much current or voltage to the circuit being tested, thus damaging semiconductors or delicate filaments; the zero-ohms control has to be adjusted separately for each range; and battery life may be short.

If any of these problems is bothering you or if you are looking for a handy 0-to-10-megohm meter with seven ranges, plus three ranges of dc millivolts (10, 100, and 1 V) at 10 megohms-per-volt (very useful for semiconductor circuits), try the circuit shown in Fig. 1. The maximum test current is 1 mA, and test voltage at full-scale deflection is only 1 V. The meter zero is so stable that no external zero control is needed. Both voltage and resistance indications are stable over a wide temperature range.

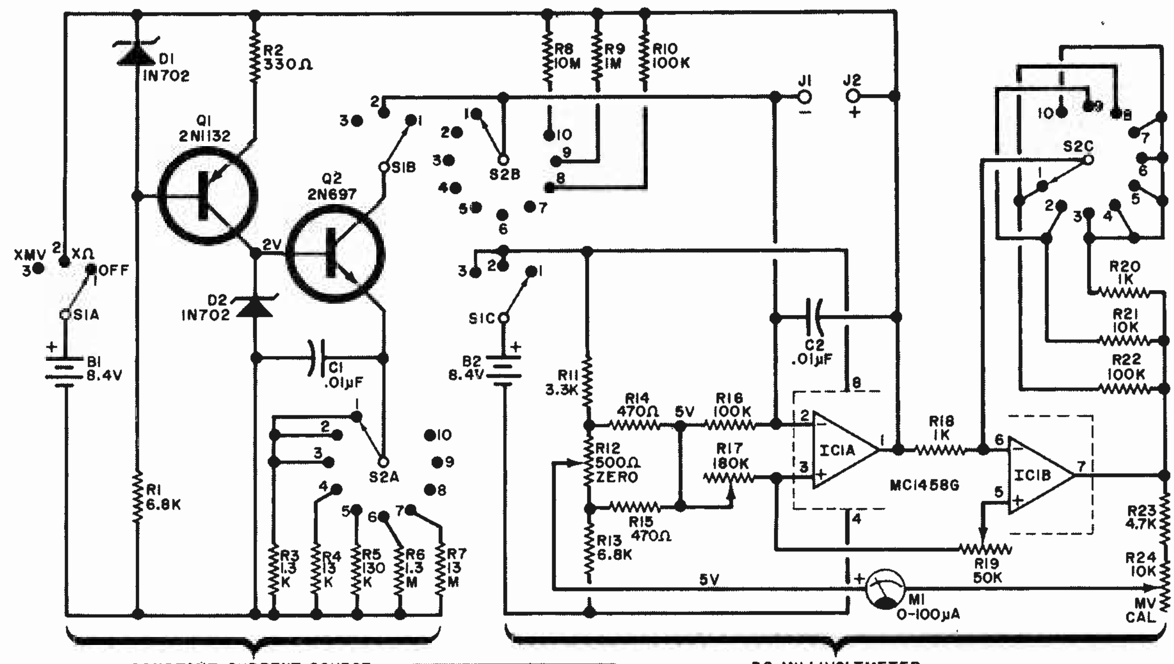
The instrument is powered by two 8.4-volt mercury batteries rated at 500 mA-hr each. Since maximum drain on either battery is under 5 mA, they can be expected to provide at least 100 hours of service. Conventional 9-volt batteries can be used, with shorter life and only a slight loss of accuracy.

How it Works. As shown in Fig. 1, the linear ohmmeter contains two sections: a constant-current source (*Q1, Q2*) and a meter driven by *IC1*. The output of the constant-current source is applied to the unknown resistance through *J1* and *J2* and the resulting voltage drop is applied to the meter through *IC1*.

Except for the input connection, the two circuits are independent. (Note that they do not even share a common ground, for reasons given later.) A position is provided on power switch *S1* to disable the constant-current source so that the instrument can be used as a millivoltmeter. With the switch in position 1 (off), both batteries are disconnected. In position 2, battery power is applied to both the constant-current source and the millivoltmeter. Also in position 2, section *S1B* applies the output of the constant-current source to the test terminals (*J1* and *J2*). In position 3, only the millivoltmeter is energized.

The constant-current source consists of a voltage regulator and a constant-current generator. The voltage regulator itself uses another constant-current generator. The voltage regulator is made up of a transistor *Q1*, resistors *R1* and *R2*, and zener diodes *D1* and *D2*. To conserve battery power, the zeners are operated at less than their rated current so that their operating potential is lower than the rated 2.6 volts. Diode *D1* provides about 1.5 volts as a reference for constant-current generator *Q1*. A constant current of about 2.5 mA, developed through *R2*, is applied from the collector of *Q1* to diode *D2*. The latter diode develops a reference voltage of about 2 volts, which is used by constant-current source *Q2* to develop a test current for the instrument. Capacitor *C1* suppresses a tendency of *Q2* to oscillate.

Resistors *R3* through *R7* are



S2 POSITION (FULL SCALE)	
1-10Ω	6-1MEG.
2-100Ω	7-10MEG.
3-1KΩ	8-10mV
4-10KΩ	9-100mV
5-100KΩ	10-1000mV

PARTS LIST

B1,B2—8.4-volt mercury battery
C1,C2—0.01-μF ceramic or Mylar capacitor
D1,D2—1N702 zener diode
IC1—Dual op-amp IC (Motorola MC1458G or equiv.)
J1,J2—Binding post (black, red)
M1—100-μA meter
Q1—2N1132 transistor
Q2—2N697 transistor
R1,R13—6800-ohm, ¼-watt resistor, 5%
R2—330-ohm, ¼-watt resistor, 5%
R3—1300-ohm, ¼-watt resistor, 5%*
R4—13,000-ohm, ¼-watt resistor, 5%*
R5—130,000-ohm, ¼-watt resistor, 5%*
R6—1.3-megohm, ¼-watt resistor, 5%*
R7—13-megohm, ¼-watt resistor, 5%*
R8—10-megohm, ¼-watt resistor, 10%
R9—1-megohm, ¼-watt resistor, 10%
R10,R16—100,000-ohm, ¼-watt resistor, 10%
R11—3300-ohm, ¼-watt resistor, 5%
R12—500-ohm potentiometer (Mallory MTC-52-L1)
R14,R15—470-ohm, ¼-watt resistor, 5%
R17—180,000-ohm potentiometer (Mallory MTC-184-L1)
R18,R20—1000-ohm, ¼-watt resistor, 1%
R19—50,000-ohm potentiometer (Mallory MTC-54-L1)
R21—10,000-ohm, ¼-watt resistor, 1%
R22—100,000-ohm, ¼-watt resistor, 1%
R23—4700-ohm, ¼-watt resistor, 10%
R24—10,000-ohm potentiometer (Mallory MTC-14-L1)
S1—3-pole, 3-position rotary switch
S2—3-pole, 10-position rotary switch
*Selected component (see text)
Misc.: Battery connectors (2); perf board with clips; knobs with indices (2); socket for 8-lead TO-5 can; suitable chassis; mounting hardware; etc.
Note: For calibration, ¼-watt, 1% resistors of 1000, 10,000, 100,000 ohms and 1 and 10 megohms are required. (Since 10-meg resistors are rare, two 5-meg units in series can be used.)

Fig. 1. Circuit consists of a constant-current generator and very stable dc millivoltmeter, which can be used as separate meter.

selected to provide five test currents in decade from 1 mA to 0.1 μA, respectively, as range switch *S2* is advanced from position 3 to position 7. To conserve battery power and keep the test current low, the value for positions 1 and 2 is held to 1 mA, but the sensitivity of the millivoltmeter is increased ten times (via *R21*) or 100 times (via *R22*), respectively.

Why is it necessary for the test current to be so incredibly small on the higher resistance ranges? Why not keep the test current at 1 mA and simply reduce the sensitivity of the millivoltmeter by decade steps? The reason is to keep the full-scale voltage at a reasonable 1-volt value. Across 10 megohms, a test current of 1 mA would develop 10,000 volts.

The effect of rising ambient temperature on *D1* and *D2* is to decrease the test current; but the effect on *Q1* and *Q2* is to increase it. The net effect is that the test current remains constant over a wide temperature range.

The total current drain on battery *B1* is 3.5 mA to 4.5 mA, depending on the resistance being tested.

The MC1458C is an inexpensive integrated circuit containing two operational amplifiers in a single case. Each is similar to the popular type 741 but without offset adjustment. One of the op-amps (*IC1A*) is used as a very-high-resistance voltage-follower, while the other (*IC1B*) is connected as an inverting dc amplifier.

The usual practice is to use a plus-and-minus supply to power an op-amp. To avoid the need for another battery, however, an "artificial ground" is created for *IC1* by the voltage divider made up of *R11* through *R15*. Resistor *R11* was made smaller in value than *R13* because this particular op-amp works better if the plus supply voltage is slightly less than the minus.

Zero potentiometer *R12* is provided to compensate for offsets of the IC. It forms a bridge with resistors *R14* and

R15 so that the positive terminal of meter M1 can be set slightly above or below the artificial ground potential at the junction of R14 and R15.

On the higher resistance ranges of the instrument, accuracy would be severely degraded if the millivoltmeter measurably loaded the constant-current source because the instrument would then be measuring the input resistance of the millivoltmeter in parallel with the unknown resistance. Therefore, op-amp IC1A is connected as a voltage-follower with very high input resistance, coupling the voltage developed across the unknown to the relatively low input resistance of the inverter.

The input to a more conventional op-amp voltage-follower is applied between common (or ground) and the plus (non-inverting) input. With a typical inexpensive op-amp, however, the input resistance of this arrangement is about 2 megohms — far too low for our purposes. In the high-impedance follower circuit of IC1A, the input is applied between the output (terminal 1) and minus input (terminal 2). This arrangement provides a very high input resistance because the input voltage is matched (or bucked out) by the action of IC1A. The disadvantage is that the millivoltmeter can't share a common ground with its input, so two batteries are required. But the input resistance of the millivoltmeter is in the hundreds of megohms.

Capacitor C2 bypasses any 60 Hz or other noise pickup on the input leads that might overload the op-amp. The capacitor may be larger for severe noise but increasing the value will also slow the response of the instrument on the high resistance ranges.

On the millivolt ranges, section B of range switch S2 bridges the input with resistors R8, R9, or R10 to enhance the stability of open-circuit meter zero. The values of these resistors are arbitrarily chosen to provide an input resistance of 10 megohms-per-volt, but they can be increased if you have good luck in selecting the op-amp or if you are satisfied with some instability of meter zero on the millivolt ranges.

The effect of bias current flowing in the plus and minus input leads of IC1A is balanced by the adjustment of potentiometer R17. This pot is set to equalize the quiescent dc level on these inputs.

The second part of IC1 is connected as a conventional inverting dc amplifier. Gain is determined by the ratio of resistors R20 through R22 to input resistor R18. These values are selected to provide gains of 1, 10, and 100, respectively. The resistances of R23 and R24 in series with the meter are selected to yield a full-scale meter sensitivity of 1 volt.

The input polarity to the millivoltmeter and the connection of the meter were arranged to give up-scale deflection for a negative-going output from IC1B. With the obverse connection (positive for up-scale deflection), the IC has a tendency to lock up like a flip-flop.

Potentiometers R17 and R19 are connected in an unconventional series arrangement to avoid biasing complications. The adjustments of R17 and R19 are slightly interacting, but they provide good stability. Nevertheless, for best results, it may be necessary to select IC1 due to the wide range of bias offsets for the MC1458.

New voltage and resistance ranges can be obtained by added switching to alter the sensitivity of the millivoltmeter. For instance, S2 could have more positions so that different feedback resistors could be selected by S2C. However, the value of R18 should not be changed because doing so will upset the balance of bias currents at the input to IC1B.

Another way of getting more meter scales might be to switch smaller values of series resistance in the places of R23 and R24. This approach has economic appeal because, in one easy step, it provides seven new resistance ranges and three new voltage ranges.

It is also possible to use a meter with higher sensitivity (50 or 20 microamps full-scale) selecting values for the meter series resistors (R23 and R24) that give the desired full-scale deflection. However, it is not a good idea to use a meter with less sensitivity than 100 microamps because full-scale deflection would necessitate excessive current in the artificial ground voltage divider, which might possibly unbalance the millivoltmeter biasing network.

At the risk of getting a slight instability, you can get another full decade step of sensitivity (1 millivolt or 1 ohm, full-scale) by using one of the foregoing methods, or some combination thereof. For a 1-mV sensitivity range, however, you should arrange for S2B to select a 10,000-ohm load.

In the prototype, potentiometer R12 is an internal adjustment. Depending on the stability and sensitivity of your version, you may want to make R12 an external control. This is especially true if you use long test leads because, on the lower resistance ranges, you will then have to contend with zero shift caused by the resistance of the leads.

Total current drain on B2 is 1.7 to 2.7 mA, except when the meter is pegged, in which case it is 4 mA. (The meter normally pegs in the ohmmeter mode when the input is open.)

The accompanying table summarizes the voltage and resistance relationships in the ohmmeter.

Construction. The prototype, except for the meter and the various switches, was assembled on perf board, with the final mounting in a small metal enclosure. The batteries were mounted on clips on the rear cover. It is suggested that the battery clips be insulated from ground because an occasional battery has an internal short from one terminal to its case.

Due to differences in bias currents, it is possible that the first dual op-amp

VOLTAGE AND RESISTANCE RELATIONSHIPS

S1	S2	Test R Full Scale (ohms)	Test I Full Scale (μA)	Test E Full Scale (mV)	IC1A Input Load (ohms)	IC1B Gain
2	1	10	1000	10	open	100
	2	100	1000	100		10
	3	1000	1000	1000		1
	4	10,000	100	1000		1
	5	100.000	10	1000		1
	6	1 meg	1	1000		1
	7	10 meg	0.1	1000		1
3	8	—	—	10	100,000	100
	9	—	—	100	1 meg	10
	10	—	—	1000	10 meg	1

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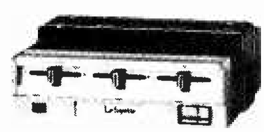

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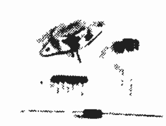

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you try might not perform satisfactorily. So it is best to have two or three of these units on hand. We tested 12 specimens consisting of an approximately equal mix of Motorola's MC1458G and MC1458CG and Texas Instruments' SN72558L. Judging by this test and the published data, just about any of the dual 741-type op-amps can be used, but 25% of any given type may not work properly. Those you can't use in this project may work well in less critical applications.

To facilitate the selection or replacement of *IC1*, it is recommended that a socket be used for mounting it. Also, get a couple of extra 1N702 zener diodes since *D1* and *D2* should be selected to provide an optimum reference voltage for the constant-current source. It is essential that capacitors *C1* and *C2* not leak — use ceramic or Mylar types.

Carbon or alkaline 9-volt batteries are perfectly suitable for *B2*. They can also be used for *B1*, but their greater range of terminal voltage will cause a slight long-term drift in resistance readings.

Resistors *R3* through *R7* must be selected to provide proper full-scale deflection on each of the resistance ranges. You may want to use potentiometers instead. If so, for *R3*, use 1000 ohms in series with a 500-ohm pot; for *R4* use 10,000 ohms in series with a 5000-ohm pot, and so forth.

Don't solder resistors *R3* through *R7* in place since the heat may change their resistance. The lead of a ¼-watt resistor can be snapped neatly into the bottom of the slot in a Vector T-28 perf-board terminal, making a firm, permanent contact without the need for solder. A connecting lead can be soldered to the tip of the terminal sticking out of the other side of the board.

Don't solder *D1* and *D2* into the circuit until you are completing the calibration process as described below.

Calibration. The millivoltmeter section must be adjusted before the constant-current source is calibrated. To adjust the meter, you will need an accurate dc source of 1 volt. If the meter circuit fails to respond as indicated, replace *IC1* and start over. Resistors needed to calibrate the constant-current souce are given in the Parts List.

Install both batteries, place test leads in *J1* and *J2* and proceed as follows:

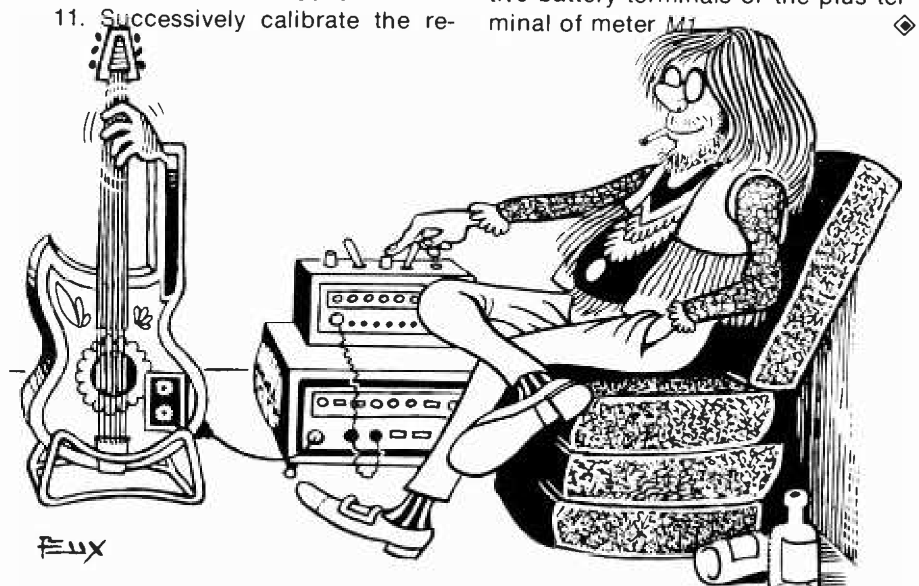
1. Set *R12* with the wiper at its maximum position. Set *R24* at mid-position.
2. Turn *S2* to position 10 and *S1* to position 3.
3. Alternately short together and open the test leads, watching meter *M1*. Adjust *R17* so that the meter reading does not change as a result of this action. Proper adjustment should occur with the meter reading somewhat above zero.
4. Leaving the test leads open, switch *S2* from position 10 to position 8 and then back to 10, watching the meter. Adjust *R19* so that the meter reading does not change as a result of this action. Again, proper adjustment should occur with the meter reading just above zero.
5. Repeat steps 3 and 4 until the meter reading is stable and no interaction occurs.
6. Adjust *R12* to zero the meter.
7. Put *S2* on position 10, and connect the test leads to a 1-volt dc source (observing polarity). Adjust *R24* for a full-scale deflection. This completes calibration of the millivoltmeter section.
8. Turn *S1* to position 2 and measure the voltage across *D2*. If it is between 1.8 and 2.0 volts, proceed to step 11. If it is not, interchange *D1* and *D2* or try different diodes. Solder *D1* and *D2* in place.
9. Set *S2* to position 3 and connect the test leads to a 1000-ohm calibration resistor. Select (or adjust) *R3* to obtain full-scale deflection.
10. Set *S2* to position 4 and connect the test leads to a 10,000-ohm calibration resistor. Select (or adjust) *R4* to obtain full-scale deflection.
11. Successively calibrate the remaining ohmmeter scales as in steps 9 and 10.

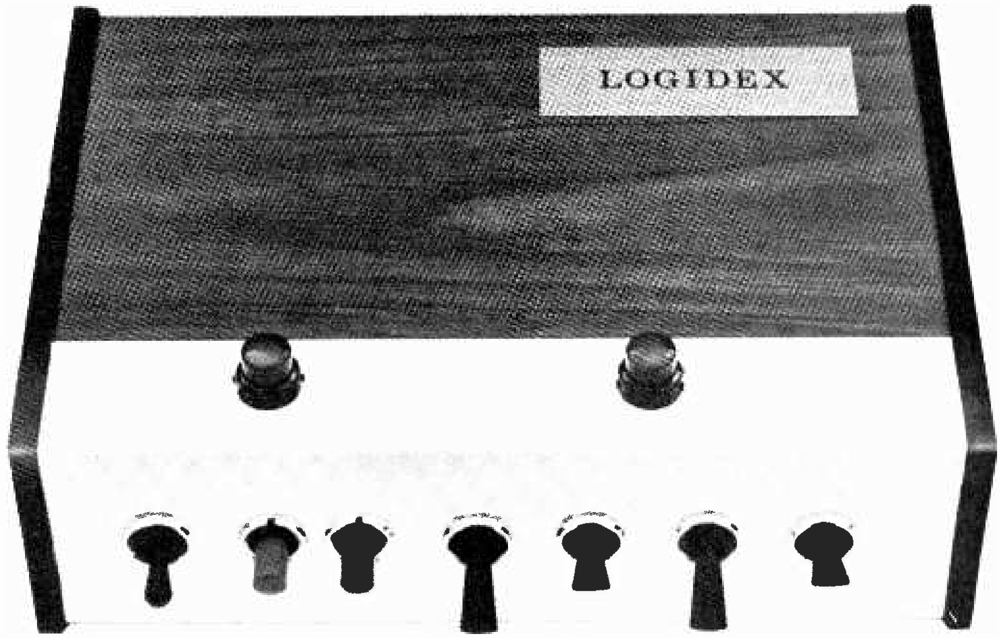
In Case of Trouble. As we have indicated, the most likely trouble spot is in the IC and the symptom is instability or difficulty in adjusting the millivoltmeter. You should suspect a marginal op-amp if it proves necessary to set *R12, R17,* or *R19* to one extreme or there is considerable interaction between the adjustment of *R17* and *R19*.

If, however, the trouble is merely a slight up-scale reading of the meter on the 10-ohm scale with test leads shorted, the first thing you should suspect is that you are really reading the resistance of the leads. If shorting directly across *J1* and *J2* with a screwdriver still doesn't drop the reading to zero, try repeating steps 2 through 6 of the adjustment procedure, but be more careful this time. With *J1* and *J2* shorted together, there should be absolutely no shift of meter zero as *S2* is switched through its ohmmeter ranges.

If you get unexpected resistance readings while using the ohmmeter for its intended purpose, don't blame the instrument until you check its linearity against a known precision resistor. Composition resistors — even brand new ones — sometimes fall outside their own rated tolerances.

If on a cold winter day, you notice intermittent jumping of the meter needle, you are probably the victim of static electricity. To preclude certain leakage problems, we have shown no connection from the circuits to the chassis. In case of static, however, you might try grounding one of the negative battery terminals or the plus terminal of meter *M1*. ◈





LOGIDEX

AN ELECTRONIC GAME FOR ALL SEASONS

Flashing light game for all ages uses digital logic.

BY HOWARD L. NURSE

CHILDREN are fascinated by switches and flashing lights, while adults are attracted to games of chance. Here is an electronic game you can build which bridges the gap and can be enjoyed by persons of all ages.

We call the game "Logidex," a combination of logic and dexterity, both of which are required to win the game.

The table describes some ways in which the Logidex can be used. The games are based on a player's ability to logically decode an unknown four-bit binary number as rapidly as possible with an array of four switches. Each time the player succeeds in finding the number with a correct combination of "up" and "down" game switches, a red light flashes for four seconds. After four seconds, the player presses the red switch to generate a new random number, and plays again. This sequence, which was started initially by pressing the green switch, continues for thirty seconds, as timed by a green flashing light.

GAMES TO PLAY

Game	Rules
Logi-Peg	1. Use a cribbage board with pegs. Assign one peg to each player.
	2. Each player advances his peg one hole per win during his 30-second play period.
	3. Play last "round" after one player reaches end.
	4. Player to advance farthest after all players have had an equal number of turns wins.
Logi-Sum	1. Assign a number (any number, including Roman numerals) to each switch.
	2. Player must sum the "up" switches after each win.
	3. First player to exceed a pre-determined total (such as 100) wins.
Logi-Lette	1. Draw a circle and divide it into 16 sections, numbering the sections from 0 to 15.
	2. Assign the game switches the numbers 1, 2, 4, and 8.
	3. Each player chooses one numbered section of the circle.
	4. One player plays Logidex and sums the "up" switches.
	5. The player whose chosen section number corresponds to the score wins.

How it Works. The schematic for

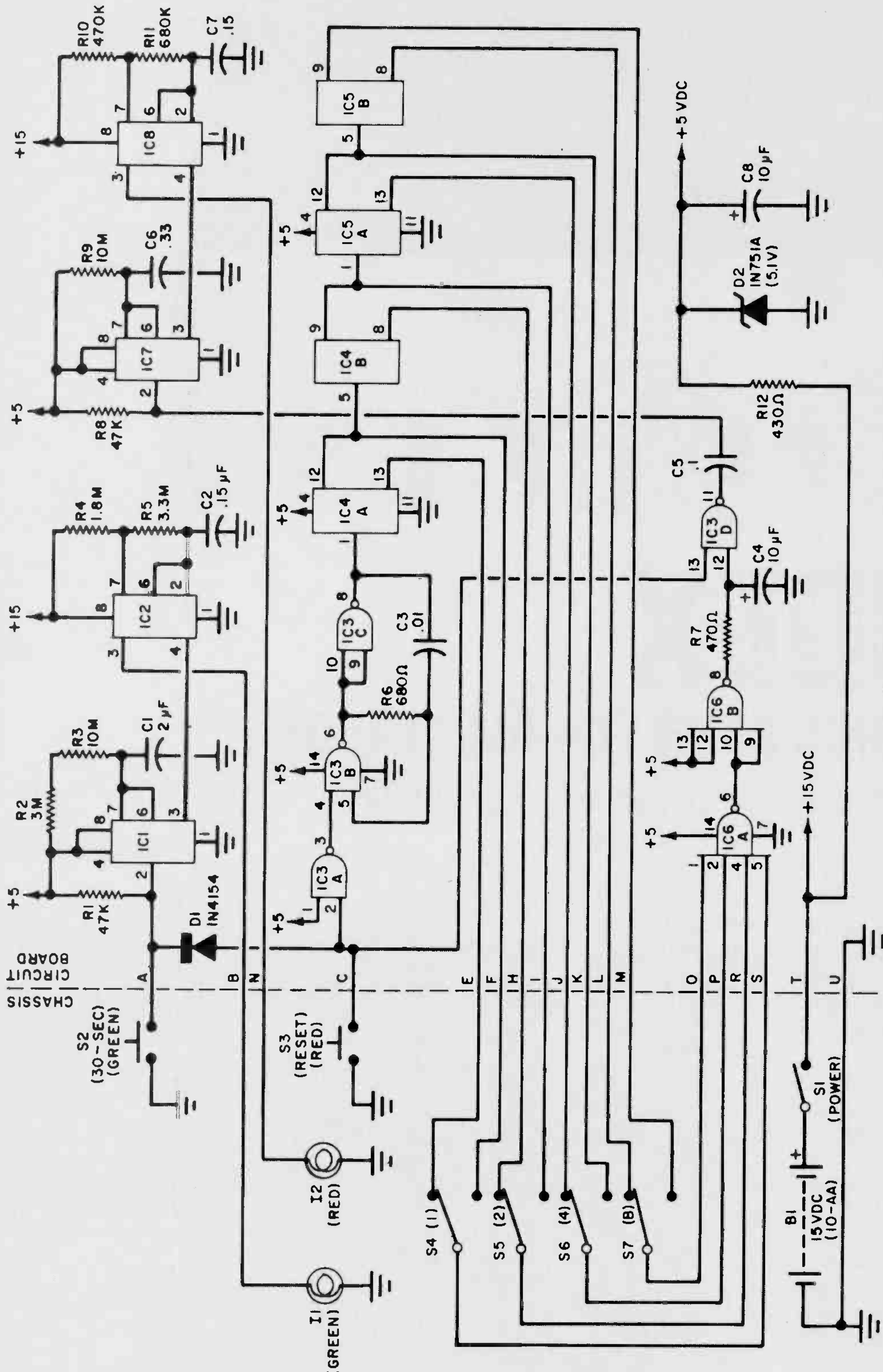


Fig. 1. The game consists of four portions: a 30-second timer and flasher, a clock, a ripple counter, and a 4-second timer-flasher.

PARTS LIST

B1—AA cell (ten required)
C1—2-µF, 200-volt capacitor (Sprague "Orange Drop" or similar)
C2, C7—0.15-µF, 100-volt capacitor (Sprague "Orange Drop" or similar)
C3—0.01-µF disc capacitor
C4, C8—10-µF, 30-volt electrolytic capacitor
C5—0.1-µF disc capacitor
C6—0.33-µF, 100-volt capacitor (Sprague "Orange Drop" or similar)
D1—Diode (1N4154 or equiv.)
D2—5.1-volt 1N751A zener diode
I1, I2—12-volt, 60 mA lamp (one red, one green)
IC1, IC2, IC7, IC8—NE555V integrated circuit
IC3—2N7400/SN74L00 (see text)
IC4, IC5—SN7473/SN74L73 (see text)
IC6—SN7420/SN74L20 (see text)
R1, R8—47,000-ohm, ¼-watt, 10% resistor
R2—3-megohm, ¼-watt, 10% resistor
R3, R9—10-megohm, ¼-watt, 10% resistor
R4—1.8-megohm, ¼-watt, 10% resistor
R5—3.3-megohm, ¼-watt, 10% resistor
R6—680-ohm, ¼-watt, 10% resistor
R7—470-ohm, ¼-watt, 10% resistor
R10—470,000-ohm, ¼-watt, 10% resistor
R11—680,000-ohm, ¼-watt, 10% resistor
R12—430-ohm, ¼-watt, 10% resistor
S1—Spst switch
S2, S3—Spst normally-open pushbutton switch (green & red)
S4—S7—Spdt switch (bathandle type)
Misc: Battery holders, cabinet (Ten-Tec JW-8), mounting hardware, etc.

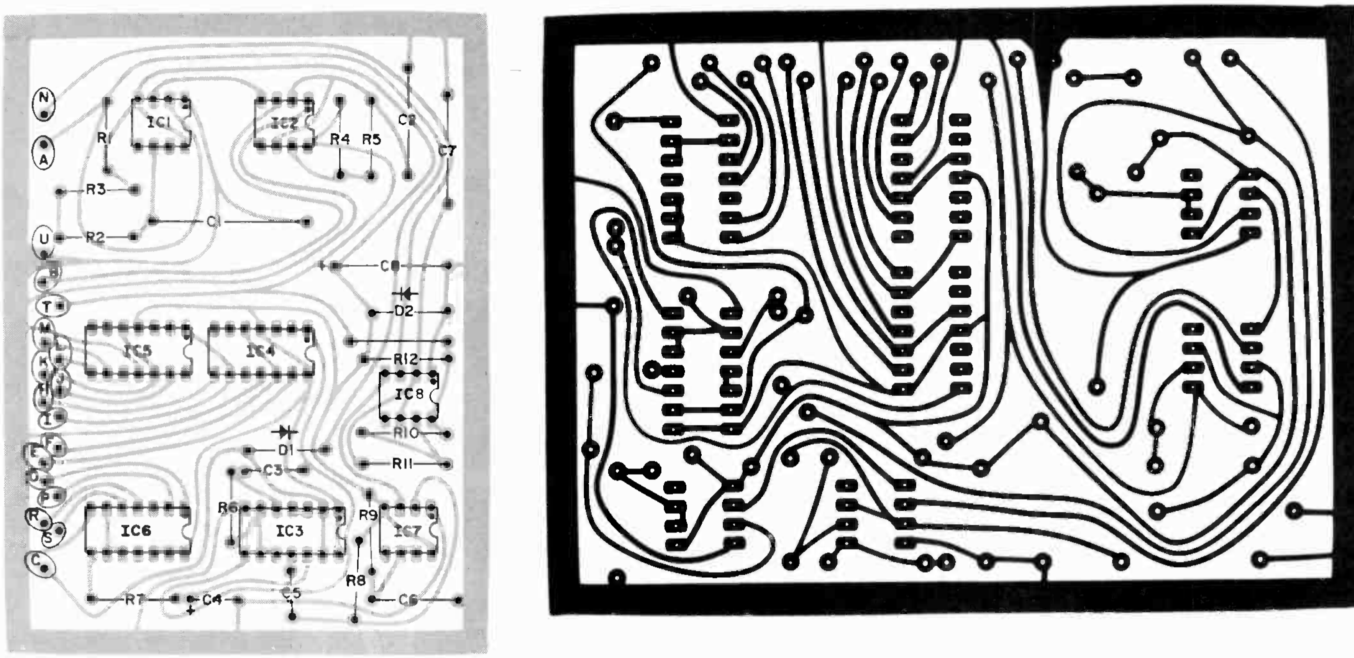


Fig. 2. Actual-size foil pattern and component installation (left). Be sure to observe polarities.

Logidex is shown in Fig. 1. The circuit can be divided into functional blocks as follows: a thirty-second timer and flasher (*IC1*, *IC2*); clock oscillator (*IC3B* and *IC3C*); counter (*IC4*, *IC5*); qualification logic (*IC6*, *IC3D*); and four-second timer and flasher (*IC7*, *IC8*). In addition, there are the power supply and controls.

Pressing the green switch (*S2*) causes the thirty-second timer to start and also enables the clock to oscillate for an unknown number of cycles. The clock oscillates at approximately 50 kHz, so that many cycles pass while the switch is pressed. The red switch (*S3*) also enables the oscillator, but it is isolated from the thirty-second timer by diode *D1*.

The clock output (from *IC3C*) is used to trigger a four-bit ripple counter (*IC4*, *IC5*). Both the normal and inverted outputs from each stage of the counter are wired to switches so that one throw of each switch will always have a logic 1 connected to it.

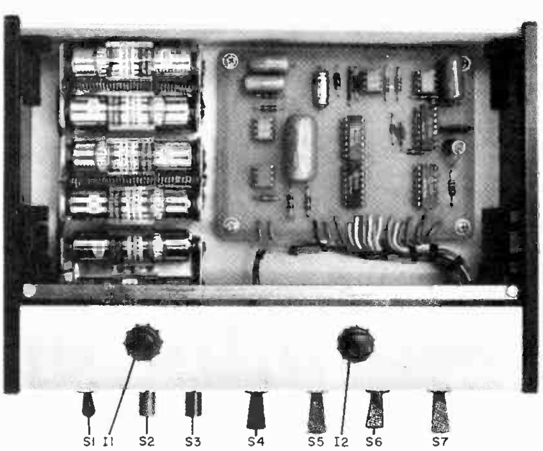
When the switches have been toggled to the correct positions, all four inputs to *IC5A* are at logic 1. The output from *IC6B* goes high under this condition, passes through a low-pass filter (*R7, C4*) to remove noise transients, and triggers the four-second timer and flasher. Integrated circuit *IC3D* allows the circuit to function only during normal play and inhibits the output of the qualification logic when the clock is causing the counter to operate.

The power supply consists of ten AA batteries with a zener regulator for the *IC* supplies.

Construction. The circuit can be constructed on a printed-circuit board, such as that shown in Fig. 2. The assembly of the prototype is shown in the photos. Note that the lamp connections from points B and N on the circuit board are routed separately from the main harness. This routing is necessary to isolate the fifteen-volt flash signals from the sensitive flip-flop inputs and outputs. Be sure to unscrew the lamps while soldering the sockets to prevent costly damage to the filaments.

When selecting components, keep in mind that high impedances are present in the timing circuits, so low-leakage capacitors are necessary for *C1* and *C6*. Sprague "Orange Drop" capacitors were used for *C1*, *C2*, *C6*, and *C7* to stabilize the flash rates and timing.

While standard TTL logic gates (such as SN7400) can be used, low-power logic (such as SN74L00) is recommended to conserve power and prolong battery life. When using an SN7400 instead of an SN74L00 for *IC3*, it is necessary to lower the values of *R6* and *R7* to 220 ohms and *R12* to 120 ohms, 2 watts. Logic speed, which is reduced when using low-power gates, is not a critical consideration in this application. ◈



Use large bathandle-type switches for easy operation. Switch S1 and lamp I1 are green, while switch S2 and lamp I2 are red.

HOW AUDIO SWEEP GENERATORS SAVE TIME & INCREASE ACCURACY

Tests amplifiers, speakers noise-reduction units, phase-locked loops.

BY JON D. PAUL

WORKING with audio equipment and circuits need not involve old-fashioned point-to-point response plotting. Audio sweep generators make the job easy and accurate.

Linear and Log Sweeps. With a linear sweep, the frequency (horizontal) scale, as displayed on an oscilloscope, is calibrated in so many Hz or kHz per division. This is fine for indicating exact frequencies, but for audio work a logarithmic scale is necessary, because the ear reacts to ratios of frequencies rather than to numerical values.

The audible sound spectrum has been roughly divided into three ranges, each spanning about three octaves or one decade (10:1 ratio). These are popularly called bass (20-200 Hz), midrange (200 Hz-2 kHz), and treble (2-20 kHz). A linear scale spanning the 0 to 20-kHz range (Fig. 1A) compresses the entire bass range into 1/100th of its length and the midrange is in the first 1/10th. However, a logarithmic scale (Fig. 1B) gives equal area to each part of the audio spectrum.

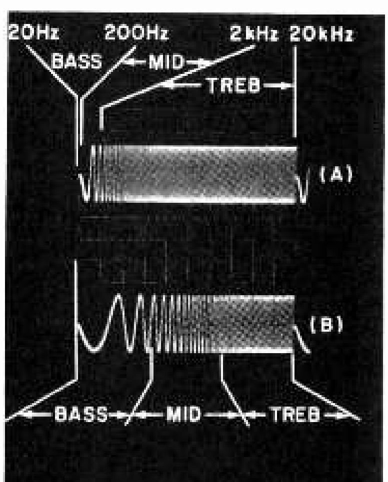


Fig 1. Basic sweeps from 20 Hz to 20 kHz. Linear sweep at (A), log at (B).

As a practical application, note the response of a low-pass filter to a linear sweep using two different sweep ranges, as shown in Fig. 2 (top). The shape of the response changes when the sweep rate is changed. With a log sweep, however, the essential shape of the response stays the same, as shown in Fig. 2 (bottom).

Using a Sweep Generator. The basic arrangement for using a sweep generator is shown in Fig. 3, with the sawtooth output of the sweeper driving the horizontal axis of the scope and the audio sweep connected to the unit being tested. The output of the tested unit (properly terminated) goes to the scope's vertical input. At slow sweep rates (longer than 100 ms), the scope should have dc coupling at the horizontal input. If an ac-coupled horizontal input is used, the sawtooth waveform will be distorted. A sweep rate that is too low will be evidenced by the uneven motion of the horizontal trace. It will be fast at first and then slow up as the sweep progresses to the right.

An alternate mode of operation is to use the time base of the scope. The external sync or trigger of the scope is connected to the sweeper's sync output, and the scope's time base is reset each time the sweeper's speed is changed. This setup is useful if the scope does not have a dc-coupled horizontal amplifier.

The scope's vertical amplifier will also affect the accuracy of an audio sweep display. When slow speeds are used with a generator that does not begin each sweep with a zero dc off-

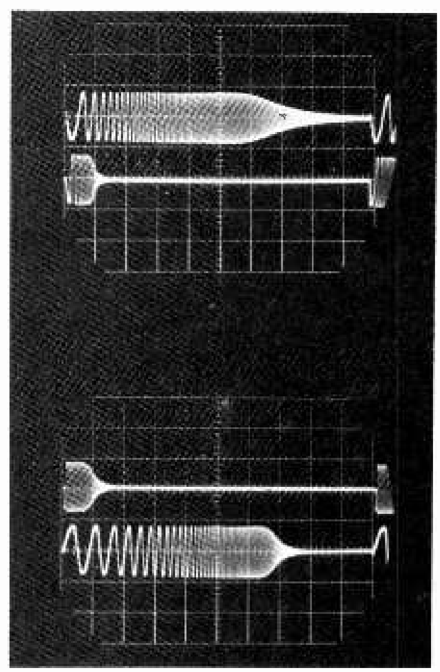
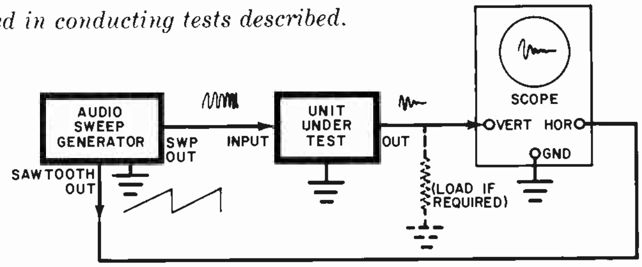


Fig. 2. Low-pass filter response for 2 different ranges of linear sweep (top). Below are log sweeps for 2 ranges of response for same low-pass filter.

Fig. 3. Test setup used in conducting tests described.



set, dc-coupling is desirable. If the generator has an initial-condition reset circuit so that the output has no dc component, either ac or dc coupling can be used.

To make a permanent recording of a sweep, it is only necessary to feed the sawtooth output of the sweeper to a

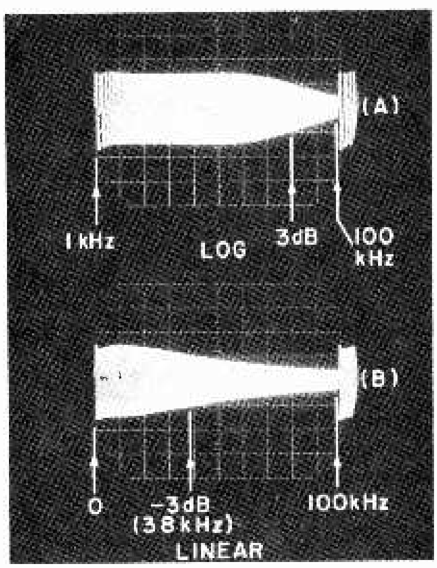


Fig. 4. Log sweep from 1 to 100 kHz of an audio amplifier (top). Below is linear sweep from 0 to 100 kHz of same.

graphic recorder's horizontal input, use a detector to convert the ac output of the circuit under test to dc, and apply this to the recorder's vertical input. The setup is similar to that shown in Fig. 3. Very slow sweeps (10 to 100 seconds) should be used to prevent the relatively slow response time of the recorder from distorting the shape of the plot.

The logarithmic mode of a sweep generator should be used for wide-band analysis and the linear mode for narrow-band (high-Q) systems—or to read exact frequencies directly off the scope trace. For example, note the audio amplifier's overall response pattern with log and linear sweeps as shown in Fig. 4. The overall response is best seen in Fig. 4A, while the exact −3-dB point is more easily identified on the linear sweep in Fig. 4B. This assumes, of course, that zero is on the left and the 100-kHz point is on the right, with each horizontal division representing 10 kHz.

In sweep generators having a fixed starting frequency, the frequency is usually 0 to 1/1000th of the full sweep frequency in the linear mode and 1/100th to 1/1000th of the sweep in the log mode. There is one other method of frequency control that may be encountered. This occurs in an r-f sweep-type generator where a start-frequency control determines the beginning of the sweep and a sweep-width control sets the frequency deviation.

The sweep-rate control on a generator determines the time that it takes a sweep to occur. There are two opposing factors involved in this setting, which means a trade-off is necessary. Slow sweeps (long sweep time and slow sweep rate) cause flicker in the CRT display which may make observation difficult. If the rate is too fast, a smearing effect occurs.

These conditions are illustrated in Fig. 5 for a low-pass filter. At top, the slow sweep produces a clear trace and the result is accurately depicted. Below, the sweep rate is too fast and a hard-to-read display results. The best way to set the sweep rate is to start off with a slow setting and watch the display while increasing the sweep speed until smearing occurs. Then back off on the speed control until the trace is clean. The sweep rate at which distortion starts depends on the bandwidth of the equipment under test and on the width of the sweep.

Sweeper Applications. There are a number of basic tests that can be performed using a sweep generator—with a scope as the readout device. We will list them by the units on which the tests are performed.

Amplifiers. There are precautions that must be observed when using a sweeper to check a power amplifier. They include: keep the output voltage low enough to prevent overloading the circuit being tested; take the response at various output levels; use a carbon resistor of the correct power-handling capability as the load—not a speaker or a wirewound resistor since its impedance varies with frequency.

The results of an amplifier test are shown in Fig. 6. The 15-watt transistorized amplifier's response with a 1-watt output is shown at the top and with a 15-watt output below. Note the reduction in high-frequency response at the rated power. Remember that some transistor power amplifiers should not be tested for prolonged periods at full power. Also make sure that the power line has the correct voltages.

Preamplifiers. Keep in mind that the various inputs to a preamplifier are designed for specific voltage levels and source impedances. Consult the preamplifier's specifications before hooking up the sweeper. Check the phono inputs for conformity to the RIAA compensation curve. The tape output (s) should be observed first since they usually bypass the volume and tone controls. The effects of these controls may be checked by making a response test from a "flat" input (usually auxiliary) to the main (power amplifier) output.

Tape Decks. Response of a tape unit

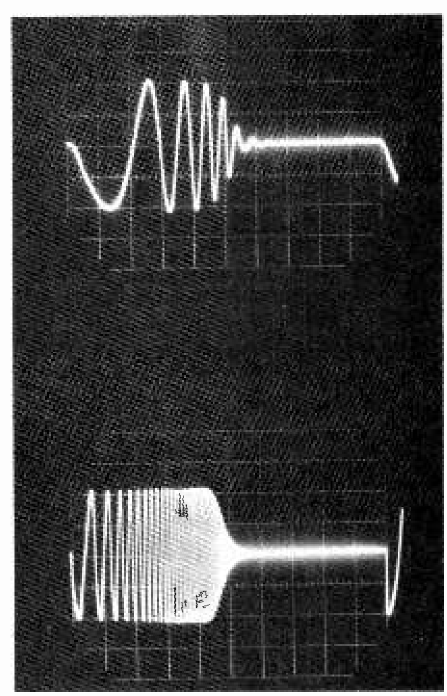


Fig. 5. Effect of sweep rates on low-pass filter response curve. Top is slow sweep, which is clear. Bottom shows fast sweep; not as readable.

depends on tape speed, bias, recording level, head alignment (especially azimuth), and, finally, the tape itself. The output of a three-head deck is delayed with respect to the input. Two-head decks, especially cassette and cartridge recorders, provide no output at all during recording. These factors complicate testing. Two techniques are available for getting around these problems. The first, which can be used only on three-head decks, is to set the sweep time to a submultiple of the delay time (head spacing/tape speed). This rate can be set by observing the swept output of the deck and adjusting the sweep rate until one complete sweep is displayed, with the starting point properly positioned and with no movement. Set the scope for Int. Sweep and Ext. Sync.

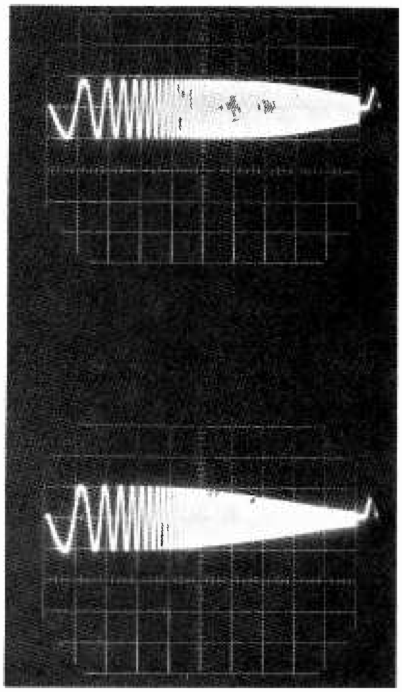


Fig. 6. Amplifier response from 1 to 100 kHz at 1-watt level (top) and 15-watt level (bottom). Note response changes with power and how 3-dB point lowers.

An alternate method, suitable for two- or three-head machines is to record a test tape of sweeps and play them back, displaying the output of the deck on a scope. The scope's internal time base is used with internal sync, and the end of each audio sweep will serve as a trigger for the scope. The scope time/cm (horizontal sweep) and triggering controls should be set for a stable display of one complete sweep. The level of the sweeper's output must be carefully adjusted so that it does not exceed the dynamic range of the deck. Although zero on the VU meter would appear to be the best level at which to run tests, many tape-deck manufacturers recommend a level that is −20 dB lower. At this level, bias and low-frequency noise may appear to be excessive, but these signals can be ignored since only the frequency response is important. If desired, a bias trap and low-frequency cutoff filter can be used at the deck's output.

Once these techniques are mastered, frequency response can be checked at various tape speeds and with different types of tapes. The channels of a stereo deck should be checked separately first and then compared for a balanced result. If the response doesn't meet the specified ratings, demagnetize and re-align the deck's bias, head azimuth, and equalization, according to the manufacturer's instructions. Use the type of tape recommended. This is the coarse alignment.

The sweep generator can now be used to make the fine alignment. Vary the bias level so that the output level is maximum with the frequency response as flat as possible (or as flat as it was after the coarse alignment). A low bias will cause a high-frequency boost while overbiasing will cut the highs. With the bias set, the response can be trimmed to get the best flatness using the equalization controls. All controls should require only small changes from the settings arrived at in the coarse alignment. Note that the head azimuth can cause a loss in high-frequency response if it were not set up properly during coarse alignment.

With the deck properly set up, a calibration sweep tape can be recorded. To check playback equalization and azimuth in the future, play the tape back and observe the scope display. The effect of using a different type of tape can now be tested. The procedure above should be repeated to test response at lower speeds on multi-speed machines, but the equalization can only be adjusted for low speed. Bias and azimuth settings are the same for all speeds. Fig. 7 shows the effects of overbiasing.

Filters. There are many applications for filters in audio equipment: speaker crossovers, multiplex FM, four-channel decoders, tone controls, scratch and rumble filters, etc. The ham uses filters in SSB transmission, code reception, SSTV, and RTTY. In

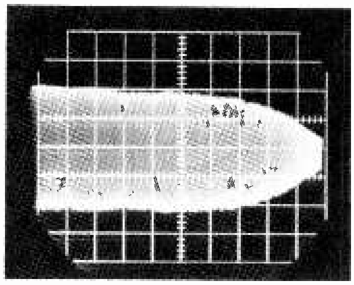


Fig. 7. Overbiased tape response curve.

testing filters, it is necessary to detach their inputs from the unit in which they are used and determine the source impedance of the driving circuit. Then use the sweep generator to drive the filter through a resistor equal to the source impedance. Observe the filter's output with a scope probe that will not load the filter. Fig. 8 shows the response of a multiplex filter used in a high-quality tuner.

Speakers. The response of a speaker is checked by applying the sweep generator signal through a power amplifier to drive the speaker. A calibrated microphone (a low-cost electret condenser unit is good) is set up a few feet in front of the speaker. The microphone's output is displayed on the scope directly (or through a preamp if necessary). The resonant point at the low-frequency, the mid-range flatness, and the high-frequency rolloff should be noted. The enclosure, the amplifier's damping ratio, and room acoustics all affect the response, so take these factors into consideration.

Microphones. In testing a microphone, the technique is identical to that used on a speaker except that a mike with calibrated response is used as a reference for comparison purposes. Then the mike to be tested is substituted for the reference mike.

After making a test with the sweep generator, the reference response is compared to the new response and the difference represents the new

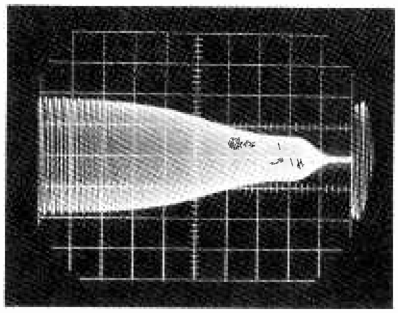


Fig. 8. FM multiplex filter response using 300-Hz to 30-kHz logarithmic sweep.

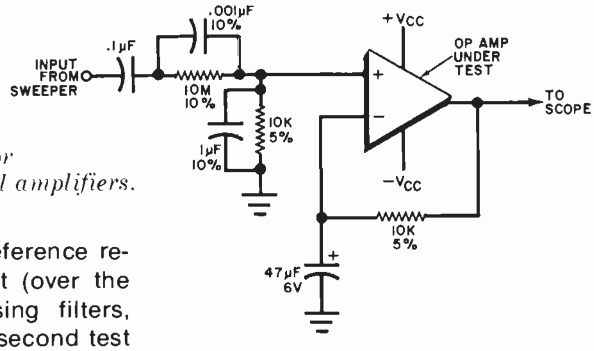


Fig. 9. Test setup for checking operational amplifiers.

mike's response. If the reference response can be made flat (over the band of interest) by using filters, equalizers, etc., then the second test will provide the test mike's curve.

Noise-Reduction Units. Dolby, dbx, and similar noise-reduction systems often create overall responses that are dependent on the input level. A flat response is desired but a full system sweep test can be done only if a coder

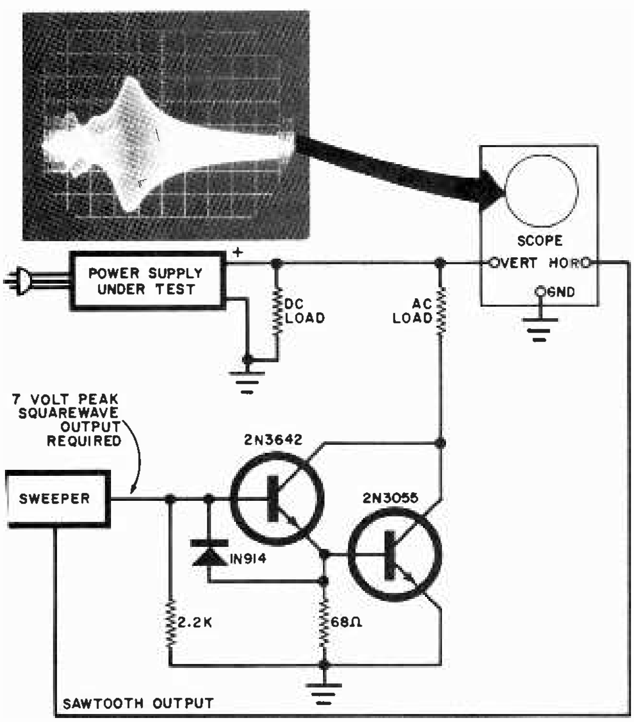


Fig. 10. Test for output impedance vs frequency of power supply can be made using this setup. Typical scope display is shown at upper left.

and decoder are connected together. If, for example, a cassette deck containing only a Dolby decoder is under test, then the Dolby decoder's family of response curves vs level must be consulted. Comparison of the swept response with these curves will verify proper operation. The noise-reduction circuit should be isolated from the surrounding circuits to avoid bias or equalization from confusing the results.

Operational Amplifiers. The setup shown in Fig. 9 can be used in testing the open-loop frequency response of an op-amp. Due to the high gain and the dc-input offset of the op-amp, attenuation of the sweep generator

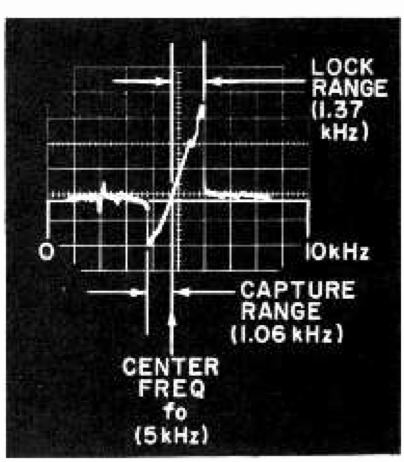


Fig. 11. Phase-locked loop response. FM output vs frequency. Note "Z" shape.

and biasing of the op-amp are necessary. The open-loop gain may be flat to a maximum of only 10 to 1000 Hz. Then, a 6-dB per octave rolloff will start. At the frequency at which the gain becomes unity, the 6-dB per octave slope must be maintained for closed-loop stability (freedom from oscillation). Maximum external compensation (for unity closed-loop gain) should be used if the op-amp is not internally compensated.

The closed-loop response will depend on the feedback. Generally, the gain and offset problems encountered in open-loop measurements will not arise and the closed-loop test will not require the setup in Fig. 9.

Power Supplies. Fig. 10 shows a simple circuit that can be used to check output impedance vs the frequency of a power supply. The sweeper switches the Darlington pair from cutoff to saturation, inserting and removing a known load resistance in the supply output at the sweep generator's frequency. The ac component in the output of the supply is displayed on the scope. This gives a direct reading of output impedance vs frequency. Power-supply oscillation problems can be debugged using the sweep generator by opening the loop and checking the open-loop response of the regulator at the opened point.

Phase-Locked Loops. Since PLL's are basically low-frequency FM detectors, the swept response will be in the shape of a Z as shown in Fig. 11. The lock and capture range of the loop and its center frequency can be measured and adjusted by viewing this display. The input level should be varied since the parameters may be functions of input voltage. A slow sweep rate is often required since some PLL circuits have a slow response to reduce noise.

Ultrasonic Transducers. In testing ultrasonic transducers, the manufacturer's drive and receiver circuits should be consulted since external inductors and/or capacitors are sometimes used for tuning and narrowbanding purposes. The power used in testing should be low. (50 mW is a typical maximum CW power.) The beamwidth of the transducers is often of interest. Move one element off-axis in a face-to-face setup. Maintain a 5-foot radius and swing the element in an arc. Measure the angles from on-axis to a point where the peak of the sweep display falls to 0.707 of its maximum. The beamwidth is twice this angle.◈

BUILD AN

IC LIGHT MODULATOR

Control up to 1 kW of lamps with a minimum of parts.

BY EDWARD M. YANDEK

LIGHT modulators are becoming increasingly popular with college students, hi-fi buffs, electric guitarists, and other musicians. Such devices expand the auditory sensations of music into a pleasing visual experience as well. Usually, however, to buy a good commercial light modulator is fairly expensive; less costly models are generally low on sensitivity and must be used at high listening levels.

The single-channel modulator described here is inexpensive, simple to build, and very sensitive. It should be possible to buy the parts for this modulator (with a 1000-watt controlling capability) for about five dollars.

Theory of Operation. The input impedance of the primary of *T1* is in series with *R1* to insure that there is no adverse loading on the audio amplifier. The stepped-up secondary voltage is controlled by sensitivity potentiometer *R2* and applied to the IC input. The key to the sensitivity and simplicity of this circuit is the use of operational amplifier *IC1*, a 741. The gain of this stage is determined by the setting of feedback control *R6* with respect to *R4;* the input is protected against overvoltage by diodes *D1* and *D2*. The output of the op-amp drives the gate of the triac through *D3*, which prevents the triac input from going negative.

The triac, which operates on both halves of the line power cycles, is connected in series with the lamp load across the power line. Protection is provided by *F1*. Each time the gate of the triac goes positive, the triac fires and remains conducting until the voltage crosses through zero. Then the triac cuts off. The triac, therefore, is turned on only during the positive peaks of the audio.

Construction. Parts layout is not critical and either a PC or a perf board can be used. A socket can be used for the IC and a heatsink is needed for the triac. Select a triac and fuse whose ratings are compatible with the lamp load being used. Keep in mind that the "common" circuit may be at power-line voltage level, so *do not* use a chassis ground for this circuit. Terminal strips are used to make power connections. Also, be sure that switch *S1* will carry the required current. For low-power operation, a line-isolation transformer is suggested. Both the op-amp and the triac are available at low cost.

Mount the completed assembly in a plastic container with only the audio-input jack, the shafts of *R2* and *R6*, and the two switches on the front panel. (*S1* can be part of *R2* or *R6*.) *Do not connect* the shield of *J1* to common ground in the circuit.

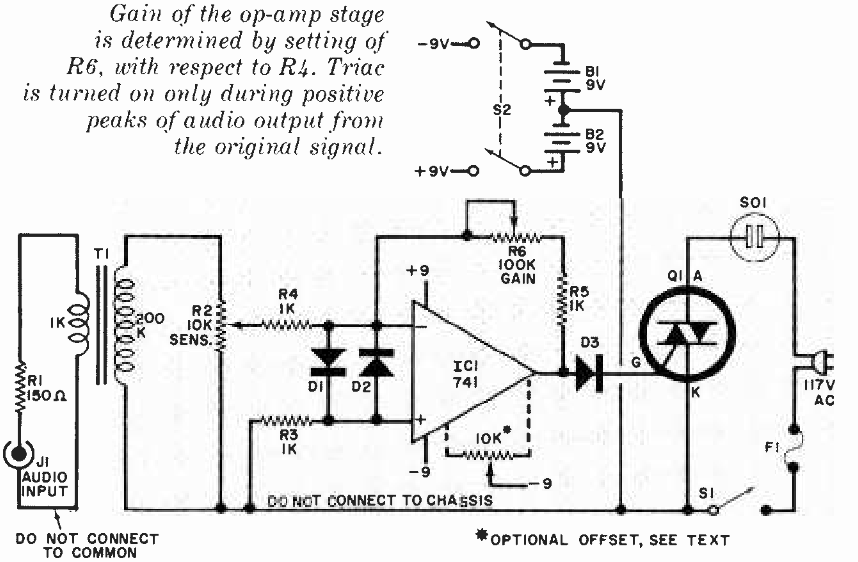
Adjustments. Connect a short-circuit across input jack *J1* and set gain control *R6* to the center of its travel. Connect a suitable lamp load to *SO1*, and turn on the modulator. The lamp should not glow. If it does, check the wiring of the circuit. If it is OK, then the op-amp may require a slight offset to bring its positive-going output below that required to fire the triac. This is accomplished by using the optional offset cricuit shown in dotted lines on the schematic. With the potentiometer connected between the offset terminals of the op-amp and with its rotor connected to −9 volts, the potentiometer can be adjusted until the light goes out. You can also use a low-voltage dc voltmeter (positive connected to the op-amp output and negative to common) to measure the op-amp output voltage. It should be less than that required to fire the triac.

Using the Modulator. Connect the audio input to *J1*. (Use two modulators for a stereo system if desired.) Set *R2* to minimum and adjust the speaker volume as desired. Turn on the ac power to the modulator; then turn on the op-amp battery power. With *R6* set at mid-position, slowly turn up *R2* until the lamp begins to pulsate with the audio level.

For increased sensitivity, rotate *R6* toward its maximum resistance. If you make the circuit too sensitive, it will be "touchy" so use less feedback and more *R2*.

Due to the high gain of the op-amp, interaction may result if there is an SCR light dimmer on the same ac circuit. Turn such devices off before using the light modulator. ◈

Gain of the op-amp stage is determined by setting of R6, with respect to R4. Triac is turned on only during positive peaks of audio output from the original signal.



PARTS LIST

B1,B2—9-volt battery
D1,D2,D3—Silicon rectifier diode
F1—Fuse and holder (see text)
IC1—741 op-amp
J1—Phono jack
Q1—Suitable triac (see text)
R1—150-ohm, ½-watt resistor
R2—10,000-ohm potentiometer
R3,R4,R5—1000-ohm, ½-watt resistor
R6—100,000-ohm potentiometer
SO1—Power socket
S1—Spst switch (may be part of R2 or R6)
S2—Dpdt or dpst switch
T1—Audio transformer: primary 1000 ohms; secondary 200,000 ohms (Lafayette AR-100 or equiv.)
Misc.: Battery holders; terminal strips; mounting hardware; heatsink for Q1; line cord.

Closed Box Speaker System Design

Here's how to match speaker to enclosure

BY DAVID B. WEEMS

ANYONE who has listened for more than ten seconds to an unmounted speaker knows that some type of enclosure is necessary. Comparing a speaker in a suitable enclosure with the same speaker in a mismatched box will also demonstrate what a bad combination can do to a speaker, In fact, choosing a box size that is right for a given woofer is the most critical decision to be made in designing a closed-box system because once construction is under way, little can be done to change it.

The fact that deciding on the speaker box design is critical to the building of a speaker system is no reason for the prospective builder to abandon his project. To be sure, some test equipment is needed to insure optimum performance; but predictable results can be realized by following generalized design charts. One virtue of the closed box is that design problems are straightforward.

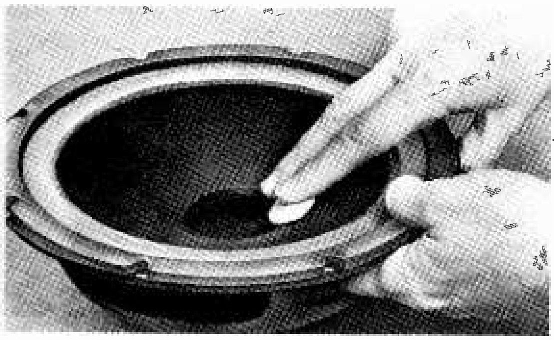
The Infinite Baffle. Closed-box speaker systems were once referred to as "infinite baffles." The name was adopted from the type of baffle that many audio men consider ideal: a flat baffle so large that the out-of-phase back wave from the speaker cone would never reach the front of the cone. Such a baffle would present equal air loading on both sides of the cone. The practical equivalent of a true infinite baffle is a speaker mounted in a room wall which acts as a barrier down to the frequency for which the path from the rear to the front of the cone is equal to one-half the wavelength of the sound. For a distance of 30 feet, the wall acts as an infinite baffle down to below 20 Hz. The frequency response of a speaker in an infinite baffle extends down to its resonant frequency, below which it rolls off at 12 dB/octave.

Chopping holes in walls for speakers has never been popular among home owners and landlords; so, the closed box came into being. The closed box is infinite in the sense that the back wave is completely isolated from the front. But there is one significant difference between the performance of a speaker in a true infinite baffle and the same speaker in a closed box. This is that the trapped air in the box acts as an added compliance, the acoustical equivalent of a capacitance.

In the mechanical circuit shown in Fig. 1, the box compliance (capacitance) is in series with the resonant circuit of the speaker, raising the frequency of resonance. This is similar to a capacitance in series with a resonant electrical circuit where it reduces the total capacitance and shifts the resonant frequency upward. In mechanical terms, the box stiffness is added to that of the speaker's suspension. To modify this effect, the early closed-box systems were made extremely large to provide high compliance (low stiffness). A typical optimum volume for a 12-inch speaker was 12 cu. ft.

To counter the large-box problem, manufacturers began making speakers with high-compliance suspensions and heavy cones. These low-resonance speakers have a much greater cone compliance (C_{ms}) than do conventional speakers. This means that they can be used with a lower box compliance (C_{mb}) and still produce a system resonance equal to the larger system. But there is a limit to how far C_{mb} can be reduced.

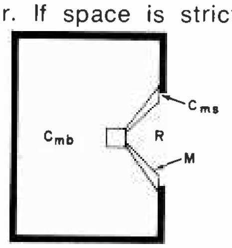
When two capacitors are connected in series, the smaller capacitance limits total capacitance no matter how much the value of the other capacitance is increased. The same principle is true of high-compliance-speaker/low-compliance-box combinations. And, in addition to the theoretical limit of total compliance due to the series circuit, there is a practical limit to how much we can increase the compliance of a speaker cone. A useful compromise between the goals of space saving and high-fidelity sound is to accept a system resonance of 50 Hz as a practical lower limit for large woofers in enclosures of moderate size and



Modeling clay is pressed against the speaker cone to change speaker mass.

70 to 100 Hz for smaller speakers in bookshelf-size cabinets.

Cabinet Size. Most published instructions for designing a speaker system begin with measurements on the woofer. If space is strictly limited,



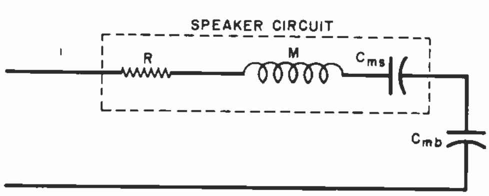


Fig. 1. Electrical schematic is equivalent of a speaker in a closed box.

there is another design decision that should be considered first. This is the choice of permissible cabinet volume. Unless this step is completed first, the builder might find that he is attempting to do the acoustical equivalent of fitting a size 10 foot into a size 9 shoe. The size of the box determines how large a woofer should be used.

To better appreciate the import of this statement, refer to Fig. 2. Notice that although each of the three speakers has the same resonance frequency (50 Hz), system resonance for any given interior volume varies greatly, especially at lower volumes. The differences are due to changing values of box compliance for the various piston areas of the speakers. The formula for box compliance is:

$$C_{mb} = V/dc^2A^2 \text{ (cm/dyne)}$$

where V is the volume of the box; d is the density of air (usually expressed as Greek letter rho); c is the speed of sound; and A is the effective cone area. All values are expressed in metric units (cgs).

Note that C_{mb} varies inversely with the square of the cone area. The air in the box acts with much more stiffness against a large piston than against a small one. This decreases the effective cone compliance of large speakers and raises their resonant frequency.

The exact enclosure volume for a woofer can be decided only after performing some tests on the woofer. However, the woofer size must still roughly match the available space

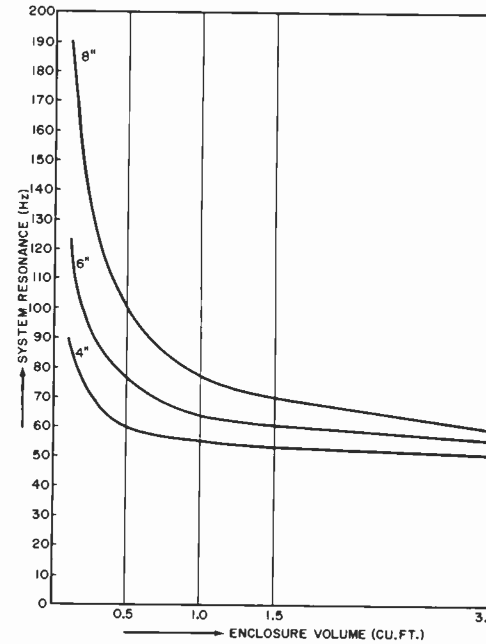


Fig. 3. Test setup for determining free-air resonance using an audio generator.

from the outset. Some suggested limits for various size woofers are given in the table. These recommendations are based on more than 200

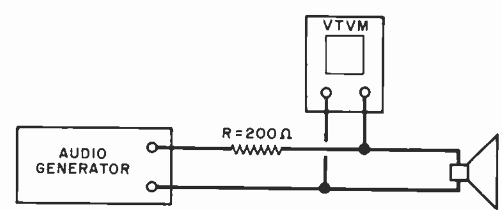


Fig. 4. Box compliance for speakers of various sizes. Sizes are nominal diameters.

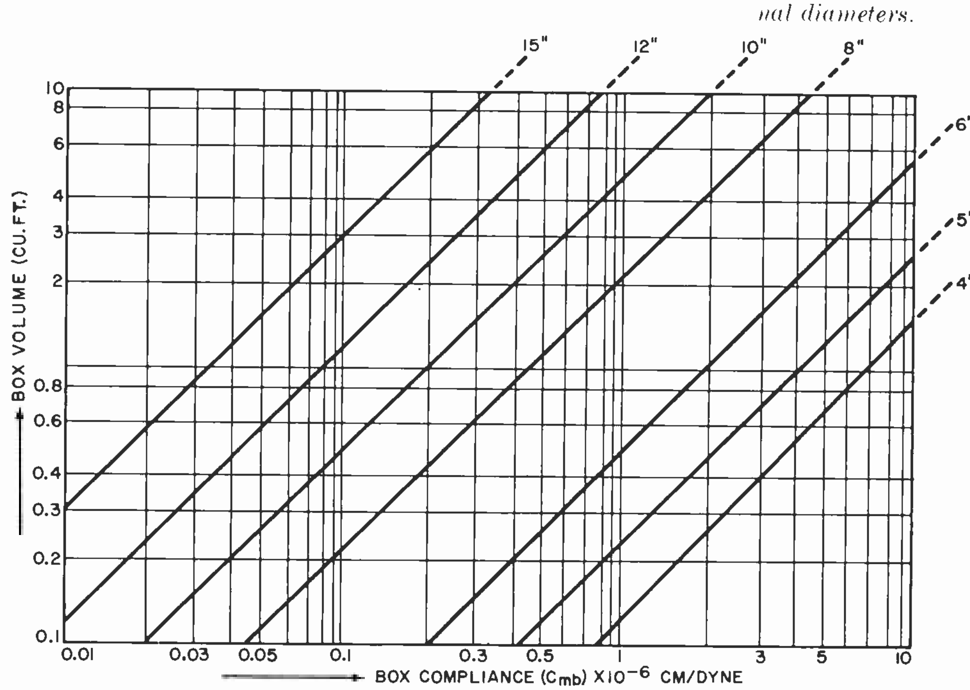


Fig. 2. System resonance vs enclosure volume for 3 different-sized woofers that have the same resonant frequency.

tests with many high-compliance woofers of various sizes. In choosing an enclosure size, volume calculations are based on *internal* dimensions.

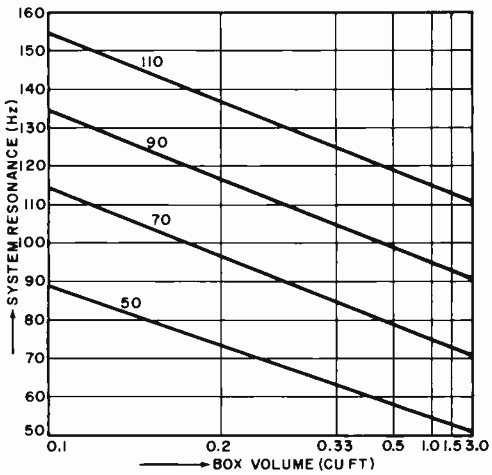


Fig. 5. Simplified design chart for 4-in. speakers. Numbers on diagonal lines indicate free-air resonance. To use, find speaker's free-air resonance. Locate point on chart for that frequency, using speaker resonance lines as guides, and draw line parallel to nearest resonance line. Read volume for desired system resonance.

There is one case in which a larger-than-normal woofer might be acceptable for a given enclosure size. This choice can be made if higher-than-normal power handling capability is desired at the expense of the low-frequency range. Small woofer cones must move much farther than large cones to deliver equal acoustical power at low frequencies. Also, small cones typically have a lower permissible range of movement before succumbing to sound distortion or serious mechanical damage. The power-handling ability of any woofer varies inversely with C_{mb}. So, for high-power operation, a compromise must be reached between the low-frequency range and the low-power-handling ability of the system by choosing a smaller-than-normal box. In addition to limiting the low-frequency response, a problem in reducing box volume below the optimum value is that Q, or resonance magnification, is increased. This factor can produce a nasty boom at resonance that is difficult to control.

ENCLOSURE SIZE/SPEAKER SIZE DESIGN TABLE

Recommended Box Volume (cu ft)	Nominal Speaker Size (in. dia.)
Less than 0.2	4
0.2-0.25	5
0.25-0.5	6
0.5-1.5	8
1.5-2.0	10
2.0 and larger	12
4.0 and larger	15

Woofer Testing. Accurate enclosure plans require knowledge of the woofer's free-air resonance, mass, and compliance. For some reason, this information—except for free-air resonance—is almost never available from the speaker manufacturer. But anyone with access to an audio generator and a VTVM can obtain this information with a few simple tests.

The first step is to find the free-air resonance. To do this, follow the hookup shown in Fig. 3. Hold the woofer in mid-air and sweep the audio generator down from about 200 Hz, noting carefully the frequency at which the voltage across the voice coil rises to a peak. This frequency is the free-air resonance (f_r).

Next, add a small known (nonmagnetic) mass to the woofer cone. Modeling clay will stick to most cones. Select enough clay to equal the mass of a nickel (5 grams), a penny (3 grams), or a dime (2.5 grams). A simple balance made from a ruler and a pencil can be used to determine how much clay you will need in each case. Press the blob of clay over the point on the front surface of the cone where the voice-coil leads protrude, providing a firm footing, until it adheres. Do this carefully, supporting the lower surface of the cone with your fingers and using only enough pressure to seat the clay. The clay, incidentally, must be of the non-drying variety.

When the added mass (M') is snug against the cone, measure the cone's resonance again. Record the new frequency as f'_r. It will be lower in frequency than the first resonance. The mass of the cone can now be calculated as follows:

$$M = M'/[(f_r/f'_r)^2 - 1] \text{ (grams)}$$

When the mass is known, the compliance (C_{ms}) can be calculated from the formula:

$$C_{ms} = 1/[(2\pi f_r)^2 M] \text{ (cm/dyne)}$$

As an example of how to put these equations to work, let us assume that a 6-inch speaker is found to have a free-air resonance of 50 Hz; a 5-gram mass is added and the new resonance is found to be 41 Hz:

$$M = 5/[(50/41)^2 - 1] = 5/(1.5-1) = 10 \text{ grams.}$$

Substituting the figure 10 for M in the formula for compliance, we have:

$$C_{ms} = 1/[(2\pi\, 50)^2\, 10] = 1/ \times 10^6 = 1 \times 10^{-6} \text{ cm/dyne.}$$

For speakers up to about 8 inches in diameter, it is practical to set the box compliance equal to that of the speaker. Two equal capacitances in series produce a new capacitance equal to half the value of one of the capacitances. When the value of the compliance in the formula for speaker resonance, $f_r = 1 / (2\pi \sqrt{M\, C_{ms}})$, is halved, the resultant resonance is equal to the square root of 2, or 1.41 times the original resonance. The same is true for closed-box speaker systems in which $C_{ms} = C_{mb}$. For large speakers, this ratio of box compliance to speaker compliance will result in enclosures that are rather large. For these speakers, it is more practical to choose a box volume that will produce a 50-Hz system resonance.

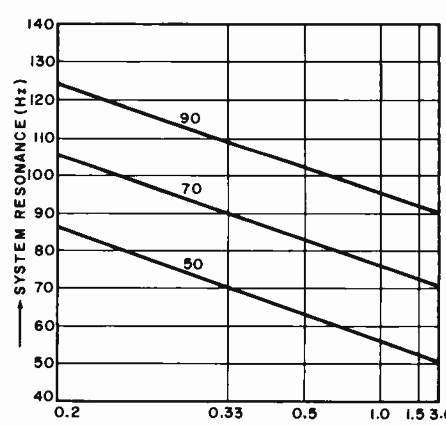


Fig. 6. Design chart for 5-in. speakers.

When the compliance is known, we can find the equal box compliance from Fig. 4. To do this, extend the vertical line at 1×10^{-6} cm/dyne until it touches the diagonal line for 6-inch speakers. Then move horizontally to the left side of the chart where you will find that a box of approximately 0.5 cubic foot is about right. At this volume, you can expect the system resonance to be 1.41 times 50 Hz, or 70 Hz.

Designing Without Test Equipment. Knowing the approximate mass and compliance of a woofer is insurance against the possibility that you have an atypical speaker. But it is not necessary to know these exact figures to obtain a reasonable degree of success. If you know the cone resonance, you can predict the system resonance by referring to the simplified design charts shown in Figs. 5 through 11. These charts show values that have been averaged from tests conducted on many different kinds of high-compliance speakers commonly available to the hobbyist.

In some charts, particularly those for larger speakers, the resonance lines are not parallel. The reason for the apparent discrepancy arises from the fact that speakers with higher resonant frequencies tend to have lighter cones than those with lower resonant frequencies. After the compliance limit is attained in low-resonance woofers, further reduction in resonant frequency must be accomplished by adding mass to the cone. Greater mass, for a given chart, produces a more horizontal line on the chart.

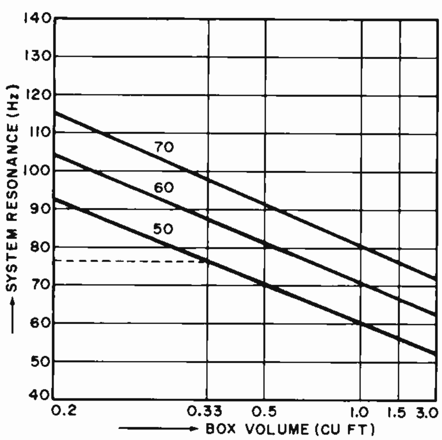


Fig. 7. Chart for 6-in. speakers Dashed line shows example in text.

To predict system resonance without knowing the mass or compliance of a woofer, you need only go to the simplified chart for the nominal diameter of your speaker. Choose the line on the chart that matches the woofer's free-air resonance or, if the correct figure is not shown, interpolate to find the location for the woofer frequency. Mark a point there and draw a line through the point, making the line parallel to the nearest resonance line.

Here's an example, using a 6-inch speaker with a free-air resonance of 50 Hz. You would refer to the 6-inch speaker chart of Fig. 7 and locate a resonance line for 50 Hz. Draw a vertical line for the enclosure volume. If the volume is to be about 0.33 cubic foot, the predicted system resonance will be about 77 Hz. If more space is available, you might set the enclosure volume at the value that will yield a box compliance equal to the speaker's compliance. To do this, multiply the speaker's free-air resonance by 1.41 to obtain a desired system resonance of 70 Hz. Now, draw a horizontal line at 70 Hz on the chart. Where this line crosses the resonance line, extend a vertical line to the bottom of the chart to find the proper enclosure volume. For our 6-inch woofer, the diagonal line crosses the 70-Hz line at 0.5 cubic foot; this would be the choice for optimum box volume.

Enclosure Details. After finding the correct enclosure volume, the next step is to make certain that this volume is available to the woofer if box compliance is to be correct. In small sealed boxes, the volume of the internal bracing can occupy enough space to become a significant factor. Also, the volumes of the tweeter and/or midrange sub-enclosures (which should be used if they are not self-enclosed) must be subtracted.

The interior depth of the enclosure should be at least 1.5 times that of the woofer (unless the woofer is unusually long). No inside dimension should be more than three times that of any other dimension. A typical ratio of external enclosure dimensions is 5:3:2. For a bookshelf speaker system, the box might be 15 inches long, 9 inches wide, and 6 inches deep. To arrive at the correct dimensions, the ratio of length to width can be set first, then enough depth can be added to provide the correct volume. If the depth is not enough to meet the requirements of 1.5 times the woofer depth, the other dimensions can be juggled a bit until the dimension ratios and the internal volume figures are right.

The best cabinet material to use for a given speaker system depends on the size of the enclosure. Plywood is the traditional choice and is quite satisfactory. Enclosures for 8-, 10-, and 12-inch speakers should be made from ¾-inch plywood that is glued and firmly screwed together with the aid of corner glue blocks and extra bracing on large panels. For small woofers, ½-inch plywood is adequate; the enclosure itself can be assembled with glue and nails. Small panels are much more rigid than are large ones of the same material and thickness. Also, small woofers are limited in their power-handling ability and will not be driven to the same sound level as large woofers.

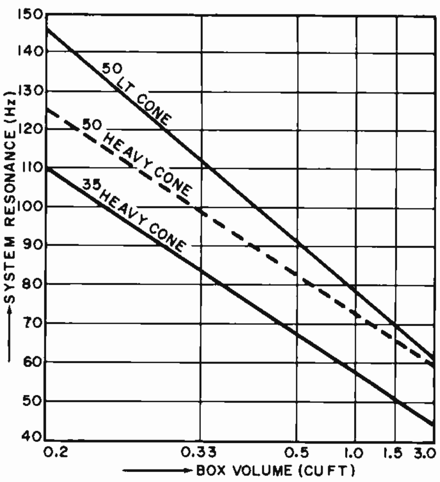


Fig. 8. For 8-in. speakers. Use the dashed line for speakers with free-air resonance near 50 Hz, with heavy cone. Use upper line for full-range speakers with free-air resonance of 50 Hz. Cone types will vary greatly.

Back panels should be installed with screws against a stop with an airtight gasket all around. A single layer of rubber tape glued around the back stop will serve well as a gasket. A perfect seal can also be obtained by running a bead of flexible caulking compound (or silicone rubber compound) around the back stop, but this makes future back removal difficult.

The inside walls of the enclosure should be covered with a damping material such as fiber glass wool to damp out-of-phase midrange reflections. The thickness of the material depends on its density. At least 1 inch of dense material is the minimum for most cabinets. The material should be added a layer at a time until the system sounds "right." Most damping materials have a limited affect on low-frequency response except to slightly increase the effective volume of the enclosure and lower the resonance of the system. The padding does this by absorbing and giving up heat, which makes the air in the enclosure operate isothermally (at a constant temperature). When sound is propagated isothermally, its velocity decreases. The formula for C_{mb}, states that this factor varies inversely with the square of the speed of sound (c). So, as c is decreased, C_{mb} increases and the enclosure appears to be larger than it actually is. The system resonance in such cases may be reduced by as much as 15 percent.

Another effect of the damping material, if enough of it is used, is that it lowers the Q of the system by damping the resonant peak. The higher the frequency of the system's resonance, the more affect the damping material will have on system Q. The optimum value of Q varies inversely with the resonant frequency of the system. For smaller high-resonance systems, a lower value of Q is desirable because a peak is more objectionable at higher frequencies. Hence, small enclosures may benefit from being loosely filled with damping material. Any loss of bass can be compensated for by boosting the bass at the amplifier—if the turnover frequency of the tone controls is not too high. But Q is more directly controlled by the magnet design of the speaker than by box stuffing. This is particularly true at the lower frequencies.

High-Compliance Woofers. High-compliance woofers are made with many types of suspension materials. A traditional material was specially treated cloth that might be formed in a sine-wave pattern, an accordion pleat, or a half-roll. Butyl rubber and

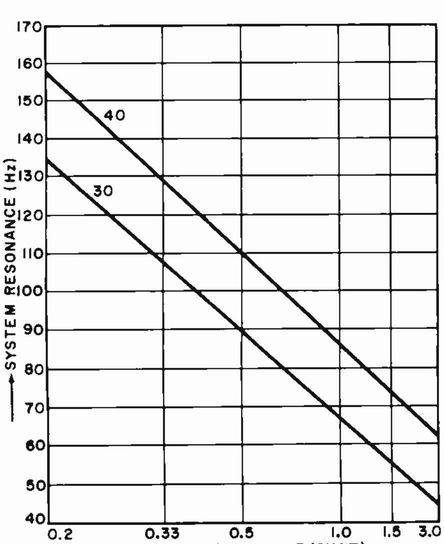


Fig. 9. Design chart for 10-in. woofers.

polyurethane foam are also used for speaker cone surrounds. Some of these materials react to climate differences or use by a change in compliance. Typically, a woofer's resonant frequency will drop after being "broken in." But in other cases, the resonant frequency rises with time.

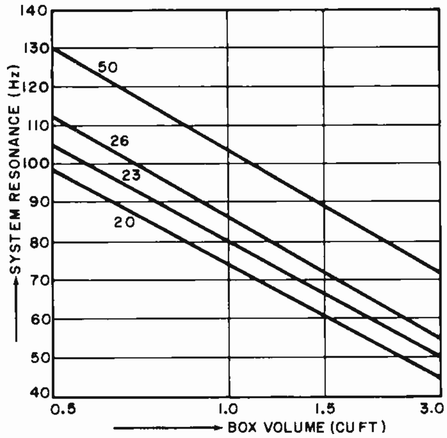


Fig. 10. For 12-in. woofers. Upper line is for a typical high-efficiency speaker with free-air resonance of 50 Hz. It shows such speakers are not suitable for compact closed box.

The experience gained in making the tests for the design charts in this article indicated that the degree of change or stability could not be determined just by observing the type of suspension material used. Polyurethane foam tended to be stable, but one such speaker was found to have changed more over a period of a few years than any other speaker tested for stability.

To guard against undue change in resonance, a new speaker should be exercised for a while before testing. The tests should be conducted under conditions of normal air humidity (similar to that of the room in which the speaker will be used). A test conducted in a humid basement may show a significantly lower resonance than one conducted in a dry attic.

If new woofers are stored in their shipping cartons for some time before use, the boxes should be turned on a side so that the cones are vertical. The heavy cones of some high-compliance large woofers will drift out of position if stored with the cone horizontal and the axis of the speaker vertical. If this occurs, it can be detected by inspecting the spider assembly for flatness. A drifted cone will have moved the voice coil away from its proper position in the region of maximum magnetic flux density, a situation that will produce more distortion than was designed into the speaker.

When a low-resonance woofer is installed behind a speaker board, the cutout should be large enough to permit the suspension to flex without hitting the board. Any friction between the suspension and the board will raise the resonant frequency and generate distortion. Front mounting of speakers is desirable, particularly for high-frequency drivers. But behind-the-panel mounting is permissible if the front edges of the holes in the speaker board are rounded off to eliminate the sharp edges.

One high-compliance woofer characteristic that is sometimes overlooked is its high-frequency response. The crossover frequency for the typical high-compliance woofer must be placed at a lower point in the audio spectrum than for a conventional woofer. There are several reasons for this. The voice coils of high-compliance woofers must be long so that the cone can move freely and yet not take the voice coil out of the high-flux range. This longer coil adds inductance to the electrical circuit and acts as a low-pass filter. Too, high-mass cones do not respond well to high frequencies.

A rule of thumb is to use no tweeter with the smaller full-range speakers. A two-way system is suggested for woofers up to 10 inches in nominal

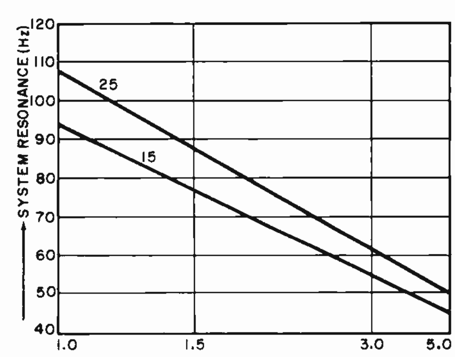
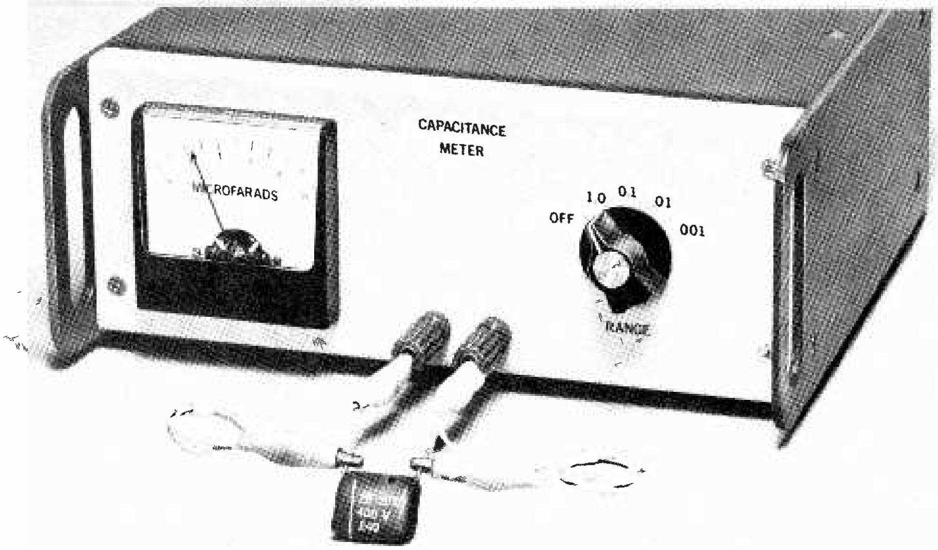


Fig. 11. Design chart for 15-in. woofers.

diameter. Most 12- and 15-inch woofers operate best when the crossover point is at 500 Hz or, at most, 1000 Hz. For systems that use these woofers, either a small full-range speaker must be selected for the tweeter or a three-way crossover network and a separate midrange speaker and tweeter must be used.

The closed-box enclosure appears to be utterly simple. It is—in the sense that there is no unique volume that is mandatory for a given speaker. But haphazard enclosure design will not provide satisfactory bass response and suitable power-handling ability. A reasonable value for box volume insures good performance, a challenge that can be met by the average audio buff. ◈



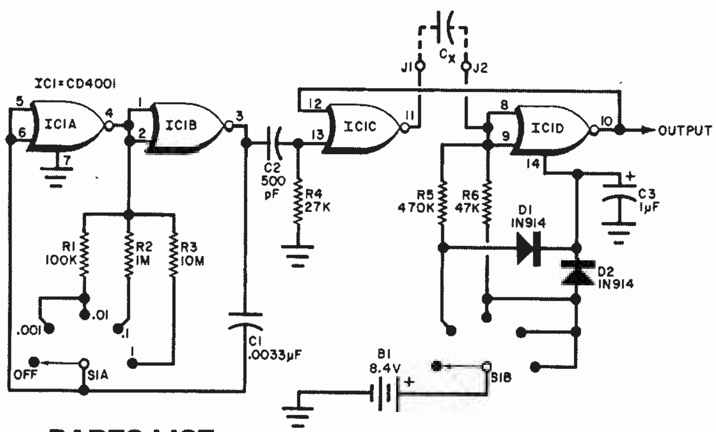
SINGLE-IC CAPACITANCE METER

Measures from 100 pF to 1 µF on a linear meter scale.

BY HARRY GARLAND AND ROGER MELEN

THERE are meters available to the electronics experimenter and service technician that can be used to measure just about any quantity in electronics. However, when it comes to measuring capacitance, the meter is probably complex and expensive. There are some instruments that have a capacitance measuring feature; but they usually have some form of bridge circuit that requires the nulling of a meter through the simultaneous operation of two or more (sometimes interacting) controls. In using this type of instrument, it is often difficult to find the correct meter null unless the approximate value of the capacitance is known before the measurement begins.

The simple capacitance meter described here is very easy to use, requiring only the connection of the unknown capacitance to a pair of binding posts and operation of a single rotary switch. When the correct range is found, the meter will indicate upscale to give the capacitance value. The indications on the meter scale are linear and the instrument covers a range from 100 pF to 1 µF. If desired, an external digital multimeter or VTVM can be used as the readout instead of the built-in meter. Since the meter is powered by batteries, its operation is independent of the power line.



PARTS LIST

B1—8.4-volt or 9-volt battery
C1—0.0033-µF capacitor
C2—500-pF capacitor
C3, C5—1-µF capacitor
C4—2000-µF electrolytic capacitor
D1, D2—1N914 silicon diode
IC1—CD4001 CMOS quad "nor" gate
J1, J2—5-way binding post
Q1—2N3565 transistor
M1—0.1-mA dc meter
R1, R11—100,000-ohm, ¼-watt resistor, 5%
R2—1-megohm, ¼-watt resistor, 5%
R3—10,000-megohm, ¼-watt resistor, 5%
R4—27,000-ohm, ¼-watt resistor, 5%
R5—470,000-ohm, ¼-watt resistor, 5%
R6—47,000-ohm, ¼-watt resistor, 5%
R7—10,000-ohm, ¼-watt resistor, 5%
R8—5000-ohm potentiometer
R9—470-ohm, ¼-watt resistor, 5%
R10—2700-ohm, ¼-watt resistor, 5%
R12—100,000-ohm potentiometer
S1—Dp 5-pos. rotary switch
Misc: Perf board; suitable chassis; switch knob; mounting hardware; etc.

Fig. 1. Unknown capacitor forms part of accurate pulse-duration generator, whose output is directly proportional to unknown value. This output is metered.

How it Works. Most of the circuit (Fig. 1) is contained on a single IC, a CMOS quad NOR gate whose extremely low power requirement ensures long battery life.

Gates *IC1A* and *IC1B* are connected to form an astable multivibrator whose frequency of operation is determined by the value of *C1* and a resistor selected by *S1A*. This signal is coupled through *C2* to trigger *IC1C* and *IC1D*, wired as a monostable pulse generator whose output pulse duration is determined by the value of the unknown capacitance (*Cx*) connected between *J1* and *J2* and the resistance value selected by *S1B*. If the selected resistor value is accurately known, the output pulse duration is then determined by the unknown capacitor.

In the prototype meter shown in the photo, the output pulse duration is measured by the circuit shown in Fig. 2A, where the readout is on a milliammeter. In this circuit, *Q1* is used as a saturating switch while *R8* is used to calibrate the meter. Since the meter

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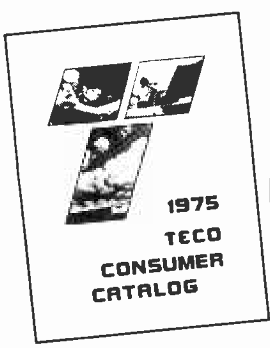

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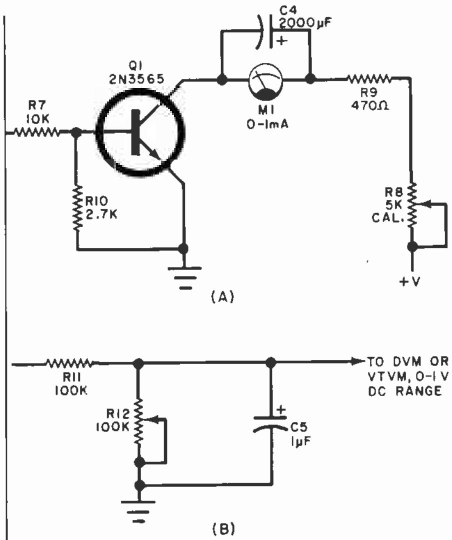

Fig. 2. (A) is used when built-in metering is desired, and (B) is for external metering.

indicates dc current flowing through *Q1*, and since the amount of dc current is directly related to the pulse duration, the meter can be calibrated directly in capacitance. Capacitor *C4* is used to integrate the dc pulses appearing across the meter; it thus removes the ac component.

The circuit in Fig. 2B is used when an external digital voltmeter or VTVM (1-volt dc range) is used as the readout instead of *M1*. In this circuit, *R11* and *R12* operate as a voltage divider while *C5* filters out the ac component.

Construction. The circuit can be assembled on a piece of perf board, using a socket for *IC1*. Switch *S1*, the two binding posts, and the meter (if used) are mounted on the front panel of the selected chassis. The battery is mounted in a holder on the perf board.

Calibration. Connect a known value of capacitance (5% or better tolerance) between *J1* and *J2*. Place the range multiplier switch, *S1*, in the appropriate position and adjust *R8* until the meter indicates the correct capacitance. If you are using the external metering device, set it to its 1-volt dc range and adjust *R12* for the correct indication. The calibration on one range suffices for all other ranges.

The accuracy of the instrument is limited by the accuracy of resistors *R1*, *R2*, *R3*, *R5*, and *R6*. Although 5% tolerance is adequate in most cases, you can use more precise resistors, or trim each range individually with small potentiometers. If you decide to trim each range separately, use a separate precision capacitor for each range. Trim *R1* before trimming *R5*. ◈

CONTINUITY TESTER

When installing a new member of an antenna farm, or trying to do some simple troubleshooting in a car, it is helpful to have a continuity tester to trace leads. An inexpensive door bell and battery can perform this function. Hook up the doorbell at either end of the antenna feedline and place the battery across the various feedlines until the bell is heard. This isolates the desired line. For auto "hot" wires, ground one side of the doorbell and search out the "hot" lead with a test lead attached to the other side of the bell.

—C. W. Hart, Jr.

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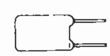

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7473	.62
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7476	62
7490	.92
7492	92
7493	.92
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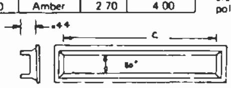

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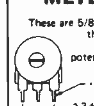

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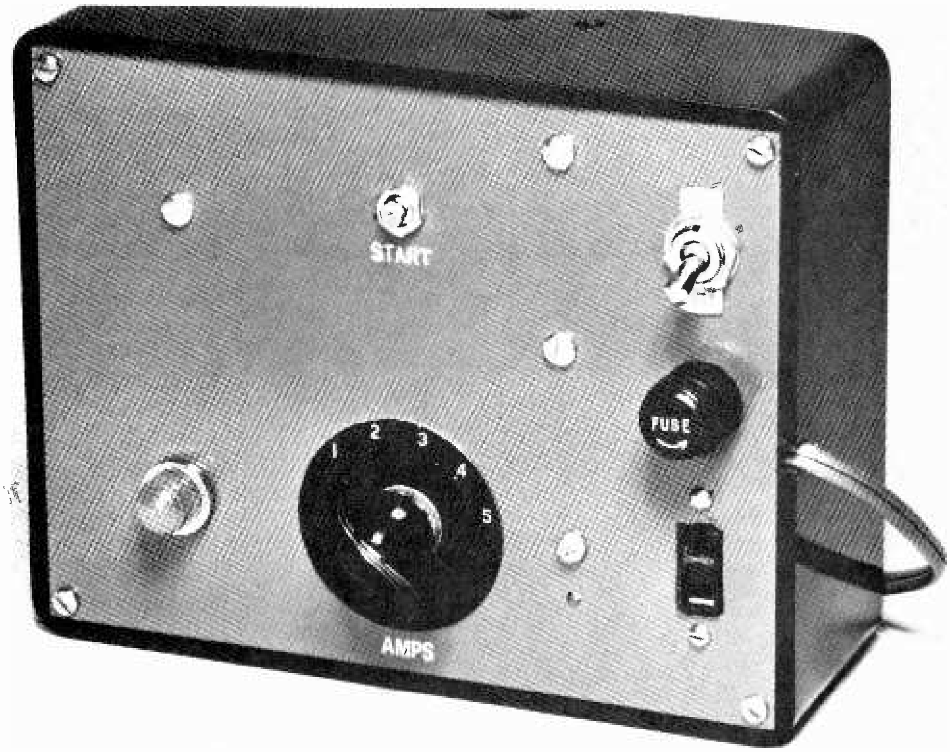


FAST-ACTING RESETTABLE ELECTRONIC FUSE

Provides adjustable (1 to 5 A) circuit-breaker action.

BY WILLIAM A. RUSSO

FOR YOUR own protection, to say nothing of your equipment, electronic apparatus should be protected by fuses or circuit breakers in their ac line inputs. But all too often, searching for the correct fuse and holder and wiring them into place are chores that are forgotten. While you can get away with doing things this way most of the time, sooner or later a puff of smoke or a nasty shock or burn are going to make you wish that you had taken the proper precautions.

The Electronic Fuse described here is designed to act as an adjustable *temporary* circuit breaker for projects undergoing tests or for any line-powered device (rated up to 600 watts) being tested or serviced on your workbench. Operating currents of from 1 to 5 amperes (in 1-A steps) can be selected. Detection of even a small overload results in fast (less than 0.5 second) interruption of both lines feeding the load. Then, instead of replacing a blown fuse, you simply flip a switch to restore normal power—after remedying the overload problem, of course.

Theory of Operation. Referring to the schematic diagram (Fig. 1), load current flows through *R1* and, depending on the setting of *S2, R2* induces a voltage drop that is sensed at the gate of the triac *Q1* through the appropriate gate resistance (*R3-R11*). When sufficient gate current flows to turn on *Q1*, the triac energizes *K1* and indicator *I1*. The contacts of *K1* are wired in a latching arrangement, while at the same time isolating the load when the relay is energized. Interrupting power by opening *S1* resets the circuit.

Pushbutton switch *S3* and resistor *R11* reduce the sensitivity of the sensing circuits to the turn-on transients of reactive or incandescent lamp loads. Potentiometers *R4* through *R8* are used for calibration.

Construction. Shown in Fig. 2 are the foil pattern and component placement diagram to be used when making the PC board and mounting components on it. Once the board is wired as shown (substitute heavy-duty perforated board if you wish), 1-inch spacers permit the board to be safely mounted on the underside of the chassis box's top. Mount the board in place before making an external connection to it.

Next mount and wire *S2* and *S3*, followed by *K1* and the remaining components. Use 18-gauge insulated wire for all leads carrying load current. When routing the power cord through a hole drilled in the chassis box, it is best to use a standard force-fit strain relief to hold it in place. However, lacking a standard strain relief, you can tie a figure-8 knot in the cord after routing it through a hole lined with a rubber grommet.

Align and drill five ¼-inch holes through the side of the chassis box to provide direct-in-line access to the adjustments slots of *R4-R8* when the Electronic Fuse is fully assembled. Be-

PARTS LIST

F1—6-ampere fuse
K1—117-volt ac relay with 5-A dpdt contacts
PL1—Neon pilot-lamp assembly
Q1—Triac (RCA 40529)
R1, R2—1-ohm, 10-watt power resistor
R3—47-ohm, ½-watt resistor
R4,R5—1500-ohm vertical-mounting trimmer*
R6,R7—3000-ohm vertical-mounting trimmer*
R8—5000-ohm vertical-mounting trimmer*
R9—1500-ohm, ½-watt resistor
R10—1800-ohm, ½-watt resistor
R11—12,000-ohm, ½-watt resistor
S1—10-ampere dpst toggle switch
S2—Dp 5-pos. non-shorting rotary switch (Mallory 173C or equiv.)
S3—Spst normally closed miniature push-button switch
SO1—Chassis-mounting, 3-conductor ac receptacle
Misc.: Chassis-mounting fuse holder for F1; three-conductor ac line cord with plug; 6¾" × 5¼" × 2¼" Bakelite (or metal) chassis box with cover; hookup wire; solder; etc.

*Pots are Mallory Type MTC-1 or equiv.

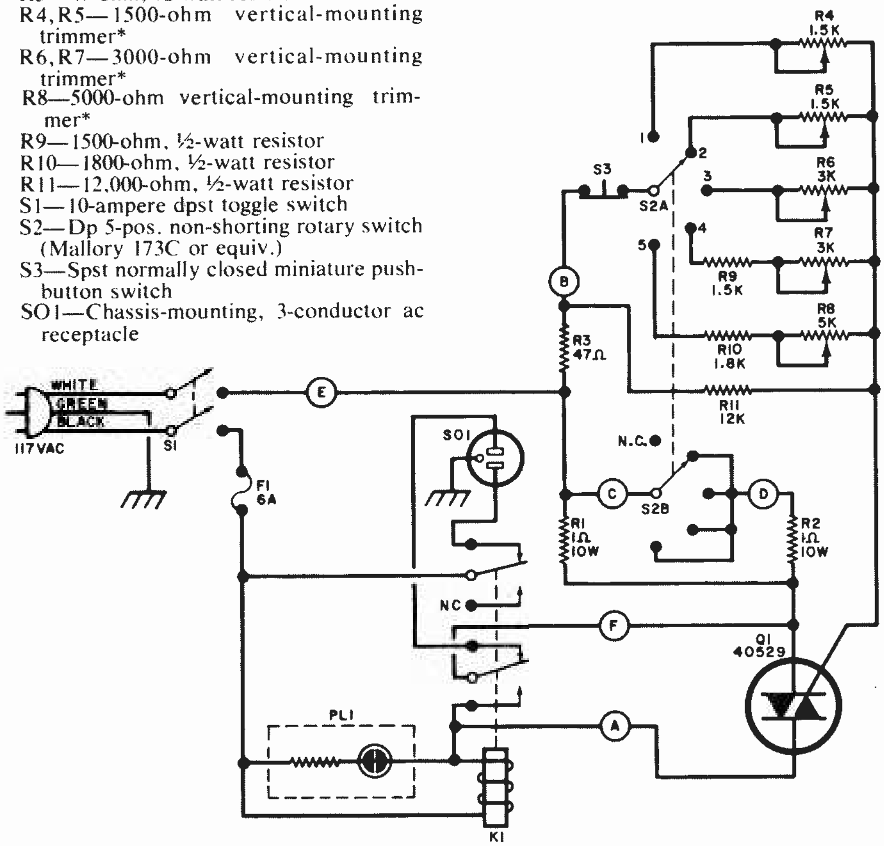


Fig. 1. In the fast-acting electronic fuse circuit, a triac is used to actuate the relay to cut off power.

fore finalizing the project, check to make certain that *K1*'s armature operates freely when the chassis box is assembled. Then, use an ohmmeter and visual inspection techniques to make absolutely certain that no current-carrying portion of the circuit touches chassis ground. The only grounded items in the circuit should be the green (neutral) wire of the power cord and the *round* contact of *SO1*. Sometimes this socket contact has a green-tinted screw for easy identification.

Calibration and Use. For calibration, you will need the following loads:

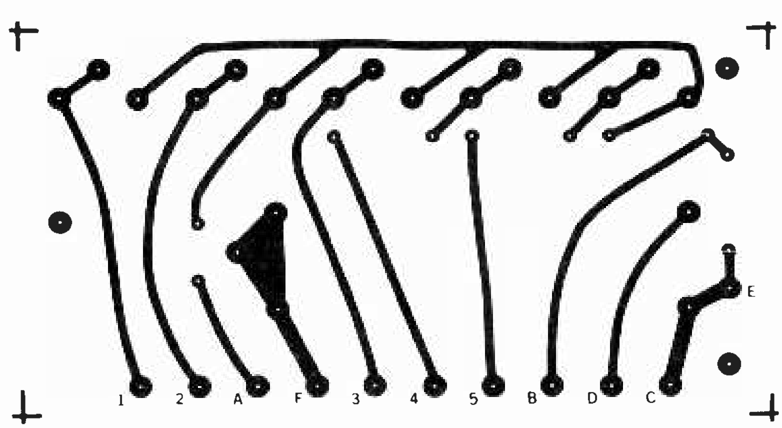
S2 Setting (A)	Output Load (W)
1	150
2	275
3	375
4	500
5	650

Incandescent lamps connected in parallel can be used in conjunction with *S3*. For heavier loads, heating coils (hair dryer, toaster, etc.) are ideal. You will also need an *insulated* screwdriver to adjust *R4-R8*. (Note: If you lack an insulated driver, you can slip a length of insulated tubing over the shank of an ordinary metal driver. Alternatively, you can drill the holes oversize and line them with rubber grommets.)

To calibrate each range, set *S2* to the proper position and plug the required load into *SO1*. Set the appropriate calibration pot to its maximum resistance setting and turn on the Fuse, using *S3* as necessary during turn-on. If *K1* pulls in, check all settings and repeat the turn-on procedure.

With the load operating, slowly adjust the pot with the insulated screwdriver until the relay just pulls in. Turn off the project and repeat the above procedure for each setting of *S2*.

Now with the Fuse plugged into an ac outlet, plug the device to be protected into *SO1* and set the range switch for the proper fuse current. Turn on the Fuse, then the load. If the Fuse trips, reset it by turning off the power and then on again. Restart the system by momentarily pressing *S3* while switching on the Fuse with the load turned on. If the Fuse's relay trips again, check the setting of *S2* to verify whether or not it is switched to the correct current range, or look for a malfunctioning load. It should be noted that, if the load is incandescent lamps, it may require three or four starting attempts. ◈



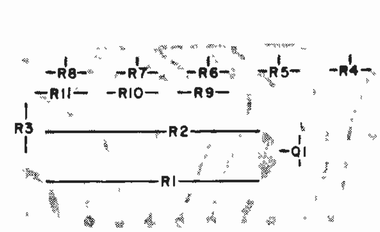


Fig. 2. Foil pattern that can be used for electronic fuse is shown above, component layout at right.

HOW TO AVOID WORKBENCH HAZARDS

Don't be careless when working with electronics.

EVERY year, thousands of electronics professionals and hobbyists suffer the painful and sometimes lethal effects of electrical shock while at their workbenches. Most are lucky enough to come away from the experience with a bruise, a broken bone, or a painful memory and a new respect for the power of electricity. Those who fail to come away from it become statistics.

These accidents need never have occurred if the victims had adopted a sensible work plan and geared themselves physically and mentally to avoid multiplying the shock hazard. You can minimize the shock hazard on your workbench by a few simple expedients and good common sense.

In this article we will be discussing some of the practices you should adopt whenever you work on line-powered and high-voltage circuits and equipment. We will detail the conditions under which you should avoid working near potentially dangerous voltages and describe what you can do to make your working environment a safer place.

Safety Practices. Let us begin with the common denominator—you. You can do everything possible to make your shop really safe, but if you are a "walking disaster," accidents will follow you on the job.

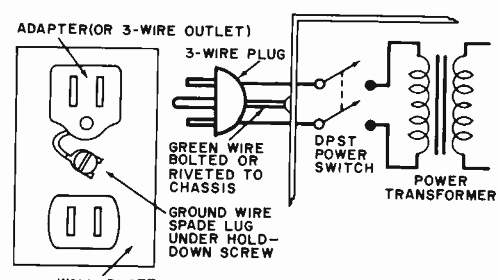
First, never work on an electronic device—powered or not—while wearing jewelry, such as a wristwatch, ring, etc. The workbench is no place for jewelry or other items like ties and dangling laces that can get hung up on the equipment in an emergency or even trigger an emergency.

Be practical about what you wear on the job. You are at your best when comfortably dressed. So, wear a long-sleeved shirt, buttoned at the wrists and open at the collar, and rubber-soled shoes.

Whenever you are working on a circuit or chassis where high voltages are present, keep your mind and eyes on what you are doing. Don't look away to observe a meter reading or a scope waveform if you are touching a test prod to a point in a powered circuit. Do your job the way a professional would: With the power to the equipment under test turned off, connect the test leads. Turn on the power, take your reading, and turn off the power. Only after the power has been turned off should you remove the test leads from the equipment. If you do the job the unsafe way, your eyes have to leave the work to take the reading, in which case the probe tip might slip. Chances are that you will overreact and get yourself into more trouble.

It takes only about 10-20 μA of current coursing through the heart to cause ventricular fibrillation, a usually fatal condition unless help and special equipment are immediately available. Currents as low as 100 mA entering a

Fig. 1. Recommended method of wiring cords and switches on electronic gear.



hand and leaving the body via the other hand or a foot can generate a fibrillatory current in the heart. So, never reach into a high-voltage circuit with both hands, and never rest one hand on the chassis while reaching into the circuit with the other hand. To avoid temptation, keep your free hand in a pocket or behind your back.

If you plan to work on unpowered equipment in which high voltages are developed, *make certain that the line cord is unplugged* and that you discharge all electrolytic capacitors in the high-voltage circuits. Electrolytic capacitors can hold a potent charge long after power is shut off; so, don't take chances. (Remember that charges too small to be lethal can inflict secondary injuries like bruises, lacerations, and broken bones as muscles violently and involuntarily contract upon contact. This can be a lifesaving move on the part of nature, by interrupting the through-the-body circuit, but it doesn't help if you crack your skull against a shelf or tear your flesh on a chassis.)

When Not to Work. Many electronics men go to work on circuits or equipment when they should be doing something else—like resting. There are definitely times when you should avoid going near electronic gear if you plan to stay healthy.

Hot, muggy environments cause a worker to perspire profusely and sap energy. A body covered with high-salinity perspiration becomes a fairly good conductor of electricity. Not only is the resistance over the surface of the skin reduced by perspiration, it provides a more direct current path between the skin and the interior of the body.

Cold environments can be equally hazardous. Cold has a numbing effect on the body, particularly in the extremities—like the fingers that hold test probes. Fingers that lose their normally acute sense of touch can easily make mistakes and do so all too often. Either heat the area or stay away.

Never approach a job if you are tired, angry, or emotionally upset. And don't try to work off excess energy at your workbench. (Go lift weights or do some jogging; it's safer.) Under these conditions, your attention is apt to wander—which is as bad as your eyes wandering.

The best time to go to work is when you are relaxed and alert. Stop working when you become fatigued or bored, and take frequent rest breaks.

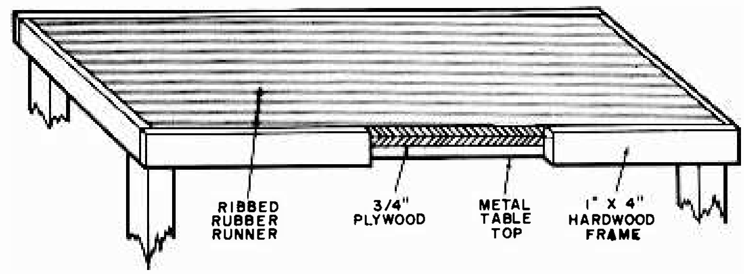


Fig. 2. How to make a workbench with a metal top safe by adding insulation.

Your Equipment and Workshop. Many electronics men who practice proper safety measures give little thought to their test equipment and workshops. This is particularly true of the hobbyist who works in a basement or attic where environmental conditions are hardly conducive to safety.

Line-powered test gear is a particularly vulnerable point. Under no circumstances can a line-powered instrument be considered safe if it is equipped with a two-conductor line cord. It is even less safe if only a single-pole, single-throw power switch is used. All two-conductor line cords should be replaced with three-conductor cords, and all instruments should be equipped with double-pole, single-throw switches. The recommended method for wiring the cords and switches into your gear is shown in Fig. 1. While you are at it, carefully inspect all power cords and plugs, replacing any that are frayed, loose, or worn.

Plug three-prong plugs into appropriate sockets or into adapters to mate them to two-conductor house wiring systems. If you use adapters, slip the spade lugs on the grounding wire under the outlet's wall-plate mounting screw and tighten down. When you have several instruments that have to be used simultaneously, your best bet is to use a circuit breaker or fuse-protected heavy-duty power-line outlet box. In this case, you need only one adapter in a two-conductor house wiring system.

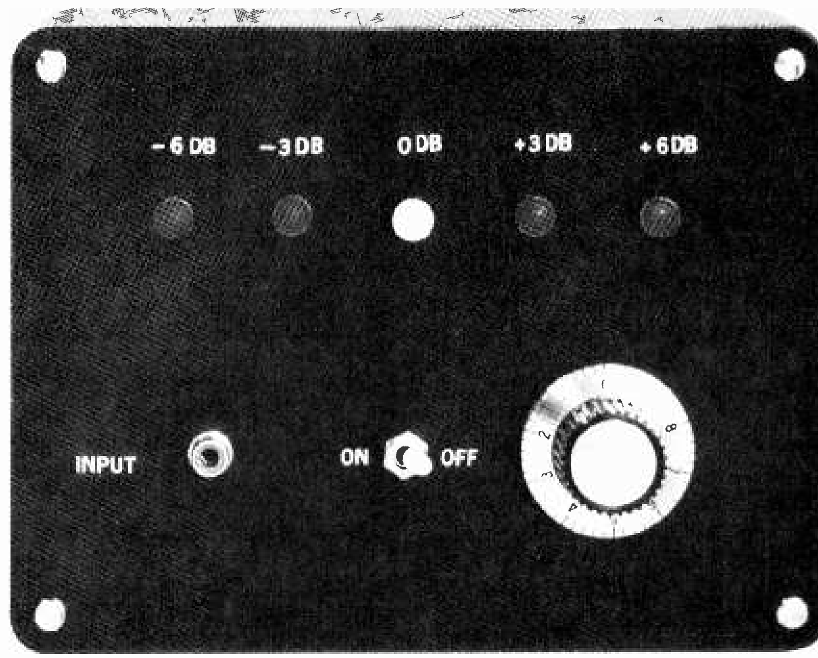
If you want to be really safe at your workbench, consider installing a ground-fault interrupter (GFI) in the bench's power system. The GFI is a fast-response device that disconnects power from the load whenever leakage current exceeds a specific amount (typically 5 mA). Don't install the GFI in the room's entire electrical system or it might extinguish the lights when it trips—a safety hazard in itself as you grope around in the dark and stumble over things.

Finally, make your work area safe and livable. In a damp basement where the floor is of raw concrete or in an attic where the floor is of unfinished lumber, lay vinyl flooring. Both areas will benefit enormously from a few sheets of hardboard nailed over exposed studs and rafters. Before installing the hardboard, however, make sure that there is adequate weather insulation between the exposed studs and rafters. A casement vent in the basement or a through-the-wall attic vent, each equipped with an exhaust fan to allow free circulation of air, will keep either area relatively dry and odor-free. While you are fixing up your work area, install adequate lighting. Any good handbook on home improvements will tell you how to do these things.

Wood is the best material for an electronics workbench, but if you must use a table with a metal top, it will have to be made safe. You will need two sheets of ¾-inch plywood cut ⅛ inch longer and wider than the dimensions of the table top. Cement the plywood sheets together and clamp overnight. Then top them with a ribbed synthetic rubber runner, held in place with contact cement, to provide a durable non-skid work surface. Finally, glue and nail a hardwood frame around this assembly, as shown in Fig. 2. When finished, the worktable surface should slip over the metal table top. Do not fasten the work surface to the top of the table.

If you do everything we have suggested, your chances of being injured (or worse) in your workshop will be minimal. But again, we must caution you. Don't relax your guard or take shortcuts. To do so is just inviting trouble. ◈

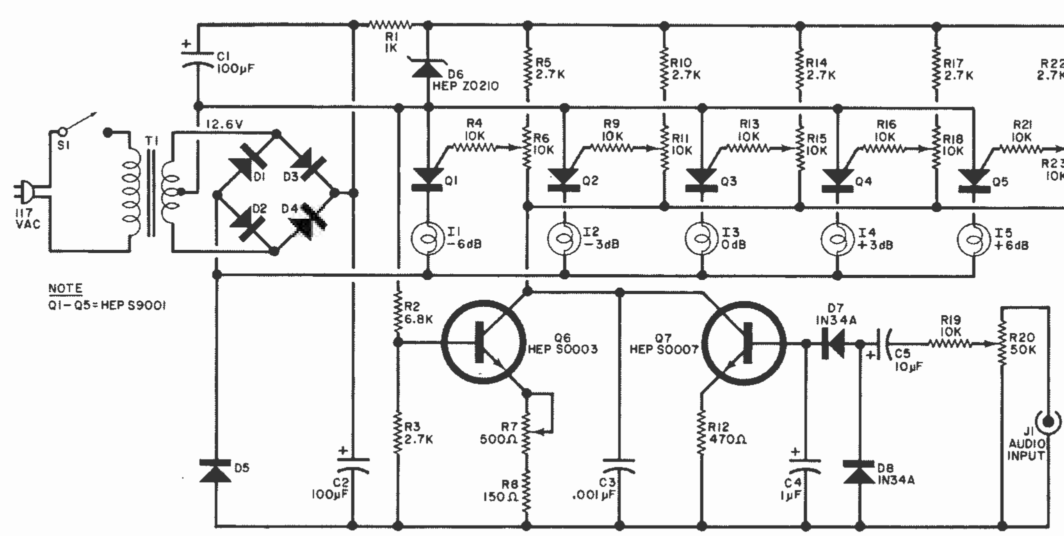
Lamp-Readout VU Meter



Audio recording simplified with discrete lamp indicators.

BY HERB COHEN

ONE of the biggest headaches involved in making an audio recording is keeping track of the signal level as indicated by a bouncing VU meter needle (or needles if you are recording in stereo). This requires close visual monitoring of the meter. By using the lamp-readout level indicator described here, you can simplify the process considerably since the indicators are positive and can be seen from a fair distance.



Gate voltages of five programmable transistors (Q1 through Q5) are preset so that they come on in sequence for each succeeding 3 dB of input.

PARTS LIST

C1, C2—100-µF, 25-volt electrolytic capacitor
C3—0.001-µF capacitor
C4—1-µF, 10-volt electrolytic capacitor
C5—10-µF, 10-volt electrolytic capacitor
D1—D5—100-V, 500 mA silicon rectifier diode
D6—4.7-volt zener diode (HEP Z0210)
D7, D8—Germanium diode (1N34A or equiv.)
I1—I5—6-volt, 50 mA lamp (Muralite L-6/50 or equiv.)
J1—Phono jack
Q1—Q5—Programmable unijunction transistor (HEP S9001)
Q6—Npn transistor (HEP S0003)
Q7—Npn transistor (HEP S0007)
R1—1000-ohm resistor
R2—6800-ohm resistor
R3, R5, R10, R14, R17, R22—2700-ohm resistor
R4, R9, R13, R16, R19, R21—10,000-ohm resistor
R6, R11, R15, R18, R23—10,000-ohm miniature potentiometer (Calectro B1-644 or equiv.)
R7—500-ohm miniature potentiometer (Calectro B1-642 or equiv.)
R8—150-ohm resistor
R12—470-ohm resistor
R20—50,000-ohm potentiometer (Calectro B1-685 or equiv.)
S1—Spst switch
T1—Transformer, 12.6 VCT secondary (Radio Shack 273-1505 or equiv.)
Misc.: Suitable cabinet, pilot lamp jewels (2 green, one clear, 2 red), knob, perf or PC board, mounting hardware.

Five lamps are mounted in a row, indicating −6, −3, 0, +3, and +6 dB. The lamp in the center is a clear jewel, the negative-dB indicators are green, and the positive lamps are red.

About the Circuit. As shown in the diagram, the circuit consists of a bank of programmable unijunction transistors (*Q1* through *Q5*) used as comparator switches. Their gates are reverse-biased by the voltage drop across *D6*.

The audio signal is applied to *J1* (with level set by *R20*). It is rectified by *D7* and *D8* and filtered by *C4*, which also determines the rate of change of the lamps. Transistor *Q7* converts the rectified signal into a current which pulls down the gates of the unijunction transistors to turn them on and illuminate their respective lamps. Which lamp is turned on is determined by the settings of potentiometers *R6, R11, R15, R18,* and *R23*. Transistor *Q6* acts as a constant-current preload for the gate line to preset the firing level and increase the sensitivity of *Q7*.

Diode *D5* isolates the filtered portion of the power supply from the unfiltered section. This allows the lamps to be turned off when the gate goes off.

Construction. Since the operation of the circuit is not critical, any type of construction can be used—perf board or printed-circuit board. All components except the lamps, input jack, and controls can be mounted on the board. For stereo operation, two VU indicators are needed.

Calibration. Before turning on the power, set pots *R6, R11, R15, R18,* and *R23* so that their movable contacts are toward the associated fixed resistors. Set *R20* at its maximum.

Turn on *S1* and rotate *R6* to the opposite end of its travel. This is the maximum sensitivity for the −6-dB lamp circuit. Adjust *R7* until the −6-dB lamp comes on, then back it off until the lamp just goes out.

Apply a 1-kHz (600-ohm, 1-mW reference) audio signal to *J1* at a −6-dB level (0.39 volt on an audio voltmeter) and adjust *R7* until the −6-dB lamp just comes on.

Increase the audio input signal to −3 dB (0.55 volt) and adjust *R11* until the −3-dB lamp just comes on. Increase the input signal level in 3-dB steps until the remaining lamps have been calibrated.

Once all lamps have been calibrated, recheck all positions to see that nothing has changed accidentally. If a zero-dB level other than the 600-ohm, 1-mW reference is used, adjust *R20* for the new reference.

Use. The lamp-readout VU meter can be connected to the line output of a tape deck or to the junction of the preamp and main amplifier of an audio system. It can also be connected between the output of a mixer circuit and the following amplifier so that all signal levels can be properly set. It can be used with ham or CB rigs by connecting it to the modulator circuit so that, when 100% modulation is applied to the rig, the 0-dB lamp will come on. ◈

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BUILD A

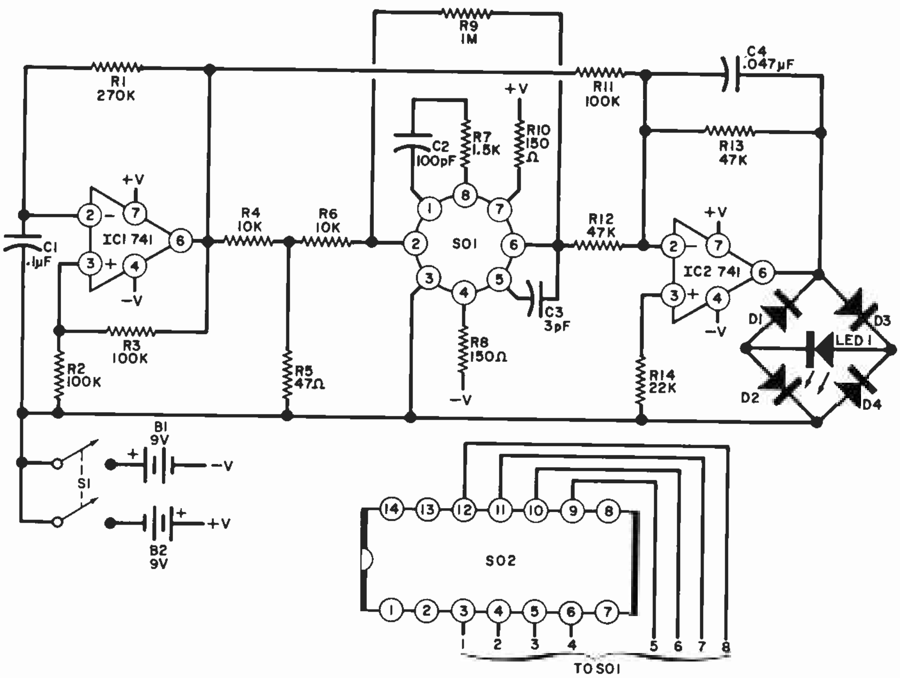
LOW-COST OP-AMP TESTER

Tests gain, stability, input offset and bias current.

BY HARRY GARLAND AND ROGER MELEN

As OP-AMPS become increasingly popular and useful, there is a growing need for a good, low-cost op-amp tester. The universal op-amp tester described here can be used to test virtually all of the popular units. It automatically checks the important parameters and has a red light-emitting diode (LED) to indicate the condition of the op-amp.

The parameters which are checked by the tester are gain, stability, input offset voltage, and input bias current. The tester, which itself uses two op-amps, can test internally compensated op-amps, such as the 741, as well as uncompensated op-amps, such as the 709 and 748. Both 8-pin TO-5 and 14-pin DIP sockets are provided.



PARTS LIST

B1, B2—9-volt battery
C1—0.1-μF capacitor
C2—100-pF capacitor
C3—3-pF capacitor
C4—0.047-μF capacitor
D1—D4—1N914 diode
IC1, IC2—741 op-amp IC
LED1—Light-emitting diode (Poly Pak "Brite Red" or equiv.)
R1—270,000-ohm, ¼-watt resistor, 10%
R2, R3, R11—100,000-ohm, ¼-watt resistor, 10%
R4, R6—10,000-ohm, ¼-watt resistor, 10%
R5—47-ohm, ¼-watt resistor, 10%
R7—1500-ohm, ¼-watt resistor, 10%
R8, R10—150-ohm, ¼-watt resistor, 10%
R9—1 megohm, ¼-watt resistor, 10%
R12, R13—47,000-ohm, ¼-watt resistor, 10%
R14—22,000-ohm, ¼-watt resistor, 10%
S1—Dpdt momentary contact pushbutton switch
SO1—8-pin, TO-5 IC socket
SO2—14-pin, DIP IC socket

Misc.: LMB type 342 cabinet; rubber feet (4); mounting hardware; etc.

If op-amp plugged into SO1 is good, LED remains off; if not, LED blinks.

Circuit Design. Integrated circuit *IC1* is used as a square-wave generator and the op-amp being tested is used as an inverting amplifier with gain of 100. The output of *IC1* is applied to both *IC2* and (through voltage divider *R4* and *R5*) to the unit being tested. The output of the latter is applied as a second input to *IC2*, which is used as a summing amplifier.

If the op-amp being tested is good, its output will exactly cancel the square wave applied to *IC2* through *R11*. When these two signals cancel, there is zero output from *IC2* and *LED1* will not light.

If the op-amp being tested is bad, the two inputs to *IC2* will not cancel and *LED1* turns on. Before *LED1* turns on, however, the output of *IC2* must exceed the threshold determined by the forward voltage drops of two of the bridge diodes (*D1* to *D4*) and *LED1*. Assuming a failure, this threshold will be exceeded if the op-amp has a gain of less than 60, an input offset voltage greater than 30 mV, or an input bias current greater than 3 microamperes. Any of the popular IC op-amps should have parameter values better than these. Similarly, the LED will turn on if the op-amp is unstable in the test circuit or has any "shorts" or "opens."

Construction. Almost any type of assembly method can be used for the tester. For the prototype, the components were assembled on two perf boards mounted in a 2⅜" × 4¼" × 3" metal cabinet. Printed-circuit boards could also be used.

The two test sockets (*SO1* and *SO2*) were mounted on one perf board with their associated components, and the board was fixed to the upper inside surface so that both sockets protruded through holes cut in the upper surface.

The second perf board (with *IC1* and *IC2* and their related components) was mounted on the bottom of the cabinet with the two batteries. The LED was glued (with epoxy) to protrude through a small hole in the front of the cabinet with *S1* mounted beside it. Four rubber feet on the bottom of the cabinet will keep the tester from slipping around when in use.

Operation. With no op-amp in either test socket, depress *S1*. The LED should flash on and off, indicating that the circuit is operating properly. To test an op-amp, plug it into the appropriate test socket and operate *S1*. If the op-amp is good, the LED will not flash.

Any of the popular IC op-amps with the same pin configuration as the 709 can be checked. This includes the 101, 301, 740, 741, and 748. Units with other pin configurations, such as dual op-amps, can also be tested if extra test sockets are wired in parallel with the existing sockets.

Since different op-amps have different specifications, a "good" indication does not necessarily guarantee that the op-amp meets all of the requirements. However, for nearly all practical applications, the test will provide a valuable go/no-go decision. You will find the tester particularly useful for sorting through bargain, untested op-amps for quickly isolating the trouble in an op-amp circuit that doesn't work. ◈

TREMOLO ADAPTER

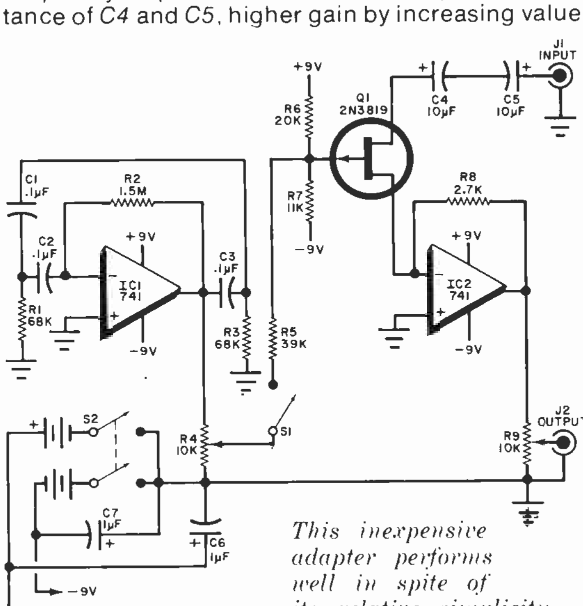
BY DEANE A. GARDNER

HERE'S an inexpensive tremolo adapter that's easy to add to an electric guitar or other electronic musical instrument. The ideal tremolo would be a low-frequency (6 to 10 Hz) sine-wave oscillator driving a non-distorting, voltage-controlled amplifier (vca). Some circuits employ triangle-wave modulation or a nonlinear vca, which can cause undesirable clicks or distortion. This adapter avoids such problems with a sine-wave modulation signal to control the channel resistance of an FET.

As shown, *IC1* and its associated components form a phase-shift oscillator. The output of this oscillator is attenuated by *R4* and *R5* and then fed to *Q1* via *S1*. You can change the value of *R5*, which affects modulation depth, to suit the gain of the FET used. A lower resistance increases depth, but avoid going below 30,000 ohms or the FET will become reverse-biased.

The oscillator output adds or subtracts from the bias level set by *R6* and *R7*. The voltage on the inverting input of *IC2* will always be very close to ground level. Therefore, the gate-source voltage of *Q1* is dependent only upon gate voltage relative to ground, resulting in a low-distortion modulation of the signal on the drain terminal. The output of *IC2* is attenuated by *R9*.

With a 1-volt peak-to-peak drive, the frequency range of the tremolo adapter is 40-50,000 Hz. Extended low-frequency response can be obtained by increasing capacitance of *C4* and *C5*, higher gain by increasing value of *R8*.



This inexpensive adapter performs well in spite of its relative simplicity.

Use of a PC or perf board and solder clips is strongly recommended. Use a low-wattage soldering iron. (*S1* can be a footswitch, but it must be sturdily mounted.)

When assembly is complete, plug the instrument to be used with the adapter into *J1* and a power amplifier into *J2*. If any clipping circuits, such as a fuzz box, are to be used, they must be placed between the instrument and adapter. Flip *S2* to power the adapter. (Note, it may take a few seconds for the tremolo oscillator to reach full output.) Put *S1* in the OUT position and adjust level control *R9* as desired. Set *S1* to IN and adjust *R4* for the desired depth of tremolo. That is all there is to it. ◈

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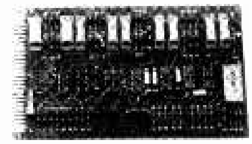

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CIRCLE NO. 8 ON FREE INFORMATION CARD

Getting the Most from your Transmitter

Some helpful hints for the ham or CB'er.

BY WILLARD R. MOODY, WA3NFU

THE POWER rating of your transmitter is one thing; but the amount of r-f that gets off the antenna is what counts. The rest is actually wasted.

The easiest way, in many cases, to make r-f power measurements, is to connect a calibrated r-f wattmeter to the transmitter output. A commercial r-f power meter is pretty expensive, however, for the occasional user, so it helps to have an inexpensive, easy-to-use device for making r-f power measurements. Once the output is measured, steps can be taken to improve efficiency.

The circuit shown in Fig. 1 is that of a commercial wattmeter, which is essentially a dc voltmeter connected across an appropriate dummy load whose resistance is the same as that required to terminate the transmitter properly. You can make your own wattmeter, using the circuit shown in Fig. 2. The dummy load, R_L, should be a noninductive resistor of the correct resistance and of sufficient wattage to withstand the expected transmitter power. Since the dummy load must be noninductive (no reactance at r-f), a wirewound resistor won't do.

The complete package should be shielded to reduce r-f radiation to a minimum while tests are being made. The diode rectifier can be any high-frequency type, while the two resistors (one a potentiometer) are used to set the meter scale.

For meter calibration, assume an expected r-f power of 3.5 watts into a 50-ohm dummy load (antenna). The measured voltage is equal to the square root of the power times the resistance which, in this case, would be 13.2 volts. Other voltage values for other r-f power and/or terminating resistors can be calculated and the meter (voltage calibrated) recalibrated in watts.

If you have a VTVM which is accurate at low frequencies and fairly accurate at high frequencies, you can measure the power indirectly by measuring the r-f voltage and using the equation $P = E^2/R$, where E is the measured voltage and R is the terminating resistance. For the example given above, $P = 13.2^2/50 = 3.5$ watts.

The R-F Ammeter. One of the handiest tools to have in making r-f power measurements and trimming up transmitting antennas is the r-f ammeter (usually a thermocouple type). Again assuming a 3.5-watt output into a 50-ohm load, the current is the square root of the power divided by the terminating resistance. The square root of 3.5 divided by 50 is about 0.26 ampere.

To use the r-f ammeter, connect it in series with the required dummy load as shown in Fig. 3. This will enable you to determine just how to tune the transmitter (and antenna network) to maximize the r-f output—the more current, the better. A table, or curve, can be plotted to relate antenna current to r-f power output.

Obviously, in the preceding calculations, you can use 52 ohms, 72 ohms, 300 ohms, or other values in place of the one used in the example. Use the value suggested by the manufacturer for that particular transmitter. In the case of a pi-network output, which can work into a wide variety of antenna loads, use the load value for your antenna. In any case, tune the transmitter and any associated antenna tuning network for maximum power in the dummy load, since this means maximum voltage (for the voltmeter approach) or maximum current (using an ammeter).

If for some reason you do not care to use either approach, load the transmitter with an ordinary light bulb whose wattage is about the same as the expected r-f output of the transmitter and tune for the brightest glow. Using a similar bulb operating at its specified voltage, a comparison can be made between light levels to closely approximate the power output.

Tuning the Antenna. After testing with a dummy load, it is important to make sure that the maximum power is transferred to the antenna. This entails a proper impedance match for maximum power transfer.

Assume that the transmitter is fed to a quarter-wavelength vertical antenna provided with the proper radials and transmission line. Connect the 50-ohm dummy load at the antenna end of the transmission line (Fig. 4A). Connect the ammeter in series with it (at the base of the antenna) and tune the transmitter for maximum current. After this is done, connect the transmission line to the base of the antenna. The antenna should have a sliding portion for adjusting the length; or it may have a variable capacitor at the

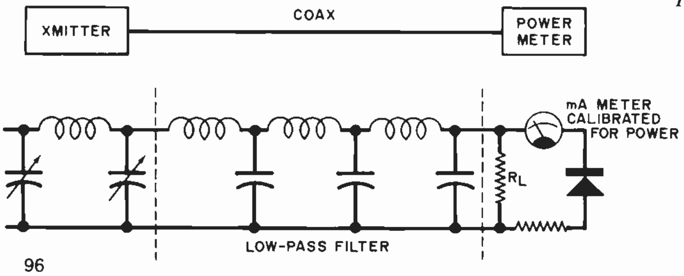


Fig. 1. A commercial power meter is a dc voltmeter measuring drop across dummy load resistor. (Filter represents coax.)

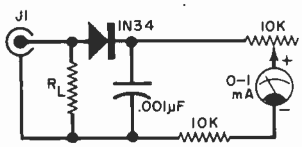


Fig. 2. In homemade power meter, potentiometer is adjusted to calibrate meter.

base (Fig. 4B) if the vertical is slightly longer than ¼ wavelength. A capacitor value of about 100 pF (at the appropriate voltage rating) can be used here. The capacitor is adjusted for maximum antenna current (at the base).

Theoretically, the resistance of a ¼-wavelength antenna (measured at the base) should be 37.5 ohms. The ratio of 50 to 37.5 gives a measure of the mismatch between the transmission cable and the antenna impedance. This comes out to 1.33:1; and although 1:1 is desirable, anything less than 3:1 is acceptable.

Refinements in impedance matching for a single operating frequency can be made by removing the center lead of the coax from the antenna connection lug (leaving the braid at the ground point) and sliding it up from ground until the meter indicates maximum antenna current. This approach is best done with a field-strength meter located several wavelengths away from the antenna, and with a second person using either a walkie-talkie or telephone calling out the changes in field strength as the antenna is trimmed.

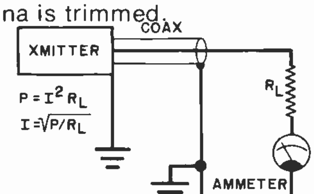


Fig. 3. Use ammeter to measure current flow through a dummy load to check output.

Transmission Line Tests. To determine how much power is lost in the transmission line, measure the transmitter power output with the dummy load connected directly to the transmitter antenna terminals. Then measure the power when the dummy load is connected across the far end of the transmission line, with the dummy load mounted very close to the portion of the antenna to which it will be connected. In the case of a dipole, the dummy load is suspended from the center insulator. For a base-fed antenna, the dummy is placed near the base connection points. The difference in measured power between the two tests will show just how much power is being lost in the transmission system.

(In using the above technique, neither the dummy load nor transmission line is connected to the antenna itself.)

If you are making measurements on a dipole antenna, then, by inserting the r-f ammeter in each leg of the transmission line, you can trim each side of the antenna for maximum current. These measurements can be made at a low r-f level to avoid r-f burns or transmitter damage in the event of accidental shorts. With a power of 1 watt and a center impedance of 70 ohms, the current at the center of the dipole would be 0.12 A. This is based on the premise that the antenna center impedance is almost 70 ohms but can vary due to height above ground and other conditions. It is preferable to use a balun so that the transmission line will not be unbalanced at the antenna feedpoint. An SWR bridge and/or a field-intensity meter can be used to make the measurements much easier.

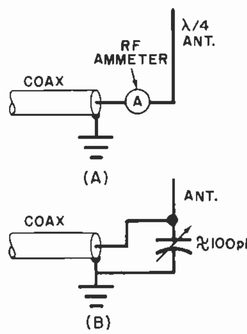


Fig. 4. Ammeter is used to check antenna current (A). A capacitor adjusts the antenna length (B).

It should be noted that every facet of r-f measurements cannot be covered in this brief article, but the role of the r-f ammeter deserves this attention since current is a fundamental parameter. ◈

ADD LIFE TO PILOT LIGHT BULBS

Replacing burnt-out pilot light bulbs in electronic gear can become tedious and expensive. Here's a simple way to extend bulb life many times. Insert a diode rectifier in the bulb voltage supply. Pulsating dc will be supplied to the bulbs, at about half the current previously drawn. The bulbs will operate at lower temperatures and last much longer. Light output will also decrease, but this is seldom a problem. If more brightness is required, use a higher wattage bulb.

CIRCLE NO. 28 ON FREE INFORMATION CARD

low-cost millivolter

Measures between 1 mV and 1 volt full-scale, and can be used as AC voltmeter.

BY RALPH TENNY

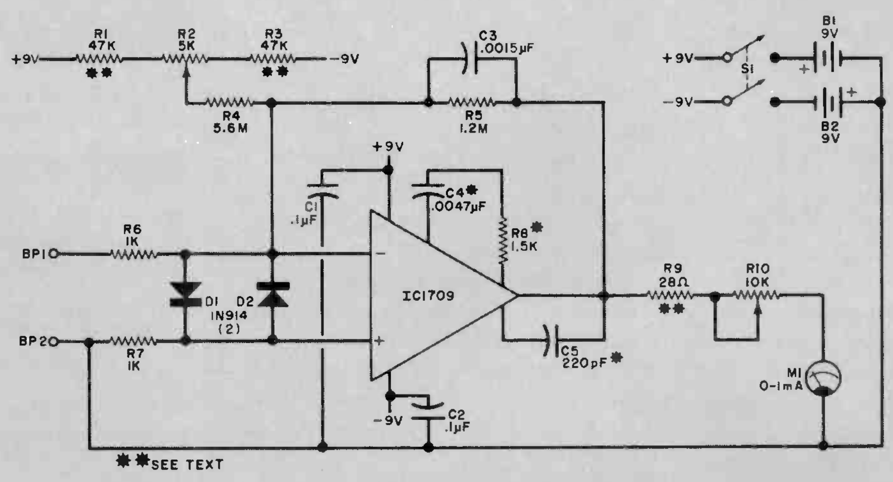
WITH the increasing trend toward the use of semiconductors and the necessity for measuring voltage in the range of 1 mV, it is apparent that the trusty old VOM has about reached its limit. However, if your VOM has a low-current range (preferably 1 mA full-scale) or if you have a 1-mA meter, all you need are a few low-cost components to make a dc voltmeter having a range of 1 mV to 1 volt full-scale with an input resistance of 10,000 ohms/mV.

Using the meter with a zero-center scale, accurate voltage settings, independent of the voltage level, can be made. Or, if desired, the circuit can easily be converted into a handy ac voltmeter.

As shown in Fig. 1, op amp *IC1* is connected as a current amplifier; that is, a change in current at the input causes a change in voltage at the output. The combination of *R9* and *R10* regulates the circuit sensitivity. The network, consisting of *R1, R2, R3,* and *R4*, furnishes an offset bias to balance out any static differential voltage in the input. This insures that the meter will indicate zero with no signal input. As an added feature, this network has sufficient range to set the meter for a zero indication at center scale. The final values of *R1* and *R3* can be between 18,000 and 68,000 ohms, depending on which type of op amp is used.

Capacitors *C1* and *C2* are power-supply filters and should be mounted as close to the op amp as possible. Resistor *R8* and capacitors *C4* and *C5* are used to compensate the 709 op amp used. If you use a 741 op amp, these three components are not required since the 741 is internally compensated.

Construction. The circuit can be assembled on perf board or on a small etched PC board. Do not use the chassis for a ground; the entire circuit should be "floating" with *BP2* as the common. Mount *D1* and *D2* as close to the op amp pins as you can. Check to make sure that you are using the correct pins on the IC and that both diodes are installed properly.



PARTS LIST

B1, B2—9-volt battery
BP1, BP2—5-way binding post
C1, C2—0.1-μF, 25-volt ceramic capacitor
C3—0.0015-μF ceramic capacitor
C4—0.0047-μF ceramic capacitor*
C5—220-pF capacitor*
D1, D2—1N914 or 1N662 diode
IC1—709 or 741 op amp
M1—0-1-mA meter
R1, R3—47,000-ohm resistor (see text)
R2—5000-ohm potentiometer
R4—5.6-megohm resistor
R5—1.2-megohm resistor
R6, R7—1000-ohm resistor
R8—1500-ohm resistor*
R9—28-ohm resistor (see text)
R10—10,000-ohm potentiometer
S1—Dpdt switch
*Use for 709 op amp only

Misc: Suitable case, battery holders mounting hardware, etc.

Fig. 1. Schematic and parts list for millivolter. The integrated circuit is connected as a current amplifier to drive 1-mA meter.

Checking. Set *R10* at its maximum resistance. Then set *R2* to its electrical center. With the power turned on, the meter should deflect very little from zero. This can be compensated for by adjusting *R2*. If the meter should deflect very much in either direction, shut off power and recheck circuit.

If the meter behaves correctly, slowly reduce the value of *R10* and adjust *R2* for meter zero. Repeat this procedure until *R10* is at its minimum and the meter is set for zero.

Calibration. Connect the meter as shown in Fig. 2A. With the values shown, the meter should deflect just about full-scale. Trim the value of *R10* for an exact indication.

Sensitivity is set by changing *R11* as follows: 10,000 ohms produces 1 mV full-scale; 100,000 ohms, 10 mV full-scale; 1 megohm, 10 mV full-scale; and 10 megohms, 1 volt. The meter scale can be calibrated accordingly.

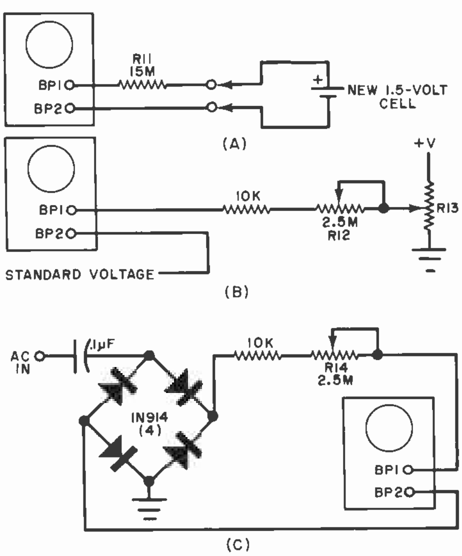


Fig. 2. Circuits for calibration (A), null indicator (B), and ac meter (C).

Uses. To use the meter as a null indicator for setting a precise voltage, use the circuit shown in Fig. 2B. Set the meter for zero using *R2*. Then start with *R12* at maximum resistance and adjust *R13* for zero indication. Keep reducing the value of *R12*, resetting *R13* for zero each time, until *R12* is at minimum resistance. The voltage at the rotor of *R13* will now be within a few microvolts of the standard.

To make the meter function as an ac voltmeter, use the circuit of Fig. 2C. Apply the ac voltage and, starting with *R14* at a high resistance, reduce it to get a more sensitive indication. Using an accurate source of ac, *R14* can be calibrated in range. ◈

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BUILD A PAIR OF SIMPLE ALARMS

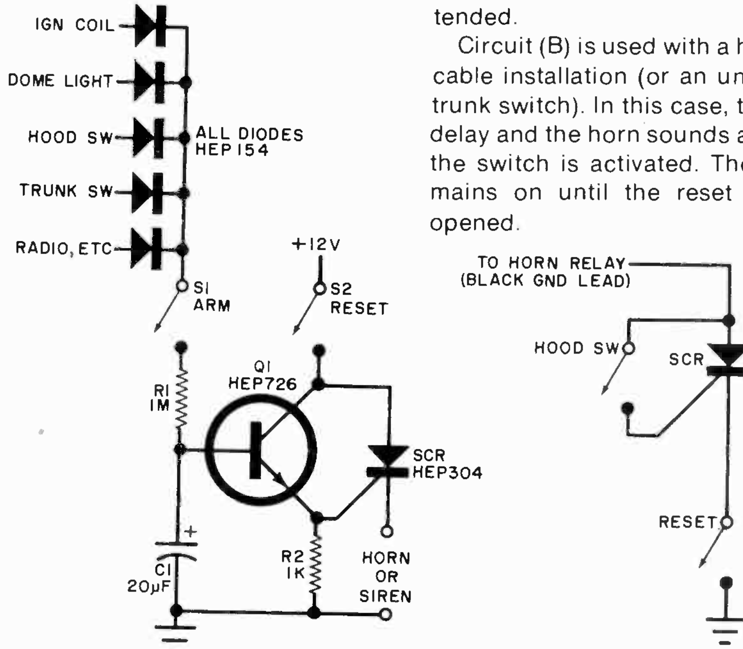
BY ANTHONY C. CAGGIANO

HERE are two simple automotive alarm circuits that can be assembled at very low cost, yet they work as well as many of the more complex systems currently available.

Circuit (A) uses a number of conventional silicon diodes whose anodes are connected to various points that have the 12-volt battery voltage applied to them when activated. These include the ignition coil, dome light, radio, tape player, etc. If normally open switches are installed also, the list can include the trunk, hood, or a pressure-sensitive switch under mat.

The operation of the circuit is quite simple. If both the arming switch (*S1*) and the reset switch (*S2*) are closed and the ignition is turned on (either by a key or a jumper wire), current flows through the associated diode and *R1*, putting a charge on *C1*. When the charge on *C1* is sufficient to cause *Q1* to turn on, the current through the transistor also turns on the *SCR*. The latter supplies current to a horn or siren. Once the *SCR* is on, it remains on regardless of the condition of the diodes or the transistor. Only the opening of the reset switch (which is concealed within the vehicle) will turn off the *SCR*. The values of *R1* and *C1* in the timing network are selected to provide sufficient delay for the owner to enter the vehicle. The arming switch is left open when the vehicle is in use and is closed when the vehicle is unattended.

Circuit (B) is used with a hood-latch cable installation (or an ungrounded trunk switch). In this case, there is no delay and the horn sounds as soon as the switch is activated. The *SCR* remains on until the reset switch is opened. ◈



(A) (B)

(A) If both switches are closed and potential applied at one of diodes at left, transistor turns on by charge on C1. (B) Horn sounds when switch is closed.

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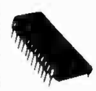

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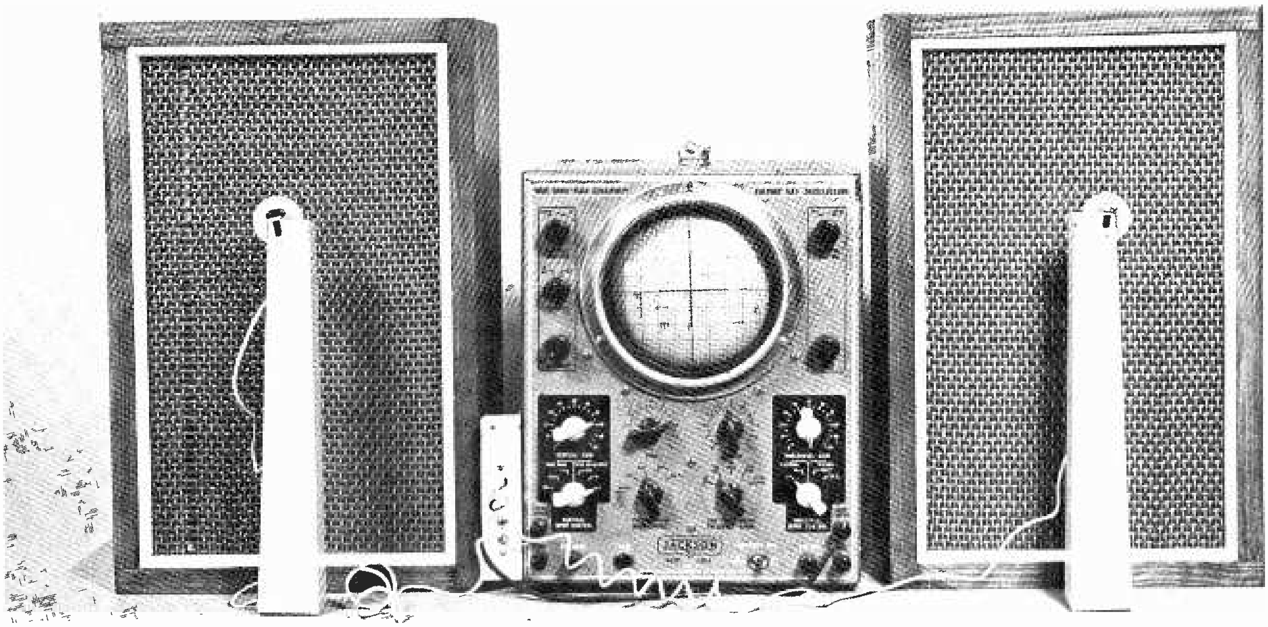


ARE YOUR SPEAKERS IN PHASE ?

Using a simple scope-microphone method to get proper connections.

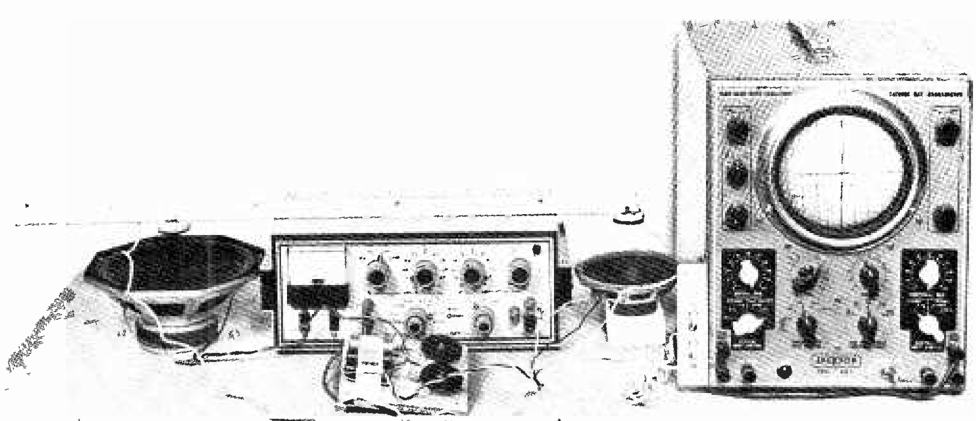
BY DAVID B. WEEMS

WE HEAR and read much about the necessity of properly phasing the speakers in a stereo system. The usual method prescribed is that of simply trying one hookup and then its reverse, settling on the speaker connection that produces the better bass response. This takes only a few seconds. Even so, many stereo speaker systems are hooked up with reversed polarity. Some "hi-fi" shops proudly demonstrate equipment that is operating out-of-phase.

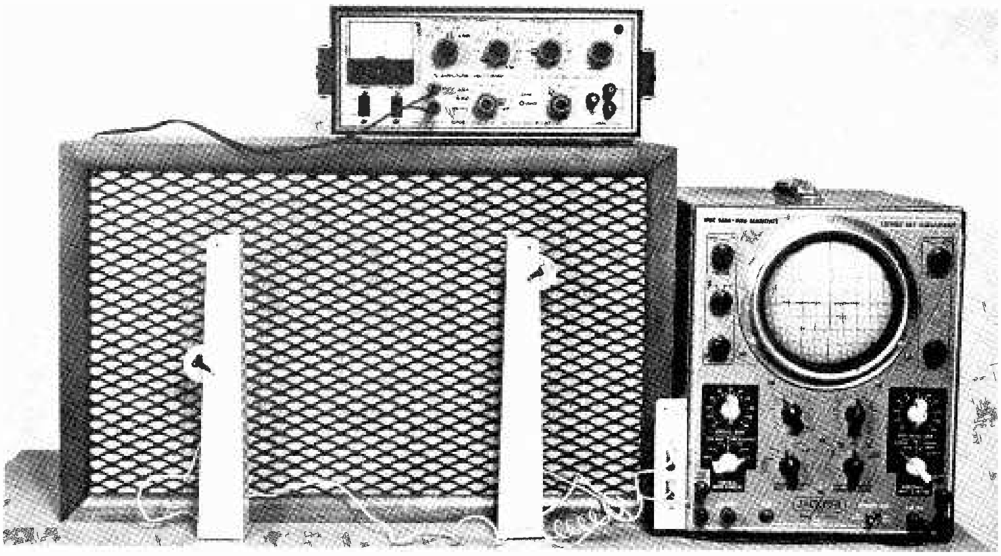
If you do not trust your ears, there is a visual test that leaves no doubt about phasing. It can be used as a final phase check on stereo speaker systems, but it is even more useful in determining the proper polarity of woofer and tweeter or woofer and midrange driver in a two- or three-way speaker system. (It should be noted that hookup errors in multi-unit speaker systems are common occurrences.)

Test Setup. Most loudspeakers have coded terminals—usually a red dot but sometimes a "+" sign stamped near, or a red fiber washer under the positive terminal. The difficulty with most faulty speaker systems involves the 12-dB/octave networks they employ. This type of network, containing two filter components in each leg (see Fig.1), is popular because it offers a compromise between the gentle frequency-response slope of the 6-dB/octave and the sharp cut-off 18-dB/octave types of networks. The 6-dB/octave network provides too little separation for some speaker systems, while the 18-dB/octave crossover can cause transient distortion.

In wiring together a 12-dB/octave network, the problem is that it pro-



Checking phase difference in crossover network. Audio generator is set at crossover frequency. Tweeter is raised above benchtop so that its cone is on plane with woofer. In photo above, two mikes are used to phase stereo speakers which are fed a low-frequency signal from stereo amp set for mono.



This test setup is used to determine phase difference between woofer and tweeters in completed speaker system.

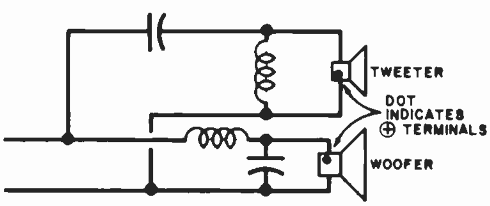


Fig. 1. Typical 12dB/octave 2-way crossover network. Phase difference between two legs of circuit is 180 degrees so speakers must be wired out-of-phase as shown.

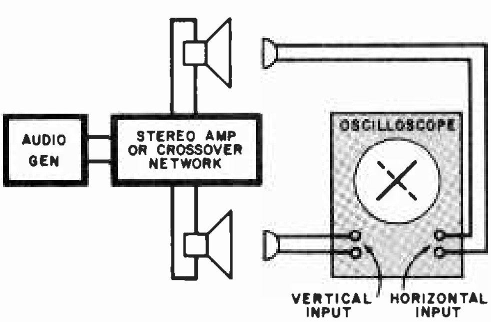


Fig. 2. Setup for phasing speakers using an oscilloscope.

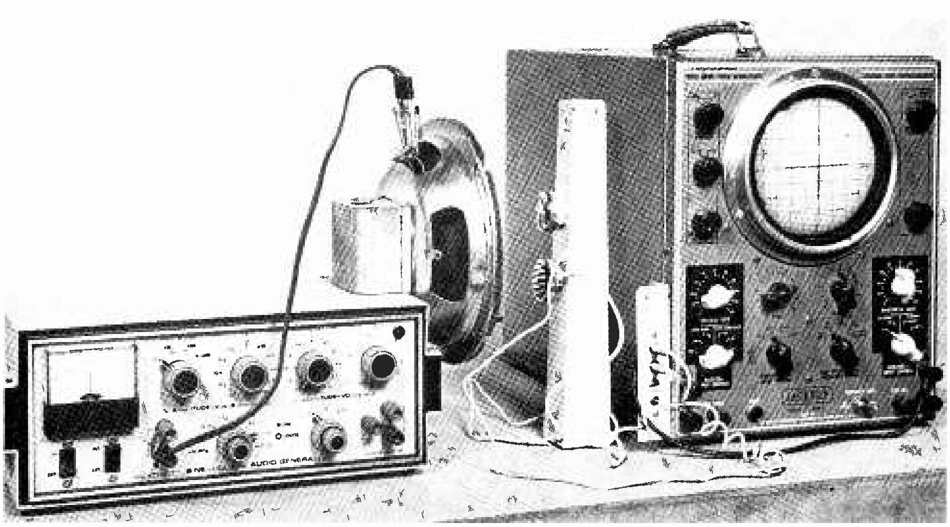
duces 180° phase differences between legs. Hence, to make the woofer and tweeter work in-phase, the two should be wired out-of-phase. In a three-way system, the midrange driver should be wired out-of-phase with both the tweeter and the woofer.

You can probably draw the schematic diagram of a crossover network that is not sealed in a can and figure out the proper connections for each driver simply by tracing out the circuit. However, if you have an oscilloscope, you can avoid this time-consuming step. In addition to the scope, you will also need two crystal microphones of the type used as "lapel mikes" with most inexpensive tape recorders. Connect one mike to the horizontal and the other to the vertical input of the scope.

Now, place the mikes at equal distances from a single-cone loudspeaker while you feed a low-frequency signal to the speaker. Adjust the scope's vertical and horizontal gain controls to obtain a line, inclined at 45° to the right or left of the vertical axis, on the CRT screen.

If the mikes are in-phase with each other and the scope is properly phased, the line will incline to the right (solid line in Fig. 2), while an out-of-phase condition will yield a line inclined to the left (dashed line). Whether the mikes are in-phase or out-of-phase is unimportant; what is important is the slope of the line when the mikes are placed before a single speaker. Make a note of the slope either on the graticule or on a card pasted to the side of the scope.

With the above described setup, you can test any pair of speakers for proper polarity, positioning one mike before each speaker and using, preferably, a low-frequency signal to drive the speakers. The mikes should be placed at equal distances from their respective speakers. Thus set up, the speakers are operating in-phase when the trace on the CRT screen is the same as that produced by a single speaker.



Put two inexpensive mikes equal distance from single speaker which is fed by low-frequency signal from generator to get phase reference curve on scope.

Multi-Driver Systems. When testing speakers in a two- or three-way system, set the test signal at the crossover frequency so that the woofer and tweeter receive equal power. In three-way systems, always test the phasing between the woofer and midrange driver (s). At high frequencies, the shorter wavelengths make positioning of the microphones just too critical to obtain reliable results. Just remember to make the polarity of the tweeter the same as that of the woofer.

If during your tests you obtain a scope pattern of a circle instead of a line, the speakers under test are out-of-phase by 90° or 270°, indicating a crossover network with either a 6-dB/octave or, less likely, an 18-dB/octave slope. In either case, the speakers should be wired in-phase. Theoretically, you should correct this situation by moving the plane of the tweeters at the crossover frequency one-quarter wavelength to the front or the rear of the plane of the woofer. Unless the crossover frequency is rather low, however, this is unnecessary because the difference in depth between the woofer and tweeter cones automatically injects a compensating factor by putting the tweeter's cone ahead of that of the woofer.

The oscilloscope test can quickly settle any arguments or indecision about the phasing of your speakers. And, if you are like most audio buffs, knowing that your speakers are wired correctly will also make them sound better. ◈

PICTURE TUBE TESTER AND REJUVENATOR

BY WILLIAM R. SHIPPEE

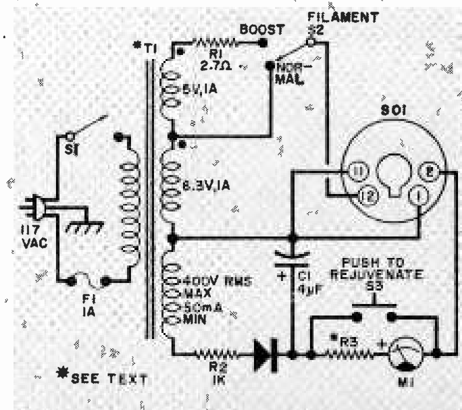
TV PICTURE TUBES are expensive —especially the ones for color. If you have one that has seen better days and is getting a bit dim and dark, you can probably give it new life by using the tester-rejuvenator described here. You will not have the equal of a new tube, but you may be able to keep the old one going for a while.

The circuit shown is used to test the emission of a cathode-ray tube; and if it is low, give the CRT a "shot" to revitalize it. The latter consists of raising the emission of the CRT by increasing the filament voltage. This "boils" the inner electrons of the cathode structure, bringing them to the outside where they can do the most good. The circuit will also remove some cathode/control-grid shorts. After using this circuit, a conventional picture-tube brightener can be used if the rejuvenated CRT does not exceed 50% emission as shown on the meter.

Construction. Any type of vacuum-tube transformer can be used for *T1* as long as the high-voltage winding does not exceed 400 volts rms. This winding should deliver at least 50 mA and the current ratings of the two filament windings should be at least one ampere. The filament windings must be properly phased so that approximately 11.3 volts appear across the series combination. If you don't get 11.3 volts, reverse the connections to one of the filament windings.

The meter should not have a full-scale reading over 5 mA. To calculate the value of *R3* for your particular meter, use Ohm's law to determine the resistance through which 5 mA will flow at the high-voltage dc obtained from the rectifier-filter circuit. As an alternative, you can start with a value of several kilohms and reduce it until the meter indicates exactly full-scale. Take care not to contact the high voltage when working with the resistor.

You can connect other sockets to *SO1* to suit your own particular type of tube, using pins 1 and 12 of *SO1* for each of the filaments of the tube and pin 2 switched to each of the control grids of the tube



PARTS LIST

C1—4-μF, 600-volt electrolytic capacitor
D1—1-A, 800-V silicon rectifier
F1—1-A slow-blow fuse and holder
M1—5-mA meter (see text)
R1—2.7-ohm, 5-watt resistor
R2—1000-ohm, 2-watt resistor
R3—½-watt resistor (see text)
S1—Spst, 1-A, 117-V switch
S2—Spdt, 1-A switch
S3—Spst pushbutton switch, normally open 600-V contact rating
T1—Power transformer; secondaries: 400-V at 50 mA, 6.3-V at 1A, 5-V at 1A
Misc.: Suitable chassis; line cord; CRT socket(s); high-voltage cables for sockets

Operation. With filament switch *S2* in the NORMAL position and pushbutton switch *S3* open, note the meter indication. To rejuvenate the CRT, momentarily depress *S3*. You may note a small arc in the neck of the tube. Release *S3* and note whether the meter indication has increased.

If the meter still indicates low, place *S2* in the BOOST position, wait a second for the filament to get hotter, and then depress *S3*. Return the filament switch to NORMAL and press the rejuvenate pushbutton a couple of times. The meter should show a marked increase. It may be necessary to repeat this operation several times, but do not leave the filament switch in the BOOST position for any length of time.

Remember that, on color tubes, there are usually three guns, so an adapter socket must be used with the circuit shown below. ◈

NOISE AND INTERFERENCE FILTER FOR SHORTWAVE RECEIVER

Construction of a simple, inexpensive active audio filter.

BY JOSEPH B. WICKLUND, JR.

HIGH-GRADE voice communication requires an audio frequency range (band) of from 330 Hz to about 3000 Hz. If some degradation of the voice tone can be tolerated, a narrower band of, say, 500 to 2000 Hz is possible. On the other hand, if a wider bandwidth is allowed, noise and adjacent signals can make the voice difficult to hear or understand. Many inexpensive and moderately priced communications receivers have insufficient selectivity—too wide a band—to provide optimum performance. To improve selectivity would generally require extensive modification, but a good audio filter can help considerably if it is used at the receiver's output.

With the availability of high-quality, low-cost operational amplifiers, active filters are generally the least expensive and easiest type to build for operation at low audio frequencies. Active filters differ from passive filters in that the former employ active elements, usually op-amps, to obtain the desired filter response. One very effective type of active filter is the "voltage-controlled voltage source" (VCVS). It is the construction of a VCVS filter that we will be discussing.

Theory of Operation. The schematic diagram of the VCVS active filter is shown in Fig. 1. A high-impedance buffer circuit made up of *IC1*, *R1*, and *C1* drives *IC2* and *IC3* which form a four-pole active filter. The filter was designed so that resistors *R2*, *R3*, *R5*, and *R6* are all of the same value. As long as all four resistors have the same value, changing that value changes the cut-off frequency but has no effect on the shape of the filter's response. The values of the resistors can be calculated from the equation: $R = (168{,}000)/F$, where F is the desired cut-off frequency in kilohertz.

To adequately control the shape of the filter's response, all frequency-controlling elements (*C1*-*C7*, *R2*, *R3*, *R5*, and *R6*) should have a 5% tolerance. With the values specified in the Parts List and given on the schematic diagram, the filter's cut-off frequency will be 3 kHz.

The filter can be operated from a single power supply with between 6 and 30 volts output if assembled as shown. Single power-supply operation requires the use of *R8* and *R9* to provide the necessary bias voltage to operate the op-amps (*IC1*-*IC3*) and V_B is not used. If two supplies (or a dual) are used, their outputs can range from a low of ±3 volts to a high of ±15 volts; in this case, the filter can be built without *R8*, *R9*, and *C9* and V_B is common.

Construction and Use. Owing to the fact that integrated circuits are used in the VCVS active filter, printed-circuit board assembly is recommended. (An actual-size etching and drilling guide

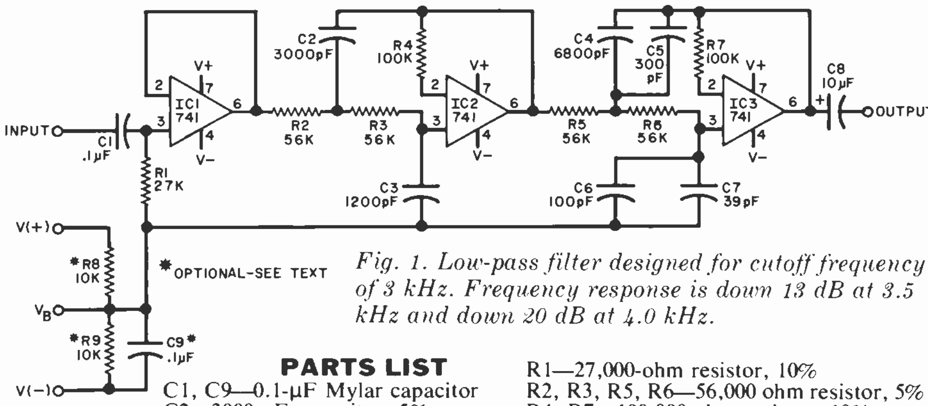


Fig. 1. Low-pass filter designed for cutoff frequency of 3 kHz. Frequency response is down 13 dB at 3.5 kHz and down 20 dB at 4.0 kHz.

PARTS LIST

C1, C9—0.1-µF Mylar capacitor
C2—3000-pF capacitor, 5%
C3—1200-pF capacitor, 5%
C4—6800-pF capacitor, 5%
C5—300-pF capacitor, 5%
C6—100-pF capacitor, 5%
C7—39-pF capacitor, 5%
C8—10-µF, 25-volt electrolytic capacitor
IC1, IC2, IC3—741 op-amp IC
R1—27,000-ohm resistor, 10%
R2, R3, R5, R6—56,000 ohm resistor, 5%
R4, R7—100,000-ohm resistor, 10%
R8, R9—10,000-ohm resistor, 10%
Misc.: Printed-circuit board or perf board; sockets; flea clips; solder; mounting hardware; etc.

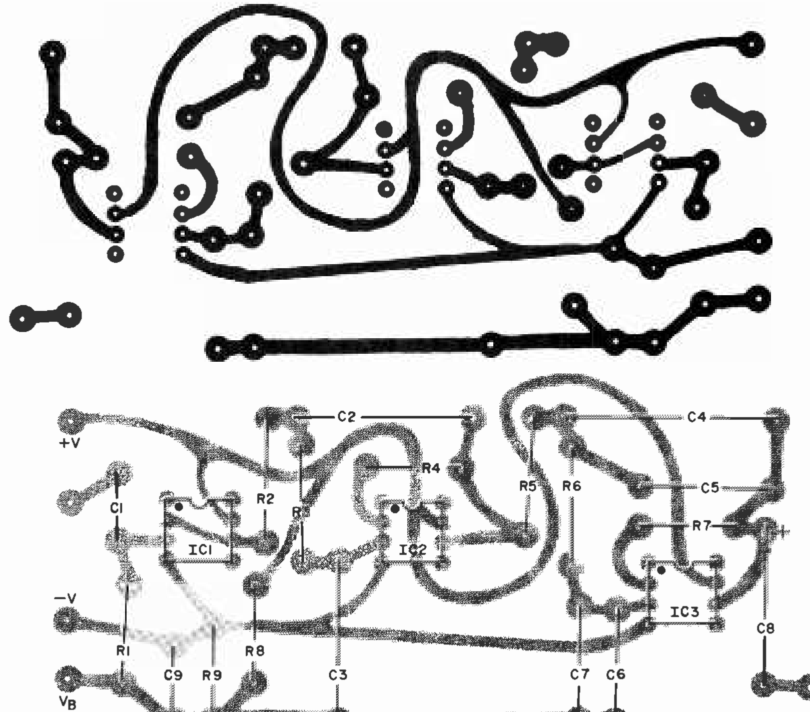


Fig. 2. Actual-size foil pattern (top) and component layout guide.

and a components installation diagram are given in Fig. 2.) However, if sockets or Molex "Soldercons" are used, the filter can readily be assembled on perforated phenolic board.

The filter can be used wherever voice-band filtering is required. For shortwave receivers, the headphone output can be used to drive the filter, with the filter driving a standard high-impedance headset. Alternatively, the filter can be permanently installed between the detector and the audio-output amplifier in the receiver.

Another possibility would be to use the filter to limit the frequency range of a microphone's output before modulation in a transmitter. This effectively concentrates more of the available output power into the critical 300-3000-Hz voice band. ◈

THE FIELD-EFFECT TRANSISTOR

What it is and how it has revolutionized electronics.

BY WILLIAM R. SHIPPEE

EVER since its introduction, the field-effect transistor has been creating quite a stir in electronics. Devices and systems heretofore impossible to produce with bipolar transistors had to be built around vacuum tubes—if at all. Now, the FET is changing the situation.

The FET has many of the qualities and advantages of both the vacuum-tube triode and the bipolar transistor. It is as compact as most small-signal transistors. It operates at low voltages, thus eliminating most of the bulk and expense of the power supply. Its input impedance can be rigged to fall into the desirable multi-megohm category. Recent developments have produced FET's which are capable of dissipating several watts of power: and since they exhibit the property of having a negative temperature coefficient, it is hard to make them succumb to thermal runaway.

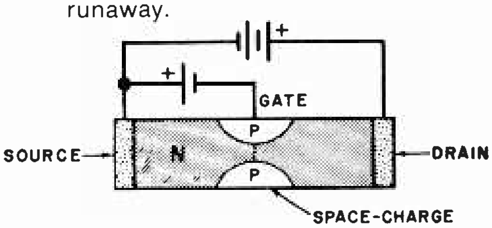


Fig. 1. FET acts as variable resistor in which the gate field has a direct effect on source-to-drain current flow.

Viewed as a design element, the FET is a semiconductor device which behaves in the manner of a variable resistor. As shown in Fig. 1, current flow between the source and drain is controlled by the gate voltage which is applied to both p sections simultaneously. As the reverse bias increases, the space charge area starts to pinch off, causing the source-to-drain current to fall almost to zero. Thus, the gate "field" has a direct "effect" on the source-to-drain current—hence the term "field-effect" transistor.

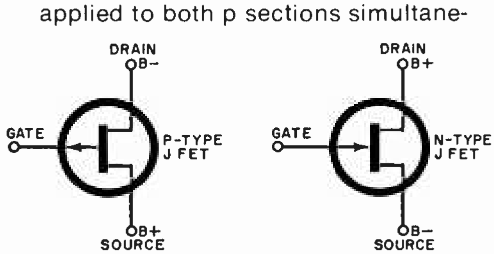


Fig. 2. Shown here are schematic symbols for a p-type (left) and an n-type junction field-effect transistor.

Types of FET's. There are basically two types of field-effect transistors in regular use today. The most common is the junction field-effect transistor, or JFET, which has a direct ohmic contact at the gate. The MOSFET, or metal-oxide field-effect transistor (sometimes known as an IGFET for insulated-gate field-effect transistor) has an electrically isolated gate.

In the JFET category, there are p- and n-channel types (see Fig. 2). The n-channel FET is very similar in voltage polarities and biasing to the vacuum-tube triode as shown in Fig. 3.

The MOSFET, a long-needed semiconductor device, even more closely approximates the input impedance of the typical vacuum tube. It can be fabricated to yield gate impedances well into the several hundred megohm region—beyond the usual capabilities of the common JFET. As shown in Fig. 4, there are two types of MOSFET's available. The one on the left is a single-gate type, while the one on the right has two gates.

The MOSFET's substrate is usually connected internally to the source; if not, the substrate is externally connected to the source or to ground. Great care must be exercised in handling the MOSFET since the gate input impedance is so high and the gate insulation so thin that any static charge introduced at the gate can perforate the oxide insulator barrier and destroy the device.

The dual-gate MOSFET finds its most popular application as the mixer stage in AM, FM, and TV tuners where it provides a convenient means of "beating" two frequencies in a nonlinear device while maintaining isolation between the two signals. Also the MOSFET appears to exhibit less noise and cross-modulation problems than do conventional transistors and vacuum tubes.

Virtually all MOSFET's produced for large current conditions are contained in single packages—not integrated circuits. The reason for this is that the FET needs roughly ten times the active area required by bipolar transistors to provide the same current capabilities.

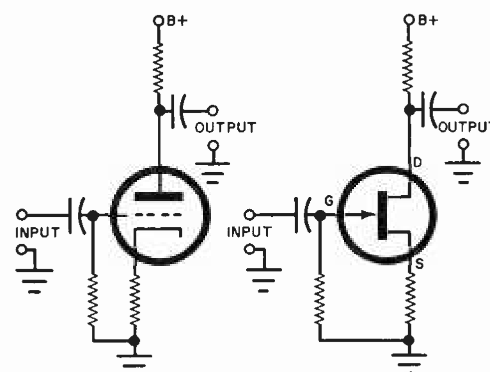


Fig. 3. Biasing arrangements for n-type JFET and vacuum-tube triode are same.

It is well to note that as r-f amplifiers, FET's are immune to strong-signal overloads. Some FET's are so symmetrically constructed that their drain and source leads are interchangeable.

The past few years have witnessed some remarkable developments in the semiconductor scene. It will be interesting to see which directions research and development will take in the future. ◈

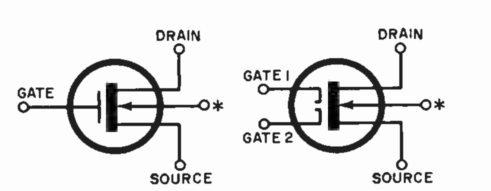
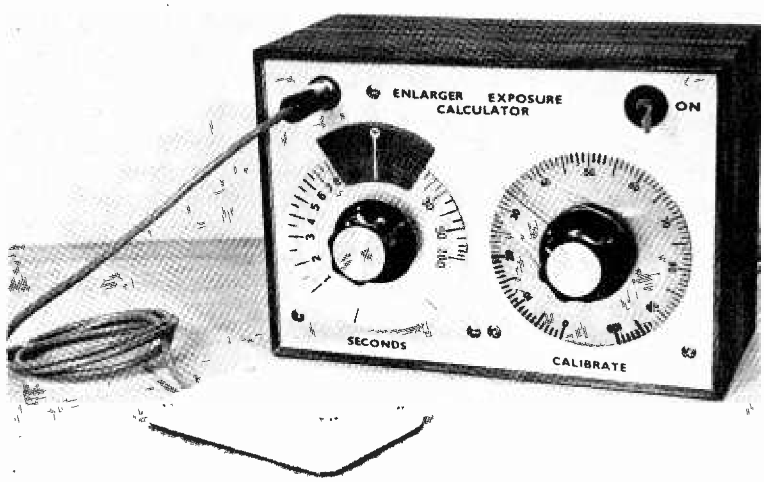


Fig. 4. Schematic symbols for single-gate and dual-gate MOSFET's are shown here.

ENLARGER EXPOSURE CALCULATOR



Direct indication of exposure time and image contrast.

BY ADOLPH A. MANGIERI

ANY well-equipped darkroom should include an enlarger exposure calculator. Its use can practically eliminate costly paper loss and greatly reduce darkroom time by eliminating test strips. Both exposure time and image contrast are directly indicated on a large, easily read, back-lit dial. A variable sensitivity control calibrates the calculator, accounting for projection paper speed and mode of operation.

Constructed at low cost and with performance exceeding that of available devices, this calculator uses a linear high-speed remote CdS light sensor in a dc comparator bridge. A very-high-gain op-amp, driving a LED fully on or off, sharply detects bridge balance with precise repeatability. Voltage changes have no effect on dial readout.

Circuit Operation. As shown in Fig. 1, potentiometer *R9* forms two arms of a variable ratio bridge. Linear high-speed sensor, *LDR1*, forms the third arm and *R8* and *R1* are the fourth. Potentiometer *R9* is calibrated in exposure time in seconds, while *R8* is used to calibrate the calculator. Operating open-loop, op-amp *IC1* senses bridge voltage through low-pass filter *C1R2*. As *R9* is rotated through bridge balance or null, *LED1* lights up when pin 2 of *IC1* goes negative and vice versa. Because *IC1* has very high gain, only slight movement of *R9* about the null point effects turn-on or turn-off of *LED1*, resulting in accurate and repeatable detection of the trip point.

Markings on the dial of *R9* are similar to those of a comparator bridge. Since *LDR1* is highly linear over the light range of interest, the dial of *R9* is calibrated using resistors substituting for *LDR1* and forming known ratios with *R1*. A split zener supply provides the op-amp supply voltage. Potentiometer *R10* adjusts the op-amp bias current, although the op-amp input offset voltage (a few millivolts) has negligible effect on the bridge balance point. A three-wire line eliminates stray fields which would otherwise degrade sharpness of the trip point, depending on the direction of ac plug insertion.

Construction. All components except *T1* are assembled on the panel of a 3" × 4½" × 6½" metal case. Begin by cutting a 2¾" diameter disc from 1/16-inch clear plastic sheet. Ream the center hole to ¼ inch and mount the disc on a drill or arbor to true up the edge. Cement the disc to a knob using a ¼-inch rod to insure alignment. Install *R9* and the disc dial on the panel and mark the window cutout. Make the cutout ¾ inch high and arced over 80 degrees of the dial face. Temporarily install the circuit board on spacers and locate the hole position for a rubber grommet which supports lamp *I1* on the board. Use shoulder washers to insulate jack *J1* from the panel.

Assemble the circuit board as shown in Fig. 2, using flea clips and point-to-point wiring. Use a socket for *IC1*. Capacitors *C1* and *C2* are installed below the board at the IC socket. Defer installation of *R6* and *R7* if you have substituted for transformer *T1*. Potentiometer *R9* may have any value from 10,000 to 50,000 ohms. Make the window pane a bit larger than the window cutout, using translucent red plastic. Mark an opaque vertical center index line on the pane and drill a 0.071-inch hole near the upper end. Using clear epoxy, cement the LED in the hole on the bottom side of the pane. Solder a length of twisted phono wire, less shield, to LED leads with the red wire to LED anode.

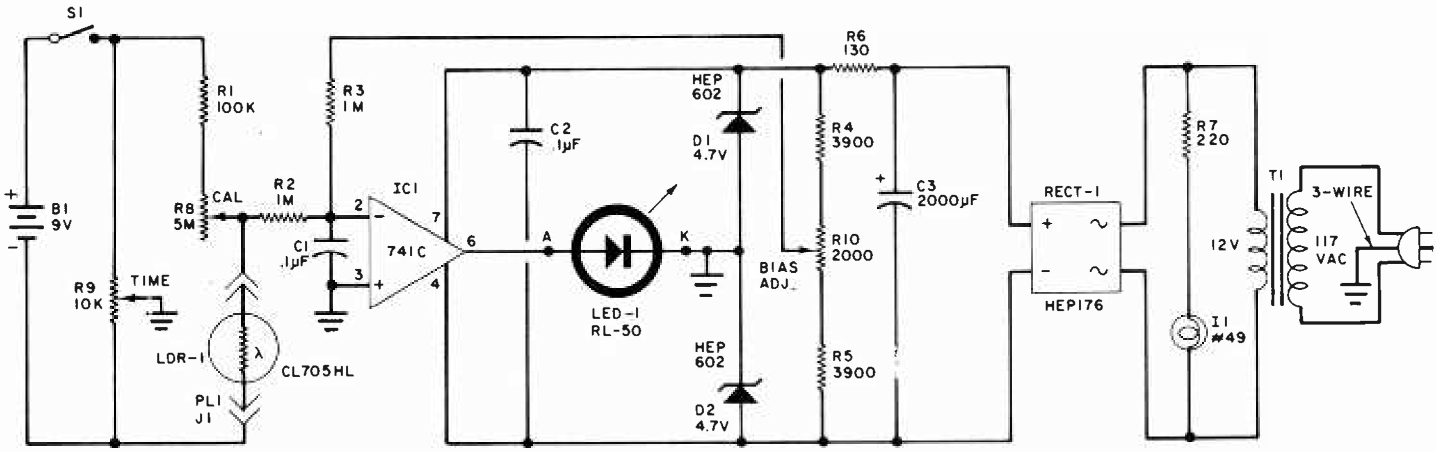
Connect the right-hand lug of *R8*, as viewed from the rear, with terminals down, to resistor *R1*. Connect the right-hand lug of *R9*, similarly viewed, to the other end of *R1*. Install *T1* and a lug strip for the line cord at one end of the case. You may add a separate spst switch ahead of *T1* in one of the ac line wires.

Connect both the circuit ground and line ground to the metal case. Install a battery compartment made of sheet aluminum.

Mount the photo sensor (*LDR1*) between two pieces of thin plastic or phenolic sheet so that the light-sensitive face of *LDR1* fits through a suitable hole in the upper sheet. Paint the top piece white. Use a length of miniature shielded and jacketed mike cable for the connecting cable. Insulate the connections to *LDR1* with bits of vinyl tape and run the center wire to *R2* through *PL1* and *J1*.

Checkout and Calibration. Connect *R6* to the circuit using clip leads with a dc milliammeter in series. Unplug *IC1* and insert the ac line cord. If necessary, alter the value of *R6* to obtain current of about 70 mA. Measure the ac voltage across *I1* and alter the value of *R7*, if required, to obtain a voltage of, preferably, 1.5 volts but not in excess of 2 volts.

Install *IC1* with power off, observing proper direction of installation. With the photo sensor unplugged, *S1* off, and *R8* set to maximum resistance, adjust *R10* very slowly to the setting where *LED1* initially lights up. If this cannot be done, increase *R4* or *R5*, as required, by about 1000 ohms to bring *R10* within adjustment range.



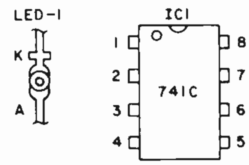


Fig. 1. When preselected amount of light unbalances the bridge (via LDR1), the op-amp switches causing LED1 to come on.

PARTS LIST

B1—9-volt transistor battery
C1,C2—0.1-µF, 25-volt disc capacitor
C3—2000-µF, 25-volt electrolytic capacitor
D1,D2—4.7-volt, 1-watt zener diode (HEP 602 or equiv.)
I1—#49 pilot light
IC1—741C op amp
J1—Phone jack
LDR1—CdS photo sensor (Clairex CL705HL. Do not substitute)
LED1—2-volt, 20 mA light-emitting diode (Litronix RL-50 or equiv.)
PL1—Phone plug
R1—100,000-ohm, ½-watt resistor, 2%
R2,R3—1-megohm, ½-watt resistor, 10%
R4,R5—3900-ohm, ½-watt resistor, 10%
R6—130-ohm, 2-watt resistor (see text)
R7—220-ohm, 2-watt resistor (see text)
R8—5-megohm audio-taper #1 potentiometer (Mallory U65 or equiv.)
R9—10,000-ohm, 5-watt wirewound potentiometer
R10—2000-ohm trimmer (Centralab V-2000 or equiv.)
RECT1—1-A, 200-PIV bridge rectifier (HEP-176 or equiv.)
S1—Spst switch
T1—12-V, 0.3-A filament transformer (Radio Shack 273-1385 or equiv.)
Misc.: Perf board; flea clips; IC socket; suitable enclosure (Vector W30-66-46 or equiv.); 3-wire ac line cord; clear and translucent-red plastic sheet; dial plate; knobs; battery clip; mike cable; calibrating resistors; hardware; etc.

The dial of *R9* is calibrated by plugging in resistors having a known ratio with bridge arm *R1*. Use either fixed resistors or a potentiometer, set to required values, connected to a phone plug with short leads. Ground the instrument to eliminate stray pickup. Jumper the two wired lugs of *R8*, insuring zero resistance. Plug 10,000 ohms into *J1* and turn *S1* on. Adjust *R9* to the point where *LED1* first lights up. Mark the dial above the index line as the 1-second index.

Similarly, use 20,000 ohms for the 2-second index, 30,000 ohms for the 3-second index, and so forth, up to 200,000 ohms for the 20-second index. Spot additional indices at 5-second intervals to 50 seconds (500,000 ohms) and at 10-second intervals to 100 seconds (1 megohm). Half-second indices between 1 and 10 seconds may be linearly interpolated or spotted, using suitable resistances. Remove the dial and label it, using dry transfers. Plug 100,000 ohms into *J1* and replace the dial for *LED* turn-on at 10 seconds. Remove the jumper from *R8*.

Application. If darkroom receptacles lack a ground wire, determine direction of plug insertion as follows and color-code both plug and receptacle. Set *R8* to maximum resistance and turn *S1* on. At room light level, invert or cover the sensor with sheets of paper so that dial readout is at least 10 seconds. Try the line cord both ways. Use that direction which produces a sharp and sudden turn-on of *LED1* as *R9* is slowly rotated. Turn off all darkroom light when using the calculator. At very low light levels, check for possible contribution of light from the dial itself by placing a blackened cardboard tube from a 35-mm film carton over the sensor.

Correct settings of calibration pot *R8* for projection papers in use are obtained by means of a test print. Using a negative of average contrast and content, make the best possible test print conventionally. Let us assume that the best print was exposed for 10 seconds at *f*/8 lens aperture.

For the integrated light method, use a 3-inch square ground-glass plate as a light scatterer. With enlarger settings undisturbed, place the sensor at the center of the projected image. Turn *S1* on and set *R9* to 10 seconds. Hold the light scatterer up to the lens and adjust *R8* slowly until *LED1* first turns on. Rock *R9* slightly to verify turn-on at 10 seconds and trim *R8* if required. Record *R8* setting and paper data. To use the calculator, and at any print magnification and lens aperture, set *R8* as previously recorded, use the light scatterer, and rotate *R9* to the point of *LED1* turn-on. Expose for the indicated time at the selected aperture.

If you cannot calibrate *R8* (advanced to its limit) open up the lens

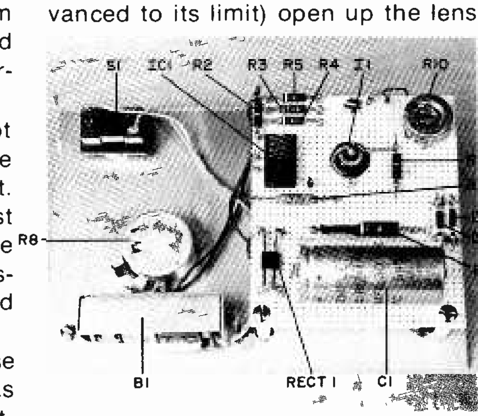


Fig. 2. Circuit can be built on perf board as shown. Lamp I1 illuminates "seconds" dial through hole in front panel. LED1 is built into window panel, as shown in the lead photograph.

one (or even two) full stops and try again for 10 seconds readout on *R9*. In using such calibration settings, you would naturally close down one (or two) full stops from any selected measuring aperture to compensate before exposing. Or, you may use one-half (or one-fourth) of the indicated exposure and expose at the measuring aperture. Record the mode you have selected along with *R8* settings and paper type.

Next, the spot method takes a single measurement at the brightest portion of the projected image (print shadows). To calibrate *R8* for this mode, place the sensor at print shadows and *R9* at 10 seconds. Adjust *R8* for 10 seconds' readout on *R9* with lens aperture set to that of the test print. In use, merely set *R8* as recorded, place the sensor at print shadows, select any lens aperture, and turn *R9* to the point of lamp turn-on for the required exposure time.

Negative contrast is related to the brightness range of the projected image. To measure contrast, place the sensor at the brightest portion of the image. Set *R9* to 1 second. Adjust lens aperture and/or *R8* for indicator turn-on at 1 second. Rock *R9* to check. Then move the sensor to the darkest portion of the image and adjust *R9* to a second, and always higher, indication. If the second reading is 12 seconds, the contrast is 12:1 or merely 12. Contrasts of 8 to 15 print well on normal contrast paper but you should make up your own grading system since contrast also depends on print development time.

You can readily conceive of other modes of operation and applications. By installing the sensor in a probe or handle, you can take measurements from printboxes and ground-glass viewing screens. Depending on your needs, you can interpret the 10-second index as 0.1, 1, or 100 seconds. ◈

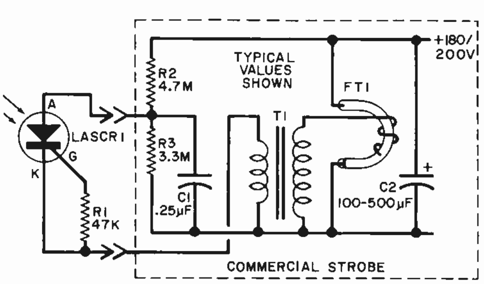
LIGHT-ACTIVATED SLAVE STROBE TRIGGER

BY ADOLPH A. MANGIERI

MANY photo enthusiasts have several strobe lights, which they often want to operate at the same time. The trigger circuit described here uses a light-activated SCR to trip a slave strobe when a master strobe is fired.

The trigger circuit, consisting of the light-activated SCR (*LASCR1*) and resistor *R1*, is shown in the schematic the way it should be connected to the standard commercial slave strobe. When used by itself with a camera, the commercial strobe is fired when an external switch (usually on the camera) is closed to discharge *C1* through the primary of *T1* and apply a surge of voltage to the flash tube.

With the slave strobe trigger, *LASCR1* acts as the switch since it starts to conduct when the light from a master strobe strikes it. (Under normal light conditions, *LASCR1* is turned off.) When *C1* and *C2* are discharged by the firing of the flash tube, *LASCR1* returns to the off state. Resistor *R1* bypasses the slight internal leakage, which may otherwise cause self turn-on of *LASCR1*.



PARTS LIST

LASCR1—1-A, 300-PIV light-activated SCR*
R1—47,000-ohm, ½-watt resistor (see text)
Misc.: Perf board; opaque container; two-conductor cable with suitable connector

*Available from Poly Paks as 87U666 ($1.75); 200-PIV units may be substituted if desired: Available from Delta Electronics, P.O. Box 1, Lynn Mass. 01903 (P/N P4118, $2.00); Poly Paks (87U666 200-PIV, $1.49); or Solid State Systems ($1.75).

Fig. 1. LASCR replaces switch in strobe.

Construction. As shown in Fig. 2, *LASCR1* and *R1* are mounted on a small piece of perf board with the sensitive end of the light-activated SCR at the end of the board. Do not solder *R1* permanently into place at this time. The case is made of any opaque tubing, such as a pill container painted black, with a small hole in the closed end to accept the face of *LASCR1*. Its sensitive surface should be about ⅜" from the end of the container. A short length of two-conductor cable, terminated with a suitable connector, is used to connect the trigger to the strobe.

Using a VTVM, check the voltage level and polarity at the strobe PC cord plug or socket. This may range up to 200 volts.

Checkout. Connect the trigger to the slave strobe. Aim the slave trigger at the main strobe, turn the main strobe on and depress the test button. If the slave strobe does not fire, use the VTVM to measure the voltage across *LASCR1* anode and cathode. If this voltage is about 1 volt, *LASCR1* is already on. Replace *R1* with a smaller ohmic value until the VTVM indicates the previously measured PC cord voltage with *LASCR1* off. Make *R1* as high a value as possible for maximum sensitivity of the light-activated SCR.

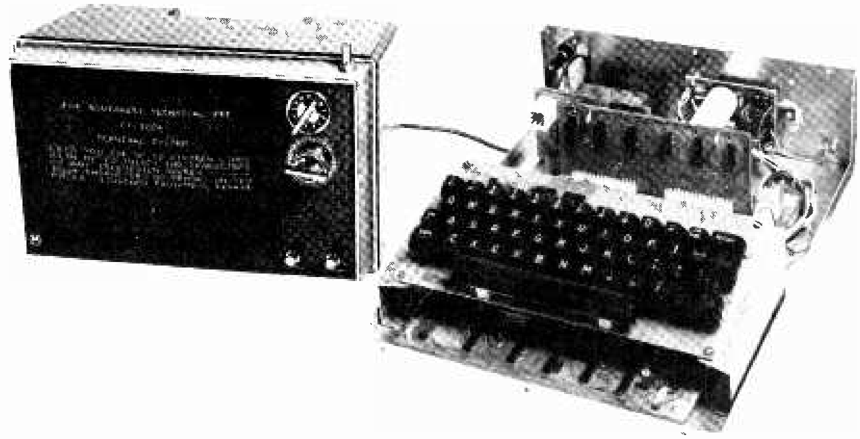
If *LASCR1* cannot be made to fire, the strobe may have unusually high resistance values as *R2* and *R3*. (Typical values are shown in Fig. 1.) In such cases, a slight leakage current may be pulling down the triggering voltage across *C1*. This can be checked by measuring the PC cord voltage with *LASCR1* connected and disconnected.

Always test-fire the master-slave strobe combination a few times before actual use, making sure that you aim *LASCR1* toward the main strobe. For use in a slightly high light level, a neutral density filter may be placed in front of *LASCR1* so that ambient light will not cause it to operate, but the much brighter flash from the main strobe will cause operation. ◈

Fig. 2. Photo of the prototype assembly.

SWTPC

CT-1024 TERMINAL SYSTEM



When we designed the CT-1024 we knew that there were many applications for an inexpensive TV display terminal system. Even so, we have been surprised at the many additional uses that have been suggested by our customer in the last four months since we introduced this kit.

The basic kit, consisting of the character generator, sync and timing circuits, cursor and 1024 byte memory gives you everything you need to put a sixteen line message on the screen of any TV monitor, or standard set with a video input jack added to it. Input information to the CT-1024 may be any ASCII coded source having TTL logic levels. Two pages of memory for a total of up to one thousand and twenty four characters may be stored at a time. The CT-1024 automatically switches from page one to page two and back when you reach the bottom of the screen. A manual page selector switch is also provided. The main board is 9½ x 12 inches. It has space provided to allow up to four accessory circuits to be plugged in. If you want a display for advertising, a teaching aid, or a communication system then our basic kit and a suitable power supply is all you will need.

CT-1 TERMINAL SYSTEM with MEMORY KIT........................$175.00 ppd

Power supply kit to provide + 5 Volts @ 2.0 Amps and - 5 Volts, -12 Volts @ 100 Ma. required by the CT-1 basic display system.

CT-P POWER SUPPLY KIT........$15.50 ppd

A very nice convenience feature at a very reasonable cost is our manual cursor control plug-in circuit. The basic kit allows you to erase a frame and to bring the curscr to the upper left corner (home up). By adding this plug-in, you can get Up, Down, Left, Right, Erase to End of Line and Erase to End of Frame functions. These may be operated by pushbutton switches, or uncommitted keyswitches on your keyboard. Although not essential to terminal operation, these features can be very helpful in some applications.

CT-M MANUAL CURSOR CONTROL KIT..$11.50 ppd

If you plan to use your terminal with a telephone line modem, or any other system that requries a serial data output; you will need our serial interface (UART) plug-in circuit. This circuit converts the ASCII code from a parallel to a serial form and adds "Start" and "Stop" bits to each character. The standard transmission rate for this circuit is 110 Baud, but optional rates of 150, 300, 600 and 1200 Baud may be obtained by adding additional parts to the board. The output of this circuit is an RS-232 type interface and may be used to drive any type modem, or coupler system using this standard interface.

CT-S SERIAL INTERFACE (UART) KIT..$39.95 ppd

If you are using the CT-1024 as an IO (input - output) device on your own computer system, you will probably want to connect it to the computer with a parallel interface system. A direct parallel interface allows for much faster data transmission and reception and is basically a simpler device than a serial interface system. Our parallel interface circuit contains the necessary tristate buffers to drive either a separate transmitt and receive bus system, or a bidirectional data bus system. TTL logic levels are standard on this interface. Switch selection of either full, or half duplex operation is provided. The terminal may write directly to the screen, or the computer may "echo" the message and write to the screen.

CT-L PARALLEL INTERFACE KIT..$22.95 ppd

We would be happy to send you a complete data package describing the CT-1024 and a achematic. If you want this additional information, circle our number shown below on your reader information service card. The CT-1024 kit has complete assembly instructions with parts location diagrams and step-by-step wiring instructions. If you would like to check the instruction manual before you purchase the kit, please return the coupon with $1.00 and we will rush you the manual and the additional data mentioned above.

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MAKE YOUR DIGITAL CLOCK "FAIL SAFE"

BY CALVIN DILLER

A MAJOR problem encountered with electronic digital clocks which operate from the ac power line is that they become useless if they are disconnected from the power line for any reason, even if the disconnect lasts for only a small portion of a second. Once this happens, and when the power is restored, the readouts will indicate the wrong time and the clock must be reset.

What is needed is a power source that will automatically take over if there is a power-line outage and some type of "clock" signal that closely approximates the 60-Hz timing signal normally derived from the power line. This should also come on automatically (preferably without losing a second) when the power goes off.

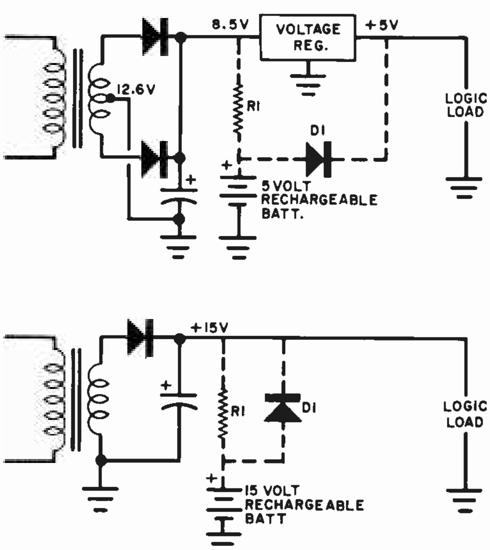


Fig. 1. Typical digital clock power supplies. Modifications are shown in the dotted lines.

Automatic Power Supply. Two possible dc power supplies are shown in Fig. 1. In normal operation, the supply delivers operating power to the clock logic and readouts, while *R1* supplies a small charging current to the rechargeable battery. Diode *D1* is reverse-biased in this mode so that it doesn't "see" the load. When the ac power fails, *D1* conducts and permits the battery to carry the load. The battery selected should have the correct voltage and capacity to drive the load for as long as desired. To determine the battery current required, operate the clock from the ac power and insert a dc current meter between the regulator and the load and record the clock current requirements. Add a little to the value to be on the safe side. Diode *D1* is any silicon rectifier, while the value of *R1* is selected to provide the trickle-charge current required by the battery. This value should be listed in the battery specifications.

Automatic Timer. The circuit shown in Fig. 2 uses a 555 timer and is designed to oscillate at 60 Hz with the values shown. With the addition of the two components to pin 5, the circuit will automatically synchronize with the 60-Hz line frequency. If the power fails, the 555 will continue to oscillate at 60 Hz as long as it is supplied with dc from the rechargeable battery. The system can also be changed to operate at 50 Hz. As soon as the ac power is restored, the 555 immediately jumps into exact line synchronism.

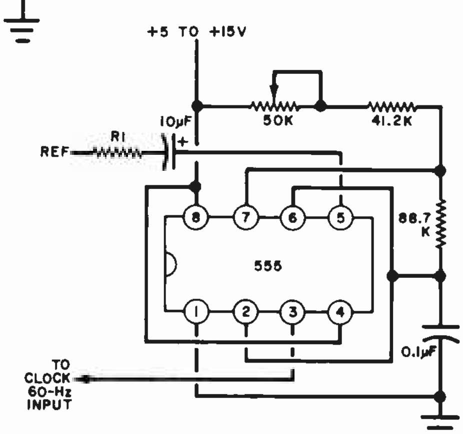


Fig. 2. This oscillator operates at 60 Hz when battery powered and it automatically synchronizes to the 60-Hz power line when on.

The line reference signal required should be sinusoidal and is usually obtained from the clock's power transformer. It is referenced to ground. Measure the ac rms voltage available, and use Fig. 3 to determine the value of *R1*.

The auxiliary timer can be set to exactly 60 Hz by connecting a frequency counter to pin 3 and adjusting the potentiometer. If you do not have access to a frequency counter, try the phase detector scheme shown in Fig. 4, using silicon diodes for the bridge and any type of 15-volt dc meter as the indicator.

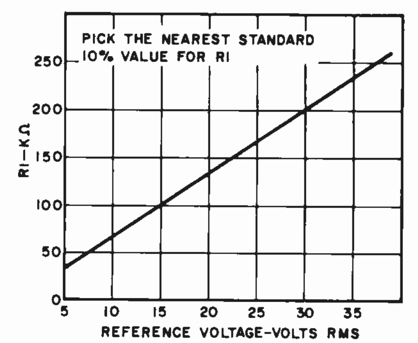


Fig. 3. Use this chart to determine value of R1 (Fig. 2) when 60-Hz reference voltage level in rms is known.

With the rechargeable battery and auxiliary timer connected to the digital clock, disconnect the clock from the ac line. The readouts should still indicate the time (showing that the rechargeable battery circuit is working), but the actual time may start running fast or slow (showing that the auxiliary

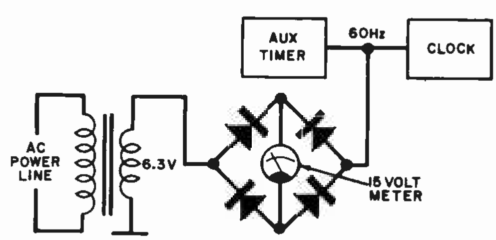


Fig. 4. Simple phase detector is used to set auxiliary timer to exactly 60 Hz.

timer is working but is not exact). With the phase detector circuit connected as shown in Fig. 4, note that the voltmeter needle has a quivering motion, indicating that there is a phase (frequency) difference between the auxiliary timer and the power-line reference. Adjust the auxiliary timer potentiometer until the motion of the needle slows down and comes to a stop. The auxiliary timer is now operating at the power-line frequency. The meter needle may drift about a little indicating a slow change in the free-running frequency of the timer.

Reconnect the digital clock to the ac power line and note that the meter needle deflects to some point on the scale and remains there, indicating that the auxiliary timer has locked in-phase with the power line. ◈

FLAGPOLE HAM ANTENNA

Vertical helix disguised as a flagpole produces good results.

BY ROLAND J. McMAHAN

WHEN we moved to our new house, my usually gentle and permissive wife issued an unusual ultimatum: "Please do not put up a spider web of antennas around our new home." For a moment, my world reeled; 40 years of spider webs—uh, amateur radio—had taught me that a good antenna is the first requirement of a good station.

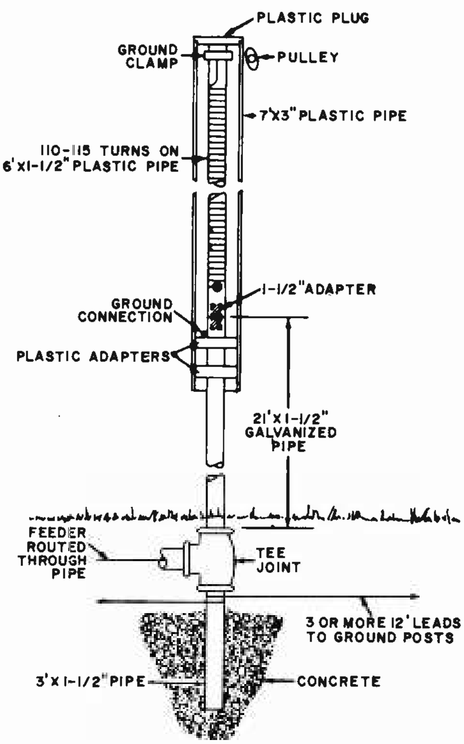
I ruminated for a while, rejecting the idea of an antenna camouflaged to look like a lightning rod in a location where lightning rods are a rarity. A flagpole? We're patriotic enough to have had one at our last location, so I was sure my wife wouldn't object. Now, I became obsessed with the idea of designing a flagpole antenna.

Trial and Error. The various handbooks cover vertical antennas very well—except for the helix, which is probably the easiest to build, consisting of a half-wavelength of wire wound on a form. Convinced that this was the design I needed, I wound 70 feet of #12 insulated stranded wire on a 6-foot length of 1½-inch inner diameter rigid plastic pipe. The turns stayed in place; so I drilled a hole at each end of the pipe and passed the wires through them to secure the coil.

The turns stayed in place well without cement—at least until the antenna-ground system was resonating on the 40-meter band, at which time about 20 turns suddenly dropped a foot and completely demolished my work up to that time. Too late, I realized that I should have taped the turns at about 12-inch intervals.

I checked the resonant frequency of the antenna-ground system with a dip meter and verified the frequency with my calibrated receiver. (The depth of the dip gives an indication of the Q of the system. The resonant frequency will change by 0.2-0.4 MHz when the antenna is raised to its perch atop a flagpole; it did with mine.)

My first effort, running the antenna to a loop around my dip oscillator coil and then to the ground I intended to use, gave a high-Q dip at about 6 MHz. Removing turns of wire sometimes raised and at other times seemed to lower the resonant frequency. Adding a ground-type clamp as a top-loading capacitance connected to the antenna at the top seemed to make the antenna behave better.



I removed turns until the wire on the form was only 60 feet. A high-Q dip appeared at 6.4 MHz—still far too low. I decided to try the antenna indoors and far off resonance because I wasn't sure yet how much difference raising the antenna would make. So, with the antenna and ground connected to my transceiver, I had to increase the loading adjustment and got the plate current to increase from 100 mA to 200 mA. The point of maximum field strength and minimum plate current occurred at the same plate tank tuning setting—an indication of low SWR.

With the antenna sitting on a table and leaning against a wall, I received an S7 from Seattle, some 300 miles away. On transmit, power to the antenna system was about 80 watts. I removed a few more turns until the oscillator indicated a dip at 7.1 MHz (this was later verified). Now minimum loading caused a broad tuning, flat plate current of slightly more than 200 mA. I tuned for maximum field-strength indication and received some S8 reports.

Finalizing the Design. Now I decided to try my antenna outside, feeding it with a 50- or 75-ohm line. Fitting a plastic adapter into the lower end of the plastic pipe form, I screwed the antenna into an iron coupling on a 3-foot iron pipe and planted the pipe in the ground. One lead of a length of flat ac power cord ran to the antenna, and the other I clamped to the pipe and to a wire about 30 feet long which terminated in a pipe in the ground.

Now the frequency was 7.4 MHz; too high. However, I tried the system out and received an S7 from a ham aboard a ship some 1000 miles away. Adding some turns and mounting the antenna on a 21-foot length of galvanized iron pipe, with the feedline running down the center of the pipe, I raised the antenna and leaned the pipe against the roof of the house to minimize capacitance problems. The antenna checked out 0.2 MHz higher than it was when on the ground.

That antenna went up and came down so many times that our neighbors must have thought I was signaling with my flagpole. Eventually, I got the antenna to resonate at 7.1 MHz. Readings of S9 became common, and I got an S6 from one of the Japanese islands on 40 meters. Using the third harmonic and with no changes to the antenna, I logged an S9 from Germany—no mean feat in Idaho. ◈

UNDERSTANDING UNGROUNDED OSCILLOSCOPE MEASUREMENTS

Making scope measurements across ungrounded components can present some problems. Here are the reasons— and some answers.

BY RAYMOND E. HERZOG

WHEN we measure a voltage in a circuit, we don't always take into account what we are actually measuring. For instance, we might say that a power supply's output is 50 V dc with 0.75 V ac ripple; or the output signal at a transistor amplifier's collector is 5 V ac. In these, and just about all voltage measurements, what we really mean is that the power supply's output is 50 V dc with respect to ground (or chassis common); the ripple is 0.75 V ac with respect to ground; and the amplifier output is 5 V ac with respect to ground. Thus, what we are really measuring is the voltage at a given point—with respect to a common point.

Since a voltage is the potential difference between two points, the two points have to be identified. For convenience, we generally use chassis ground as the second point.

But what if we want to measure the voltage across a component both sides of which are above ground? This presents a problem—many problems, in fact. Obviously, one difficulty is the lack of a convenient, easy-to-get-at chassis for a connecting point. More important, however, are the possible bad effects of connecting both leads of a test instrument to ungrounded points.

Of course, occasions such as this do not occur often; but when they do, knowing the proper procedure can make the job easier and prevent undesirable effects such as overloaded circuits and noise pickup when making measurements.

Not a Simple Test. Measuring a voltage between two ungrounded points is not always a simple matter. Assuming that an oscilloscope is being used, one does not merely connect the test probe and ground lead at opposite ends of the ungrounded component—certain precautions must be observed. Consider the following examples.

Assume that we have a conventional scope which has a three-wire power cord. For safety, the scope chassis is tied to the third wire and ground. Because the signal input "ground" terminal is also common with the chassis ground, it too is tied to the third wire in the power cord. Now, let's say we're testing an ac/dc radio or a transformerless TV receiver; the chassis being tested is tied to the low side (ground) of the power cord. So we have the situation shown in Fig. 1. The chassis are tied together through the powerline system.

As long as the scope's ground test lead is connected to the tested chassis (point Z in Fig. 1), the chassis are tied together, the grounds are tied, and we have a good safe test setup. Notice, however, what happens when the scope's ground test lead is connected to a point above ground potential (point Y in Fig. 1). The portion of the circuit between Y and Z is effectively shorted out by the ground circuit through the two chassis and the power-line ground. This, of course, could severely disturb the circuit op-

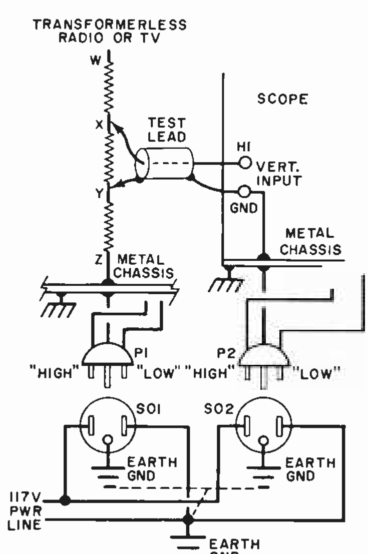


Fig. 1. In transformerless circuit, scope ground return on point Y can create a short across Y-Z element.

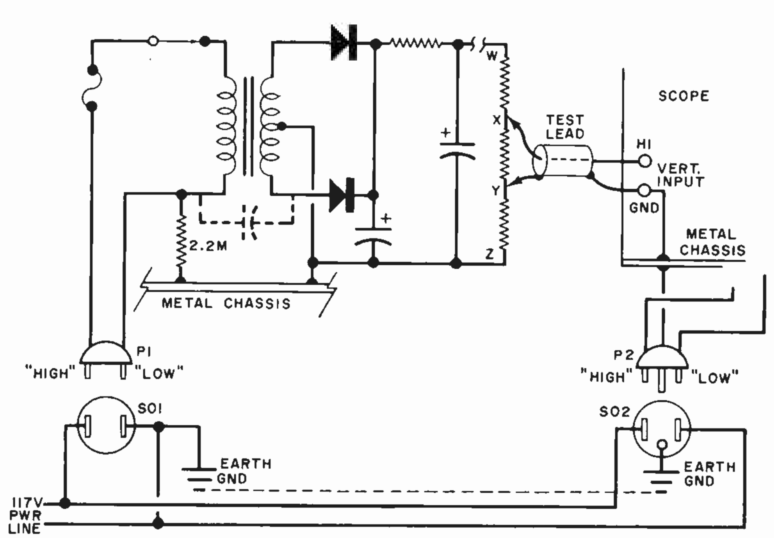


Fig. 2. With power transformer in circuit, connecting scope ground to point Y can put an ac shunt across X-Z.

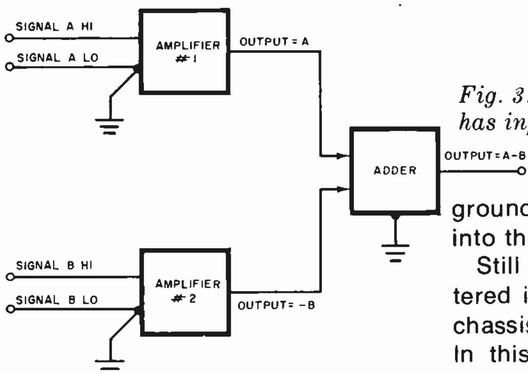


Fig. 3. Simple differential amplifier has inputs A and B and output A - B.

eration and possibly damage the components in the network. The same ground problem could also occur with a scope that has two-wire power cord if the low side is tied to chassis ground.

Now assume that we have a scope with a three-wire power cord and we're testing a TV receiver with a power transformer and a conventional two-wire ac connection. As shown in Fig. 2, the low side of the ac line is connected to the chassis through a large resistance (commonly 2.2 megohms). Of course, not all equipment has this resistance—some are entirely isolated—but it is important to know whether it is there or not. With the circuit shown in Fig. 2, there is an ac shunt effectively placed across part of the circuit under test.

The ac path inside the receiver is from point Z to the chassis, to the transformer secondary, through stray primary-to-secondary capacitance, to the transformer primary, to the ac line. This ac shunt can cause problems, especially with high-frequency measurements. (For all of the above situations, tying the ac line to an ungrounded point can introduce noise into the circuit.)

Still another problem is encountered if the scope does not have its chassis tied to the power-line ground. In this case, connecting the scope ground lead to a point above ground could make the chassis "hot."

So there are several undesirable effects that we want to avoid—dc shunt, ac shunt, noise pickup, hot scope, etc. Let's examine ways to make ungrounded measurements.

Conventional Scopes. The only way to get ideal ungrounded measurements is with a special scope with a differential-amplifier input. Next best is to use a scope with a dual-trace amplifier input and an "A-B" mode. But even with a conventional scope, there are ways to reduce some of the undesirable effects—if not completely eliminate them.

First, we must know our scope. Is the chassis grounded to the ac line? Also, note the grounding on the device under test. If it has been determined that the scope will not cause a dc shunt, we must realize that the circuit being tested can still be disturbed by the scope's ac loading. Noise and hum pickup should be taken into account. Even if the scope chassis is not directly tied to the line ground, it can put noise into the tested circuit; the

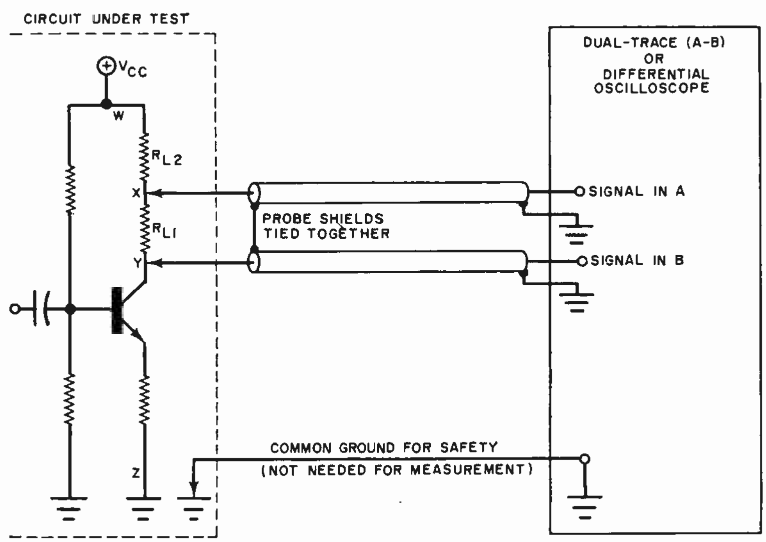


Fig. 4. Using a differential scope to make a measurement across an ungrounded component.

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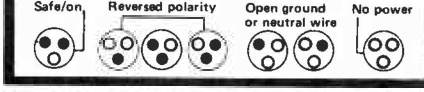

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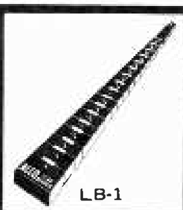

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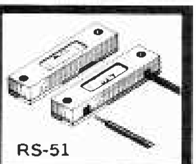

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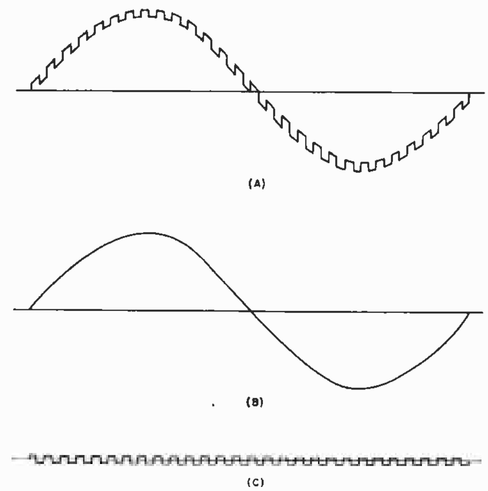

Fig. 5. Conventional scope (A) displays both signal and hum, while a differential scope (B) displays only signal across load resistor.

scope's large metal cabinet acts as an antenna and picks up noise.

We can use an isolation transformer with the scope to prevent dc shunts. This also reduces the ac shunt; but because of the transformer's primary-to-secondary capacitance, ac loading and noise will still occur. However the capacitive reactance does reduce loading and noise—as compared to a direct line without a transformer. In using an isolation transformer, remember that the scope chassis can be hot when the signal ground lead is connected to an ungrounded point.

Oscilloscope Differential Amplifier. Because the output of a differential amplifier is the difference of its inputs, we can use it to measure the voltage across a component—which is a potential difference. The significance of a differential voltage measurement is that it can be used for an ungrounded component without encountering the bad effects noted previously.

Fig. 3 shows a simplified block diagram of a differential amplifier. It consists of two identical amplifiers. They have the same gain, but one inverts its input. The outputs are then combined by algebraic addition; and since one output is inverted, the total is A − B.

Differential measurements are less common than conventional single-input measurements so scopes with differential-amplifier inputs are few. Many scopes of the plug-in type have differential-amplifier inputs. The electronics enthusiast, experimenter, or service technician could build a differential amplifier to feed a conventional single-input scope.

In practice, the signal high leads are connected to the points to be measured. No common ground connection is required so the two low leads are not used. Usually, they are shields on the test probes.

For safety reasons, however, the scope and device under test may be connected with a grounding wire. However, this connection must not be the signal low leads or the shields, to avoid ground loops.

It should be pointed out that the differential-amplifier scope is not the same as a dual-trace scope. The latter has two amplifiers, each with its own input. Electronic switching alternately feeds the output of each amplifier to the scope's vertical deflection section. The result is a simultaneous display of the two inputs.

Under certain conditions, a dual-trace scope can provide some of the benefits of a differential-amplifier scope. For instance, if there is an A − B mode and the amplifiers are well matched, the difference of the two inputs is displayed. The scope manufacturer's operation manual will explain this function where applicable.

Making Differential Measurements. Let's see how differential scope measurements are made for ungrounded tests. In the circuit shown in Fig. 4, we want to measure the signal across load resistor R_{L1}. Only the high signal leads are connected to the circuit under test (at points X and Y). The shields are connected together and grounded; but they are not grounded at the tips. This connection reduces the impedance of the loop formed by the shields and equalizes the currents through the loop, thereby allowing the differential amplifiers to nullify loop current effects by common-mode rejection.

It is not correct to tie both shields

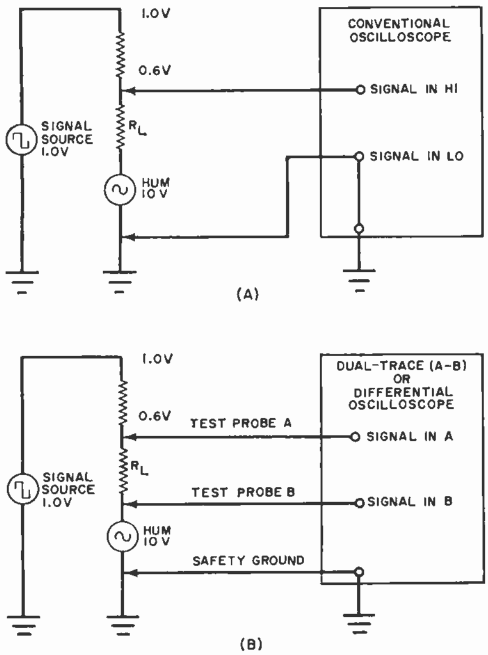

Fig. 6. Signal with hum (A), hum alone (B), and signal with hum rejected (C).

together at the probe end and also connect them to the chassis. This makes a circuit for ground currents through the shield and can create measurement errors because of voltage difference at the scope. It would also be wrong to leave both shields unconnected at the probe ends. This would permit the shields to act like antennas to pick up noise.

The probe tips represent a high impedance to the circuit being tested and do not introduce excessive loading as did the conventional test circuits shown in Figs. 1 and 2. Since the scope chassis is not tied to a signal high point, there is no ac line noise introduced.

Reject Common-Mode Signals. One of the more important uses of a differential measurement is to reduce the effects of a common-mode signal such as hum. Common-mode signals are identical with respect to amplitude and time. Since the output of a differential amplifier is the difference between its two inputs, a common (identical) signal on each input will be reduced (but not eliminated) in the output. There is a limit as to just how effective a differential amplifier can be. Its ability to reject common-mode signals is known as common-mode rejection. The ratio of the common-mode input signal amplitudes to the amplitude of the difference signal displayed on the scope is known as the common-mode rejection ratio. The higher the ratio, the better the differential amplifier.

For example, if the common-mode signal on both inputs is 10 volts and the signal produces a scope display of only 0.01 volt, the common-mode rejection ratio is 10/0.01 or 1000.

Note what happens with a conventional scope measurement as shown in Fig. 5A. There is an unwanted 60-Hz hum signal in the circuit along with the desired signal. Assume the hum is 10 volts and the square-wave source is 1 volt, of which 0.6 volt appears across the component being tested. With the conventional scope setup, both the 0.6-volt signal and the 10-volt hum would be displayed, as in Fig. 6A. The desired signal rides on the bothersome hum, making the measurement difficult.

But notice what occurs when a differential scope is used, as in Fig. 5B. In test probe A is the combined signal and hum, while the hum alone is in probe B. The scope displays A minus B or only the desired signal across the resistor, as shown in Fig. 6C. The amount of hum that is rejected depends on the scope's common-mode rejection ratio; and if the latter is good, the resultant signal would have negligible hum.

As we have pointed out, the A – B mode of a dual-trace scope can be used for differential measurements. The common-mode rejection of such a scope, however, is less than that for a differential-amplifier scope. Nevertheless, the ability to reduce common-mode signals to even a small degree would be all that is needed for making a good measurement. ◈

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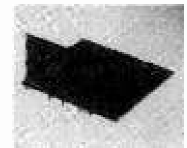

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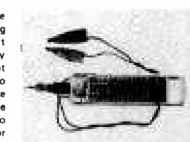

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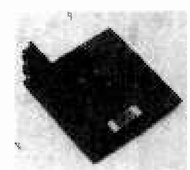

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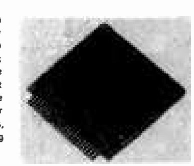

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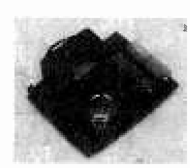

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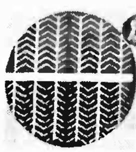

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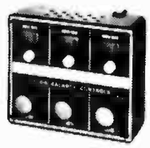

Easy to build low-cost kit needs no technical knowledge. Completed unit has 3 bands of audio frequencies to modulate 3 independent strings of colored lamps (i.e. "lows"-reds, "middies"-greens, "highs"-blues. Just connect hi-fi, radio, power amp etc. & plug ea. lamp string into own channel (max 300w ea.). Kit features 3 neon indicators, color intensity controls, controlled individ SCR circuits; isolation transformer; custom plastic housing; instr.

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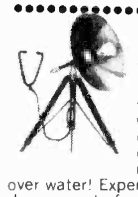
ELECTRONIC SOUND COLLECTOR



The Big Ear ™ is a science toy that's fun for all, to tune in a whisper at 20 ft., normal conversations up to 100 ft., bird calls & high freq. sounds at almost 200 ft. Nearly twice as far over water! Experiment and hobby-ize with it indoors or out - from baby sitting to bird watching. Big 18" plastic reflector disc pulls sound waves into transistor unit, then amplifies them. Mike system, stethoscope earphones, tripod (33½" extended). Sets up in secs.

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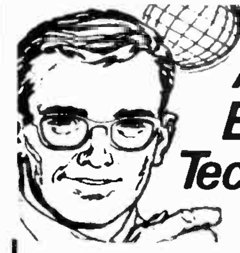

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1976 ELECTRONIC EXPERIMENTERS HANDBOOK ADVERTISERS INDEX

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five new breadboard testers from CSC

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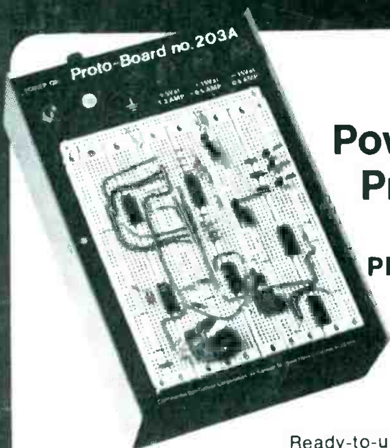

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New Proto Boards PB-203 and PB-203A with built-in regulated short-proof power supplies!

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PB-203

- 3 QT-59S Sockets
- 4 QT-59B Bus Strips
- 1 QT-47B Bus Strip
- Fuse • Power Switch
- Power-On Light
- 9.75"L x 6.6"W x 3.25"H
- Weight: 5 lbs.
- 5V, 1 AMP regulated power supply

75.

Add $2.50 shipping/handling

OUTPUT SPECIFICATIONS

Output Voltage	5V ± ¼ V
Ripple & Noise	@ ½ AMP 10 millivolts
Load Regulation	Better than 1%

PB-203A

- 3 QT-59S Sockets
- 4 QT-59B Bus Strips
- 1 QT-47B Bus Strip
- Fuse • Power Switch
- Power-On Light
- 9.75"L x 6.6"W x 3.25"H
- Weight: 5 lbs.
- 5V, 1 AMP regulated power supply (same as PB-203)
- +15V, ½ AMP regulated power supply
- −15V, ½ AMP regulated power supply

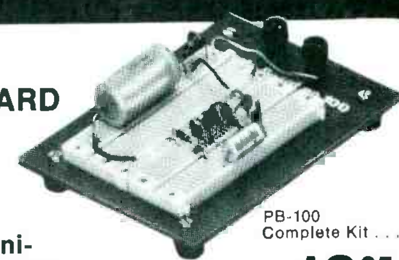
120.

Add $2.50 shipping/handling

OUTPUT SPECIFICATIONS

Output Voltage	15V, internally adjustable
Ripple & Noise	@ ¼ AMP, 10 millivolts
Load Regulation	Better than 1%

PROTO BOARD 100



A complete mini-breadboard budget kit with full IC capacity

PB-100 Complete Kit . . . **19.95**

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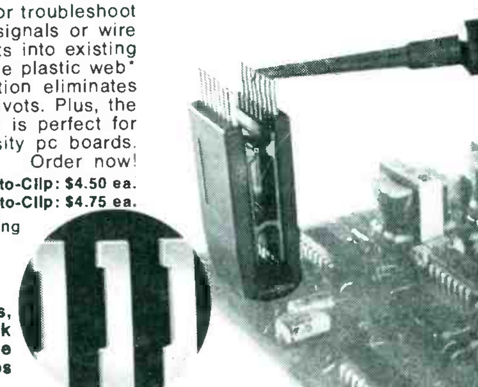
CSC PROTO-CLIP offers power-on . . . hands-off signal tracing . . . under $5!

Trace signals or troubleshoot fast. Inject signals or wire unused circuits into existing boards. Flexible plastic web* construction eliminates springs and pivots. Plus, the narrow throat is perfect for high density pc boards. Order now!

PC-14 14-pin Proto-Clip: $4.50 ea.
PC-16 16-pin Proto-Clip: $4.75 ea.

Add $1.00 shipping and handling



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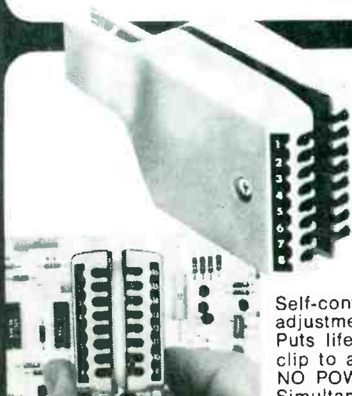

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