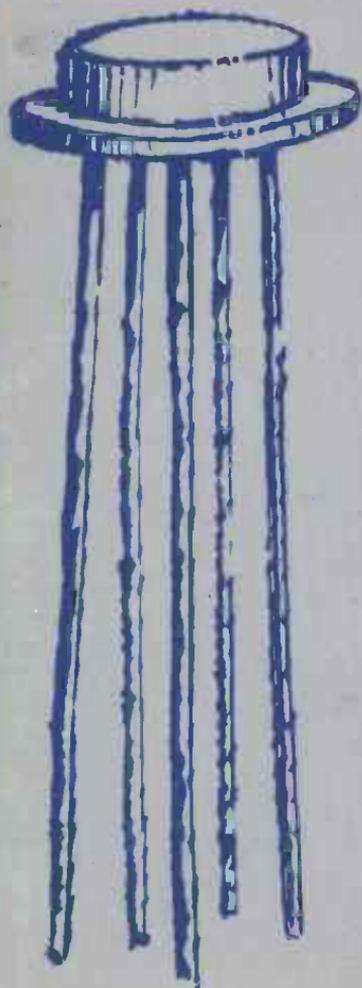


125

ONE TRANSISTOR PROJECTS

BY RUFUS P. TURNER



AUDIO AMPLIFIERS

RF, IF, & DC AMPLIFIERS

OSCILLATORS

CONTROL & ALARM DEVICES

TEST INSTRUMENTS

**POWER SUPPLY
APPLICATIONS**

**RECEIVERS, TRANSMITTERS
& ACCESSORIES**

125
One-Transistor
Projects

TAO BOOKS
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Tulsa, OK 74103
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One-Transistor
Projects

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By Rufus P. Turner



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Preface

One of the many virtues of the transistor is its ability to function single-handedly in all kinds of circuits. The 125 single-transistor circuits shown in this book are examples of this usefulness. Only a few inexpensive transistor types are used here; however, all varieties are represented: germanium, silicon, NPN, PNP, unijunction, and field-effect. The reader should be able to set up and test these circuits with a minimum of effort and equipment. The circuit designs take into consideration the comparatively wide spread in transistor ratings; manufacturers' typical values were used (or where only maximum and minimum values were specified, the minimum value was used). Thus, where a 30K-ohm resistor is specified, a choice between standard EIA values of 27K and 33K may be indicated, and it may be necessary to try both values to determine which one works best. However, in the case of frequency-determining RC networks, the exact values shown must be used. By following these instructions, the reader should have little trouble duplicating or bettering the performance claimed. These circuits are offered to the student who wants to set up and test typical circuits for his own instruction, to the designer who wants ready-made building blocks for incorporation into a system, and to the hobbyist whose amateur interests will be served by the predesigned, pretested circuits. Many of the units may be combined (as in cascaded amplifiers, oscillator-amplifier combinations, and so on) to suit individual system requirements.

RUFUS P. TURNER

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ABOUT THE AUTHOR

Rufus P. Turner has had wide experience in the semiconductor field, dating back to the crystal detector of the 1920s and to his engineering association with Sylvania when the famed 1N34 germanium diode was developed there in the 1940s. He earned his B. A. degree from California State College at Los Angeles, and his M. A. and Ph. D. degrees from the University of Southern California.

Dr. Turner wrote two of the first practical transistor books: *Transistors, Theory & Practice* (1954) and *Transistor Circuits* (1957). He is the author of 30 books and more than 2500 articles.

Dr. Turner is licensed as a registered professional engineer in California and Massachusetts.

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Chapter 1

Audio Amplifiers

The single-stage AF amplifier ranks among the simplest of transistor circuits and was one of the first. In this application, the transistor typically offers many advantages: small size, light weight, good efficiency, and low power supply demands. Here are 20 audio amplifier circuits, ranging from conventional aperiodic units to frequency-selective units. In most instances, these circuits may be cascaded as desired to obtain performance beyond the capability of a single stage. All are biased for Class A operation.

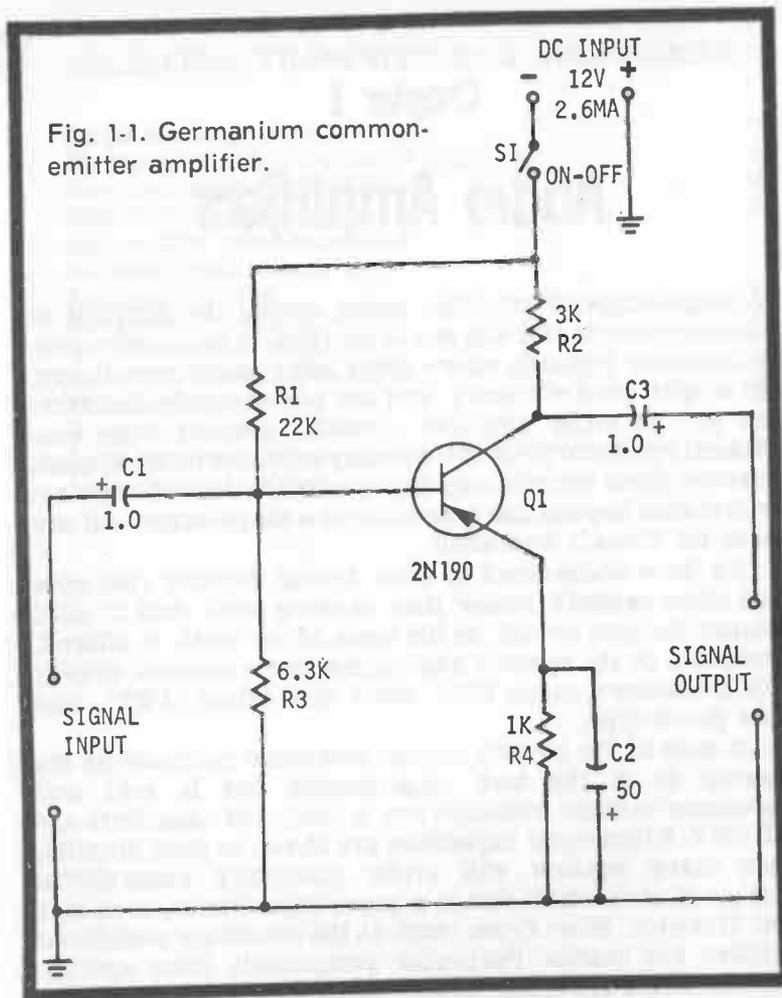
We have endeavored to show typical circuits (the ones most often needed), rather than obscure ones. And in each instance the best circuit, on the basis of our tests, is offered. Depending on the specific application, some circuits employ NPN transistors, some PNP, some field-effect (FET), and some power-type.

In each of the circuits, unless otherwise indicated on the drawing or in the text, capacitances are in mfd and resistances in ohms, resistors are $\frac{1}{2}$ watt, and capacitors are 25 DCWV. Electrolytic capacitors are shown in most circuits, since many readers will prefer miniature electrolytics because of their small size in a given capacitance, such as 1 mfd. However, other types (such as the miniature metallized tubular) are usable. Particular components from specific manufacturers are indicated only when those components (or exact equivalents, when available) seem essential to the correct performance of the circuit. In all other instances, any component having the specified ratings may be used.

1.1—GERMANIUM COMMON-EMITTER AMPLIFIER

Fig. 1-1 shows an RC-coupled common-emitter stage employing a 2N190 germanium PNP transistor. Operated from a 12-volt DC supply, this circuit provides an open-circuit voltage gain of 100 at 1 kHz. The maximum input-signal amplitude before the onset of output-signal peak clipping is 10 mv RMS and the corresponding maximum output-signal amplitude is 1v RMS. Voltage gain is down 3 db at 100 Hz and 10

Fig. 1-1. Germanium common-emitter amplifier.



kHz. The input impedance is approximately 1000 ohms at 1 kHz.

DC base bias is obtained from emitter resistor R4 and voltage divider R1-R3. The circuit draws 2.6 ma at 12v DC.

1-2-SILICON COMMON-EMITTER AMPLIFIER

The common-emitter stage shown in Fig. 1-2 employs an inexpensive 2N2712 silicon transistor. In most respects, this circuit is similar to the equivalent germanium circuit in Fig. 1-1, but the silicon transistor offers several worthwhile advantages: higher input impedance than that of equivalent germanium transistors, lower collector current, better

frequency response, and higher temperature tolerance. A particular advantage is the simpler DC base-biasing arrangement which may be used with the silicon transistor—the single dropping resistor, R1.

Operated from a 6-volt DC supply at a 100 µa current drain, this amplifier stage provides an open-circuit voltage gain of 125 at 1 kHz. The maximum input-signal amplitude before the onset of output-signal peak clipping is 8 mv RMS and the corresponding maximum output-signal amplitude is 1v RMS. Voltage gain is 3 db down at 70 Hz and 140 kHz. The input impedance is approximately 2500 ohms at 1 kHz.

1-3-DEGENERATIVE COMMON-EMITTER AMPLIFIER

In a common-emitter stage, negative feedback through current degeneration is easily obtained by omitting the

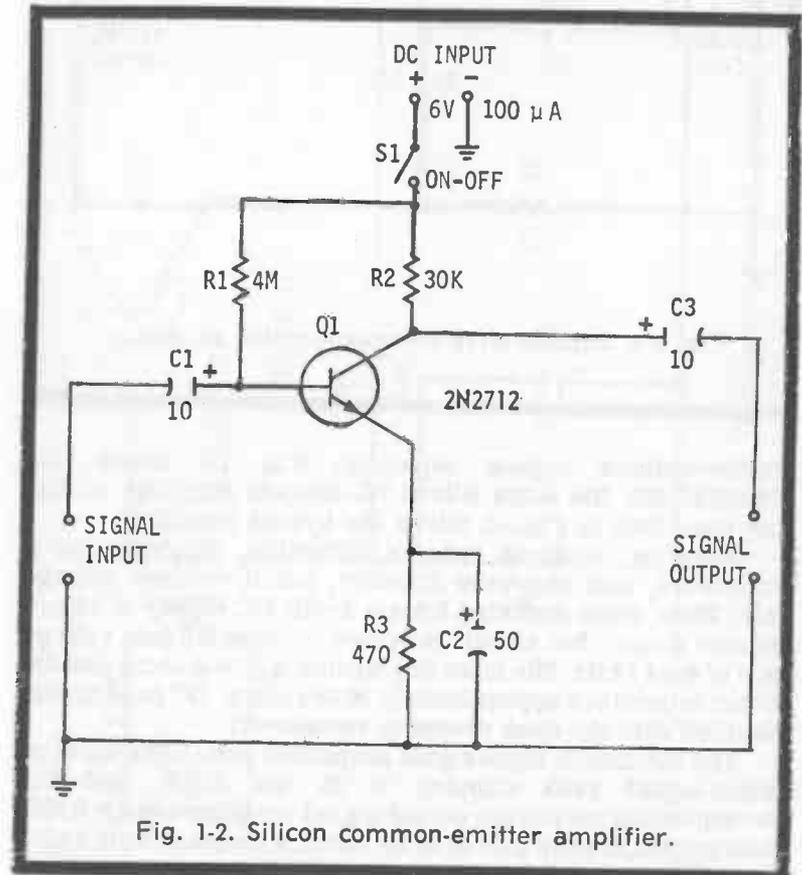


Fig. 1-2. Silicon common-emitter amplifier.

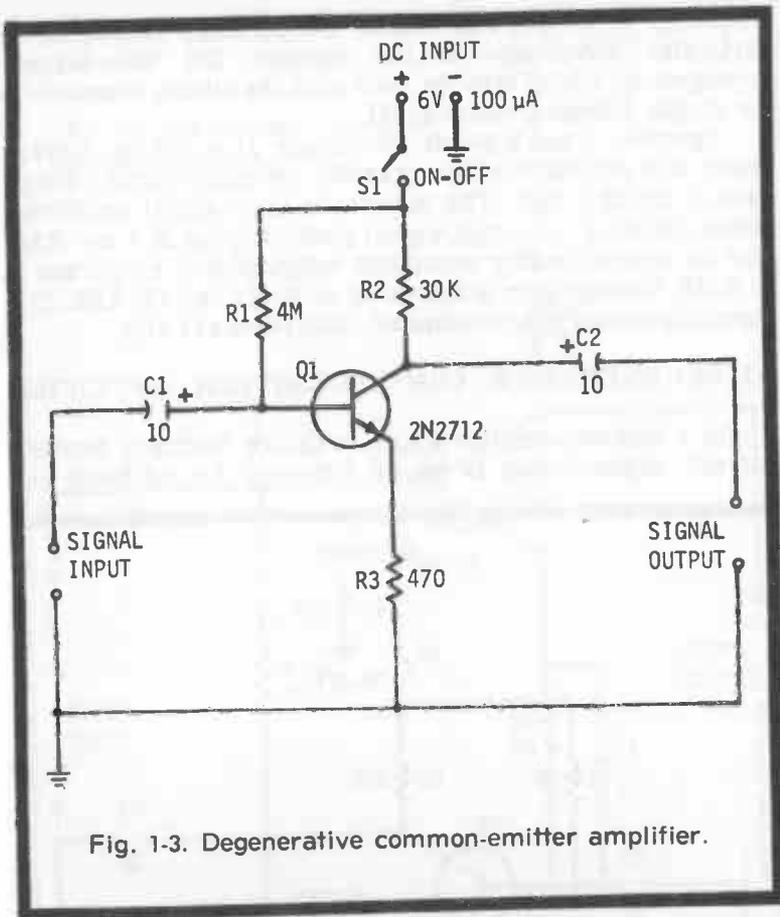


Fig. 1-3. Degenerative common-emitter amplifier.

emitter-resistor bypass capacitor. Fig. 1-3 shows this arrangement, the same silicon RC-coupled amplifier circuit just described in Fig.1-2 minus the bypass capacitor.

Negative feedback reduces distortion, increases input impedance, and improves linearity, but it reduces voltage gain. Thus, when operated from a 6-volt DC supply at 100 μ A current drain, this circuit provides an open-circuit voltage gain of 40 at 1 kHz. The input impedance is 50,000 ohms and the output impedance approximately 30,000 ohms. DC base bias is obtained entirely from dropping resistor R1.

The maximum input-signal amplitude before the onset of output-signal peak clipping is 25 mv RMS, and the corresponding maximum output-signal amplitude is 1v RMS. Voltage gain is down 3 db at 50 Hz but it is constant from 1 kHz to 100 kHz.

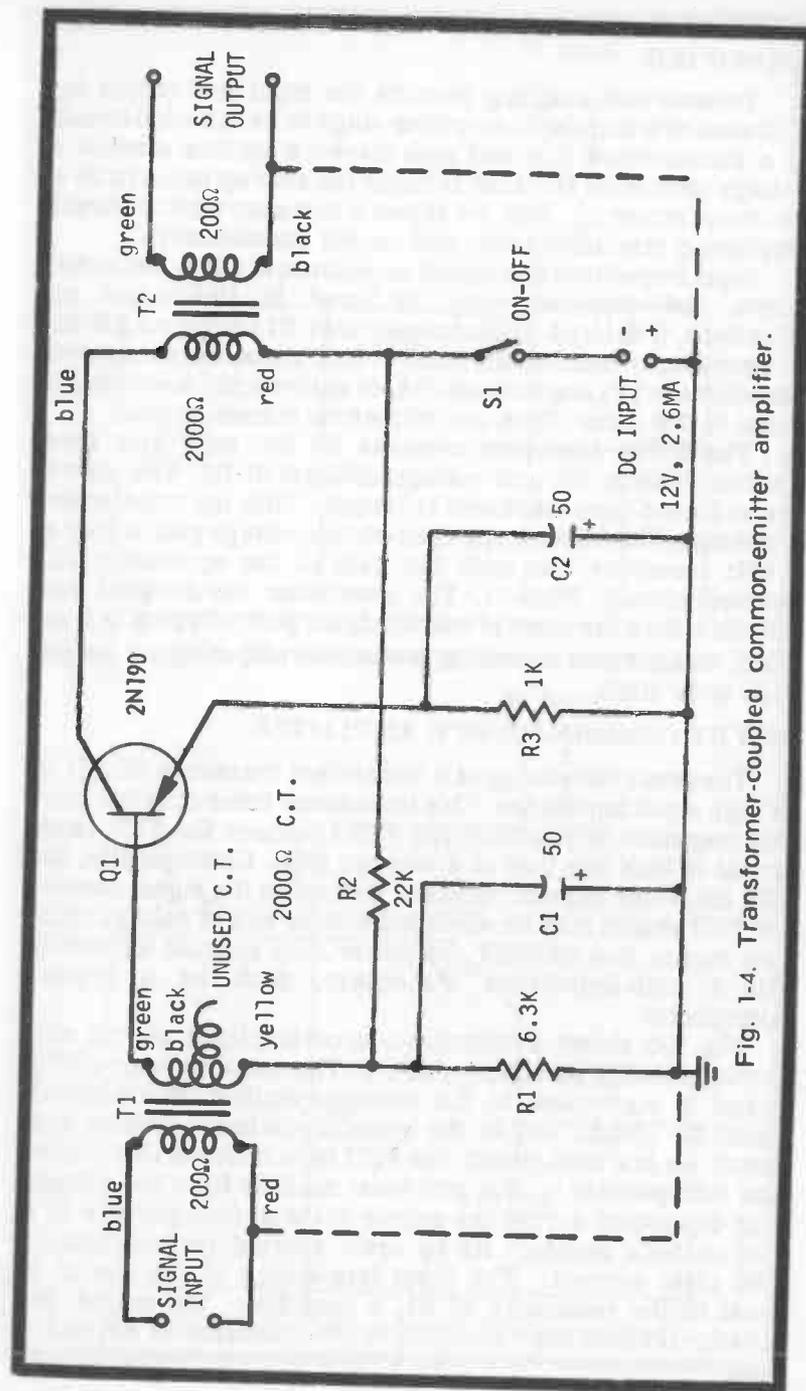


Fig. 1-4. Transformer-coupled common-emitter amplifier.

1-4—TRANSFORMER-COUPLED COMMON-EMITTER AMPLIFIER

Transformer coupling permits the input and output impedances of a transistor amplifier stage to be matched closely to a transmission line and also allows a certain amount of voltage gain to be obtained through the step-up turns ratio of the transformer(s). Fig. 1-4 shows a common-emitter circuit employing miniature input and output transformers.

Input impedance and output impedance are low (200 ohms); hence, the amplifier may be used in line-to-line applications, if desired. Input transformer T1 (Argonne AR-123, or equivalent) matches 200 ohms to 2000 ohms, whereas output transformer T2 (Argonne AR-116, or equivalent) matches 2000 ohms to 200 ohms. Both are miniature transformers.

The 2N190 transistor receives its DC base bias from emitter resistor R3 and voltage divider R1-R2. The circuit draws 2.6 ma from the 12-volt DC supply. With the transformer impedance ratios shown, the open-circuit voltage gain is 2000 at 1 kHz (compare this with the gain of the equivalent RC-coupled circuit, Fig.1-1). The maximum input-signal amplitude before the onset of output-signal peak clipping is 1 mv RMS, and the corresponding maximum output-signal amplitude is 2v RMS.

1-5—FET COMMON-SOURCE AMPLIFIER

The great advantage of a field-effect transistor (FET) is its high input impedance. This impedance (characteristically, 1000 megohms in junction-type FETs) causes the FET input circuit to look like that of a vacuum tube. Consequently, the FET amplifier imposes virtually no load on the signal source, and FET stages may be cascaded with no loss of voltage gain. This means that an FET amplifier may operate efficiently with a high-impedance transducer, such as a crystal microphone.

Fig. 1-5 shows a common-source amplifier circuit employing a 2N3823 junction-type FET. The common-source FET circuit is equivalent to the common-emitter conventional-transistor circuit and to the common-cathode vacuum-tube circuit. As in a tube circuit, the FET here receives its DC gate bias (comparable to tube grid bias) entirely from the voltage drop developed across the source resistor (comparable to a tube cathode resistor) R3 by drain current (comparable to tube plate current). The input impedance of the circuit is equal to the resistance of R1, 5 megohms; the output impedance is equal approximately to the resistance of R2, 10,000 ohms.

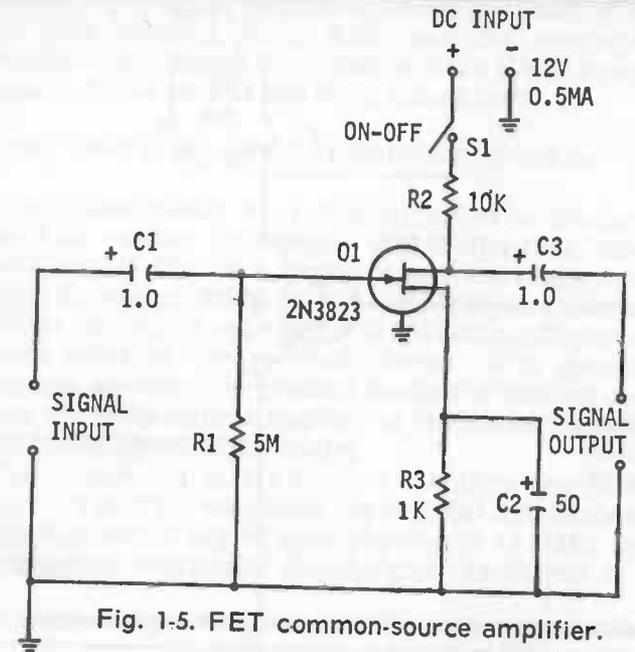


Fig. 1-5. FET common-source amplifier.

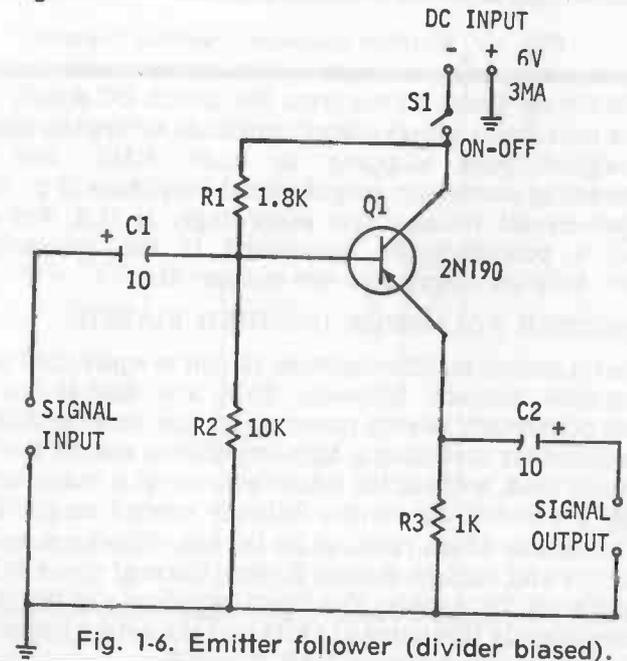


Fig. 1-6. Emitter follower (divider biased).

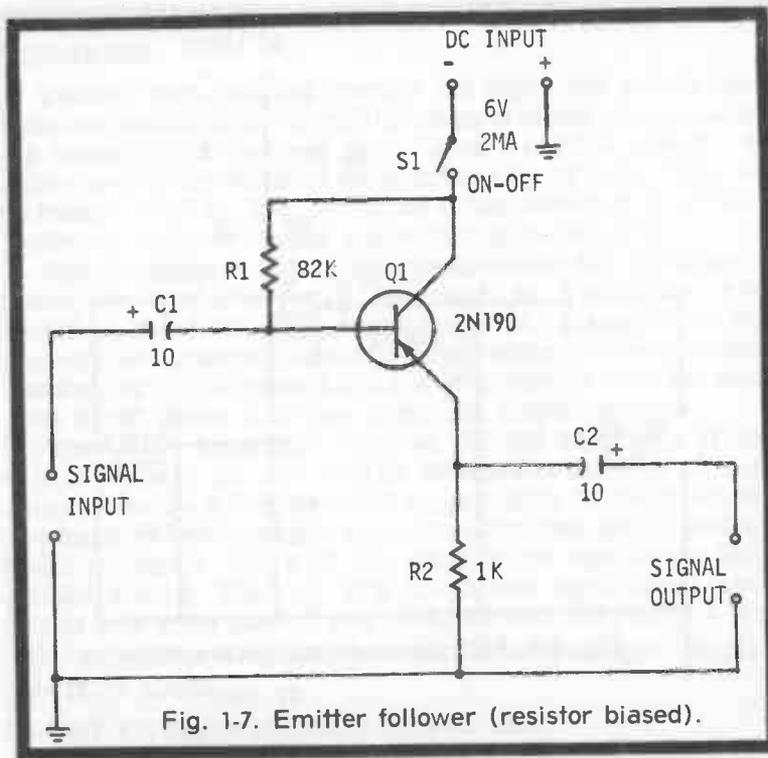


Fig. 1-7. Emitter follower (resistor biased).

The circuit draws 0.5 ma from the 12-volt DC supply. At 1 kHz, the maximum output-signal amplitude before the onset of output-signal peak clipping is 0.15v RMS, and the corresponding maximum output-signal amplitude is 2v RMS. The open-circuit voltage gain accordingly is 13.3. For gain control, a potentiometer, connected in the conventional manner, may be substituted for resistor R1.

1-6—EMITTER FOLLOWER (DIVIDER BIASED)

The transistor emitter-follower circuit is equivalent to the vacuum-tube cathode follower; both are high-to-low impedance converters having power gain. The emitter follower is convenient for matching a high-impedance source to a low-impedance load, without the inconvenience of a transformer.

Fig. 1-6 shows an emitter-follower circuit employing a 2N190 transistor which receives its DC base bias from emitter resistor R3 and voltage divider R1-R2. Current drain is 3 ma from the 6-volt DC supply. The input impedance of the circuit is approximately 1500 ohms at 1 kHz and the output impedance is 800 ohms.

The open-circuit voltage gain is 0.95. At 1 kHz the

maximum input-signal amplitude before the onset of output-signal peak clipping is 1v RMS and the corresponding maximum output-signal amplitude is 0.95v RMS. Frequency response is flat to 100 kHz and down 3 db at 100 Hz.

1-7—EMITTER FOLLOWER (RESISTOR BIASED)

A somewhat higher input impedance can be obtained in a germanium emitter follower if the divider-type base-bias circuit shown in Fig. 1-6 is replaced with the single dropping resistor, R1, shown in Fig. 1-7. The 1-kHz input impedance of this latter circuit, for example, is 20,000 ohms, compared with the 1500 ohms of the previous circuit. With germanium transistors, however, the resistor method of biasing does not provide the temperature stability of the operating point afforded by voltage-divider biasing.

The circuit in Fig. 1-7 draws 2 ma from the 6-volt DC supply. At 1 kHz the maximum input-signal amplitude before the onset of output-signal peak clipping is 1v RMS and the corresponding maximum output-signal amplitude is 0.96v

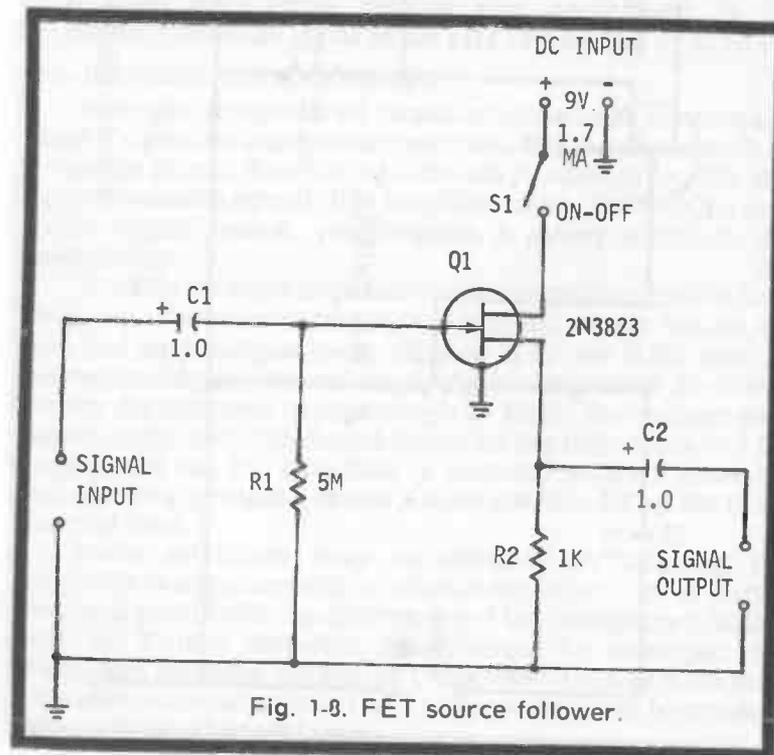


Fig. 1-8. FET source follower.

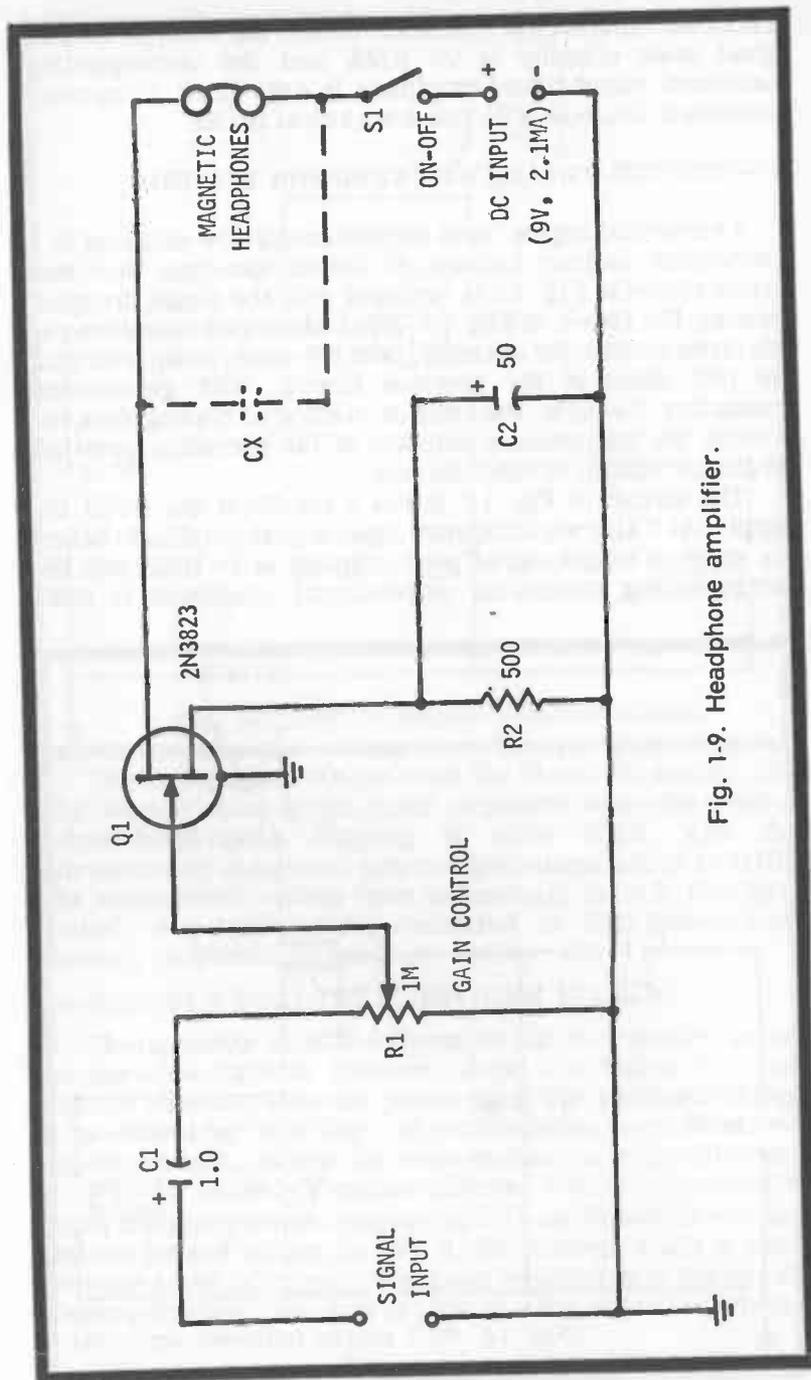


Fig. 1-9. Headphone amplifier.

RMS. The open-circuit voltage gain accordingly is 0.96. Frequency response is flat to 100 kHz and down 3 db at 100 Hz. Output impedance is 520 ohms.

1-8-FET SOURCE FOLLOWER

When a field-effect transistor is employed in a high-to-low impedance converter, the arrangement is a source-follower circuit which is equivalent to the emitter follower and cathode follower. The advantage of the FET is its very high input impedance, comparable to that of a vacuum tube.

Fig. 1-8 shows a source-follower circuit employing a 2N3823 junction-type FET. At 1 kHz the input impedance of this circuit is equal to R_1 , 5 megohms, and the output impedance is 560 ohms. The circuit draws 1.7 ma from the 9-volt DC supply. The DC gate bias is obtained entirely from the voltage drop produced across source resistor R_2 by the flow of drain current. For gain control, a potentiometer, connected in the conventional manner, may be substituted for resistor R_1 .

At 1 kHz the maximum input-signal amplitude before the onset of output-signal peak clipping is 1v RMS and the corresponding maximum output-signal amplitude is 0.7v RMS. The open-circuit voltage gain accordingly is 0.7. Frequency response is flat to 100 kHz and down 3 db at 50 Hz.

1-9-HEADPHONE AMPLIFIER

A simple, low-powered amplifier sometimes is needed to boost a signal for headphone operation. Fig. 1-9 shows such an amplifier circuit. Employing a 2N3823 junction-type FET in a common-source circuit, this amplifier offers virtually no load to the signal source, yet develops a strong signal in the headphones.

At 1 kHz the input impedance of the circuit is equal to R_1 , 1 megohm. The maximum input-signal amplitude before the onset of output-signal peak clipping is 50 mv RMS and the corresponding maximum output-signal amplitude developed across the 2000-ohm headphones is 3v RMS. The voltage gain accordingly is 60. The circuit draws 2.1 ma from the 9-volt DC supply and the DC gate bias is obtained entirely from the voltage drop produced across source resistor R_2 by the drain current flow.

Audio selectivity may be obtained by shunting the magnetic headphones with a suitable capacitor (C_x) to form a resonant circuit with the inductance of the headphones. With a pair of Trimm 2000-ohm headphones, for example, the frequency response peaked at 1 kHz when C_x was 0.005 mfd. However, some adjustment of the capacitance will be required with individual headphones.

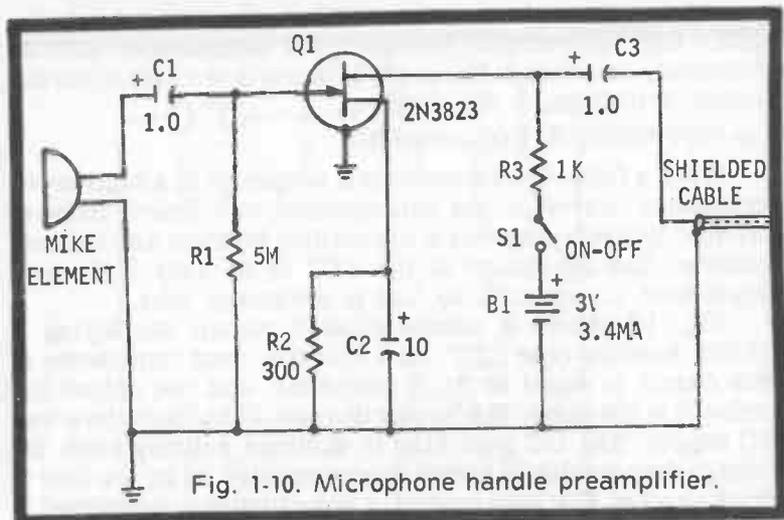


Fig. 1-10. Microphone handle preamplifier.

1-10—MICROPHONE-HANDLE PREAMPLIFIER

Due to its small size and ability to operate from one or two penlight cells, the transistor has made it practicable to enclose a self-powered preamplifier in the handle of a low-output microphone—a decided convenience in adapting certain low-level microphones to existing speech amplifiers.

Fig. 1-10 is a schematic of such an amplifier, operated from two 1.5v penlight cells connected in series. The high input impedance of the 2N3823 field-effect transistor (Q1) enables this amplifier to be used with any type of microphone without loading the latter. The circuit draws 3.4 ma from the 3-volt DC supply.

At 1 kHz the input impedance of the circuit is equal to R1, 5 megohms. The output impedance is equal to R3, 1000 ohms, and is low enough to minimize problems of stray-signal pickup and of coupling into the main amplifier. At 1 kHz the maximum input-signal amplitude before the onset of output-signal peak clipping is 0.4v RMS and the corresponding maximum output-signal amplitude is 2v RMS. The open-circuit voltage gain is 5.

1-11—DRIVER FOR CLASS-B STAGE

Fig. 1-11 shows the circuit of a Class A driver for a Class B pushpull amplifier stage. This driver delivers 12 mw output at 1 kHz and draws 2.6 ma from the 12-volt DC supply.

In this arrangement, the 2N190 transistor receives its DC base bias from emitter resistor R3 and voltage divider R2-R4.

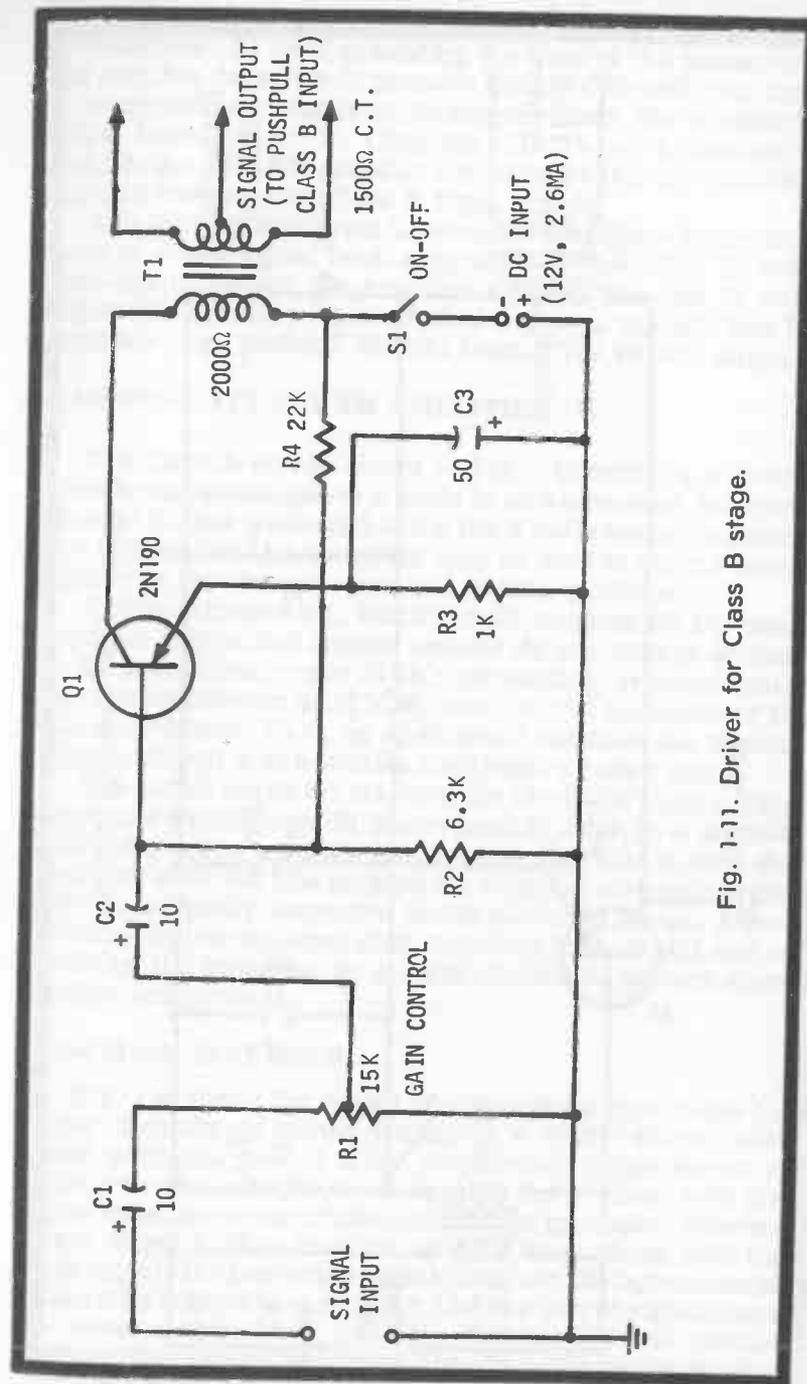


Fig. 1-11. Driver for Class B stage.

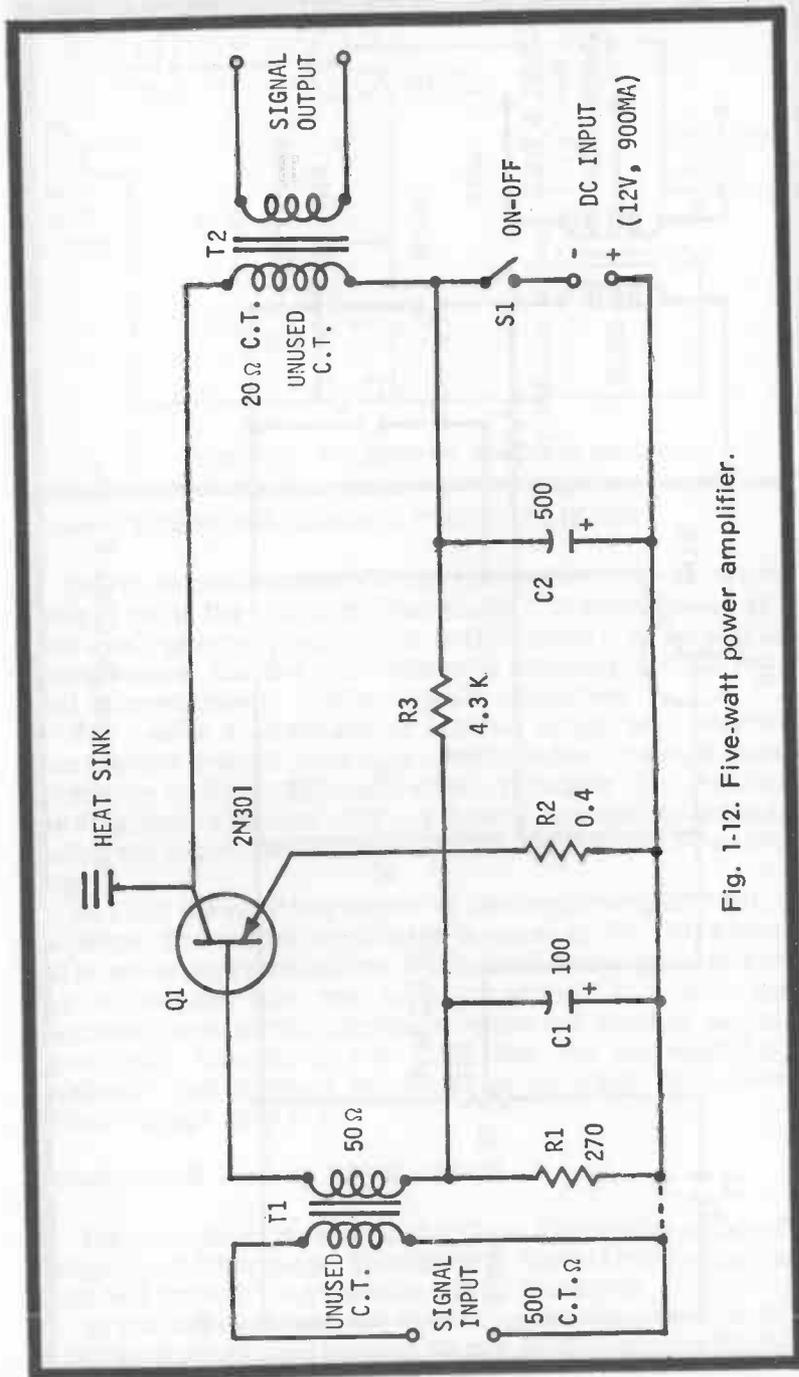


Fig. 1-12. Five-watt power amplifier.

Blocking capacitor C2 is needed to prevent gain-control potentiometer R1 from grounding the base of the transistor, and coupling capacitor C1 protects the input circuit from any DC component arriving from the signal source. The miniature output transformer, T1 (Argonne AR-175, or equivalent), matches the 2000-ohm collector circuit to the typical 1500-ohm pushpull base-emitter Class B input circuit.

At 1 kHz the maximum input-signal amplitude before the onset of output-signal peak clipping is 10 mv RMS. At this input-signal voltage the amplifier delivers the full 12 w output which is more than sufficient to drive a typical Class B amplifier (e.g., pushpull 2N109s) from 175 to 200 mw output.

1-12—FIVE-WATT POWER AMPLIFIER

The Class A circuit shown in Fig. 1-12 employs a 2N301 power transistor to deliver 5 watts to an 8-ohm load. An input signal of 2.5 mw will result in the full 5 watts output, a power gain of 2000. Such an amplifier may be used to drive a loud-speaker or for various servo and control purposes.

In this arrangement, transistor Q1 receives its DC base bias from unbypassed emitter resistor R2 and voltage divider R1-R3. Input transformer T1 (Knight 54A2373, or equivalent) matches the 500-ohm input to 50 ohms; output transformer T2 (Chicago-Stancor TA-12, or equivalent) matches the 20-ohm collector circuit to an 8-ohm load. All resistors are 1 watt.

The circuit draws 901 ma from the 12-volt DC supply. This represents a continuous DC power level of 10.8w, so a suitable heat sink must be provided (as shown in Fig. 1-12) to keep the transistor cool. For this purpose the collector electrode of the 2N301 is internally connected to the mounting flange, a fact that must be remembered when attaching the heat sink and in mounting the transistor on a metal chassis to prevent short circuits and grounds.

1.13—PHASE INVERTER

Fig. 1-13 shows the circuit of a paraphase-type phase inverter. This simple circuit, employing a 2N2712 silicon transistor, performs well at audio frequencies. Signal output 1, taken from the collector, is 180 degrees out of phase with the signal input, due to the triode action of the transistor, whereas signal output 2, taken from the emitter, is in phase with the input signal. The two output signals thus are 180 degrees out of phase with respect to each other. The two output signals have the same amplitude if collector resistor R2 and emitter resistor R3 are closely matched (the absolute resistance value

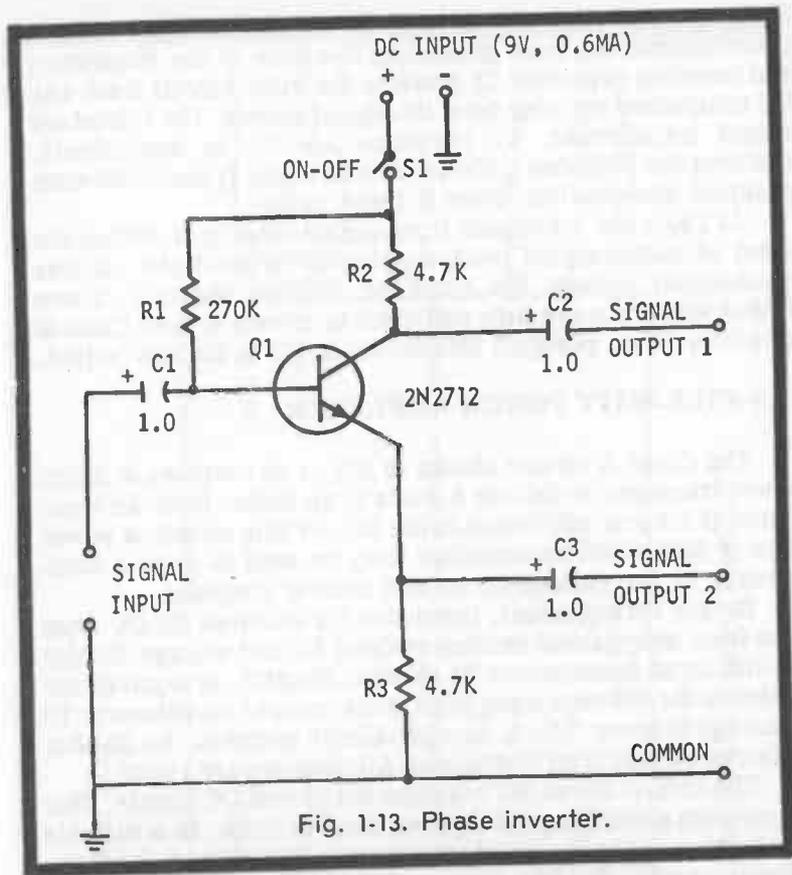


Fig. 1-13. Phase inverter.

is relatively unimportant, so long as both resistors have the same resistance). This type of phase inverter is basically a common-emitter amplifier for signal output 1 and an emitter follower for signal output 2, with both amplifiers having identical voltage gain.

The circuit draws 0.6 ma from the 9-volt DC supply. DC base bias is obtained through 270K dropping resistor R1. At 1 kHz the maximum input-signal amplitude before the onset of output-signal peak clipping is 1.7v RMS and the corresponding maximum output-signal amplitude is 1.67v RMS for either signal output 1 or signal output 2. The open-circuit voltage gain is 0.98. Input impedance is 50,000 ohms at 1.kHz.

1-14—AUDIO MIXER

A 3-channel-input audio mixer circuit is shown in Fig. 1-14.

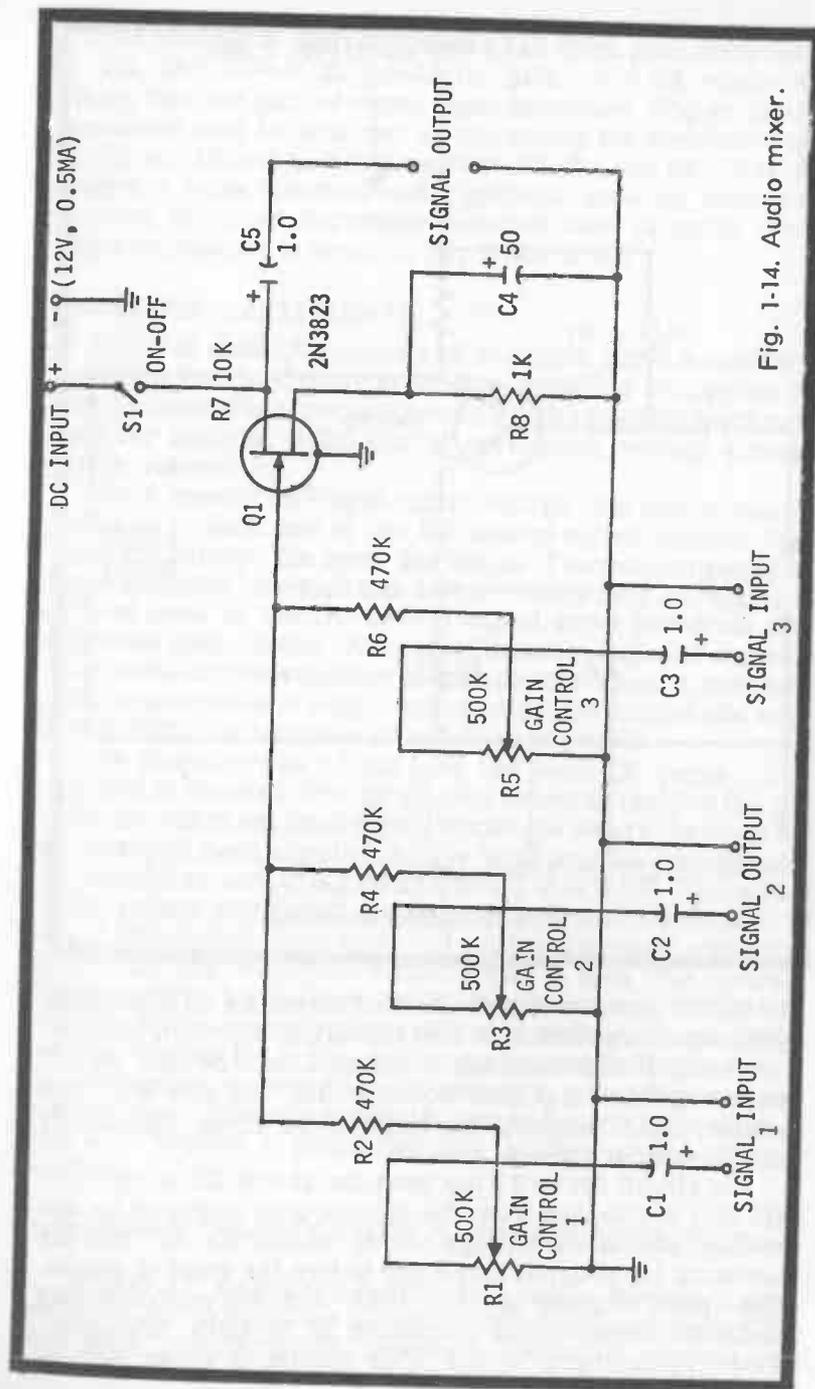


Fig. 1-14. Audio mixer.

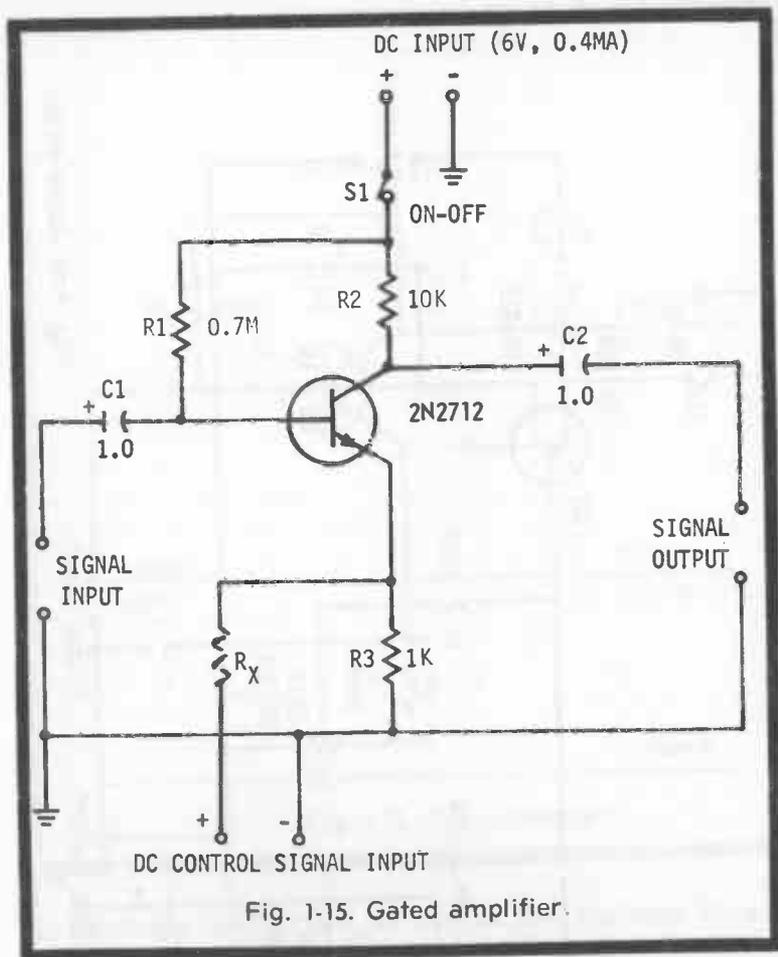


Fig. 1-15. Gated amplifier.

The 2N3823 junction-type field-effect transistor (Q1) provides a high input impedance, so that the circuit offers virtually no load to signal sources (such as microphones) beyond that of the $\frac{1}{2}$ -megohm input gain controls (R1, R3, and R5). The common signal output is the blend of the three individually controlled input signals.

The circuit draws 0.5 ma from the 12-volt DC supply. DC gate bias is obtained from the voltage drop produced by the flow of drain current through source resistor R8. At 1 kHz the maximum input-signal amplitude before the onset of output-signal peak clipping is 0.15v RMS and the corresponding maximum output-signal amplitude is 2v RMS. The open-circuit voltage gain is 13.3. This applies to either pair of signal input terminals.

The maximum input impedance (all three gain controls—R1, R3, and R5—set to maximum gain) is 0.328 megohm looking into any pair of signal input terminals. Higher input impedance may be obtained by increasing the resistance of R1, R3, and R5 and isolating resistors R2, R4, and R6. Thus, if 1-megohm potentiometers and 1-megohm isolating resistors are used, the input impedance becomes 0.667 megohm. The output impedance is equal to R7, 10,000 ohms.

1-15—GATED AMPLIFIER

Fig. 1-15 shows the circuit of a simple gated amplifier, actually a standard common-emitter amplifier employing a 2N2712 silicon transistor, but one in which provision has been made for applying a DC gating, or control, voltage across emitter resistor R3.

For a constant AF input-signal voltage, the output-signal voltage is proportional to the DC control-signal voltage; the higher the voltage, the lower the output. Thus, the output of a larger amplifier in which this unit is a stage may be rectified and fed back to the DC control signal input terminals for automatic gain control. Also, a sufficiently high DC control voltage will cut the amplifier off completely. Thus, a positive-going square wave or step function of proper magnitude will gate (switch) the amplifier on and off as required.

The circuit draws 0.4 ma from the 6-volt DC supply. DC base bias is obtained through 0.7-meg dropping resistor R1. At 1 kHz the maximum input-signal amplitude before the onset of output-signal peak clipping is 0.15v RMS and the corresponding maximum output-signal amplitude is 1v RMS. The open-circuit voltage gain is 6.67. Input impedance is 50,000 ohms at 1 kHz. A DC control signal of 6v will cut the transistor collector current off and reduce the output signal to zero. The control function draws 3.6 ma from the DC control-voltage source. If isolation of the control-voltage source is desired, an isolating resistor (dotted Rx in Fig. 1-15) may be installed. However, this will cause a proportionate reduction in applied control voltage, due to the voltage division across Rx and R3 in series, and will require a proportionately higher control-voltage source.

1-16—TUNED-TRANSFORMER BANDPASS AMPLIFIER

Tuned AF amplifiers peaked at a single frequency are useful as active bandpass filters in bridge null detectors, CW and MCW radiotelegraph reception, electronic musical instruments, and electronic control applications. Amplifiers of

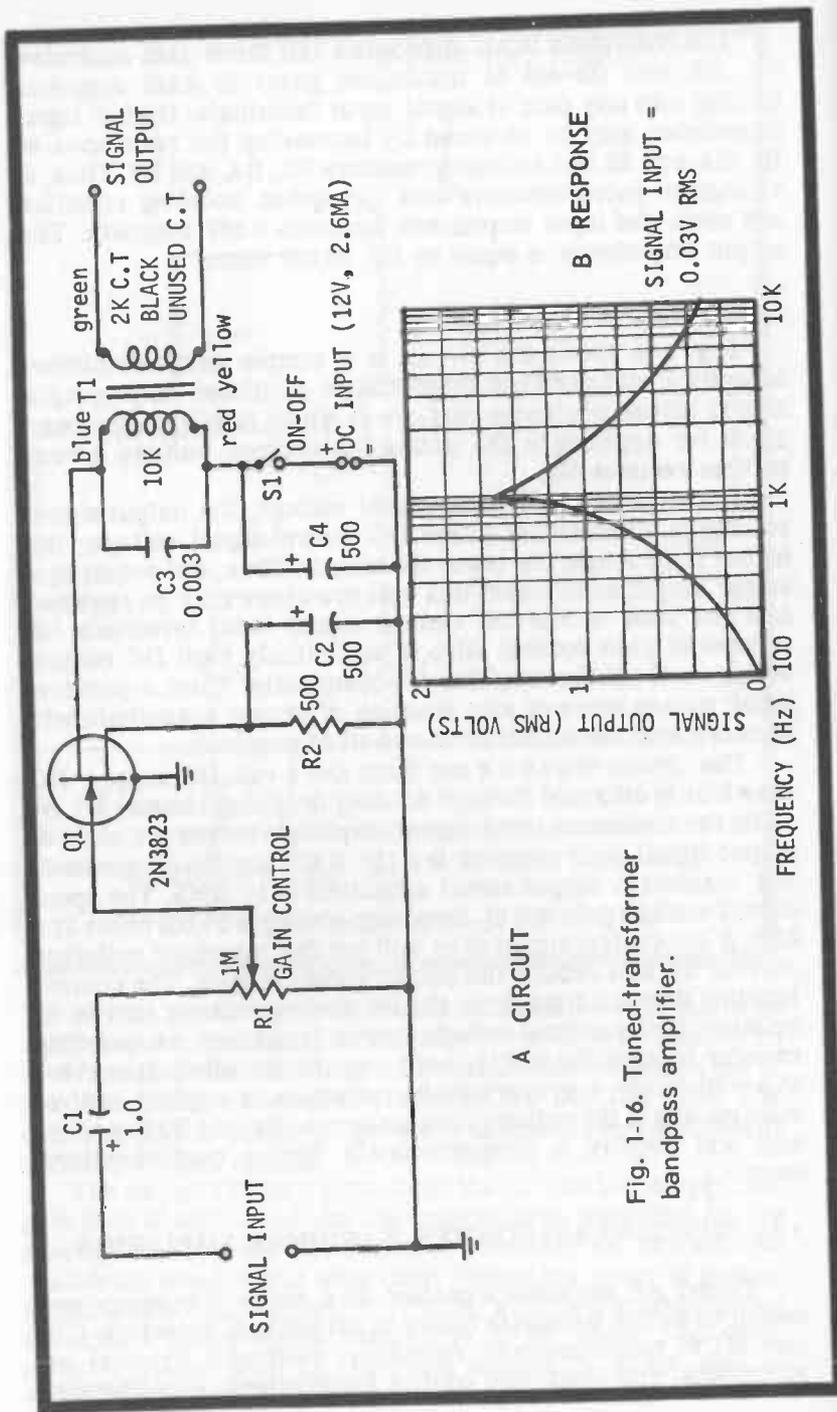


Fig. 1-16. Tuned-transformer bandpass amplifier.

this type are shown in Figs. 1-16, 1-17, and 1-19 and are described in the corresponding sections.

In Fig. 1-16A, the amplifier is tuned by a parallel resonant circuit consisting of capacitor C3 and the inductance of the primary winding of miniature output transformer T1 (Argonne AR-109, or equivalent). The 0.003-mfd capacitor shown here tunes the transformer to approximately 1 kHz (Fig. 1-16B shows the amplifier frequency response), but the capacitance will require some adjustment with an individual transformer. For higher frequencies, reduce the capacitance; for lower frequencies, increase it.

The circuit employs a 2N3823 field-effect transistor in a common-source amplifier. Current drain is 2.6 ma from the 12-volt DC supply. DC gate bias is obtained from the voltage drop developed across source resistor R2 by the drain-current flow. The input impedance is equal to the resistance of gain control R1, 1 megohm.

At the 1-kHz resonant peak the maximum input-signal amplitude before the onset of output-signal peak clipping is 30 mv RMS and the corresponding maximum output-signal amplitude is 1.5v RMS. The open-circuit voltage gain at the resonant peak is 50.

1-17—INDUCTOR-CAPACITOR-TUNED BANDPASS AMPLIFIER

Somewhat better selectivity than that afforded by the amplifier described in the preceding Section may be obtained with the bandpass amplifier circuit shown in Fig. 1-17A. The frequency response is shown in Fig. 1-17B.

In this circuit, the parallel resonant circuit which tunes the amplifier consists of capacitor C2 and a miniature, high-Q, adjustable, 5-henry inductor, L1 (U. T. C. Type VI-C15, or equivalent). This LC combination is inserted into the negative feedback loop between the collector output circuit and the base input circuit of the 2N190 transistor. The wavetrap action of the parallel-resonant combination (L1C2) removes one frequency (actually, a narrow band of frequencies) from the fed-back signal voltage. Consequently, there is no negative feedback at that frequency and the amplifier readily transmits a signal of that frequency. At all other frequencies, however, negative feedback reduces the amplifier gain and effectively prevents transmission of signals at those frequencies. The net result is the sharply peaked bandpass response shown in Fig. 1-17B.

Since the inductor is screwdriver adjustable, the amplifier may be tuned precisely. The specified inductor, for example,

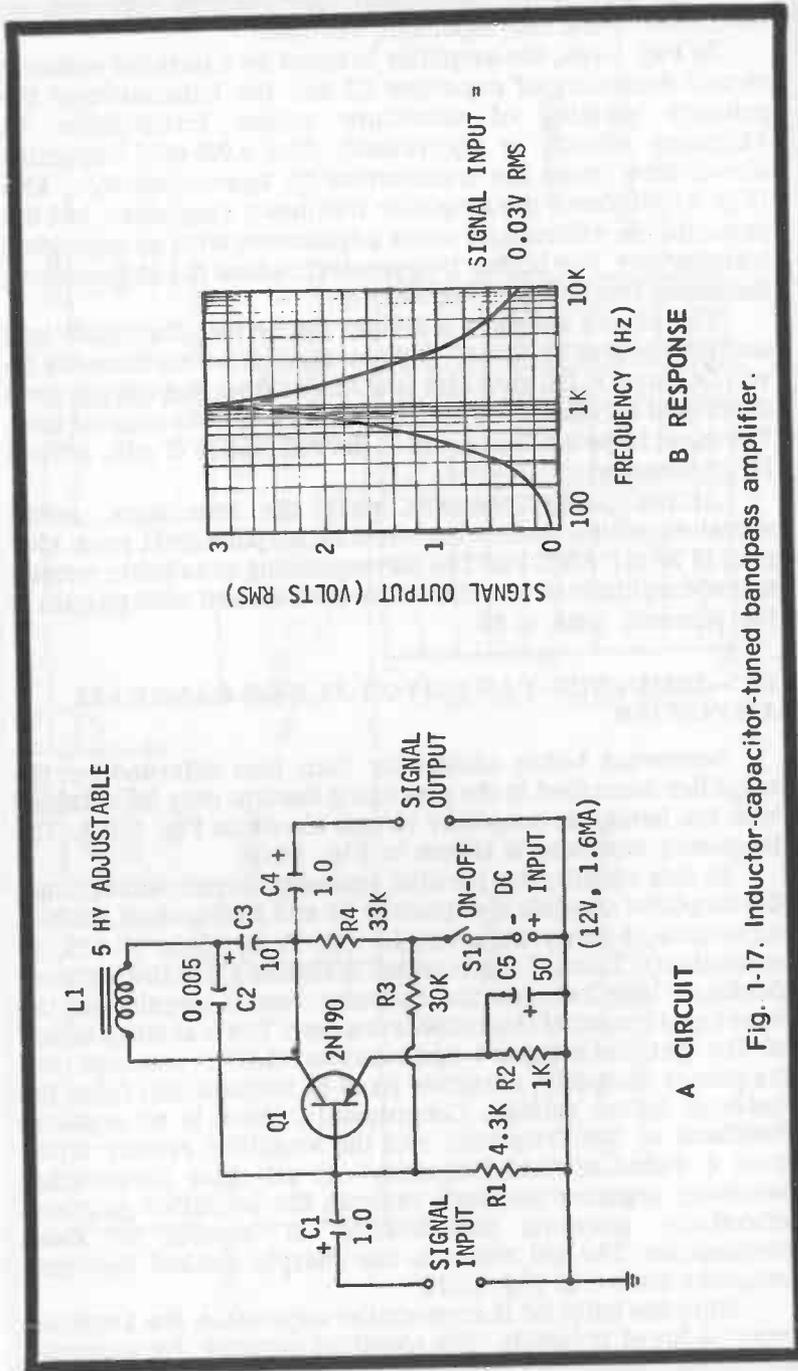


Fig. 1-17. Inductor-capacitor-tuned bandpass amplifier.

is adjustable from 2.7 hy to 6 hy, approximately, which means that if C2 is 0.005 mfd, the obtainable frequency range is 919 Hz to 1370 Hz. For higher frequencies, decrease C2; for lower frequencies, increase it. The 10-mfd coupling capacitor (C3) is not a part of the resonant circuit, but serves to block the passage of DC from the collector back to the base of the transistor.

The circuit draws 1.6 ma from the 12-volt DC supply. DC base bias is obtained from emitter resistor R2 and voltage divider R1-R3. At the 1-kHz resonant peak, the maximum input-signal amplitude before the onset of output-signal peak clipping is 30 mv RMS, and the corresponding output-signal amplitude is 3v RMS. The open-circuit voltage gain at the resonant peak accordingly is 100.

1.18—INDUCTOR-CAPACITOR-TUNED BANDSTOP AMPLIFIER

AF amplifiers tuned to reject a single frequency are useful as active bandstop filters (also called notch, band-suppression, or band-elimination filters) in harmonic distortion meters, interference eliminators, CW and MCW radiotelegraphy, and electronic control applications. Amplifiers of this type are shown in Figs. 1-18 and 1-20 and are described in the corresponding sections.

In Fig. 1-18A, the amplifier is tuned by a series resonant circuit consisting of capacitor C2 and a miniature, high-Q, adjustable, 5-henry inductor, L1 (U. T. C. Type VI-C15, or equivalent). This LC combination is inserted into the negative feedback loop between the collector output circuit and the base input circuit of the 2N190 transistor. The pass action of the series-resonant combination (L1C2) transmits one frequency (actually, a narrow band of frequencies) through the feedback path, and ideally removes all others. Consequently, the amplifier gain is cancelled at that frequency and its output ideally reduced to zero, whereas signals at all other frequencies are transmitted by the amplifier. The net result is the notch-type frequency response shown in Fig. 1-18B.

As in the bandpass amplifier described in the preceding Section 1.17, since the inductor is screwdriver adjustable, the amplifier may be tuned precisely. The specified inductor, for example, is adjustable from 2.7 hy to 6 hy, approximately, which means that if C2 is 0.005 mfd, the obtainable frequency range is 919 Hz to 1370 Hz. For higher frequencies, decrease C2; for lower frequencies, increase it. In this circuit, tuning capacitor C2 also serves as a blocking capacitor to prevent the passage of DC from the collector back to the base of the transistor.

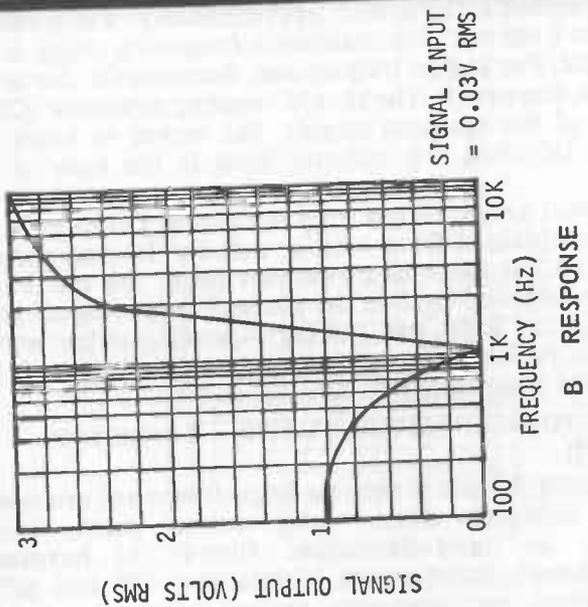
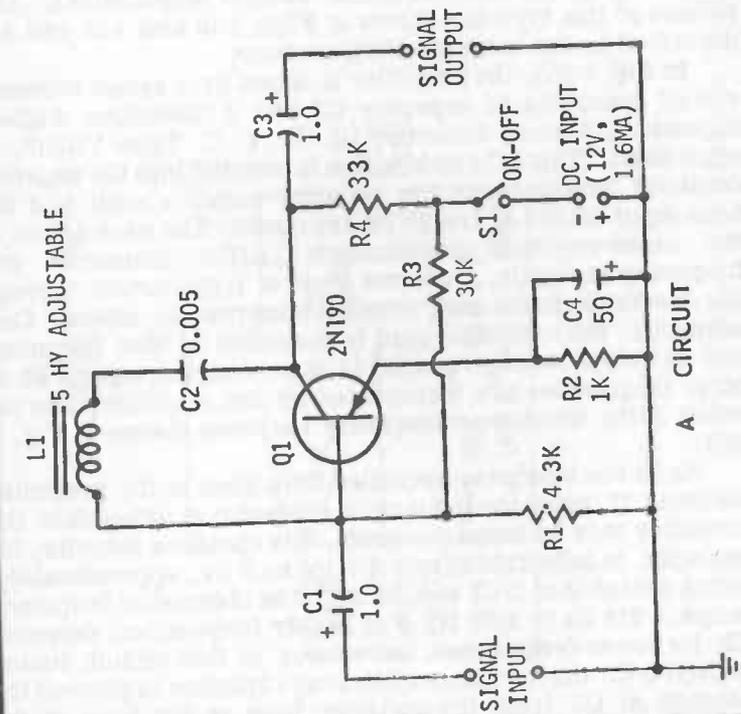


Fig. 1-18. Inductor-capacitor-tuned bandstop amplifier.

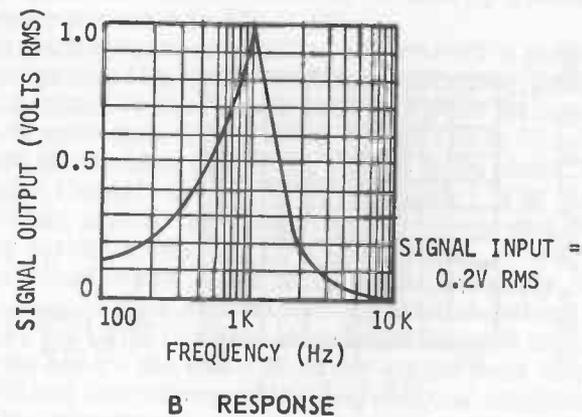
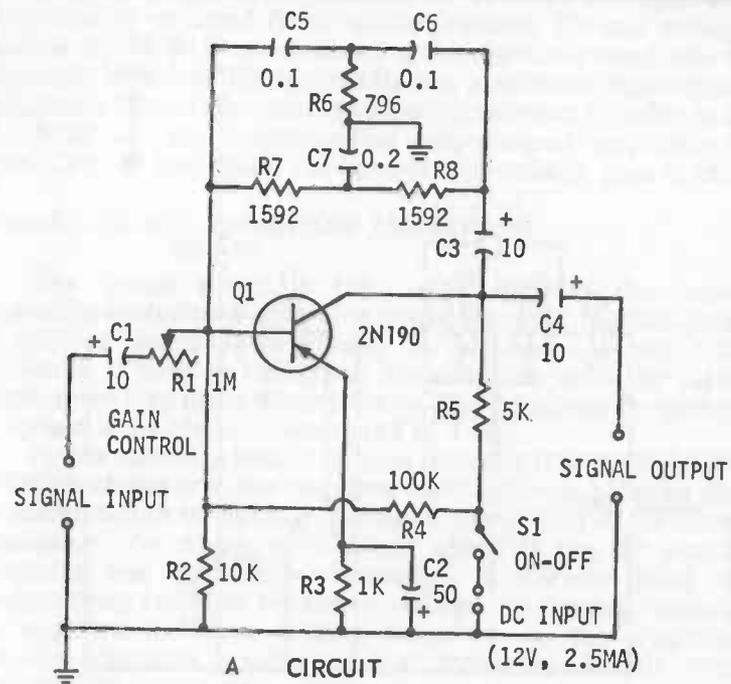


Fig. 1-19. RC-tuned bandpass amplifier.

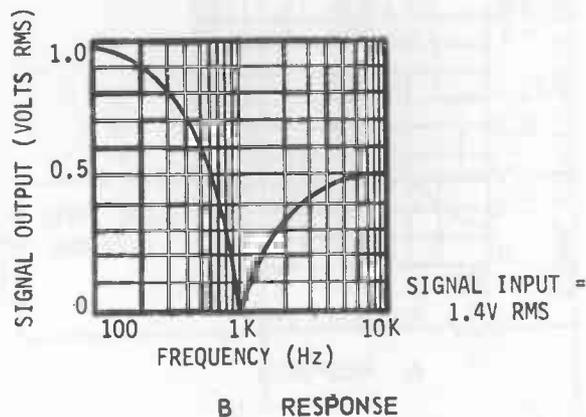
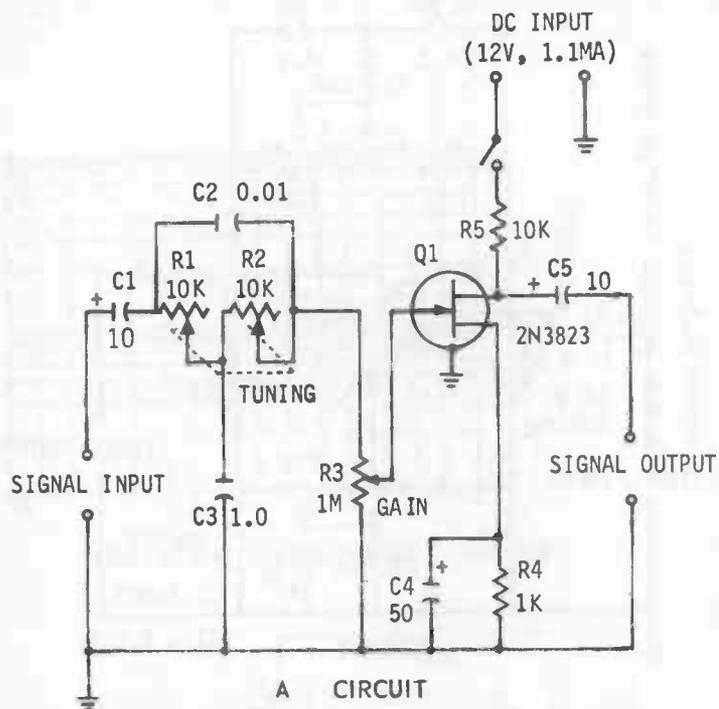


Fig. 1-20. RC-tuned bandstop amplifier.

The circuit draws 1.6 ma from the 12-volt DC supply. DC base bias is obtained from emitter resistor R2 and voltage divider R1-R3. At some distance to the high-frequency side of the 1-kHz resonant dip, say 10 kHz, the maximum input-signal amplitude before the onset of output-signal peak clipping is 30 mv RMS and the corresponding output-signal amplitude is 2.9v RMS. At that point, the open-circuit voltage gain is 96.7.

1-19—RC-TUNED BANDPASS AMPLIFIER

The circuit shown in Fig. 1-19A operates the same, basically, and serves the same applications as the bandpass amplifiers described previously in Sections 1.16 and 1.17. However, it uses no inductors; instead, it is tuned by three resistances and three capacitances. The bandpass frequency response of the circuit is shown in Fig. 1-19B.

In this circuit, a twin-T RC null network (C5-C6-C7-R6-R7-R8) is inserted into the negative-feedback loop between the collector output circuit and the base input circuit of the 2N190 transistor. The notch (or nulling) action of the RC circuit removes one frequency (actually, a narrow band of frequencies) from the feed-back voltage. As a result, there is no negative feedback at that frequency, so the amplifier readily transmits a signal of that frequency. At all other frequencies, however, negative feedback reduces the amplifier gain and effectively prevents transmission of signals at those frequencies. The net result is the sharply peaked bandpass response shown in Fig. 1-19B.

The notch frequency of the twin-T network is governed by the capacitances (C5, C6, C7) and the resistances (R6, R7, R8). Here, C5 equals C6 equals $\frac{1}{2}C7$, and R7 equals R8 equals $2R6$. For 1-kHz operation, C5 is 0.1 mfd, C6 is 0.1 mfd, C7 is 0.2 mfd, R7 is 1592 ohms, R8 is 1592 ohms, and R6 is 796 ohms. For any frequency, f equals $10^6 / (6.28C5R7)$, where f is in Hz, C5 in microfarads, and R7 in ohms. The capacitors and resistors must be carefully selected for exact values.

The circuit draws 2.5 ma from the 12-volt supply. DC base bias is obtained from emitter resistor R3 and voltage divider R2-R4. At the 1-kHz resonant peak the maximum input-signal amplitude before the onset of output-signal peak clipping is 0.2v RMS and the corresponding output-signal amplitude is 1v RMS. The open-circuit voltage gain at the resonant peak is 5.

1-20—RC-TUNED BANDSTOP AMPLIFIER

The circuit in Fig. 1-20A serves the same basic applications as the bandstop amplifier described in Section 1.18.

However, it uses no inductors; rather, it is tuned by two resistances (R1, R2) and two capacitances (C2, C3) connected into a bridged-T null (notch) network. Resistances R1 and R2 are the two sections of a ganged 10,000-ohm wirewound potentiometer. Capacitors C2 and C3 must be carefully selected for exact values.

In this circuit, the bridged-T null network (R1-R2-C2-C3) is inserted at the input of the 2N3823 field-effect transistor which operates as a common-source amplifier, and this network removes one frequency (depending upon the setting of dual potentiometer R1-R2) from the applied input signal. The network output is amplified and the notch-type frequency response of the complete circuit resembles Fig. 1-20B. If C2 is 0.01 mfd and C3, 1 mfd, varying potentiometer R1-R2 from maximum to minimum resistance varies the notch frequency from 200 Hz to 2000 Hz. This tuning range may be changed to 20-to-200 Hz by making C2 1 mfd and C3 10 mfd, and to 2000-to-20,000 Hz by making C2 0.001 mfd and C3 0.01 mfd. Since the selective RC network is followed by a field-effect transistor, there is virtually no loading to upset the selectivity (loading resulting from the 1-megohm gain-control potentiometer (R3) is negligible). The amplifier partially compensates for the inherent attenuation of the bridged-T network at frequencies off resonance.

The circuit draws 1.1 ma from the 12-volt DC supply. DC gate bias is obtained from the voltage drop across source resistor R4, resulting from drain-current flow. At some distance to the low-frequency side of the 1-kHz resonant dip (notch), say 100 Hz, the maximum input-signal amplitude before the onset of output-signal peak clipping is 1.4v RMS and the corresponding output-signal amplitude is 0.9v. At that point, the open-circuit voltage gain is 0.643.

Chapter 2

RF, IF & DC Amplifiers

The transistor is eminently useful in single-stage radio-frequency, intermediate-frequency, and direct-current amplifiers. The conventional (bipolar) transistor is especially interesting in some DC applications (as shown in Sections 2.6 and 2.7) since, unlike a vacuum tube or field-effect transistor, it is a true current amplifier. Modern RF transistors are capable of operation far into the radio spectrum.

In this Chapter is a collection of amplifier circuits selected from many tested. In each circuit, unless otherwise indicated on the drawing or in the text, capacitances are in mfd and resistances in ohms, resistors are $\frac{1}{2}$ watt, and all capacitors are rated to withstand a DC working voltage of 25v.

Particular components are specified only when those units from specific manufacturers (or exact equivalents, when available) seem essential to correct performance of the circuit. In all other instances, any component having the specified ratings may be used.

2.1—BROADCAST BAND RF AMPLIFIER (PRESELECTOR)

A tuned radio-frequency amplifier suitable for use as a preselector in the standard broadcast band is shown in Fig. 2-1. This unit is continuously tunable from 500 to 1700 kHz and has a 20-kHz bandwidth (between 3 db 0.707 points) at 1000 kHz.

Employing a 2N169A NPN transistor, the circuit draws approximately 10 ma from the 6-volt DC supply. Maximum input-signal amplitude is 300 mv RMS for an undistorted power gain of 20 db in the amplifier. Input and output impedances are approximately 500 ohms each.

Shielding of the input (L1-L2) and output (L3-L4) transformers is necessary to prevent oscillation in the amplifier, which otherwise would need to be neutralized. Commercial, shielded, tapped-coil RF transformers may be used or the

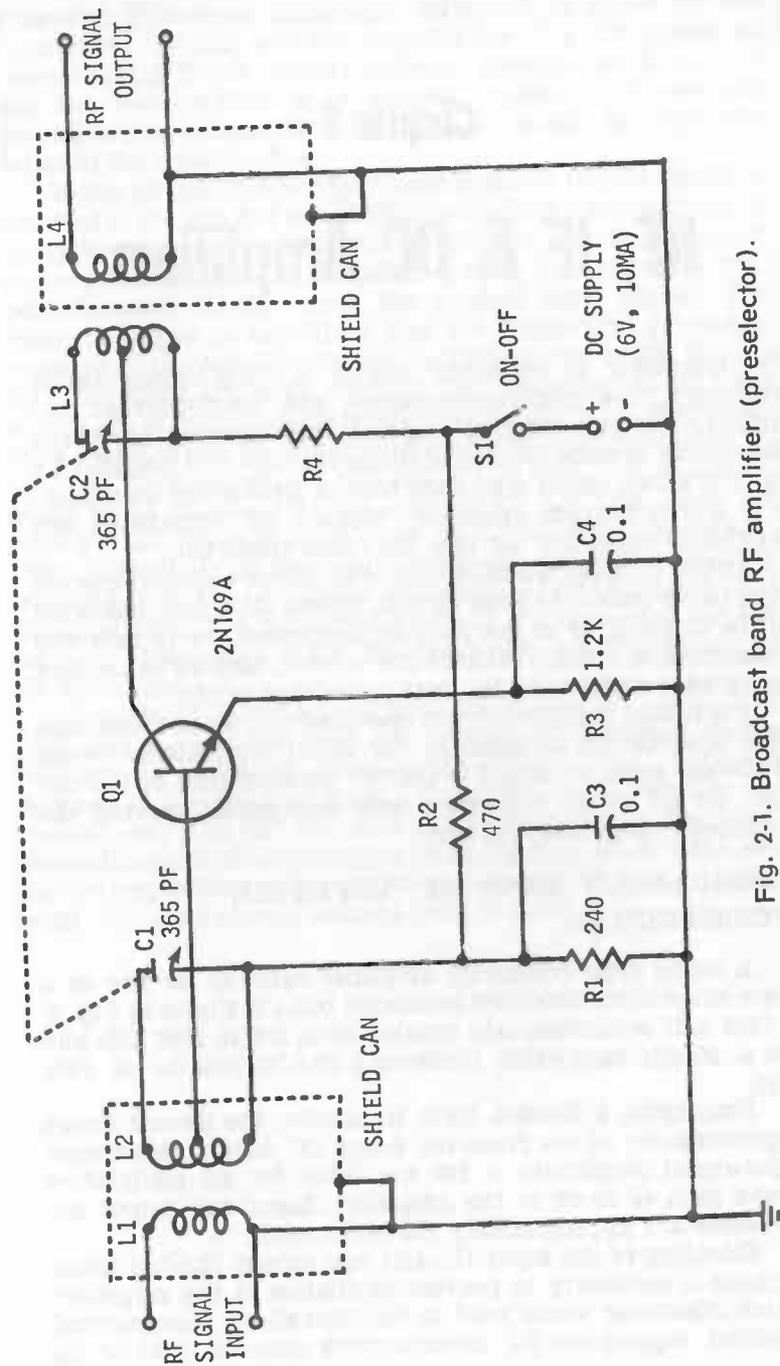


Fig. 2-1. Broadcast band RF amplifier (preselector).

coils may be wound with No. 32 enameled wire on 1-inch diameter forms according to Table 2-1.

The two sections (C1, C2) of the tuning capacitor are ganged; however, the rotors cannot be connected together electrically, or the DC circuit will be short-circuited. Instead, the rotors must be linked by means of an insulated shaft coupling. The entire tuning capacitor assembly must be insulated from a metal chassis, if one is used as a foundation for the amplifier. It is for these reasons that a standard 2-gang tuning capacitor cannot be used.

This RF amplifier circuit may be operated at higher frequencies if the tuned circuits are proportioned accordingly and a high-frequency transistor type selected.

TABLE 2-1. COIL TABLE FOR R-F AMPLIFIER

L1—10 turns of No. 32 enameled wire closewound around the lower end of L2 and insulated from the latter.

L2—130 turns of No. 32 enameled wire closewound on a 1"-diameter form. Tap 15th turn from the lower end.

L3—130 turns of No. 32 enameled wire closewound on a 1"-diameter form. Tap 80th turn from the lower end.

L4—12 turns of No. 32 enameled wire closewound around the lower end of L3 and insulated from the latter.

2.2—CONVENTIONAL 455-kHz IF AMPLIFIER

Fig. 2-2 shows the circuit of a conventional, transformer-coupled intermediate-frequency amplifier for 455-kHz operation. Employing a 2N394 PNP transistor, the circuit draws approximately 10 ma from the 6-volt DC supply. DC base bias is obtained from emitter resistor R4 and voltage divider R2-R3.

The miniature, tapped, stepdown, 455-kHz IF transformers (T1, T2) are core tuned. (T1 input: J. W. Miller Type 2041, or equivalent, T2 output: J. W. Miller 2042, or equivalent.) Neutralization, to prevent IF oscillation, is provided by a 30-pf miniature ceramic capacitor (C4) when connected to the correct end of the T2 secondary winding.

This amplifier circuit may be used at higher frequencies (say, 10.7 MHz) by substituting high-frequency IF transformers for the 455-kHz units and by selecting a suitable high-frequency transistor type.

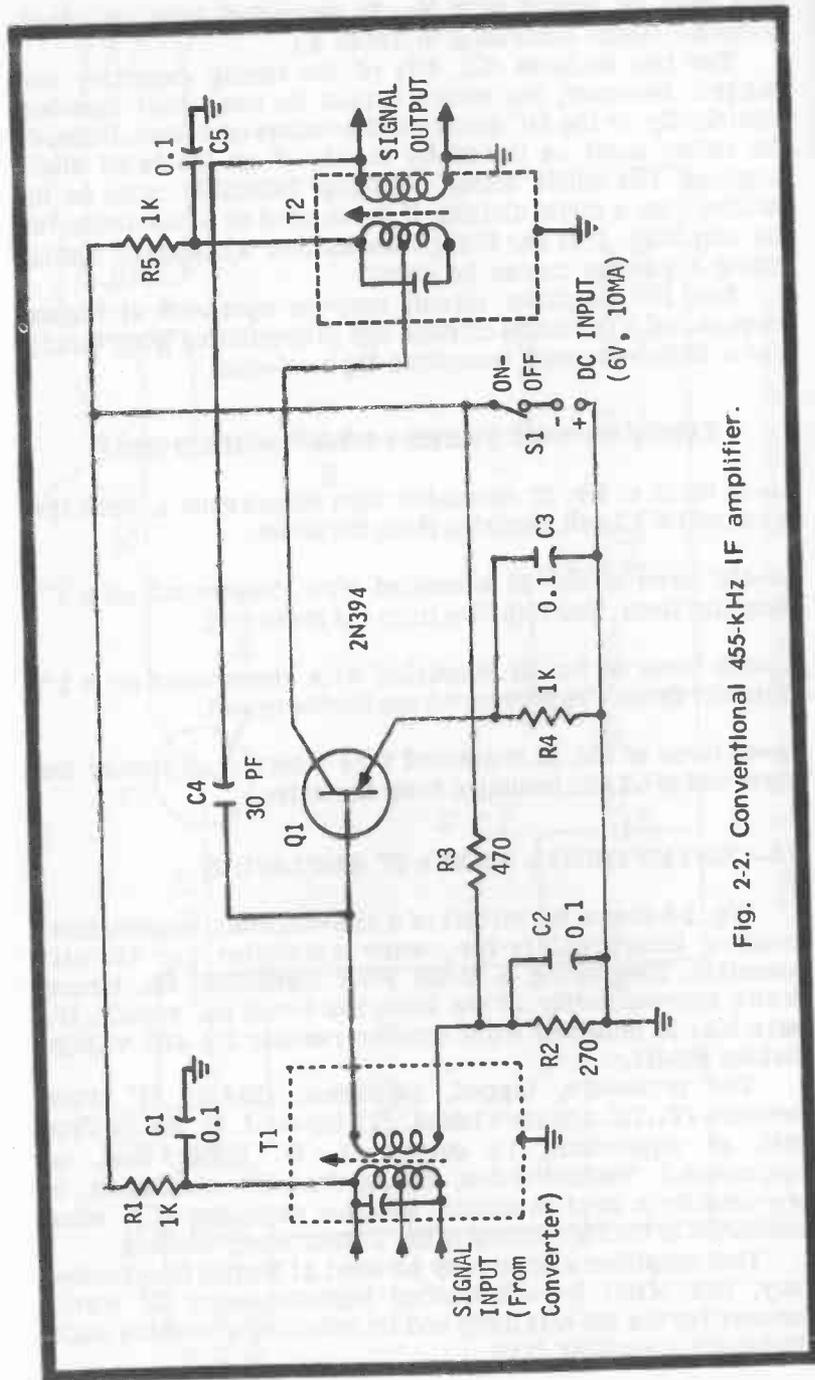


Fig. 2-2. Conventional 455-kHz IF amplifier.

2.3—SINGLE-TUNED 455-kHz IF AMPLIFIER

A 455-kHz IF amplifier which is somewhat simpler than the one described in Fig. 2-2 is shown in Fig. 2-3. This latter circuit employs only one tuning unit, output transformer T1; and although it has lower selectivity than that of the double-tuned circuit, its response is adequate for a number of applications (simple radio receivers, heterodyne test instruments, etc.).

Employing a 2N394 PNP transistor, the circuit draws approximately 1.5 ma from the 6-volt DC supply. DC base bias is obtained from emitter resistor R3 and voltage divider R1-R2. The miniature trimmer capacitor (C2) allows the circuit to be neutralized to prevent IF oscillation, provided the capacitor is connected to the correct end of the T1 secondary winding. The miniature stepdown 455-kHz output transformer, T1 (J. W. Miller Type 2042, or equivalent) is core tuned.

This amplifier circuit also may be used at higher frequencies (say, 10.7 MHz), provided a suitable high-frequency transformer is substituted for the 455-kHz unit and a suitable high-frequency transistor is selected.

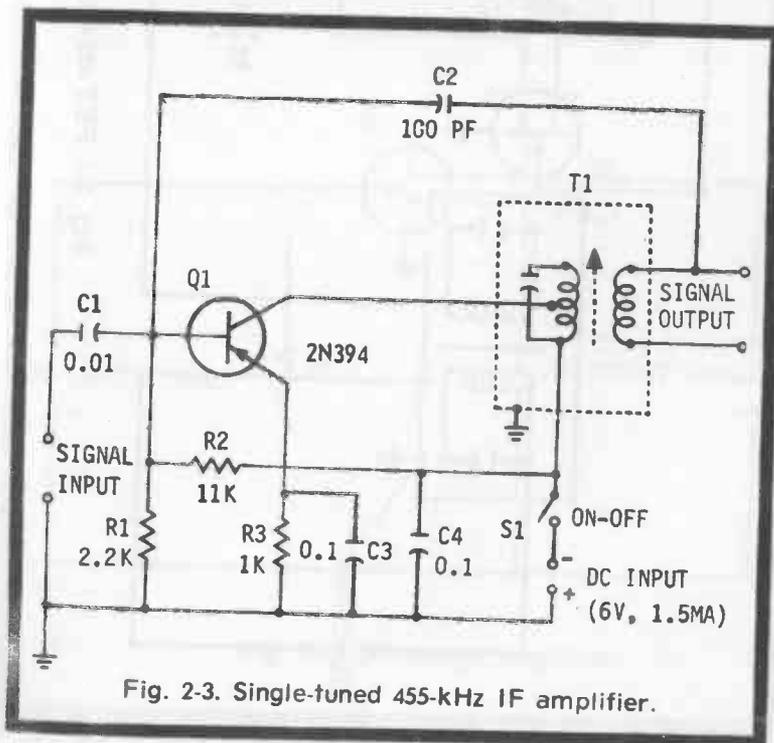


Fig. 2-3. Single-tuned 455-kHz IF amplifier.

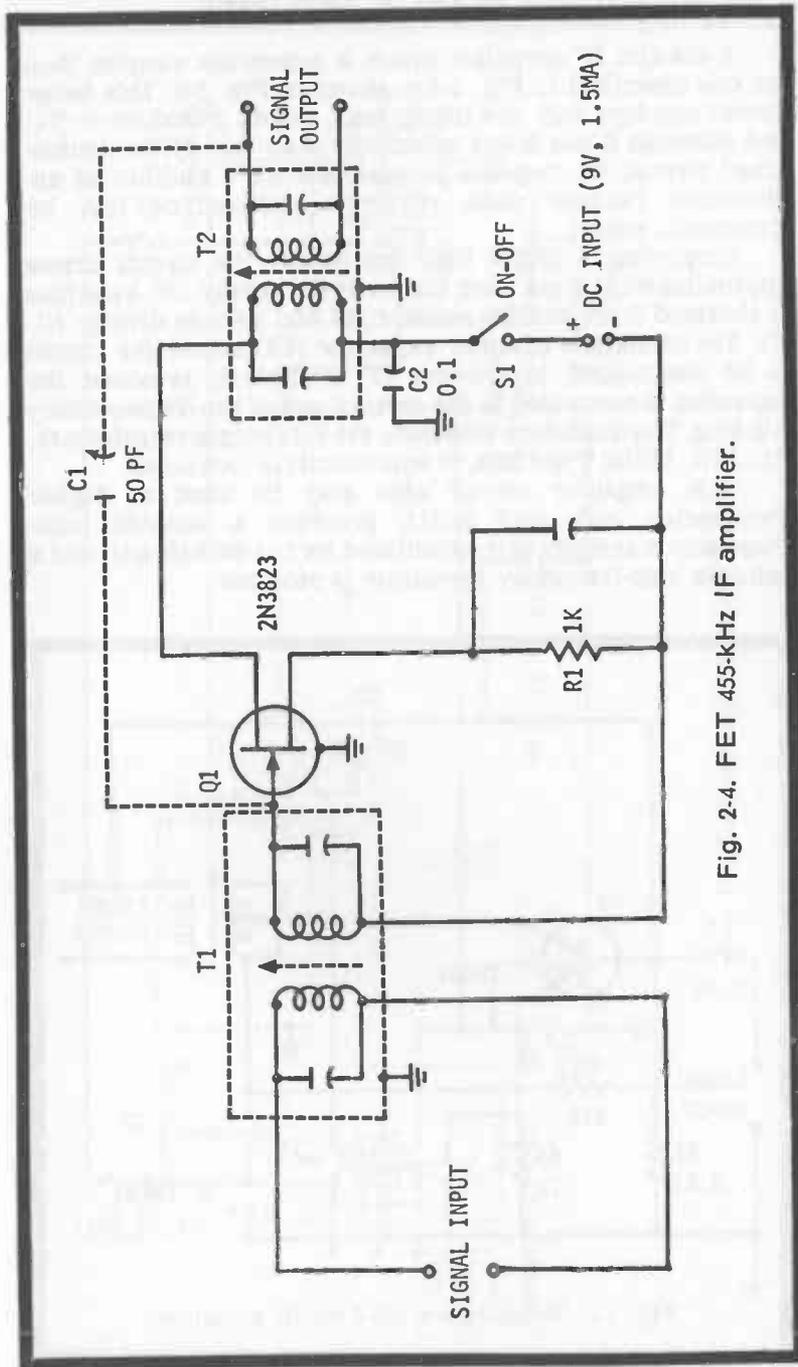


Fig. 2-4. FET 455-kHz IF amplifier.

2.4-FET 455-kHz IF AMPLIFIER

The high input impedance of field-effect transistors (typically, 1000 megohms in junction-type FETs) allows it to be used with standard IF transformers intended for tube circuits. It also permits a simpler control-electrode DC biasing scheme than the three-resistor arrangement required by the conventional transistor. Fig. 2-4 shows an FET IF amplifier.

In this circuit, employing a 2N3823 junction-type FET, the current drain is 1.5 ma from the 9-volt DC supply. DC gate bias is obtained entirely from the voltage drop produced across source resistor R1 by drain-current flow. The input and output IF transformers (T1 and T2, respectively) are regular, tube-type units (standard midget or miniature in size).

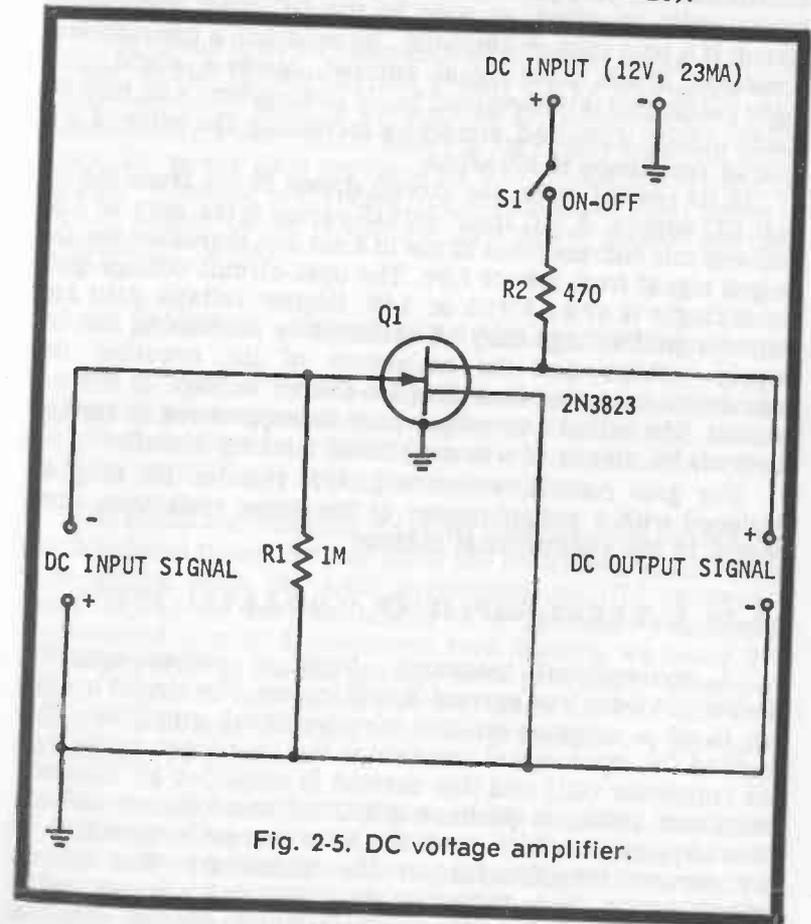


Fig. 2-5. DC voltage amplifier.

Most setups of this circuit will require no neutralization if the transformers are well shielded and if input and output wiring is separated as much as possible. If IF oscillation occurs, however, a simple neutralizing capacitor (C1) may be added to the circuit. This is a miniature 50-pf compression-type trimmer, rigidly mounted and connected to the correct end of the T2 secondary for out-of-phase feedback to the gate of Q1.

2.5—DC VOLTAGE AMPLIFIER

The relatively simple circuit in Fig. 2-5 is designed to amplify DC voltage signals. Since it employs a field-effect transistor, Q1 (Type 2N3823)—which is essentially electrostatically operated, as seen by the DC input signal—this circuit is a true voltage amplifier. By contrast, a conventional transistor draws input-signal current, however slight. The input resistance is determined by R1 (1 megohm) and may be made higher, if desired, simply by increasing the value of R1. Output resistance is 470 ohms.

In its resting state, the circuit draws 23 ma from the 12-volt DC supply. A DC input signal varied from zero to 1.5v reduces this current from 23 ma to 8 ma and increases the DC output signal from 0.6v to 7.6v. The open-circuit voltage gain accordingly is $(7.6-0.6)/1.5$ or 4.66. Higher voltage gain and output-signal voltage may be obtained by increasing the DC supply voltage and the resistance of R2, provided the maximum recommended drain-to-source voltage is not exceeded. The initial 0.6v output may be suppressed to zero, if desired, by means of a conventional bucking circuit.

For gain control, where required, resistor R1 may be replaced with a potentiometer of the same resistance, connected in the conventional manner.

2.6—DC CURRENT AMPLIFIER (LOW-LEVEL TYPE)

A conventional transistor, being a current-actuated device, provides true current amplification. The circuit in Fig. 2-6 is of a common-emitter direct-current amplifier. The applied DC input signal constitutes the base-input current of the transistor (Q1) and this current is amplified by inherent transistor action to produce a DC collector output current which approaches beta times the base current (where beta is the current amplification of the transistor). The silicon transistor (Q1, Type 2N2712) is especially well adapted to this class of service, since its static collector current (Ico) is

virtually zero (maximum $0.5 \mu\text{a}$ for the 2N2712) and since temperature has only a slight effect upon operation of silicon transistors.

The actual current gain obtained depends upon the beta of the individual transistor and upon the resistance (RL) of the load device, the gain decreasing as RL increases. For example, an input-current (Ii) change of 0-130 μa produces an output-current (Io) change of 0-30 ma when RL equals 1 ohm (this action represents a current gain of 231), whereas for the same 0-30 ma output-current change the input current must increase from zero to 195 μa when RL equals 150 ohms (this latter action represents a current gain of 154). A maximum output-current level of 30 ma has been chosen here to keep within the allowable collector power dissipation of the 2N2712 transistor.

The circuit draws a maximum of 30 ma from the 6-volt DC supply; initial resting current is only $0.5 \mu\text{a}$ maximum. The input resistance seen by the applied DC input signal varies from 4838 ohms when RL is 1 ohm to 3395 ohms when RL is 150 ohms. DC power gain provided by this circuit is a matter of interest: When RL is 150 ohms, for example, the input power is $136 \mu\text{w}$, output power is 135 mw, and power gain is 993 (approximately 30 db).

2.7—DC CURRENT AMPLIFIER (POWER TYPE)

Fig. 2-7 shows a DC amplifier circuit which delivers higher output current than the 30 ma limit of the circuit described in Fig. 2-6. The higher level circuit shown here employs an HEP 230 PNP power transistor.

The actual current gain obtained depends upon the beta of the individual transistor and upon the resistance (RL) of the load device, with the gain decreasing as RL increases. However, for the low resistances (1 to 10 ohms) commonly encountered in most high-current load devices, we found the DC output current (Io) to change from practically zero (actually, 1 ma) to 250 ma as the DC input current (Ii) was changed from zero to 10 ma. The corresponding current gain accordingly is 24.9.

The circuit draws a maximum of 250 ma from the 6-volt DC supply; resting current is only 1 ma. Higher output current levels may be obtained by replacing Q1 with a higher-powered transistor. Type 2N173, for example, has a maximum collector-current rating of 15 amperes.

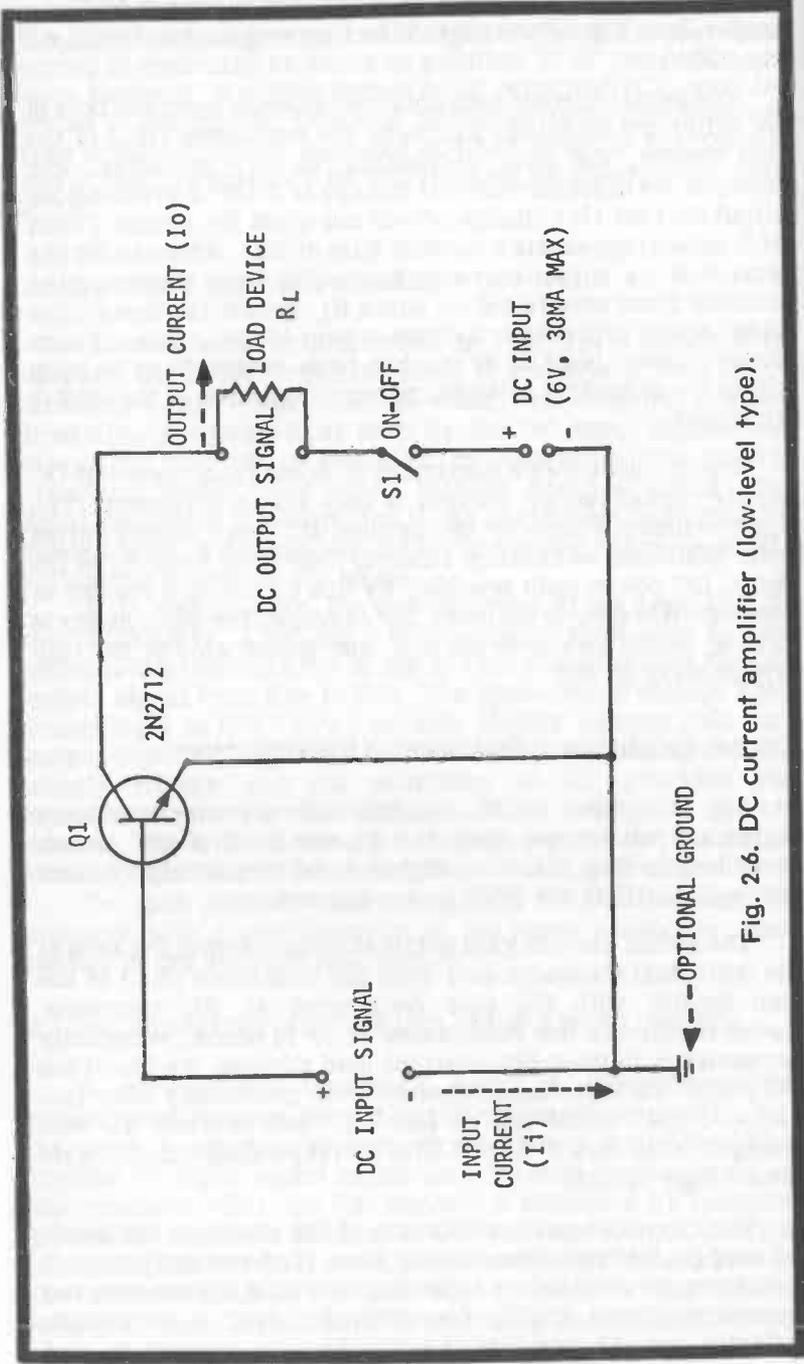


Fig. 2-6. DC current amplifier (low-level type).

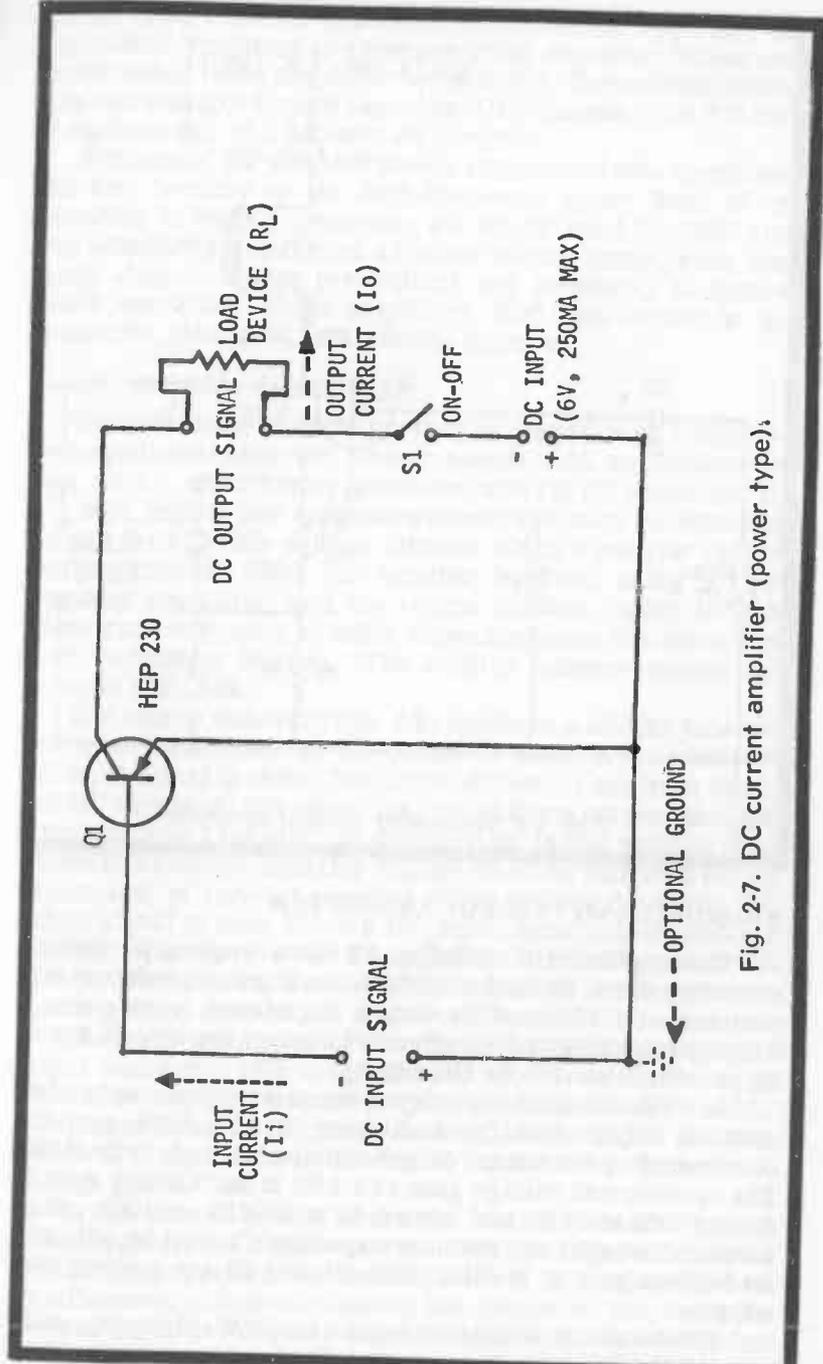


Fig. 2-7. DC current amplifier (power type).

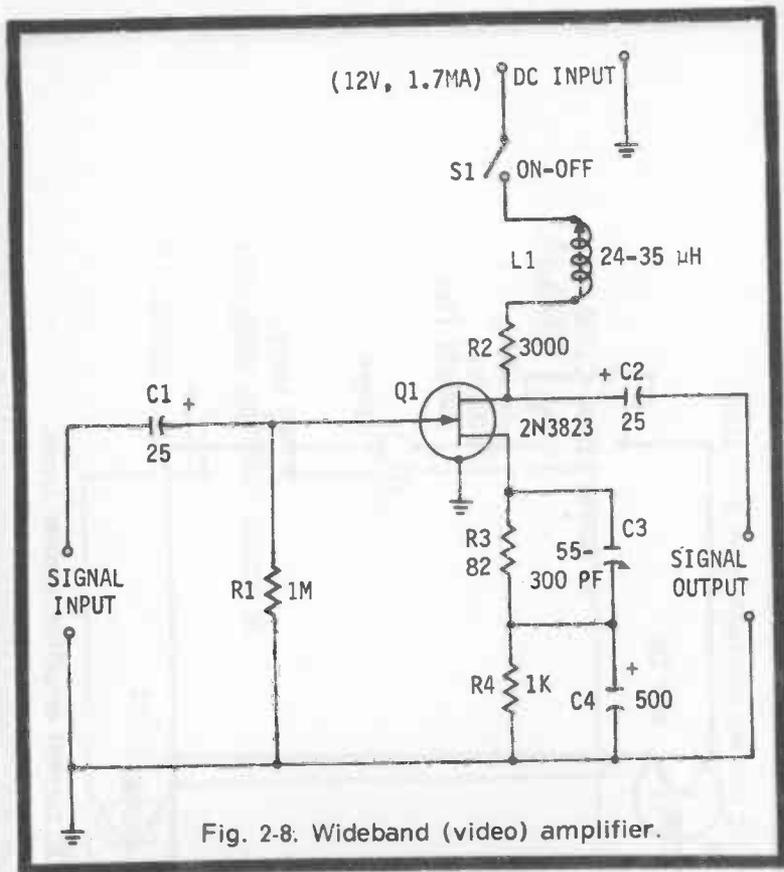


Fig. 2-8. Wideband (video) amplifier.

2.8—WIDEBAND (VIDEO) AMPLIFIER

The amplifier circuit in Fig. 2-8 has a frequency response extending from 50 Hz to 10 MHz. Its input impedance is 1 megohm at 1 kHz and its output impedance is 3000 ohms. Employing a 2N3823 field-effect transistor, the circuit draws 1.7 ma from the 12-volt DC supply.

At 1 kHz the maximum input-signal amplitude before the onset of output-signal peak clipping is 0.2v RMS, and the corresponding maximum output-signal amplitude is 2v RMS. The open-circuit voltage gain at 1 kHz is 10. Voltage gain is down 2.5 db at 50 Hz and down 6 db at 10 MHz. Initially, slug-tuned inductor L1 and trimmer capacitor C3 must be adjusted for highest gain at 10 MHz. Both L1 and C3 are screwdriver adjusted.

The miniature 24-35- μ h inductor, L1 (J. W. Miller No. 4508, or equivalent), is employed as a peaking coil, and the 55-300-pf

midget mica trimmer capacitor, C3 (Elmenco Type 427, or equivalent), functions as a compensating capacitor. These are conventional video amplifier components. The subminiature, high-value source bypass capacitor, C4 (Sprague Type TE-106, or equivalent), is a 500-mfd electrolytic.

Because of the wide frequency response of this amplifier, and also because of the high-frequency upper limit of its operation in the RF spectrum, all wiring must be rigid and kept widely separated from adjacent wiring, components, and metal chassis. These precautions are necessary to insure stable operation. Video amplifiers find application in instruments, television, and control systems.

2.9—DC SOURCE FOLLOWER

The usefulness of the follower-type amplifier as a high-to-low impedance converter, already noted for AC applications in Figs. 1-6, 1-7, and 1-9, may be utilized also for DC operation. In this way, high-to-low resistance conversion may be obtained without the extreme voltage division which a resistor circuit would introduce. Both the emitter follower, using a conventional transistor, and the source follower, using a field-effect transistor, may be used, depending upon the amount of input resistance desired. (The emitter follower circuit appears in Fig. 2-10.)

The source follower (Fig. 2-9) employs a 2N3823 junction field-effect transistor. In the quiescent state (i. e., when the DC input signal is zero), the circuit draws 7.5 ma from the 3-volt DC supply. At this point, the flow of drain current through source (output) resistor R3 produces an initial voltage drop across this resistor, and this voltage must be balanced out by adjustment of zero-set rheostat R2 to reduce the static DC output signal to zero. When a DC input signal subsequently is applied, with the gate of Q1 negative, it reduces the drain current and causes the DC output signal to rise proportionately from zero in the negative direction.

The input resistance of the circuit is 1 megohm and the output resistance 1000 ohms. If desired, however, the input resistance may be increased simply by raising the value of R1. A DC input signal of 1.5v produces a DC output signal of 1.25v, while reducing the drain current to 5.2 ma. The open-circuit voltage gain is 0.83.

In this circuit, S1 and S2 are the two sections of a double-pole, single-throw toggle switch. This arrangement allows both the DC supply and the bucking battery, B1, to be switched simultaneously, thus eliminating the danger of one being on without the other. The zero-adjuster control, R2, is a 5000-ohm wirewound rheostat.

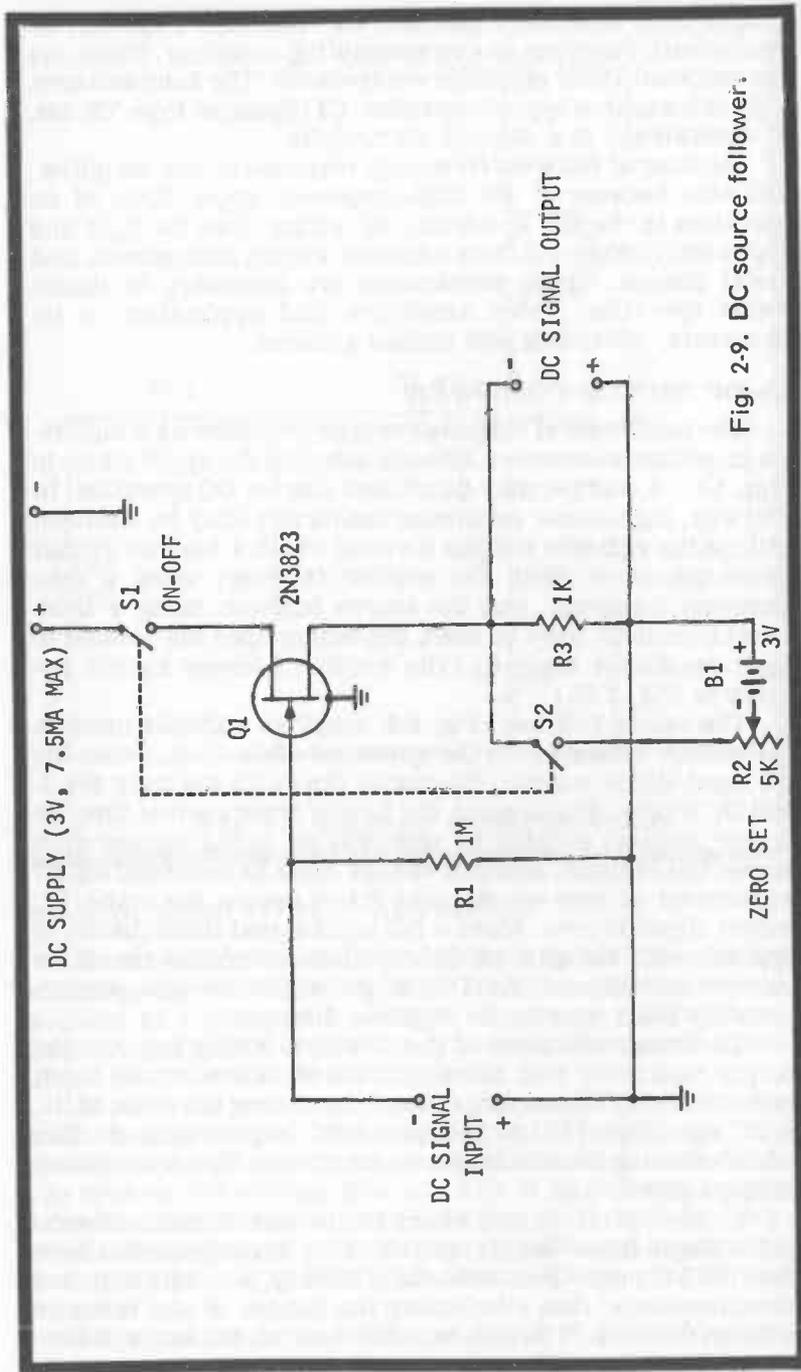


Fig. 2-9. DC source follower.

2-10-DC EMITTER FOLLOWER

When the very high input resistance of the FET DC source follower (Fig. 2-9) is not needed, a simpler DC emitter follower, based upon a conventional transistor, may be used. Fig. 2-10 shows a DC emitter-follower circuit employing a 2N2712 silicon transistor.

Because the static collector current (I_{co}) of this silicon transistor is so tiny (less than $0.5 \mu a$), the static voltage drop across emitter (output) resistor R2 is inconsequential; therefore, the zero-set circuit required by the FET source follower (Fig. 2-9) is not needed here. The initial DC signal output, consequently, is zero when the input signal is zero.

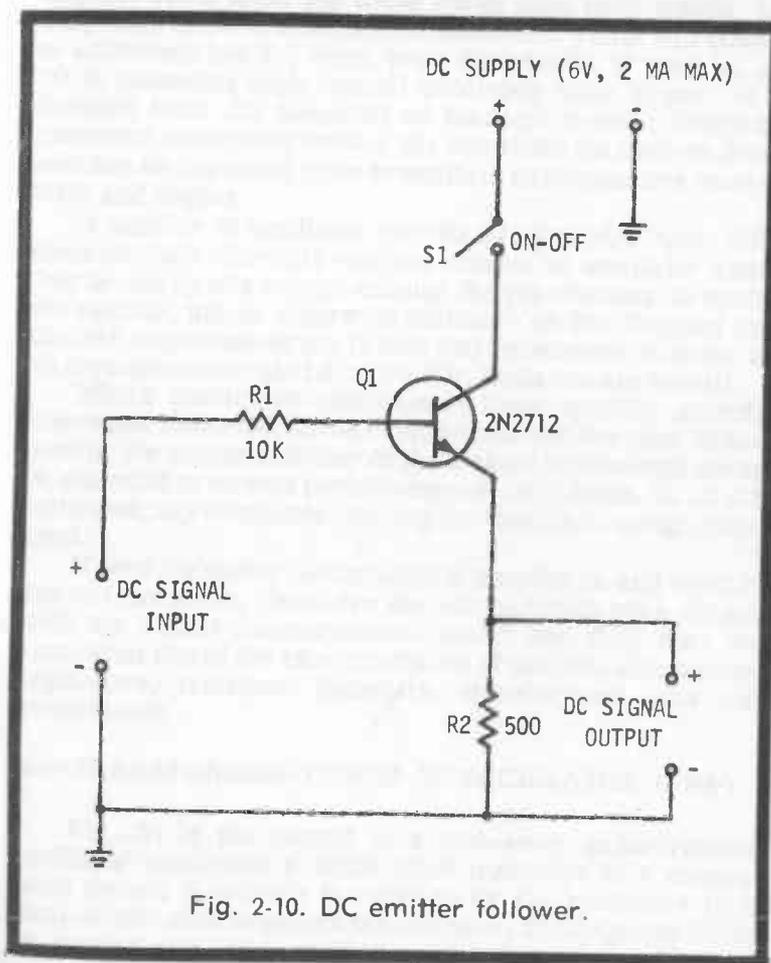


Fig. 2-10. DC emitter follower.

When a DC input signal is applied, with the base of Q1 positive, the DC output signal rises proportionately from zero in the positive direction. A DC input signal of 1.5v produces a DC output signal of 0.8v. The open-circuit voltage gain is 0.53. At this 1.5-volt point, the input resistance is greater than 80,000 ohms and the output resistance is 500 ohms. Under these maximum-signal conditions, the circuit draws 2 ma from the 6-volt DC supply. The 10,000-ohm input resistor (R1) is a protective resistor.

Chapter 3

Oscillators

The transistor oscillator is attractive in two important respects aside from the usual small size, light weight, and simplicity found in transistorized devices: (1) It will operate on extremely low DC input, even microwatts, if need be, and (2) it possesses high overall efficiency (due in part to its freedom from the necessity of filament power). Originally restricted to low frequencies, the transistor (in modern dress) now can be depended upon to oscillate at frequencies up to 400 MHz and higher.

A number of oscillator circuits are included here. While these circuits illustrate various classes of oscillator action, they do not by any means exhaust the possibilities. In each of the circuits, unless otherwise indicated on the diagram or in the text, capacitances are in mfd and resistances in ohms, and all capacitors are rated at 25 DCWV. Resistors are $\frac{1}{2}$ watt.

Where particular components from specific manufacturers are indicated, these components are the ones actually used by the author and they or their exact equivalents seem to be essential to correct performance of the circuit. In all other instances, any component having the specified ratings may be used.

Where frequency, output-signal amplitude, and operating current are given, these are the values which were obtained with the author's experimental model and they may vary somewhat due to the characteristics of individual transistors, capacitors, resistors, inductors, transformers, and other components.

3.1—TRANSFORMER-TUNED AF OSCILLATOR (PNP)

Fig. 3-1 is the circuit of a sine-wave audio-frequency oscillator employing a 2N107 PNP transistor in a common-base circuit. Feedback is supplied by the miniature 10,000-ohm-to-2000-ohm stepdown transformer, T1 (Argonne AR-109, or equivalent). For oscillation, the transformer must be

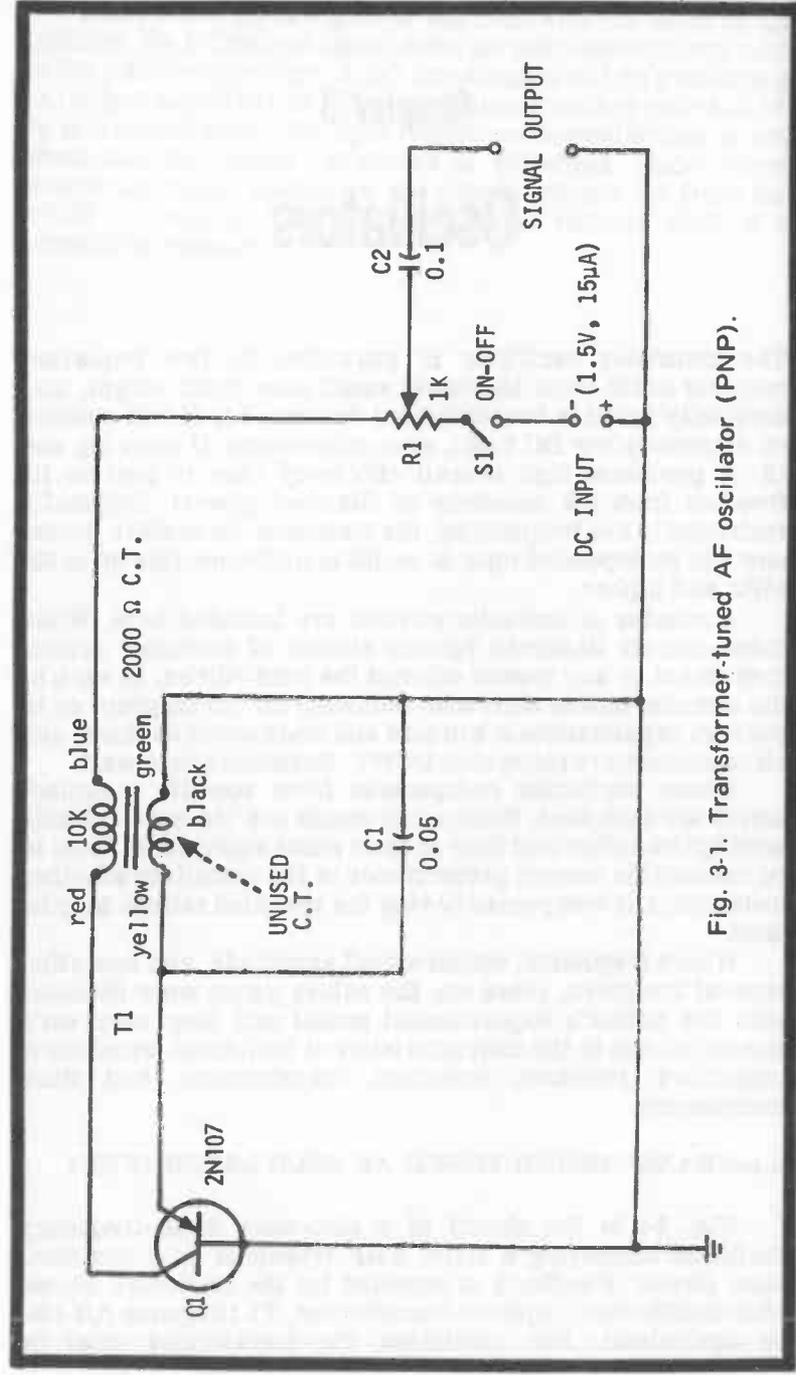


Fig. 3-1. Transformer-tuned AF oscillator (PNP).

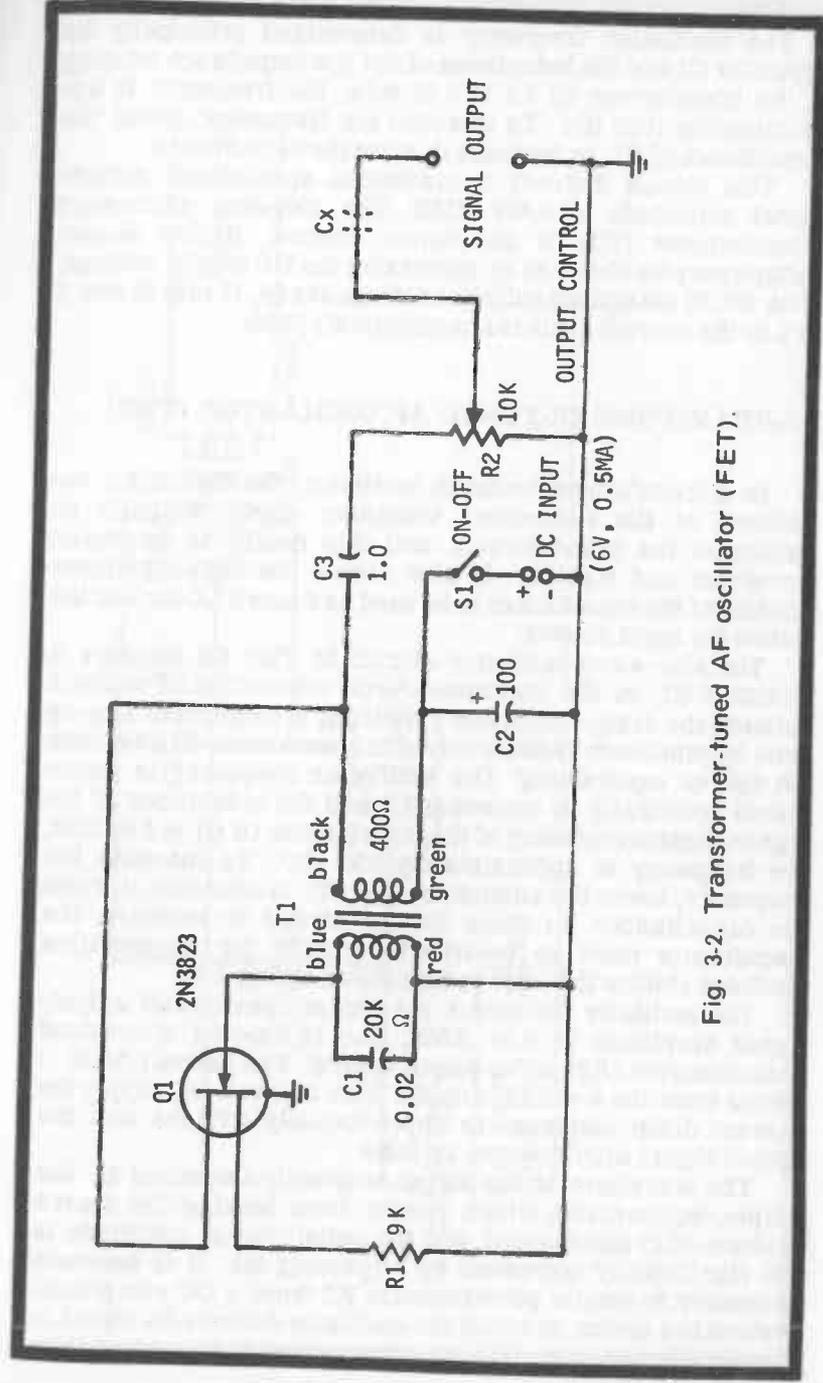


Fig. 3-2. Transformer-tuned AF oscillator (FET).

polarized correctly, as indicated by the color coding in Fig. 3-1. The oscillation frequency is determined principally by capacitor C1 and the inductance of the low-impedance winding of the transformer (if C1 is 0.05 mfd, the frequency is approximately 1000 Hz). To increase the frequency, lower the capacitance of C1; to decrease it, raise the capacitance.

This circuit delivers a maximum open-circuit output-signal amplitude of 0.65v RMS. The 1000-ohm wirewound potentiometer (R1) is the output control. Higher output voltage may be obtained by increasing the DC supply voltage. (The 2N107 maximum collector ratings are 6v, 10 ma, 50 mw.) At 1.5v the current drain is approximately 15 μ a.

3.2—TRANSFORMER-TUNED AF OSCILLATOR (FET)

In a transformer-feedback oscillator, the high input impedance of the field-effect transistor offers virtually no loading of the tuned circuit, and this results in improved waveform and stability. It also allows the high-impedance winding of the transformer to be used as a tuned LC tank in the transistor input circuit.

The sine-wave oscillator circuit in Fig. 3-2 employs a 2N3823 FET in the common-source connection. Feedback between the drain-output and gate-input is supplied in step-up ratio by miniature 20,000-to-400-ohm transformer T1 (Argonne AR-105, or equivalent). The oscillation frequency is determined principally by capacitor C1 and the inductance of the high-impedance winding of the transformer (if C1 is 0.02 mfd, the frequency is approximately 150 Hz). To increase the frequency, lower the capacitance of C1; to decrease it, raise the capacitance). In order for the circuit to oscillate, the transformer must be polarized correctly for regenerative feedback (follow the color coding shown in Fig. 3-2).

The oscillator delivers a maximum open-circuit output-signal amplitude of 0.6v RMS. The 10,000-ohm wirewound potentiometer (R2) is the output control. The current drain is 0.5 ma from the 6-volt DC supply. With a 12-volt DC supply the current drain increases to approximately 0.75 ma and the output-signal amplitude to 1v RMS.

The waveform of the signal is greatly improved by the ample degeneration which results from leaving the source resistor (R1) unbypassed, and the output signal amplitude is not significantly increased by bypassing R1. If it becomes necessary to isolate potentiometer R2 from a DC component present in a device to which the oscillator delivers its signal, a 1- or 2-mfd capacitor (CX) may be inserted in the output line.

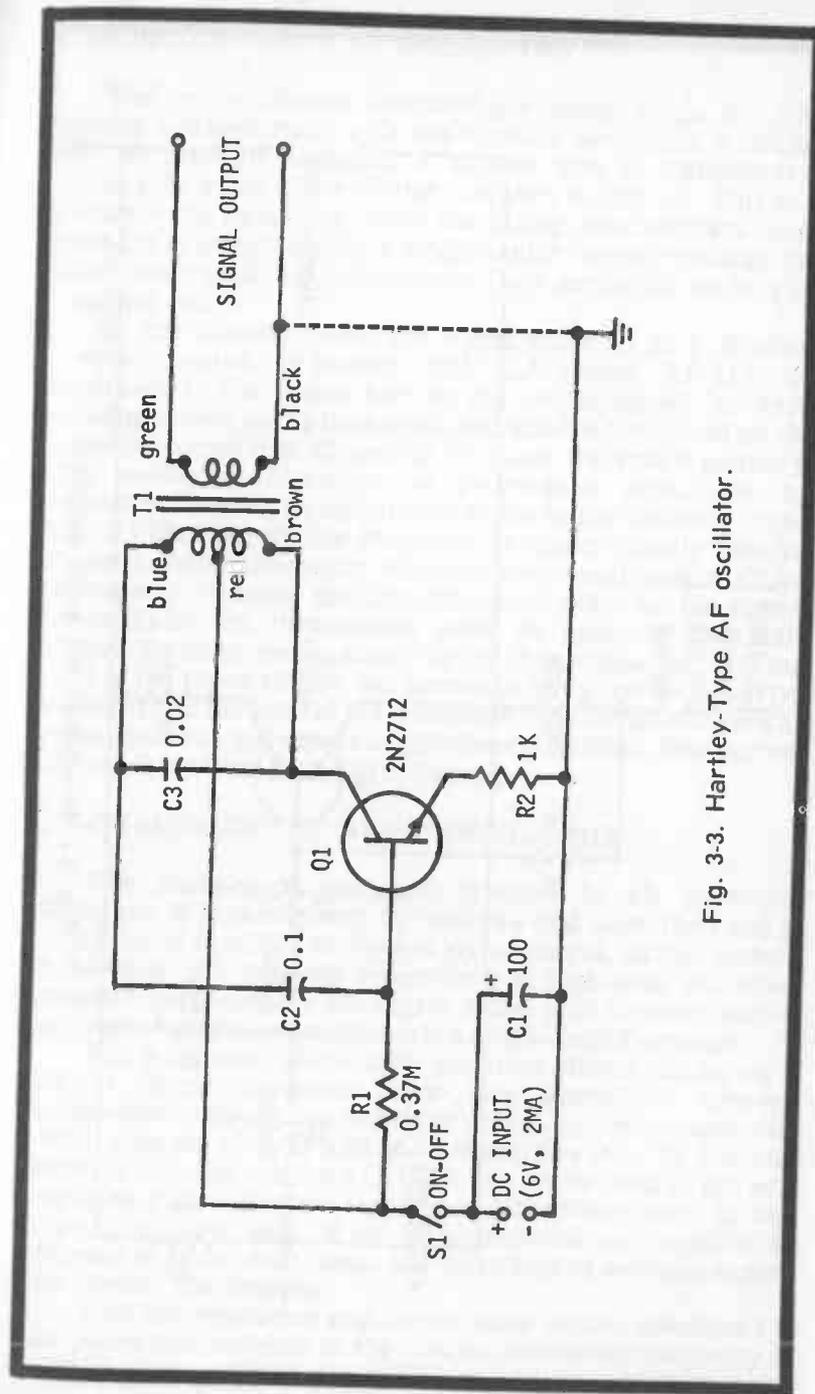


Fig. 3-3. Hartley-Type AF oscillator.

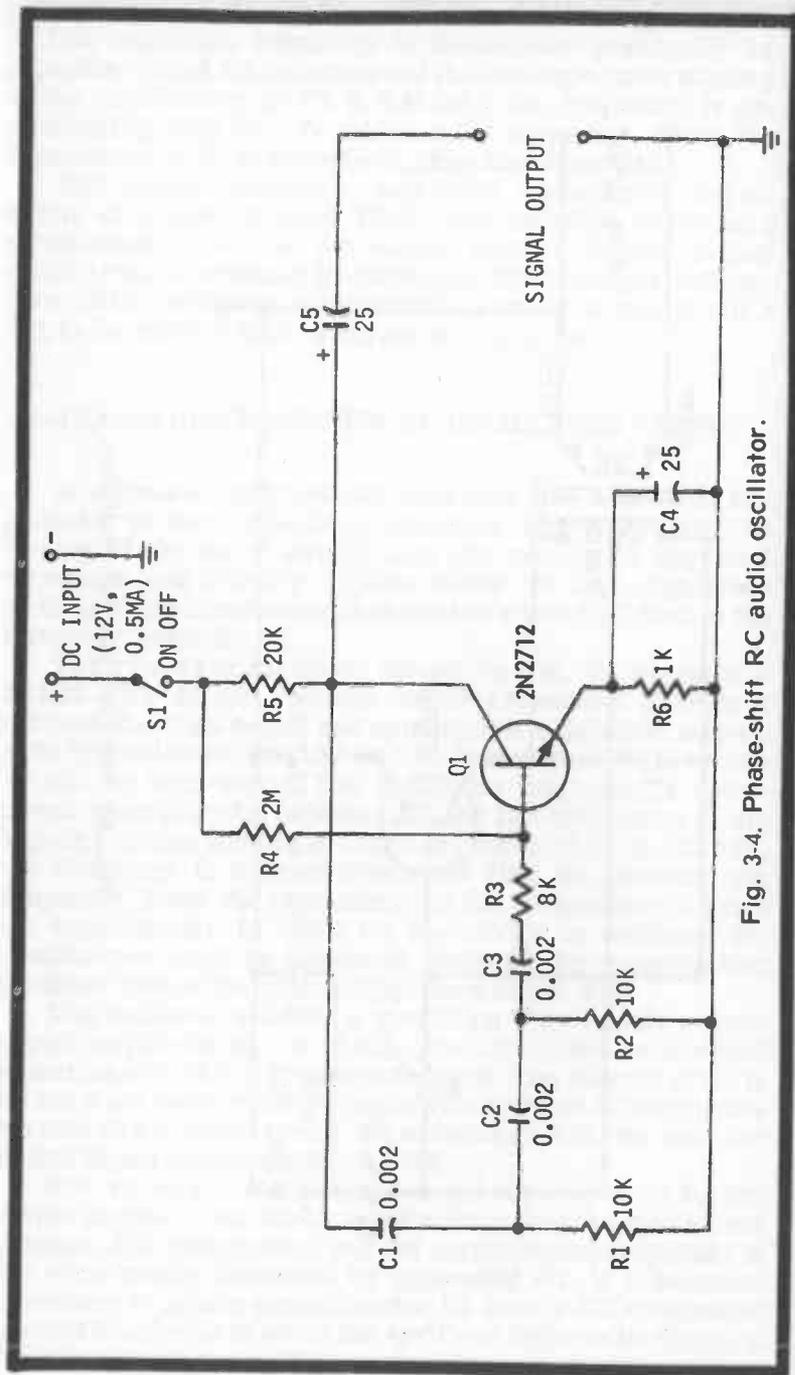


Fig. 3-4. Phase-shift RC audio oscillator.

3.3—HARTLEY-TYPE AF OSCILLATOR

The two oscillators described previously (Figs. 3-1, 3-2) employ a transformer with one winding serving as a tickler coil for positive feedback. A second type of transformer-tuned, sine-wave, AF oscillator is shown in Fig. 3-3. This is a Hartley-type circuit, in which the tuning and feedback functions are accomplished by a single center-tapped winding; the other winding of the transformer then serves as an output-coupling coil.

In this circuit, miniature transformer T1 is a 500-ohm (center-tapped)-to-30-ohm unit (Argonne AR-117, or equivalent). The upper half of the center-tapped primary winding serves as the base-input coil and the lower half as the collector-output coil. Capacitor C3 tunes the entire primary. The oscillation frequency is determined principally by capacitor C3 and the inductance of the entire primary winding. If C3 is 0.02 mfd, the frequency is approximately 2000 Hz. To increase the frequency, decrease the capacitance of C3; to decrease it, increase the capacitance. In order for the circuit to oscillate, the transformer must be polarized correctly (follow the color coding shown in Fig. 3-3). Capacitor C2 is not part of the tuned circuit, but serves to block the collector DC voltage from the base of the transistor. The circuit delivers an open-circuit output-signal amplitude of 0.8v RMS. The current drain is 2 ma from the 6-volt DC supply.

3.4—PHASE-SHIFT RC AUDIO OSCILLATOR

The phase-shift oscillator, familiar to all electronic designers, is tuned entirely by resistors and capacitors and is noted for its excellent sine-wave signal output. In this circuit, a 3-legged RC network resembling a high-pass RC filter supplies the necessary 180-degree phase shift between output and input to produce oscillation in a single amplifier stage.

Fig. 3-4 shows a phase-shift oscillator circuit employing a 2N2712 silicon transistor. Here, the phase-shift network (connected between the collector-output and base-input circuits) consists of three 0.002-mfd capacitors (C1, C2, C3) and three 10,000-ohm resistors (R1, R2, the series total of R3, and the base-ground internal resistance of the transistor). In this type of network, each of the three identical legs supplies 60 degrees of phase shift; thus, the three legs in cascade supply the needed 180 degrees.

With the resistance and capacitance values specified for the phase-shift network in Fig. 3-4, the oscillation frequency is 2000 Hz. (Some adjustment of R3 is necessary according to the

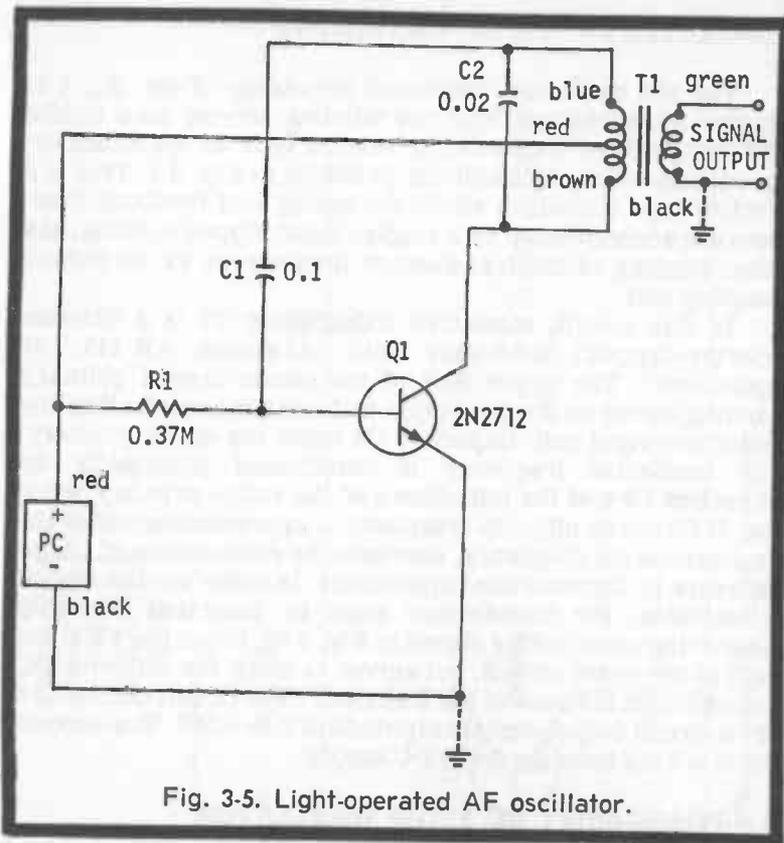


Fig. 3-5. Light-operated AF oscillator.

characteristics of the transistor in use and this will affect frequency.) For any phase-shift network of the type shown here, the oscillation frequency f is equal approximately to $10^6 / (20 RC)$, where f is in Hz, R in ohms, and C in mfd. The frequency may be changed by simultaneously changing all three capacitances, all three resistances, or both, by the same amount. Increasing R or C decreases f , and vice versa.

The circuit delivers an open-circuit output-signal amplitude of 3v RMS. The current drain is 0.5 ma from the 12-volt DC supply.

3.5—LIGHT-POWERED AF OSCILLATOR

The high efficiency of the transistor and its ability to operate at a very low DC collector voltage make possible the operation of oscillators from low-voltage DC generators. Such generators include photocells and thermopiles. Fig. 3-5 shows the circuit of a simple Hartley-type, sine-wave audio oscillator

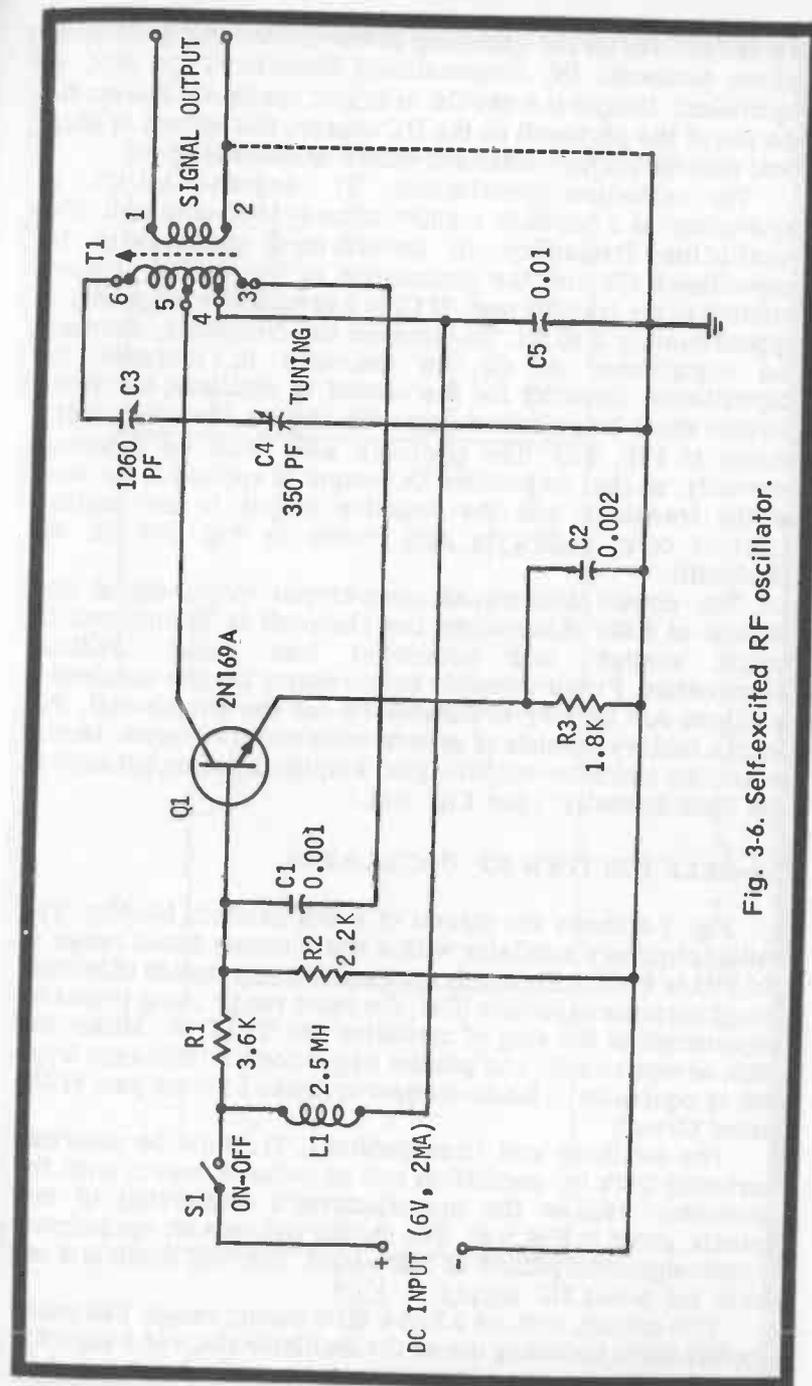


Fig. 3-6. Self-excited RF oscillator.

which receives its DC operating power from a self-generating silicon photocell, PC (International Rectifier Type S1M, or equivalent: Output is 0.55v DC in bright sunlight). Except for the use of the photocell as the DC source, this circuit is identical with the Hartley oscillator shown earlier in Fig. 3-3.

The miniature transformer, T1 (Argonne AR-117, or equivalent) is a 500-ohm (center-tapped)-to-30-ohm unit. The oscillation frequency is determined principally by capacitance C2 and the inductance of the entire primary winding of the transformer. If C2 is 0.02 mfd, the frequency is approximately 2000 Hz. To increase the frequency, decrease the capacitance of C2; to decrease it, increase the capacitance. In order for the circuit to oscillate, the transformer must be polarized correctly (follow the color coding shown in Fig. 3-5). The photocell also must be polarized correctly, so that its positive DC output is applied to the base of the transistor and the negative output to the emitter (correct color coding is also shown in Fig. 3-5 for the photocell).

The circuit delivers an open-circuit output-signal amplitude of 0.09v RMS when the photocell is illuminated by bright sunlight and somewhat less under artificial illumination. Proportionately higher output may be obtained if a silicon sun battery is substituted for the simple cell, PC. Such a battery consists of several cells wired in series. In any event, the oscillator output-signal amplitude is proportional to the light intensity. (See Fig. 3-14.)

3.6—SELF-EXCITED RF OSCILLATOR

Fig. 3-6 shows the circuit of a self-excited, Hartley-type radio-frequency oscillator with a continuously-tuned range of 500 kHz to 4 MHz. Tuning is accomplished by means of midget 350-pf variable capacitor (C4), the exact range being preset by adjustment of the slug of oscillator coil T1 (J. W. Miller No. 2023, or equivalent) and padder capacitor C3 (Elmenco Type 308, or equivalent). Radio-frequency choke L1 is not part of the tuned circuit.

The oscillator coil (transformer), T1, must be polarized correctly both for oscillation and impedance match with the transistor (follow the manufacturer's numbering of terminals, given in Fig. 3-6). The circuit delivers an open-circuit output-signal amplitude of 0.65v RMS. Current drain is 2 ma from the 6-volt DC supply.

This circuit, with its 0.5-to-4-MHz tuning range, has many applications, including use as the oscillator stage of a superhet receiver. Its operating frequency may easily be changed by

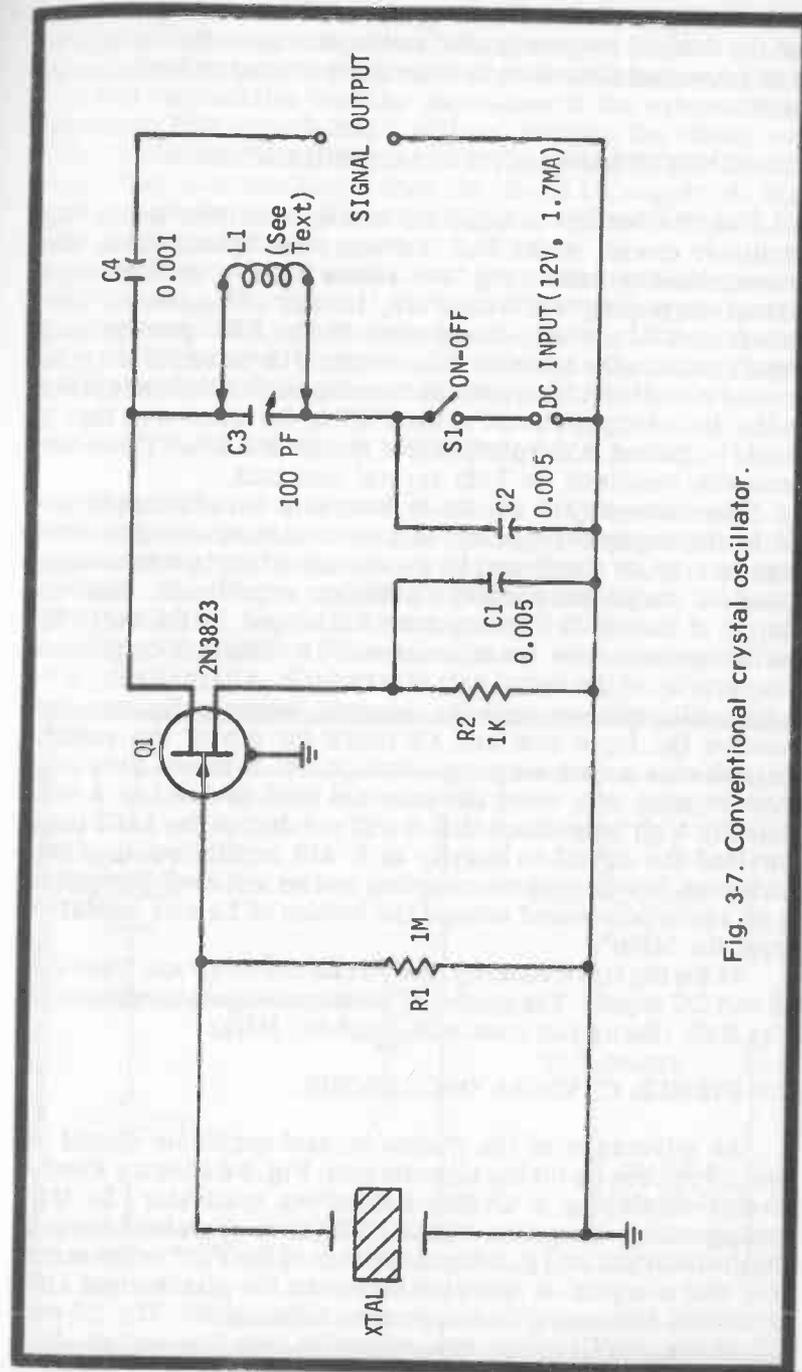


Fig. 3-7. Conventional crystal oscillator.

substituting for T1 a coil having an inductance rating suitable for the desired frequency. Such coils, for use with tuning C4, may be wound according to instructions found in radio handbooks.

3.7—CONVENTIONAL CRYSTAL OSCILLATOR

The field-effect transistor makes possible a crystal oscillator circuit which has the same configuration as the vacuum-tube version. Fig. 3-7 shows such a conventional circuit employing a 2N3823 FET. In this arrangement, the quartz crystal (XTAL) is operated in the FET gate/ground input circuit in the same way that it would be placed in the grid/ground circuit of a tube, and the tuned tank (L1C3) is operated in the drain/output circuit of the FET in the same way that it would be placed in the plate/output circuit of a tube. These are favorable positions for both crystal and tank.

The value of L1 is chosen to tune with variable capacitor C3 to the crystal frequency. Either a commercial inductor may be used, or a coil may be wound according to instructions found in radio handbooks. Oscillator adjustment consists simply of tuning C3 for maximum RF output, as indicated by an RF vacuum-tube (or transistorized) voltmeter connected temporarily to the signal output terminals. Alternatively, a 0-5 DC milliammeter may be inserted temporarily into the positive DC input line and C3 tuned for dip of the meter. Capacitance output coupling (through C4) is shown here and may be used only when the external load device has a sufficiently high impedance that it will not detune the L1C3 tank nor load the circuit so heavily as to kill oscillation. In other instances, low-impedance coupling can be achieved through a 2- or 3-turn coil wound around the bottom of L1 and insulated from the latter.

At the crystal frequency, the circuit draws 1.7 ma from the 12-volt DC supply. The open-circuit output-signal amplitude is 1.1v RMS (the author's test was made at 7 MHz).

3.8—PIERCE CRYSTAL OSCILLATOR

An advantage of the Pierce crystal oscillator circuit is that it requires no tuning adjustments. Fig. 3-8 shows a Pierce circuit employing a 2N3823 field-effect transistor. In this arrangement, the quartz crystal (XTAL) is operated between the drain/output and gate/input sections of the FET in the same way that a crystal is operated between the plate/output and grid/input sections of the equivalent tube circuit. The 2.5-mh RF choke (RFC1) does not ordinarily tune the circuit, but

serves only to keep RF energy out of the DC supply. The circuit oscillates as soon as switch S1 is closed.

Capacitive output coupling is provided through capacitor C1, and necessitates that the impedance of the external load device be high enough that it will not overload the circuit and kill oscillation. When the circuit is oscillating at the crystal frequency, it draws 2.3 ma from the 12-volt DC supply. At that point, the open-circuit RF output-signal amplitude is 6.2v RMS (the author's test was made at 7 MHz).

The Pierce circuit oscillates at the fundamental frequency of the crystal. Therefore, if you have a harmonic-type crystal, oscillation will take place at the basic frequency of the crystal, not at the labeled (harmonic) frequency. Also, the Pierce oscillator demands a very active crystal.

3.9—MULTI-FREQUENCY CRYSTAL OSCILLATOR

The utility of a crystal oscillator may be extended by including a crystal-switching section. The capability enables signals at several frequencies to be generated, all with crystal stability. If a Pierce oscillator circuit is employed, the complication of LC tuned tanks is avoided.

Fig. 3-9 shows a Pierce oscillator circuit adapted for

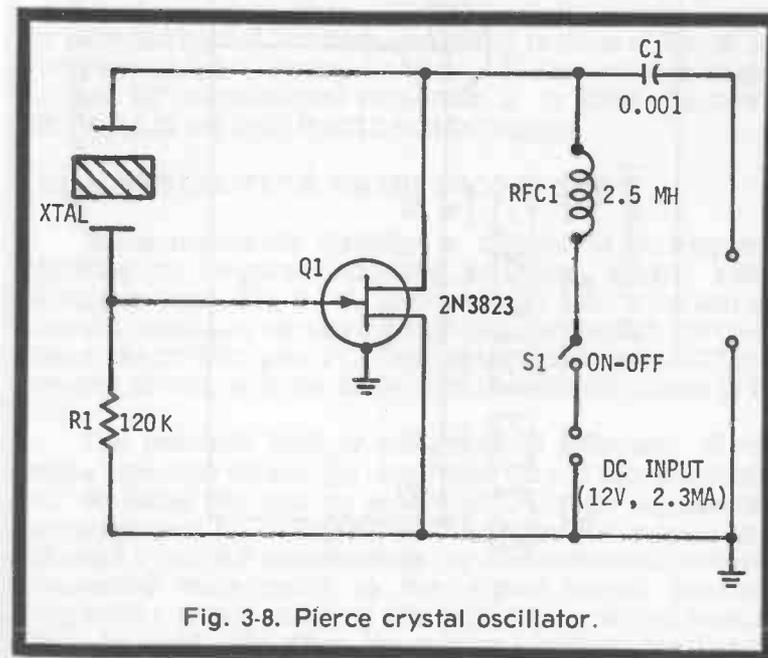


Fig. 3-8. Pierce crystal oscillator.

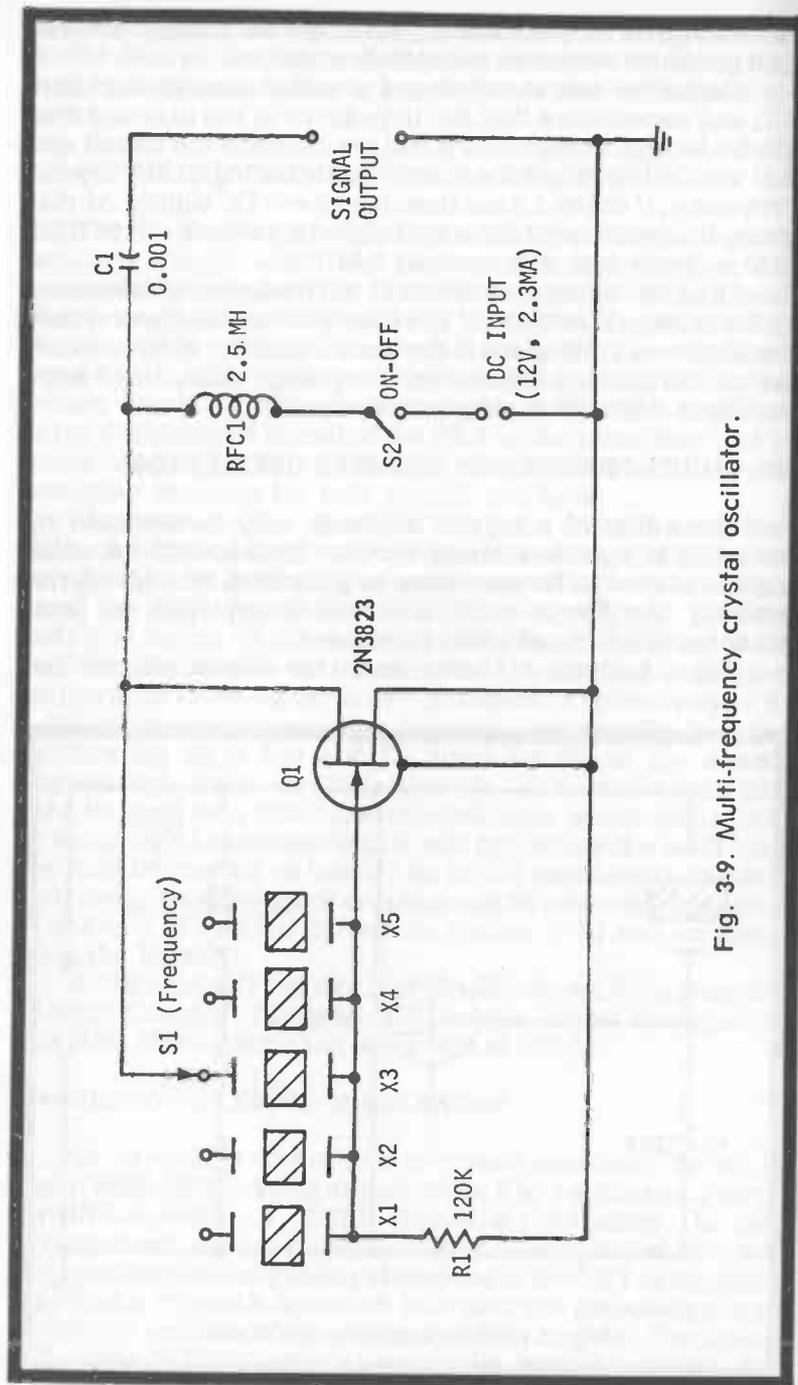


Fig. 3-9. Multi-frequency crystal oscillator.

crystal switching. Except for the addition of several crystals and the switch (S1), the circuit is identical with the basic Pierce oscillator described previously. Switch S1 is a single-gang, nonshorting, rotary selector switch having as many positions as there are crystals (X1, X2, etc.). While five crystals are shown here, any number may be used.

The RF output-signal amplitude will vary somewhat, depending upon frequency and especially upon crystal activity, but it will be on the order of 6.2v RMS. Current drain likewise will vary somewhat, but will be on the order of 2.3 ma from the 12-volt DC supply. See Section 3.8 for additional pointers on the Pierce oscillator.

3.10—SELF-EXCITED 100-KHZ OSCILLATOR

Fig. 3-10 is the schematic of a self-excited 100-kHz oscillator suitable for many applications which do not demand the high stability of a crystal oscillator. This is a version of the Colpitts oscillator circuit, which is characterized by the split-capacitor (C3-C4) tuning of the frequency-determining tank (L1-C3-C4).

The oscillator frequency may be set exactly to 100 kHz by adjusting the tuning slug of the 0.5 to 5-mh inductor, L1 (J. W. Miller No. 6313, or equivalent) while zero beating against a standard-frequency station WWV signal. RF output coupling is provided by coil L2 which consists of 15 turns of No. 22 DCC wire wound tightly around L1. This oscillator delivers an open-circuit RF output-signal amplitude of 1v RMS. Its current drain is 1.25 ma from the 22.5-volt DC supply.

3.11—CRYSTAL-TYPE 100-KHZ OSCILLATOR

When maximum stability is desired in a frequency-spotting or frequency-standard oscillator, crystal control should be used. Fig. 3-11 is the circuit for such a crystal-type 100-kHz oscillator employing a 2N3819 field-effect transistor. Here, the 100-kHz quartz (XTAL) is operated in the FET gate/ground circuit, as in the basic crystal oscillator shown in Fig. 3-7.

The resonant tank is composed of 0.002-mfd silvered mica capacitor C3 and the slug-tuned 0.3 to 3-mh inductor, L1 (J. W. Miller No. 6318, or equivalent). Circuit adjustment is accomplished by tuning L1 for maximum RF output, as indicated by an RF vacuum-tube (or transistorized) voltmeter connected temporarily to the signal output terminals. Capacitive output coupling (through C4) is shown here and may be used only when the external load device has suf-

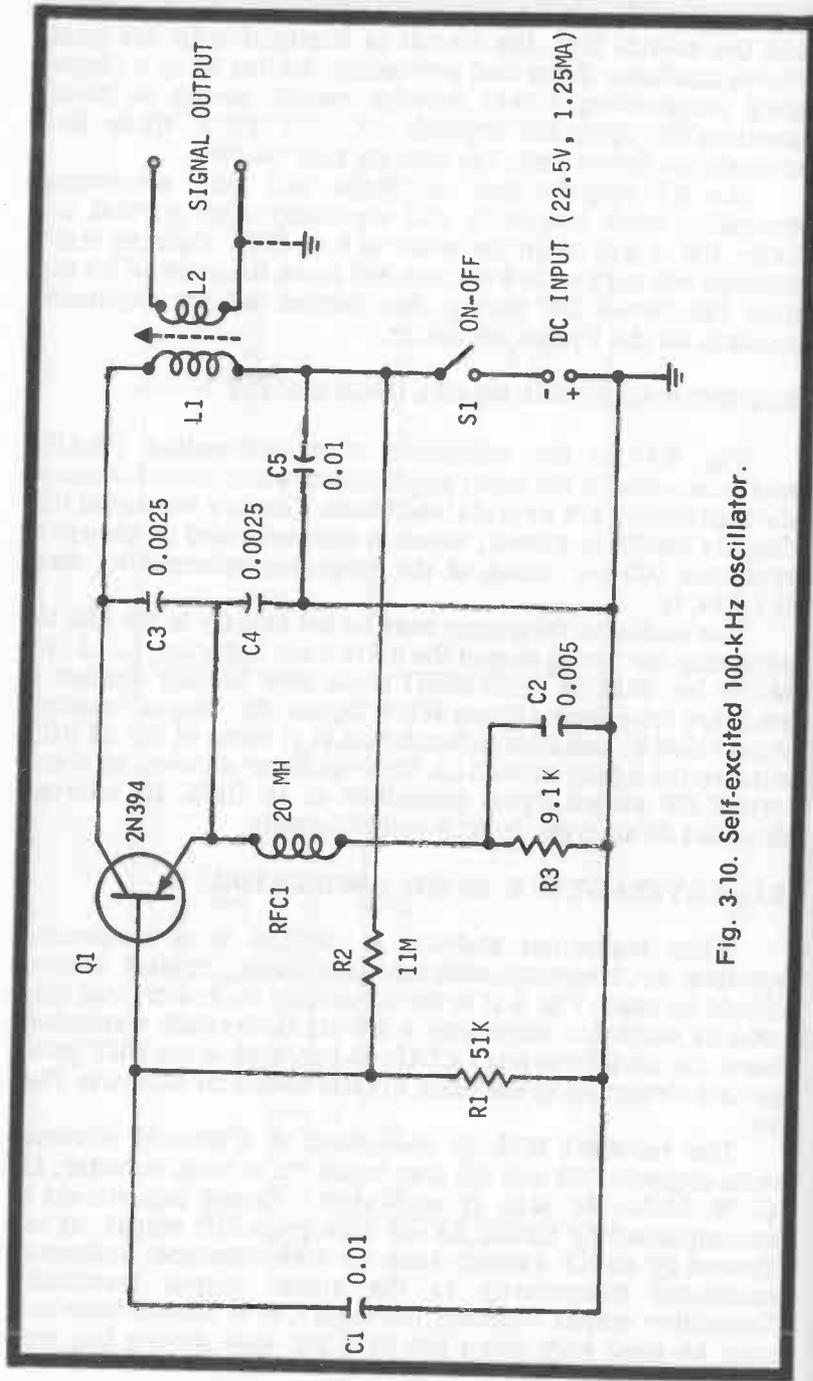


Fig. 3-10. Self-excited 100-kHz oscillator.

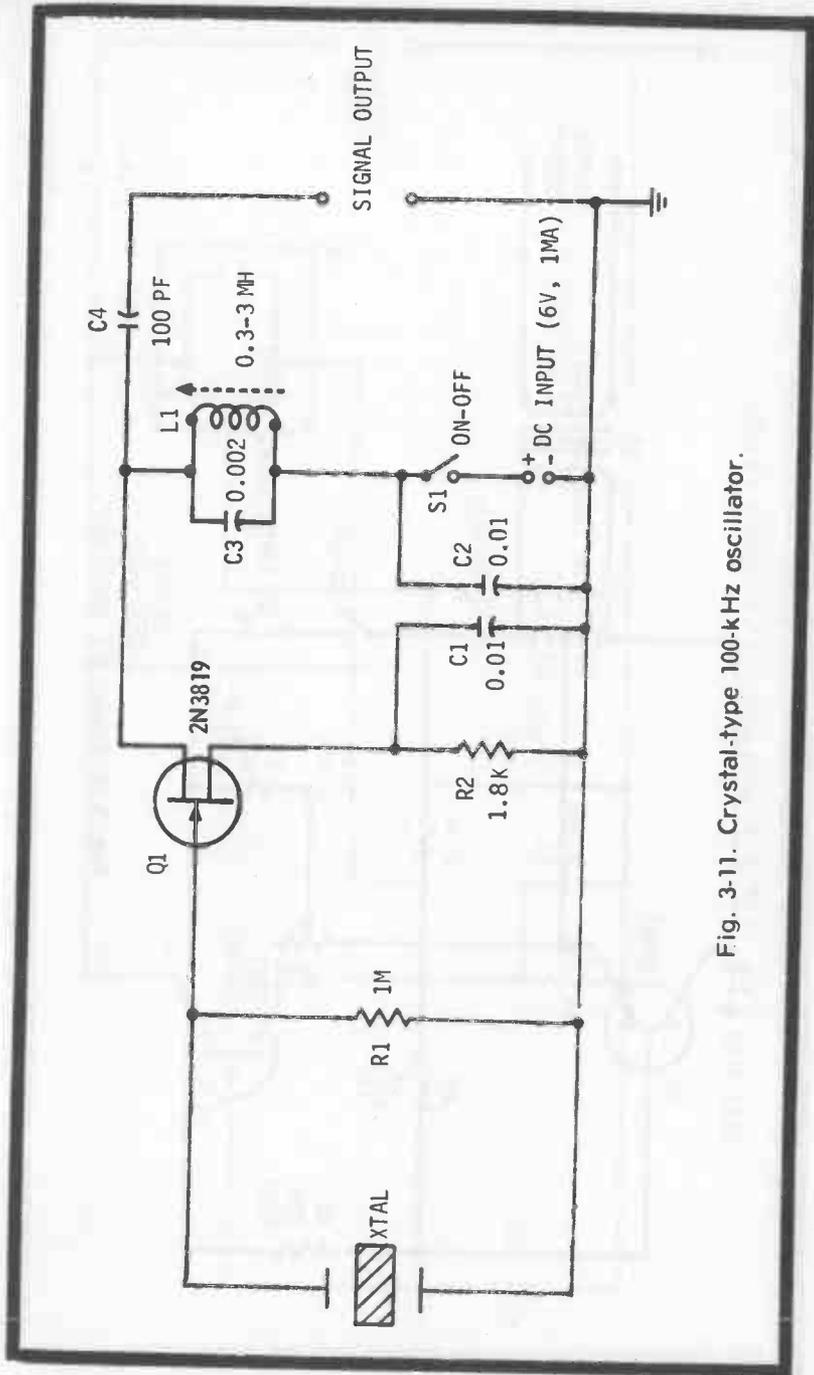


Fig. 3-11. Crystal-type 100-kHz oscillator.

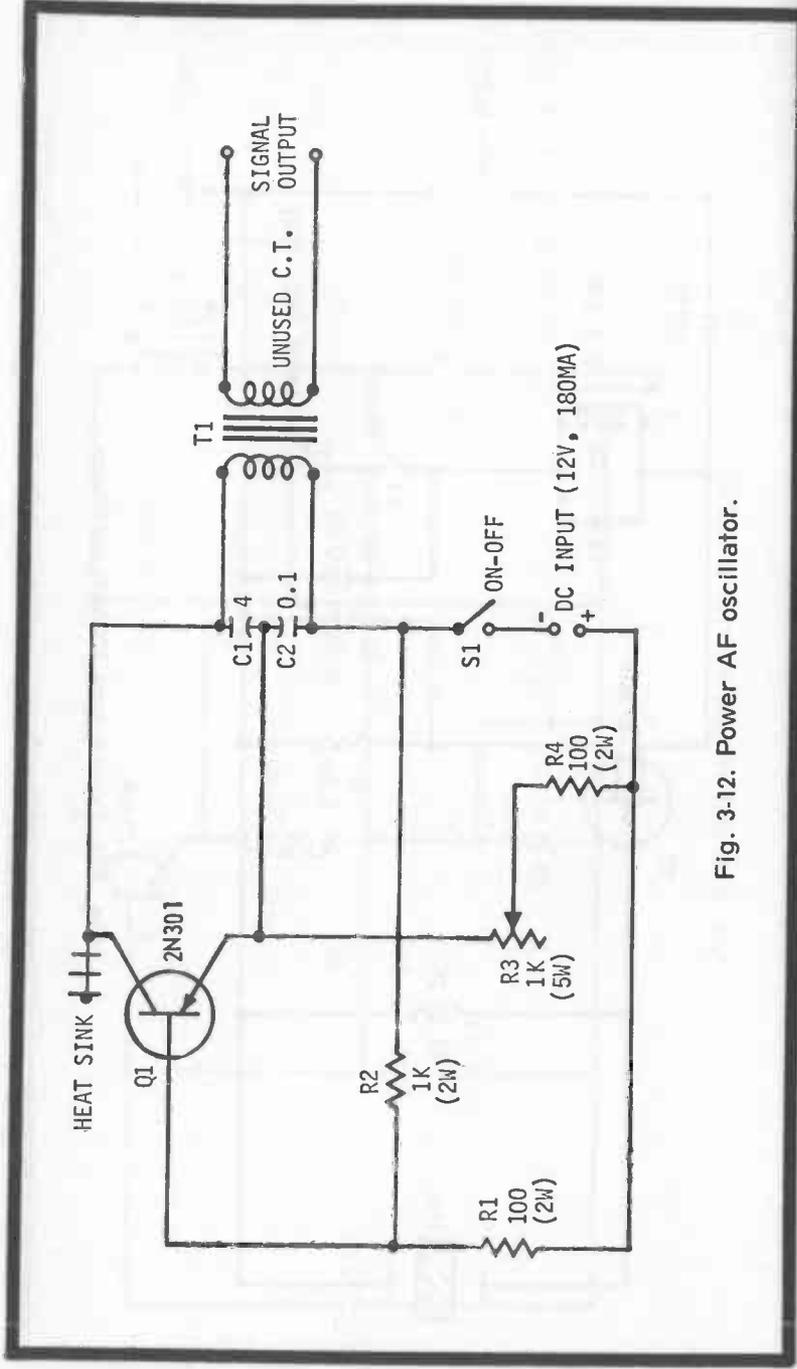


Fig. 3-12. Power AF oscillator.

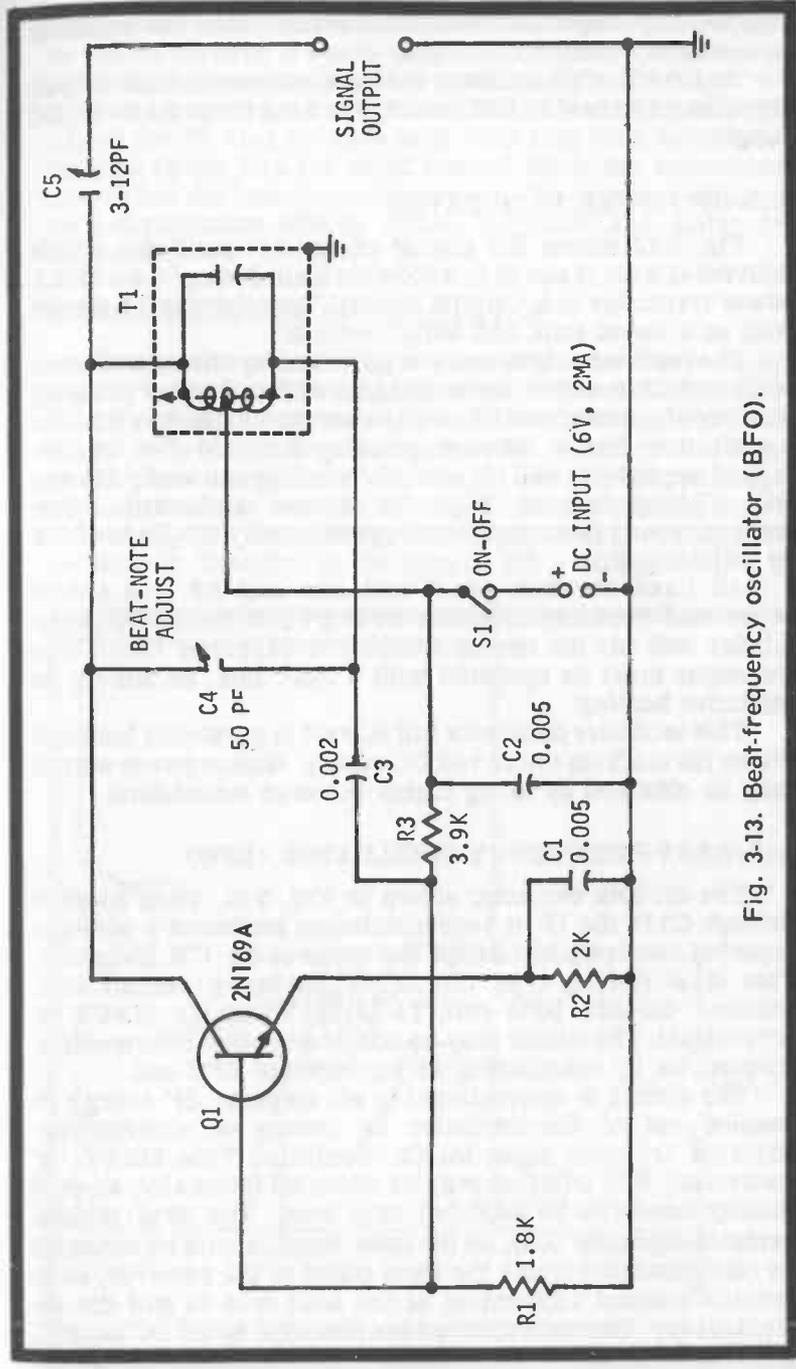


Fig. 3-13. Beat-frequency oscillator (BFO).

ficiently high impedance that it will not overload the oscillator and severely reduce its output.

At 100 kHz this oscillator delivers an open-circuit output-signal amplitude of 5v RMS and draws 1 ma from the 6-volt DC supply.

3.12—AF POWER OSCILLATOR

Fig. 3-12 shows the circuit of an AF oscillator which delivers $\frac{1}{2}$ watt of power to a 200-ohm load. It employs a 2N301 power transistor in a Colpitts circuit. Transformer T1 serves both as a tuned tank and output coupler.

The oscillation frequency is governed by the capacitance of C1 and C2 in series, the inductance of the 200-ohm primary winding of transformer T1, and the setting of rheostat R3. The transformer has a 200-ohm primary and 200-ohm center-tapped secondary, and its primary winding can carry 200 ma DC (Chicago-Stancor Type TA-58, or equivalent). The frequency may be varied from approximately 400 Hz to 3 kHz by adjusting R3.

All fixed resistors are 2-watt size and R3 is a 5-watt wirewound rheostat. Capacitor C1 is a 4-mfd metalized paper tubular unit (do not use an electrolytic capacitor here). The transistor must be operated with a heat sink, as shown, to minimize heating.

This oscillator delivers a full $\frac{1}{2}$ watt to a 200-ohm load and draws 180 ma from the 12-volt DC supply. Higher power output may be obtained by using higher-powered transistors.

3.13—BEAT-FREQUENCY OSCILLATOR (BFO)

The 455-kHz oscillator shown in Fig. 3-13, when coupled through C5 to the IF or second-detector section of a straight superhet receiver, will adapt the receiver for CW reception. This is a Hartley-type circuit employing a commercial, shielded, 455-kHz BFO coil, T1 (J. W. Miller No. 912-C5, or equivalent). The circuit may be adapted to other intermediate frequencies by substituting an appropriate BFO coil.

The circuit is conventional in all respects. IF energy is coupled out of the oscillator by means of screwdriver-adjusted trimmer capacitor C5 (Centralab Type 822-FZ, or equivalent). The trimmer may be mounted internally, since it usually needs to be adjusted only once. The 50-pf midget variable capacitor (C4), on the other hand, should be mounted on (and insulated from) the front panel of the receiver, as it permits manual adjustment of the beat note to suit the individual ear. Current drain is 2 ma from the 6-volt DC supply.

The BFO unit (T1), has a screwdriver-adjusted tuning

slug, a convenience that permits the audible beat note range to be preset through a combination of T1 and C4 adjustments. (The usual procedure is to tune-in a signal "on the nose" with the receiver; then, with C4 set to its maximum capacitance, adjust the T1 slug for zero beat. The slug then need not be touched again. As C4 is tuned toward minimum capacitance, the pitch of the beat note will increase.) To prevent deleterious body-capacitance effects, follow the rotor and stator connections shown for C4 in Fig. 3-13.

3.14—MICROPOWER AF OSCILLATOR

Transistors are noted for their ability to operate from very low DC voltage and current, a feature that has made possible the powering of a transistor oscillator from the feeble DC output of a thermopile, makeshift battery, low-output photocell, or any similar DC supply. In some applications, such an oscillator is useful as a DC-to-AC transducer, since its output-signal amplitude is proportional (though not necessarily linearly) to the applied DC voltage.

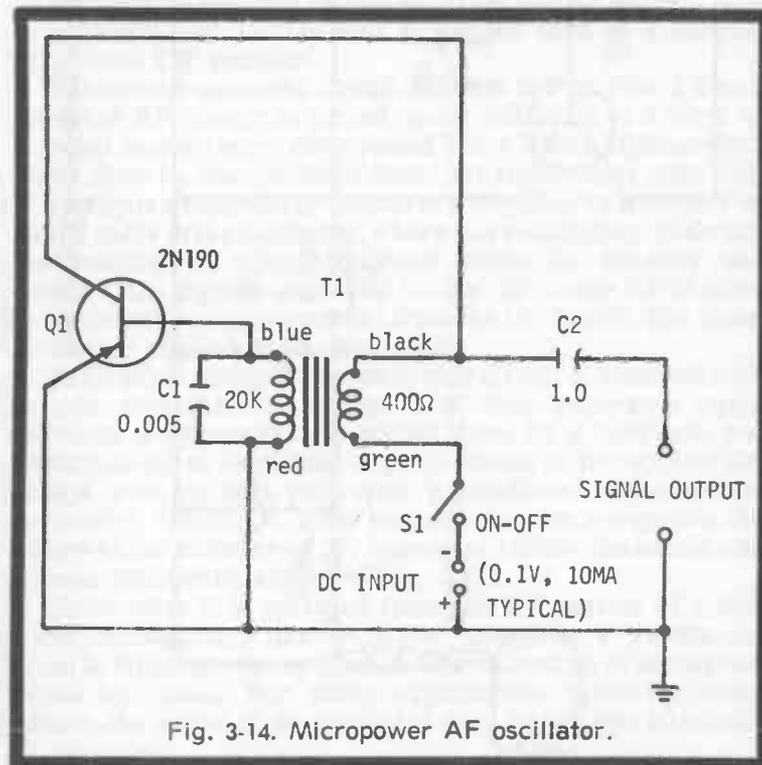


Fig. 3-14. Micropower AF oscillator.

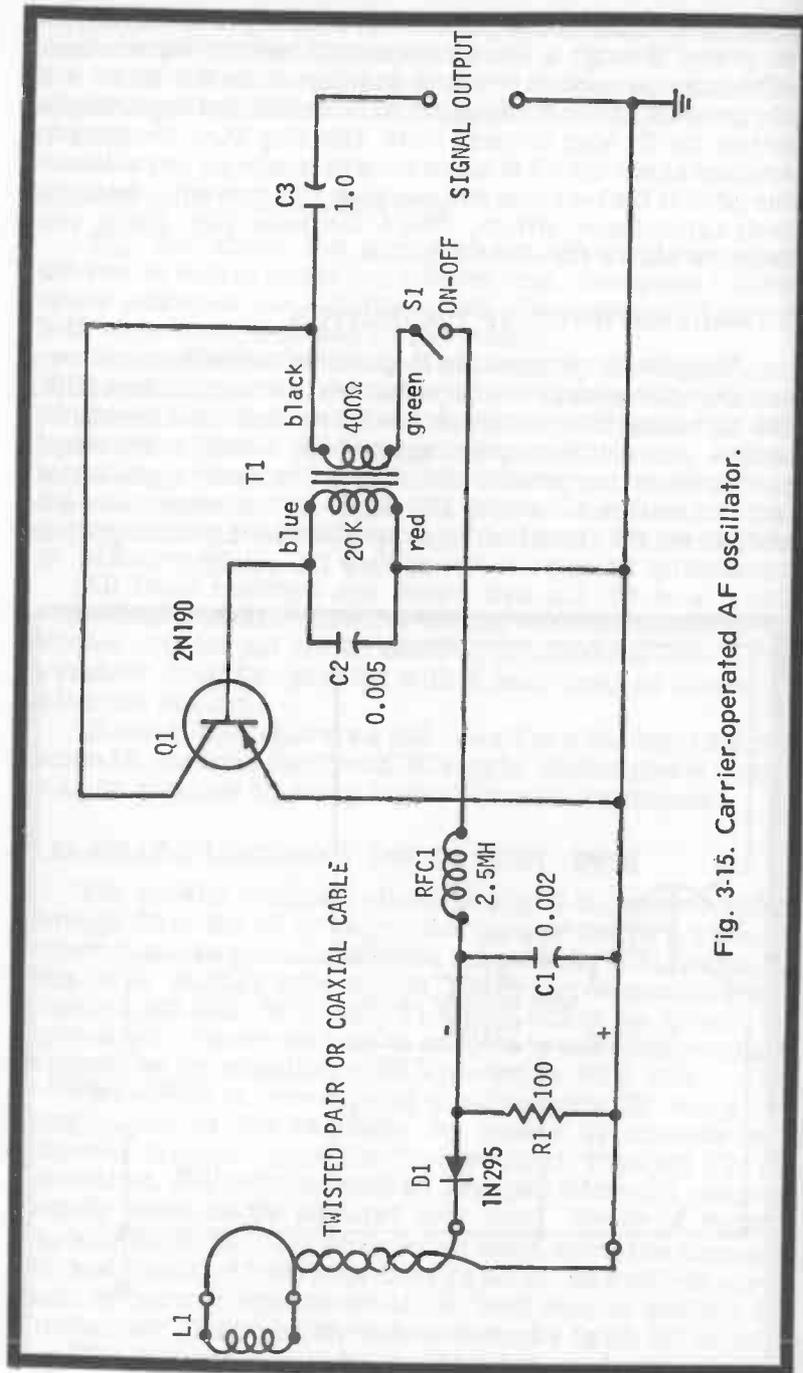


Fig. 3-15. Carrier-operated AF oscillator.

Fig. 3-14 shows a sine-wave audio-frequency oscillator circuit for operation from a low DC supply. With 0.1v applied to the DC input terminals, the current drain is $10 \mu\text{a}$ (representing a DC input power of 1 microwatt), and the open-circuit output-signal amplitude is 1.7 mv RMS. At 0.2v DC, the current is very little more than the original $10 \mu\text{a}$, but the open-circuit output-signal amplitude is 10 mv RMS.

This is a transformer-feedback oscillator with a miniature 20,000 ohms-to-400 ohms transformer, T1 (Argonne AR-105, or equivalent) serving both as the feedback coupler and as the tuned tank. The oscillation frequency is determined principally by capacitance C1 and the inductance of the high-impedance winding of the transformer. If C1 is 0.005 mfd, the frequency is approximately 400 Hz. For higher frequencies, decrease C1; for lower frequencies, increase C1. The transformer must be polarized correctly for regenerative feedback; follow the color coding shown in Fig. 3-14.

3.15-CARRIER-OPERATED AF OSCILLATOR

The micropower oscillator described in the foregoing section may be adapted to operate from the RF carrier of a radio transmitter. In this way, it can be used as a simple, batteryless, CW monitor.

The carrier-operated circuit appears in Fig. 3-15. A small amount of RF energy is picked up by coil L1 (2 or 3 turns of insulated hookup wire, closewound 2 or 3 inches in diameter) placed close to, but insulated from, the transmitter tank coil. RF energy is conducted by means of a length of twisted pair or coaxial cable to the oscillator, where it is rectified by diode D1. The resulting DC voltage appears across the 100-ohm load resistor (R1). Bypass capacitor C1 and RF choke RFC1 filter the radio-frequency component from the DC output. The clean DC then is applied to the oscillator.

In all other respects, the oscillator circuit is identical with the one described in Section 3.14. The generated audio frequency is approximately 400 Hz when C2 is 0.005 mfd and the output-signal amplitude is proportional to the applied DC voltage (and, by back reference, proportional to the carrier amplitude). Diode, D1, must be polarized for a negative DC voltage at the collector of the transistor (follow the anode and cathode positioning shown in Fig. 3-15).

Even when it is operated from the RF carrier of a low-power transmitter, this oscillator produces a substantial signal in high-impedance headphones connected to the signal-output terminals. For other applications requiring more volume, the output of the oscillator may be fed into a suitable AF amplifier.

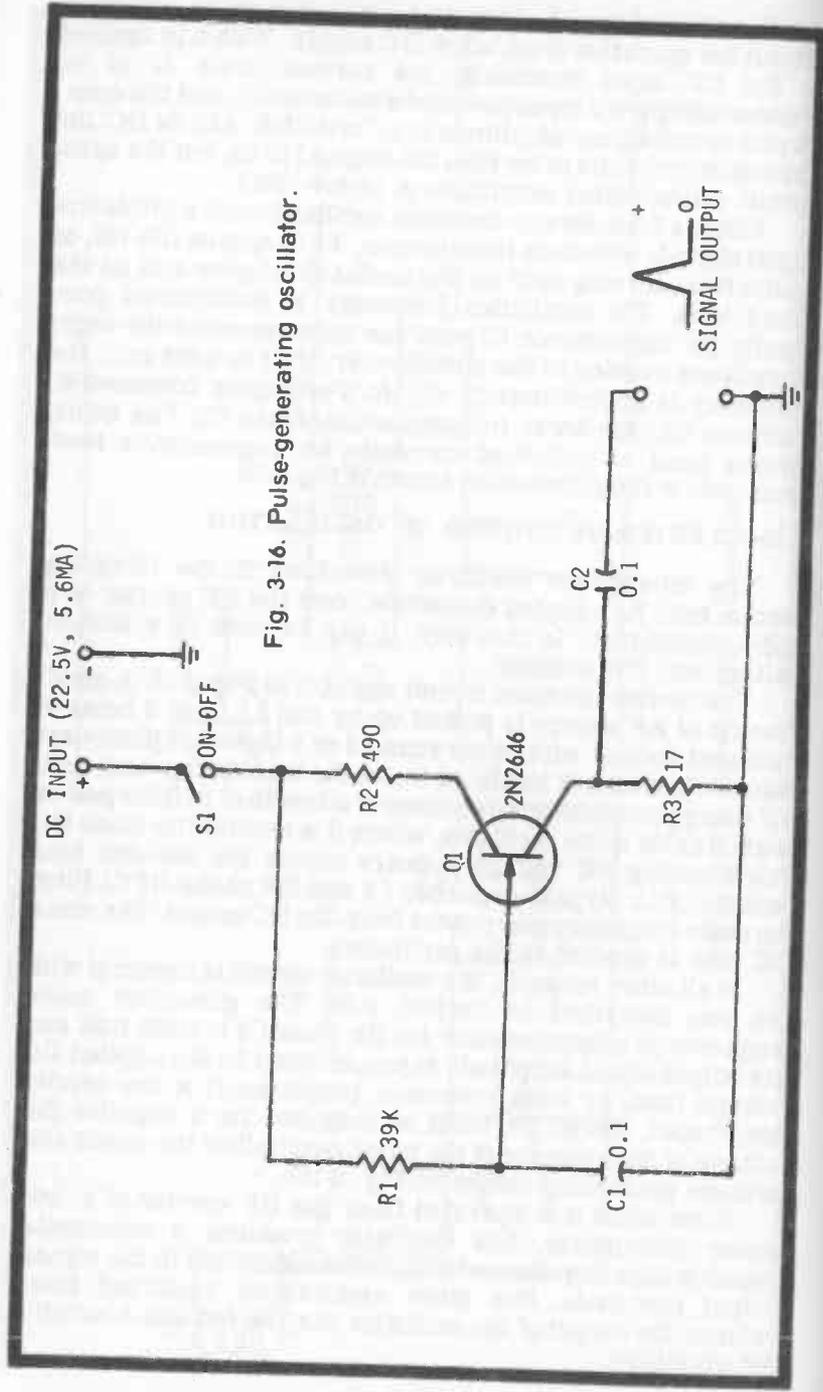


Fig. 3-16. Pulse-generating oscillator.

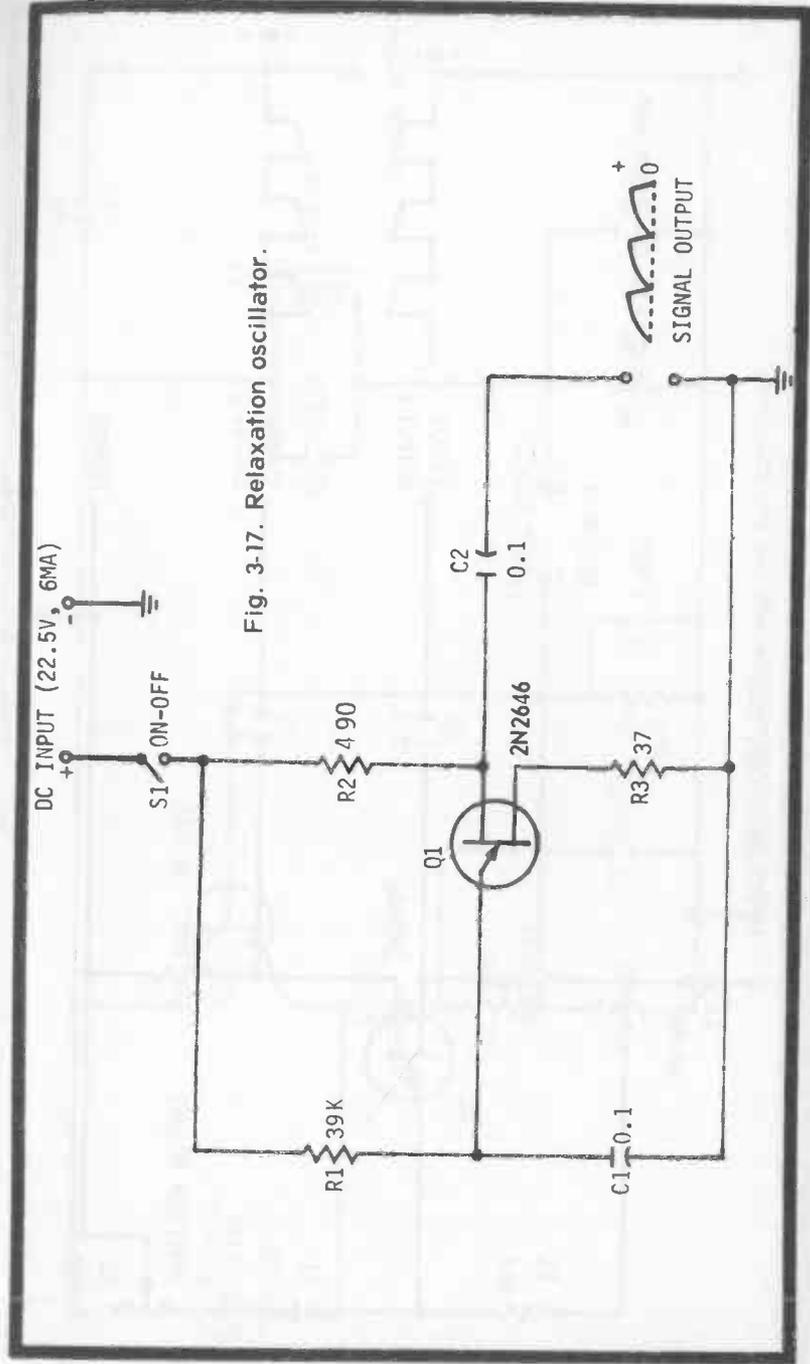


Fig. 3-17. Relaxation oscillator.

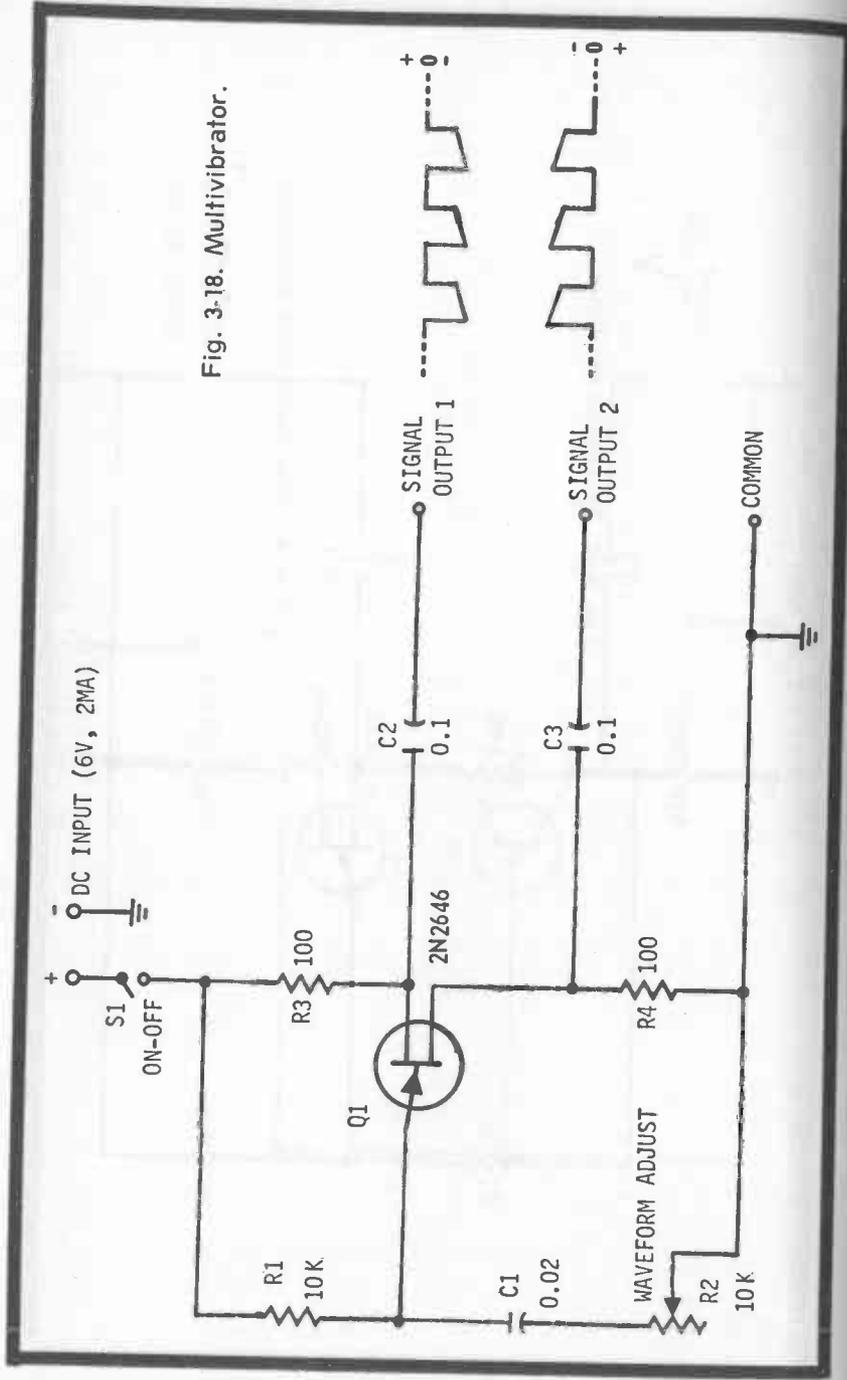


Fig. 3-18. Multivibrator.

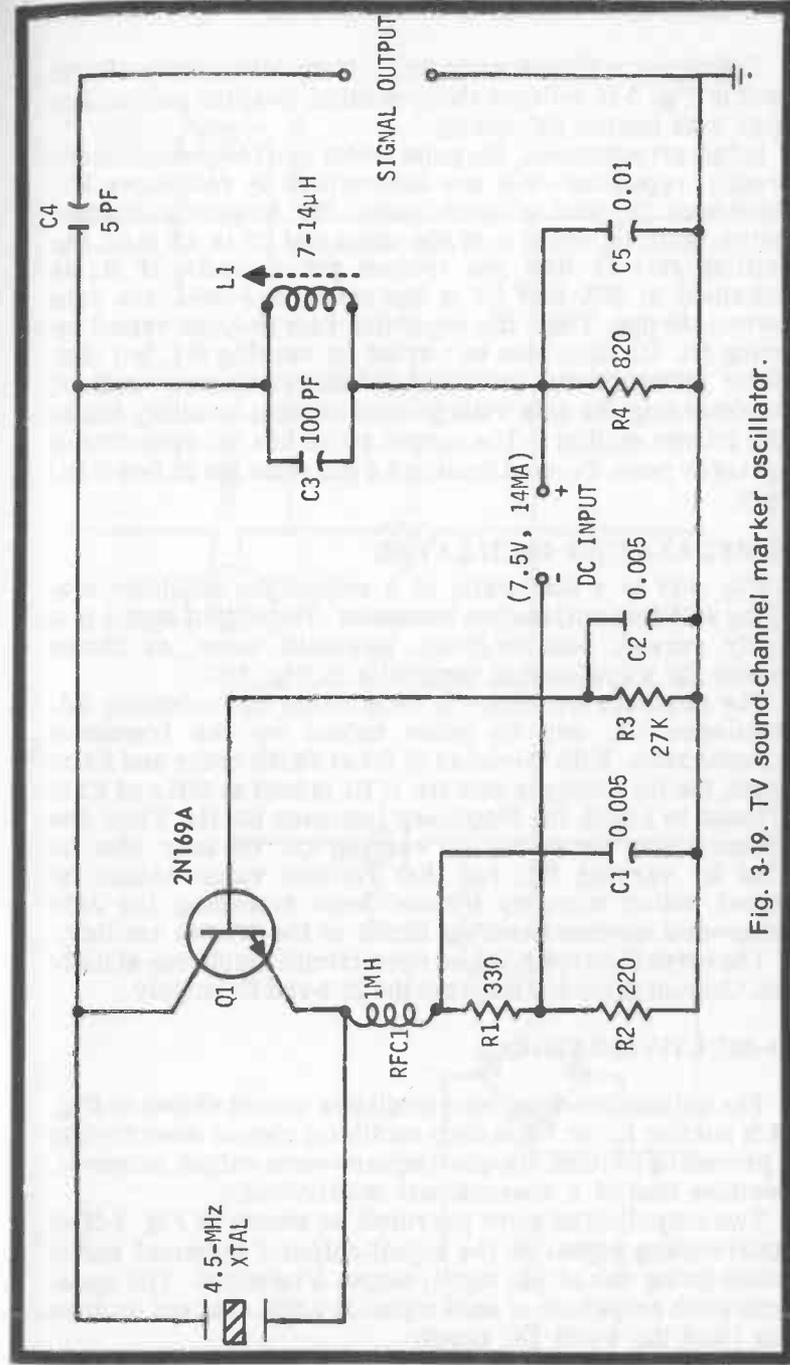


Fig. 3-19. TV sound-channel marker oscillator.

3.16—PULSE-GENERATING OSCILLATOR

Employing a 2N2646 unijunction transistor, the oscillator circuit in Fig. 3-16 delivers sharp-pointed, positive pulses. The circuit uses simple RC tuning.

In this arrangement, the pulse width and frequency (more correctly, repetition rate) are determined by resistance R1, capacitance C1, and to some extent the transistor characteristics. With R1 equal to 39,000 ohms and C1 to 0.1 mfd, the repetition rate is 1600 pps (pulses per second). If R1 is maintained at 39K and C1 is increased to 1 mfd, the rate becomes 160 pps. Thus, the repetition rate may be varied by varying C1. (It may also be varied by varying R1, but that resistor value cannot be reduced indiscriminately without soon exceeding the safe voltage- and current-handling limits of the 2N2646 emitter.) The output pulse has an open-circuit height of 9v peak. Current drain is 5.6 ma from the 22.5-volt DC supply.

3.17—RELAXATION OSCILLATOR

Fig. 3-17 is a schematic of a relaxation oscillator employing a 2N2646 unijunction transistor. The output signal is a slightly curved, positive-going, sawtooth wave, as shown opposite the signal-output terminals in Fig. 3-17.

The sawtooth frequency is determined by resistance R1, capacitance C1, and to some extent by the transistor characteristics. With the value of R1 at 39,000 ohms and C1 at 0.1 mfd, the frequency is 1600 Hz. If R1 is held at 39K and C1 is increased to 1 mfd, the frequency becomes 160 Hz. Thus, the frequency may be varied by varying C1. (It may also be varied by varying R1, but that resistor value cannot be reduced indiscriminately without soon exceeding the safe voltage- and current-handling limits of the 2N2646 emitter.)

The output sawtooth has an open-circuit amplitude of 0.42v peak. Current drain is 6 ma from the 22.5-volt DC supply.

3.18—MULTIVIBRATOR

The unijunction-transistor oscillator circuit shown in Fig. 3-18 is similar to the relaxation oscillator circuit described in the preceding section. Its quasi square-wave output, however, resembles that of a conventional multivibrator.

Two output signals are provided, as shown in Fig. 3-18—a negative-going signal at the signal-output 1 terminal and a positive-going one at the signal-output 2 terminal. The open-circuit peak amplitude of each signal is 0.56v. Current drain is 2 ma from the 6-volt DC supply.

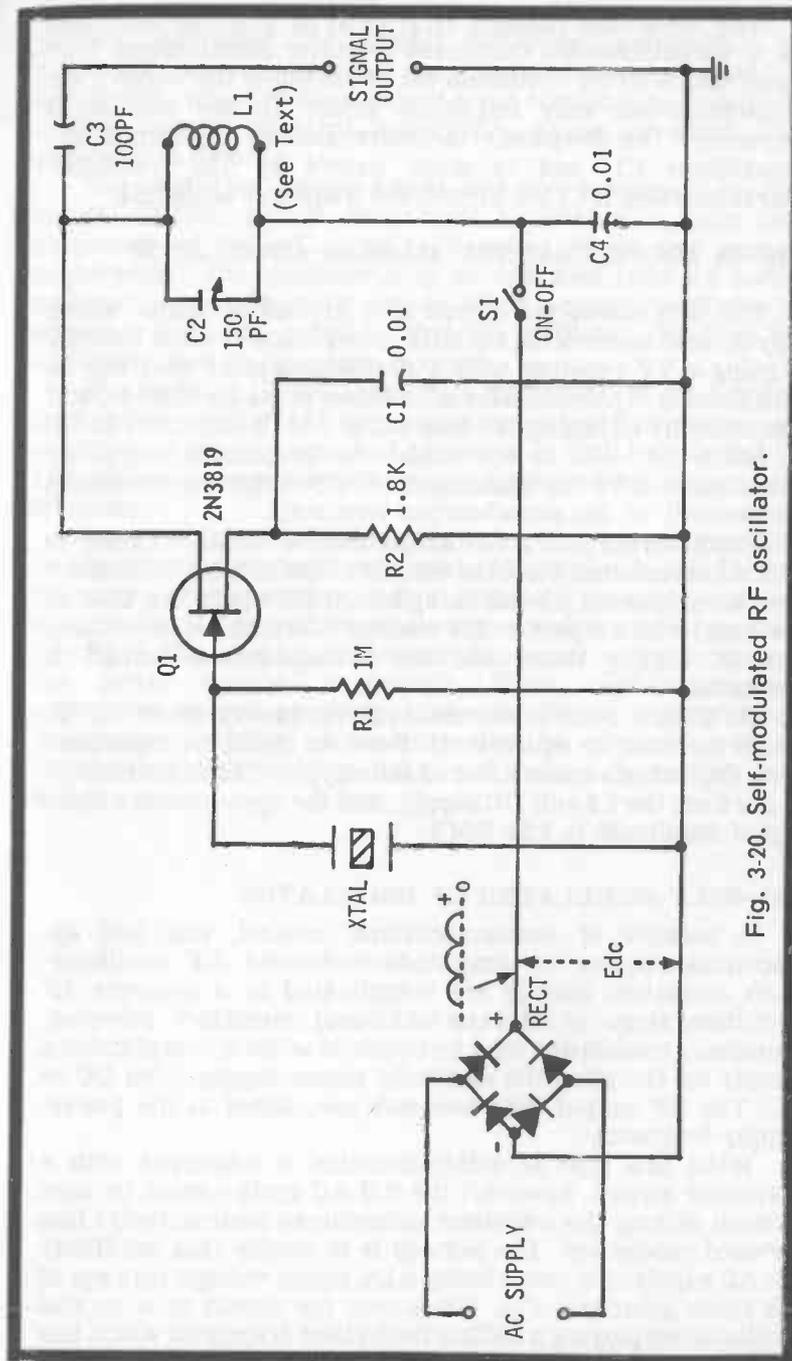


Fig. 3-20. Self-modulated RF oscillator.

The 10,000-ohm rheostat (R2) must be adjusted (with the aid of an oscilloscope connected to either signal-output 1 or signal-output 2) for minimum tilt of the top of the wave. This adjustment has only negligible effect on the oscillation frequency. The frequency is determined by resistance R1, capacitance C1, and to some extent by the transistor characteristics. If C1 is 0.02 mfd, the frequency is 2600 Hz.

3.19—TV SOUND-CHANNEL MARKER OSCILLATOR

Fig. 3-19 shows the circuit of a crystal oscillator which may be used to mark the 4.5-MHz sound channel when visually aligning a TV receiver with a sweep generator that has no such marker. This oscillator is fix-tuned to the 4.5-MHz crystal frequency by adjusting the slug-tuned 7-14 μ h inductor, L1 (J. W. Miller No. 4406, or equivalent) for maximum output, as indicated by a VT (or transistorized) RF voltmeter connected temporarily to the signal-output terminals.

Resistors R2 and R4 in series form a voltage divider to which is connected the base circuit of the 2N169A transistor. This arrangement places an optimum DC operating bias on the base, with respect to the emitter bias and collector bias. The DC supply floats and the voltage-divider tap-off is grounded.

RFC1 is a 1-mh miniature radio-frequency choke (J. W. Miller No. 4652, or equivalent). For best stability, capacitors C1 to C4, inclusive, are silvered-mica types. The circuit draws 14 ma from the 7.5-volt DC supply, and the open-circuit output-signal amplitude is 4.5v RMS.

3.20—SELF-MODULATED RF OSCILLATOR

A number of communications, control, and test applications require an amplitude-modulated RF oscillator. Such oscillators usually are complicated by a separate AF modulator stage (at least one additional transistor); however, amplitude modulation may be obtained without complications simply by changing the oscillator power supply from DC to AC. The RF output then becomes modulated at the power-supply frequency.

When this type of self-modulation is employed with a transistor circuit, however, the full AC cycle cannot be used without driving the transistor (sometimes destructively) into forward conduction. The remedy is to rectify (but not filter) the AC supply, the result being a DC ripple voltage (always of the same polarity). Fig. 3-20 shows the circuit of a crystal oscillator employing a 2N3819 field-effect transistor which has been adapted to self-modulation by means of a full-wave

bridge rectifier (RECT) that converts the AC supply to a positive, pulsating DC supply. Full-wave rectification is employed, since it provides an unbroken train of positive half-cycles, whereas simpler half-wave rectification results in a gap between adjacent half-cycles, a gap in which the oscillator is dead.

The oscillator output signal has the frequency of the crystal (XTAL) and is 100-percent modulated at twice the frequency of the AC supply. Because of its low power requirement, the oscillator may be operated from an audio test oscillator as the power supply. While a crystal oscillator is shown in Fig. 3-20, a self-excited RF oscillator also may be self-modulated in the same manner.

The rectifier (RECT) may be a miniature, 10-ma bridge-type meter rectifier. It also might consist of four silicon or germanium diodes connected as shown in Fig. 3-20. Inductor L1 is chosen to resonate with tuning capacitor C2 at the crystal frequency.

In an experimental setup of the circuit operated from the 200-Hz output of an audio oscillator (AC supply, 6v RMS), the DC output of the rectifier (Edc in Fig. 3-20) was 2.8v and the open-circuit RF output-signal average-carrier amplitude was 2.3v RMS. The signal is modulated 100 percent at 400 Hz (twice the power supply frequency). While high-impedance, capacitive output coupling (through C3) is shown, low-impedance coupling, if desired, may be obtained with a conventional link-coupling coil wound around the bottom end of L1.

Chapter 4

Control & Alarm Devices

Its high sensitivity (both DC and AC), wide frequency response, and generally low power requirements suit the transistor to use in control devices and alarm devices in which only one active element must be used. Thus, it is possible with relatively simple equipment to obtain reliable near control and remote control. The 21 circuits described in this chapter are representative of various classes of electronic control and will suggest other applications. They are intended primarily to be used as they are shown; but, as building blocks, they may be combined with amplifiers from Chapters 1 and 2 and/or oscillators from Chapter 3 to produce more complex control systems.

In each of these circuits, unless otherwise indicated on the diagram or in the text, capacitances are in mfd and resistances in ohms. Capacitors are rated at 25 DCWV and resistors at one-half watt. Where particular components from specific manufactures are indicated, they are the ones actually used by the author and they (or their exact equivalents, where available) seem to be essential to correct operation of the circuit. In all other instances, any component having the specified electrical ratings may be used. Where frequency, phase, signal amplitude, voltages, and currents are given, these are the values that were obtained with the author's experimental model and they may vary somewhat with individual transistors and circuit components.

4.1—SENSITIVE DC RELAY (NPN)

A transistor operated as a DC amplifier in conjunction with a DC relay will perform a variety of switching applications. A milliampere-type relay thus may be operated from a control signal of only a few microamperes, or an ampere-type relay may be operated on a few milliamperes. Sensitive relay circuits of this type are shown in Figs. 4-1 to 4-4.

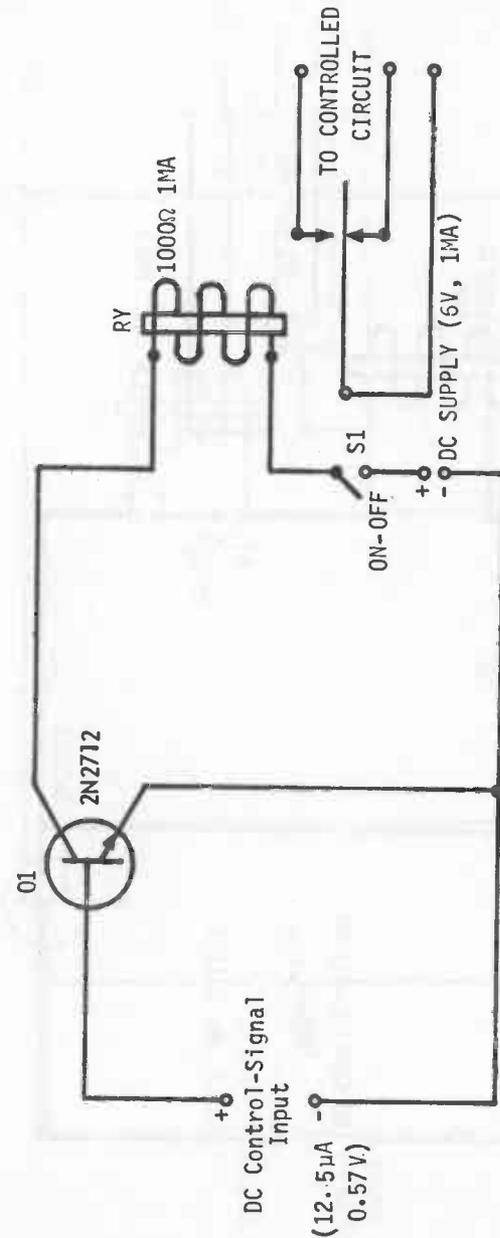


Fig. 4-1. Sensitive DC relay (NPN).

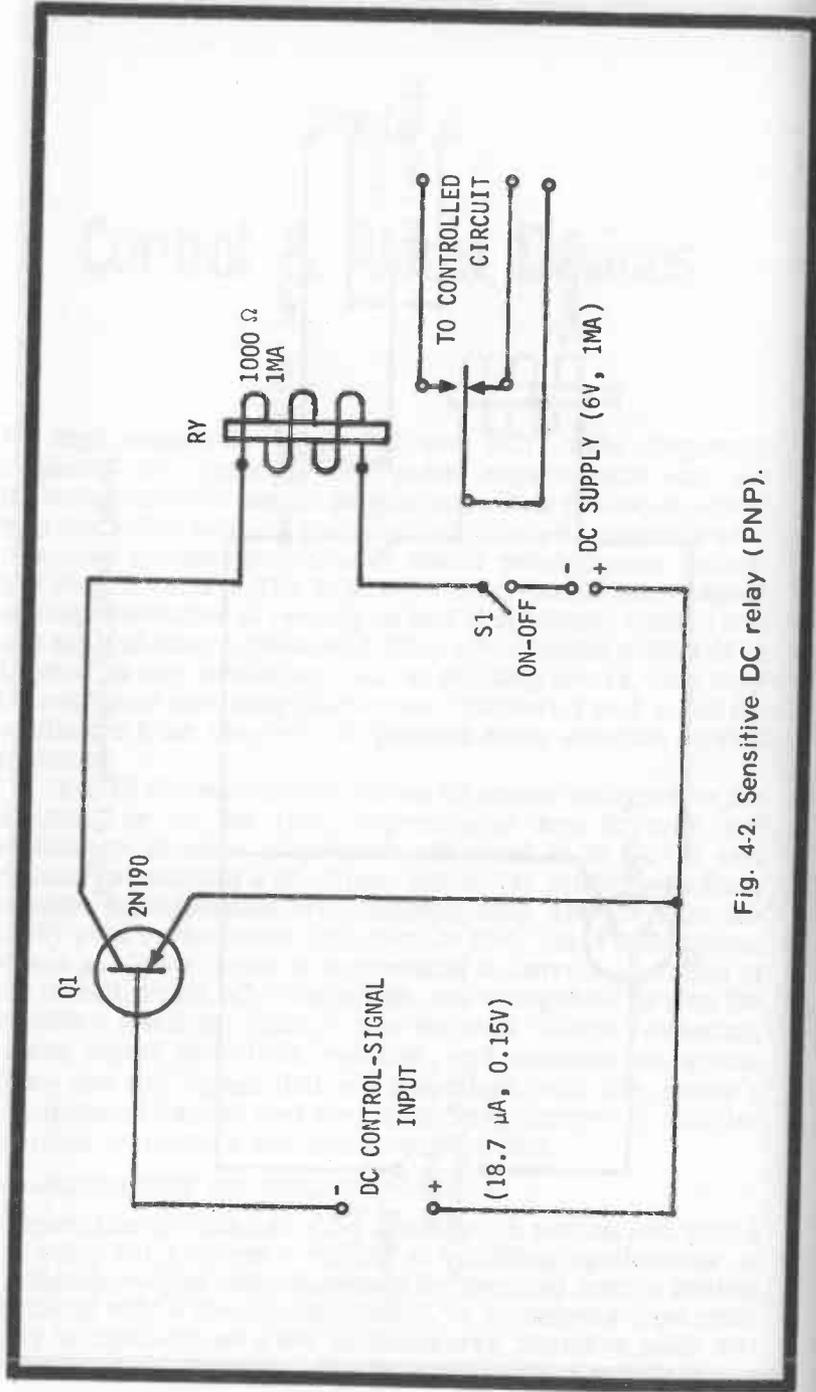


Fig. 4-2. Sensitive DC relay (PNP).

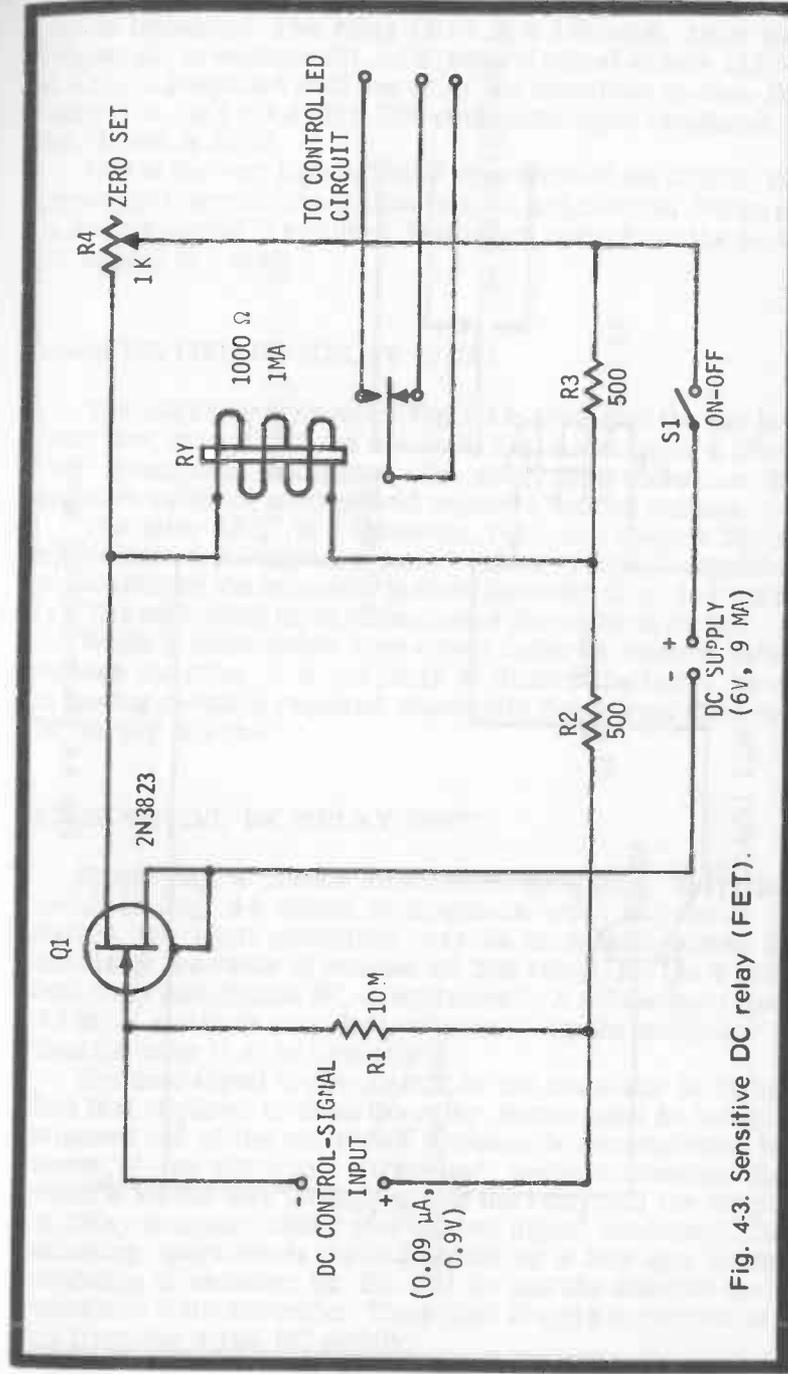


Fig. 4-3. Sensitive DC relay (FET).

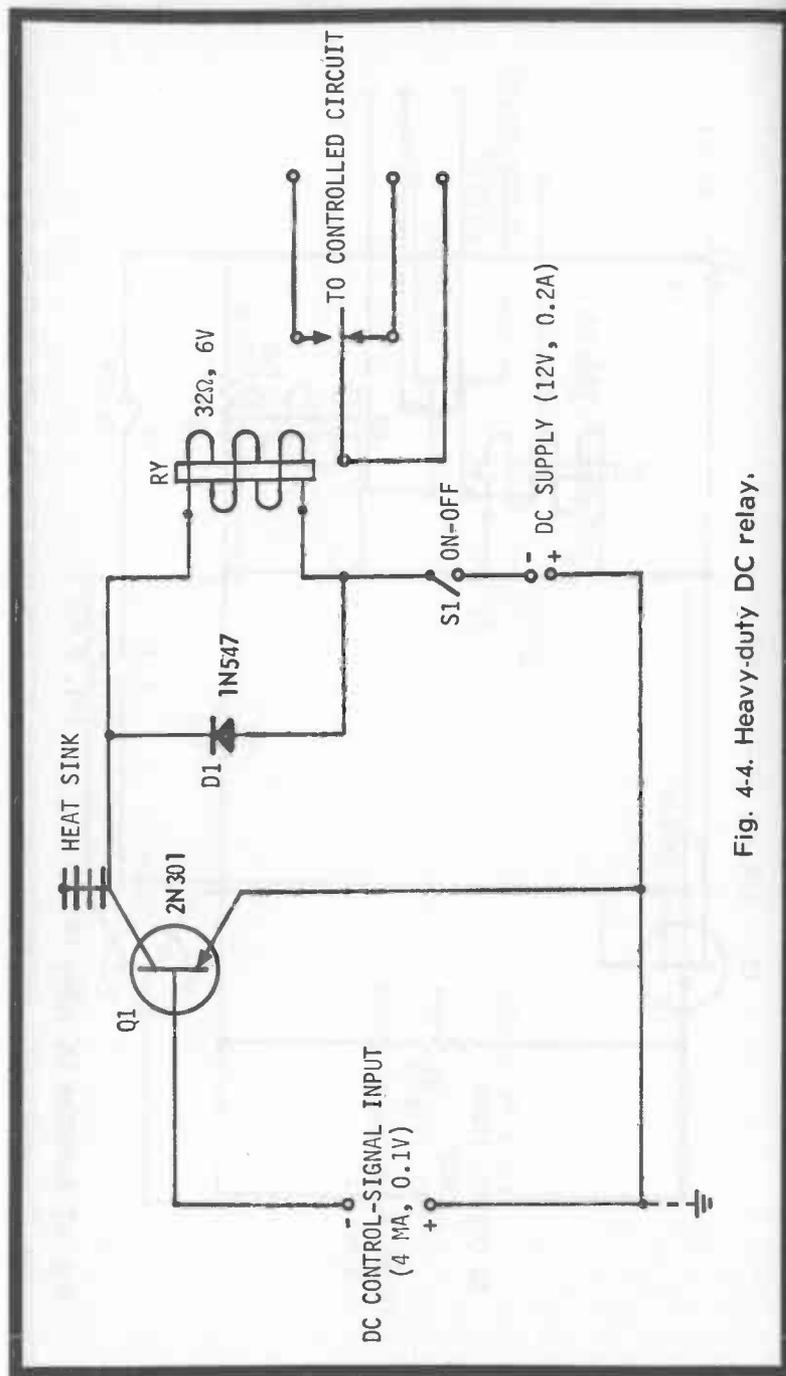


Fig. 4-4. Heavy-duty DC relay.

Fig. 4-1 shows a relay circuit employing a 2N2712 NPN silicon transistor. The relay (RY) is a 1000-ohm, 1-ma unit (Sigma 5F, or equivalent). A DC control signal of only $12.5 \mu\text{a}$ at 0.57v is amplified sufficiently by the transistor to close the relay (i. e., to 1 ma at 1v). The equivalent input resistance of the circuit is 45.6K.

Due to the very high collector resistance of the 2N2712, the zero-signal current through the relay is insignificant, hence no balancing circuit is required. Maximum drain from the 6-volt DC supply is 1 ma.

4.2—SENSITIVE DC RELAY (PNP)

The relay circuit shown in Fig. 4-2 is similar to the one just described, except that the circuit in Fig. 4-2 employs a 2N190 PNP germanium transistor. The chief differences are the negative collector voltage and negative control voltage.

The relay (RY) is a 1000-ohm, 1-ma unit (Sigma 5F, or equivalent). A DC control signal of $18.7 \mu\text{a}$ at 0.15v is amplified sufficiently by the transistor to close the relay (i. e., to 1 ma at 1v). The equivalent input resistance of the circuit is 8.02K.

While a measurable zero-signal collector current flows through the relay, it is too small to disturb the latter, so no balancing circuit is required. Maximum drain from the 6-volt DC supply is 1 ma.

4.3—SENSITIVE DC RELAY (FET)

Employing a 2N3823 field-effect transistor, the relay circuit in Fig. 4-3 offers 10 megohms input resistance. If desired, the input resistance may be made still higher by increasing the value of resistor R1. The relay (RY) is a 1000-ohm, 1-ma unit (Sigma 5F, or equivalent). A DC control signal of $0.09 \mu\text{a}$ at 0.9v is amplified sufficiently by the transistor to close the relay (i. e., to 1 ma at 1v).

The zero-signal drain current of the transistor is higher than that required to close the relay, hence must be initially balanced out of the relay coil. Balance is accomplished by means of the 1000-ohm, wirewound, zero-set rheostat, R4, which is set for zero voltage across the relay coil (or simply for relay dropout) under zero control signal conditions. The balancing operation is made possible by a four-arm bridge consisting of resistors R2, R3, and R4 and the internal drain resistance of the transistor. The circuit draws a maximum of 9 ma from the 6-volt DC supply.

4.4—HEAVY-DUTY DC RELAY

While the three preceding circuits employ sensitive, 1-ma, DC relays, the same combination of DC amplifier plus electromechanical relay may be used with a higher-current relay. Fig. 4-4 shows a circuit used to sensitize a 6-volt, 32-ohm relay (Potter & Brumfield Type KA5DY, or equivalent). In this device, a 2N301 power transistor amplifies the 4-ma, 0.1v DC control signal to the 187.5-ma, 6v level required to close the relay (this represents a power amplification of better than 2800).

The circuit draws a maximum of 200 ma from the 12-volt DC supply. Although the zero-signal collector current is measurable, it is too low to disturb the relay. Hence, no balancing circuit is required. The input resistance presented to the DC control-signal source is 25 ohms. The reverse-connected 1N547 silicon diode (D1) suppresses kickback transients generated by the relay-coil inductance, which might damage the transistor.

The same type of circuit may be employed to sensitize other heavy-current relays, provided the proper transistor is chosen to (1) handle the relay-coil current and (2) provide the desired current amplification.

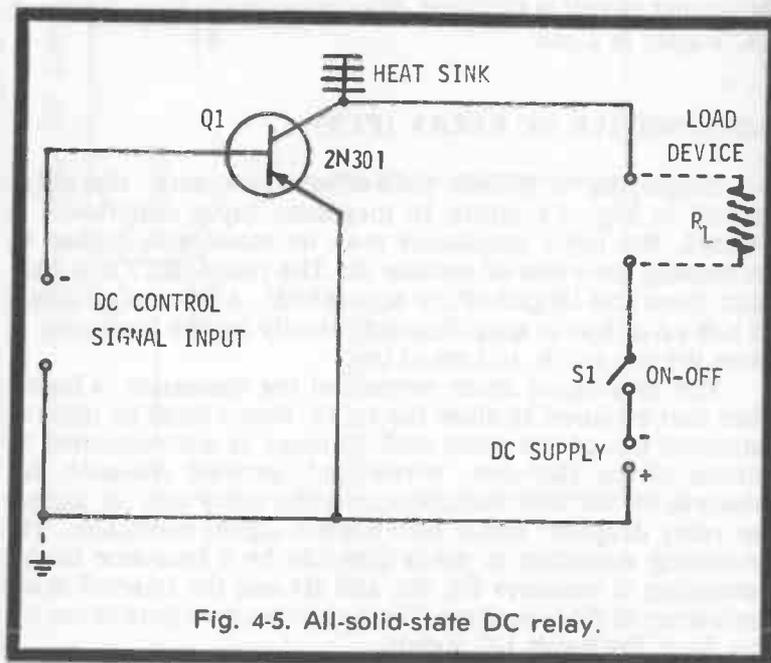


Fig. 4-5. All-solid-state DC relay.

4.5—ALL-SOLID-STATE DC RELAY

When an external load resistance (i. e., the resistance of the controlled circuit or device) is low with respect to the internal collector resistance of the transistor, the electromechanical relay may be dispensed with entirely and an all-solid-state relay circuit used. Such a circuit has the obvious advantage of freedom from moving parts.

Fig. 4-5 shows a relay circuit of this kind. Here, the load device (R_L) is connected directly in the collector output circuit of a 2N301 power transistor. The amount of current amplification obtained with an arrangement of this sort depends upon the load resistance, varying inversely with R_L . For example, when R_L is 30 ohms or less, the current amplification (for the 2N301) approaches 50 (which means that the DC control-signal input current need be only 1/50 of the required output load current). The maximum DC collector current is 1.5 amperes for the 2N301. Higher current operation is afforded by other power transistors.

In a control circuit of this type, output load current flows as long as input control signal current flows. Unlike the electromechanical relay, however, this transistor relay provides an output current which is proportional to the control current, and the output current, consequently, will not make and break sharply—in the manner of an electromechanical relay—unless the control current also does.

For any desired load current that the transistor is capable of handling (which is identical with the transistor collector current, I_c), the collector-to-emitter voltage (V_{ce}) is equal to the supply voltage (V_{cc}) minus the voltage drop ($I_c R_L$) across the load device: V_{ce} equals V_{cc} minus $I_c R_L$. And V_{ce} must never exceed the maximum safe operating value (32v for the 2N301). The DC supply voltage thus must be chosen with respect to the load resistance and the maximum safe collector/emitter voltage recommended by the transistor manufacturer.

4.6—AC/RF RELAY

Fig. 4-6 is the circuit of a sensitive relay which may be operated from AC and RF control signals. This circuit is useful over a wide frequency range, extending from low, powerline frequencies to at least 50 MHz.

The AC control signal is rectified by a shunt-diode rectifier circuit, consisting of blocking capacitor C1, diode D1, and load resistor R1. The DC output of this circuit is applied to the gate/input circuit of the 2N3823 field-effect transistor (Q1) which as

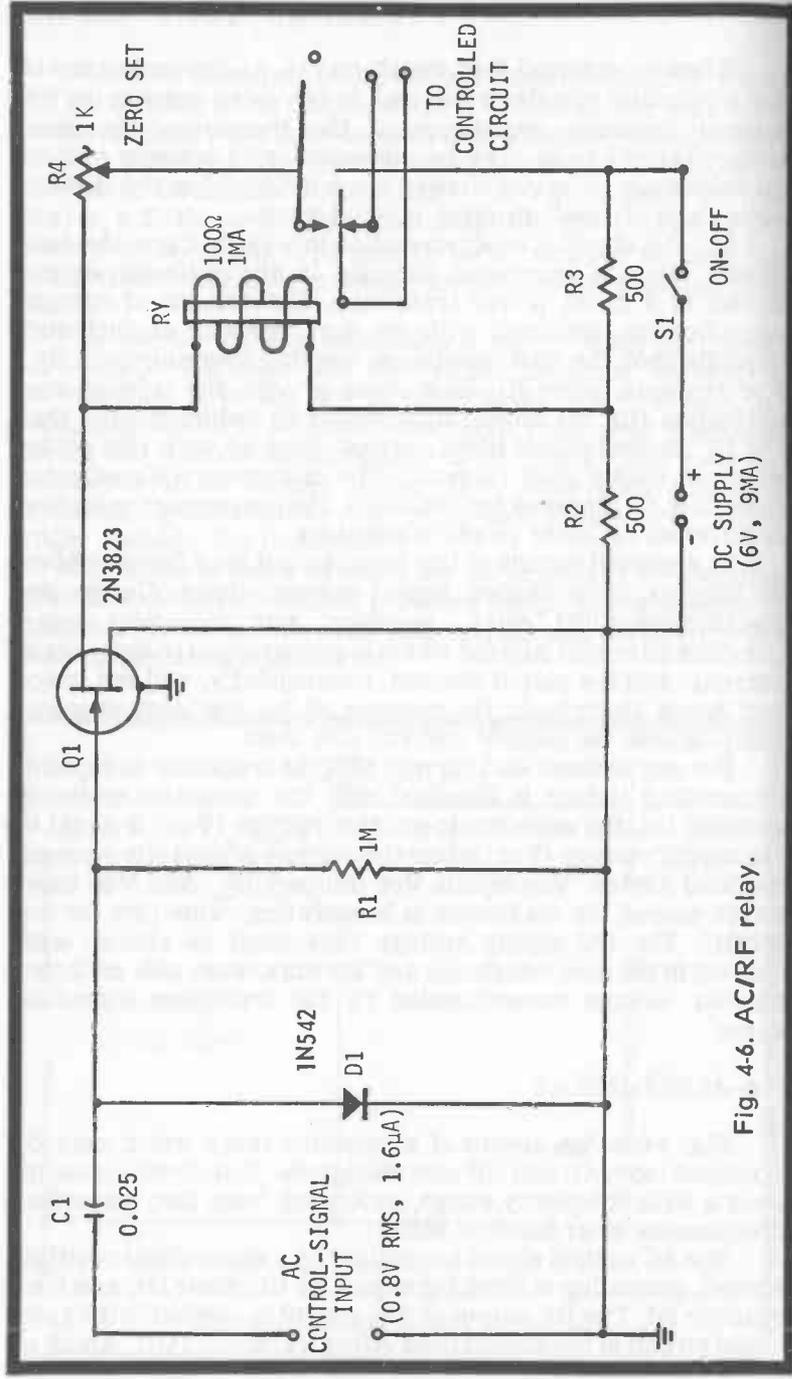


Fig. 4-6. AC/RF relay.

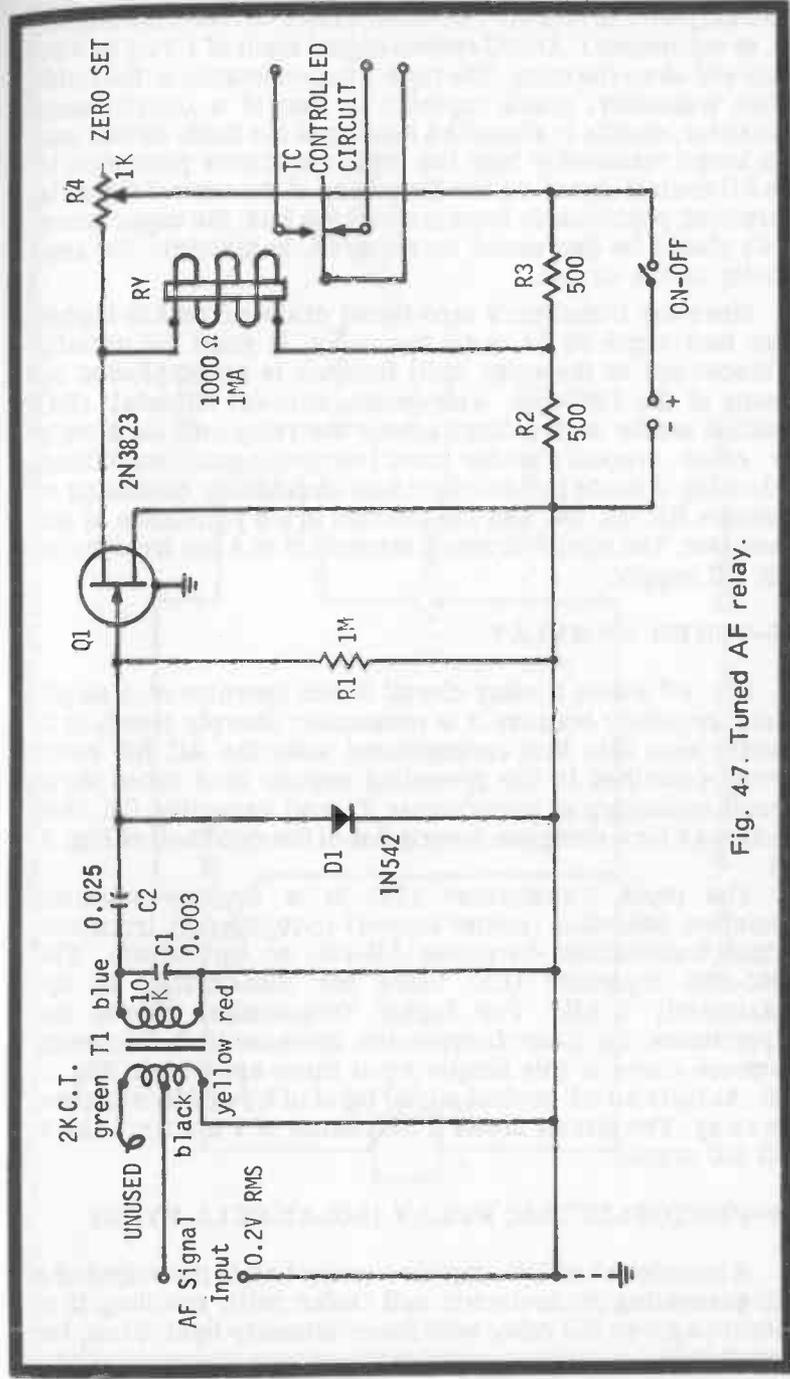


Fig. 4-7. Tuned AF relay.

a DC amplifier drives the 1000-ohm, 1-ma DC/relay, RY (Sigma 5F, or equivalent). An AC control-signal input of $1.6 \mu\text{a}$ at 0.8v RMS will close the relay. The high input resistance of the field-effect transistor, much superior to that of a conventional transistor, results in almost no loading of the diode circuit and this keeps reasonably high the input resistance presented to the AC control signal. As the frequency of the control signal is increased, particularly beyond about 500 kHz, the capacitance of C1 should be decreased, as required, to maintain the sensitivity of the circuit.

Since the transistor's zero-signal drain current is higher than that required to close the relay, it must be initially balanced out of the relay coil. Balance is accomplished by means of the 1000-ohm, wirewound, zero-set rheostat (R4) which is set for zero voltage across the relay coil (or simply for relay dropout) under zero control-signal conditions. Balancing is made possible by a four-arm bridge consisting of resistors R2, R3, R4, and the internal drain resistance of the transistor. The circuit draws a maximum of 9 ma from the 6-volt DC supply.

4.7-TUNED AF RELAY

Fig. 4-7 shows a relay circuit which operates at a single audio frequency because it is reasonably sharply tuned. It is readily seen that this arrangement adds the AC RF relay circuit described in the preceding section to a tuned input circuit consisting of transformer T1 and capacitor C1. (See Section 4.6 for a complete description of the right half of Fig. 4-7.)

The input transformer (T1) is a reverse-connected miniature 2000-ohm (center-tapped)-to-10,000-ohm transistor output transformer (Argonne AR-109, or equivalent). The 0.003-mfd capacitor (C1) tunes the transformer to approximately 1 kHz. For higher frequencies, reduce the capacitance; for lower frequencies, increase it. A frequency response curve of this simple input tuner appears in Fig. 1-16B. At 1kHz an AF control-signal input of 0.2v RMS will close the relay. The circuit draws a maximum of 9 ma from the 6-volt DC supply.

4.8-PHOTOELECTRIC RELAY (SOLAR-CELL TYPE)

A transistorized DC amplifier easily boosts the output of a self-generating photoelectric cell (solar cell), enabling it to operate a given DC relay with lower-intensity light. Thus, the circuit in Fig. 4-8 will operate from a comparatively dim light.

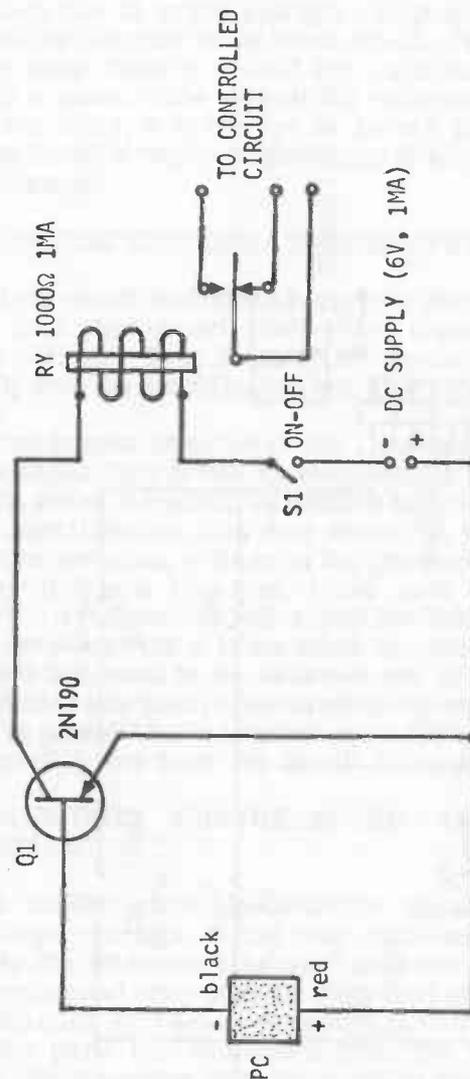


Fig. 4-8. Photoelectric relay (solar-cell type).

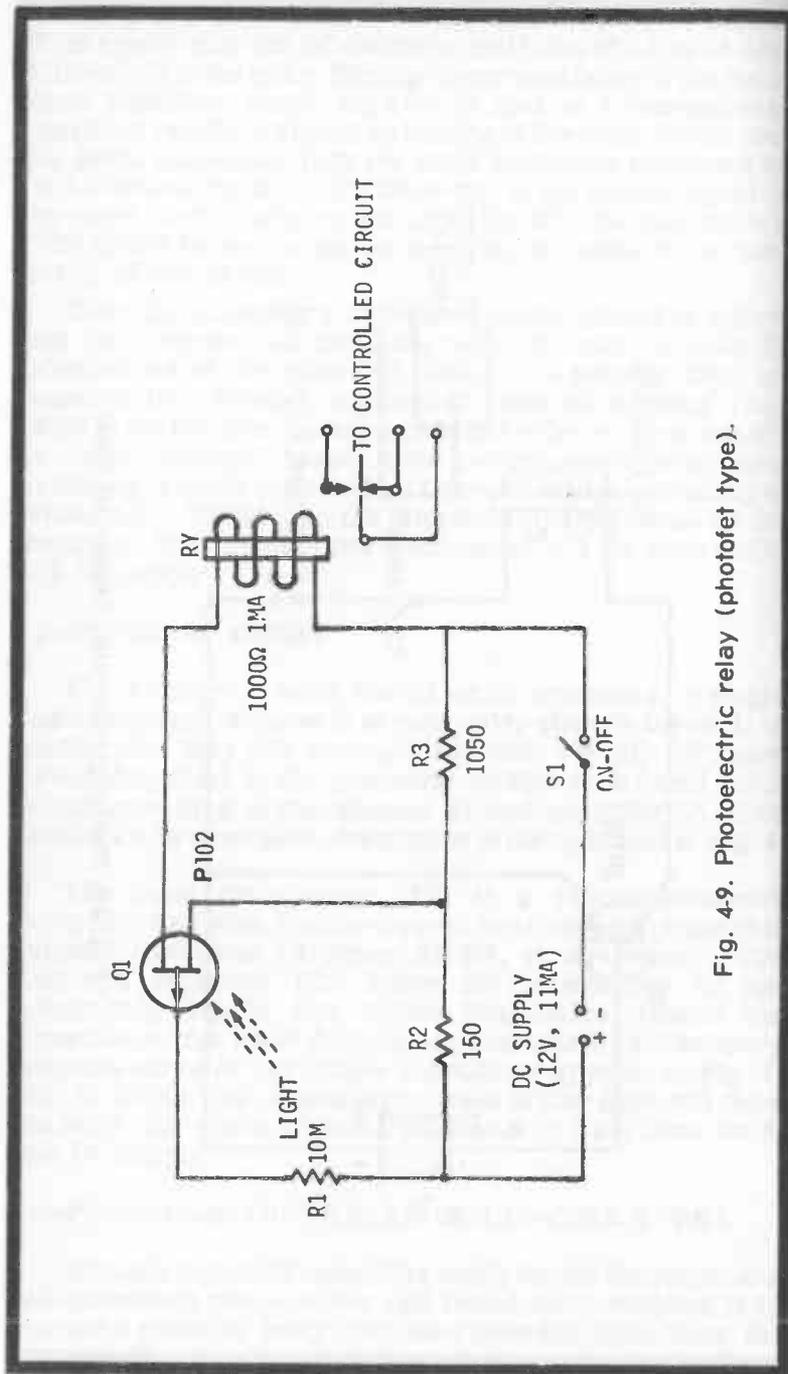


Fig. 4-9. Photoelectric relay (photofet type).

The DC output of the cell, PC (International Rectifier Type S1M, or equivalent), is amplified by a 2N190 transistor (Q1) to actuate the 1000-ohm, 1-ma DC relay, RY (Sigma 5F, or equivalent). Only a 0.15v output from the cell is required to close the relay in this circuit (the S1M cell delivers approximately 0.5v in bright sunlight and proportionately less under artificial illumination at lower levels. (The cell could not drive the relay directly without the transistor.)

While a measurable zero-signal collector current flows through the relay, it is too low to disturb the latter, so no balancing circuit is required. Maximum drain is 1 ma from the 6-volt DC supply.

4.9—PHOTOELECTRIC RELAY (PHOTOFET TYPE)

The light-sensitive device employed in the circuit shown in Fig. 4-9 is a photoelectric field-effect transistor, called a photofet (Siliconix Type P-102). This device combines light sensitivity with the amplification and high input impedance of the FET.

A 10 ma current from the 12-volt DC supply flows through resistors R2 and R3 in series. Consequently, a 1.5v drop occurs across R2, which is applied as positive bias to the gate of the photofet, and there is a 10.5v drop across R3, which is applied as negative operating voltage to the photofet drain.

Relay RY is a 1000-ohm, 1 ma unit (Sigma 5F, or equivalent). A measurable zero-signal (no-light) current flows through the relay, but it is too small to cause false closure, hence does not need to be balanced out of the relay coil. Illumination of less than 10 foot-candles will cause the relay to close. (The photofet has a lens in its nose.) The circuit draws a maximum of 11 ma from the 12-volt DC supply.

4.10—HEAVY-DUTY PHOTOELECTRIC CONTROL (ALL SOLID-STATE)

In a transistorized photoelectric circuit designed to control large currents, if the load resistance is low with respect to the transistor's internal collector resistance, the electromechanical relay may be dispensed with entirely and an all-solid-state light-sensitive control circuit obtained. Thus all moving parts are eliminated from the control circuit. However, the transistor collector must be able to handle the load current.

Fig. 4-10 shows a photoelectric circuit of this kind. Here, the load device (R_L) is connected directly in the output (collector) circuit of a 2N1906 power transistor. The amount of

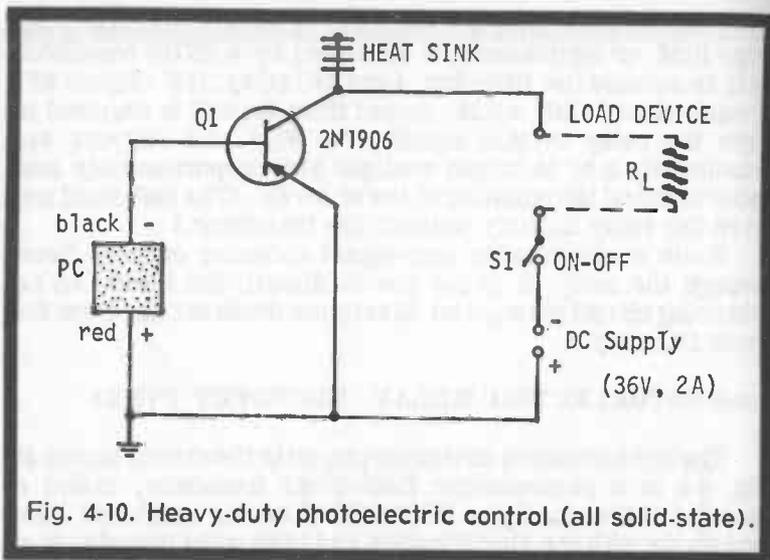


Fig. 4-10. Heavy-duty photoelectric control (all solid-state).

current amplification obtained depends upon the load resistance, varying inversely with R_L . For example, when R_L is 10 ohms or less, the current amplification (for the 2N1906) approaches 125 (which means that the output of the photocell (PC) need be only 1/125 of the required output load current). The maximum DC collector current is 10 amperes for the 2N1906.

In a control circuit of this type, output load current flows as long as the photocell is illuminated. Unlike the electromechanical relay, however, this circuit provides output current proportional to the illumination, and the output will not make and break sharply (in the manner of the electromechanical relay) unless the illumination does also.

For any desired load current (which is identical with the transistor collector current, I_c), the collector-to-emitter voltage (V_{ce}) is equal to the supply voltage (V_{cc}) minus the voltage drop ($I_c R_L$) across the load device: V_{ce} equals V_{cc} minus $I_c R_L$. And V_{ce} must never exceed the maximum safe operating value (40v for the 2N1906). Therefore, the DC supply voltage must be chosen with respect to the load resistance and the maximum safe collector-to-emitter voltage recommended by the transistor manufacturer.

With a self-generating silicon solar cell, PC (International Rectifier Type S1M, or equivalent), in the circuit in Fig. 4-10, 16 ma output from the cell will produce a load current of 2 amp through a 12-ohm load device, R_L . This places 24v across the load device and 12v across the transistor when the DC supply is 36 volts as shown.

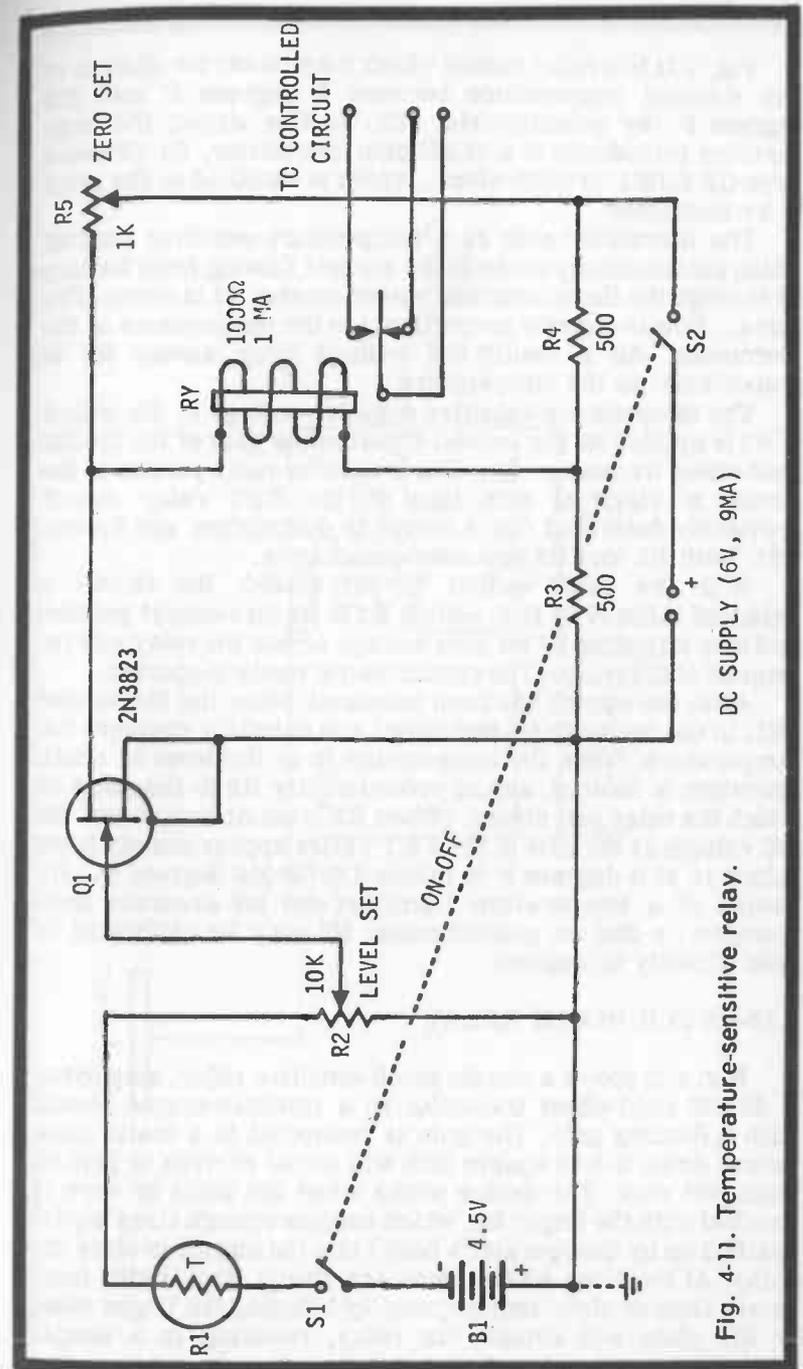


Fig. 4-11. Temperature-sensitive relay.

4.11—TEMPERATURE-SENSITIVE RELAY

Fig. 4-11 is a relay circuit which may be set for closure at any selected temperature between 0 degrees F and 250 degrees F (by potentiometer R2). In this circuit the temperature transducer is a 15,000-ohm thermistor, R1 (Fenwal Type GB42JM1, or equivalent), which is installed at the point to be monitored.

The thermistor acts as a temperature-sensitive resistor which automatically controls the current flowing from battery B1 through the thermistor and potentiometer R2 in series. The current flow is directly proportional to the temperature of the thermistor. As a result the voltage drop across R2 is proportional to the temperature.

The temperature-sensitive negative voltage at the output of R2 is applied as the control signal to the gate of the 2N3823 field-effect transistor (Q1). The transistor-relay portion of the circuit is identical with that of the FET relay circuit previously described (for a complete description, see Section 4.3). Both R2 and R5 are wirewound units.

With the on-off switch (S1-S2) closed, the circuit is balanced initially by first setting R2 to its zero-output position and then adjusting R5 for zero voltage across the relay coil (or dropout of the relay). The circuit then is ready to operate.

After the circuit has been balanced, place the thermistor (R1) in the region to be monitored and carefully measure the temperature. When the temperature is at the level at which operation is desired, adjust potentiometer R2 to the point at which the relay just closes. (When R2 is set at maximum, the DC voltage at the gate of the FET varies approximately from minus 1v at 0 degrees F to minus 4.6v at 250 degrees F.) By means of a temperature chamber and an accurate thermometer, a dial on potentiometer R2 may be calibrated to read directly in degrees.

4.12—TOUCH-PLATE RELAY

Fig. 4-12 shows a simple touch-sensitive relay, employing a 2N3823 field-effect transistor in a common-source circuit with a floating gate. The gate is connected to a metal plate (areas down to 0.06 square inch will work) or even to just an insulated wire. The device works when the plate or wire is touched with the finger tip, which couples enough stray signal (picked up by the operator's body) into the circuit to close the relay. At locations where there are strong stray fields from power lines or other sources, merely bringing the finger close to the plate will actuate the relay, resulting in a simple capacitance relay circuit.

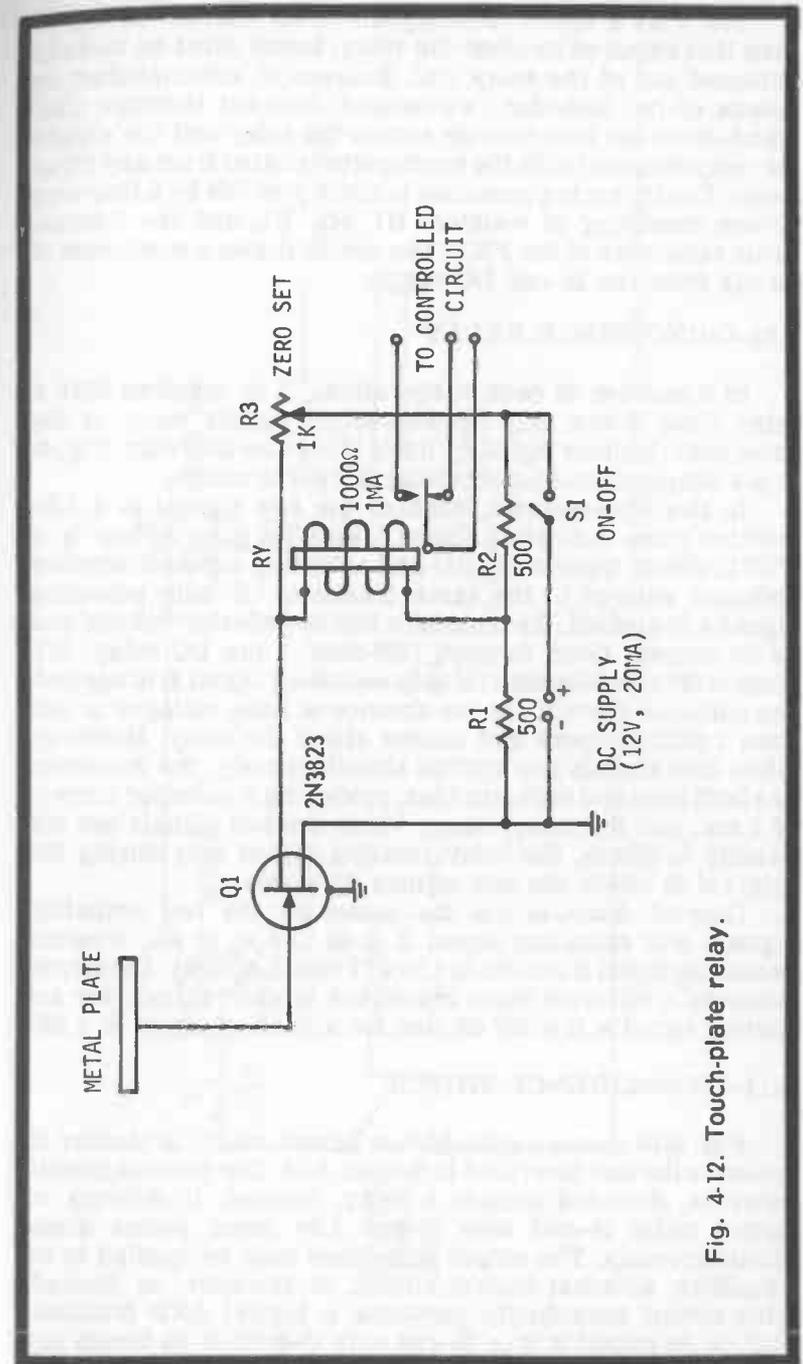


Fig. 4-12. Touch-plate relay.

The FET's static (zero-signal) drain current is higher than that required to close the relay, hence must be initially balanced out of the relay coil. Balance is accomplished by means of the 1000-ohm, wirewound, zero-set rheostat (R3) which is set for zero voltage across the relay coil (or simply for relay dropout) with the touch-plate isolated from any stray fields. The balancing operation is made possible by a four-arm bridge consisting of resistors R1, R2, R3, and the internal drain resistance of the FET. The circuit draws a maximum of 20 ma from the 12-volt DC supply.

4.13-COINCIDENCE RELAY

In a number of control operations, it is required that a relay close if—and only if—two control signals occur at the same time. Neither signal by itself is to close the relay. Fig. 4-13 is a simple circuit for obtaining this performance.

In this arrangement, each of the two signals is a 1.5v positive pulse. Actuating signal A supplies base voltage to a 2N2712 silicon transistor (Q1) and actuating signal B supplies collector voltage to the same transistor. If only actuating signal A is applied, the transistor has no collector voltage and so no current flows through 1000-ohm, 1 ma DC relay, RY (Sigma 5F, or equivalent); if only actuating signal B is applied, the collector current (in the absence of base voltage) is less than 1 microampere and cannot affect the relay. However, when both signals are applied simultaneously, the transistor has both base and collector bias, producing a collector current of 1 ma, and the relay closes. When the two signals are not exactly in phase, the relay remains closed only during the interval in which the two signals do overlap.

Current drain is not the same for the two actuating signals. For actuating signal A it is 1.5v at 80 μ a, whereas actuating signal B results in 1.5v at 1 ma. Similarly, the circuit presents a different input resistance to each signal; for actuating signal A it is 187.5K and for actuating signal B, 1.5K.

4.14-COINCIDENCE SWITCH

Fig. 4-14 shows a coincidence circuit which is similar in action to the one described in Section 4.13. The present circuit, however, does not include a relay. Instead, it delivers an output pulse if—and only if—two 1.5v input pulses occur simultaneously. The output pulse then may be applied to an amplifier, external control circuit, or recorder, as desired. This circuit accordingly performs a logical AND function; that is, its output is true if—and only if—both of its inputs are true.

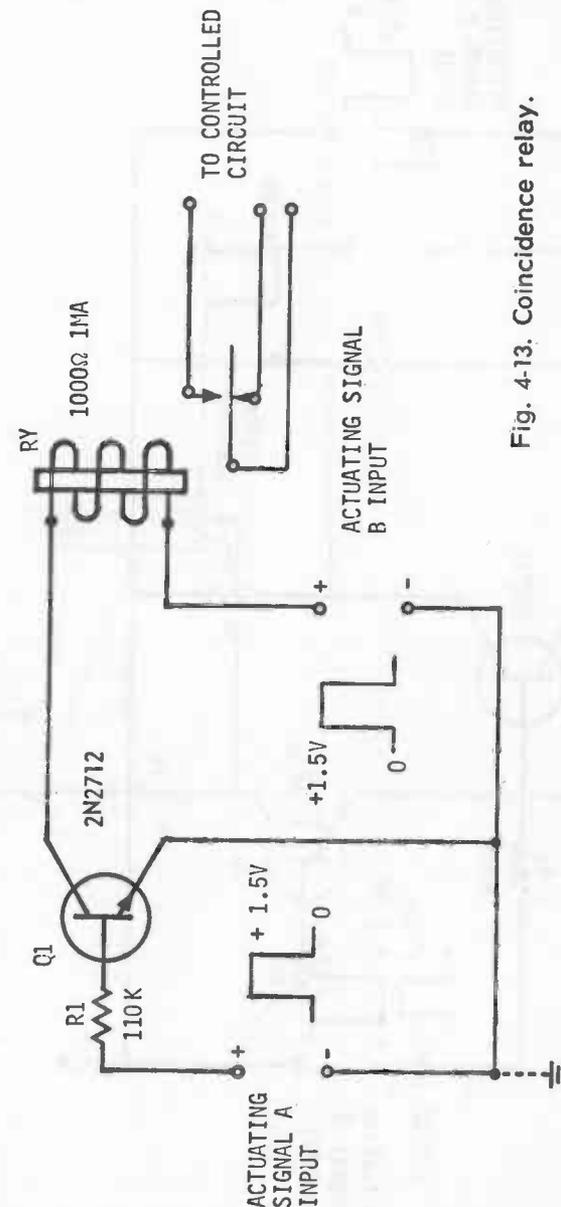


Fig. 4-13. Coincidence relay.

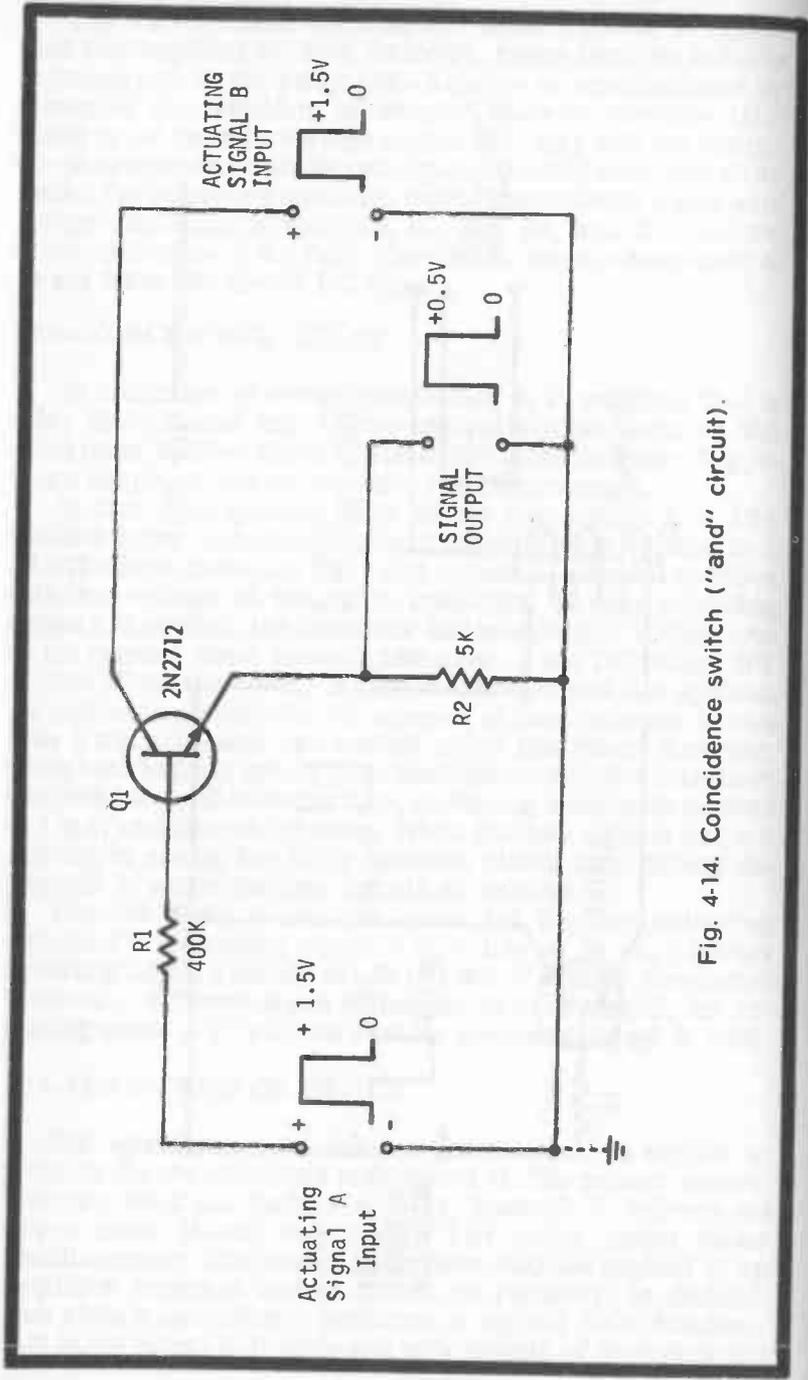


Fig. 4-14. Coincidence switch ("and" circuit).

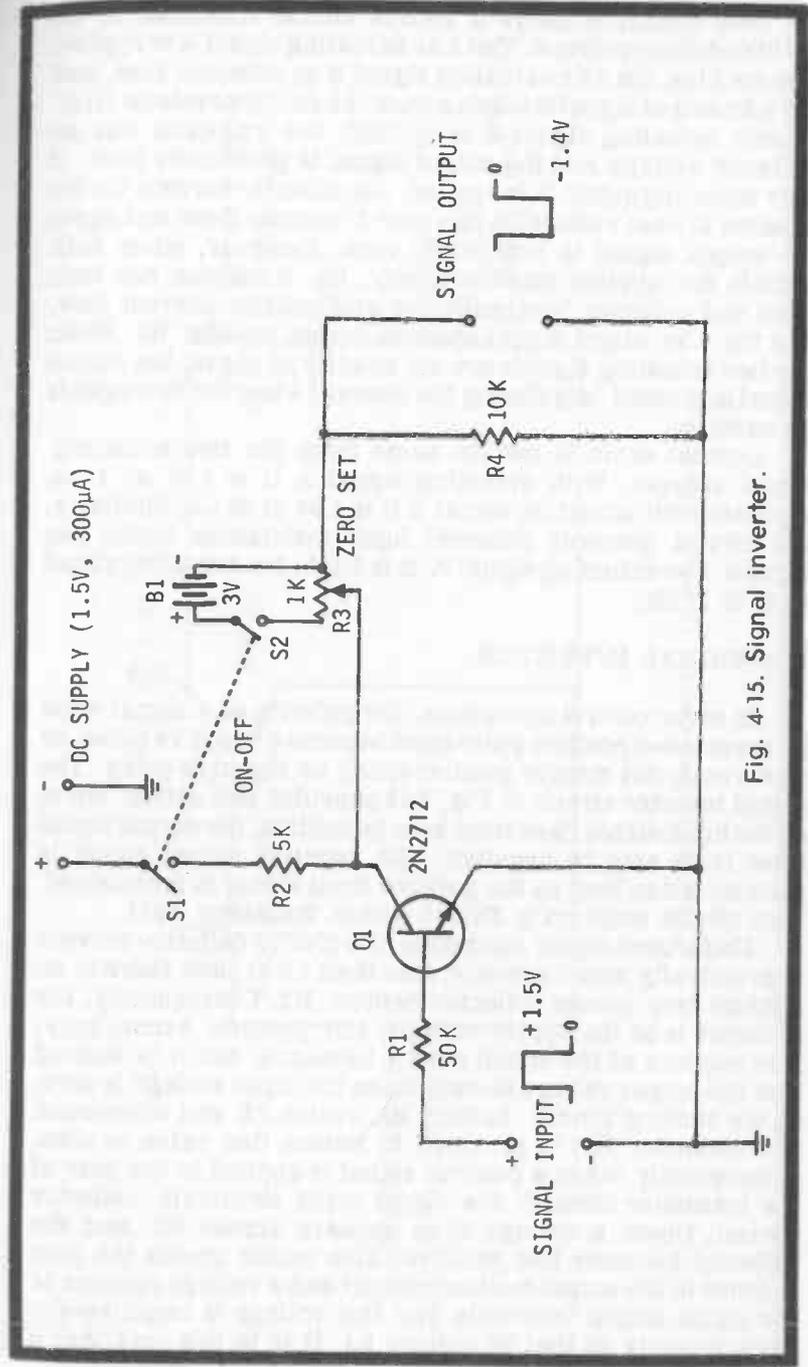


Fig. 4-15. Signal inverter.

This circuit employs a 2N2712 silicon transistor in an emitter-follower circuit. The 1.5v actuating signal A is applied as base bias, the 1.5v actuating signal B as collector bias, and the 0.5v output signal is taken across the emitter resistor (R2). If only actuating signal A is applied, the transistor has no collector voltage and the output signal is practically zero; if only actuating signal B is applied, the collector current (in the absence of base voltage) is less than 1 microampere and again the output signal is practically zero. However, when both signals are applied simultaneously, the transistor has both base and collector bias, collector and emitter current flow, and the 0.5v output signal appears across resistor R2. When the two actuating signals are not exactly in phase, the output signal is present only during the interval when the two signals do overlap.

Current drain is not the same from the two actuating-signal sources. With actuating signal A it is 1.5v at 1 μ a, whereas with actuating signal B it is 1.5v at 85 μ a. Similarly, the circuit presents different input resistances to the two signals. For actuating signal A, it is 1.5M; for actuating signal B, it is 17.6K.

4.15—SIGNAL INVERTER

In some control operations, the polarity of a signal must be reversed—a positive pulse must become a negative pulse, or vice versa, not merely positive-going or negative-going. The signal inverter circuit in Fig. 4-15 provides this action. Here, as the input signal rises from zero to positive, the output signal rises from zero to negative. The negative output signal is maintained as long as the positive input signal is maintained. This circuit employs a 2N2712 silicon transistor (Q1).

Under zero-signal conditions, the 2N2712 collector current is practically zero (actually, less than 1 μ a), and there is no voltage drop across collector resistor R2. Consequently, the collector is at the supply voltage, 1.5v positive. Accordingly, 1.5v appears at the signal output terminals, but it is desired that the output voltage be zero when the input voltage is zero, so the bucking circuit (battery B1, switch S2, and wirewound potentiometer R3) is provided to reduce this value to zero. Subsequently, when a positive signal is applied to the base of the transistor through the signal input terminals, collector current flows, a voltage drop appears across R2 and the collector becomes less positive. This action upsets the zero balance in the output bucking circuit and a voltage appears at the signal output terminals. But this voltage is negative—the same polarity as that of battery B1. It is in this way that a

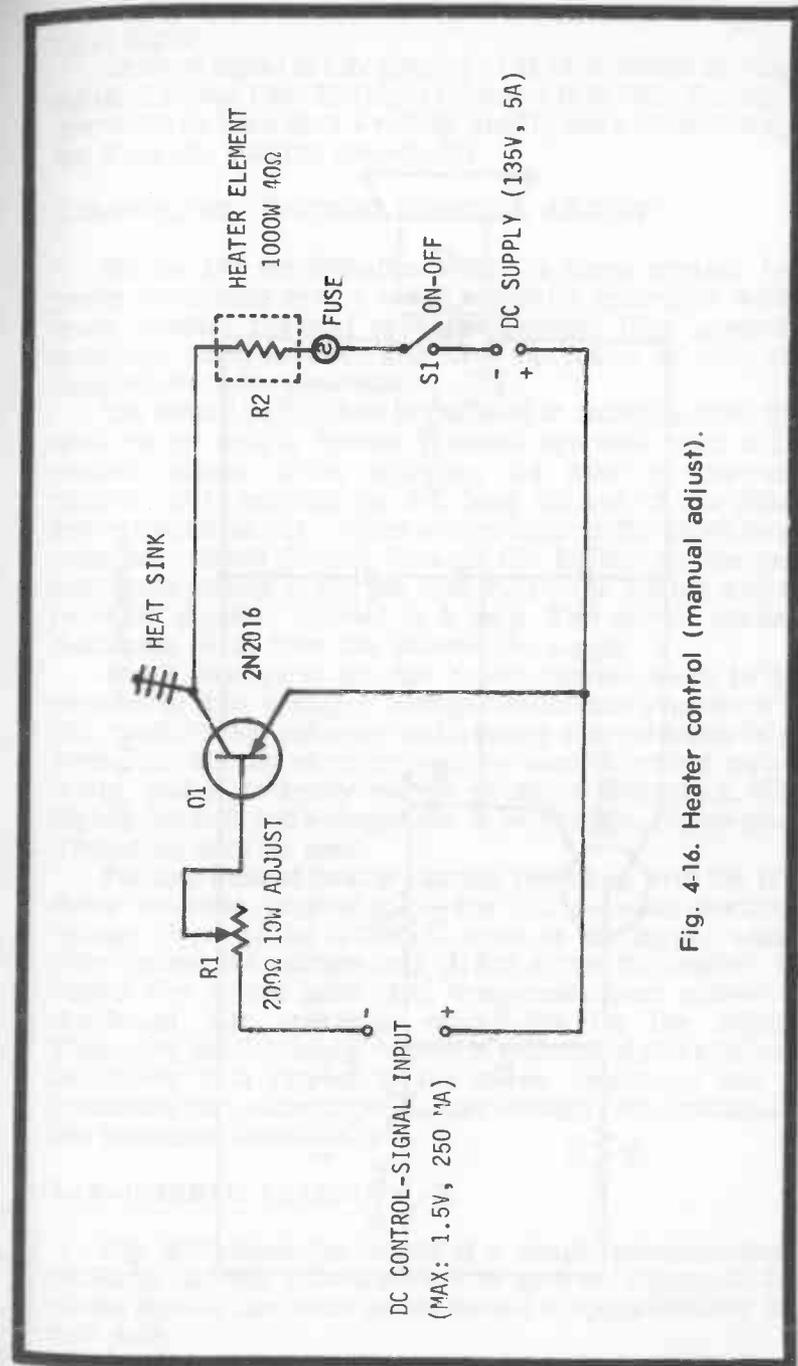


Fig. 4-16. Heater control (manual adjust).

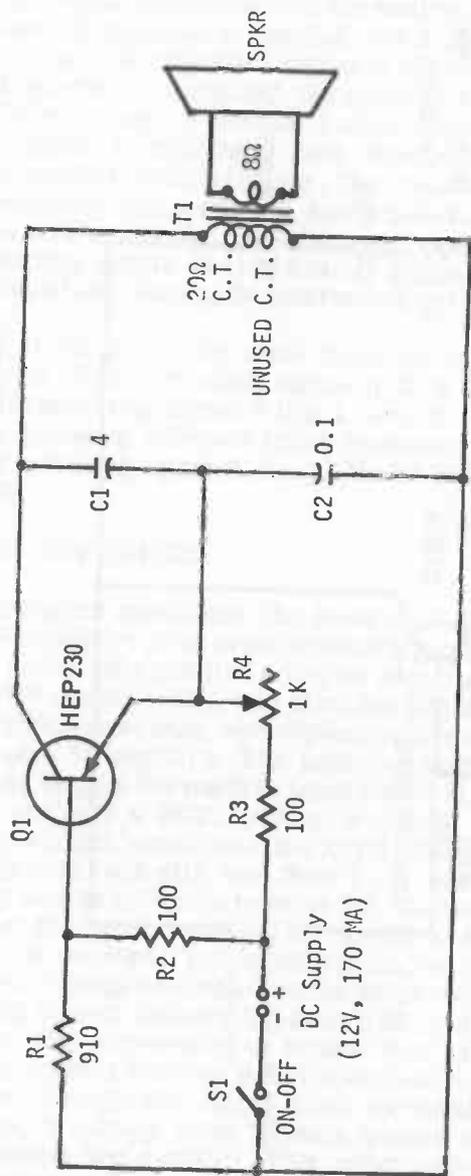


Fig. 4-17. Audible alarm.

negative output signal is obtained in response to a positive input signal.

An input signal of 1.5v positive at 16 μ a produces an output signal of minus 1.4v. The input resistance is 93.75K. The circuit draws 300 μ a from the 1.5-volt DC supply and approximately 3 ma from the bucking battery, B1.

4.16-HEATER CONTROL (MANUAL ADJUST)

Simple DC amplification allows a large current (and power-consuming device to be manually controlled with a much smaller rheostat or potentiometer, thus promoting economy, compactness, and cool operation of the continuously-variable controller.

The circuit in Fig. 4-16 is designed to control a 200v, 1000-watt heater with a 10-watt rheostat operated from a 1.5v control voltage. Here, adjusting the 200-ohm wirewound rheostat (R1) controls the DC base current of the 2N2016 power transistor (Q1), which in turn controls the much larger collector current flowing through the heater. At the zero-resistance setting of R1, the base current is 250 ma and the collector (heater) current is 5 amp. The circuit draws a maximum of 5A from the 135-volt DC supply.

At all settings of R1, the heater current tends to hold constant in spite of supply voltage fluctuations because of the flat, pentode-like, collector volt/ampere characteristic of the transistor. Similar circuitry may be used to control motors, lamps, and other heavy-current electrical equipment. When higher currents and voltages are to be handled, higher-power transistors may be used.

For any desired heater current (identical with the transistor collector current, I_c), the DC collector-to-emitter voltage (V_{ce}) of the 2N2016 is equal to the supply voltage (V_{cc}) minus the voltage drop ($I_c R_2$) across the heater: V_{ce} equals V_{cc} minus $I_c R_2$. And V_{ce} must never exceed the maximum safe operating value (65v for the 2N2016). Therefore, the DC supply voltage in a circuit of this kind must be chosen with respect to the heater resistance and the maximum safe collector-to-emitter voltage recommended by the transistor manufacturer.

4.17-AUDIBLE ALARM

Fig. 4-17 shows the circuit of a simple power-transistor oscillator driving a loudspeaker to give an adjustable-tone alarm signal. The audio power output is approximately one-half watt.

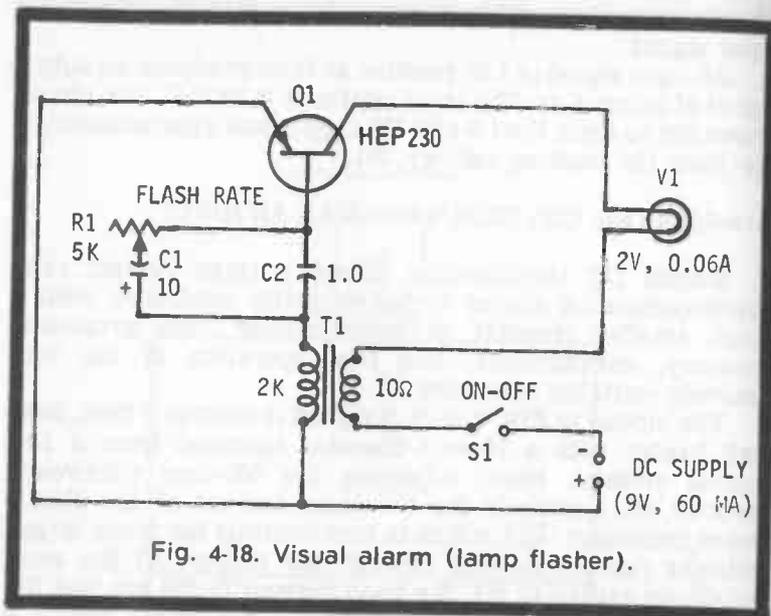


Fig. 4-18. Visual alarm (lamp flasher).

A Colpitts oscillator circuit is employed with an HEP 230 transistor. In this circuit the operating frequency is determined principally by capacitances C1 and C2 and the inductance of the primary winding of the 20-ohm (center tapped)-to-8-ohm output transformer T1 (Chicago-Stancor TA-12, or equivalent). The frequency is adjustable also from approximately 300 Hz to 3 kHz by positioning the tap on 1000-ohm, 50-watt, wirewound resistor R4. R3 is a 50-watt and capacitor C1 must be a non-electrolytic. The circuit draws approximately 170 ma from the 12-volt DC supply.

4.18-VISUAL ALARM (LAMP FLASHER)

The circuit in Fig. 4-18 provides bright flashes from a small radio-type pilot lamp. Backed up with a reflector, the lamp can give flashes that are visible for an appreciable distance. The flashing rate is continuously variable from 1 to 20 flashes per second by adjustment of the 5000-ohm wirewound rheostat, R1.

A blocking oscillator circuit is employed, with an HEP 230 power transistor (Q1) operated in conjunction with a miniature 2000-ohm-to-10-ohm output transformer, T1 (Lafayette 99T6101, or equivalent). The lamp (V1) is a 2v, 60-ma (brown-bead) pilot lamp. The circuit draws 60 ma from the 9-volt DC supply.

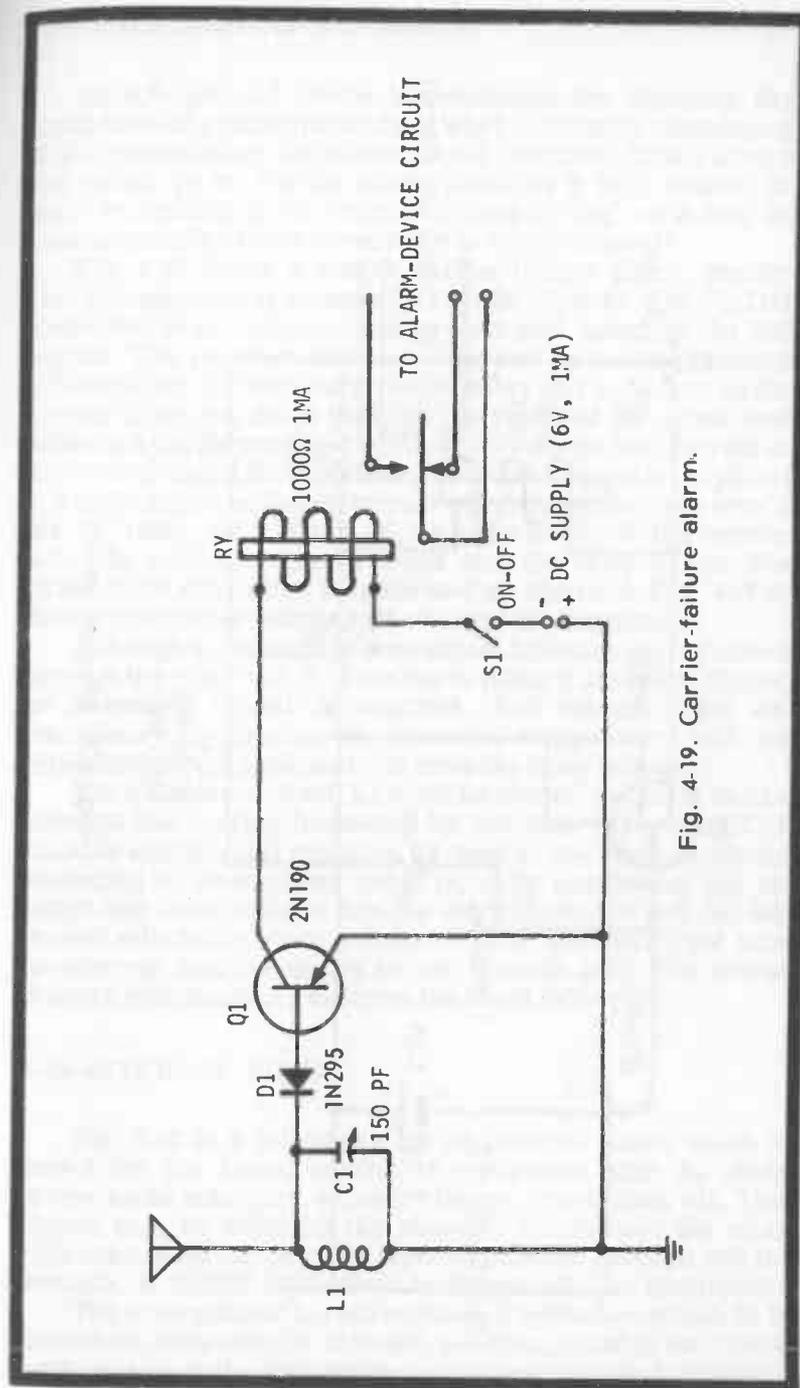


Fig. 4-19. Carrier-failure alarm.

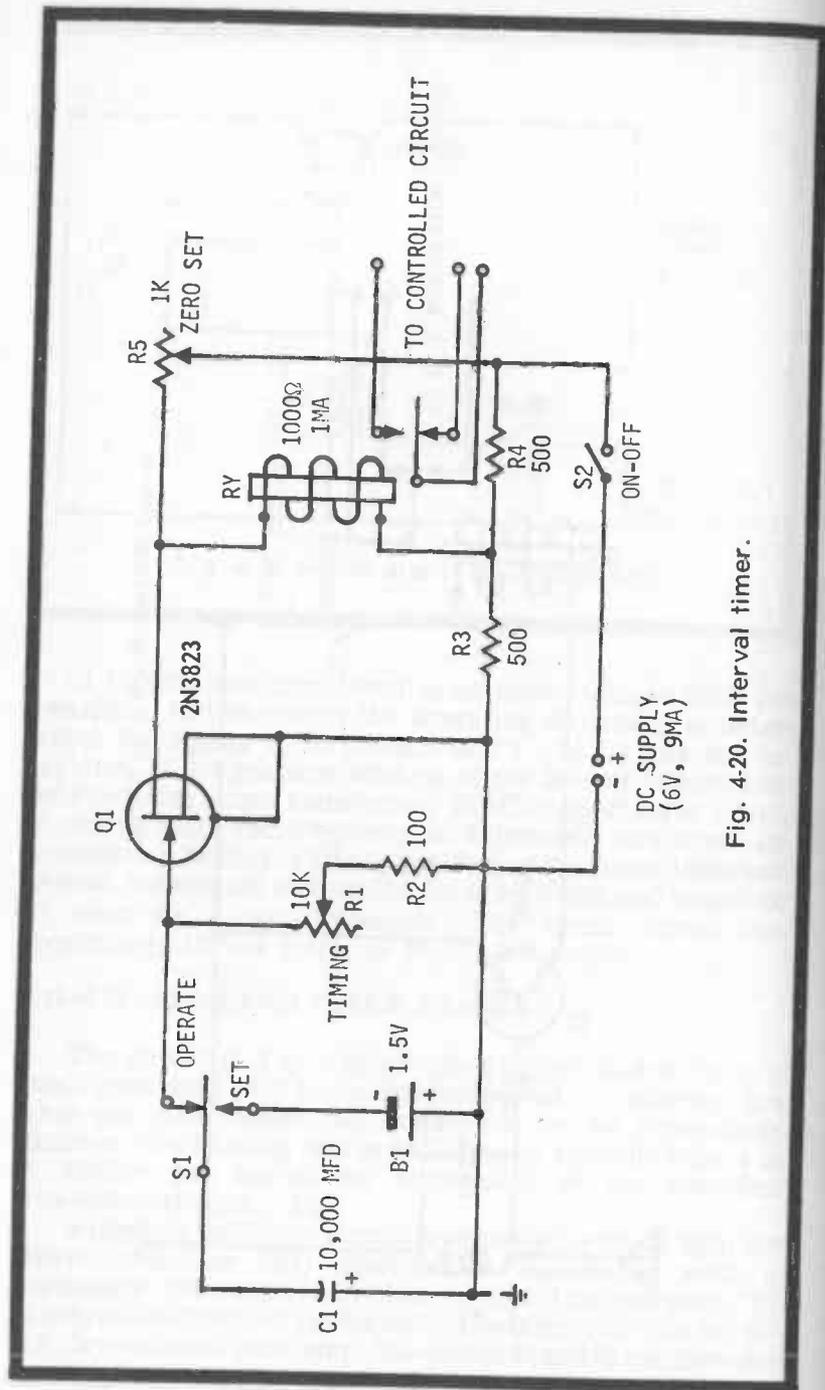


Fig. 4-20. Interval timer.

4.19-CARRIER-FAILURE ALARM

An RF-operated device is convenient for signaling the breakdown of a radio transmitter when continuous monitoring of the transmissions themselves is not desirable. Such a device can switch on an audible alarm (such as a bell, buzzer, or horn) or a visual device (such as a lamp or flag), or it may be used to operate a timer or recorder to log the time off.

Fig. 4-19 shows a simple carrier-failure alarm circuit. This arrangement is essentially a diode receiver (L1, C1, D1) connected to an antenna and ground and tuned to the RF carrier. The receiver section is followed by a transistorized DC amplifier (Q1) and sensitive DC relay (RY). As long as the carrier is on, the diode rectifies the received RF signal and delivers a corresponding negative DC voltage and current to the base of the 2N190 transistor. The base current is amplified to a much larger collector current which closes the 1000-ohm, 1 ma DC relay, RY (Sigma 5F, or equivalent). If the carrier fails, the collector current drops and the relay opens. The 1N295 diode (D1) must be polarized as shown in Fig. 4-19 to deliver a negative voltage to the base of the transistor.

Although a measurable zero-signal collector current flows through the relay coil, it is too low to disturb the relay. Hence, no balancing circuit is required. And except under extraordinary circumstances, increased temperature will not raise this current sufficiently to close the relay falsely.

The inductance of coil L1 must be chosen so that it can be tuned to the carrier frequency by variable capacitor C1. A suitable commercial coil may be used or one may be wound according to instructions found in radio handbooks. No attempt has been made to tap the diode down the coil for improved selectivity, since it is better that this circuit not tune too sharply, lest the station be lost through drift. The circuit draws a maximum of 1 ma from the 6-volt DC supply.

4-20-INTERVAL TIMER

Fig. 4-20 is a schematic for an interval timer which is useful for the timed control of equipment such as photo printers and enlargers, exposure lamps, irradiators, etc. This circuit may be adjusted (by rheostat R1) to hold the relay (RY) closed for any selected interval between 1 second and 101 seconds. A 2N3823 field-effect transistor (Q1) is employed.

The arrangement is conventional. Pushbutton switch S1 is depressed momentarily to its set position, causing the 10,000-mfd capacitor (C1, Mallory Type 10100, or equivalent) to charge

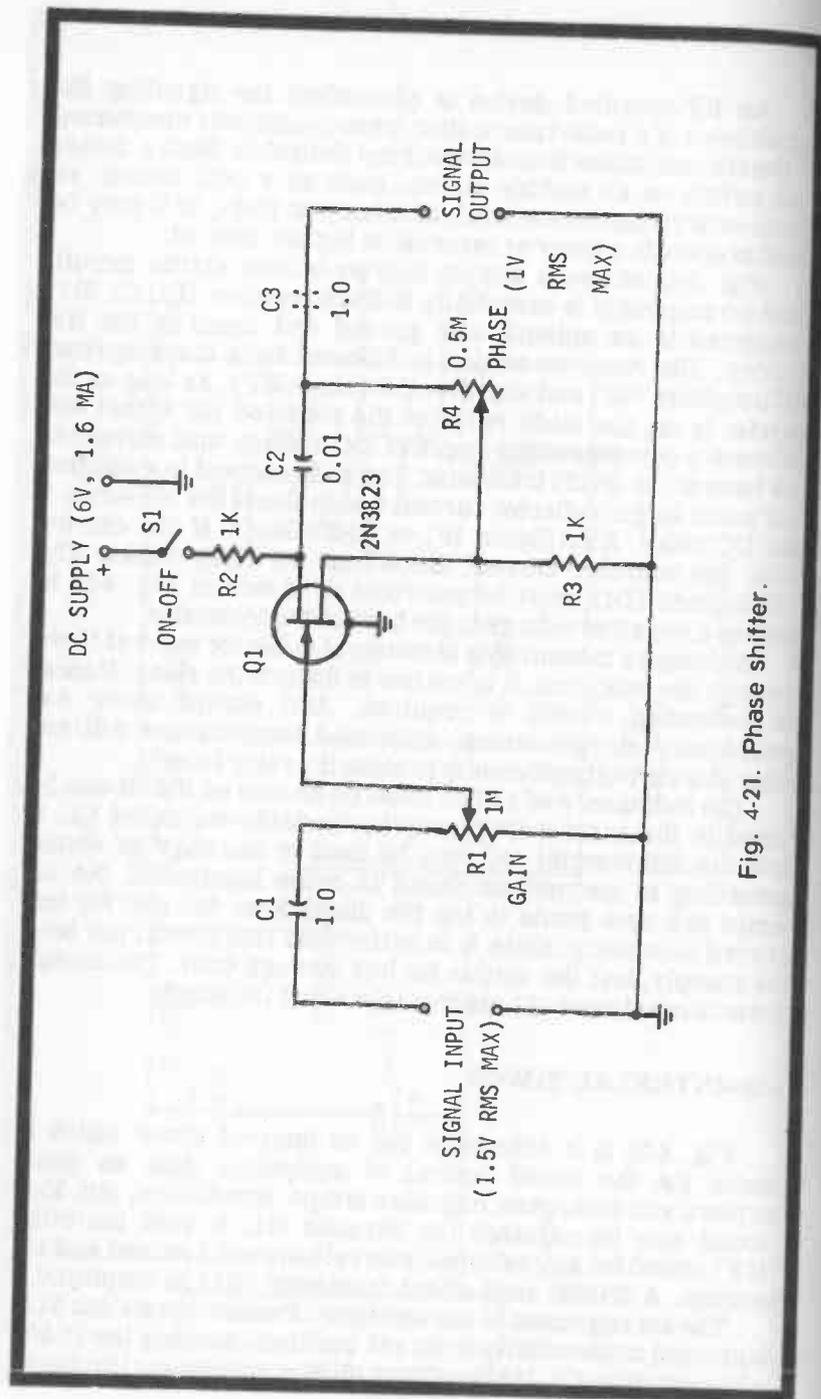


Fig. 4-21. Phase shifter.

from the 1.5v Size-D flashlight cell, B1. When S1 is released, it springs back to its operate position and C1 discharges through the 10,000-ohm wirewound rheostat (R1) and 100-ohm resistor (R2) in series. The discharge time constant is t equals 10^{-6} (R1 plus R2)C seconds, where the Rs are in ohms and C in microfarads. The FET monitors the DC voltage drop produced across R1 and R2 by the capacitor discharge current which actuates the relay time. Thus, the relay is picked up as soon as S1 returns to operate and drops out when C1 has discharged sufficiently to reduce the R1 plus R2 voltage drop to a critical value. The interval of closure depends upon the setting of rheostat R1 and may be varied from 1 sec when R1 is set to zero (leaving only R2 in the circuit) to 101 sec when R1 is set to 10,000 ohms. The high input resistance of the FET insures that the RC timing circuit will not be adversely shunted by the relay amplifier.

In the zero-signal state (C1 completely discharged), the static drain current of the FET, flowing through the coil of the 1000-ohm, 1 ma DC relay, RY (Sigma 5F, or equivalent), would keep the relay closed falsely. Therefore, a balancing circuit is provided to buck this current initially out of the relay: With S1 at operate and C1 completely discharged, 1000-ohm wirewound rheostat R5 is adjusted for zero voltage across the relay coil (or simply for relay dropout).

The circuit draws a maximum of 9 ma from the 6-volt DC supply. The 1.5v cell (B1) will give long service, owing to its intermittent use only to charge capacitor C1.

A precision stopwatch or other time standard may be used to calibrate the dial of the timing control (R1) directly in seconds, so that the rheostat may be set quickly to any desired time interval within its range.

4.21—PHASE SHIFTER

Fig. 4-21 shows the circuit of a continuously-variable phase shifter. Unlike the non-electronic shifter circuits, this one maintains a constant output amplitude over its entire phase-angle and frequency range. Adjustment of the 0.5-megohm phase control rheostat (R4) shifts the frequency over the range 0 degrees to 180 degrees at any frequency between 100 Hz and 5 kHz. (At frequencies outside of that range, stray reactances in the components and wiring restrict the phase range.)

The 1-megohm gain control potentiometer (R1) permits smooth control of the signal output amplitude. With R1 set for maximum gain, the maximum input-signal amplitude before

output-signal peak clipping is 1.5v RMS and the corresponding maximum no-load output-signal amplitude is 1v RMS.

Resistors R2 and R3 must be precisely matched. When the circuit must be DC-isolated from the load (external device), a 1-mfd blocking capacitor (C3) may be inserted in the output line. The phase shifter draws 1.6 ma from the 6-volt DC supply.

Chapter 5

Test Instruments

Since its very first appearance, the transistor has intrigued the designers of test equipment. Its size, weight, and DC power advantages—together with its high overall efficiency—made possible for the first time self-contained portable instruments capable of economical battery operation. And these same advantages of the transistor have effected dramatic reductions of size, weight, and heating in many stationary, power-line-operated instruments, as well. Today, there is scarcely any type of electronic test instrument which has not been transistorized in some model.

The circuits described in this chapter give some idea of the range of possibilities attainable with single-transistor instruments. Each circuit is useful in the form in which it is presented; nevertheless, any one of them may be combined in a number of ways with circuits given in other chapters to obtain special-purpose systems.

Current, voltage, signal amplitude, frequency, and phase values given in this chapter are those obtained in tests of the author's models and may vary somewhat with individual transistors and circuit components. Where particular components from specified manufacturers are listed, those components (or their exact equivalents, when obtainable) appear to be necessary to proper performance of the circuit. In all other instances, any component having the specified electrical characteristics may be used. Except where shown otherwise on the diagram or in the text, all resistances are in ohms and capacitances in microfarads; resistors are one-half watt, and capacitors are 25 DCWV. Other circuits which might also be used as test instruments are found in Sections 1.9, 1.13, 1.14, 1.18, 1.19, 1.20, 2.5, 2.6, 3.1-3.4, 3.6-3.11, 3.16, 3.19, 3.20, 4.14, 4.20, and 4.21.

5.1—DC MICROAMMETER (NPN)

A transistor DC amplifier ahead of a DC milliammeter converts the latter into a microammeter. With an electronic

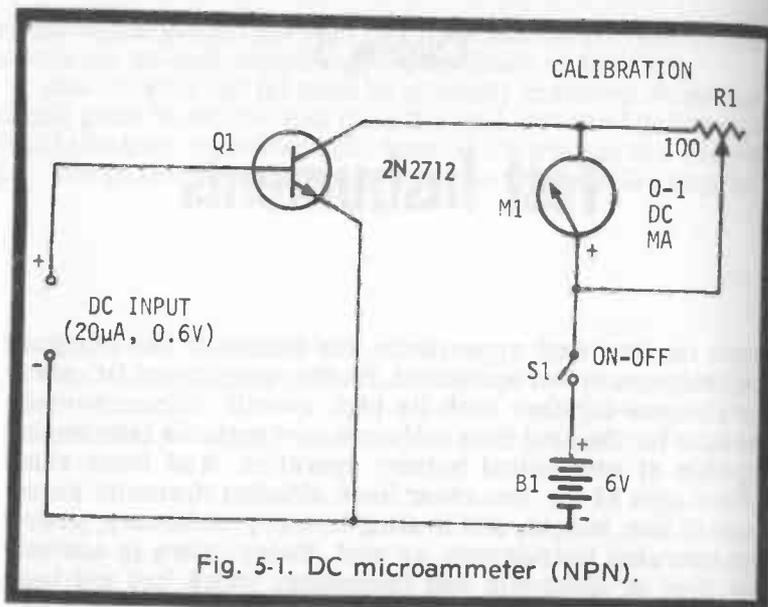


Fig. 5-1. DC microammeter (NPN).

microammeter of this type, the indicating meter is comparatively rugged. Transistorized DC microammeter circuits are shown in Figs. 5-1 and 5-2.

The circuit in Fig. 5-1 employs a Type 2N2712 NPN silicon transistor. A DC input of $20\ \mu\text{A}$ at 0.6V will deflect 0-1 DC milliammeter M1 to full scale. Response of the circuit is linear.

The static (zero-signal) collector current of the transistor is so low (less than $1\ \mu\text{A}$) that it cannot be seen on the meter scale, even when the circuit is operated at extremes of temperature. Hence, no zero-set control is required.

The 100-ohm wirewound rheostat (R1) is a calibration control. To adjust this control, feed an accurately known $20\ \mu\text{A}$ direct current into the DC input and adjust R1 for exact full-scale deflection of the meter.

The input resistance of this circuit is 30K , a comparatively high value which limits the instrument to uses where such a resistance and the accompanying 0.6-volt drop can be tolerated. The circuit draws a maximum of $3\ \text{ma}$ from the 6-volt battery (B1).

5.2—DC MICROAMMETER (PNP)

Fig. 5-2 is a microammeter circuit employing a Type 2N190 PNP germanium transistor. In this arrangement, a DC input of $20\ \mu\text{A}$ at 0.13V will deflect 0-1 DC milliammeter M1 to full scale. The circuit response is linear.

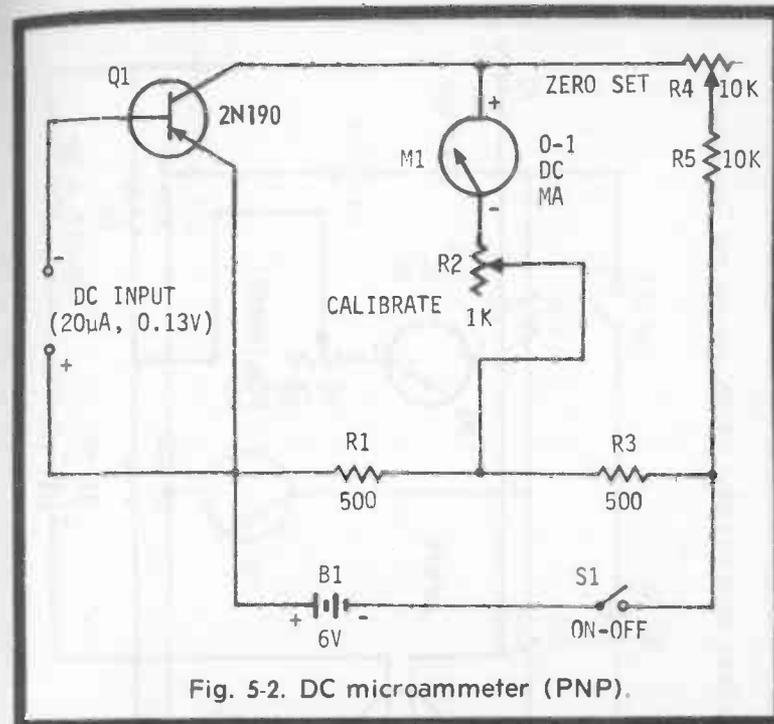


Fig. 5-2. DC microammeter (PNP).

With the DC input terminals open, the meter is set to zero by the $10,000\text{-ohm}$ wirewound rheostat (R4), which balances the static collector current of the transistor out of the meter.

After being zeroed, the circuit is calibrated by feeding an accurately known $20\ \mu\text{A}$ into the DC input and adjusting the 1000-ohm wirewound rheostat (R2) for exact full-scale deflection of the meter. It may be necessary to work back and forth between the R2 and R4 adjustments.

The input resistance of the circuit is 6.5K , a condition which limits the instrument to uses where this resistance and the accompanying 0.13-volt drop can be tolerated. The circuit draws a maximum of $6.5\ \text{ma}$ from the 6-volt battery (B).

5.3—ELECTRONIC DC VOLTMETER

Fig. 5-3 shows the circuit of a transistorized voltmeter which is comparable to a VTVM, offering as it does a constant input resistance of better than $12\ \text{megohms}$ (11.11M in the input divider, R2 through R5, and 1M in the probe). This instrument provides four ranges: 0-1, 0-10, 0-100, and 0-1000 volts, indicated by a 0-1 DC milliammeter, M1.

The high input resistance is achieved through the use of a 2N3578 field-effect transistor (Q1). The voltage under test is

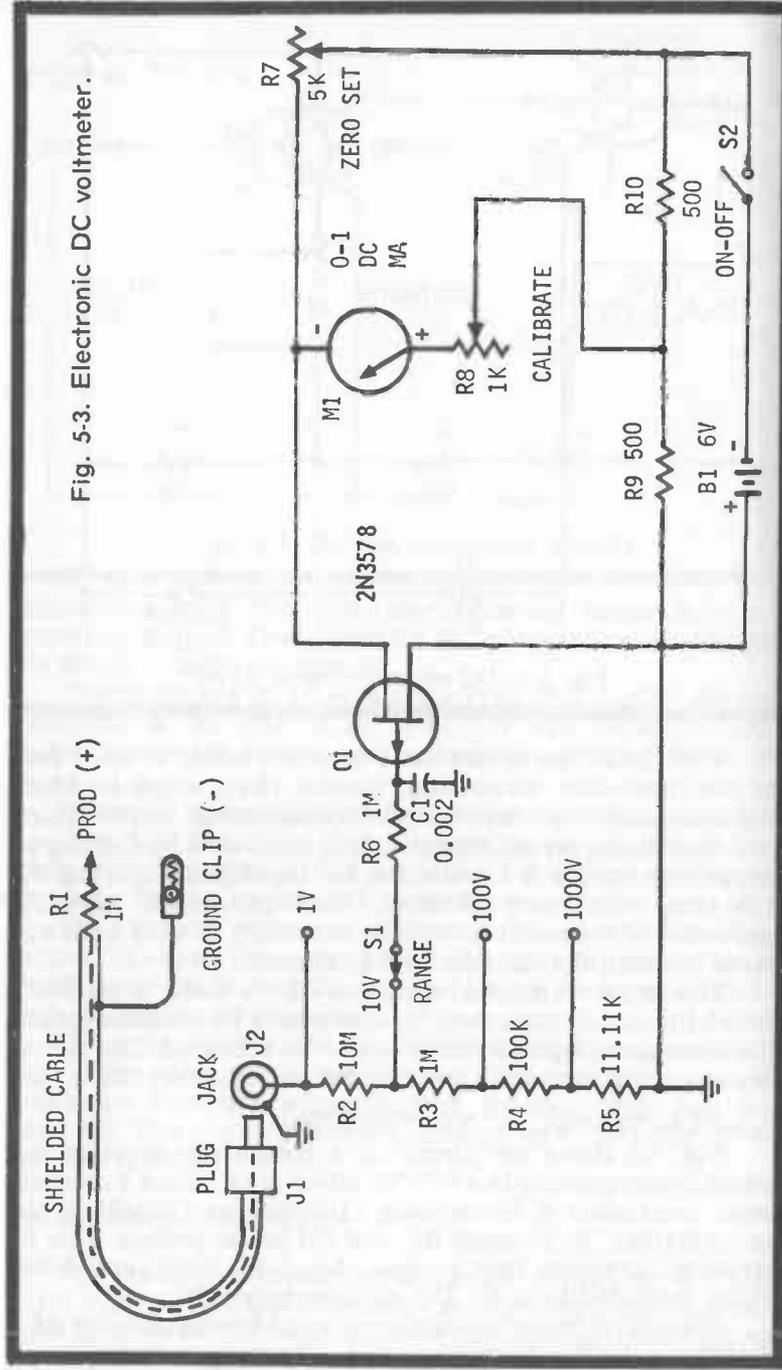


Fig. 5-3. Electronic DC voltmeter.

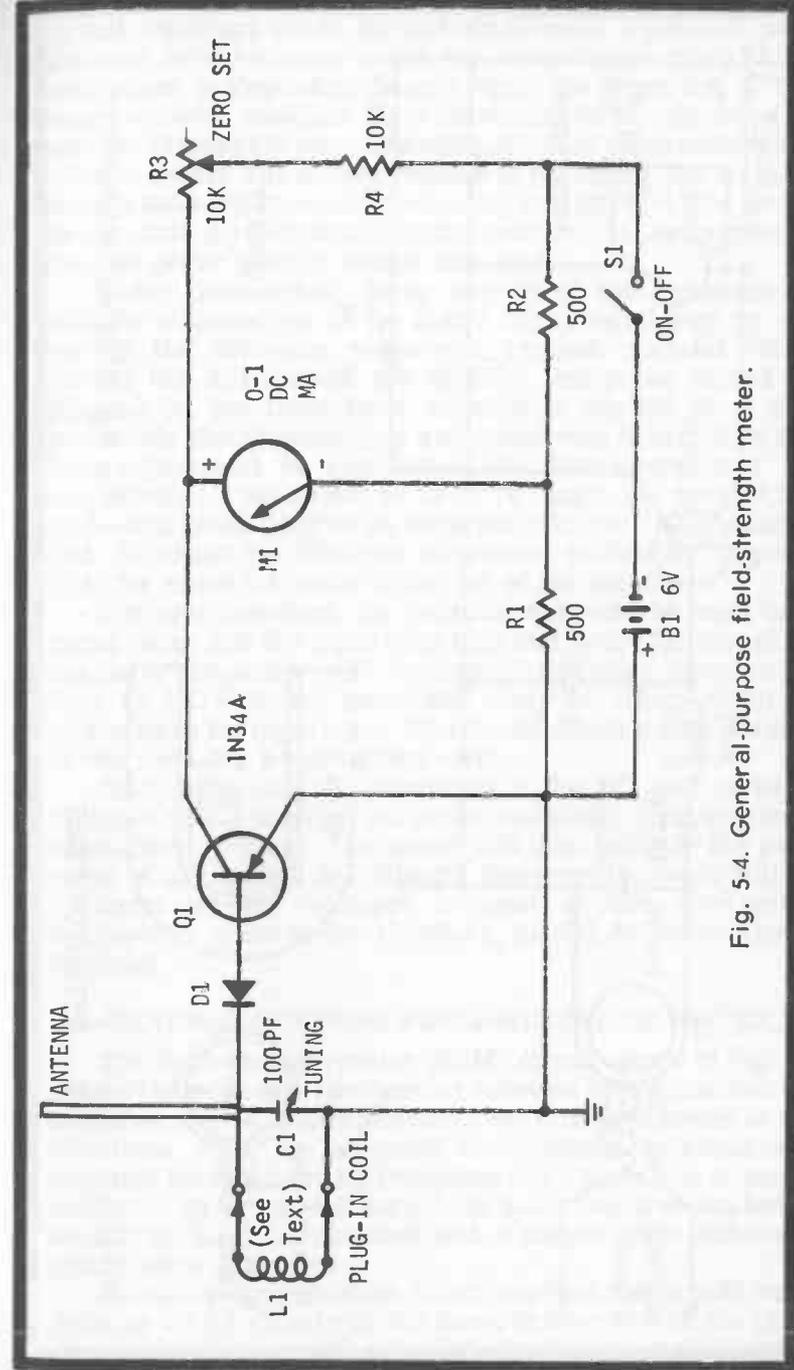


Fig. 5-4. General-purpose field-strength meter.

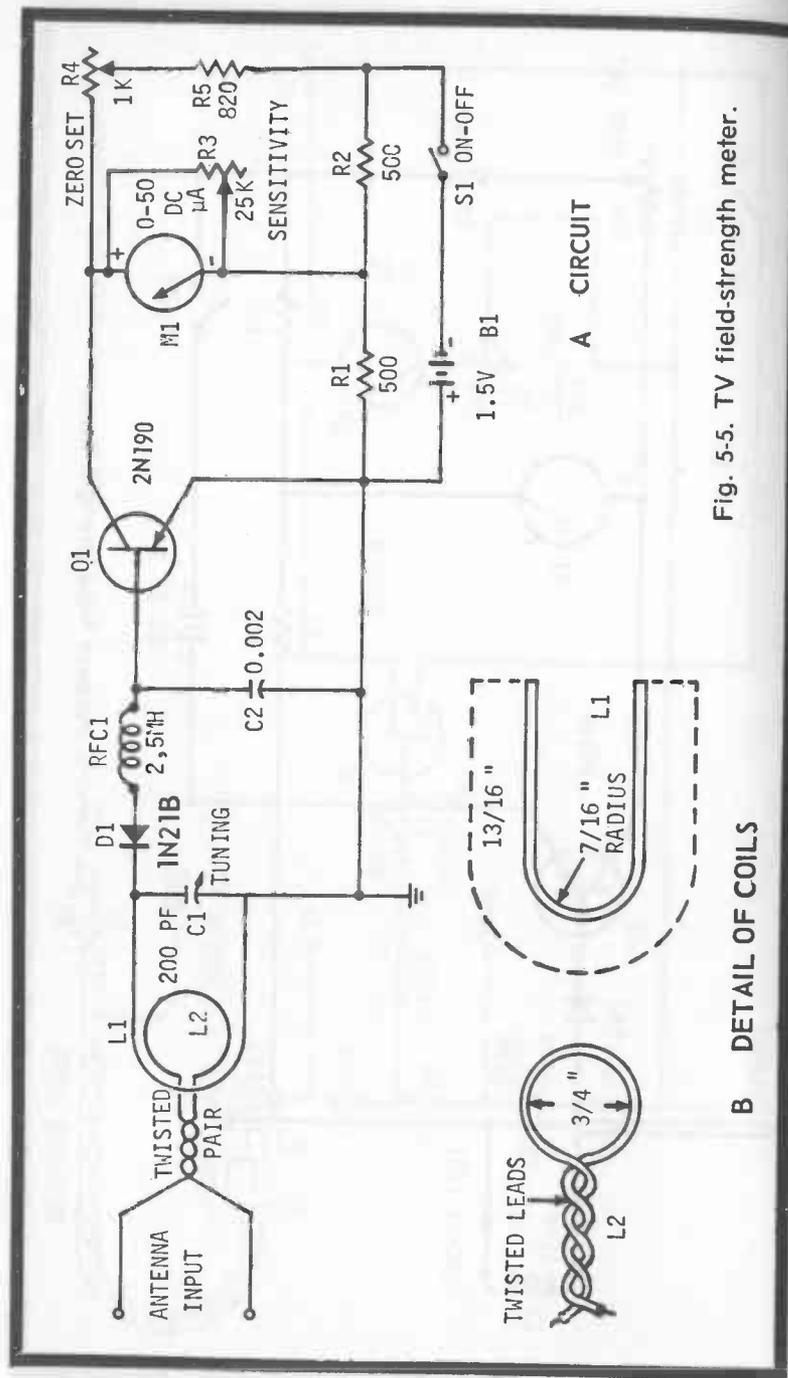


Fig. 5-5. TV field-strength meter.

applied to the transistor gate through the range-switching circuit (resistors R2 to R5 and single-pole, 4-position, non-shorting, rotary selector switch S1). Since the accuracy of the instrument is dependent largely upon the precision of the range selector, resistors R2 to R5 should be closely rated (1 percent tolerance is recommended). Each of these resistors is rated at 1-watt. The bottom resistor in the string (R5, 11.11K) may be made up by series-connecting an 11,000-, a 100-, and a 10-ohm unit. A 1-megohm isolating resistor (R1) is inserted in the test probe directly behind the prod.

Static (zero-signal) drain current of the transistor is initially balanced out of the meter in the usual way by adjusting the 5000-ohm wirewound zero-set rheostat (R7). During this adjustment, the shielded test probe should be plugged in, but must have no voltage applied to it and preferably should be clear of any interfering fields. After the zero adjustment is completed, the instrument may be calibrated: (1) Set switch S1 to its 1v range; (2) connect the probe and ground clip to an accurately known 1v DC source, and (3) adjust the 1000-ohm wirewound calibration rheostat (R8) for exact full-scale deflection of the meter.

For best shielding, the instrument should be built in a metal case, and the input plug (J1) and jack (J2) should be coaxial or concentric units. Resistor R6 and mica capacitor C1 form an RC filter for removing stray AC which might be picked up by the input leads. This circuit draws a maximum of 10 ma from the 6-volt battery (B1).

AC voltage may be measured if the DC test probe is replaced with a shielded AC probe containing a conventional shunt-diode rectifier. The meter will then indicate the peak value of the applied AC voltage; however, the scale will be nonlinear at low voltages because of the low-current nonlinearity of the diode; therefore, special AC calibration is required.

5.4—GENERAL-PURPOSE FIELD-STRENGTH METER

The field-strength meter (FSM) circuit shown in Fig. 5-4 consists of a tuned diode detector followed by a transistor DC amplifier and DC milliammeter. This circuit is useful in applications, such as adjusting transmitting antennas and checking the approximate frequency of a transmitter or power oscillator, in which high sensitivity is not required but better sensitivity than that obtained with a simple diode-and-meter combination is needed.

A small pickup antenna (short vertical rod or stiff wire) picks up the RF signal which is tuned-in sharply with the L1-C1 circuit. Diode D1 rectifies the RF signal voltage and applies

the resulting negative DC voltage to the base of the 2N190 transistor (Q1). The base current is amplified by the transistor which then deflects the 0-1 DC milliammeter (M1).

The zero-signal (static) collector current of the transistor is initially balanced out of the meter by adjusting the 10,000-ohm wirewound zero-set rheostat (R3) in the absence of any RF pickup. The circuit then will remain zeroed for long periods.

Inductor L1 is selected to resonate, with 100-pf tuning capacitor C1, to the carrier frequency of interest. This may be a commercial coil or a home-made one wound according to directions given in Table 5-1.

TABLE 5-1

COIL A	1.8 to 4 MHz	57 turns	No. 32 enameled wire closewound on a 1-inch diameter form.
COIL B	3.8 to 4.6 MHz	25 turns	No. 26 enameled wire on a 1-inch diameter form. Space to a winding length of one-half inch.
COIL C	8 to 18 MHz	12 turns	No. 22 enameled wire on 1-inch diameter form. Space to a winding length of one-half inch.
COIL D	15 to 34 MHz	5½ turns	No. 22 enameled wire on a 1-inch diameter form. Space to a winding length of one-quarter inch.
COIL E	30 to 68 MHz	2½ turns	No. 22 enameled wire on a 1-inch diameter form. Space to a winding length of one-quarter inch.

5.5—TV FIELD-STRENGTH METER

The field strength meter circuit in Fig. 5-3A is designed for orienting TV receiving antennas. The circuit is similar to the one described in Section 5.4, except for the special coupler (L1-L2) and the DC microammeter (M1). It makes a lightweight instrument for rooftop use and is entirely adequate when the field strength of the received signals is 1 millivolt and higher.

The input circuit is continuously tunable, by means of 200-pf midget variable capacitor C1, from Channel 2 through Channel 13 (specifically, from 40 MHz to 260 MHz). L1 is a hairpin loop made from No. 20 bare copper wire and L2 is a single turn of insulated hookup wire (dimensions of these inductors are given in Fig. 5-5B). Coupling coil L2 is mounted rigidly inside L1, as shown in Fig. 5-5A.

For operation at the high TV frequencies, D1 is a point-contact silicon microwave diode, Type 1N21B, but it is inexpensive. The negative DC output current of the diode is fed into the base of the 2N190 DC amplifier transistor (Q1) through a radio-frequency filter (RFC1-C2). This current is amplified by the transistor which then deflects the microammeter.

The zero-signal (static) collector current of the transistor is initially balanced out of the meter by adjusting the 1000-ohm

wirewound zero-set rheostat (R4) in the absence of any RF pickup. The circuit should remain zeroed for long periods. The 25,000-ohm rheostat (R3) serves as a sensitivity (or gain) control. DC operating power is supplied by a 1½-volt, size-D flashlight cell (B1). Maximum drain is 3 ma.

5.6—DIP OSCILLATOR (PNP)

The utility of the grid-dip oscillator (GDO) is well known to electronic workers of all classes, and it is increased by the transistor which divorces the dip oscillator (DO) completely from the power line and often permits a dramatic reduction in size and weight. Transistorized dip-oscillator circuits are shown in Figs. 5-6 and 5-7.

Fig. 5-6 shows a DC circuit employing a 2N1178 high-frequency PNP transistor in a special adaptation of the common-base circuit. In this instrument, tuning (C1), meter (R2), and sensitivity (R5) adjustments are performed in the usual manner. In fact, the instrument is used in every application in the same way as the usual GDO. Here, however, the microammeter (which usually is connected in a tube grid circuit and deflected by the DC resulting from grid rectification of the RF oscillator signal voltage) is deflected by rectified RF sampled across the tuned circuit (L1-C1). For this purpose, C3, D1, R2, and M1 form a shunt-diode rectifier/meter circuit.

A plug-in inductor (L1) is chosen to resonate with a 100-pf midget variable capacitor (C1) over a desired frequency band. A set of plug-in inductors covers the entire desired frequency range of the instrument. Commercial coils may be used, or a set may be wound according to instructions given in Table 5-1 to cover the 1.8-MHz to 68-MHz range. Additional coils will extend the range above and below these frequency limits.

This dip oscillator is somewhat less sensitive and the tuning is more critical than the equivalent tube circuit, chiefly because of the loading of the tuned circuit by the diode/meter circuit, but it is inexpensive and easy to build.

5.7—DIP OSCILLATOR (FET)

The dip oscillator circuit in Fig. 5-7 has the sensitivity and tuning sharpness which are characteristic of the tube-type GDO, thanks to the high input impedance of the 2N3823 field-effect transistor (Q1).

This circuit is seen to be equivalent in configuration also to the conventional tube circuit. That is, a Colpitts type RF oscillator is employed, tuning is accomplished with a midget dual 100-pf variable capacitor (C1), sensitivity is adjusted by

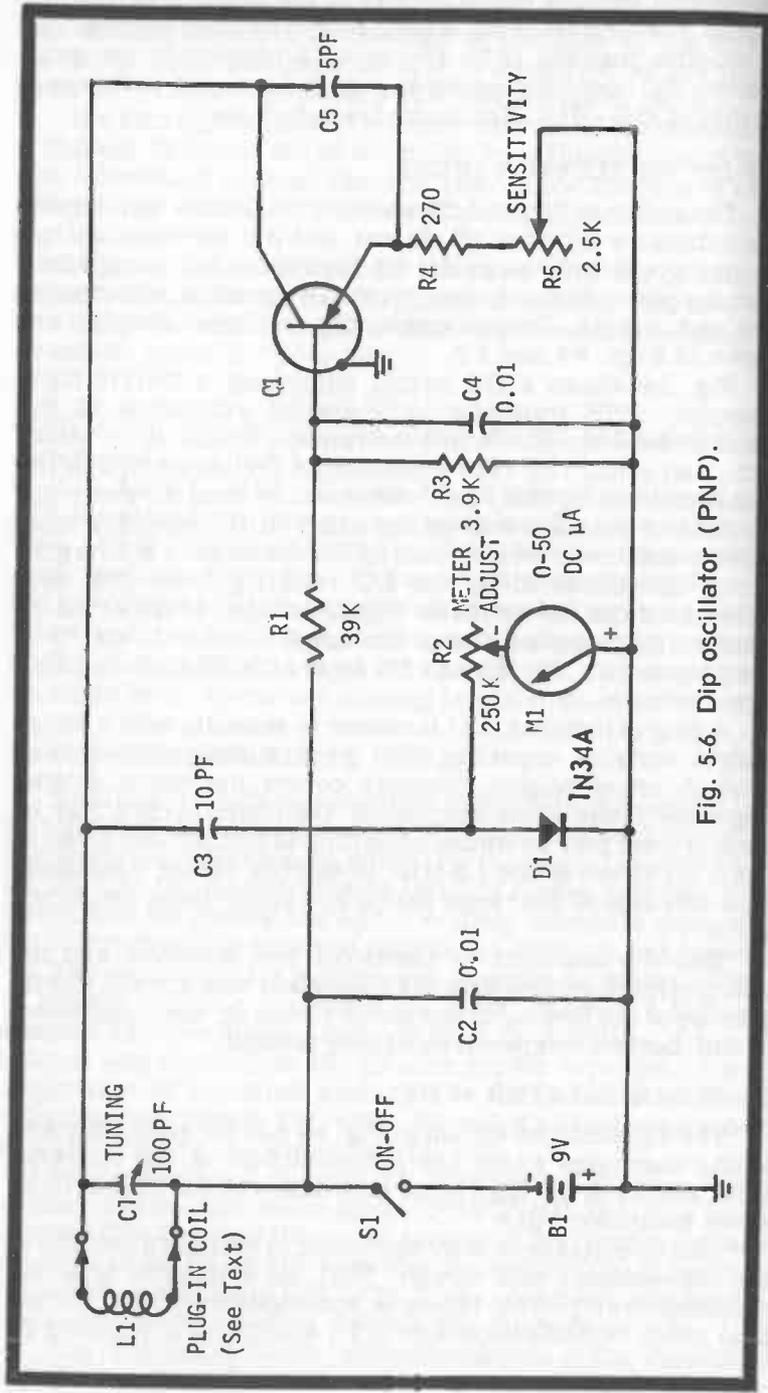


Fig. 5-6. Dip oscillator (PNP).

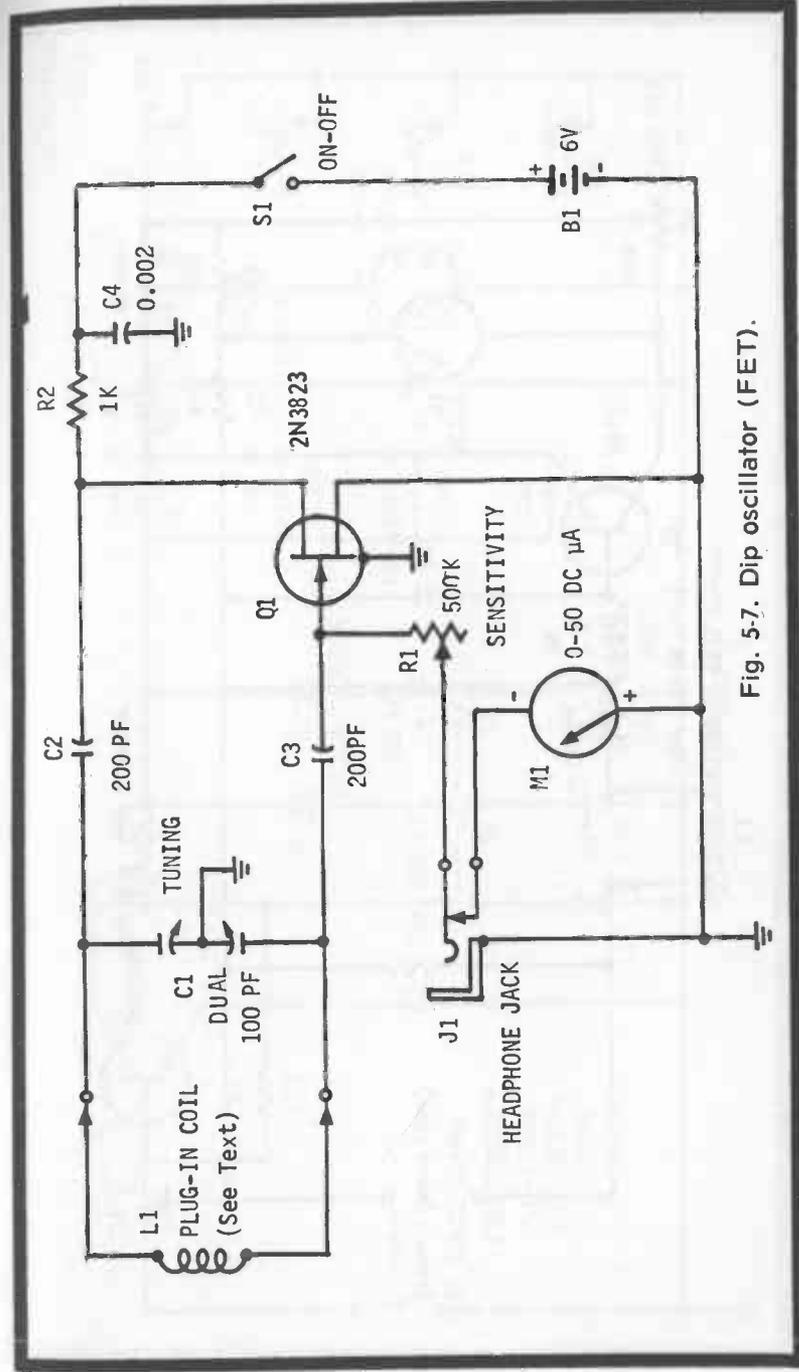


Fig. 5-7. Dip oscillator (FET).

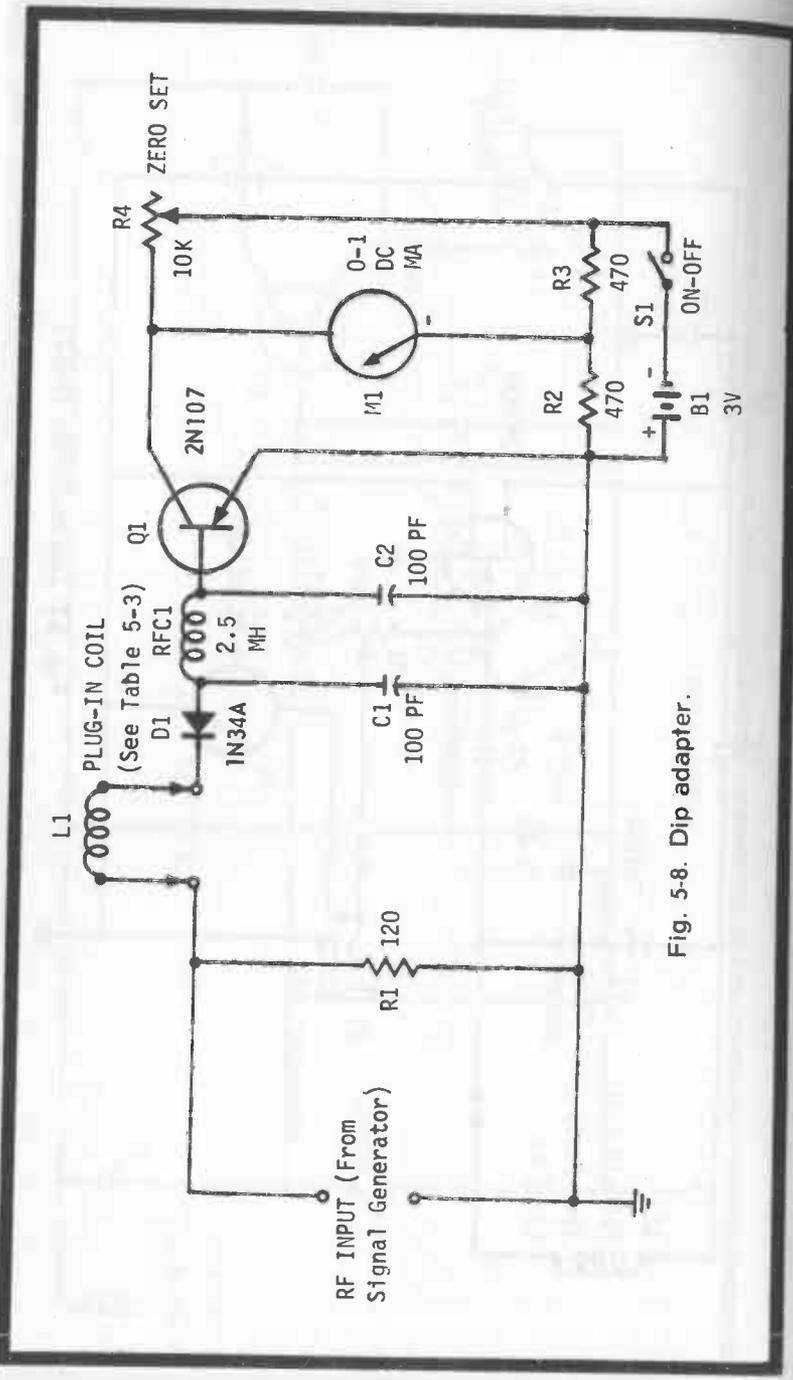


Fig. 5-8. Dip adapter.

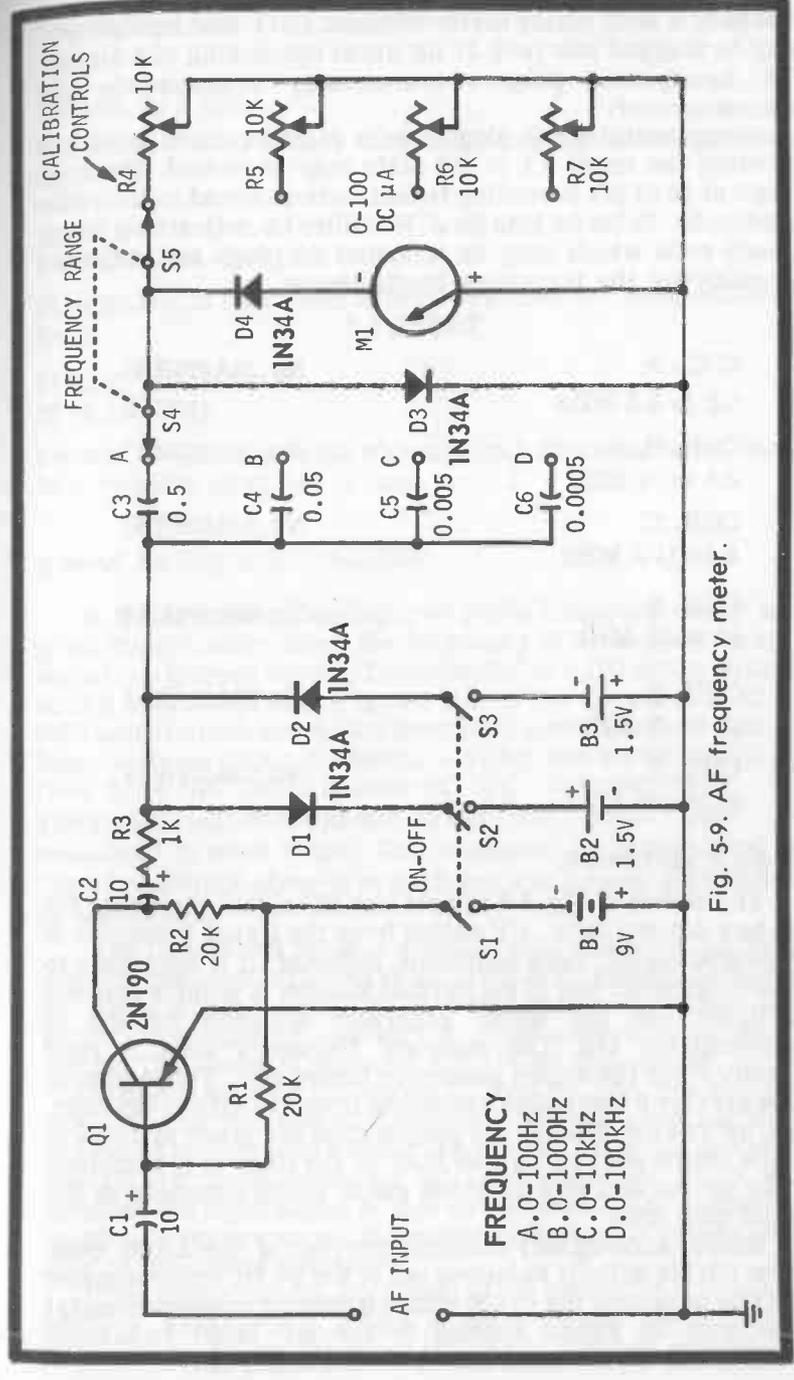


Fig. 5-9. AF frequency meter.

means of a 500K series meter rheostat (R1), and headphones may be plugged into jack J1 for aural monitoring of a signal; the headphone plug automatically disconnects the microammeter.

Commercial GDO plug-in coils may be used, or a set covering the range 1.1 to 100 MHz may be wound (for a C1 range of 10-50 pf) according to instructions found in the radio handbooks. Table 5-2 lists six J. W. Miller Co. adjustable (slug-tuned) coils which may be mounted on plugs and adjusted precisely for the frequency bands shown.

TABLE 5-2

COIL A 1.1 to 2.5 MHz	No. 20A474RBI
COIL B 2.5 to 5 MHz	No. 20A104RBI
COIL C 5 to 11.5 MHz	No. 20A225RBI
COIL D 10 to 25 MHz	No. 20A476RBI
COIL E 20 to 45 MHz	No. 20A156RBI
COIL F 40 to 100 MHz	No. 20A337RBI

5.8—DIP ADAPTER

The circuit in Fig. 5-8 adapts any RF signal generator for use as a dip oscillator. RF output from the signal generator is applied to the RF input terminals, inductor L1 is held close to the circuit under test in the normal manner of using a grid-dip oscillator, and the signal generator is tuned for dip of milliammeter M1. The resonant frequency then is read directly from the signal generator tuning dial. Three plug-in coils provide a test range extending from 100 kHz to 250 MHz. Instructions for winding the plug-in coils are given in Table 5-3. The entire adapter can be built in the form of a hand-held probe on the end of a shielded cable which connects to the signal generator.

Static (zero-signal) collector current of the 2N107 transistor (Q1) is initially balanced out of the 0-1 DC milliammeter (M1) by adjusting the 10,000-ohm wirewound zero-set rheostat (R4) with no signal applied to the RF input terminals. Maximum drain is 6 ma from the 3-volt battery (B1).

TABLE 5-3

COIL A
100 kHz to 6 MHz
86 turns No. 32 enameled wire closewound on 1-inch diameter form.

COIL B
5 to 35 MHz
11 turns No. 24 enameled wire closewound on 1-inch diameter form.

COIL C
30 to 250 MHz
2½ turns No. 24 enameled wire on 1-inch diameter form. Space to a winding length of ¼ inch.

5.9—AF FREQUENCY METER

A direct-reading, analog-type audio frequency meter is a great convenience, since the frequency of an unknown audio signal can be read directly from the dial of a DC meter without tuning or calculations. Fig. 5-9 shows the circuit of such an instrument which covers the frequency spectrum 0-100 kHz, in four ranges: 0-100Hz, 0-1000Hz, 0-10kHz and 0-100 kHz. All are read from the 0-100 µa scale of DC microammeter M1. The meter response is linear, so only one point need be calibrated in each range. The response is independent of signal amplitude above 0.7v RMS and it is largely independent of waveform.

The instrument utilizes a well-known operating principle which is common to tube-type AF frequency meters. The input AF signal higher than 0.7v overdrives the 2N190 amplifier (Q1) which, consequently, delivers a quasi square-wave output signal. The latter is squared further by the limiter circuit consisting of resistor R3, diodes D1 and D2, and 1½v penlight cells B2 and B3, and the resulting square-wave signal is presented to the meter circuit (capacitors C3 to C6, diodes D3 and D4, DC microammeter M1, and rheostats R4 to R7). Because the input signal to this circuit consists of a constant-amplitude square wave, the meter deflection is proportional only to the signal frequency.

Frequency ranges are selected by means of a 2-pole, 4-position, nonshorting, rotary selector switch (S4-S5). Each position selects one capacitor (C3 to C6) and one correspond-

ing 10,000-ohm wirewound calibration rheostat (R4 to R7). The capacitors need not be high-precision units because the rheostats allow adjustment of each range. All diodes, electrolytic capacitors, and the microammeter must be polarized exactly as shown.

The best point for calibration is full-scale deflection of the meter—100 Hz for Range A, 1000 Hz for Range B, 10 kHz for Range C, and 100 kHz for Range D. Calibration Procedure: (1) With the 3PST on-off switch (S1-S2-S3) closed, set the range switch (S4-S5) to A. (2) Apply an accurately known 100-Hz signal (higher than 0.7v RMS) to the AF input terminals and adjust rheostat R4 for exact full-scale deflection of meter M1. (3) Switch S4-S5 to B, change the input-signal frequency to 1000 Hz, and adjust R5 for an exact full-scale deflection of M1. (4) Switch S4-S5 to C, change the input-signal frequency to 10 kHz, and adjust R6 for exact full-scale deflection of M1. (5) Set S4-S5 to D, change the input-signal frequency to 100 kHz, and adjust R7 for exact full-scale deflection of M1.

5.10—AUDIO SIGNAL INJECTOR

Fig. 5-10 shows the circuit of a miniaturized 1-kHz transformer-feedback oscillator which may be built into a probe handle and used to troubleshoot audio amplifiers and associated equipment by the signal injection method. A 2N190 transistor is operated here from a 1½-volt penlight cell (B1).

The frequency is determined by the inductance of the 2000-ohm winding of miniature transformer T1 (Argonne AR-109, or equivalent), and the 0.05-mfd capacitance (C1). The transformer must be polarized correctly for oscillation; follow the color coding shown in Fig. 5-10.

The output-signal amplitude is 0.4v RMS across an external load of 39,000 ohms; the amplitude drops to approximately 18 mv RMS when the load is reduced to 1000 ohms. When testing sensitive audio devices, it sometimes is not necessary to connect directly into the circuit—merely placing the prod close to the test point will couple sufficient signal into the circuit. Current drain is 8 µa from the 1½-volt cell.

5.11—STEP-TYPE AUDIO OSCILLATOR

Fig. 5-11 is the circuit of a Hartley-type, sine-wave AF oscillator which delivers eight fixed frequencies selected individually by means of a rotary switch. The frequencies are 250, 400, and 500 Hz and 1, 2, 5, 10, and 20 kHz. A 2N2712 silicon transistor is employed.

A miniature 500-ohm (center tapped)-to-30-ohm transformer, T1 (Argonne AR-117, or equivalent) provides the tuning, feedback, and output functions. The upper half of the

500-ohm winding serves as the transistor base-input coil and the lower half as the collector-output coil. The entire winding is tuned by a capacitor (C1 to C8) selected by means of a single-pole, 8-position, non-shorting, rotary selector switch (S2). The 30-ohm transformer winding is used solely for output coupling.

The oscillation frequency is determined principally by the selected capacitor in the C1 to C8 group and the inductance of the 500-ohm winding of transformer T1. Thus, with the range switch (S2) in position D, as shown in Fig. 5-11, the 0.08-mfd capacitor is selected and the frequency is 1 kHz. Three of the capacitances must be built up by parallel-connecting several smaller capacitances. Thus, C1 (1.281 mfd) is made up from the following values: 1.0, 0.2, 0.08, and 0.001 mfd; C3 (0.32 mfd) from 0.22 and 0.1; C4 (0.08 mfd) from 0.047 and 0.033; C6 (0.0032 mfd) from 0.0022 and 0.001; and C7 (0.0008 mfd) from 0.00047 and 0.00033. Other step frequencies are obtainable with other appropriate capacitances.

For oscillation to occur, the transformer must be polarized correctly; follow the color coding shown in Fig. 5-11. Capacitor C9 is not part of the tuned circuit, but serves only to block the collector DC voltage from the base of the transistor. Negative feedback provided by the unbypassed emitter resistor (R2) acts to minimize distortion.

The 10,000-ohm wirewound potentiometer (R3) serves as

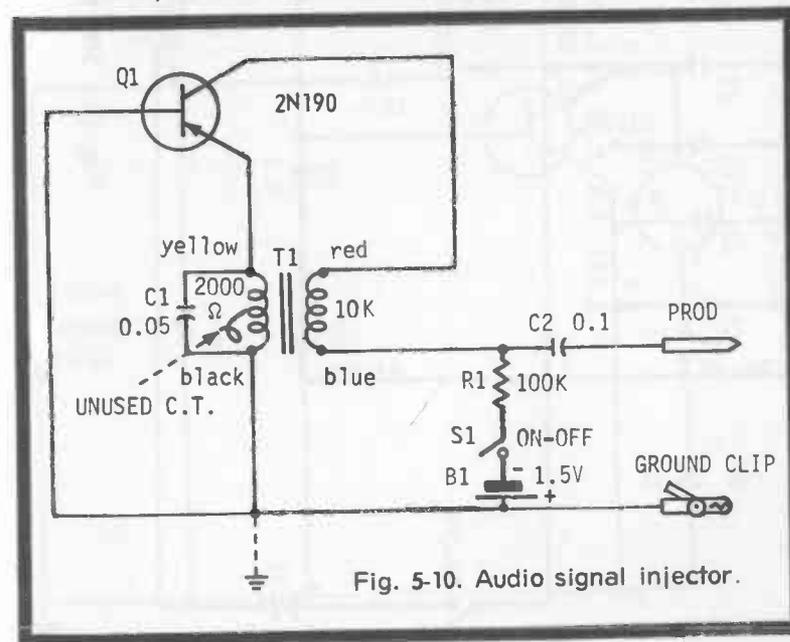


Fig. 5-10. Audio signal injector.

FREQUENCY STEPS

- A. 250 Hz
- B. 400 Hz
- C. 500 Hz
- D. 1kHz
- E. 2kHz
- F. 5kHz
- G. 10kHz
- H. 20kHz

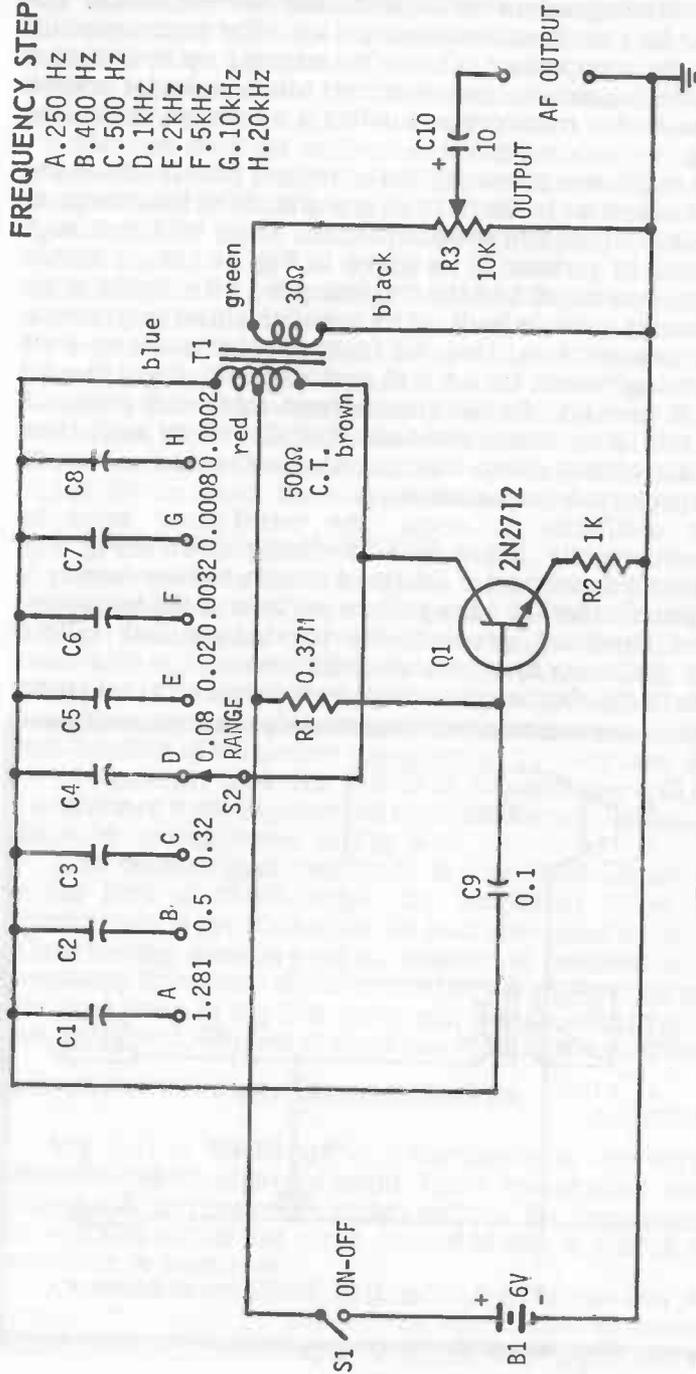


Fig. 5-11. Step-type audio oscillator.

an output-amplitude control and gives smooth variation of the signal from zero to maximum. Capacitor C10 serves as a blocking capacitor to protect R3 and T1 from any DC component in the external circuit to which the oscillator supplies a signal. Output-signal amplitude is 0.8v RMS on Ranges E, F, G, and H, and is somewhat lower on Ranges A, B, C, and D. Maximum drain is 2 ma from the 6-volt battery (B1).

5.12-INDUCTANCE/CAPACITANCE CHECKER

The circuit of a simple device which adapts a test oscillator or signal generator for inductance and capacitance checks is shown in Fig. 5-12. Inductances from 63 uh to 6300 hy and capacitances from 150 pf to 5000 mfd can be checked with the aid of a common, service-type audio oscillator (connected to terminals 1 and 2) which tunes from 20 Hz to 200 kHz.

With this device, a series-resonant circuit is set up between an unknown inductance (Lx) and standard capacitor C1, or between an unknown capacitance (Cx) and standard inductor L1. The combination of diode (D1), transistor (Q1), and DC milliammeter (M1) acts as an AC microammeter to check

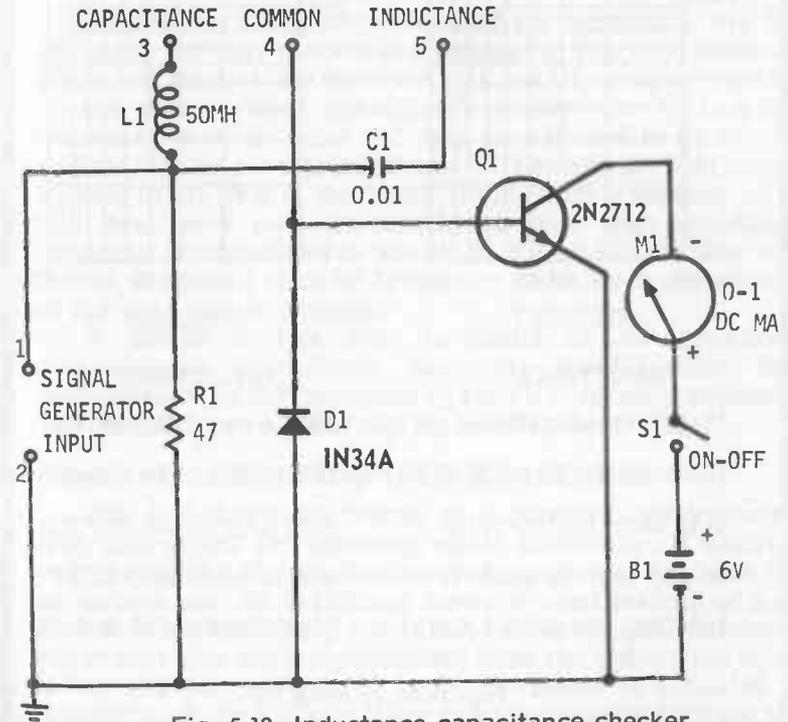


Fig. 5-12. Inductance capacitance checker.

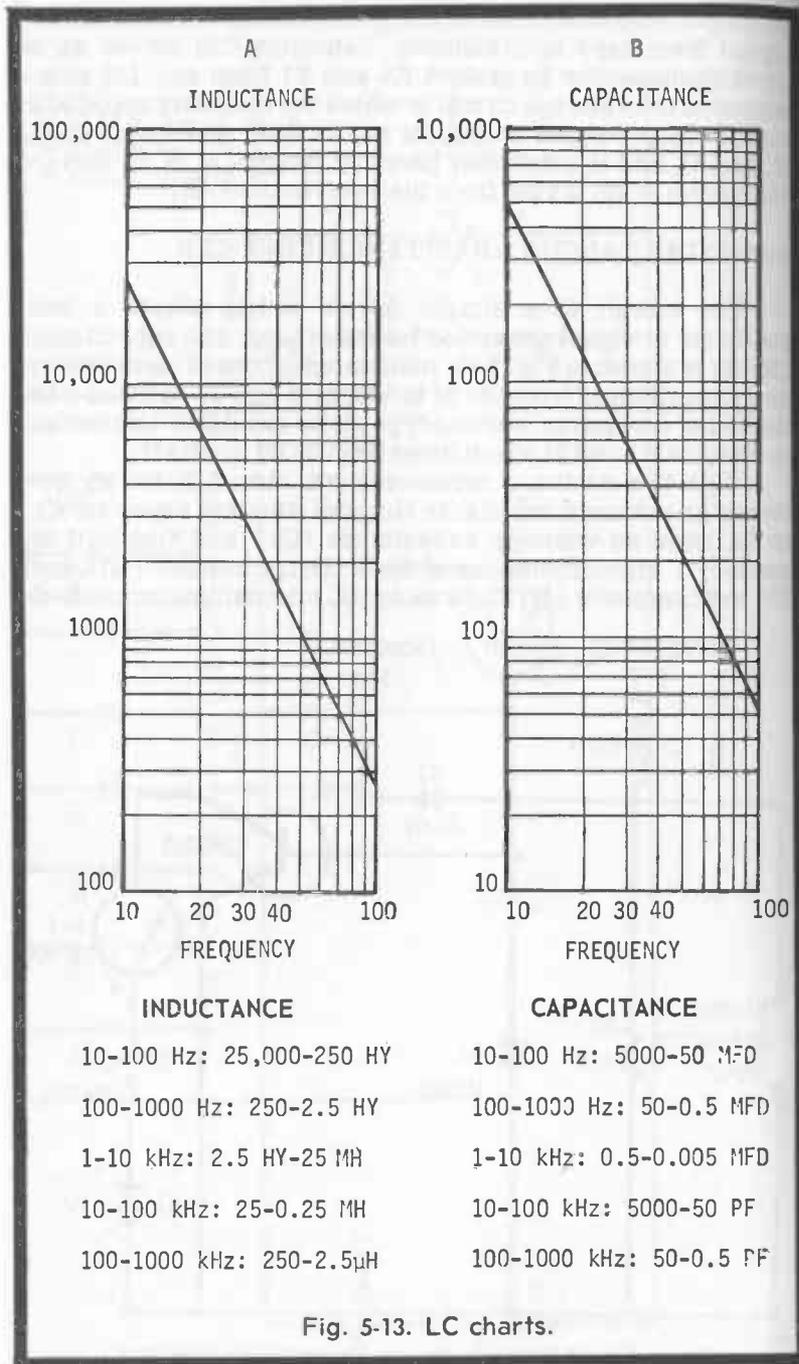


Fig. 5-13. LC charts.

the resonant current. An unknown capacitance is connected to terminals 3 and 4, or an unknown inductance is connected to terminals 4 and 5. Then, the input frequency is varied for a peak deflection of meter M1, indicating resonance. At this point, an unknown capacitance C_x equals $10^6/1.975 f^2$, or an unknown inductance L_x equals $10^9/395 f^2$, where C_x is in microfarads, L_x in henrys, and f in Hz. These tedious calculations may be avoided by using the inductance-vs-frequency chart given in Fig. 5-13A and the capacitance-vs-frequency chart in Fig. 13B. These charts are approximately correct and will give the L and C values at least as closely as the frequency can be read from most oscillator dials and as closely as f , C_1 , and L_1 usually can be obtained in inexpensive units. The accuracy of this device (and of the method) depends upon the precision of 0.01-mfd capacitor C_1 , 50-mh RF inductor L_1 (J. W. Miller No. 758, or equivalent), and the oscillator.

An oscillator output voltage of approximately 0.75v RMS (applied to terminals 1 and 2) will drive milliammeter M1 to full scale at resonance. The circuit draws 3 ma from the 6-volt battery (B1).

5.13—RF SIGNAL COMPARATOR

An untuned mixer/amplifier unit for comparing two RF signals by the beat-note method appears in Fig. 5-14. Separate inputs and shunt-diode detectors (D1 and D2) are provided for the two signals. When one signal is within a few Hz to a few thousand Hz of the other (or may be adjusted to be so), an audible beat note is set up between the two and the beat note is applied to the gate of the 2N3823 field-effect transistor (Q1). The beat note may be monitored with high-impedance magnetic headphones (as shown in Fig. 5-14), further amplified, or applied to an AF frequency meter for measurement of the beat-note frequency.

A device of this type is useful in RF frequency measurement procedures, harmonic identification, and calibration of an RF oscillator against a frequency standard. The circuit draws 2.1 ma from the 6-volt battery (B1).

5.14—STATIC DETECTOR (ELECTROMETER)

Fig. 5-15 shows the circuit of a sensitive, electrometer-type instrument for detecting static electricity. It behaves very much like the old-fashioned electroscope, but with the advantage that an indicating meter is used instead of gold leaves. When the pickup rod (3 to 5 inches of rigidly mounted rod or stiff wire superbly insulated from the instrument case) is brought near a charged body, the meter is deflected in proportion to the static voltage and the distance between the

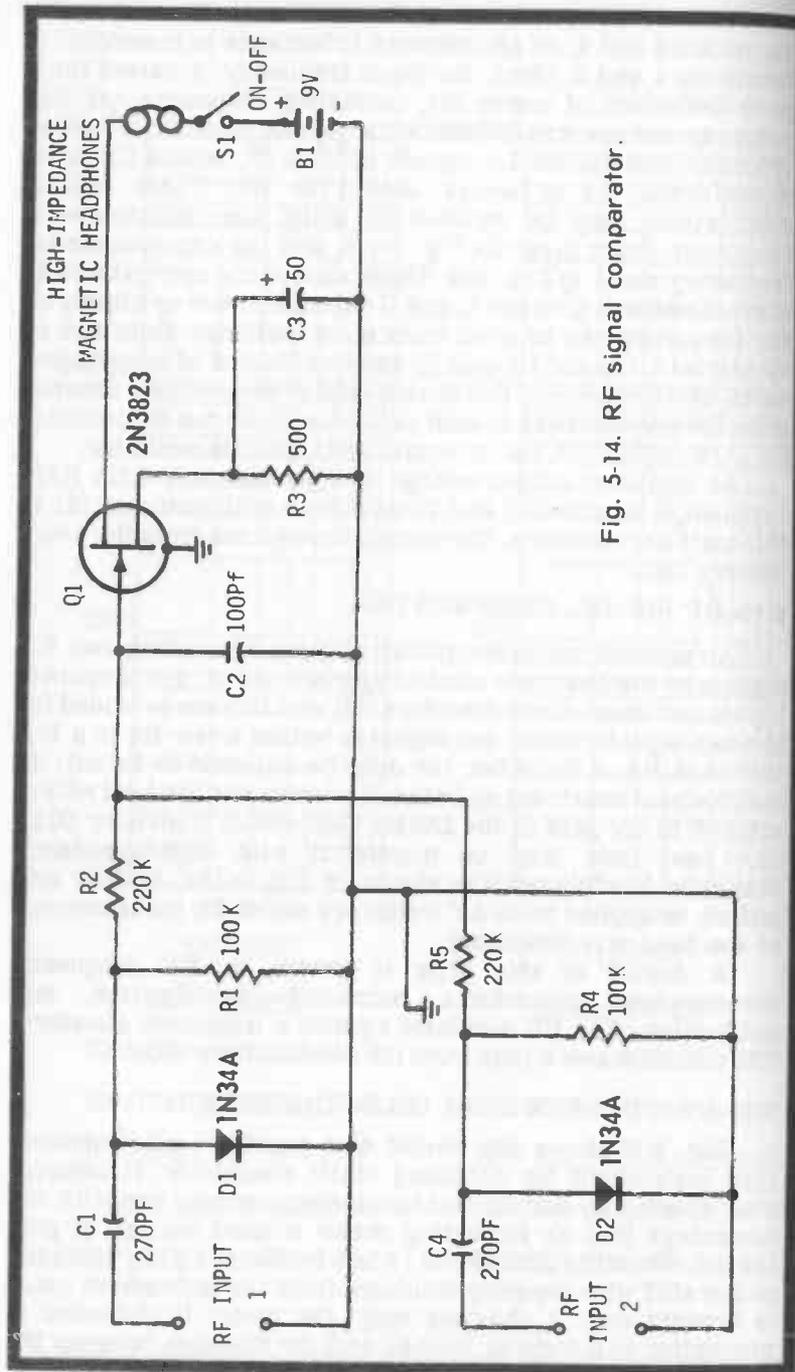


Fig. 5-14. RF signal comparator.

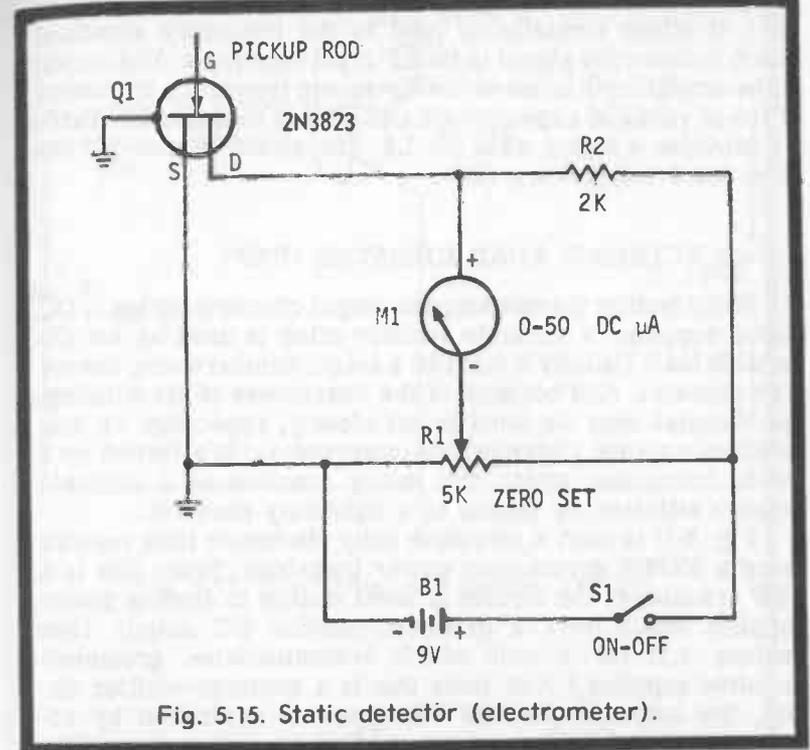


Fig. 5-15. Static detector (electrometer).

rod and the body. A dry comb charged by running it briskly through one's hair will deflect the meter to full scale when the comb is about one inch from the tip of the rod. The tip of the rod should be polished to a smooth roundness (some designers prefer to cap the rod with a polished brass ball).

This detector employs a 2N3823 field-effect transistor (Q1) in a floating-gate circuit, equivalent to a floating-grid vacuum-tube circuit. Current drain is 2.5 ma from the 9-volt battery (B1). The instrument is initially balanced by setting meter M1 to zero (by adjusting wirewound potentiometer R1) with the pickup rod clear of all fields and the operator's body.

5.15—HARMONIC AMPLIFIER FOR FREQUENCY STANDARDS

100-kHz and 1-MHz frequency standards are widely used for RF calibrations (see Sections 3.10 and 3.11). Sometimes, however, a desired check-point harmonic is too weak to be useful. When this occurs, an external amplifier tuned to the harmonic frequency will boost the signal. Fig. 5-16 is the circuit of such an amplifier.

Since this circuit employs a 2N3823 field-effect transistor (Q1), it offers virtually no load to the frequency standard which delivers the signal to its RF input terminals. The output of the amplifier is tuned to the harmonic frequency by means of 100-pf variable capacitor C4 and plug-in inductor L1. Table 5-1 provides winding data for L1. The circuit draws 1.5 ma from the 9-volt battery (B1).

5.16—ELECTRONIC LOAD RESISTOR (PNP)

While testing the volt/ampere output characteristics of DC power supplies, a variable resistor often is used as an adjustable load. Usually it must be a large, cumbersome, heavy-duty rheostat. And because of the coarseness of its winding, the rheostat may be hard to set closely, especially at low-resistance points. Considerable improvement is afforded by a power transistor, which can easily function as a variable resistor adjusted by means of a light-duty rheostat.

Fig. 5-17 is such a circuit—a fully electronic load resistor using a 2N1906 germanium power transistor. Since this is a PNP transistor, the circuit is most usable in testing power supplies which have a grounded-positive DC output. (See Section 5.17 for a unit which accommodates grounded-negative supplies.) And since this is a common-emitter circuit, the output (collector) resistance is controlled by adjustment of the DC base current. The 1½-volt battery (B1)

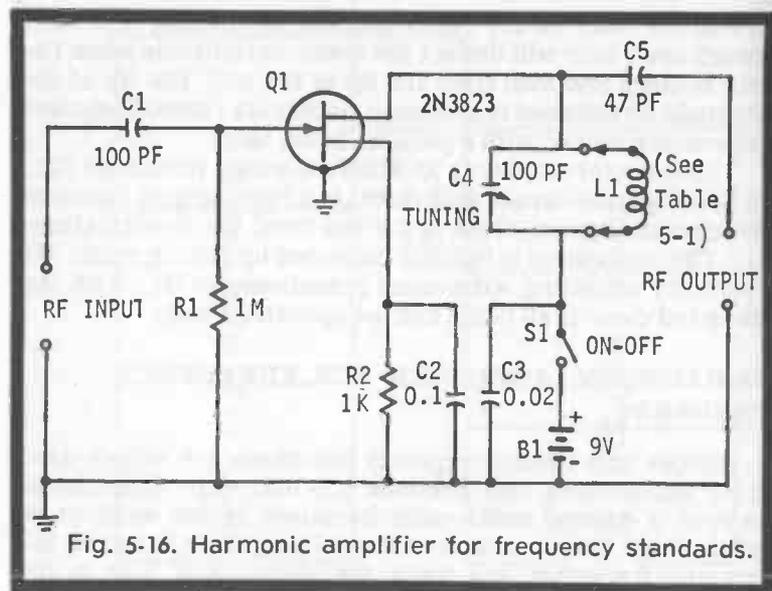


Fig. 5-16. Harmonic amplifier for frequency standards.

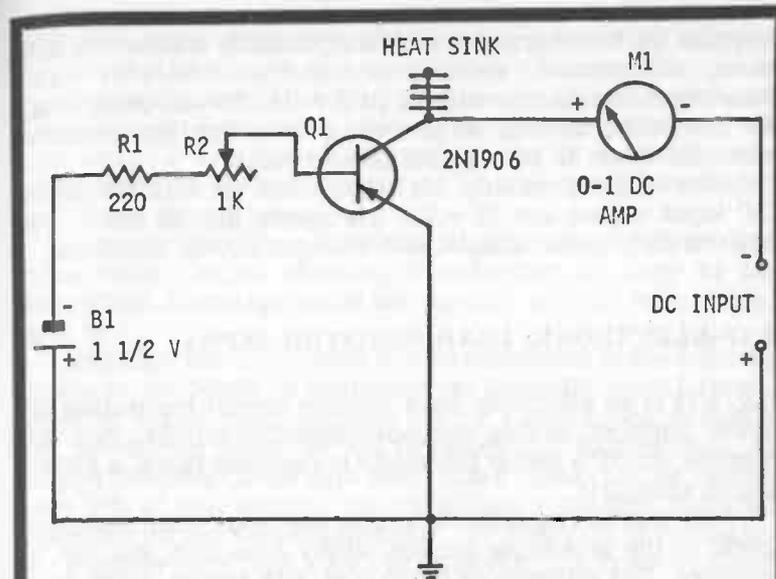


Fig. 5-17. Electronic load resistor (PNP).

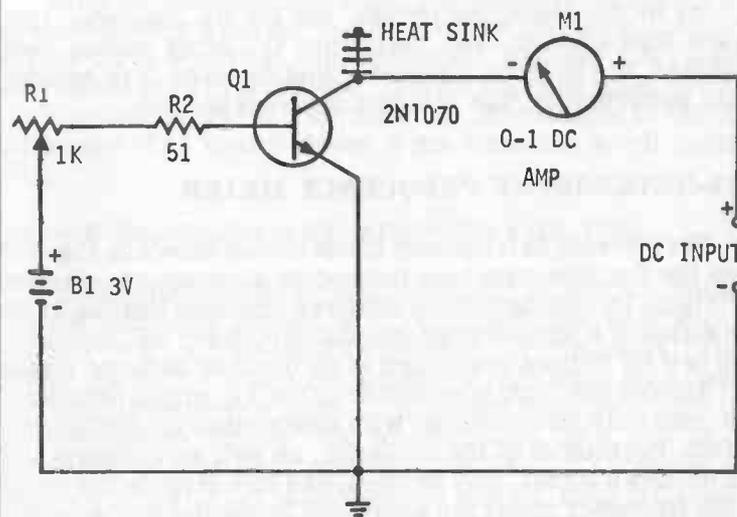


Fig. 5-18. Electronic load resistor (NPN).

supplies the base current which is adjusted by means of a 1000-ohm, wirewound, volume-control-type rheostat (R2). Maximum current drawn from B1 is 7 ma. The 220-ohm, 1-watt R1 is a safety resistor to prevent overdriving the transistor when R2 is set to its zero-resistance point.

The 0-1 DC ammeter (M1) reads load current. Maximum DC input values are 36 volts, 1 ampere, and 50 watts. For heavier-duty performance, use a larger power transistor.

5.17—ELECTRONIC LOAD RESISTOR (NPN)

Fig. 5-18 is an electronic load resistor circuit for testing DC power supplies having grounded-negative output. For this purpose, an NPN power transistor is required (here, a 2N1070 silicon unit, Q1).

The overall remarks regarding electronic load resistors, made in the preceding section, apply here and will not be repeated. The differences in the Fig. 5-18 circuit, aside from the polarity difference in the DC input, are: R1 is a 1000-ohm, 5-watt, wirewound rheostat; R2 is a 51-ohm, 2-watt safety resistor; and B1 is a 3-volt battery. The maximum current drawn from B1 is 35 ma.

As in the preceding circuit, the 0-1 DC ammeter (M1) reads load current. The maximum DC input values here, however, are 50 volts, 1 ampere, and 50 watts. For heavier-duty performance, use a larger power transistor.

5.18—HETERODYNE FREQUENCY METER

The heterodyne frequency meter circuit shown in Fig. 5-19 uses the familiar beat-note method to measure an unknown RF signal by comparing the unknown (by zero beating) with the signal of a single-range variable-frequency RF oscillator. The two RF signals are mixed in an untuned detector (diode D1) and the beat note is available at the AF output terminals. The note may be monitored with headphones or further amplified. Harmonics of the oscillator, as well as harmonics of the unknown signal, may be used, and this extends the range of the frequency meter far above and below the fundamental-frequency band of the RF oscillator in the instrument.

The circuit given in Fig. 5-19 may be used to measure frequencies from 50 kHz to 30 MHz. Above and below these limits, the harmonics become too weak for practical use. The tuning range of the oscillator (determined by the L1-C3-C4 combination) is 1000 to 2000 kHz. The main tuning capacitor

(C4) is a 350-pf midget variable; a second midget variable capacitor (C3) is used as a fixed-tuned trimmer. Inductor L1 consists of 58 turns of No. 26 enameled wire closewound on a 1-inch diameter form, with a tap at the 30th turn from the bottom end; L2 consists of 15 turns of No. 26 enameled wire closewound on the same form as L1 and spaced one-eighth inch from the bottom of L1.

For best stability, capacitors C1, C2, C5, and C6 should be mica units. Output coupling transformer T1 may be any convenient interstage audio unit having a turns ratio of 2 to 1 or 3 to 1.

Initially, the instrument can be calibrated in the following manner: (1) Feed an unmodulated 1000-kHz signal from a good RF signal generator into the RF signal input terminals. (2) Set tuning capacitor C4 to maximum capacitance. (3) Adjust trimmer C3 for zero beat. The C4 dial may be marked 1000 kHz at this setting. (4) Substitute a 100-kHz frequency standard for the signal generator. (5) Reset C3, if necessary, for a closer zero beat with the standard. (6) Tune C4 slowly from its 1000-kHz setting until the next higher standard-frequency harmonic is tuned in at zero beat. Mark this point 1100 kHz on the C4 dial. (7) Repeat at each standard harmonic, marking the dial 1200, 1300, 1400, etc., to 2000 kHz.

After completing calibration, the unknown frequency can be read directly from the C4 dial. It will be necessary then only to tune C4 to zero beat with the signal. Whether this reading then needs to be multiplied or divided depends upon whether the unknown signal is a harmonic or submultiple of the dial frequency. The circuit draws 2 ma from the 6-volt battery (B1).

5.19—AF SIGNAL-TRACER ADAPTER FOR VOM

The simple adapter circuit shown in Fig. 5-20 will convert the AC voltmeter portion of a VOM (5000 ohms-per-volt, or higher) into a high-input-impedance (5 megohms) audio signal tracer. To obtain the necessary high input impedance, this circuit employs a 2N2823 field-effect transistor (Q1) in a source-follower circuit. Its amplification, like that of all followers, is less than unity—in this case, 0.7. Output impedance of the circuit is 560 ohms at 1000 Hz. Gain is smoothly adjustable from zero to maximum by means of 5-megohm potentiometer R1.

When the signal is being traced through an audio system, the tracer sensitivity may be adjusted in two ways: (1) by means of gain control R1, and (2) by means of the voltmeter range switching. Comparative gain and loss measurements may be made in either volts or decibels.

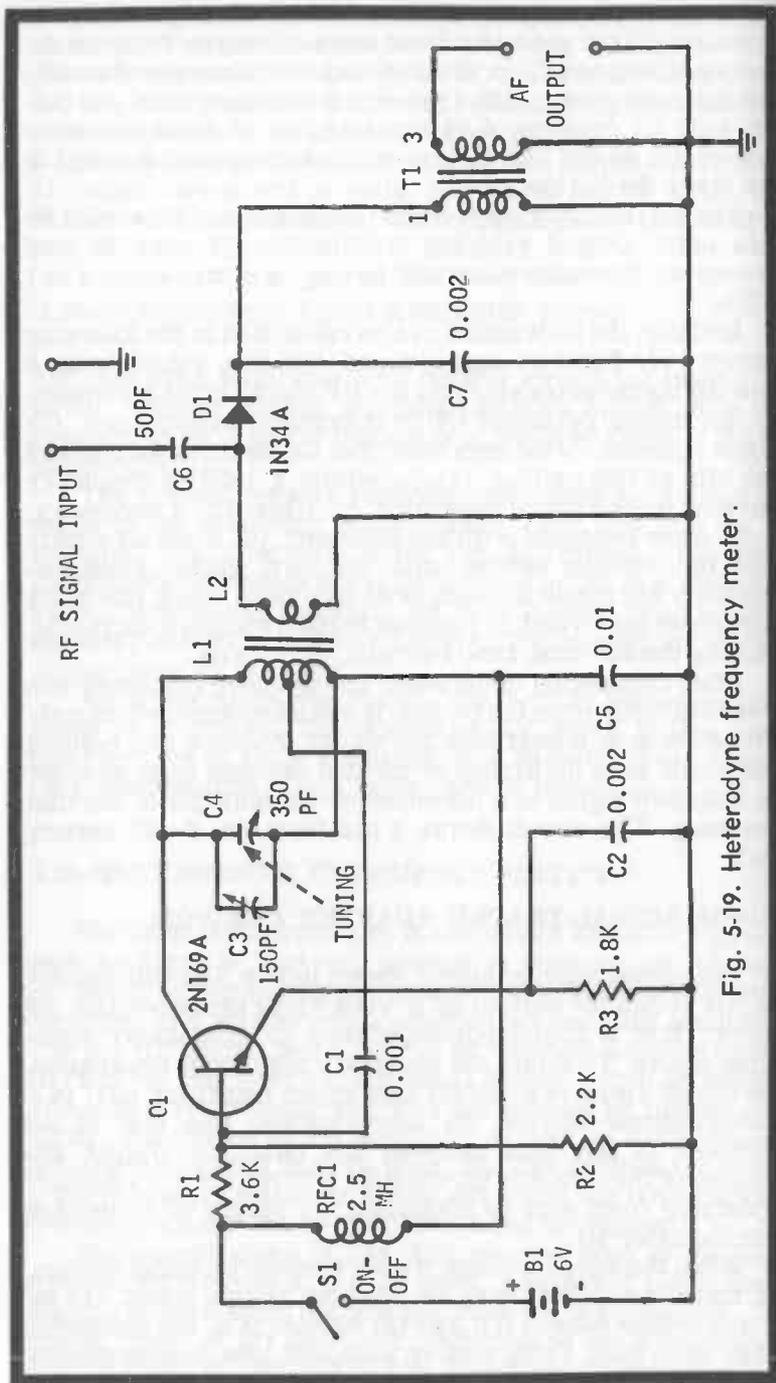


Fig. 5-19. Heterodyne frequency meter.

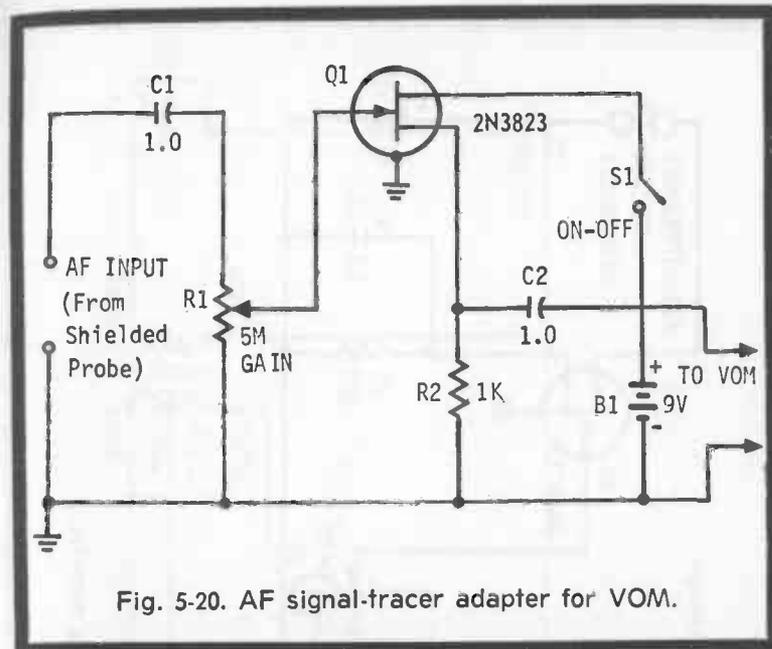


Fig. 5-20. AF signal-tracer adapter for VOM.

With R1 set for maximum gain, the maximum input-signal amplitude before output-signal peak clipping is 1v RMS and the corresponding maximum output-signal amplitude is 0.7v RMS. The circuit draws 1.7 ma from the 9-volt battery (B1).

5.20-PHONE MONITOR

Fig. 5-21 shows the circuit of a tuned monitor for checking amplitude-modulated radio transmitters and other power-type modulated RF generators. The RF signal is picked up by a suitable pickup antenna (usually short and vertical) and is tuned in by the L1-C1 combination. The signal then is detected by diode D2 and the resulting audio output is presented to a transistor audio amplifier through coupling transformer T1. The amplified signal then is monitored with high-impedance magnetic headphones or is amplified further. Audio gain is continuously adjustable by means of 1-megohm potentiometer R1.

Inductor L1 and 100-pf variable capacitor C1 are selected to tune the RF carrier frequency. (See Table 5-1 for L1 winding instructions.) Transformer T1 may be any convenient interstage audio transformer having a step-up turns ratio of 2 to 1 or 3 to 1. The circuit draws 2.1 ma from the 9-volt battery (B1).

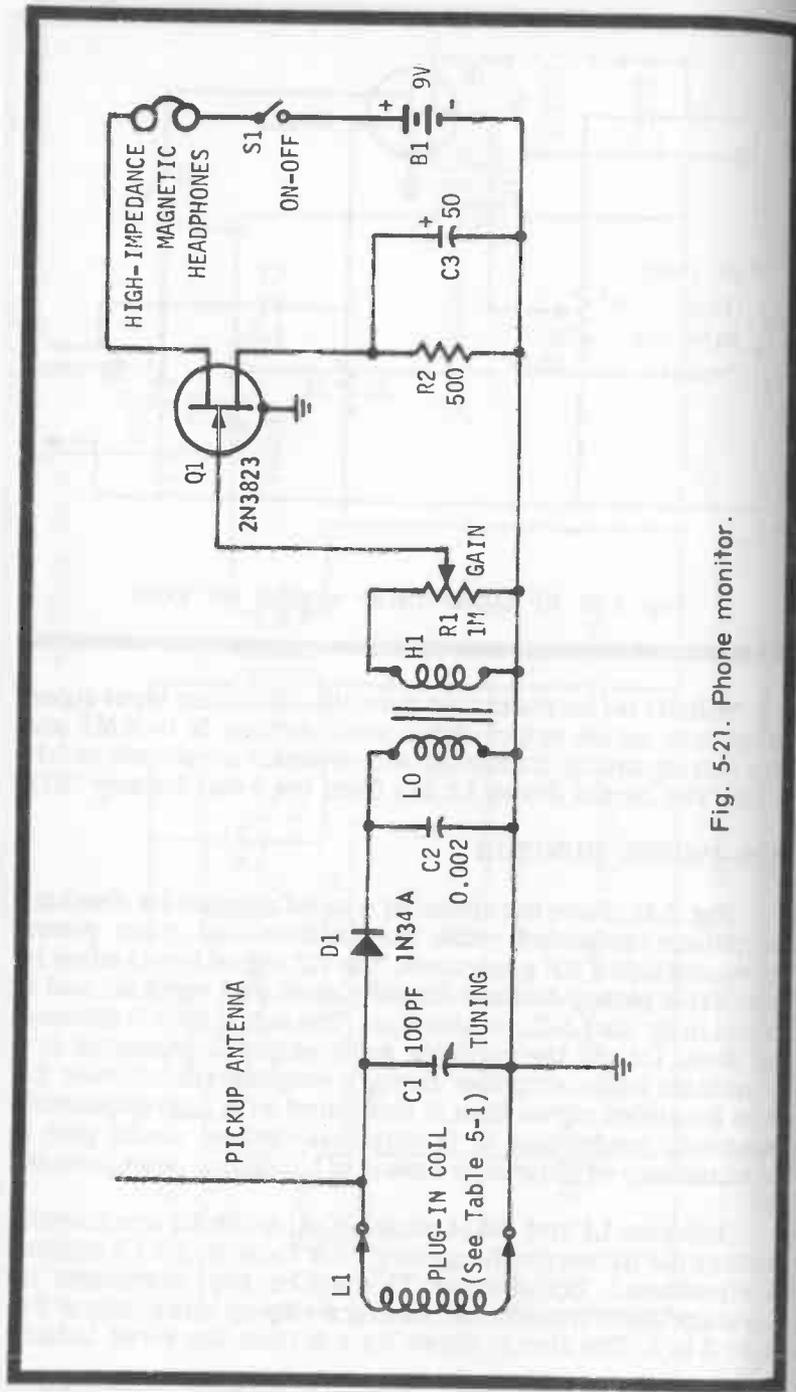


Fig. 5-21. Phone monitor.

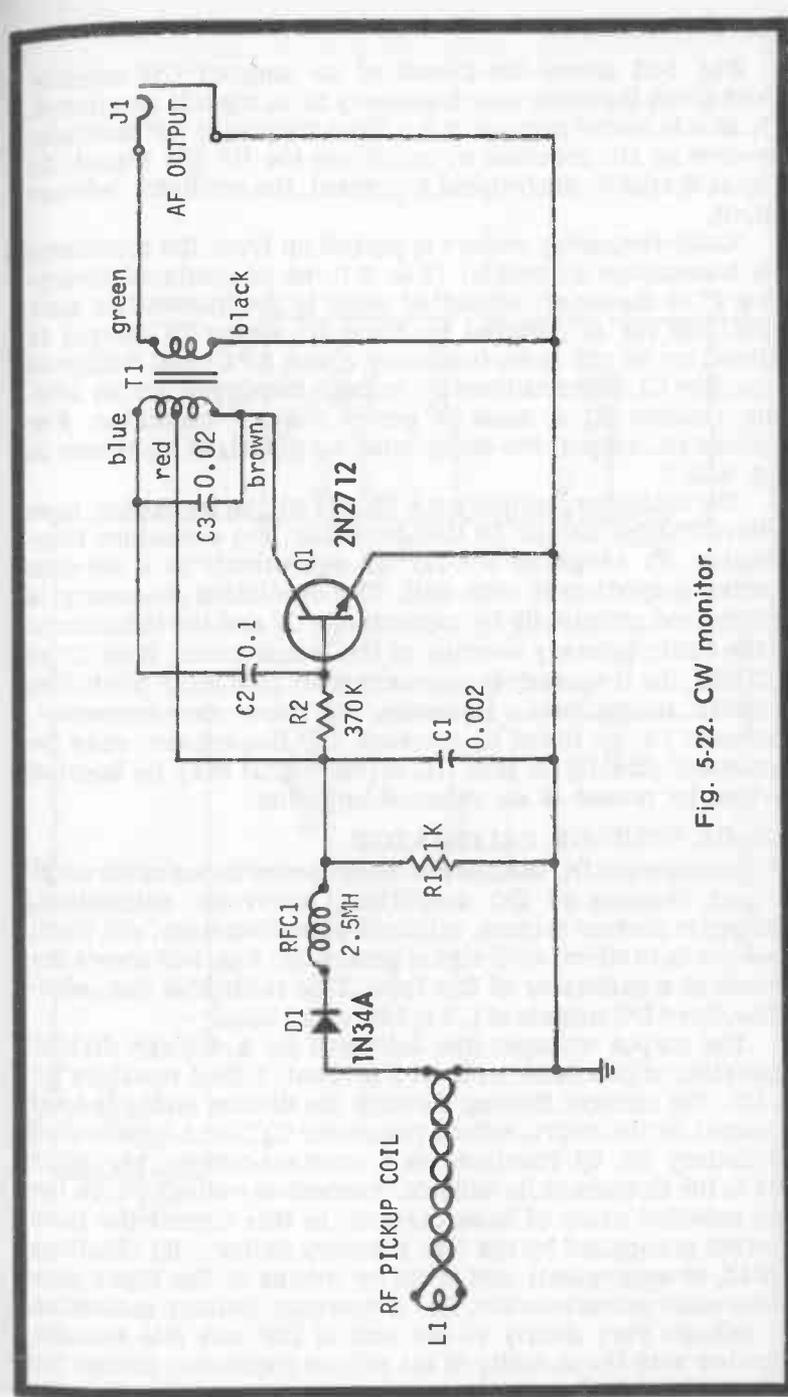


Fig. 5-22. CW monitor.

5.21—CW MONITOR

Fig. 5-22 shows the circuit of an untuned CW monitor which gives the same tone frequency to all signals monitored. It is able to do this because it is a fixed-frequency AF oscillator powered by DC obtained by rectifying the RF CW signal. As long as the dot or dash signal is present, the oscillator delivers output.

Radio-frequency energy is picked up from the monitored CW transmitter by coil L1 (2 or 3 turns of insulated hookup wire 2" in diameter, supported close to the transmitter tank coil). The RF is rectified by diode D1 whose DC output is filtered by 2½ mh radio-frequency choke RFC1 and 0.002-mfd capacitor C1. The resulting DC voltage developed across 1000-ohm resistor R1 is used to power the AF oscillator. For positive DC output, the diode must be polarized as shown in Fig. 5-22.

The oscillator, employing a 2N2712 silicon transistor, uses a Hartley-type circuit. In this oscillator, the miniature transformer, T1 (Argonne AR-117, or equivalent) is a 500-ohm (center tapped)-to-30 ohm unit. The oscillation frequency is determined principally by capacitance C3 and the inductance of the entire primary winding of the transformer. With C2 at 0.02 mfd, the frequency is approximately 2000 Hz, a pitch that satisfies many ears. However, to raise the frequency, decrease C3; to lower it, increase C3. Headphones may be connected directly to jack J1, or the signal may be boosted further by means of an external amplifier.

5.22—DC VOLTAGE CALIBRATOR

An accurate DC voltage source is needed in the calibration of such devices as DC amplifiers, electronic voltmeters, electronic current meters, millivolt potentiometers, etc. Such a source is, in effect, a DC signal generator. Fig. 5-23 shows the circuit of a calibrator of this type. This calibrator has selectable, fixed DC outputs of 1, 0.1, 0.01, and 0.001v.

The output voltages are delivered by a voltage divider consisting of precision (at least 1 percent) 1-watt resistors R3 to R7. The current flowing through the divider string is held constant by the 2N2712 silicon transistor (Q1) and is delivered by battery B2. Q1 functions as a constant-current regulator due to the flatness of its collector current-vs-voltage curve for any selected value of base current. In this circuit the base current is supplied by the 1.4v mercury battery, B1 (Mallory RM4Z, or equivalent) and is set by means of the 10,000-ohm wirewound potentiometer, R2. A mercury battery maintains its voltage very nearly to the end of life and this feature, together with the stability of the silicon transistor, makes the calibrator highly stable.

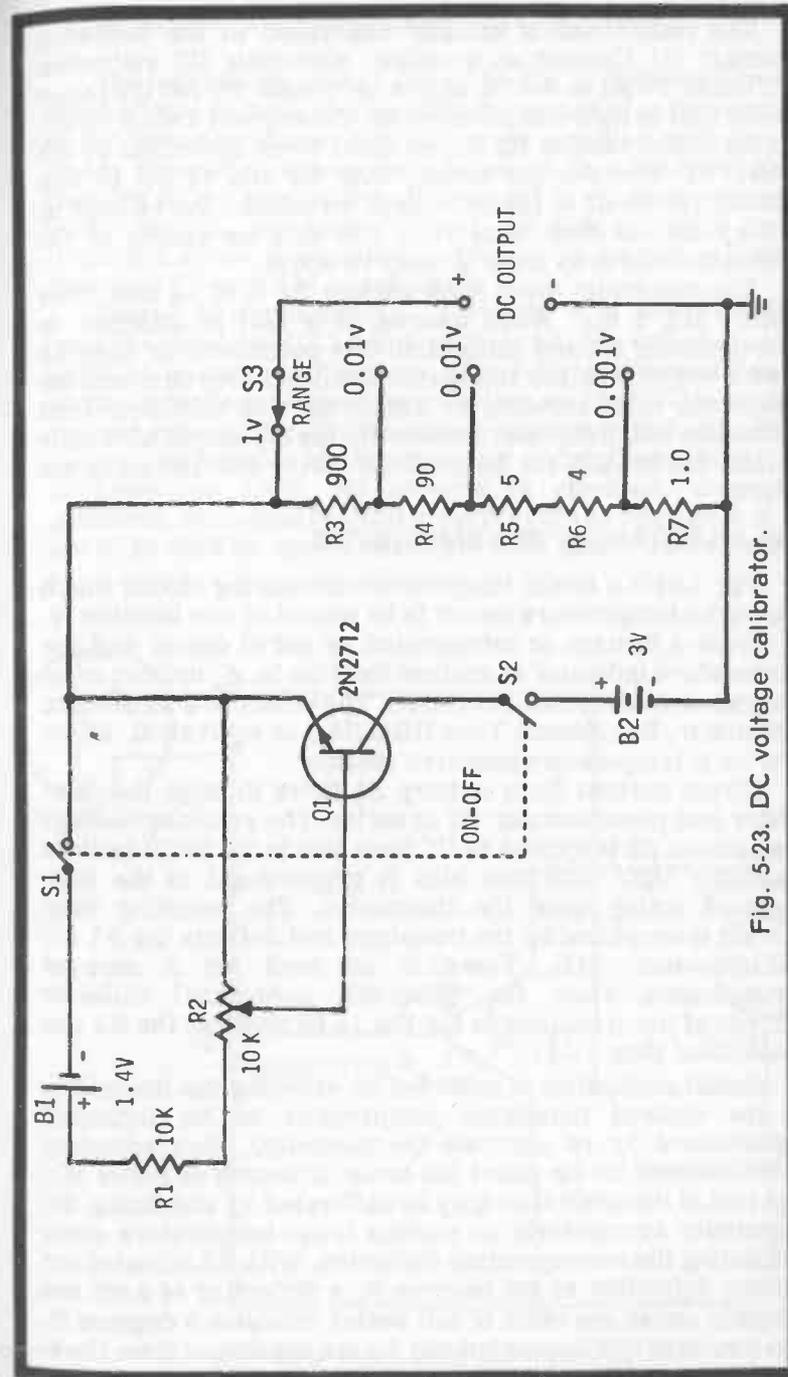


Fig. 5-23. DC voltage calibrator.

The instrument is initially calibrated in the following manner: (1) Connect an accurate, electronic DC voltmeter (VTVM or TVM) to the DC output terminals. (2) Set the range switch (S3) to its 1-volt position. (3) Close on-off switch S1-S2. (4) Set potentiometer R2 for an exact 1-volt deflection of the meter. (5) Remove the meter. With R2 and S3 set in this manner for 1-volt at the DC output terminals, the voltage of battery B2 can drop from 3v to 1.5v and the output of the calibrator will drop from 1v only to 0.975v.

The maximum drain from battery B1 is 77 μ a and from battery B2, 1 ma. When desired, if a dial is attached to potentiometer R2 and calibrated, this potentiometer may be used to vary the output voltage smoothly between zero and the maximum value selected by switch S3. The minimum load resistance which may be connected to the DC output terminals is 100K for 1v, 10K for 0.1v, 1K for 0.01v, and 100 ohms for 0.001v.

5.23—ELECTRONIC THERMOMETER

Fig. 5-24 is a useful temperature-measuring circuit which allows the temperature sensor to be placed at one location (e. g., inside a furnace or refrigerator, or out of doors) and the temperature indicator at another location (e. g., indoors or on a workbench or instrument panel). The sensor is a 15,000-ohm thermistor, R1 (Fenwal Type GB42JM1, or equivalent, which acts as a temperature-sensitive resistor.

Direct current from battery B1 flows through the thermistor and potentiometer R2 in series. The resulting voltage drop across R2 is applied as DC base bias to the 2N2712 silicon transistor (Q1). And this bias is proportional to the temperature acting upon the thermistor. The resulting base current is amplified by the transistor and deflects the 0-1 DC milliammeter (M1). There is no need for a zero-set arrangement, since the quiescent, zero-signal collector current of the transistor is too low to be read on the 0-1 ma scale (less than 1 μ a).

Initial calibration is achieved by exposing the thermistor to the desired maximum temperature to be indicated (determined by an accurate thermometer), then adjusting potentiometer R2 for exact full-scale deflection of meter M1. The rest of the scale then may be calibrated by stabilizing the thermistor successively at various lower-temperature steps and noting the corresponding deflection. With R2 adjusted for a 1 ma deflection at 350 degrees F, a deflection of 0.085 ma (slightly under one-tenth of full scale) indicates 0 degrees F. The circuit draws approximately 1.1 ma maximum from the 6-volt battery (B1).

5.24—SENSITIVE ABSORPTION WAVEMETER

The wavemeter circuit shown in Fig. 5-25 is used in the conventional manner to check the frequency of a transmitter or RF power oscillator, but it is more sensitive than the ordinary absorption wavemeter. Hence, it does not need close coupling to the transmitter. It is useful in all applications in which neither the sensitivity of the field-strength meter (Section 5.4) nor the oscillating feature of the dip oscillator (Sections 5.6 and 5.7) are needed.

Tuning is accomplished by means of a 100-pf midget variable capacitor (C1) and a suitable plug-in coil (L1). (See Table 5-1 for coil specifications.) The RF signal is rectified by germanium diode D1, and the resulting positive DC is applied as base bias to the 2N2712 silicon transistor (Q1). The base current is amplified by the transistor and deflects the 0-1 DC milliammeter (M1). No zero-set is required, since the quiescent, zero-signal collector current of the transistor is too low to be read on the 0-1 ma scale (less than 1 μ a). C1's dial

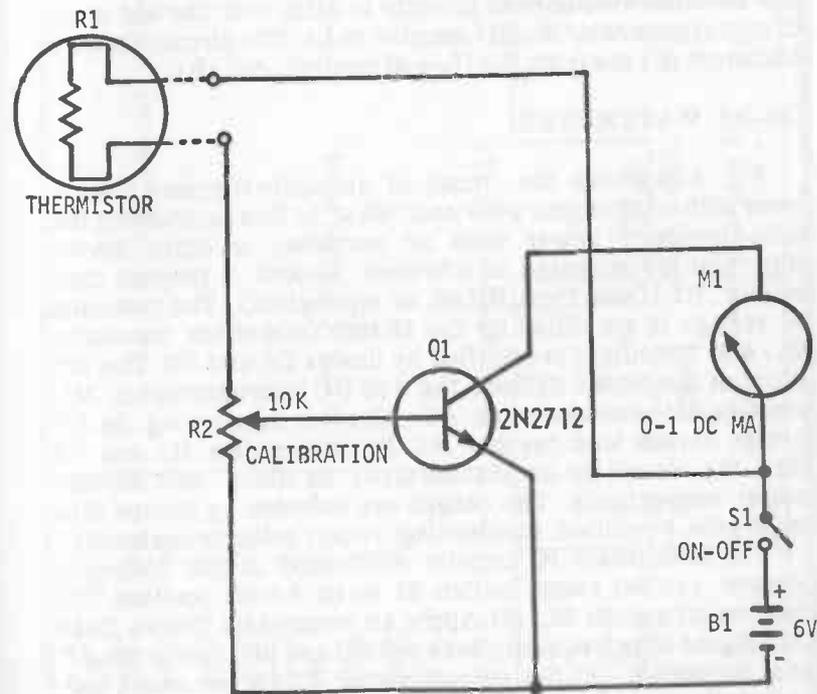


Fig. 5-24. Electronic thermometer.

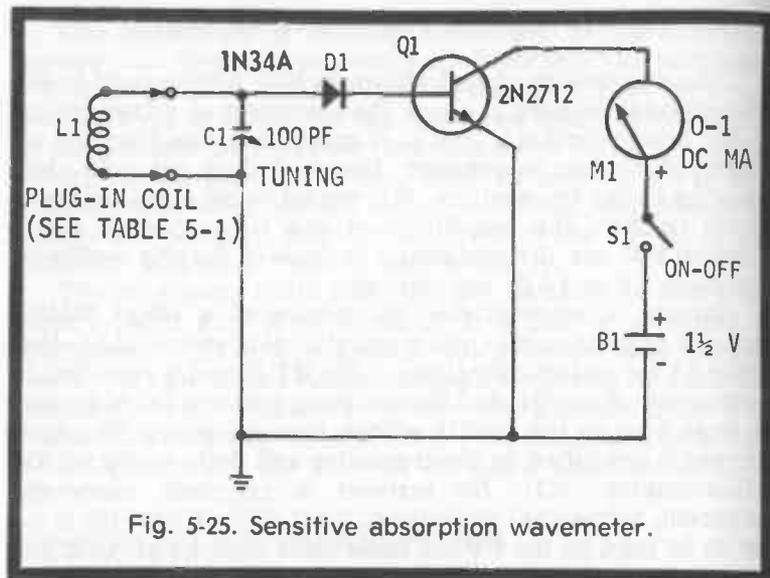


Fig. 5-25. Sensitive absorption wavemeter.

may be calibrated to read directly in MHz with the aid of an RF signal generator loosely coupled to L1. The circuit draws a maximum of 1 ma from the 1½-volt penlight cell (B1).

5.25—AF WATTMETER

Fig. 5-26 shows the circuit of an audio-frequency wattmeter with two ranges: 0-5 w and 0-50 w. In this instrument the audio-frequency power from an amplifier or other device under test is dissipated in a 50-watt, 50-ohm, 1 percent load resistor, R1 (Dale Type RH-50, or equivalent). The resulting AF voltage is amplified by the 2N4868 field-effect transistor (Q1) and the output is rectified by diodes D1 and D2. The DC output of the diodes deflects the 0-50 DC microammeter, M1. Actually, M1 is an electronic AC voltmeter monitoring the AF voltage across load resistor R1. Potentiometers R2 and R3 allow the circuit to be standardized on the 5- and 50-watt ranges, respectively. The ranges are selected by means of a single-pole, 2-position, nonshorting, rotary selector switch (S1).

The instrument is initially calibrated in the following manner: (1) Set range switch S1 to its 5-watt position. (2) Close on-off switch S2. (3) Apply an accurately known 15.8v RMS signal (any frequency between 100 and 1000 Hz) to the AF input terminals. (4) Set potentiometer R2 for an exact full-scale deflection of meter M1, and mark this point 5 w on the meter scale. (5) Reduce the voltage to 12.2v, noting the meter deflection, and mark this point 4 w. (6) Repeat at the various

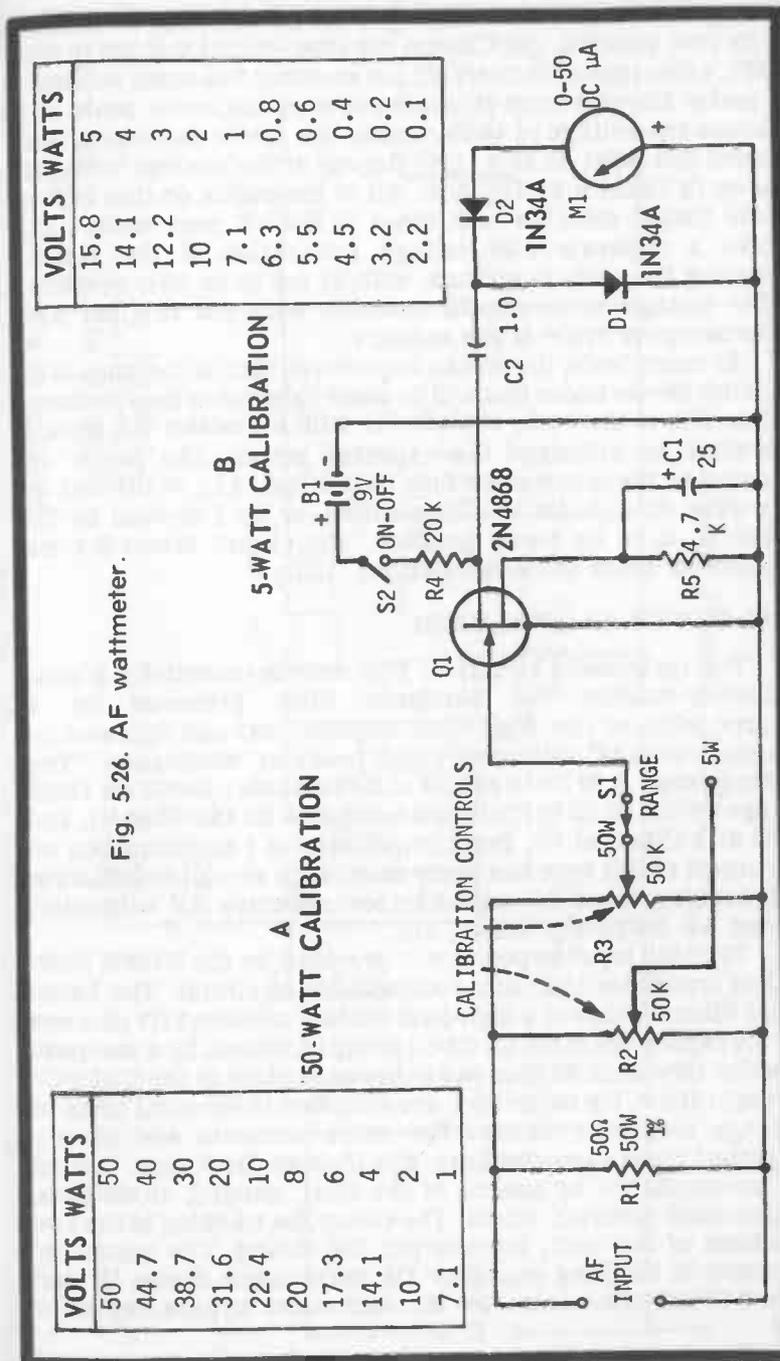


Fig. 5-26. AF wattmeter.

voltages shown in Table B in Fig. 5-26. (7) Set range switch S1 to its 50-w position. (8) Change the input-signal voltage to 50v RMS, adjust potentiometer R3 for an exact full-scale deflection of meter M1, and mark this point 50 w on the meter scale. (9) Reduce the voltage to 44.7v, noting the meter deflection, and record this point as 40 w. (10) Repeat at the various voltages shown in Table A in Fig. 5-26. All of the points on this 0-50 w scale should coincide with those of the 0-5 watt scale. (11) Make a separate 0-50 voltage calibration of the meter, marking the scale in voltage, with S1 set to its 50-w position. (The voltage scale should coincide with the original 0-50 microampere scale of the meter.)

In many tests, the output impedance (Z_o) of the amplifier or other device under test will be some value other than 50 ohms. When this is the case, replace R1 with a resistor R_x equals Z_o rated to withstand the expected power. The power indicated by the meter must then be multiplied by 50 divided by R_x when S1 is in its 50-watt position, or by 5 divided by R_x when S1 is in its 5-watt position. The circuit draws 0.5 ma maximum from the 9-volt battery (B1).

5.26-TUNED AF ANALYZER

The instrument circuit in Fig. 5-27 is essentially a continuously-tunable RC bandpass filter preceded by a source follower (for high input impedances) and followed by an electronic AC voltmeter which peaks at "resonance." The tuning range is 20 Hz to 20 kHz in three bands: 20-200 Hz (with range switch S2-S3 in Position A), 200-2000 Hz (S2-S3 at B), and 2-20 kHz (S2-S3 at C). Input impedance is 1 megohm. An instrument of this type has many uses, such as null detector for AC bridges, tuned AF signal tracer, selective AF voltmeter, tuned AF frequency meter, etc.

The high input impedance is provided by the 2N3823 field-effect transistor (Q1) in a source-follower circuit. The band-pass filter consists of a high-pass section (rheostat R3 plus one of the capacitors in the C2-C3-C4 group) followed by a low-pass section (rheostat R5 plus one of the capacitors in the C5-C6-C7 group). Here, the capacitors are switched in identical pairs to change frequency bands. The more accurate and closely matched these capacitors are, the sharper the tuning. Tuning is accomplished by means of the dual, ganged, 10,000-ohm, wirewound rheostat, R3-R5. The closer the tracking of the two sections of this unit, the sharper the tuning. The voltmeter consists of blocking capacitor C8, germanium diodes D1 and D2, 0-50 DC microammeter M1, and meter bypass capacitor C9.

The R3-R5 tuning rheostat dial may be calibrated to read

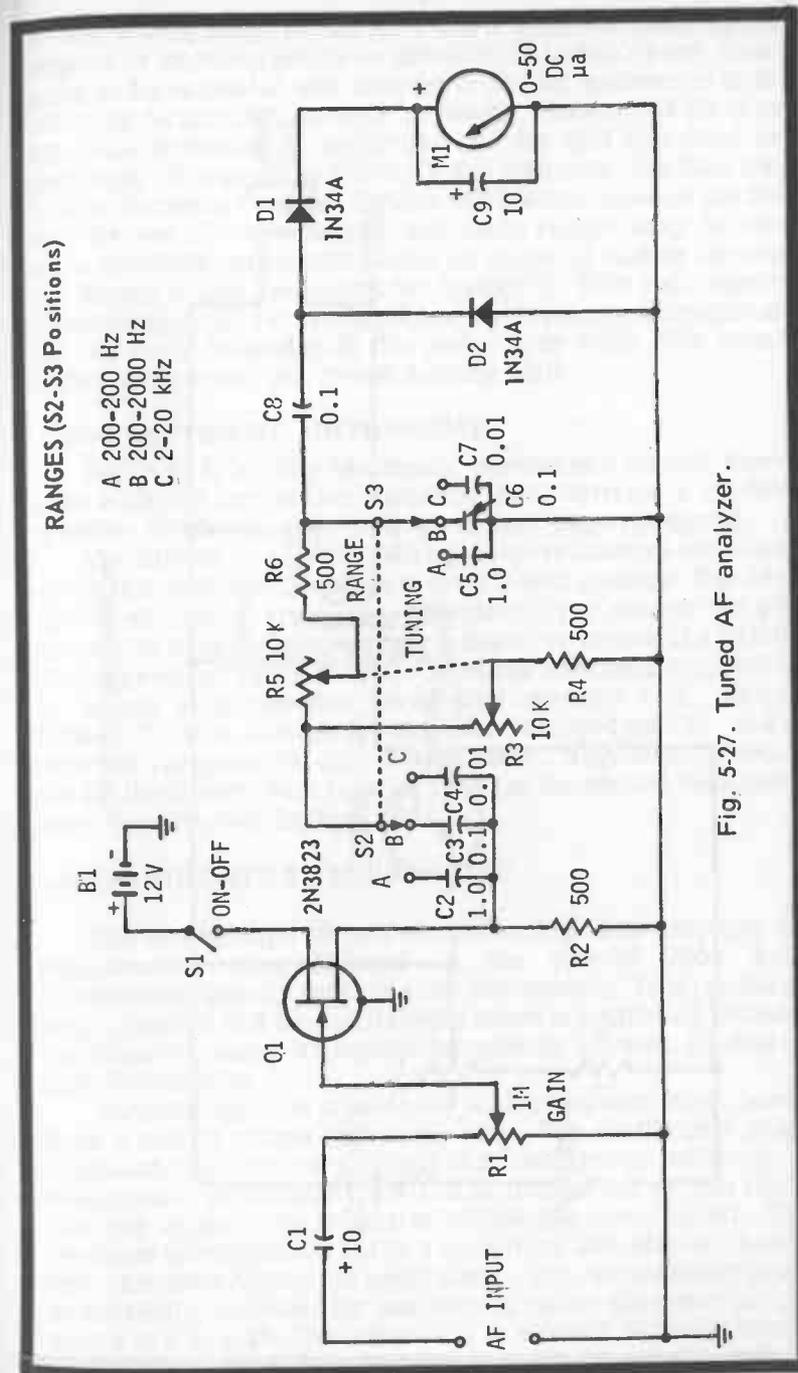


Fig. 5-27. Tuned AF analyzer.

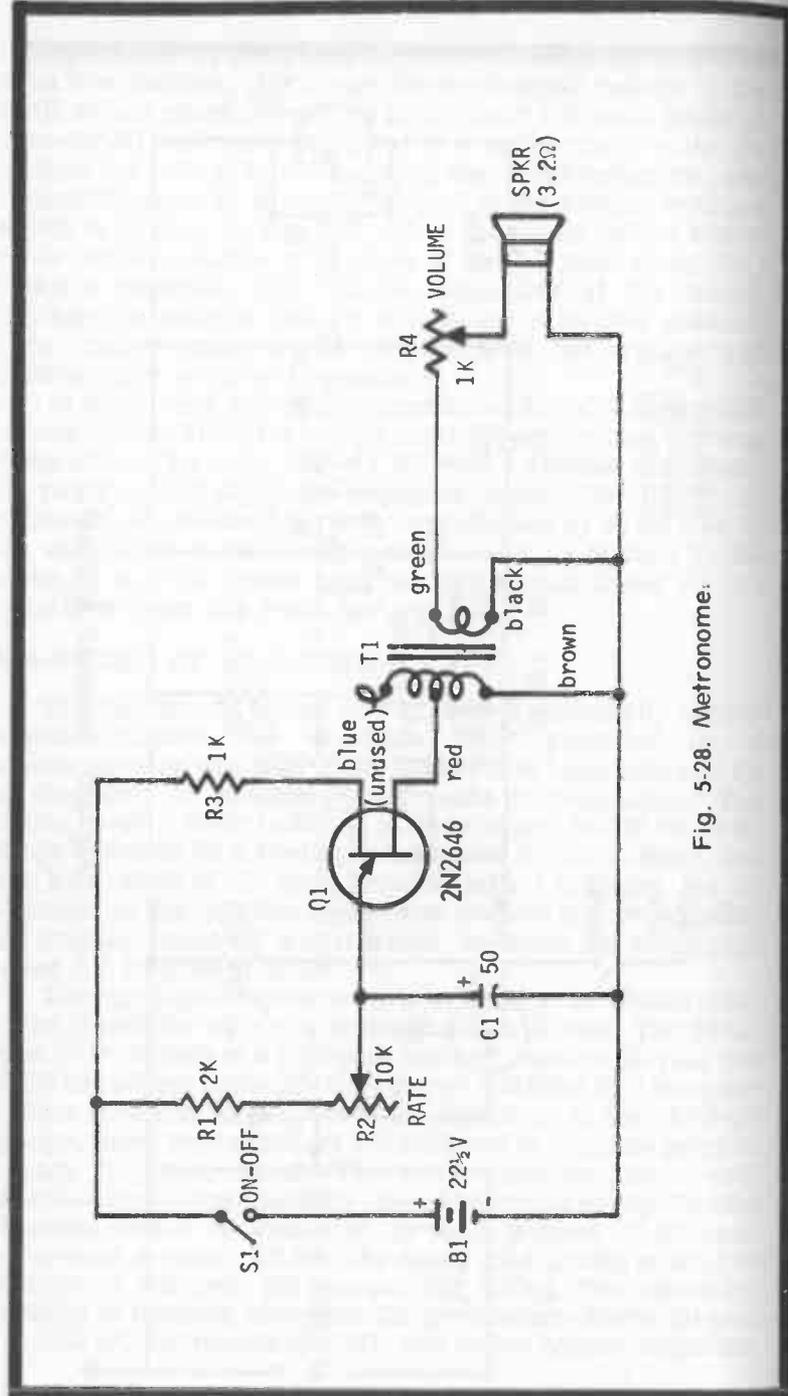


Fig. 5-28. Metronome.

20-200 Hz with range switch S2-S3 set to Position A and signals applied at as many points as practicable in this range (use a good audio oscillator with sine-wave output, connected to the AF input terminals). At each frequency, rheostat R3-R5 is set for peak deflection of meter M1 and the dial inscribed accordingly. If capacitors C2 to C7 are accurate, the dial then will be accurate (without further calibration) also on the 200-2000-Hz and 2-20 kHz ranges, and these ranges may be read from the 20-200 calibration simply by mentally adding one zero for Range B and two zeros for Range C. With gain control potentiometer R1 set at maximum, the input-signal amplitude for full-scale deflection of the meter is 2v RMS. The circuit draws 2 ma from the 12-volt battery (B1).

5.27—ELECTRONIC METRONOME

Fig. 5-28 is a fully electronic metronome circuit based upon a 2N2646 unijunction transistor (Q1). Driving a 2½-inch speaker, this device provides a good, loud, pop-type signal.

The circuit is a variable-frequency relaxation oscillator which is transformer-coupled to the 3.2-ohm speaker. The beat rate is adjustable from approximately 1 per second (60 per minute) to 10 per second (600 per minute) by means of a 10,000-ohm wirewound rheostat (R2). Speaker volume is adjustable by means of a 1000-ohm wirewound rheostat (R4). Transformer T1 is a miniature, 250-ohm (center-tapped)-to-3.2-ohm unit (Argonne AR-122, or equivalent). The circuit draws 4 ma (at the slowest beat rate) to 7 ma (at the fastest beat rate) from the 22½-volt battery (B1).

5.28—BRIGHT-BULB STROBOSCOPE

The stroboscope circuit shown in Fig. 5-29 employs an incandescent lamp instead of the special neon bulb (Strobotron) usually found in such instruments. This results in bright flashes, but the top flashing speed is restricted because the filament cannot extinguish completely between flashes at high frequencies.

The arrangement is basically a blocking oscillator, based upon a 2N2712 silicon transistor (Q1). The oscillator's pulse repetition rate, and accordingly the flash rate of the lamp, is continuously adjustable from 1 to 25 flashes per second (60 to 1500 per minute) by means of 50,000-ohm rheostat R1. The feedback transformer (T1) is a miniature 2500-ohm-to-11-ohm unit (Argonne AR-114, or equivalent). This transformer must be correctly polarized for oscillation; follow the color coding shown in Fig. 5-29. The lamp (V1) is a 2-volt, 60-milliamper (brown-bead) pilot light operated on collector-current pulses.

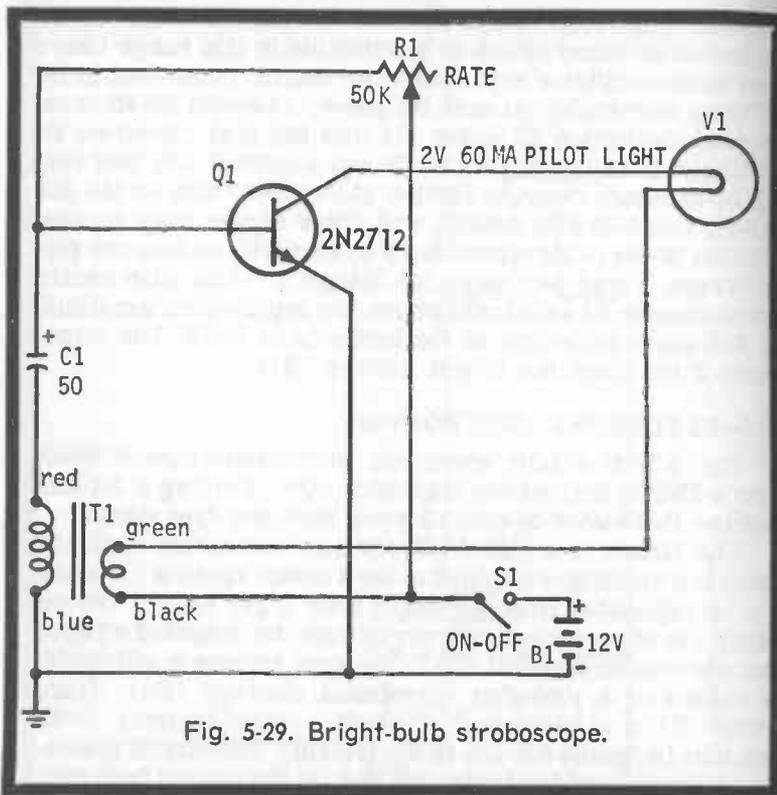


Fig. 5-29. Bright-bulb stroboscope.

For increased brilliance, the lamp may be mounted in a suitable reflector. The circuit draws 60 ma on instantaneous pulses, from the 12-volt battery (B1).

5.29—SENSITIZER FOR DC MILLIAMMETERS

Occasionally, a sensitive DC milliammeter or microammeter is needed and only a relatively high-range milliammeter (e. g., 0-10 or 0-100 ma) is available. In this case, an inexpensive silicon transistor (2N2712), 1½-volt battery, and 500-ohm rheostat can convert the more rugged milliammeter into a sensitive instrument.

Fig. 5-30 shows a simple sensitizer circuit. In this arrangement, an input current of 50 μ a deflects the 0-10 DC milliammeter (M1) to full scale when the miniature 500-ohm wirewound calibration rheostat (R1) is set to zero. (The DC signal input voltage is 0.75v and the corresponding input resistance of the circuit is 15K.) When R1 is set to approximately 125 ohms, a DC signal input of 1 ma at 0.69v deflects the meter to full scale (the input resistance is 690 ohms).

The circuit will also operate with a 0-100 DC milliammeter (M1). In this instance, with R1 set to zero and B1 delivering 1½v, an input current of 10 ma at 1v deflects the 0-100 DC milliammeter to full scale (input resistance is 100 ohms). If B1 is raised to 6v and R1 is set approximately to 50 ohms, an input current of 1 ma at 0.67v deflects the meter to full scale (input resistance is 670 ohms).

No zero set is required, as the quiescent, zero-signal collector current of the transistor is too low (less than 1 μ a) to be seen on the meter scale. Also because of this low drain, no on-off switch is needed.

5.30—CIRCLE GENERATOR FOR OSCILLOSCOPE WHEEL PATTERNS

Wheel patterns are useful for oscilloscopic comparison of two frequencies. One of these patterns is the familiar dot wheel. In the dot-wheel method, a signal of known frequency (from a frequency standard or accurate variable-frequency oscillator) is applied simultaneously to the horizontal and vertical inputs of the oscilloscope, but with the horizontal and vertical portions of the signal 90 degrees out of phase with each other. The result is a circle (or wheel) pattern on the

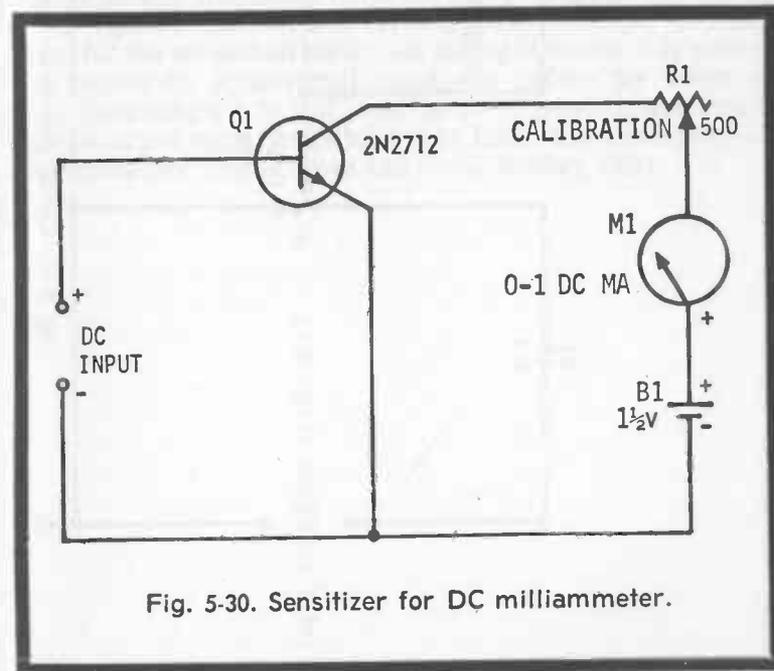


Fig. 5-30. Sensitizer for DC milliammeter.

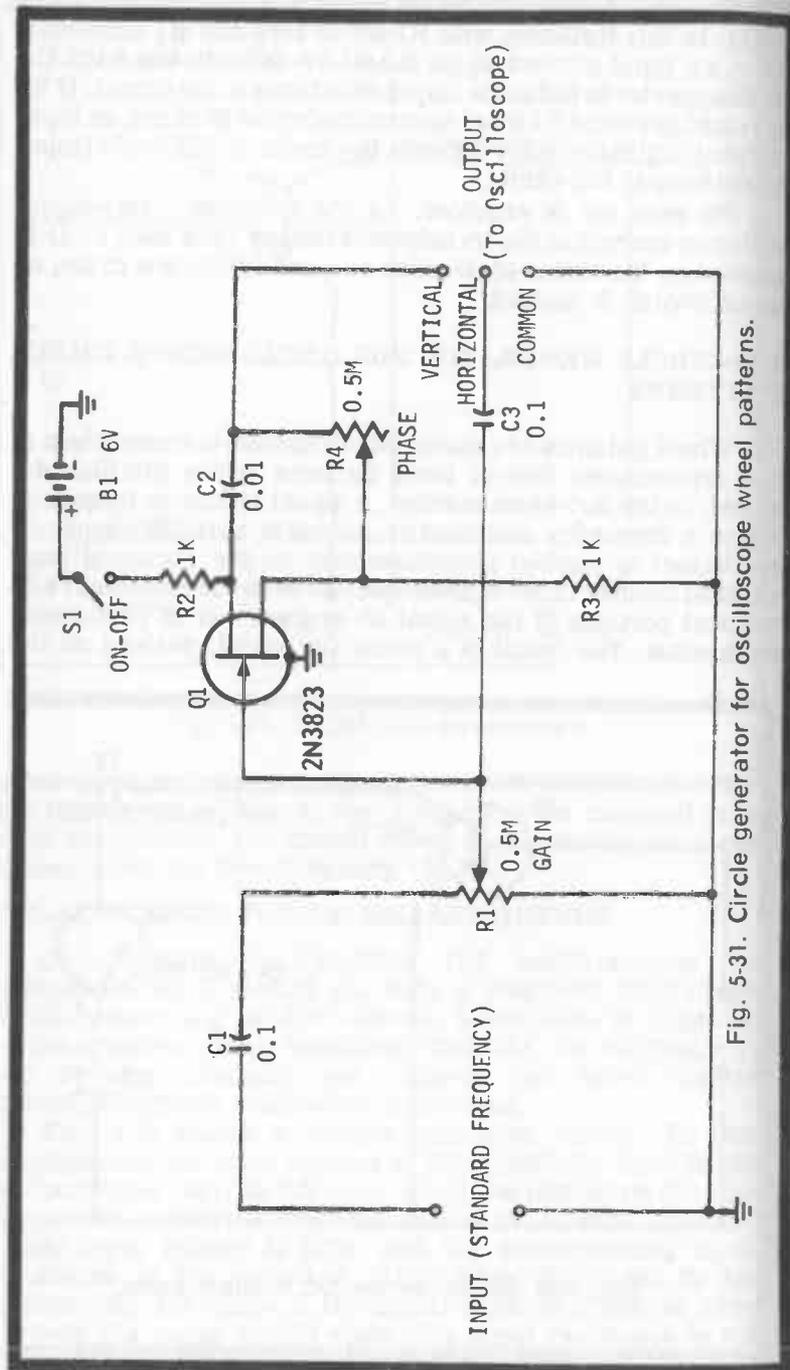


Fig. 5-31. Circle generator for oscilloscope wheel patterns.

screen. The signal of unknown frequency then is applied to the intensity-modulation (Z-axis) input of the oscilloscope, which causes the previously solid line of the circle or wheel to be broken up into dots or segments. The operator then multiplies the standard frequency by the number of dots to obtain the unknown frequency.

Fig. 5-31 shows a continuously-variable phase-splitter circuit for providing the separate horizontal and vertical signals from a single, known-frequency source. This arrangement has several advantages over the simple RC circuit usually employed: (1) A single control (R4) permits the 90-degree phase shift to be obtained at any frequency between 20 Hz and 20 kHz. (2) No range switching is required. (3) The outputs have a constant amplitude, regardless of the setting of R4. (4) A common ground is provided between the input and the two outputs. (5) The gain (and accordingly the circle diameter) is continuously variable (potentiometer R1). (6) Employing a 2N3823 field-effect transistor (Q1), the circuit offers virtually no load to the standard-frequency source.

In this circuit, 1000-ohm resistors R2 and R3 must be closely matched. The horizontal and vertical output-signal amplitudes then will be very nearly equal. Connections to the oscilloscope must be kept as short as possible to prevent stray pickup. At any frequency (with the oscilloscope gain controls set for equal horizontal and vertical gain), rheostat R4 must be set for the smoothest circle; its tuning is reasonably sharp. The maximum input-signal amplitude before the onset of circle flattening is 1.5v RMS and the corresponding maximum no-load output-signal amplitude is 1v RMS. The circuit draws approximately 1.6 ma from the 6-volt battery (B1).

Power Supply Applications

Transistors are employed in various types of power supplies and power-supply accessories. All operating constants given in this chapter are those obtained in tests with the author's models and may vary somewhat due to individual transistor and circuit component characteristics. Where particular components from named manufacturers are listed, these components (or their exact equivalents, when obtainable) appear to be necessary to the proper performance of the circuit. In all other instances, however, any component having the specified electrical characteristics may be used. Except where shown otherwise on the diagram or in the text, all resistances are in ohms and capacitances in microfarads; resistors are one-half watt; and capacitors are 25 DCWV. Other circuits, described in earlier chapters, which might also have possible use in power supplies are found in Sections 2.7, 3.12, 4.10, 4.16, 5.17, and 5.18.

6.1—VARIABLE DC POWER SUPPLY

Fig. 6-1 is a variable-output DC supply suitable for powering transistorized equipment and other low-voltage gear. Its maximum DC output is 50v, 400 ma. The circuit employs a 2N1482 silicon power transistor as a variable resistor in series with the DC output of a solid-state power supply (T1-D1-D2-C1). When Q1's DC base-bias current is varied by adjustment of the light-duty 50K wirewound rheostat, R1 (Clarostat Type 43C2, or equivalent), the internal collector-emitter resistance is varied accordingly.

Transformer T1 is a 140v to 40-0-40v, 1.2-amp unit (Knight 54 D 3988, or equivalent). Diodes D1 and D2 are 1N3544, 600 ma silicon rectifiers. Filter capacitor C1 is a 100-mfd, 150 DCWV electrolytic. The advantage of this circuit lies in its ability to control a high output current with a light-duty rheostat.

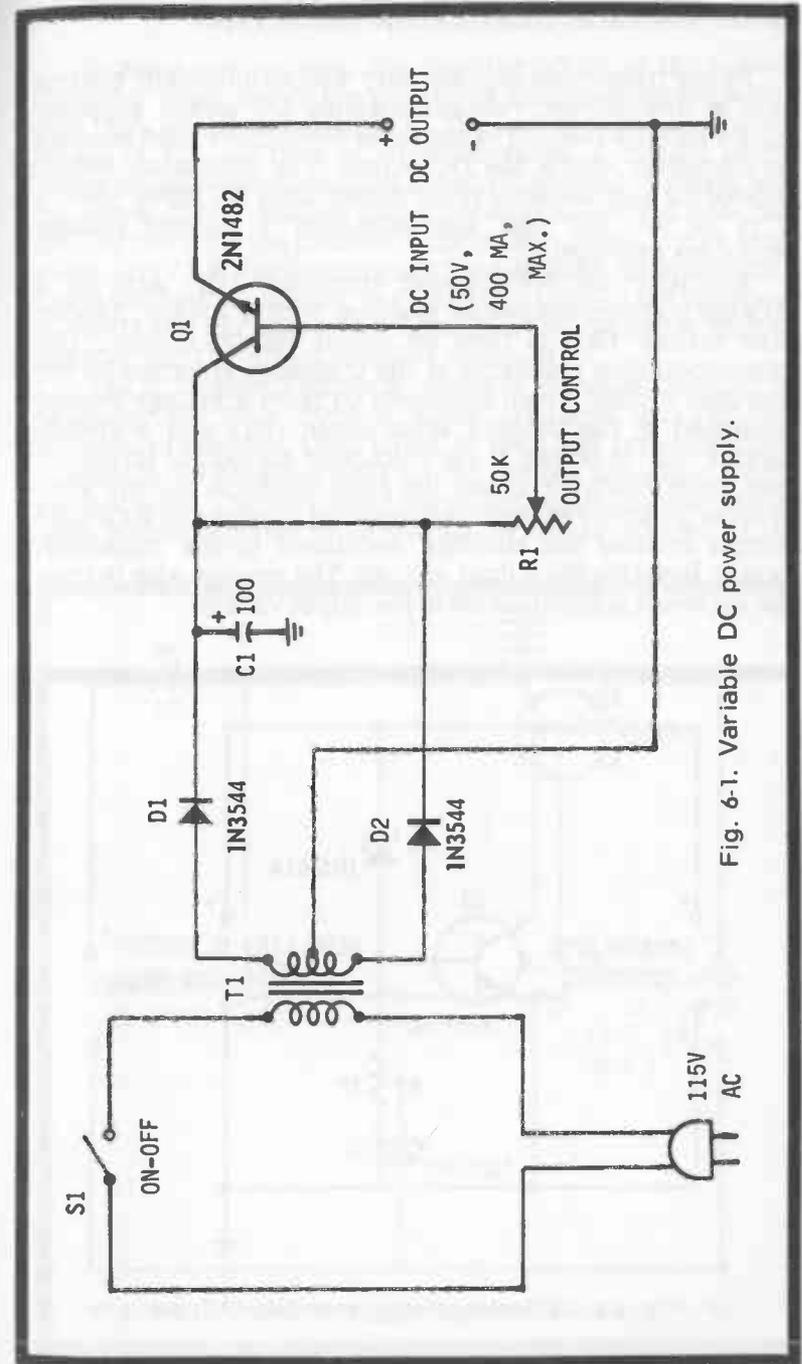


Fig. 6-1. Variable DC power supply.

6.2—DC VOLTAGE REGULATOR (Shunt Type)

Power transistors function very well as automatic control units in low-voltage, voltage-regulated DC power supplies. Fig. 6-2 shows a shunt-type regulator circuit, so called because the transistor shunts the DC output. The maximum output (depending upon the level of the unregulated DC input: 40-50v) is 28 to 30v, 500 ma. Approximately 2 percent voltage regulation provided.

The HEP 247 NPN power transistor (Q1) acts as a variable resistor, connected in series with a 10-ohm, 10-watt fixed resistor (R1) to form an output voltage divider. The collector-emitter resistance of the transistor is varied by DC base-bias current which flows into Q1 from a voltage divider comprised of the 1N5361A zener diode (D1) and a 47-ohm resistor (R2) in series. If the voltage at the output terminals rises, from whatever cause, the diode conducts heavily after its zener point is reached; the resulting increase of Q1's base current reduces the shunting resistance of the transistor, thereby lowering the output voltage. The reverse also is true. The net result is stabilization of the output voltage.

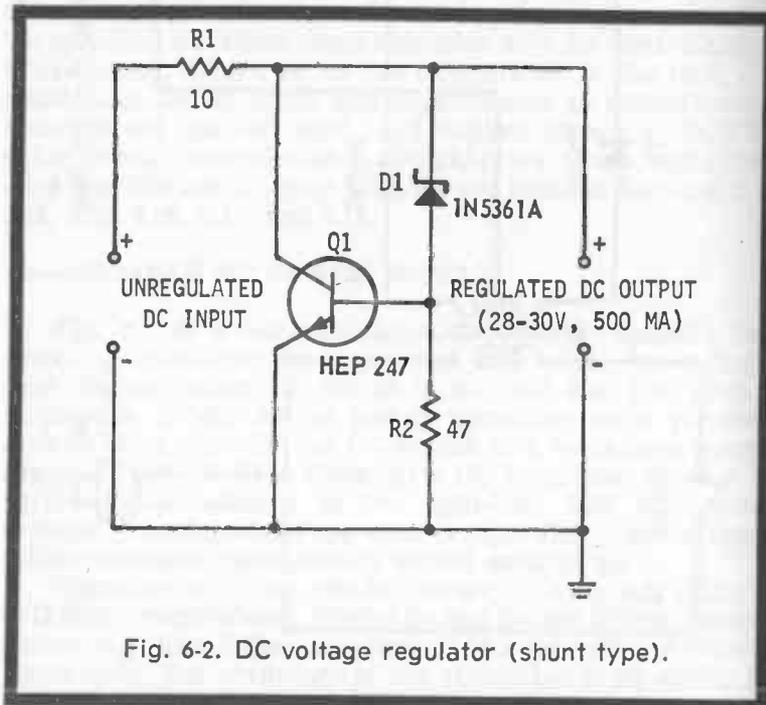


Fig. 6-2. DC voltage regulator (shunt type).

6.3—VOLTAGE REGULATOR (Series Type)

In some applications, the shunting effect of the voltage regulator described in the preceding section is incompatible with the powered equipment. The series-regulator circuit shown in Fig. 6-3 then is more desirable. A 2N1070 silicon power transistor functions as a variable resistor in series with the power supply and the output terminals. The DC base bias of Q1 arises from the constant voltage drop across the 3v zener diode, D1 (TRW Semiconductors Type PD6001, or equivalent). Since this bias is constant, the operating point of the transistor is fixed; and because of the flatness of the collector characteristic, the DC output remains constant.

The 10,000-ohm wirewound rheostat (R2) is set initially for zener-current flow through diode D1, and the 1000-ohm wirewound rheostat (R3) is set to limit the transistor base current to the required level for the desired regulated output-current level. The maximum regulated DC output (depending upon the unregulated DC input: 40-55v) is 50v, 1 amp. R1 is a

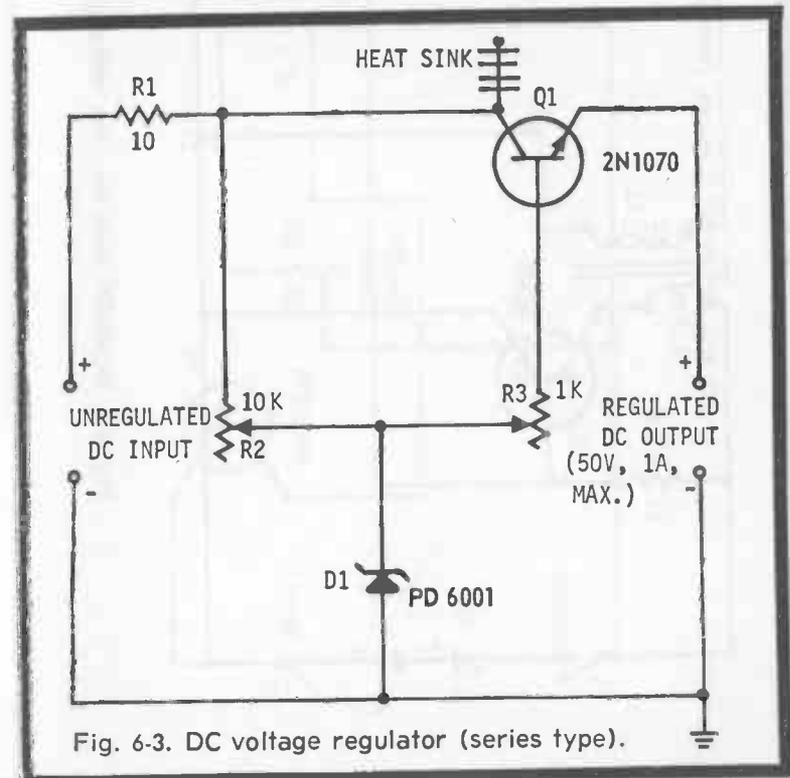


Fig. 6-3. DC voltage regulator (series type).

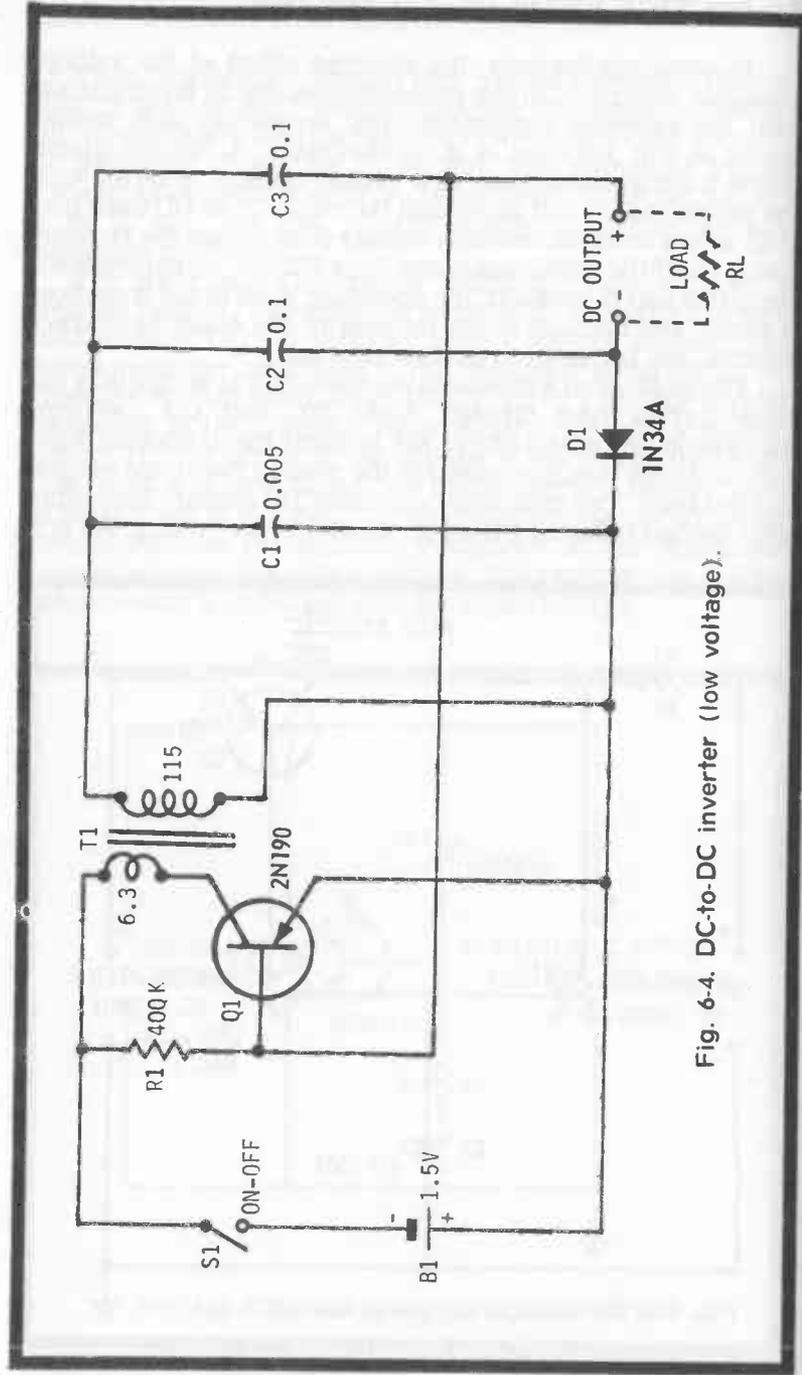


Fig. 6-4. DC-to-DC inverter (low voltage).

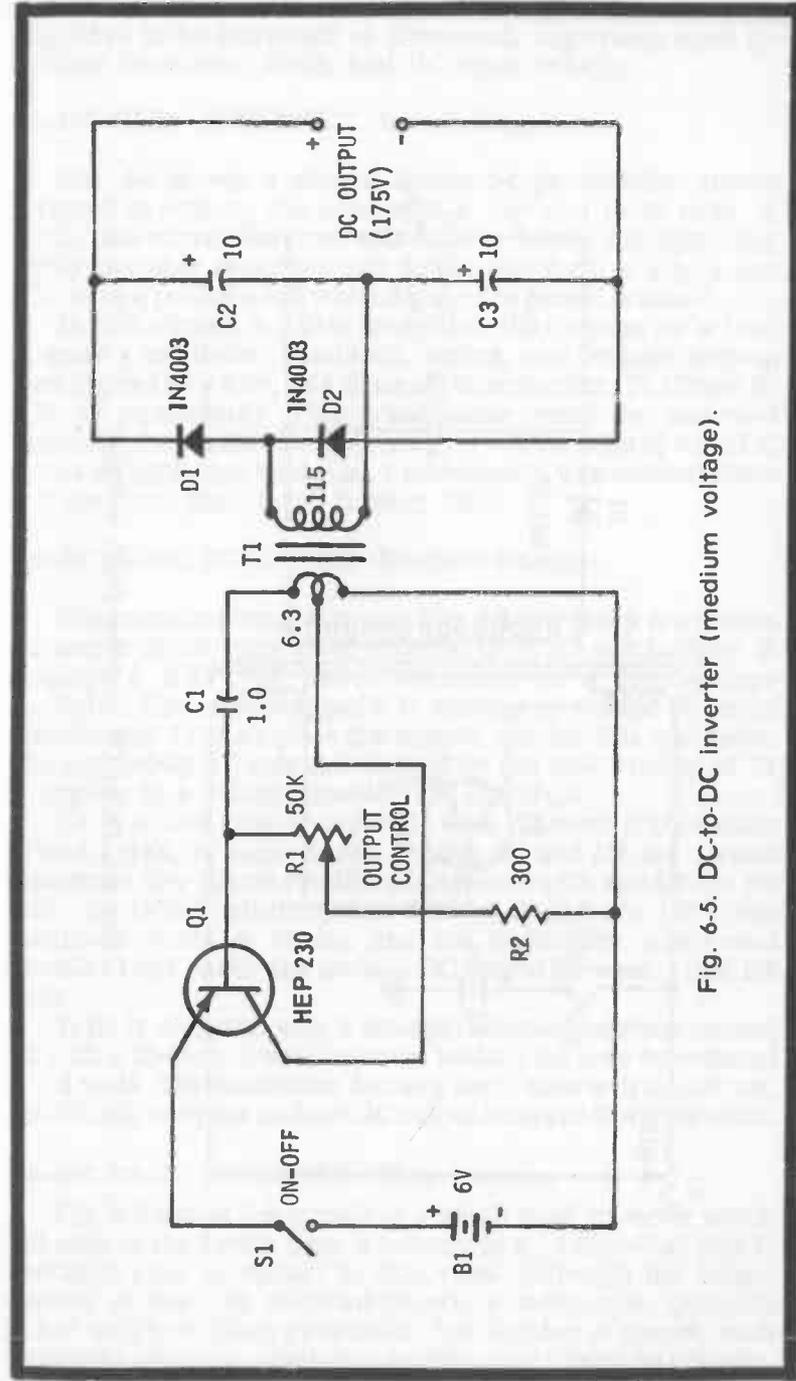


Fig. 6-5. DC-to-DC inverter (medium voltage).

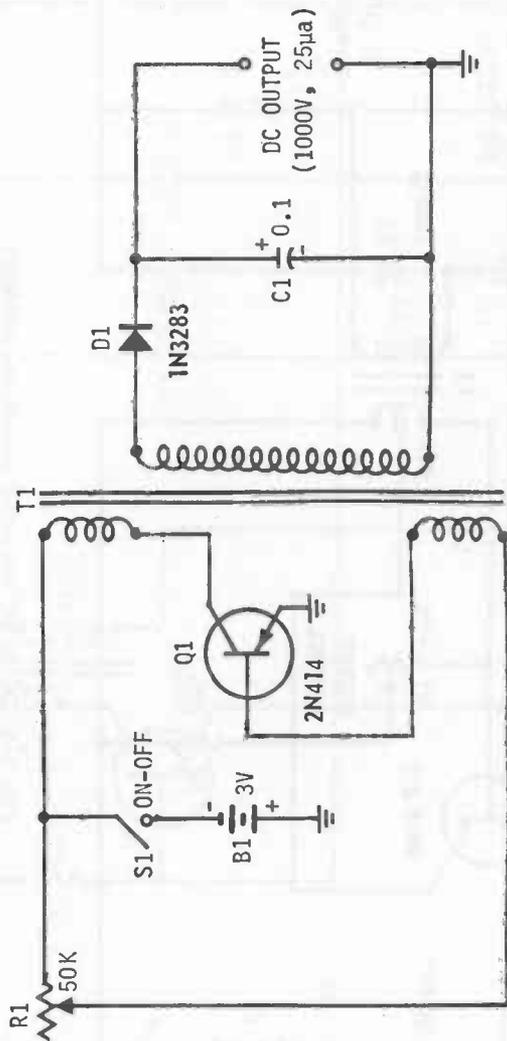


Fig. 6-6. DC-to-DC inverter (high voltage).

10-ohm, 10-watt resistor. The maximum value of rheostat R2 may have to be increased or decreased, depending upon individual transistor, diode, and DC input voltage.

6.4—DC-TO-DC INVERTER (Low-Voltage)

Fig. 6-4 shows a simple Kundert-type inverter circuit designed to step up the voltage of a 1.5v cell to 10 volts. A small, low-current unit of this sort is handy for operating highly portable transistorized equipment from a single dry cell (even a penlight cell when the service period is short).

In this circuit, a 2N190 transistor (Q1) serves as a low-frequency oscillator. Feedback, tuning, and voltage step-up are supplied by a 6.3v, 0.6A filament transformer, T1 (Triad F-13X, or equivalent). This transformer must be polarized correctly, for oscillation. This inverter will develop 10 volts DC across a 10,000-ohm load (i.e., 1 ma output). The current drain is 9 ma from the 1.5-volt battery (B1).

6.5—DC-TO-DC INVERTER (Medium-Voltage)

The inverter circuit shown in Fig. 6-5 delivers a maximum DC output of 175 volts when powered from a 6-volt battery. It employs a HEP 230 power transistor in a Hartley-type oscillator. The center-tapped 6.3v winding of a small filament transformer (T1) supplies the tapped coil for this oscillator. The stepped-up AC output delivered by the 115v winding of T1 is applied to a voltage doubler (D1-D2-C2-C3).

T1 is a 6.3v (center-tapped), 1.2A filament transformer (Triad F-14X, or equivalent). Diodes D1 and D2 are 1N4003 miniature 200v silicon rectifiers. Capacitors C2 and C3 are 10-mfd, 150 DCWV electrolytics. With a 6v battery (B1), the maximum drain is 80 ma and the 50,000-ohm wirewound rheostat (R1) varies the no-load DC output between 2 and 175 volts.

If R1 is replaced with a 200-ohm wirewound rheostat and R2 with a 20-ohm, 1-watt resistor, battery B1 may be reduced to 1.5 volts. The maximum battery drain then will be 100 ma, and R1 will vary the no-load DC output between 60 and 80 volts.

6.6—DC-TO-DC INVERTER (High Voltage)

Fig. 6-6 shows the circuit of a small-sized inverter which will step up the 3 volts from a battery (e.g., 2 Size-C or Size-D flashlight cells in series) to 1000 volts. Although the output current is low (25 microamperes), a miniature, portable power supply of this type is useful in a number of places, such as Geiger counters, insulation testers, and flashtube circuits.

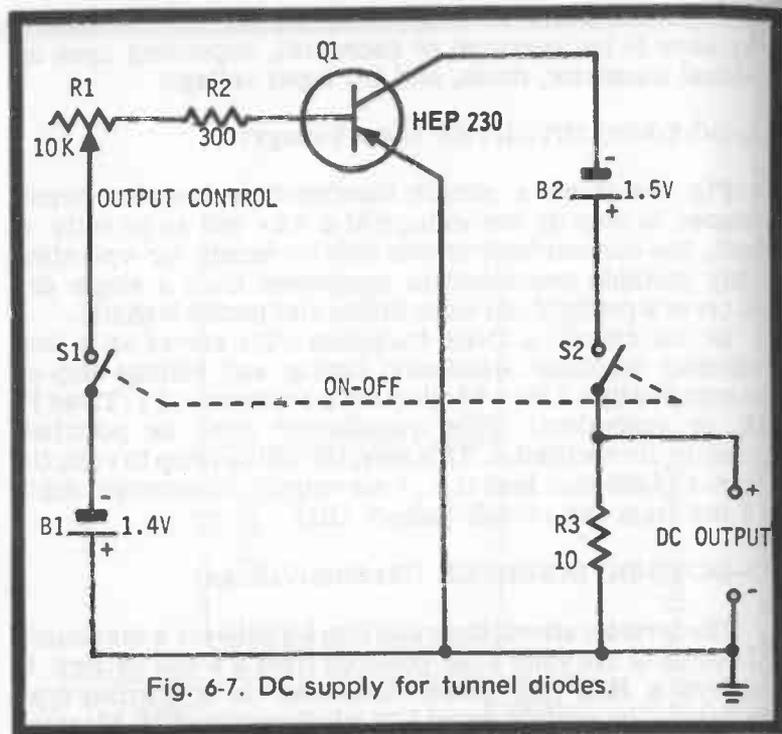


Fig. 6-7. DC supply for tunnel diodes.

The circuit is actually a blocking oscillator (operating frequency is approximately 400 Hz) employing a 2N414 transistor. The feedback and voltage step-up functions are provided by a special miniature high-voltage transformer, T1 (Microtran 8051, or equivalent). The two low-voltage windings of T1 must be polarized correctly, for oscillation. The high output voltage is rectified by the 1N3283, 1500v silicon diode (D1) and filtered by the 0.1-mfd, 1000v tubular capacitor, C1 (Sprague 10TM-P10, or equivalent). The 50,000-ohm wirewound rheostat (R1) is set for reliable oscillation and the highest voltage output. The circuit draws approximately 70 ma from the 3-volt battery (B1).

6.7—DC SUPPLY FOR TUNNEL DIODES

Tunnel diodes require a low-voltage DC supply having good regulation and low output resistance. An ordinary voltage divider operated from batteries or other power supply capable of delivering the required millivolts output usually is not "stiff" enough for correct operation of the tunnel diode.

Fig. 6-7 shows the circuit of a DC supply which has the desired characteristics for tunnel-diode powering. This

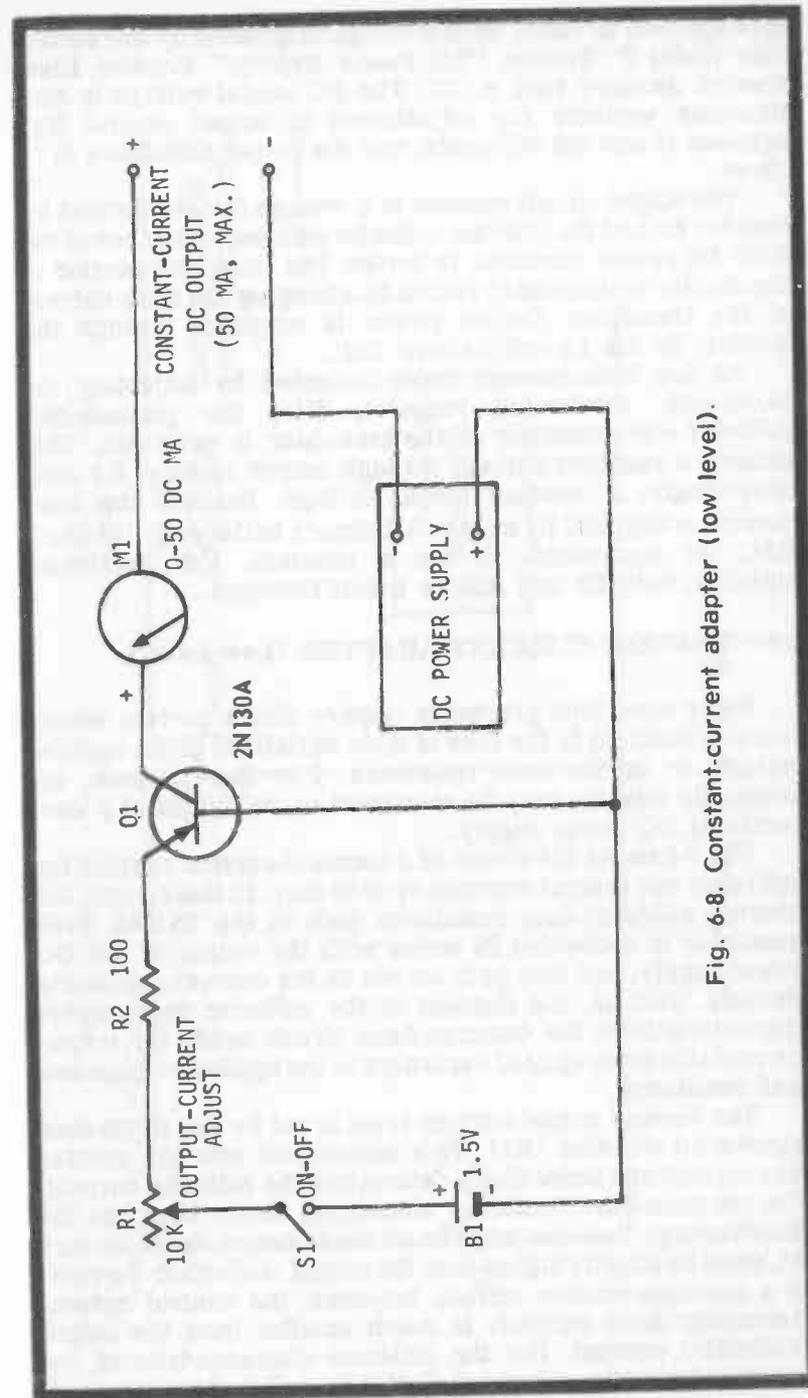


Fig. 6-8. Constant-current adapter (low level).

arrangement is based upon a design originated by the author (see Rufus P. Turner, "TD Power Supply," *Popular Electronics*, January 1962, p. 72). The DC output voltage is continuously variable (by adjustment of output control R1) between 10 and 500 millivolts, and the output resistance is 10 ohms.

The output circuit consists of a voltage divider formed by resistor R3 and the internal collector-emitter resistance of the HEP 230 power transistor in series. The transistor portion of the divider resistance is varied by changing the base current of the transistor. Output power is supplied, through the divider, by the 1.5-volt battery (B2).

At any base-current level (selected by adjusting the 10,000-ohm wirewound rheostat, R1), the pentode-like collector characteristic of the transistor is very flat. This insures a constant current through output resistor R3 and, accordingly, a constant output voltage. Because the base current is supplied by a 1.4-volt mercury battery, B1 (Mallory RM1, or equivalent), it too is constant. For maximum stability, both R2 and R3 are 1-watt resistors.

6.8-CONSTANT-CURRENT ADAPTER (Low Level)

Some electronic processes require direct current which remains constant in the face of wide variations in the applied voltage or in the load resistance. For this purpose, an automatic adapter may be connected to the output of a conventional DC power supply.

Fig. 6-8 shows the circuit of a constant-current adapter for light-duty use (output currents up to 50 ma). In this circuit, the internal collector-base resistance path of the 2N130A PNP transistor is connected in series with the output of the DC power supply, and that path serves as the current-regulating element. That is, the flatness of the collector volt-ampere characteristic in the common-base circuit holds the output current steady, in spite of variations in the applied voltage and load resistance.

The desired output-current level is set by the 10,000-ohm wirewound rheostat (R1). This adjustment sets the emitter bias current and accordingly determines the collector current. The common-base transistor connection shown here has the disadvantage that the emitter current, drawn from battery B1, must be slightly higher than the output (collector) current. In a common-emitter circuit, however, the control current (transistor base current) is much smaller than the output (collector) current. But the collector characteristic of the common-base circuit is much flatter than that of the common-

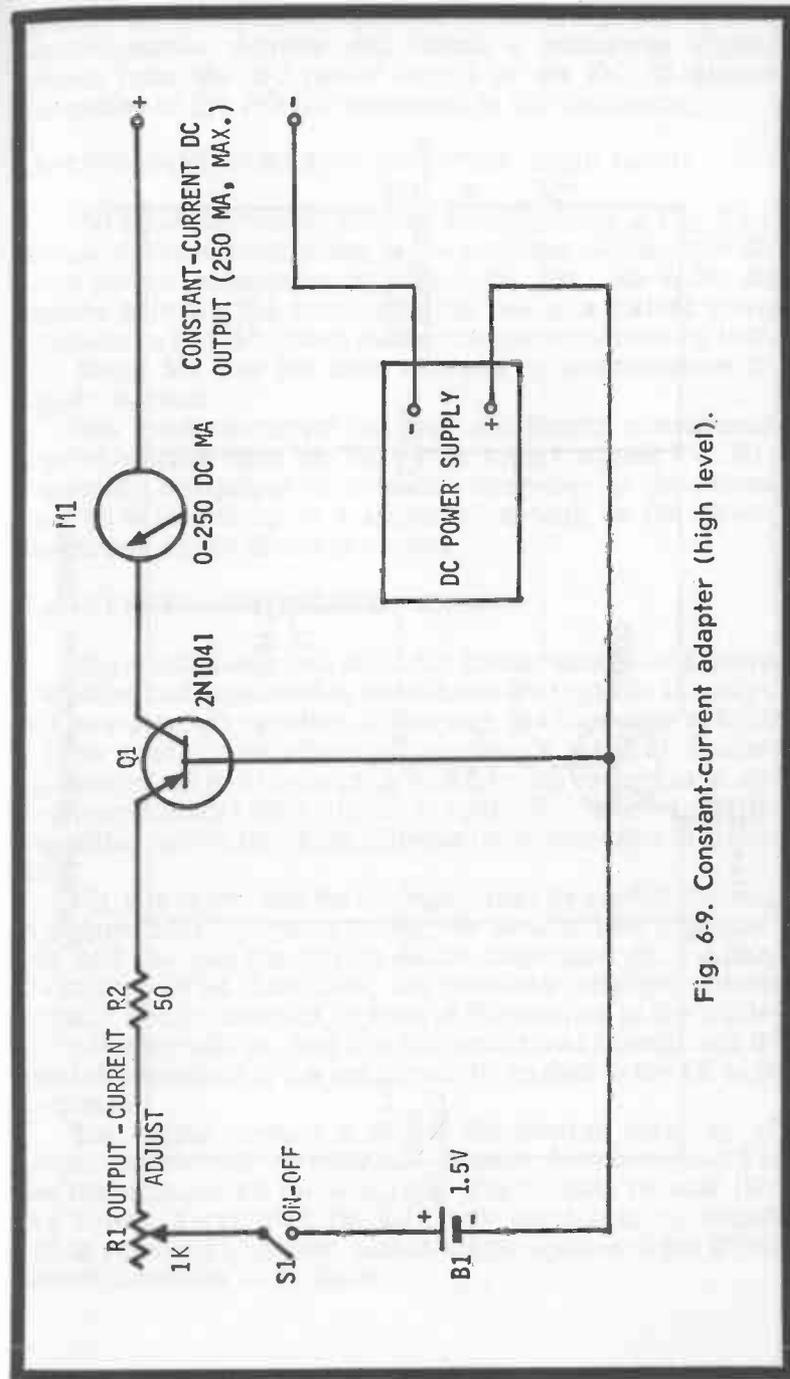


Fig. 6-9. Constant-current adapter (high level).

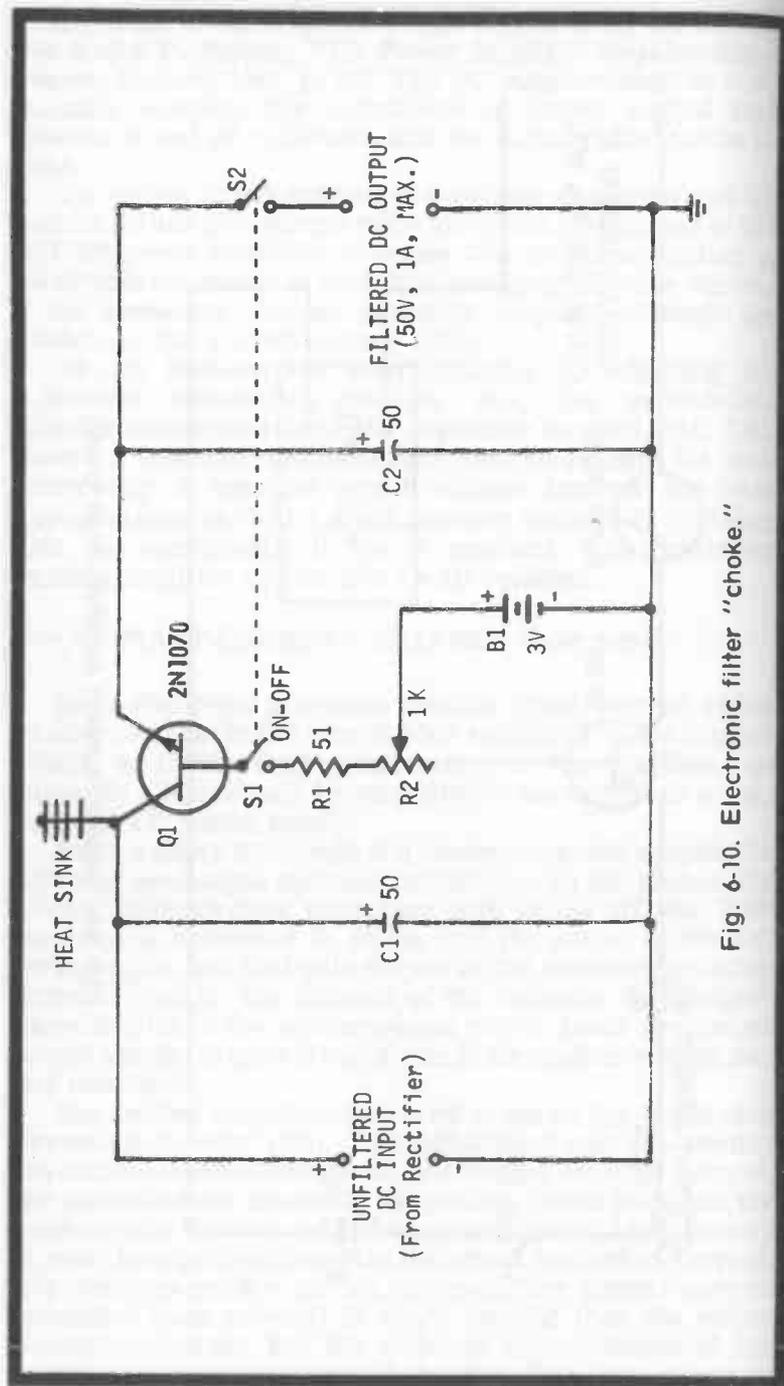


Fig. 6-10. Electronic filter "choke."

emitter and thus affords better current regulation. This constant-current adapter will handle a maximum applied voltage from the DC power supply of 40v DC. Maximum dissipation of the 2N130A transistor is 100 milliwatts.

6.9—CONSTANT-CURRENT ADAPTER (High Level)

The constant-current adapter circuit shown in Fig. 6-9 is similar to the one described in the previous section, and the same circuit explanation applies here, too. The main difference between the circuits is the use of a 2N1041 power transistor in Fig. 6-9, which allows current regulation up to 250 ma. Meter M1 also has been changed to accommodate the higher current.

This constant-current adapter will handle a maximum applied voltage from the DC power supply of 100v DC. Q1's maximum dissipation is 20 watts. Operation of this circuit may be extended up to 3 amperes, as long as the 20-watt dissipation figure is not exceeded.

6.10—ELECTRONIC FILTER "CHOKE"

The constant-current collector characteristic of a power transistor can be utilized to smooth out the ripple in the output of a power-supply rectifier. In this way, the transistor acts like a filter choke. And where high currents (several hundred milliamperes) are involved, a considerable reduction in size over conventional filter chokes is realized. Furthermore, the transistor—unlike the choke coil—has no troublesome magnetic field.

Fig. 6-10 shows how the transistor may be used in this way. A pi-type filter is formed by the two 50-mfd, 150v capacitors (C1 and C2) and the 2N1070 silicon transistor (Q1). At any given level of DC base bias, the transistor collector current tends to remain constant, in spite of fluctuations in the applied DC collector voltage. And it is this action that smooths out the ripple fluctuations in the unfiltered DC applied to the DC input terminals.

The output current is set to the desired level by appropriately setting the base-bias current. Bias is adjusted by the 1000-ohm, 5-watt rheostat (R2). The limiting resistor (R1) is a 51-ohm, 2-watt unit. The maximum obtainable DC outputs are 50 volts and 1 ampere. Maximum dissipation of the 2N1070 power transistor is 50 watts.

Chapter 7

Receivers, Transmitters & Accessories

With the birth of the transistor, the radioman's dream of a self-contained superhet receiver small enough to fit into a shirt pocket at last came true. What had previously been science fiction became everyday fact. Today, a receiver with seven transistor stages, and including battery, loudspeaker, and antenna, is no bigger than a cigarette package. In TV receivers, and in transmitters also, the transistor has found increasing use. Because of its relatively low power-handling capacity, however, the transistor has not completely supplanted the vacuum tube in transmitters. Nevertheless, some low-powered transmitters (especially CB units) are now completely solid state.

Presently, some transistorized radio receivers are so inexpensive that an experimenter could not duplicate one at twice the retail price. So nobody would build such a set just to save money. The hobbyist and student, however, will find diversion and education in simple receiver and transmitter circuits with which to experiment. Here are several receiver and transmitter applications, each of which employs a single transistor. These circuits are entirely practical. Moreover, some of them may be combined with other circuits, described earlier in this book, to form more complicated setups.

Current, voltage, signal amplitude, and frequency values given in this chapter are those obtained in tests of the author's models; your readings may vary somewhat due to individual transistor and circuit component characteristics and different receiving locations. Where particular components from named manufacturers are listed, these components (or their exact equivalents, when obtainable) appear to be necessary for proper performance of the circuit. In all other instances, however, any component having the specified electrical characteristics may be used. Except where shown otherwise on the diagram or in the text, all resistances are in ohms and capacitances in microfarads; resistors are one-half watt, and capacitors are 25 DCWV. Aside from audio amplifiers

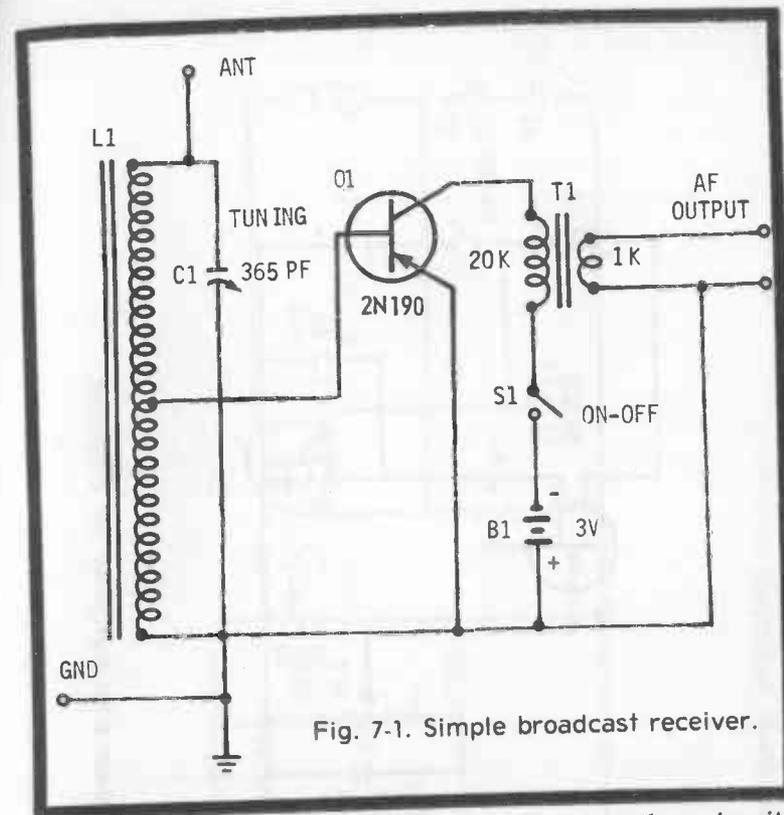


Fig. 7-1. Simple broadcast receiver.

(Chapter 1) and power supplies (Chapter 6), other circuits described in earlier chapters, which might be useful also in communications systems, may be found in Sections 2.1-2.4, 2.8, 3.6, 3.7-3.9, and 3.13.

7.1—SIMPLE BROADCAST RECEIVER

Fig. 7-1 is a simple broadcast receiver circuit employing a 2N190 transistor (Q1) as the detector. The self-contained antenna (L1) is a flat, ferrite-strip type (J. W. Miller No. 2004, or equivalent), which comes equipped with a low-impedance tap used here to match the input circuit of the transistor to the antenna. Strong, nearby stations may be picked up directly with this antenna when the receiver is rotated for maximum signal pickup. For other stations, an external antenna and ground will be required. The tuning range (with a 365-pf variable capacitor, C1) is 540 to 1650 kHz.

The circuit delivers an audio output at the secondary of transformer T1, a 20,000-to-1000-ohm unit (Argonne AR-104, or equivalent). Headphones or an external audio amplifier may

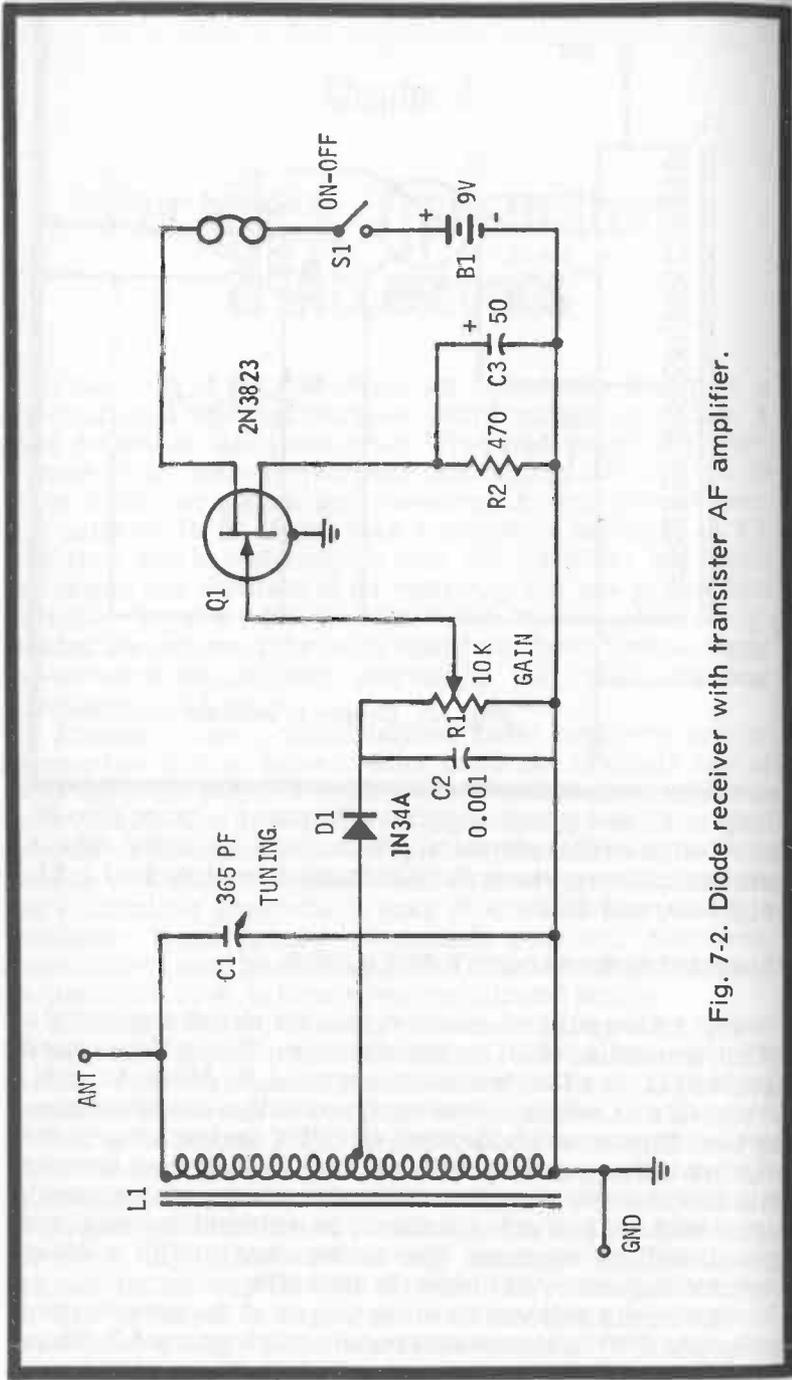


Fig. 7-2. Diode receiver with transistor AF amplifier.

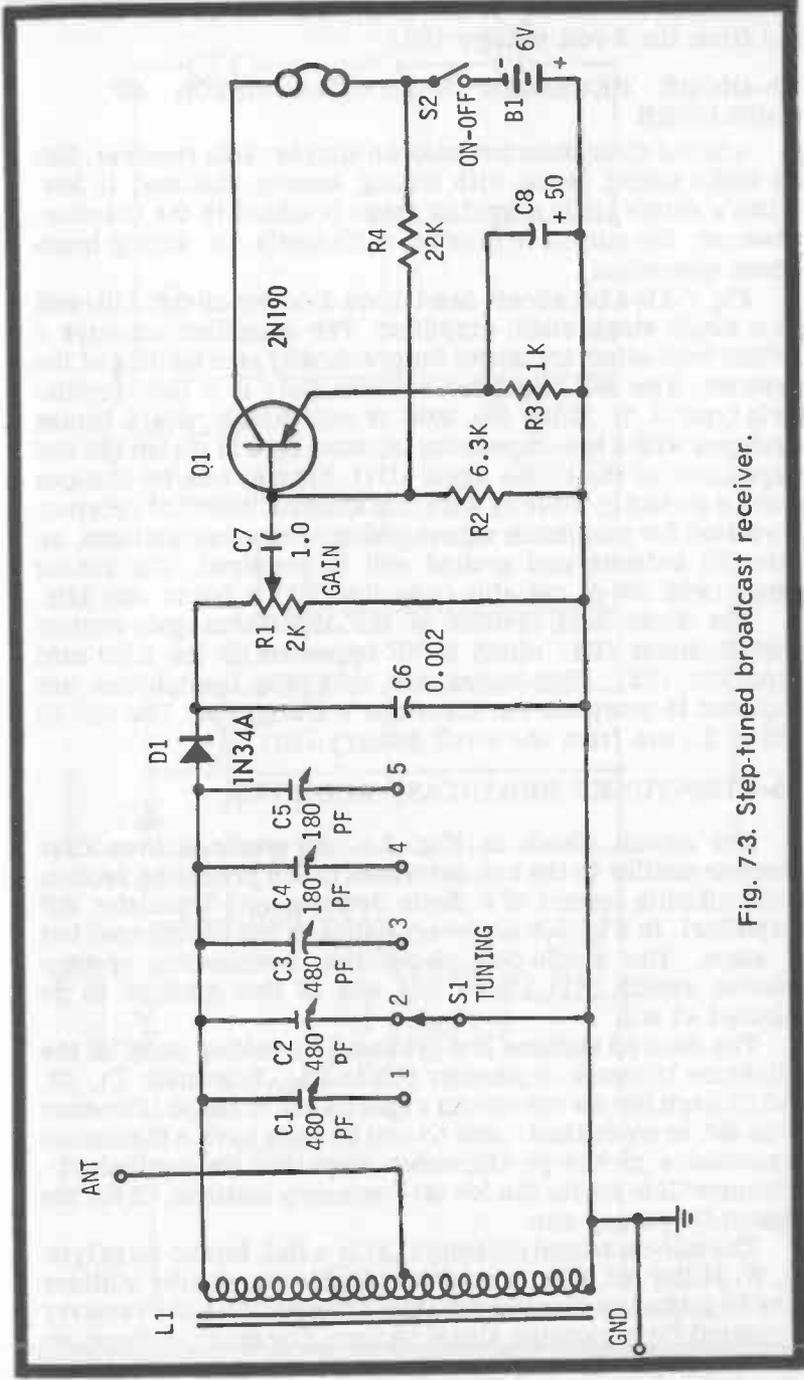


Fig. 7-3. Step-tuned broadcast receiver.

be connected to the AF output terminals. The circuit draws 1 ma from the 3-volt battery (B1).

7.2—DIODE RECEIVER WITH TRANSISTOR AF AMPLIFIER

A tuned diode detector makes a simple radio receiver. But its audio output (even with strong, nearby stations) is low. When a single audio amplifier stage is added to the detector, however, the output is boosted sufficiently for strong head-phone operation.

Fig. 7-2 is a broadcast-band diode detector circuit followed by a single-stage audio amplifier. The amplifier employs a 2N3823 field-effect transistor for practically zero loading of the detector. The self-contained antenna (L1) is a flat, ferrite-strip type (J. W. Miller No. 2004, or equivalent), which comes equipped with a low-impedance tap used here to match the low impedance of the 1N34A diode (D1). Strong, nearby stations may be picked up directly with this antenna when the receiver is rotated for maximum signal pickup. For other stations, an external antenna and ground will be required. The tuning range (with 365-pf variable capacitor, C1) is 540 to 1650 kHz.

The diode load resistor is the 10,000-ohm gain-control potentiometer (R1) which is RF-bypassed by the 0.001-mfd capacitor (C2). High-impedance magnetic headphones are required to complete the transistor's DC circuit. The circuit draws 2.1 ma from the 9-volt battery (B1).

7.3—STEP-TUNED BROADCAST RECEIVER

The circuit shown in Fig. 7-3 represents a broadcast receiver similar to the one described in the preceding section (both circuits consist of a diode detector and transistor AF amplifier). In Fig. 7-3, however, tuning is not continuous, but in steps. The single-pole, 5-position, nonshorting, rotary selector switch (S1) allows any one of five stations to be selected at will.

The desired stations are pretuned by setting each of the miniature trimmer capacitors (C1 to C5). Trimmers C1, C2, and C3 each have a maximum capacitance of 480 pf (Elmenco Type 466, or equivalent), and C4 and C5 each have a maximum capacitance of 180 pf (Elmenco Type 463, or equivalent). Trimmer C1 is set for the lowest-frequency stations, C5 for the highest-frequency one.

The self-contained antenna (L1) is a flat, ferrite-strip type (J. W. Miller No. 2004, or equivalent). Strong, nearby stations may be picked up directly with this antenna when the receiver is rotated for maximum signal pickup. For other stations, an external antenna and ground will be required.

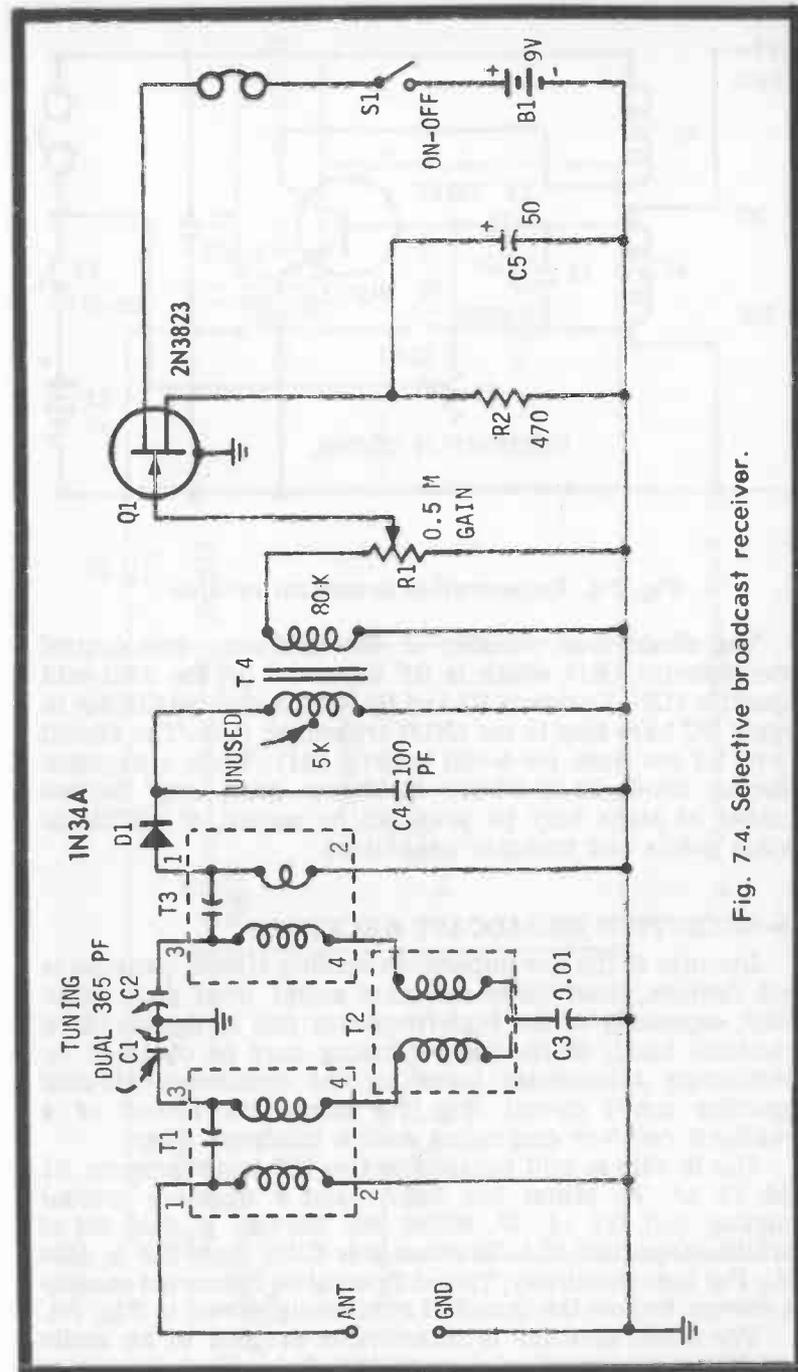


Fig. 7-4. Selective broadcast receiver.

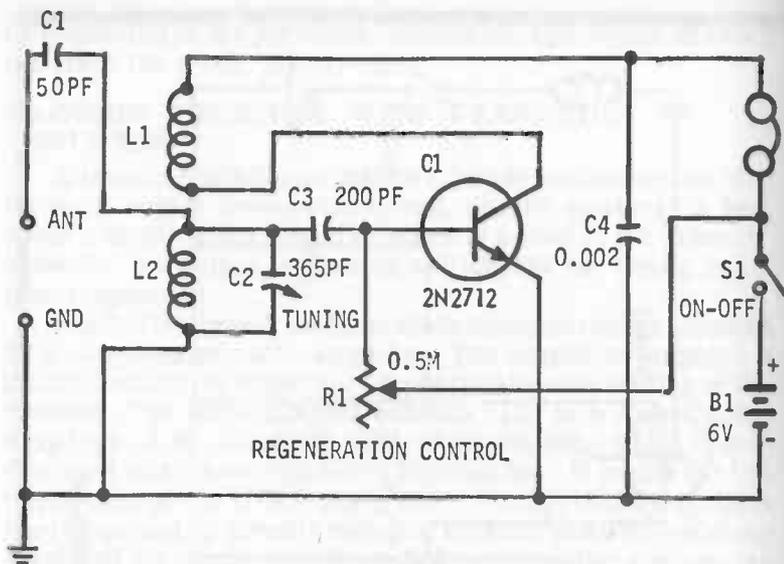


Fig. 7-5. Regenerative broadcast receiver.

The diode load resistor is the 2000-ohm gain-control potentiometer (R1) which is RF-bypassed by the 0.001-mfd capacitor (C6). Resistors R2 and R4 form a voltage divider to supply DC base bias to the 2N190 transistor (Q1). The circuit draws 2.1 ma from the 6-volt battery (B1). While a 5-station selector (S1-C1-C2-C3-C4-C5) is shown here, any desired number of steps may be provided by means of additional switch points and trimmer capacitors.

7.4—SELECTIVE BROADCAST RECEIVER

Because of the low-impedance loading effects common to such devices, most diode detectors suffer from poor selectivity, especially at the high-frequency end of the standard broadcast band. Much sharper tuning may be obtained by substituting a bandpass tuner for the common single-coil/capacitor tuned circuit. Fig. 7-4 shows the circuit of a broadcast receiver employing such a bandpass filter.

The bandpass unit consists of two RF transformers: T1 and T3 (J. W. Miller No. 242-A) and a negative mutual coupling coil, T2 (J. W. Miller No. EL-56). A dual 365-pf variable capacitor (C1-C2) tunes this filter from 540 to 1650 kHz. For best selectivity, T1 and T3 must be connected exactly as shown: Follow the terminal numbering shown in Fig. 7-4.

The diode detector is transformer-coupled to an audio amplifier stage employing a 2N3823 field-effect transistor

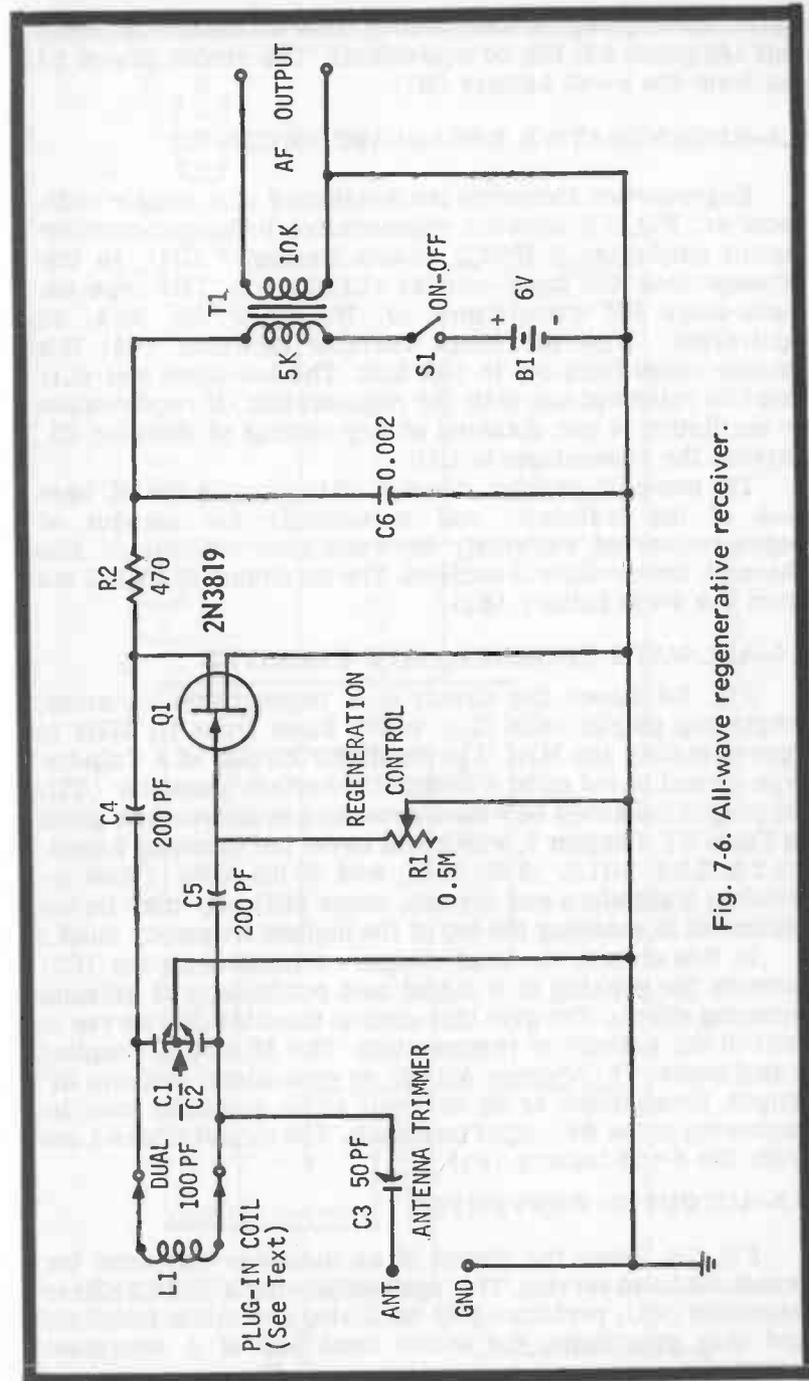


Fig. 7-6. All-wave regenerative receiver.

(Q1). The coupling transformer, T4, is a miniature 5K-to-80K unit (Argonne AR-158, or equivalent). The circuit draws 2.1 ma from the 9-volt battery (B1).

7.5—REGENERATIVE BROADCAST RECEIVER

Regeneration increases the sensitivity of a simple radio receiver. Fig. 7-5 shows a regenerative broadcast-receiver circuit employing a 2N2712 silicon transistor (Q1). In this arrangement, the input coupler (L1-L2) is a TRF-type antenna-stage RF transformer (J. W. Miller No. 20-A, or equivalent). With the 365-pf variable capacitor (C2) this coupler tunes from 540 to 1700 kHz. The low-turns coil (L1) must be polarized correctly for regeneration (if regeneration or oscillation is not obtained at any setting of rheostat R1, reverse the connections to L1).

The one-half-megohm rheostat (R1) controls the DC base bias of the transistor and accordingly the amount of regeneration. At extremely low-resistance settings of this rheostat, the circuit will oscillate. The maximum drain is 2 ma from the 6-volt battery (B1).

7.6—ALL-WAVE REGENERATIVE RECEIVER

Fig. 7-6 shows the circuit of a regenerative receiver, employing plug-in coils (L1) which tunes from 1.1 MHz to approximately 100 MHz. The oscillator circuit is a Colpitts-type circuit based upon a 2N3819 field-effect transistor (Q1). Six plug-in coils may be wound according to instructions given in Table 5-2, Chapter 5, which will cover the following bands: 1.1-2.5, 2.5-5, 5-11.5, 10-25, 20-45, and 40-100 MHz. (With individual transistors and layouts, some difficulty may be experienced in reaching the top of the highest frequency band.)

In this circuit, the 50-pf midjet variable capacitor (C3) permits the peaking of a signal and minimizing of antenna detuning effects. The gate bias-control rheostat (R1) serves to control the amount of regeneration. The 5K-to-10K coupling transformer, T1 (Argonne AR-155, or equivalent) delivers AF output. Headphones or an external audio amplifier may be connected to the AF output terminals. The circuit draws 1 ma from the 6-volt battery (B1).

7.7—AUTODYNE CONVERTER

Fig. 7-7 shows the circuit of an autodyne converter for broadcast-band service. This unit, employing a 2N2712 silicon transistor (Q1), performs both oscillator and mixer functions and thus constitutes the entire front end of a broadcast superhet.

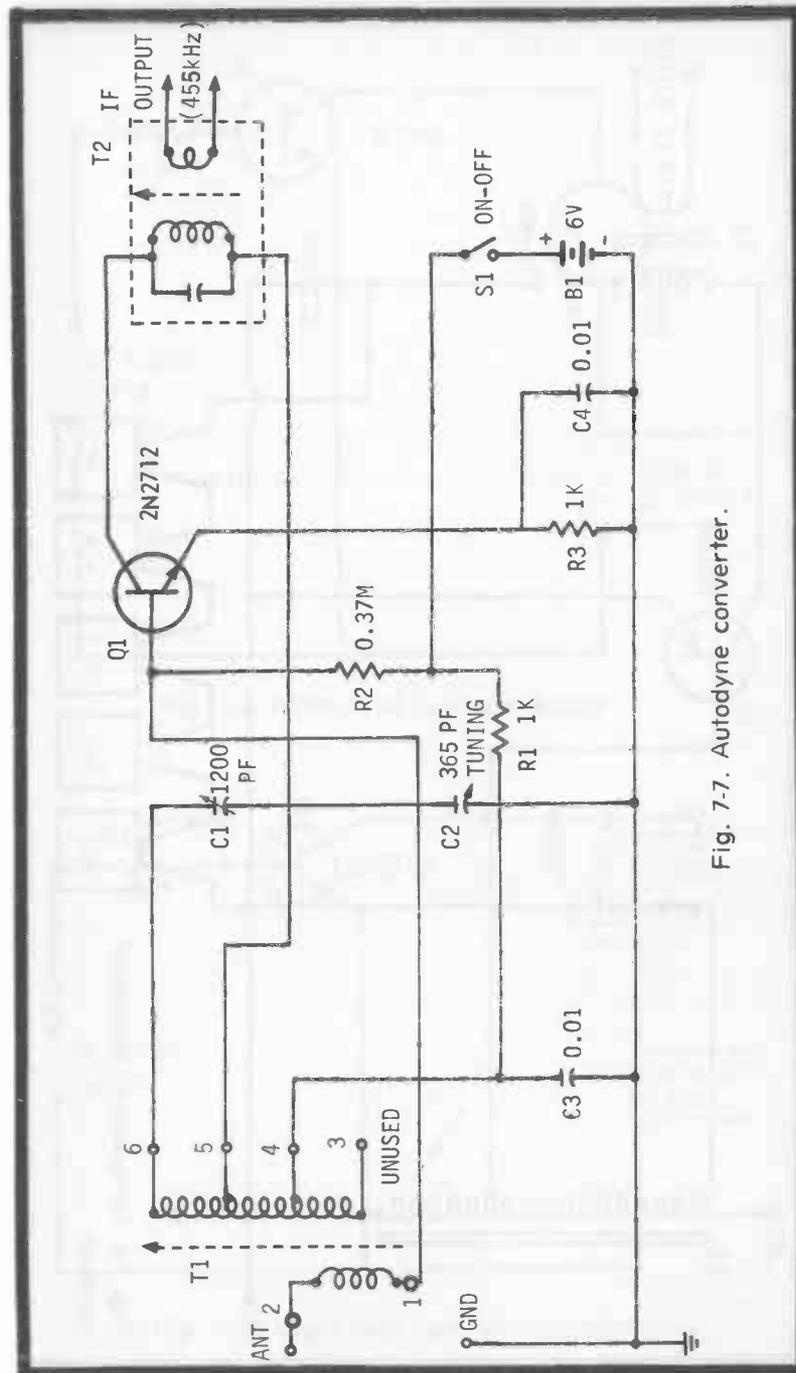


Fig. 7-7. Autodyne converter.

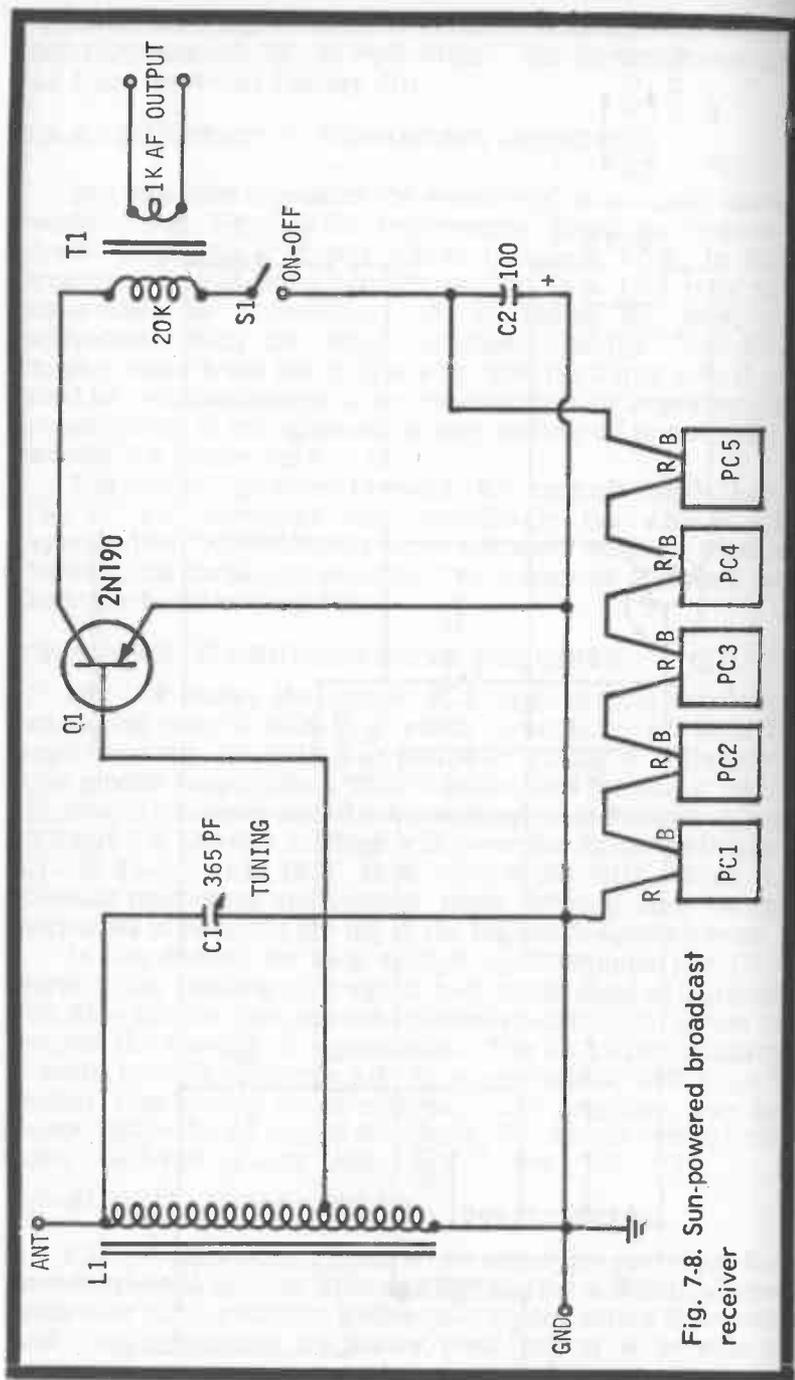


Fig. 7-8. Sun-powered broadcast receiver

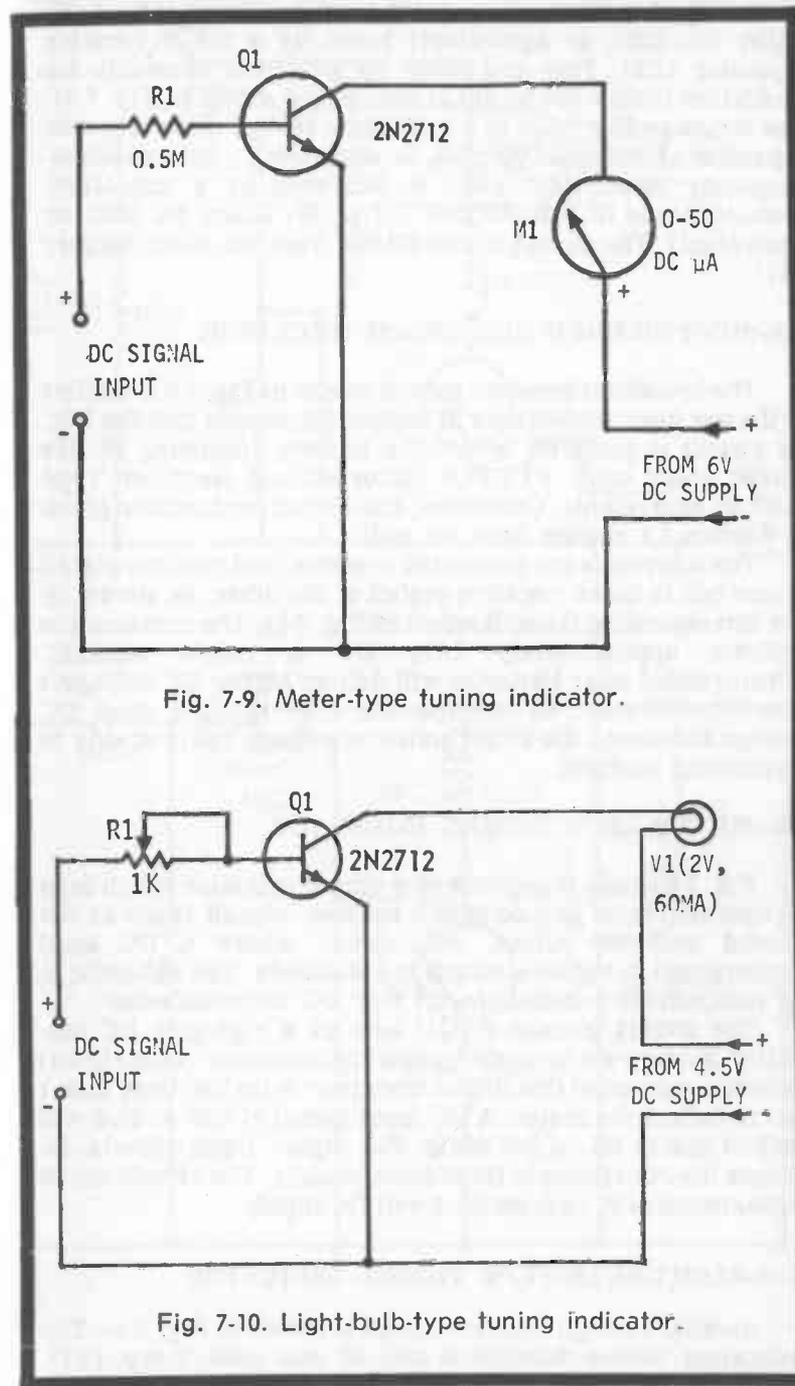


Fig. 7-10. Light-bulb-type tuning indicator.

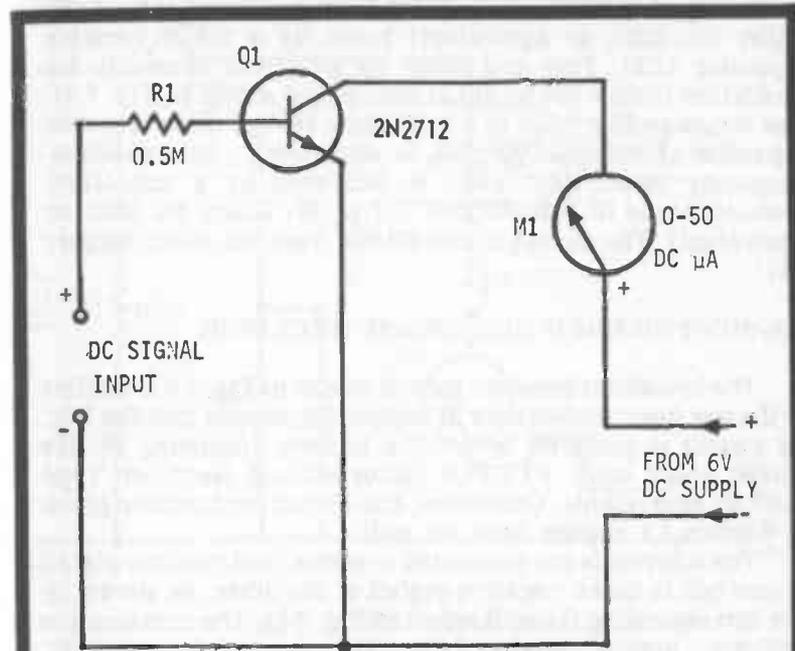


Fig. 7-9. Meter-type tuning indicator.

Maximum signal is indicated by the brightest glow of the lamp.

The 2N2712 transistor (Q1) acts as a DC amplifier. No zero-set (lamp extinguisher) is needed, since the quiescent (zero-signal) collector current of this silicon transistor is too low (less than 1 μ a) to light the lamp. With rheostat R1 set to its maximum resistance of 1000 ohms, a DC input signal of 1.5v at 1 ma will produce maximum brilliance. The lamp goes out when the signal drops to 0.7 volt. The circuit draws a maximum of 60 ma from the 4½-volt DC supply.

7.11—CW TRANSMITTER

Fig. 7-11 shows the circuit of a simple crystal oscillator-type CW transmitter, employing a U222 field-effect transistor (Q1). While the RF power output of this transmitter is only 0.6 watt, it can be effective when communication conditions are favorable.

The circuit is tuned and operated in the conventional manner. The combination L1-L2 is a commercial low-power transmitter "coil" chosen to resonate with 100-pf variable capacitor C3 at the crystal frequency. (If desired, the reader may wind his own coils according to instructions found in amateur radio handbooks.)

After being tuned to crystal resonance, as indicated by the dip of the 0-50 DC milliammeter (M1), the circuit then may be loaded by the antenna to 50 ma. At 24 volts, this represents a DC power input of 1.2 watt. Higher output power may be obtained with heavier and more expensive RF power transistors. For example, Type 2N5644 (Motorola) will deliver 12 watts at 12.5v DC, and Type SRD54117 (Solitron Devices) 60 watts at 28v DC.

7.12—FREQUENCY DOUBLER

The circuit in Fig. 7-12 is a frequency doubler. Multipliers of this type are common in the exciter sections of transmitters. For doubling, the input frequency f is converted to an output frequency $2f$. (The L1-C4 tank is tuned to $2f$.) This circuit employs a U222 field-effect transistor operated at 24v, 50 ma, and its RF output is approximately 0.6 watt. The circuit is tuned and operated in the conventional manner.

The tank inductor (L1) is a commercial, low-power transmitter coil chosen to resonate with 100-pf variable capacitor C4 at twice the input frequency. (If desired, the reader may wind his own coil according to instructions found in amateur radio handbooks.) After being tuned to frequency

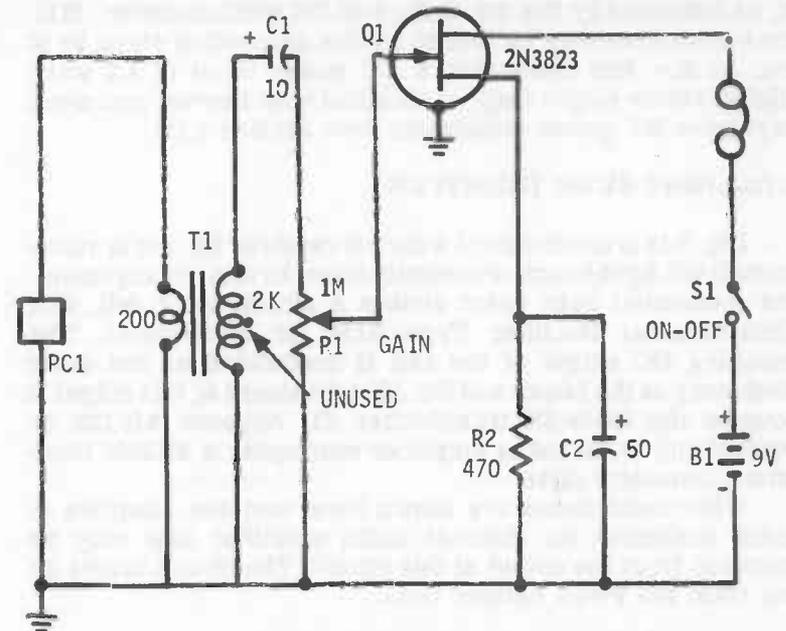


Fig. 7-13. Light-beam receiver.

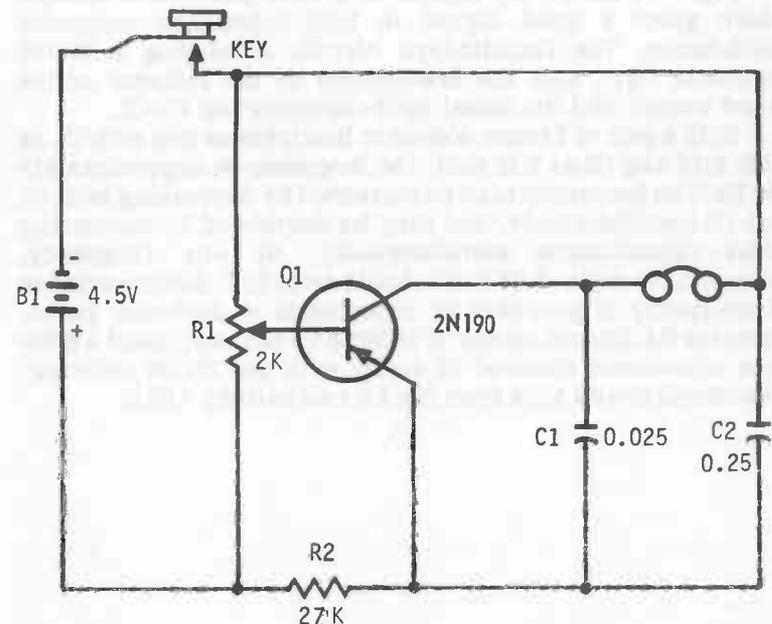


Fig. 7-14. Code-practice oscillator.

2f, as indicated by the dip of the 0-50 DC milliammeter (M1), the circuit then may be loaded by the succeeding stage to 50 ma. At 24v, this represents a DC power input of 1.2 watt. Higher power output may be obtained with heavier and more expensive RF power transistors (see Section 7.11).

7.13—LIGHT-BEAM RECEIVER

Fig. 7-13 is the circuit of a simple receiver for use in voice-modulated light-beam communications. In this arrangement, the modulated light beam strikes a silicon solar cell, PC1 (International Rectifier Type S1M, or equivalent). The resulting DC output of the cell is modulated at the same frequency as the beam, and the AF component of this output is coupled (by 200-to-2K transformer T1: Argonne AR-123, or equivalent) to an audio amplifier employing a 2N3823 field-effect transistor (Q1).

While headphones are shown here, and are adequate in many instances, an external audio amplifier also may be operated from the output of this circuit. The circuit draws 2.1 ma from the 9-volt battery (B1).

7.14—CODE-PRACTICE OSCILLATOR

Fig. 7-14 shows the circuit of a code-practice oscillator which gives a good signal in high-impedance magnetic headphones. The Colpitts-type circuit, employing a 2N190 transistor (Q1), uses the headphones as the inductor of the tuned circuit and the usual split-capacitor leg C1-C2.

With a pair of Trimm 2000-ohm headphones and with C1 as 0.025 mfd and C2 as 0.25 mfd, the frequency is approximately 750 Hz. The frequency may be increased by decreasing both C1 and C2 simultaneously, and may be decreased by increasing these capacitances simultaneously. At any frequency, however, the ratio of C2 to C1 should be 10 to 1. Some variation of frequency is provided by adjustment of 2000-ohm potentiometer R1. Should control of volume be desired, insert a 5000-ohm wirewound rheostat in series with the 2N190 collector. The circuit draws 3 ma from the 4.5-volt battery (B1).

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