

transistor circuits

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contents

CHAPTER 1 . . . AUDIO AMPLIFIERS 7

Single-stage, R-C-coupled, common-base amplifier. Single-stage, transformer-coupled, common-base amplifier. Single-stage, R-C-coupled, common-emitter amplifier. Single-stage, transformer-coupled, common-emitter amplifier. Single-stage, R-C-coupled, common-collector amplifier. Single-stage transformer-coupled, common-collector amplifier. Multistage, R-C-coupled amplifier. Multistage, transformer-coupled amplifier. Push-pull output circuits. Phase inverters. Class-B loudspeaker amplifier with conventional transistors. Class-B loudspeaker amplifier with power transistors. Class-A loudspeaker amplifier with power transistor. Complementary-symmetry power amplifier with conventional transistors. Complementary-symmetry power amplifier with power transistors. Single-stage low-noise preamplifier. Preamplifier with high-impedance input. Boosting transistor input impedance.

CHAPTER 2 . . . RF AND IF AMPLIFIERS 29

Broadcast-band rf amplifier and preselector. Single-stage, single-tuned 455-kc if amplifier. Two-stage, 455-kc if amplifier. 50-kc if amplifier. Transistors for high-frequency rf amplifiers.

CHAPTER 3 . . . DC AMPLIFIERS 37

Single-stage current amplifier. Single-stage current amplifier with fixed-resistance input. Cascaded p-n-p current amplifier. Cascaded p-n-p-n-p-n current amplifier. Single-stage, dc voltage amplifier. Cascaded dc voltage amplifier. Single-stage, power-type current amplifier. Two-stage, high-gain, power-type current amplifier. Alpha-beta conversion.

CHAPTER 4 . . . OSCILLATORS 45

Single-frequency audio oscillator. Simple code-practice oscillator. Loudspeaker-operating code-practice oscillator. Phase-shift audio oscillator. Step-type variable-frequency audio oscillator. Tuned, variable-frequency audio oscillator. 100-kc crystal oscillator. Self-excited 100-kc oscillator. High-frequency crystal oscillator. Self-excited rf oscillator. Wide-range rf oscillator. Multivibrator. Blocking oscillator. Light-powered oscillators. Oscillator operating notes.

CHAPTER 5 . . . POWER SUPPLIES 59

Variable low-current dc power supply. Variable high-current dc power supply. Full-wave variable dc power supply. Light-duty dc voltage regulator. Heavy-duty dc voltage regulator. Light-duty constant-current adaptor. Heavy-duty constant-current adaptor. High-voltage dc-to-dc power supply.

CHAPTER 6 . . . RADIO RECEIVERS	69
Single-transistor broadcast receiver. Two-transistor broadcast receiver. Miniaturization of simple broadcast tuners. Regenerative broadcast receiver. Step-tuned diode-transistor broadcast receiver. Push-pull transistor broadcast receiver, Loop-operated diode-transistor broadcast receiver. Sun-powered broadcast receiver. Superheterodyne broadcast receiver. All-wave regenerative receiver. High-frequency cueing receiver.	
CHAPTER 7 . . . TRIGGERS AND SWITCHES	81
Flip-flop. One-shot multivibrator. Pulse delay circuit. Simple coincidence circuit. Pulse inverter.	
CHAPTER 8 . . . CONTROL DEVICES	89
Sensitive dc relay. Ultra-sensitive dc relay. Sensitive switching circuit for light-duty contacts. Sensitive ac relay. Sensitive rf relay. Ultra-sensitive rf relay. Heavy-duty relay booster. Sensitized heavy-duty relay booster. Sound-operated relay. Light-operated relay. Photoelectric counter. Phototimer. Miniature-motor control. General-purpose radio control relay. Carrier failure alarm. Impulse counter.	
CHAPTER 9 . . . TEST INSTRUMENTS	105
Low-gain single-transistor dc microammeter. Zero-setting circuits for microammeter. High-gain dc microammeter. Supersensitive dc microammeter. Electronic dc voltmeter. Audio voltmeter-millivoltmeter. Millivolt adaptor for ac vacuum-tube voltmeter. Sensitizer for high-range dc milliammeter. TV antenna compass. TV field-strength meter. Bridge null detector. Sharply tuned null detector. Crystal-type sound marker generator. Audio signal injector. Crystal-type set aligner. Rf signal generator. Radio-frequency comparator. Heterodyne frequency meter. Sound-level [noise] meter. Af-rf signal tracer. "Grid"-dip oscillator. Low-frequency "grid"-dip oscillator. "Grid"-dip adaptor. Sensitive light meter.	
CHAPTER 10 . . . AMATEUR DEVICES	135
Keying monitor. Phone monitor. Speaker-type dynamic microphone. Signal peaker (CW filter). Field-strength meter. Modulation monitor. CW transmitter. Conelrad receiver. Transistors in conventional amateur receivers.	
CHAPTER 11 . . . MISCELLANEOUS CIRCUITS	145
Simple Geiger counter. Geiger counter with transistorized high-voltage supply. Phono oscillator. Wireless microphone. Photoelectric intrusion alarm. Transistorized 0.6-watt intercom. Hearing aid. Diode-type transistor power supplies. Safety diode for transistor circuits.	
INDEX	157

introduction

APPPLICATION of the transistor to electronic circuitry is a striking example of technological progress. Eight years prior to this writing, very few knew how the transistor worked and some were openly skeptical of its future. Today, many people need transistor circuit information. They range all the way from the scientist and engineer to the casual experimenter.

This book is a collection of practical transistor circuits, each of which has been tested by the author in his own laboratory. These circuits are presented with the expectation that they will save many hours of design time. As building blocks with values already worked out, they may be employed singly or (in systems) in combination with each other or with other circuitry. Slight modifications of component values will allow the reader to obtain results different from those originally intended. A study of the chapter headings will show that an effort has been made to cover a wide range of interests.

The circuits in this book have been designed by many persons or are based upon their inventions. No claim is made or implied that they are of the author's invention. Transistorized circuits have become so numerous that already their use is becoming "customary." That is, a certain circuit becomes so useful and commonplace (even as with vacuum-tube circuits) that eventually the originator's name is dropped, although no less respected.

No space is devoted to theory, except in occasional instances where it appears necessary for a clear understanding of the circuit under discussion. For a general treatment of the subject of tran-

sistors, the reader is referred to the author's book *Transistors, Theory and Practice* from the same publisher.

Several of the circuits have been taken from articles which have appeared previously under the author's name in various magazines. These are listed as references at the end of each chapter so that the reader might explore the subject more extensively. For permission to use this material, the author is grateful to the editors of *Audio-craft*, *Electronics*, *Popular Electronics*, *RADIO-ELECTRONICS Magazine*, *Radio & Television News* and *Tele-Tech & Electronic Industries*.

RUFUS P. TURNER
Los Angeles, Calif.

PRELIMINARY NOTES

1. All resistances are specified in ohms, all capacitances in microfarads, unless labeled otherwise.
2. All resistors are $\frac{1}{2}$ -watt rating, unless labeled otherwise.
3. P-n-p transistors are shown in most of the circuits. N-p-n transistors having the same characteristics may be substituted, provided the battery connections are reversed as well as the connections of dc meters, electrolytic capacitors and photocells.
4. Where a ground (chassis) connection is shown with dotted lines, a ground might improve the circuit stability but cannot always be guaranteed to do so. The reader therefore should test the circuit with and without the dotted-line grounds. Results vary with different layouts and workmanship.
5. Use the exact component values specified. When the circuit is to be modified to suit individual requirements, wire it first according to the book and verify its operation against the author's representations. Then, undertake the changes.

audio amplifiers

THE first widespread practical application of transistors was in audio amplifiers. The reason for this is easy to understand when it is remembered that the early transistor essentially was a low-frequency device. Its small size, low dc voltage and current requirements and high efficiency suit the transistor naturally to use in hearing aids, which are small-sized audio amplifiers.

Since the early developmental period, great progress has been made in the design and manufacture of transistors and allied circuit components and in amplifier circuit design. Numerous circuits have been offered to the engineer, technician and hobbyist.

Representative audio-amplifier circuits are described in this chapter.

Single-stage, R-C-coupled, common-base amplifier

Fig. 101 shows the circuit of a single-stage resistance-capacitance-coupled amplifier employing the common-base configuration. This circuit sometimes is called "grounded base."

This amplifier provides a voltage gain of 30 when operated into a high-impedance load. Power output is 1.8 milliwatts. Its input impedance is 130 ohms and output impedance 5,000 ohms. The maximum input signal voltage which may be applied before output-voltage peak clipping appears is 0.1 volt rms. The corresponding maximum output signal voltage is 3 volts rms. Fig. 102 shows the frequency response.

Two batteries are required in the common-base amplifier, B1 (1.5 volts) for the emitter bias and B2 (6 volts) for collector bias.

Battery B2 supplies approximately 0.8 milliamperes dc and battery B1, 0.85 ma. The dpst switch (S1-S2) makes and breaks connections to both batteries simultaneously.

An alternative, single-battery circuit is shown in Fig. 103. Here, a voltage divider R3-R4 is operated with a single 7.5-volt battery. Current flowing through the divider develops the emitter voltage as a drop across R3 and collector voltage across R4. The bleeder current is 9.62 ma.

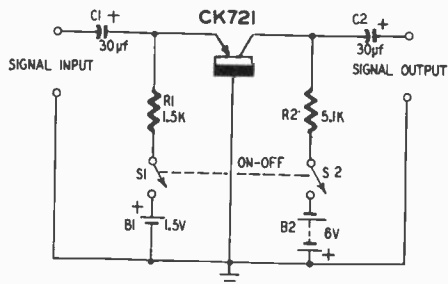


Fig. 101. Single-stage, R-C-coupled, common-base amplifier.

All resistors in Figs. 101 and 103 are $\frac{1}{2}$ watt. Capacitors C1 and C2 may be miniature, low-voltage tantalum electrolytics if subminiaturization is desired.

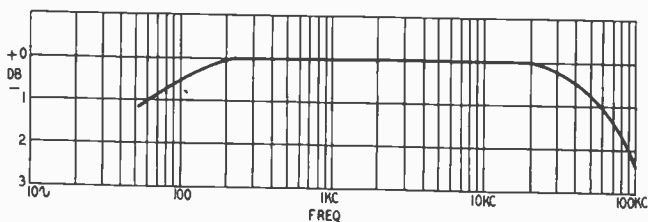


Fig. 102. Frequency response of single stage, R-C-coupled, common-base amplifier.

Single-stage, transformer-coupled, common-base amplifier

In any transistorized amplifier, the highest per-stage power gain is obtained only with transformer coupling between stages or between input and output.

Fig. 104 shows the circuit of a typical common-base, transformer-coupled, single-stage amplifier employing a General Electric 2N45 transistor. This amplifier has been designed for 50,000 ohms input impedance and 500 ohms output impedance. Power output is 2 milliwatts and power gain is 500 times, or 27 db. This means that an input-signal driving power of 4 microwatts will give full output.

Miniature transformers are employed for input (T1) and output

(T2) coupling. If these transformers are mounted close together, they must be oriented in such a way that their cores are at right angles to prevent feedback.

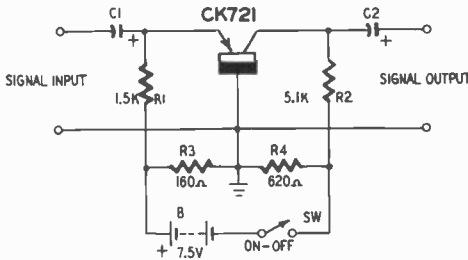


Fig. 103. Alternative arrangement using single battery and voltage divider.

Bypass capacitors C1 and C2 may be standard 25-volt electrolytics or miniature tantalum electrolytics. Variable resistor R is a 2,000-ohm miniature wirewound rheostat. Initially, this control is set for a dc collector current of 1 milliampere.

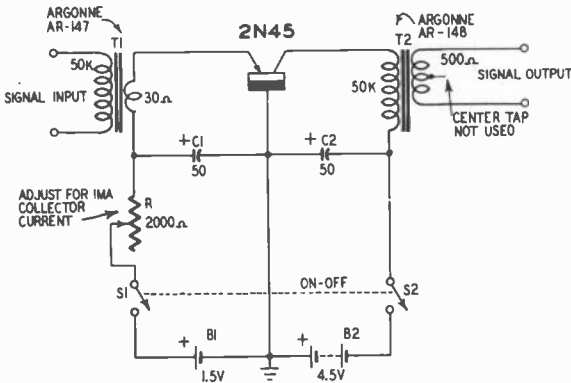


Fig. 104. Single-stage, transformer-coupled, common-base amplifier.

Single-stage, R-C-coupled, common-emitter amplifier

The common-emitter circuit, also called grounded-emitter, provides high voltage and high power gain and is adapted readily to single-battery operation.

Fig. 105 gives the circuit of a single-stage, R-C-coupled, common-emitter amplifier employing a Sylvania 2N34 transistor. Fig. 106 shows its frequency response. The input impedance, measured at 1,000 cycles, is 780 ohms; the output impedance 10,000 ohms. The voltage gain is 80 when the amplifier is operated into a high-im-

pedance load. (Higher values of gain may be obtained with individual transistors.) Maximum input-signal voltage before output-voltage peak clipping is 20 millivolts rms. Corresponding maximum output-signal voltage is 1.7 volts rms. Total current drain is 500 microamperes dc.

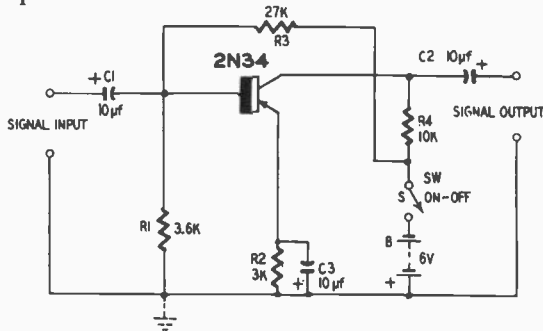


Fig. 105. Single-stage, R-C-coupled, common-emitter amplifier.

The transistor operating point is stabilized by the steady value of base bias voltage supplied by voltage divider R1-R3 and emitter resistor R2. The latter is bypassed by capacitor C3 to prevent degeneration.

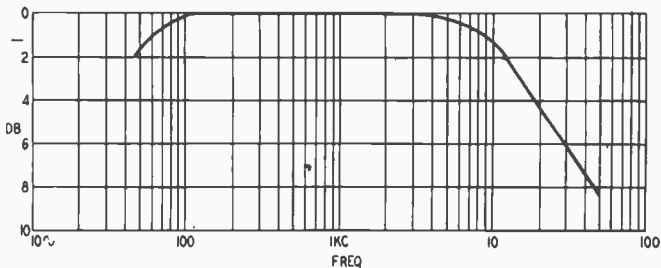


Fig. 106. Frequency response of circuit shown in Fig. 105.

The electrolytic coupling capacitors C1 and C2 and bypass capacitor C3 may be standard-size 25-volt components or miniature, tantalum electrolytics.

Single-stage, transformer-coupled, common-emitter amplifier

Transformers are employed for input and output coupling in the common-emitter circuit shown in Fig. 107. This allows very nearly the full 39-db power gain of the transistor (Raytheon CK722) to be obtained.

Miniature transformers are used. While the 200-ohm secondary of T1 and the 15,000-ohm primary of T2 do not match the transistor

input and output impedances exactly, the match is close enough for good power transfer. The secondary of T2 may be connected directly to the base-input circuit of a similar amplifier stage.

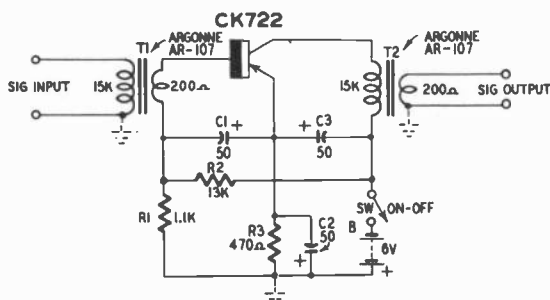


Fig. 107. Single-stage, transformer-coupled, common-emitter amplifier.

Dc base bias is supplied by the R1-R2 voltage divider. Further stabilization of the operating point is provided by emitter resistor R3 which is bypassed by C2 to prevent degeneration.

Single-stage, R-C-coupled, common-collector amplifier¹

The common-collector circuit (also known as the grounded collector) has the highest input impedance of the three transistor amplifier configurations. Its operation and characteristics resemble somewhat those of the vacuum-tube cathode follower and, for this reason, the common collector often is referred to as an "emitter follower."

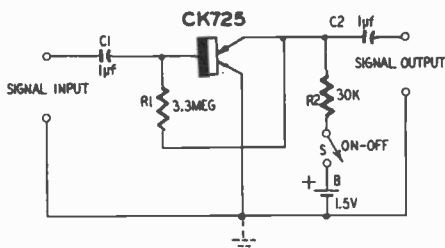


Fig. 108. Single-stage, R-C-coupled, common-collector amplifier.

Fig. 108 shows a single-stage, resistance-capacitance-coupled, common-collector amplifier employing a Raytheon CK725 transistor. At 1,000 cycles, the input impedance of this amplifier is 1 megohm. Output impedance is 30,000 ohms. The input impedance varies with the signal frequency, being 1.2 megohms at 20 cycles (Fig. 109-a) and 160,000 ohms at 50 kc.

The voltage gain of the amplifier is constant at 0.96 from 20

cycles to 10 kc (Fig. 109-b) and falls slowly to 0.88 at 50 kc. The maximum input-signal voltage before positive-peak clipping ap-

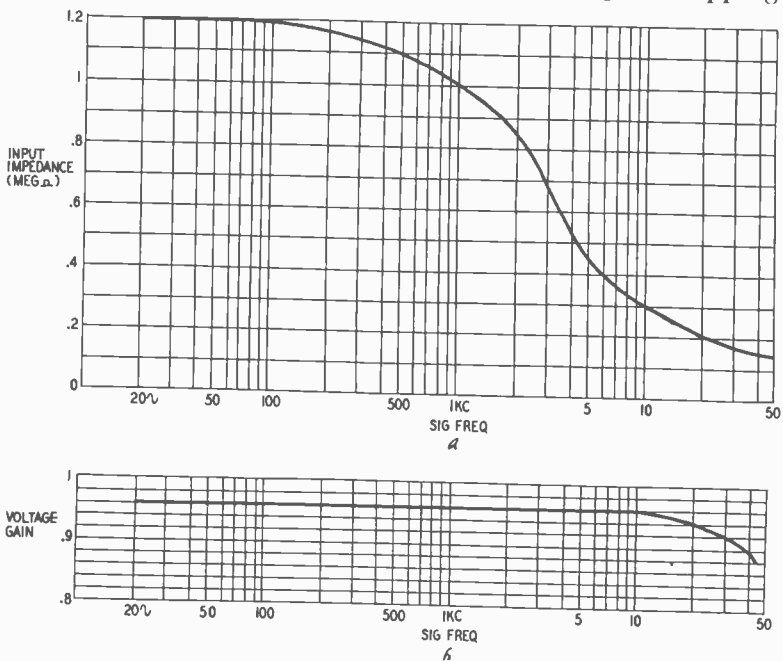


Fig. 109-a, -b. Characteristics of common-collector amplifier: (a) input impedance variation; (b) frequency response.

pears in the output signal is 0.52 volt rms. The corresponding maximum output-signal voltage is 0.499 volt rms when the external load is 300,000 ohms or higher. Power gain is 30.74, corresponding to 14.87 db.

Single-stage, transformer-coupled, common-collector amplifier

The common-collector amplifier is convenient for coupling into a low-impedance line. However, Z_i varies with the output impedance (Z_o), with the result that Z_i drops to approximately 20,000 ohms when $Z_o = 500$ ohms.

The most satisfactory operating conditions therefore are obtained when the common-collector output is transformer-coupled to the lower-impedance line. Fig. 110 shows a common-collector amplifier with transformer output. Here, the input impedance follows the curve given in Fig. 109-a and the output impedance is constant at 100 ohms.

The common-collector amplifier may be transformer-coupled to

the base-input circuit of a common-emitter amplifier by employing a transformer having a 1,000-ohm secondary.

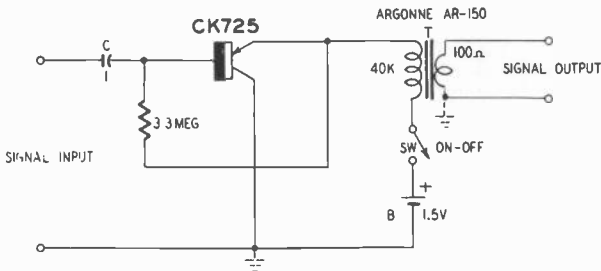


Fig. 110. Single-stage, transformer-coupled, common-collector amplifier.

Multistage R-C-coupled amplifier

Fig. 111 shows the circuit of a four-stage, resistance-capacitance-coupled amplifier employing Sylvania 2N34 transistors. Each stage utilizes the common-emitter circuit.

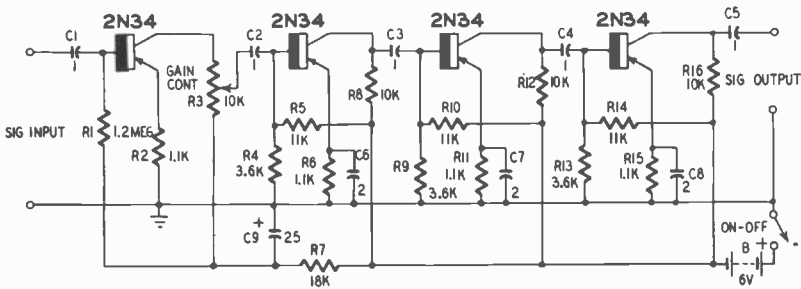


Fig. 111. Multistage R-C-coupled amplifier.

With the GAIN CONTROL set at maximum, this amplifier provides a voltage gain of 4,000. The maximum input-signal voltage before output-voltage peak clipping is 0.2 millivolt rms. The corresponding maximum output-signal voltage is 1.2 volts rms. At 1,000 cycles, the input impedance is approximately 1,000 ohms. The noise level was measured as 5 millivolts, with the input terminals of the amplifier short-circuited (72 db below maximum output voltage). Fig. 112 shows the frequency response.

Stabilizing bias is applied to the bases of the transistors by voltage-divider networks R4–R5, R9–R10 and R13–R14. To prevent degeneration, emitter resistors R6, R11 and R15 are bypassed by capacitors C6, C7 and C8. Emitter resistor R2 in the input stage is left unbypassed for a small amount of degeneration in this stage.

Capacitor C9 and resistor R7 form a decoupling network to suppress motorboating.

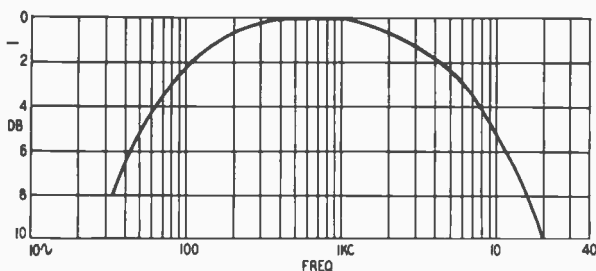


Fig. 112. Frequency response of the circuit shown in Fig. 111.

Total current drain is approximately 8 milliamperes dc from the 6-volt battery.

Multistage transformer-coupled amplifier

A three-stage transformer-coupled amplifier employing Raytheon CK722 transistors is shown in Fig. 113. This circuit provides an overall power gain of 80 db (with GAIN CONTROL R5 set to maximum) and a power output of 6 milliwatts. The input impedance is approximately 1,000 ohms and the output impedance 1,200 ohms.

The dc base bias is stabilized in each stage by means of voltage dividers R1-R2, R6-R7 and R9-R10. Emitter current-limiting resistors R3, R8 and R11 are bypassed adequately by C2, C5 and C7 to prevent degeneration. Capacitor C3 and resistor R4 form a decoupling network to suppress motorboating. The total current drain from the 3-volt battery is 7.3 milliamperes dc.

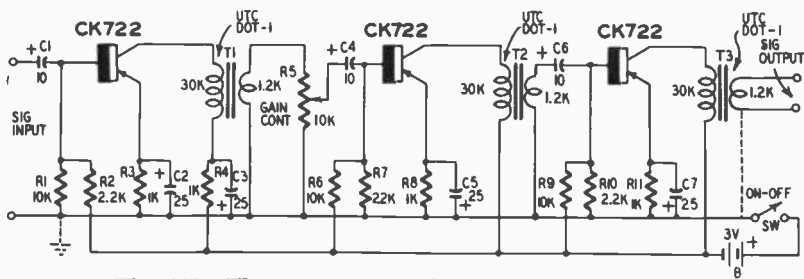


Fig. 113. Three-stage transformer-coupled amplifier.

In the construction of this amplifier, the miniature transformers T1, T2 and T3 must be mounted far enough apart that their magnetic fields do not interact. When a compact layout necessitates close spacing, the transformers must be oriented in such a way that their cores are at right angles.

Push-pull output circuits

As in vacuum-tube practice, push-pull output amplifier stages and drivers are employed when more signal power is desired than can be supplied by a single-ended transistor stage. Fig. 114 shows typical class-A and class-B push-pull output stages.

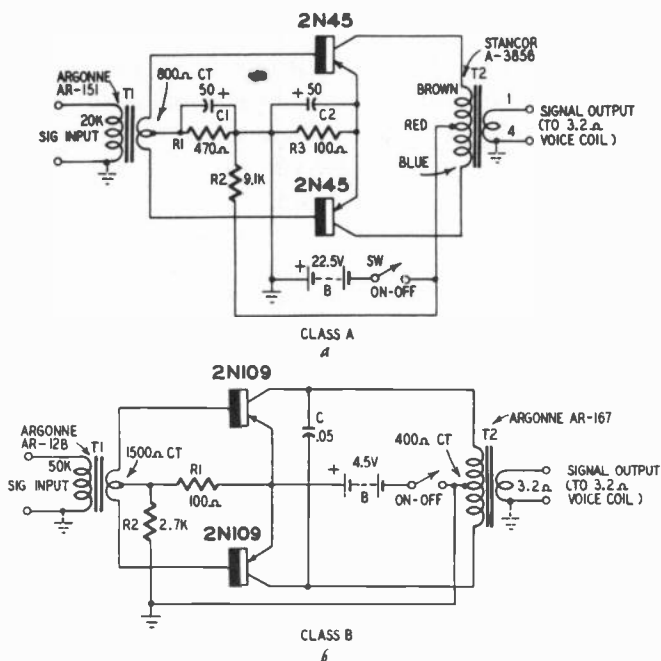


Fig. 114-a, -b. Push-pull amplifiers: (a) class A; (b) class B.

Class A

(Fig. 114-a). Here, two General Electric 2N45 transistors are employed. The input impedance is 20,000 ohms. (The 20,000-ohm primary of the input transformer T1 can act as the collector load of a single-ended transistor input stage.) The output impedance matches the 3.2-ohm voice coil of a loudspeaker. At 5% total harmonic distortion (measured at 1,000 cycles), the power output is 75 milliwatts and the overall power gain 30 db, or 1,000 times.

The total continuous current drain is 12 milliamperes dc from the 22.5-volt battery. The collector-circuit efficiency is 42%. Stabilizing dc bias is supplied to the bases of the transistors by the voltage divider R1-R2. The emitter current-limiting resistor R3 is bypassed by C2 to prevent degeneration.

Class B

As in tube practice, class-B transistor output amplifiers are preferred to class-A units because of their higher collector-circuit efficiency. In the class-A circuit (Fig. 114-a), a rather large collector current flows continuously, while in class-B circuits large values flow only on audio peaks and at other times the "resting" collector current is negligible.

Fig. 114-b shows a typical class-B circuit employing two RCA 2N109 transistors. The input impedance of this amplifier is 50,000 ohms and the output impedance 3.2 ohms to match the voice coil of a loudspeaker. At approximately 8% total harmonic distortion (measured at 1,000 cycles), the power output is 75 milliwatts and overall power gain is 30 db, or 1,000 times. The collector-circuit efficiency is 64%.

The zero-signal total collector current is approximately 4 ma dc, and the maximum-signal total collector current approximately 26 milliamperes. The base-bias voltage divider R1-R2 draws approximately 1.6 milliamperes.

Phase inverters

The interesting properties of transistors permit the design of ingenious phase inverter circuits. Fig. 115 shows two types.

Single-transistor type

(Fig. 115-a). This is a simple circuit employing one Raytheon CK721 transistor and is similar to the so-called "hot-cathode" phase inverter sometimes used with triode vacuum tubes. A portion of the output-signal voltage (OUTPUT 1) is developed across the collector load resistor R4 and a second portion (OUTPUT 2) across the emitter resistor R3. Due to phase shift, which is inherent in the common-emitter circuit, OUTPUT 1 is out of phase with the input-signal voltage while OUTPUT 2 is in phase with the input. The two output-signal voltages accordingly are out of phase with each other.

A 20-millivolt rms input-signal voltage will give output voltages of approximately 0.8 volt when the phase inverter feeds into high-impedance loads. Actually, these two output voltages are not exactly equal because the circuit is not perfectly balanced. However, they may be made very nearly equal by adjustment of R3 either above or below the specified 4,700-ohm value, as required.

Complementary-symmetry type

The circuit shown in Fig. 115-b makes use of the principle of com-

plementary symmetry which is evidenced when n-p-n and p-n-p transistors with identical characteristics (other than polarity) are operated in the same circuit. The n-p-n unit here is a Sylvania 2N35 and the p-n-p a Sylvania 2N34.

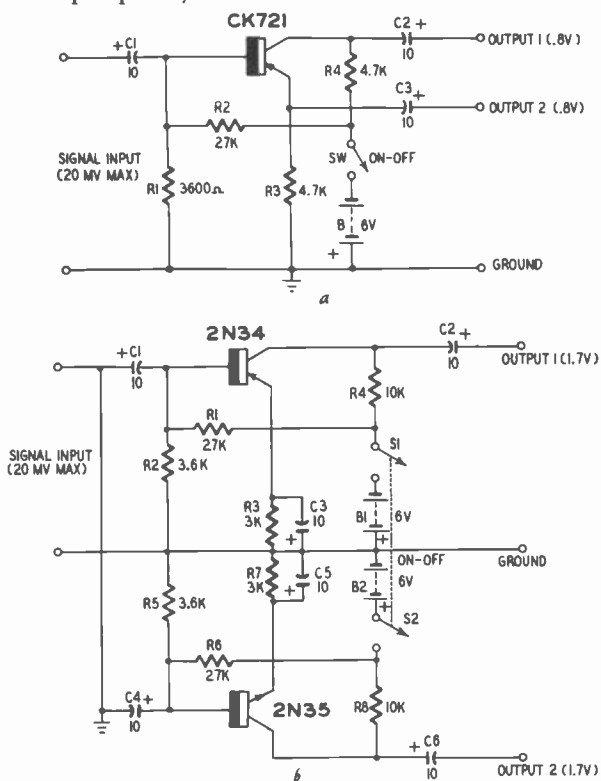


Fig. 115-a, -b. Phase inverters: (a) single-transistor type; (b) complementary-symmetry type.

Notice that the circuit above the ground line is identical with that below the line except for the transistor type and the polarities of batteries and capacitors. Since the input circuits of the two halves are connected in parallel, the input signal is applied in the same phase to each transistor. The two output signals (OUTPUT 1 and OUTPUT 2) are out of phase with each other.

When a signal is applied with the same polarity to an n-p-n and p-n-p transistor simultaneously, the collector currents shift in opposite directions. Thus in Fig. 115-b, when the upper SIGNAL INPUT terminal is made positive with respect to the lower terminal, the 2N34 collector current decreases while the 2N35 collector current increases. The 2N34 collector therefore becomes more negative,

while the 2N35 collector becomes more positive. OUTPUT 1 accordingly swings negative while OUTPUT 2 swings positive. These conditions reverse when the upper INPUT SIGNAL terminal is made negative.

The maximum value of input-signal voltage before peak clipping in the output signal is 20 millivolts rms. The corresponding maximum output-signal voltage is 1.7 volts rms each for OUTPUT 1 and OUTPUT 2.

The two output-signal voltages may be equalized by matching the transistors, circuit components, and battery voltages closely. Base-bias stabilization is provided by voltage dividers R1-R2 and R5-R6.

Class-B loudspeaker amplifier with conventional transistors

Useful loudspeaker operation may be obtained with transistorized power amplifiers. Fig 116 shows a two-stage amplifier circuit employing small, conventional transistors.

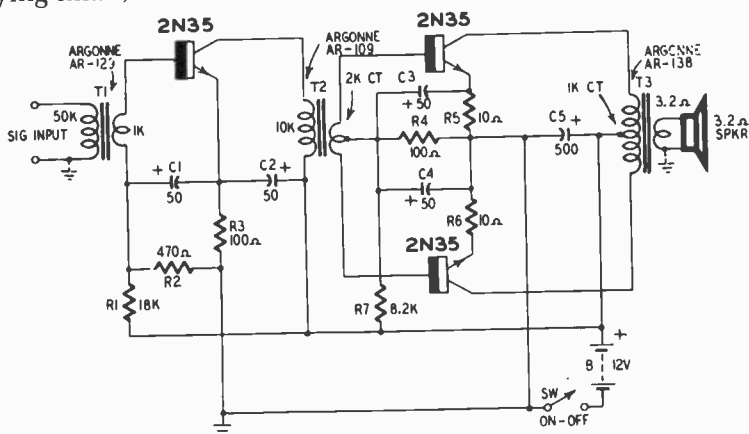


Fig. 116. Class-B loudspeaker amplifier with conventional transistors.

N-p-n transistors (Sylvania 2N35) are used both in the single-ended class-A driver and push-pull class-B output stages. Operated from a 12-volt battery, this amplifier delivers 100 milliwatts of audio power to the loudspeaker. Input power is 0.25 microwatt. This represents an overall power gain of 56 db, or 400,000 times. Full audio output is obtained with an input signal voltage of 112 millivolts rms. Total harmonic distortion at 1,000 cycles is 5%.

Input impedance of the amplifier is 50,000 ohms. This permits the unit to be driven by high-impedance microphones, phonograph pickups, or transistorized preamplifiers.

Total zero-signal current drain is 10 ma dc from the 12-volt battery. Total maximum-signal current is 23 ma. Stabilizing base bias is supplied by the voltage-divider networks R1-R2 and R4-R7. Divider resistors R1 and R7 may require adjustment for optimum performance, especially for the elimination of crossover distortion.

Class-B loudspeaker amplifier with power transistors ²

When the desired amount of audio power exceeds the capability of conventional transistors, power transistors are required.

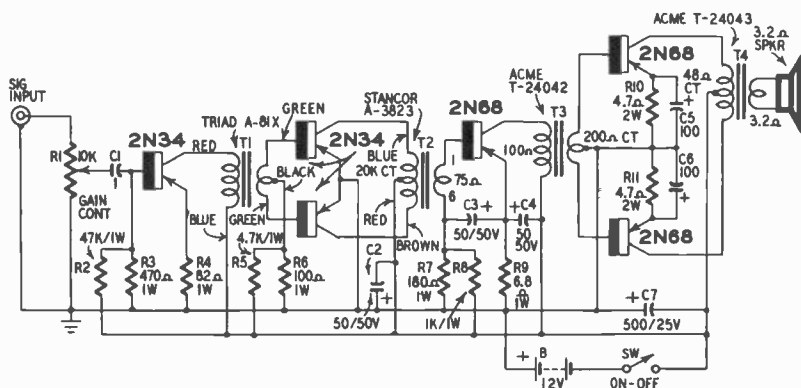


Fig. 117. Class-B loudspeaker amplifier with power transistors.

Fig. 117 shows the circuit of a complete amplifier employing conventional transistors in the low-level stages and power transistors in the driver and class-B output stages. With the GAIN CONTROL R1 set to maximum, this amplifier delivers an output power of 5 watts to the voice coil of a 3.2-ohm loudspeaker. For full audio output power, the input-signal voltage is 1 millivolt rms. Overall power gain is 68 db. Noise level is 6 mv (56 db below maximum output voltage) and total harmonic distortion at 1,000 cycles is 11%. Fig. 118 shows the frequency response.

Common emitters are used throughout. The first stage employs a Sylvania 2N34 as a class-A amplifier. The second stage employs push-pull 2N34's in class A. The third stage is the class-A driver with a Sylvania 2N68 power transistor. The output stage employs two 2N68's operated in push-pull class B.

The 2N68 driver draws a steady dc collector current of 150 ma. The total zero-signal collector current of the push-pull output stage is 1.5 milliamperes dc and the total maximum-signal collector current 550 ma dc.

To dissipate the heat generated during operation, all three power

transistors must be bolted solidly to the metal chassis of the amplifier. However, they must be insulated from the chassis electrically, since their cooling structures are at collector potential. This is accomplished by using a 1.5-mil-thick mica washer between the transistor case and the chassis.

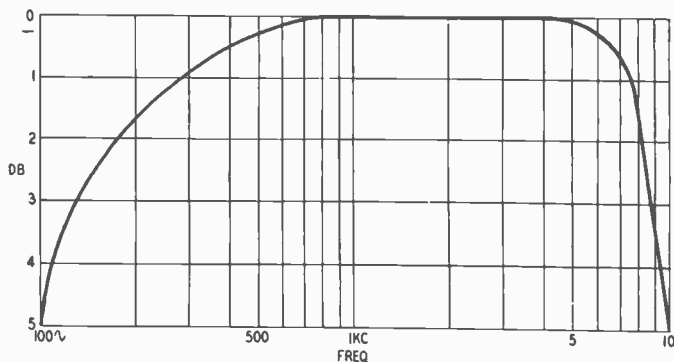


Fig. 118. Frequency response of Class-B loudspeaker amplifier with power transistors.

Class-A loudspeaker amplifier with power transistor

A power transistor operated as a class-A amplifier does not provide as high output as when a pair are operated class B, but the class-A power type gives considerably more power than conventional transistors even when the latter are operated class B.

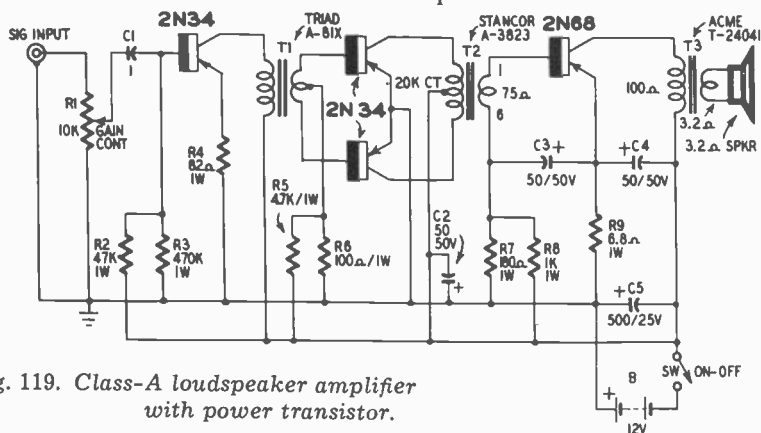


Fig. 119. Class-A loudspeaker amplifier with power transistor.

Fig. 119 shows the circuit of an amplifier in which the output stage contains a single class-A-operated Sylvania 2N68 power transistor. The audio output of this amplifier is 0.6 watt. The two low-level stages employ Sylvania 2N34 transistors (single-ended in the input stage and push-pull in the intermediate stage).

Full output power is obtained with a 1-millivolt rms input signal. Total continuous current drain is 200 milliamperes dc from the 12-volt supply. Dc base-bias stabilization is provided by voltage-divider networks R2-R3, R5-R6 and R7-R8. Emitter resistor R4 in the input stage is left unbypassed for degeneration.

Complementary-symmetry power amplifier with conventional transistors

The principle of complementary symmetry, explained in connection with the n-p-n-p-n-p phase inverter (Fig. 115-b and its accompanying section) may be utilized to obtain a novel single-ended push-pull amplifier.

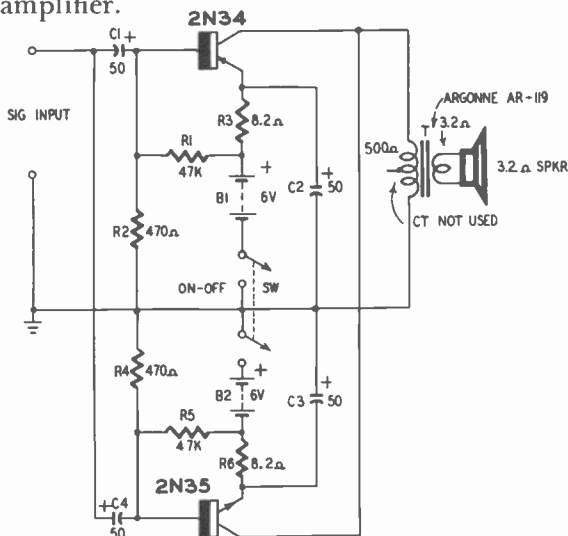


Fig. 120. Complementary-symmetry power amplifier with conventional transistors.

Fig. 120 shows the circuit of a complementary-symmetry class-A amplifier employing conventional small-sized transistors. This unit has an audio power output of 100 milliwatts. Its power gain is 40 db. For full output, the input-signal driving power is 0.01 mw. Input impedance is approximately 400 ohms.

This circuit employs a Sylvania 2N34 transistor as the p-n-p unit and a Sylvania 2N35 as the n-p-n. Notice that the halves of the circuit above and below the ground line are images of each other, except for the differences in transistor type and of the polarities of batteries and capacitors.

If the transistors, battery voltages and resistors are matched carefully, the circuit will be balanced and no dc can flow through the

nect this voice coil directly into the circuit without an output transformer, as shown, since the circuit is balanced and no dc can flow through the speaker if the transistors, battery voltages and resistors are matched carefully. If the transistors are nearly, but not exactly, matched some degree of matching may be secured by varying either R2 or R4 to balance the circuit, like a bridge, and remove any dc from the voice coil.

Single-stage low-noise preamplifier ³

The single-stage preamplifier circuit shown in Fig. 122 employs a Raytheon CK727 transistor, a low-noise type. It is particularly important to have a quiet preamplifier when the main amplifier following this unit has considerable gain.

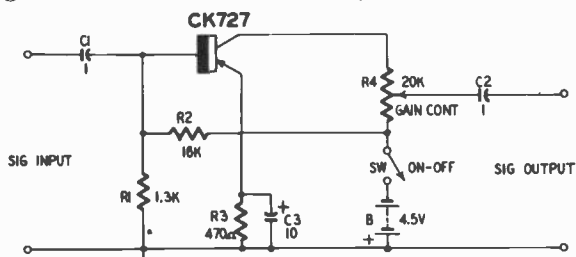


Fig. 122. Single-stage low-noise preamplifier.

Having an input impedance of approximately 1,000 ohms, this circuit provides a voltage gain of 93 with the GAIN CONTROL R4 set to maximum. The maximum input-signal voltage before output-signal peak clipping is 10 millivolts rms. The corresponding maximum output-signal voltage, across a high-impedance load, is 0.93 volt rms. Fig. 123 shows the frequency response.

The noise voltage, measured with an ac vacuum-tube millivoltmeter at the amplifier output terminals, was 3 millivolts with the amplifier input terminals short-circuited. This is 49.8 db below maximum output signal voltage.

The total drain from the 4.5-volt battery is 360 microamperes dc.

Preamplifier with high-impedance input ⁴

Through the use of a common-collector input stage, the preamplifier circuit shown in Fig. 124 achieves high input impedance. This is an invaluable property when the amplifier must impose negligible loading upon the signal source from which it is operated. The final stage also employs the common-collector circuit to provide the low-impedance output of 500 ohms while maintaining a fairly high input impedance.

Like vacuum-tube cathode followers, common-collector transistor amplifiers have voltage gains of less than 1. The gain of this preamplifier therefore is supplied by an intermediate common-emitter

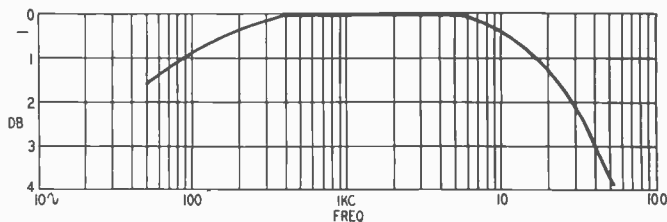


Fig. 123. Frequency response of the low-noise preamplifier.

stage. Although the common-emitter and common-collector circuits require dc bias voltages of opposite polarity, use of a single battery has been made possible in this preamplifier circuit by op-

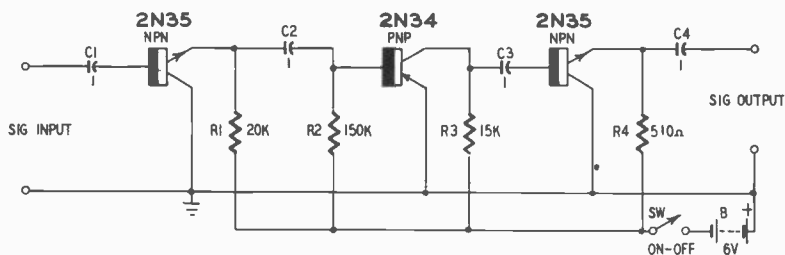


Fig. 124. Preamplifier with high-impedance input.

erating an n-p-n transistor (Sylvania 2N35) in each of the common-collector stages and a corresponding p-n-p transistor (Sylvania 2N34) in the common-emitter stage.

At 1,000 cycles, the measured input impedance of the preamplifier is 0.2 megohm. Higher values, up to 1 megohm or better, may be obtained through a special selection of the input transistor. Maximum input-signal voltage before output-voltage peak clipping is 4 millivolts rms. Corresponding maximum output voltage is 0.2 volt rms. Voltage gain is 50 times, or 34 db. Power is 56 db. Total harmonic distortion at 1,000 cycles is 1.2%. Fig. 125 shows the frequency response.

Total drain from the 6-volt battery is 600 microamperes dc. This low power drain of the preamplifier, together with the small number of components and the absence of transformers, permits miniature construction (e. g., mounting in a probe handle) and operation from penlight batteries or photoelectric cells.

If a gain control is required, a 15,000-ohm potentiometer may be substituted for the 15,000-ohm load resistor R3.

No bias-stabilization networks have been provided in this pre-amplifier. These resistors were omitted to reduce the number of components, minimize the dc drain and keep the input impedance

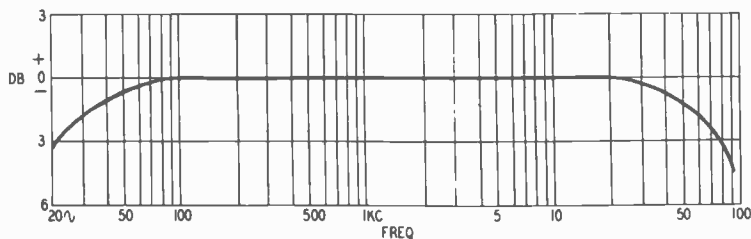


Fig. 125. Frequency response of preamplifier.

of each stage as high as possible. (The input and output stages have “floating” bases for highest input impedance). The lack of stabilization limits the amplifier to use in reasonably constant temperature environments and to low signal-input voltage levels (not exceeding the 4-mv value mentioned earlier).

Boosting transistor input impedance

Experimenters find it hard to make transistors work “like tubes” in some circuits because the input impedance of the transistor inherently is low while that of the tube is very high. The input impedance of a transistor amplifier might be 1,000 ohms or less while a tube amplifier may exhibit several megohms.

In a transistor amplifier, this difficulty may be overcome to some extent by using a stepdown input transformer. This gives the amplifier a much higher input impedance than it could offer alone. But the use of a transformer is not desirable in every case, simple resistance-capacitance coupling sometimes being preferred even to a subminiature transformer.

Three schemes for boosting transistor input impedance without a transformer immediately come to mind. One of these is to add a series resistor to the transistor input circuit, the second is to employ emitter degeneration (obtained by means of an unbypassed resistor in series with emitter and ground in a common-emitter circuit) and the third is the common-collector type of amplifier circuit which normally has many times the input impedance of the more widely-used common-emitter circuit. Each of these schemes has its place, and it is interesting to compare these circuits with the conventional common-emitter.

Fig. 126-a shows the common-emitter circuit used for comparison on a basis of experimental operation. The measured input impe-

dance is 1,500 ohms and the voltage gain is 100. The value of the collector resistor R2 was chosen experimentally, with the individual transistor, for maximum voltage gain.

Series-resistance method

The first step was to add series input resistance to the original common-emitter circuit. Thus, R_s was inserted as shown in Fig. 126-b. The presence of R_s reduces the amplification of the stage. It was found that when $R_s = 1,500$ ohms, the input resistance of the

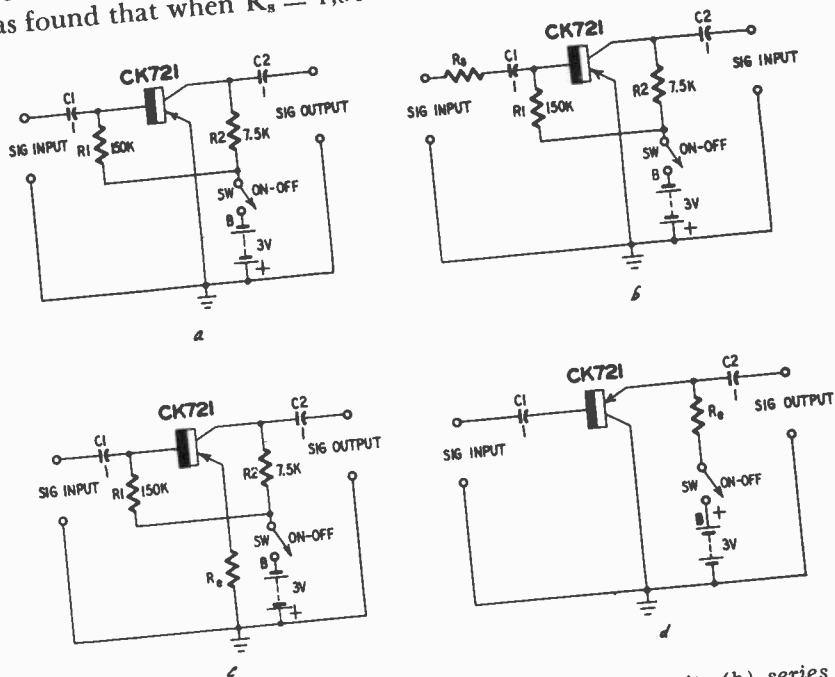


Fig. 126. (a to d). (a) original common-emitter circuit; (b) series resistance; (c) emitter degeneration; (d) common collector.

stage is 3,000 ohms and the voltage gain drops to 50, just one-half of its original value. When R_s is increased to the point at which the gain is reduced to 1, similar to the ideal gain of a cathode follower; $R_s = 90,000$ ohms, the input resistance of the stage is 91,500 ohms and the voltage gain is 1. Obviously, there is no voltage amplification in this instance; but there is a power amplification of 25, since the output impedance is less than the input impedance. Thus, various values of added resistance R_s between 1,500 and 90,000 ohms boost the original input resistance (impedance) by 2 to 61 times, but at the same time they reduce the voltage gain to a value from 0.5 to 0.01 of its original level.

Emitter degeneration method

In the common-emitter circuit, an unbypassed resistor inserted between emitter and ground will provide degeneration which raises the input impedance of the circuit. An interesting fact to note is that the resulting input impedance is *higher* than the value of the added resistor, and the voltage gain of the basic circuit is not lost nearly as fast as when using an ordinary series input resistor. Fig. 126-c shows connection of the emitter series resistor R_e into the original circuit.

Table 1 shows how the input resistance R_{in} and voltage gain vary with the value of the emitter resistance R_e . These values will differ somewhat with individual transistors.

Table 1—Input and Emitter Resistance vs. Voltage Gain

R_e (Ohms)	R_{in} (Ohms)	Voltage Gain
500	10,000	80
1,000	23,000	35
2,000	50,000	15
3,500	100,000	1

Common-collector circuit

In the common-collector amplifier circuit, the collector is the grounded or common electrode of the transistor and the emitter becomes the output electrode. The common collector normally has the highest input impedance of the three basic transistor circuits. It has somewhat less power gain than the common emitter and common base. Fig. 126-d shows the common-collector circuit tested.

Emitter resistor R_e was varied for maximum voltage gain, which was found to be 0.90. At this point, $R_e = 20,000$ ohms and the input resistance (impedance) is 0.5 megohm.

Analysis

A number of conclusions can be drawn from these simple results. The alert experimenter may find several applications for each of the three circuits in which the input impedance has been boosted. The analysis runs somewhat as follows.

1. If there is gain to spare and the input impedance need be raised only a small amount, use Fig. 126-b. Resistor R_a may be inserted on either side of capacitor C1.
2. If the voltage gain must be kept as high as practicable and still a relatively large increase in input impedance obtained, use Fig. 126-c.

3. If the highest possible input impedance is required and a voltage gain of less than 1 can be tolerated, then use the emitter follower Fig. 126-d.

4. With regard to frequency response, Fig. 126-d has the best (excellent), Fig. 126-c next best (good) and Fig. 126-b the worst (fair to poor, depending upon how high a value of R_s is employed).

REFERENCES

¹ Rufus P. Turner, *Common-Collector Transistor Amplifier*. RADIO-ELECTRONICS Magazine, September, 1955; p. 37.

² Rufus P. Turner, *5-Watt Transistorized Audio Amplifier*. Radio & Television News, April, 1956; p. 53.

³ Rufus P. Turner, *Low-Noise Transistor Preamplifier*. RADIO-ELECTRONICS Magazine, August, 1954; p. 64.

⁴ Rufus P. Turner, *Transformerless Line Output Preamplifier*. Electronics, August, 1956; p. 198.

rf and if amplifiers

COMMERCIALLY AVAILABLE high-frequency transistors extend the operating range of tubeless equipment into the radio-frequency spectrum. Transistorized radio-frequency and intermediate-frequency amplifiers now are entirely practical for applications in receivers, instruments and control devices.

The low dc power requirements of transistorized high-frequency amplifiers, as in other transistorized circuits, suit them to economical battery operation. This is invaluable when such amplifiers are to be used in conjunction with other equipment but must not draw power from the equipment.

Broadcast-band rf amplifier and preselector

Fig. 201 shows the circuit of a tuned radio-frequency amplifier suitable for use as a preselector ahead of a broadcast receiver to increase the selectivity of the latter.

Employing a Sylvania 2N94A n-p-n-type rf transistor, this amplifier may be tuned continuously from 500 to 1,700 kc and provides a power gain of 20 db, or 100 times. Input and output impedances are approximately 500 ohms. Maximum input-signal voltage is 300 millivolts. In the middle of the standard broadcast band (1,000 kc) the amplifier bandwidth is 20 kc at the 3-db points (70.7% of voltage at peak of response curve).

The coils are wound with No. 32 enameled wire on 1-inch-diameter forms according to the instructions given in Table 2. Each primary and secondary coil set (L1-L2 and L3-L4) must be enclosed in a grounded metal shield can to prevent oscillation.

While the two sections (C1 and C2) of the tuning capacitor are ganged together for simultaneous adjustment, the rotors cannot be connected together electrically. This prohibits use of the standard two-gang unit. It is best to gang together two separate 365- μmf variable capacitors by an insulated shaft coupling or belt drive and to insulate the entire assembly from the metal chassis. An insulating shaft coupling also must be provided between the capacitor and tuning dial to prevent detuning effects due to body capacitance.

Total current drain of the amplifier is approximately 10 milliamperes dc from the 6-volt battery.

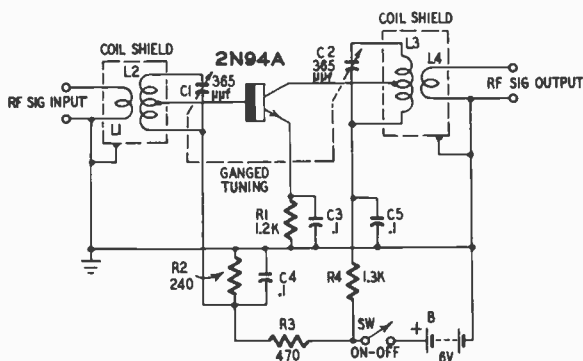


Fig. 201. Broadcast-band rf amplifier and preselector.

Table 2—Coil Table for RF Amplifier

- L1—10 turns No. 32 enameled wire closewound around “ground end” of L2 and insulated from it.
- L2—130 turns No. 32 enameled wire closewound on 1-inch-diameter form. Tap 15th turn from “ground” end.
- L3—130 turns No. 32 enameled wire closewound on 1-inch-diameter form. Tap 80th turn from “ground” end.
- L4—12 turns No. 32 enameled wire closewound around the ground end of L3 and insulated from it.

Single-stage, single-tuned 455-kc if amplifier

A simple 455-kc intermediate-frequency amplifier circuit suitable for inclusion in a small radio receiver or for general-purpose experimental applications is shown in Fig. 202. Employing a Raytheon 2N112/CK760 rf transistor, this amplifier provides a power gain of 30 db. Bandpass is 10 kc at the 6 db points.

Input and output impedances of the amplifier both are 600 ohms. These values are convenient for operation of the unit between a low-

impedance converter, such as a crystal diode, and a succeeding transistor if amplifier or second detector stage or crystal-diode second detector.

Because of the inherent regenerative nature of the common-emit-

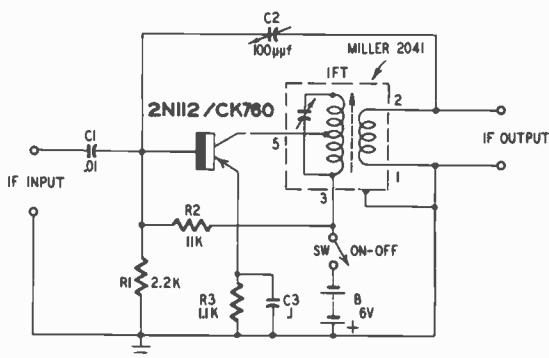


Fig. 202. Single-tuned one-stage 455-kc if amplifier.

ter circuit, oscillation will occur unless this circuit is neutralized. Variable capacitor C2 is provided for this purpose. The circuit is neutralized when C2 is set approximately to 80 μf , but the capacitance setting will depend upon the actual transistor internal capacitances and upon the stray circuit capacitances due to the layout.

The miniature, transistor type if transformer has an input (collector) impedance of 25,000 ohms and an output impedance of 600 ohms. The figures shown on the transformer leads correspond to the manufacturer's coding and must be followed in the wiring to obtain correct amplification and neutralization.

Total current drain from the 6-volt battery is approximately 1.5 milliamperes dc.

The amplifier is aligned initially in the conventional manner: (1) Connect an amplitude-modulated signal generator to the IF INPUT terminals. (2) Connect a detector (ac vacuum-tube voltmeter or high-impedance magnetic headphones in series with a 1N34 germanium diode) to the IF OUTPUT terminals. (3) Close switch SW. (4) Set the signal generator to 455 kc. (5) Tune the slug of the if transformer for maximum signal in the detector. Reduce the output of the signal generator if the amplifier or detector blocks.

Two-stage 455-kc if amplifier

In a majority of applications, the single-stage if amplifier described in the preceding section will not provide sufficient gain. A

60-db power gain may be obtained with the two-stage if amplifier circuit shown in Fig. 203. Sensitive receiver and instrument demands may be met with this amount of amplification.

This amplifier also employs Raytheon 2N112/CK760 rf transistors and transistor type miniature if transformers, as in the preceding example. Total current drain from the 6-volt battery is approximately 7 milliamperes dc.

The input impedance of the amplifier is 25,000 ohms; output impedance 600 ohms. These values are satisfactory for working out of a transistorized converter stage and into either a diode or transistor type second detector. The circuit is neutralized automatically by the fixed capacitors C1 and C2.

The figures on the leads of the if transformers in Fig. 203 correspond to the manufacturer's coding and must be followed for correct amplification and neutralization.

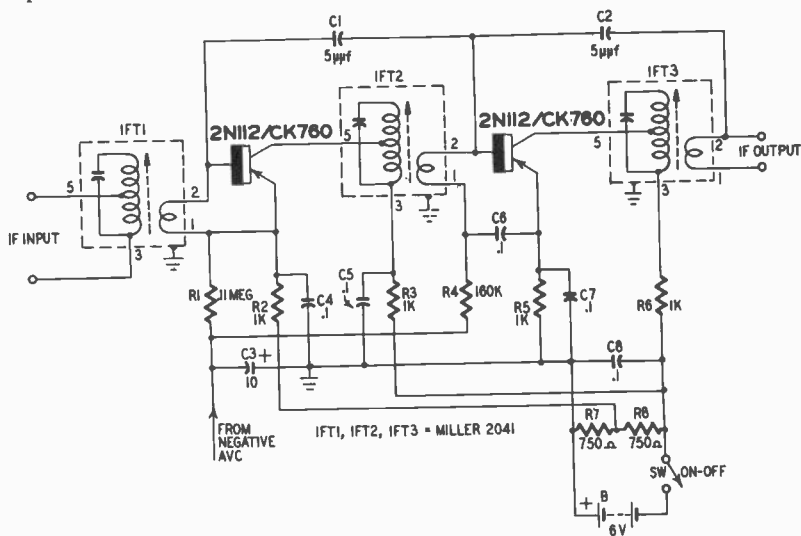


Fig. 203. Two-stage 455-kc if amplifier.

Negative avc voltage, derived from the second detector when the if amplifier is included in a superheterodyne receiver, is applied to the if transistors at the lower ends of resistors R1 and R4. If avc voltage is not available, connect R1 and R4 to the negative 6-volt line (e.g., to the right-hand end of resistor R8).

The amplifier is aligned to 455 kc initially in the conventional manner, as described in the preceding section, with an AM signal generator connected to the IF INPUT terminals and a detector to the IF OUTPUT terminals. The slugs of the transformers (T1, T2 and

T3) are tuned for peak output signal. It is necessary to jockey back and forth between these adjustments until all three transformers are "peaked."

50-kc if amplifier

Low-frequency if amplifiers are employed in heterodyne type instruments and as selectivity boosters for CW receivers. It is advantageous to transistorize such amplifiers when battery operation, especially in portable service, is desired.

Fig. 204 shows the circuit of a 50-kc peaked amplifier employing three Raytheon 2N130 transistors. Parallel-resonant collector tuning is employed, since 50-kc if transformers with the proper impedance ratings for interstage coupling between transistors are not available at this writing. For this purpose, slug-tuned inductors L1 and L2

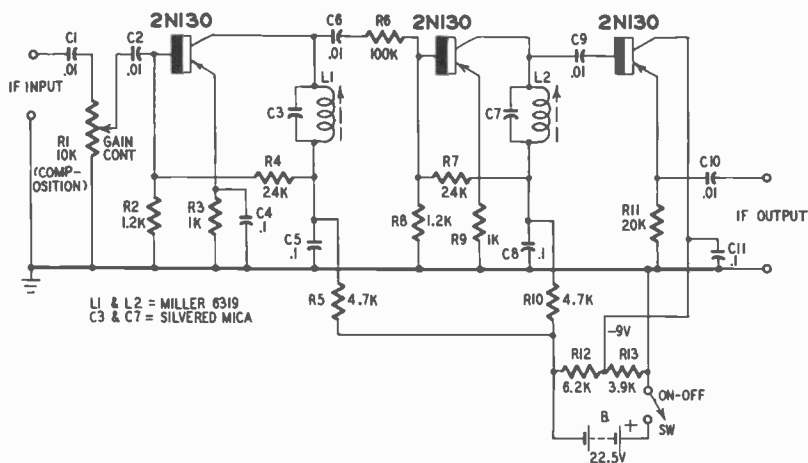


Fig. 204. 50-kc if amplifier.

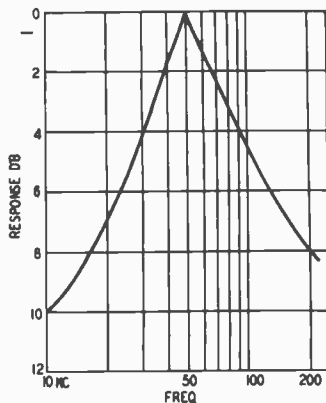
(variable from 15 to 60 mh) are connected in parallel with silvered-mica capacitors C3 and C7, respectively.

The first two stages are common-emitter amplifiers. The emitter current-limiting resistor R9 in the second stage is left unbypassed. The resulting degeneration raises the input impedance of this stage and this feature, plus the effect of the series input resistor (R6), reduces the second-stage loading on the first-stage tuned circuit (L1-C3), thereby preserving the selectivity of the latter. The output stage is a common-collector amplifier which offers a high impedance and minimum detuning to the second-stage tuned circuit (L2-C7) and provides comparatively low impedance output. This

arrangement protects the second-stage tuned circuit from fluctuating external load conditions.

At the 50-kc peak, the overall voltage gain of this amplifier is 500. With GAIN CONTROL R1 set to maximum, the maximum recommended input-signal voltage is 10 millivolts rms. The corresponding maximum output-signal voltage is 5 volts rms. Fig. 205 shows the frequency response.

Fig. 205. Frequency response of the circuit shown in Fig. 204.



Input impedance is approximately 1,000 ohms and output impedance 20,000 ohms. Total current drain is approximately 8 milliamperes dc from the 22.5-volt battery.

The 50-kc if amplifier is aligned with the aid of an amplitude-modulated signal generator: (1) Connect the generator output to the IF INPUT terminals. (2) Connect a detector (ac vacuum-tube voltmeter or high-impedance magnetic headphones in parallel with a 1N34 germanium diode) to the IF OUTPUT terminals. (3) Set GAIN CONTROL R1 to midrange. (4) Switch on the signal generator and if amplifier. (5) Tune the slugs in inductors L1 and L2 for peak output signal, working back and forth between the two adjustments until both stages are peaked. Reduce the generator signal or the amplifier GAIN CONTROL (R1) setting if the amplifier or detector blocks from excessive signal strength.

Transistors for high-frequency rf amplifiers

Radio-frequency amplifiers for operation at frequencies above the standard broadcast band may be constructed, employing essentially the same design as that given in Fig. 201. However, the values of L1, L2, L3, L4, C1 and C2 must be chosen for resonance at the desired operating frequencies and an rf transistor having the proper operating-frequency range employed. A variety of high-frequency transistors is available commercially.

Table 3 shows the alpha cutoff frequency of common rf transistors.

Table 3—Rf Transistors

Type	Manufacturer	Alpha Cutoff Frequency Mc	Class
903	Texas Inst.	3	n-p-n (silicon)
904	Texas Inst.	3	n-p-n (silicon)
904A	Texas Inst.	8	n-p-n (silicon)
905	Texas Inst.	3	n-p-n (silicon)
GT760	Genl. Transistor	5	p-n-p
GT761	Genl. Transistor	10	p-n-p
GT762	Genl. Transistor	20	p-n-p
SB-100	Philco	30 *	p-n-p (surface barrier)
2N94	Sylvania	3.5	n-p-n
2N94A	Sylvania	6	n-p-n
2N112/ CK760	Raytheon	5	p-n-p
2N113/ CK761	Raytheon	10	p-n-p
2N114/ CK762	Raytheon	20	p-n-p
2N124	Texas Inst.	3	n-p-n
2N125	Texas Inst.	5	n-p-n
2N126	Texas Inst.	5	n-p-n
2N127	Texas Inst.	5	n-p-n (silicon)
2N168	Genl. Electric	6	n-p-n
2N168A	Genl. Electric	8	n-p-n
2N169	Genl. Electric	4	n-p-n
2N169A	Genl. Electric	5	n-p-n
2N170	Genl. Electric	2.5	n-p-n

* Maximum operating frequency.

The maximum operating frequency shown in table 3 above is 30 mc. At the time of this writing, this is the highest frequency rating of *commercially available* triode-type transistors but this is not to be construed as the ultimate frequency of transistor operation.

The thinner the base layer, the higher the maximum operating frequency of a junction transistor. In recent developmental transistors produced in the laboratory, base layers only 50 millionths of an inch thick have been formed in transistors having cutoff frequencies between 500 and 600 mc (Bell Telephone Laboratories). But at the time of this writing no predictions have been made as to when these uhf transistors will be available to the experimenter and hobbyist. It may be reasonably expected, however, that new fabrication techniques which make possible the production of microscopic layers in a tiny semiconductor bar will bring uhf junction transistors within the reach of all users within the next few years.

Another significant development, still in the pilot production stage at this writing, is the high-frequency tetrode transistor. One such n-p-n unit (General Electric) has a cutoff frequency of 120 mc and a collector dissipation rating of 50 milliwatts.

With transistors now commercially available with frequency ratings up to 30 mc, and with power transistors available for the audio frequencies, uhf transistors are all that is required to complete the applications picture in receivers and test instruments other than the microwave type.

Uhf transistors with reasonable power dissipation ratings will make possible transistorized portable transceivers (especially "handie-talkies"), compact transmitters and receivers for garage-door opening and similar control purposes, and uhf signal generators and "grid"-dip oscillators. Conventional, low-powered, uhf transistors will be invaluable in TV front ends and high-frequency video amplifiers, and in some radar and altimeter applications.

dc amplifiers

THE transistor basically is a current-operated device. Current amplification therefore is inherent in it. The base-to-collector amplification factor (beta) expresses the extent to which common-emitter or common-collector transistor circuits will amplify currents.

Direct current and voltage amplifiers using transistors are simple and straightforward, often more so than ac amplifiers. In most such basic circuits, high, true current gain is possible.

This chapter describes several practical dc amplifier circuits. The resistance-coupled circuits will be seen to bear a close resemblance to direct-coupled vacuum-tube amplifiers. The direct-coupled transistor amplifier, on the contrary, has no counterpart in tube practice.

Single-stage current amplifier

Fig. 301 shows the circuit of a single-stage direct current amplifier in which amplification is provided directly by the common-emitter beta of the transistor. Because beta is a short-circuit parameter, the load device connected to the DC OUTPUT terminals must have low resistance. High values of load resistance decrease the current gain. The load resistance should be 100 ohms or less.

With a dc supply (B) of 1.5 volts a current gain of 40 is obtained: 25 microamperes input will give an output-current increase of 1 ma. (This is very nearly the full beta value of 45 for the Raytheon CK721 transistor.) The steady direct current (i_{co}) flowing when the DC INPUT terminals are open is 100 μ a.

The current gain and the zero-signal collector current vary with the resistance of the source to which the DC INPUT terminals are connected. The zero-signal collector current and the current gain both decrease as the input resistance decreases.

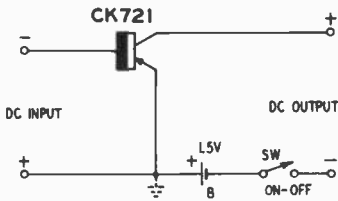


Fig. 301. *Single-stage current amplifier.*

Since i_{co} and beta both vary with temperature, this rudimentary amplifier is suitable only in applications in which the ambient temperature can be stabilized. In spite of this shortcoming, however, its simplicity and high gain fit this amplifier for use between a high-resistance dc source and a low-resistance load device (such as relay or model actuator).

Transistors with higher beta ratings will provide higher current amplification. Raytheon CK725 and 2N65, for example, have maximum current gains of 90.

Single-stage current amplifier with fixed-resistance input

The tendency of current gain to vary with input (generator) resistance in the simple direct-current amplifier may be minimized to a large extent by fixing the input resistance by means of a relatively low value of shunting resistance, R (Fig. 302).

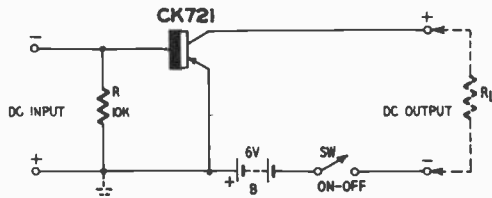


Fig. 302. *Single-stage current amplifier with fixed input resistance.*

While the presence of this resistor reduces the current gain to 25 (i. e., 40- μ a input gives a 1-ma output-current increase in an external load R_L not exceeding 100 ohms) this gain is stabilized against the large changes usually encountered when the generator resistance varies from a few ohms to several megohms.

At the measured current gain of 25, the input resistance of the stage is approximately 10,000 ohms, dc input voltage 0.5 volt, dc input current 40 μ a, zero-signal collector current 100 μ a and maxi-

zero-signal collector current increase is 1.0 ma (that is, from 100 μ a to 1.1 ma).

Increasing the external load resistance R_L , decreases the gain. For example, raising R_L to 1,000 ohms drops the gain 10.

Cascaded p-n-p current amplifier ¹

Fig. 303 shows how current-amplifier stages may be direct-coupled for increased gain.

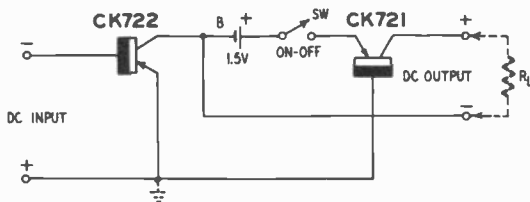


Fig. 303. Cascaded p-n-p current amplifier.

In this circuit, the collector output current of the first transistor (Raytheon CK722) flows directly through the base-emitter circuit of the second transistor (Raytheon CK721). The input current thus is amplified successively by the CK722 and CK721.

Both stages are connected as common-emitter dc amplifiers and the 1.5-volt battery (B) supplies both collectors. A dc input of 10 μ a gives an output-current increase of 1.0 ma through a load resistance R_L of 100 ohms or less. This represents a current gain of 100 times, or 40 db.

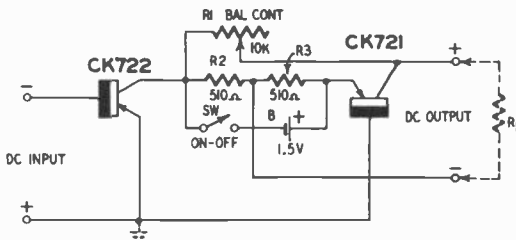


Fig. 304. Cascaded current amplifier with balance control.

The zero-signal collector current of the CK721 output transistor is the sum of the CK721 zero-signal current (i_{co}) and the amplified zero-signal current of the CK722. This initial current is large enough to be troublesome in some applications. However, it may be bucked out by means of any of the conventional balancing circuits.

Fig. 304 shows how a balancing circuit be added for setting the initial collector current to zero. Here the bucking circuit is a four-arm bridge consisting of resistors R1, R2, R3 and the internal collector resistance of the CK721. Arm R1 of the bridge is made variable for adjustment. When the bridge is set to null by adjustment of R1, the current through the load resistor R_L is reduced to zero and only the current change due to application of the input signal will flow through it.

All common-emitter, direct-coupled, current amplifier circuits employing a single battery have the slight disadvantage that there is no ground terminal common to both input and output. Thus, the input or output may be grounded but not both simultaneously.

To employ a common ground, a separate battery must be used in each stage. Otherwise resistance coupling must be employed.

Cascaded p-n-p-n-p-n current amplifier ²

The effect of the amplified zero-signal collector current of the input transistor in a two-stage direct-coupled current amplifier is minimized by employing a p-n-p (Sylvania 2N34 in Fig. 305) in the input stage and an n-p-n transistor (Sylvania 2N35) in the output stage. The zero-signal collector currents in the two transistors are in opposite directions. This circuit sometimes is called a complementary-symmetry dc amplifier.

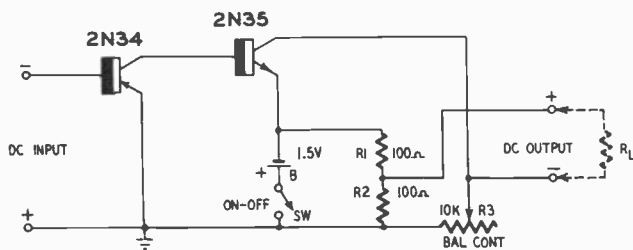


Fig. 305. Cascaded p-n-p-n-p-n current amplifier.

Here, as in the previous two circuits, a single 1.5-volt battery (B) supplies the two transistors. Since the n-p-n and p-n-p collectors require voltages of opposite polarity, the battery is connected in series with the output-stage emitter. Negative voltage thus is applied to the 2N35 emitter and (through the emitter-base circuit of the 2N35) to the 2N34 collector.

The balancing circuit is similar to the one explained previously in connection with Figs. 303 and 304. It consists of a four-arm

bridge (R_1 , R_2 , R_3 and the internal collector resistance of the 2N35). When R_3 is set for null, no dc flows through the external load resistance.

With $0.4 \mu\text{a}$ dc input, this amplifier delivers an output of $100 \mu\text{a}$ through a 1,000-ohm load resistance R_L . This represents a current gain of 250 times, or 48 db. With the amplifier delivering full output, drain from the 1.5-volt battery is 7.5 ma dc.

Single-stage dc voltage amplifier

Fig. 306 shows the circuit of a single-stage, resistance-coupled, dc voltage amplifier employing a Raytheon CK721 transistor. With a

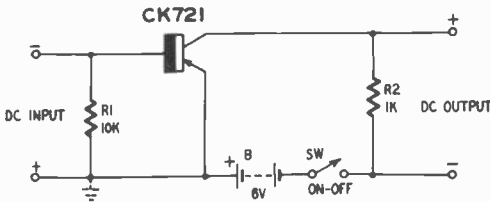


Fig. 306. Single-stage dc voltage amplifier.

measured input resistance of approximately 2,550 ohms, this amplifier provides a voltage gain of 5. For an input signal of 0.2 volt, the output-voltage change across collector resistor R_2 is 1 volt. The zero-signal voltage drop across R_2 is 2.4 volts, which may be nulled by applying a bucking voltage in series with the top of R_2 and the positive DC OUTPUT terminal. Current input is $80 \mu\text{a}$ and the corresponding output-current change is 1 ma in R_2 .

Cascaded dc voltage amplifier

Resistance-coupled dc voltage amplifier stages of the type described in the preceding section may be cascaded, as shown in Fig.

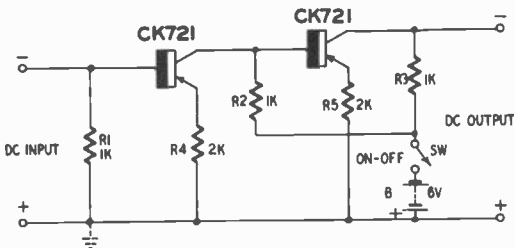


Fig. 307. Cascaded dc voltage amplifier.

307, for higher voltage gain. Transistor noise is the factor limiting the number of stages which may be coupled in this manner.

Employing two Raytheon CK721 transistors, this resistance-coupled circuit provides a dc voltage gain of 25. With an output of 1 volt, the dc input signal is 40 mv.

Single-stage power-type current amplifier

The circuits described up to this point, while providing high gain, handle relatively low current levels. By substituting a power transistor for the conventional type, larger currents may be accommodated. Fig. 308 shows such a circuit.

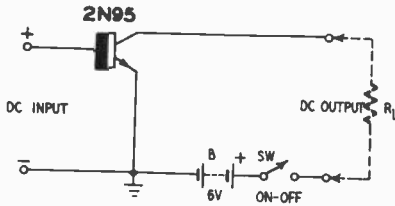


Fig. 308. Single-stage power-type current amplifier.

Here a Sylvania 2N95 power transistor is operated at 6 volts. Providing a current gain of 35, this single stage provides a dc output-signal increase of 250 ma in response to a dc input signal of 10 ma. The initial, zero-signal dc collector current is less than 10 ma. To utilize most fully the rated beta of the 2N95, the external load resistance R_L must be 1 ohm or less. This low resistance requirement restricts the amplifier load devices to solenoids, lamps, high-current relays, heater coils and similar low-ohmic equipment. Proportionately lower amplification will be obtained with higher R_L values.

Two-stage, high-gain, power-type current amplifier

When higher current amplification is required than can be obtained with the power type amplifier described, a conventional-transistor input stage may be prefixed.

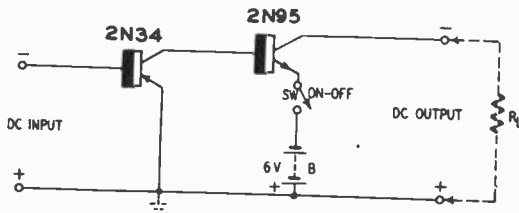


Fig. 309. High-gain power-type current amplifier.

This arrangement is shown in Fig. 309. A Sylvania 2N34 conventional transistor is used in the input stage and a Sylvania 2N95

power transistor in the output stage. This gives a p-n-p-n-p-n type of circuit similar to the complementary-symmetry amplifier shown in Fig. 305.

The zero-signal output is 10 ma dc. An input-signal of 150 μ a dc gives an output current increase of 260 ma. This represents an overall current gain of 1,670 times, or 64 db. As in the preceding circuit, maximum current amplification is obtained only with low values of external load resistance R_L . Resistances of 1 ohm or less are ideal. Proportionately lower gain is obtained with higher R_L values.

Alpha-beta conversion ³

Since the maximum current gain obtainable with a transistor in the common-emitter amplifier circuit is equal to beta (β , the base-to-collector amplification factor) or to alpha (a , the emitter-to-collector current amplification factor) in a common-base circuit, it is necessary to know the a and β values when designing transistorized dc amplifiers. Unfortunately, some tables of transistor characteristics list beta and some list alpha, but very few give both. However, if one factor is known, the other may be calculated. Thus:

$$a = 1 - (1/\beta + 1)$$

and

$$\beta = a / (1 - a)$$

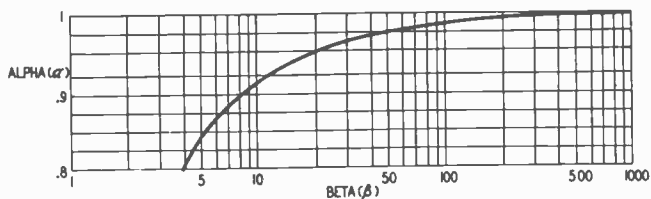


Fig. 310. Alpha-beta conversion curve.

For convenience, the graph given in Fig. 310 may be used to convert alpha to beta and vice versa. This graph saves calculations in most instances, but the curve rises so steeply at alpha values beyond about 0.991 that calculations are recommended when high accuracy is desired beyond the point at which

$$a = 0.980$$

or

$$\beta = 100.$$

REFERENCES

- ¹ Rufus P. Turner, *Transistorized Voltmeter*. RADIO-ELECTRONICS Magazine, December, 1954; p. 54.
- ² The work leading to the design of the zero-balanced p-n-p-n-p-n circuit was performed by the author for Cornell-Dubilier Electric Corp.
- ³ Rufus P. Turner, *Transistor Alpha-Beta Conversion*. Tele-Tech & Electronic Industries, March, 1956; p. 74.

oscillators

TRANSISTORIZED oscillators are unique in their ability to operate with good efficiency at dc power inputs of extremely low level. Oscillation has been obtained when the transistor bias consisted of a few microwatts of dc supplied by such an unconventional source as a thermocouple, photocell, charged capacitor or saliva-and-coin makeshift wet cell.

Practical af and rf transistor oscillators are feasible in a wide variety of output power ratings. Like transistorized amplifiers, transistor oscillators make efficient use of their dc supplies. Because of the low power drain when using conventional small transistors, these oscillators are free from the high heating effects which cause severe frequency drift in vacuum-tube oscillators.

Single-frequency audio oscillator

Fig. 401 shows the circuit of a single-frequency audio oscillator of the tickler-feedback type.

Employing a Raytheon CK722 transistor, this circuit delivers a maximum output of 0.65 volt rms into a high-impedance load. Direct-current drain is 8 to 10 μ a at 1.5 volts. Higher-voltage batteries (up to 6 volts) will give larger audio output with increased dc drain.

The operating frequency is determined by the inductance of the low-impedance winding of transformer T and the capacitance C1. A 0.05- μ f capacitor with the transformer specified in Fig. 401 gives a frequency of approximately 1,000 cycles. The frequency may be raised by decreasing C1 and lowered by increasing this capacitance.

Transformer T must be connected as shown, according to the manufacturer's color coding to obtain feedback of the proper polarity for oscillation.

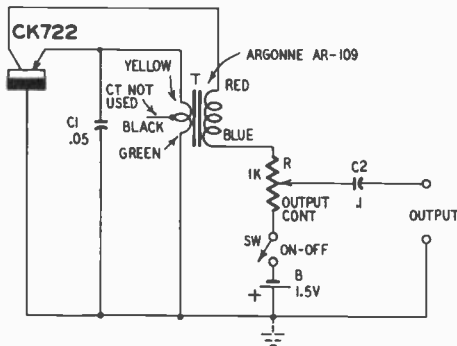


Fig. 401. Single-frequency audio oscillator.

The output waveform may be improved (distortion reduced), as required by individual demands, by a resistance connected in the external base lead of the transistor. The amount of resistance, 200 ohms or higher, must be determined experimentally for an individual transistor and transformer.

Simple code-practice oscillator ¹

The magnet coils of a pair of headphones supply the required inductance in the audio oscillator circuit shown in Fig. 402. This arrangement makes a convenient code-practice oscillator of the Col-

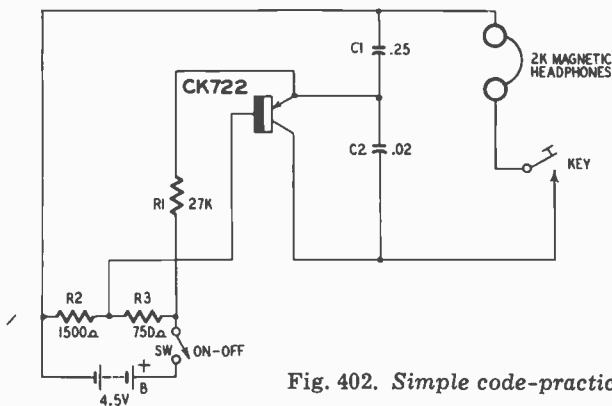


Fig. 402. Simple code-practice oscillator.

pitts type, in which the operating frequency is determined by the capacitance combination C1-C2 and the headphone inductance. A Raytheon CK722 transistor is employed.

The headphones must be of the magnetic type. Crystal units are inoperative in this circuit. With a pair of Trimm 2,000-ohm head-

phones and with $C1 = .25 \mu\text{f}$ and $C2 = .02 \mu\text{f}$, the operating frequency was measured as 750 cycles. The frequency may be raised by decreasing both $C1$ and $C2$ and lowered by increasing their values. At any chosen frequency, the ratio between the two capacitances must be approximately 10 to 1, as shown in Fig. 402.

Total direct-current drain is approximately 3 ma. A 50,000-ohm potentiometer may be connected as a volume control, if desired, in series with the collector lead of the transistor.

Loudspeaker-operating code-practice oscillator ²

Classroom code instruction often requires the use of a loudspeaker. This enables the instructor to give verbal directions while demonstrating sound signals—a task not easily performed when students are wearing headphones.

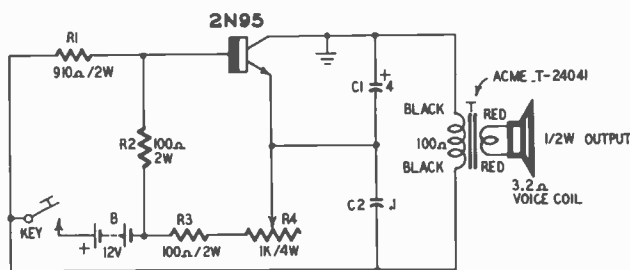


Fig. 403. Loudspeaker-operating code-practice oscillator.

Fig. 403 shows the circuit of a code-practice oscillator which will deliver 0.5 watt of audio power to a loudspeaker. This output is adequate for small- and medium-sized classrooms. A Sylvania 2N95 power transistor is operated from a 12-volt battery.

This oscillator employs a Colpitts-type circuit, similar to the low-powered oscillator described earlier. The frequency is determined by the inductance of the primary winding of transformer T and the capacitance of the $C1$ - $C2$ combination. Using the transformer and the capacitances specified in Fig. 403, the frequency is approximately 360 cycles at the lowest-resistance setting of potentiometer R4. The frequency increases to approximately 3,500 cycles when R4 is set to its full resistance of 1,000 ohms. This gives a frequency range of nearly 10 to 1.

The dc drain when the key is closed is 170 ma. Because of the intermittent nature of code-practice oscillator service, the operating current may be supplied by a battery of size-D flashlight cells if compactness is a requirement.

Phase-shift audio oscillator ⁸

The vacuum-tube phase-shift oscillator is noted for its simplicity and low distortion. In this circuit, the operating frequency is determined by the resistance and capacitance values in a three-leg R-C network which transmits feedback voltage in the proper phase for oscillation. This phase-shift network is connected between plate output and grid input of a single-tube stage.

Many experimenters have failed to apply the phase-shift principle successfully to transistorized oscillators. The main reasons for poor results have been (1) failure to take into consideration the influences of the transistor input impedance on the phase-shift network parameters and (2) insufficient amplification in the transistor stage.

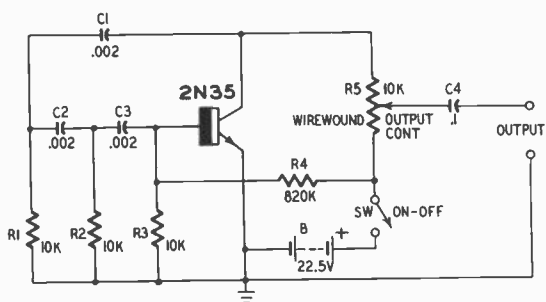


Fig. 404. Phase-shift audio oscillator.

In the oscillator circuit shown in Fig. 404, the required 180 degrees of phase shift are provided by the R-C network C1-C2-C3-R1-R2-R3. The resistance values in this network have been chosen with respect to the transistor input resistance shunting resistor R3. The R and C values have been selected for 2,000-cycle oscillation. Adequate gain is provided by the Sylvania 2N35 transistor in the common-emitter circuit.

The maximum signal-output voltage across a high-impedance load is 3.5 volts rms when OUTPUT CONTROL R5 is set to maximum. Total harmonic distortion is 0.26%. Total dc drain is 1.1 ma from the 22.5-volt battery. Resistor R4 acts with the frequency-network resistor R3 to form a voltage divider for base-bias stabilization.

The oscillator frequency may be altered by changing the value of each network capacitor (C1, C2, C3) by the same amount. Increasing the capacitance lowers the frequency and vice versa.

For low-impedance output with approximately 12-db power gain, a common-collector amplifier stage may be added to the phase-shift oscillator.

Step-type variable-frequency audio oscillator

The basic tickler-feedback circuit given in Fig. 401 may be modified to deliver several selected frequencies by providing switched capacitors to tune the transformer winding. Fig. 405 shows such a simple, step-type, variable-frequency oscillator circuit.

Capacitors C1, C2, C3 and C4 are selected experimentally for the desired frequencies. The capacitances will vary somewhat for individual transformers, depending upon the primary inductance of T, hence are not specified here. The capacitors are switched successively into the circuit in parallel with the transformer primary by switch S1. All other circuit constants are the same as those given in Fig. 401.

Tuned variable-frequency audio oscillator

Step-type frequency control is limited to those applications calling for discrete test frequencies only. Most applications require, not spot frequencies, but a continuous smooth variation of oscillator frequency. Simple step-type oscillators such as the circuit shown in Fig. 405 or a smaller switched-capacitor version of Fig. 404, accordingly are special-purpose devices.

Fig. 406 shows the circuit of a smoothly tuned variable-frequency audio oscillator covering the range 200 to 10,000 cycles in one complete turn of the tuning dial. This is a resistance-tuned oscil-

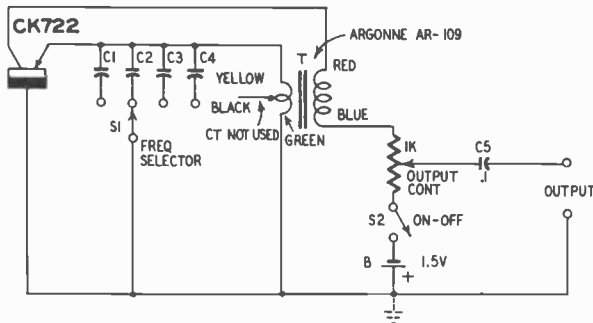


Fig. 405. Step-type variable-frequency audio oscillator.

lator based upon the Villard selective amplifier. Employing four General Electric 2N107 transistors, this oscillator delivers 0.75 volt rms open circuit. Its output is low-impedance. Higher output may be obtained with an output booster amplifier stage.

Tuned phase-shift networks (C1-R3 and C3-R8) are included in the first two transistor amplifier stages (V1 and V2). The

adjustable arms of these networks are supplied by the dual, ganged 100,000-ohm potentiometer R3–R8. The two sections of this control

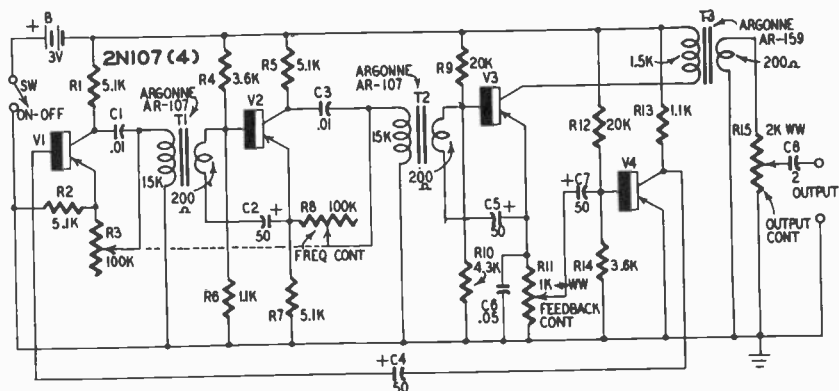


Fig. 406. Tuned variable-frequency audio oscillator.

must track closely and each must have an audio taper. This potentiometer is the frequency control and its dial may be calibrated to read directly in cycles per second.

The final transistor stage (V4) provides the regenerative feedback, through capacitor C4, required for oscillation. Oscillation occurs at the single frequency at which zero phase shift is provided by the C1R3–C3R8 network. The degree of oscillation is governed by the setting of the feedback-control potentiometer R11. By setting this control to the point at which the circuit just begins to oscillate stably, output-signal distortion may be held to 1% or less. Under these conditions it may be necessary to readjust R11 for oscillation as the frequency is shifted throughout the tuning range.

The polarity of connections to the transformers T1 and T2 is very important. If the circuit fails to oscillate or does not cover the 200–10,000-cycle range, the primary connections of T1 or T2 (but not both) should be reversed. All three transformers in the circuit must be mounted far enough apart and oriented so that no interaction occurs between their magnetic fields.

100-kc crystal oscillator

Fig. 407 shows the circuit of a 100-kc crystal oscillator suitable for use as a secondary frequency standard or general-purpose experimental signal source. This oscillator employs a Sylvania 2N94 transistor as a common-base oscillator and a Sylvania 2N34 as a common-collector output amplifier. Single-battery operation is made possible through this use of an n-p-n transistor in the common-base stage and a p-n-p in the common-collector stage.

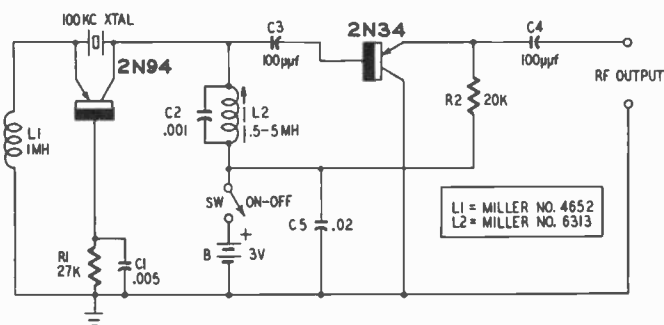


Fig. 407. 100-kc crystal oscillator.

The oscillator is tuned to the crystal frequency by adjusting the slug in inductor L2 for peak deflection of an rf vacuum-tube voltmeter connected to the RF OUTPUT terminals. For maximum stability, capacitors C2 and C3 must be of the silvered mica type.

The open-circuit output-signal voltage is approximately 0.8 volt rms. The total, dc drain is less than 1 ma from the 3-volt battery B. For long service at constant voltage, it is recommended that B consist of two mercury cells (Mallory RM1R) connected in series (total voltage = 2.69 volts).

Self-excited 100-kc oscillator

Fig. 408 shows the circuit of a simple, self-excited 100-kc oscillator employing a single Raytheon 2N112/CK760 rf transistor. Drawing a total direct current of 1.25 ma, this oscillator possesses good stability and is recommended for applications not demanding the high stability of a crystal oscillator.

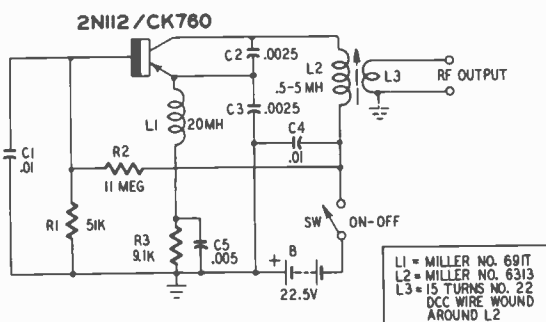


Fig. 408. Self-excited 100-kc oscillator.

A Colpitts-type oscillator circuit is employed. The operating frequency may be set exactly to 100 kc by the conventional zero-beat

method using WWV transmissions, by adjustment of the tuning slug in inductor L2.

The signal-output coupling coil L3 consists of 15 turns of No. 22 dcc wire wound tightly around the outside of L2. This coupling coil delivers an open-circuit rf signal output voltage of approximately 1 volt rms.

Stabilizing dc base bias is provided by the voltage divider R1-R2. The radio-frequency choke L1, prevents grounding the rf component of emitter current.

High-frequency crystal oscillator

The oscillator circuit shown in Fig. 409 may be used with crystals having fundamental frequencies up to 7.5 mc. Employing a Sylvania 2N94A rf transistor, this is a conventional, collector-tuned, common-base circuit.

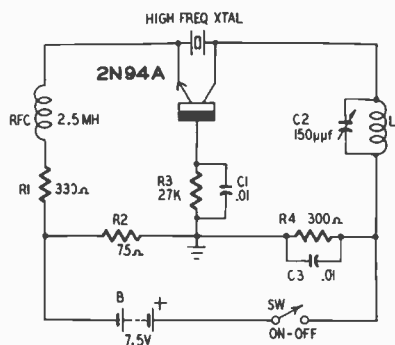


Fig. 409. High-frequency crystal oscillator.

A 7.5-volt battery (B) supplies the bias currents. A voltage divider is formed by resistors R2 and R4 in series. Emitter bias of 1.5 volts is developed by the voltage drop across resistor R2. Collector bias of 6 volts is developed by the voltage drop across R4. The bleeder (voltage-divider) direct current is 20 ma. This high value is required for stabilization of the 4-ma emitter and collector currents. Total dc drain is 28 to 30 ma.

The collector-tuned circuit L-C2 is resonated to the crystal frequency by means of variable capacitor C2. The coil may be wound for a particular crystal frequency according to tables and directions found in amateur radio handbooks. Tuning is indicated by peak deflection of an rf vacuum-tube voltmeter or crystal-diode meter inductively coupled to the tank coil L. The rf voltage developed across coil L at resonance is approximately 5 volts rms.

Energy may be coupled out of the collector tank by conventional capacitance coupling, inductive coupling or link coupling to coil L.

Self-excited rf oscillator

Fig. 410 shows the circuit of a self-excited Hartley-type oscillator having a continuous tuning range of 500 kc to 4.5 mc. This stabilized circuit, employing a Sylvania 2N94 rf transistor, has a radio-frequency output of 0.65 volt rms (open circuit) and a total drain of 2 ma dc.

Tuning is accomplished with a single-section 350- or 365- μf variable capacitor C2. The tuning range may be set with the aid of the 1260- μf padder capacitor C1 and the tuning slug of inductor L.

Inductor L is a miniature, commercial, transistor-type, high-Q oscillator coil assembly, consisting of a two-tap main coil and a secondary coupling coil. The figures shown on the coil terminals in Fig. 410 correspond to the manufacturer's labeling and must be followed to obtain oscillation.

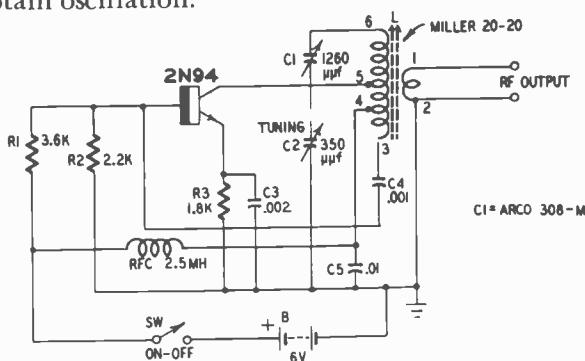


Fig. 410. Self-excited rf oscillator.

Base bias stabilization is provided by the voltage divider R1-R2. The radio-frequency choke (RFC) prevents the grounding through the voltage divider of rf energy from the output tank circuit. Emitter current-limiting resistor R3 is bypassed by capacitor C3 to prevent reduction of gain from degeneration.

Wide-range rf oscillator

Radio-frequency transistors have made possible the design of oscillators operating at higher frequencies than the standard broadcast band. The circuit shown in Fig. 411 employs a Philco SB100 surface-barrier transistor and oscillates as high as 44 mc. Specially picked transistors possibly will oscillate at somewhat higher frequencies.

This tickler-feedback circuit employs either plug-in coils or coil switching. L2 is the tuned collector coil and L1 the low-impedance

base tickler. For each frequency band, both coils are wound on the same form. The frequency coverage provided by the oscillator is

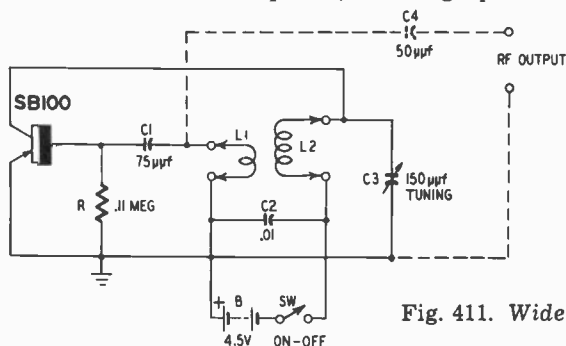


Fig. 411. Wide-range rf oscillator.

1 to 44 mc in four bands: 1–2.7, 2.5–7, 6–17 and 15–44. Table 4 gives winding instructions for the coils.

Rf output may be taken from the oscillator through capacitance coupling to the tickler coil, as shown by the dashed lines, or link coupling may be provided by two or three turns of the same size wire used in L2 and wound as close as possible to the ground end of L2.

Total dc drain of the oscillator is approximately 1.1 ma from the 4.5-volt battery B.

Table 4—Coil Table for Rf Oscillator

Band A 1–2.7 mc	L1: 25 turns No. 30 enameled wire closewound over ground end of L2. L2: 100 turns No. 30 enameled wire closewound on 1-inch-diameter form.
Band B 2.5–7 mc	L1: 9 turns No. 24 enameled wire closewound over ground end of L2. L2: 35 turns No. 24 enameled wire closewound on 1-inch-diameter form. Space to winding length of $\frac{3}{4}$ inch.
Band C 6–17 mc	L1: 7 turns No. 24 enameled wire closewound over ground end of L2. L2: 23 turns No. 24 enameled wire closewound on $\frac{1}{2}$ -inch diameter form.
Band D 15–44 mc	L1: 7 turns No. 20 enameled wire closewound adjacent to ground end of L2. L2: 9 turns No. 20 enameled wire on $\frac{1}{2}$ -inch diameter form. Space to winding length of $\frac{1}{2}$ inch.

Multivibrator

The multivibrator circuit shown in Fig. 412 is analogous to the symmetrical vacuum-tube multivibrator. Like the tube circuit, it

consists of a two-stage R-C-coupled amplifier with positive feedback (through capacitor C4) from the second-stage output to the first-stage input. The n-p-n transistors are Sylvania 2N35.

With the circuit constants shown, the natural frequency of this multivibrator is approximately 2,000 cycles. For this symmetrical

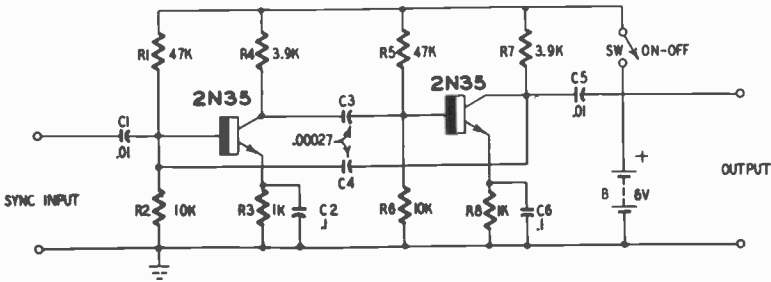


Fig. 412. Multivibrator circuit.

circuit the approximate operating frequency $f = 1,000 / (2RC)$, where f is the frequency in kilocycles, C the capacitance in microfarads of either $C3$ or $C4$ (both have the same capacitance value) and R the transistor input resistance. The latter will vary between 900 and 1,200 ohms with individual transistors. To change the frequency, both $C3$ and $C4$ must be changed by the same amount; increase the capacitance to decrease the frequency and vice versa.

The peak-to-peak output is approximately 2.5 volts. Total dc drain is 1.6 ma from the 6-volt battery B . The multivibrator may be synchronized readily by means of a 7-volt peak signal (sinusoidal or nonsinusoidal) applied to the SYNC INPUT terminals.

Blocking oscillator

Fig. 413 shows the circuit of a simple blocking oscillator designed for 1- μ sec output pulses at a repetition rate of 10 kc. For fast rise time, an n-p-n rf transistor (Sylvania 2N94) is employed.

Operation of this circuit is somewhat similar to that of a vacuum-tube blocking oscillator. The transistor collector current normally is held at cutoff by the negative base bias due to the charge in capacitor $C2$. A positive pulse applied to the TRIGGER INPUT terminals momentarily discharges $C2$ through resistors $R1$ and $R2$ and causes collector current to flow. The circuit oscillates during this interval, which is determined by the $C2$ - $R1$ - $R2$ time constant because of positive feedback through the pulse transformer T . The peak amplitude of the positive input-trigger pulse must be better than 3 volts. The output-pulse duration is adjustable by means of rheostat $R2$.

Pulse-output voltage is delivered by the third winding of transformer T through capacitor C3.

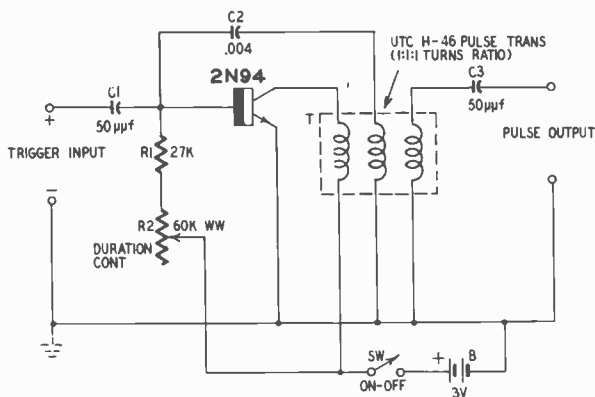


Fig. 413. Blocking oscillator circuit.

Light-powered oscillators ^{4, 5}

The high efficiency of the junction transistor and its ability to operate at extremely low collector voltage make possible practical os-

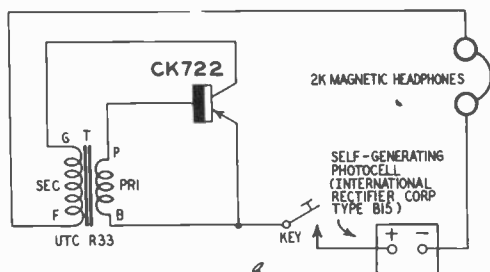
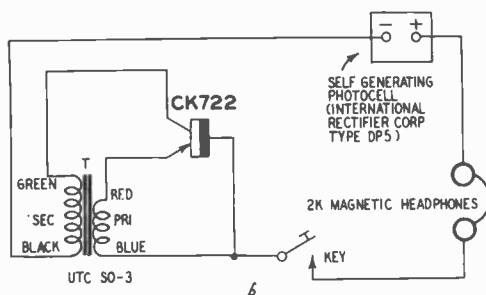


Fig. 414-a, -b. Light-operated oscillators: (a) conventional transformer; (b) subminiature transformer.



cillators operated from microwatt dc supplies. A typical example is the light-powered oscillator in which the dc collector bias is supplied by a self-generating photocell.

Two light-operated oscillator circuits are shown in Fig. 414. While these are wired as code-practice oscillators, they may (by replacing the telegraph key with an on-off switch) be employed in any other application utilizing a tone signal generated by the action of light rays. The circuit becomes a telegraph when the headphones are removed to the distant end of a two-wire line running from the oscillator.

Self-generating selenium photocells are used in each circuit. Fig. 414-a uses a standard-size audio transformer for feedback while Fig. 414-b employs a subminiature unit. In Fig. 414-a, the tone frequency is approximately 2,000 cycles when the photocell is illuminated by moderate sunlight. The frequency decreases as the light intensity increases and vice versa. In Fig. 414-b, the frequency is approximately 900 cycles. In each circuit, the frequency may be lowered by connecting a capacitor in parallel with the secondary of the transformer.

Each circuit will operate with artificial illumination as well as with sunlight. In subdued room light, the arrangement shown in Fig. 414-b will develop a signal of 0.02 mv rms across 2,000-ohm magnetic headphones. A 100-watt incandescent lamp 1 foot from the photocell produces a 0.5-mv rms signal. From 1 to 2 millivolts can be obtained when the photocell is in direct sunlight.

Oscillator operating notes

To obtain maximum performance from transistorized oscillators and to protect the transistors from accidental burnout, some precautions are necessary that often can be neglected in amplifier operation.

Keeping within maximum dissipation ratings is very important. When no power is drawn from the oscillator, the transistor is called upon to dissipate the entire dc power input. The heating can destroy the transistor if it is excessive. When the oscillator *must* be unloaded, check the collector current and voltage values to be sure that the product $V_c \times I_c$ does not exceed the maximum allowable dissipation. At elevated ambient temperatures, the allowable dissipation falls. Follow the derating procedure recommended by the transistor manufacturer.

Heat sinks should be provided when power transistors are operated at their full ratings in oscillators. This will prevent heat destruction of the transistor. While bolting the power transistor to the metal chassis will suffice in most instances, a special large-area sink may be required for some types of operation where dis-

sipation is high. Consult the transistor manufacturer's literature for definite requirements.

Use "stiff" dc bias networks to prevent collector current runaway. Where required, such networks are shown in the circuits in this book. An oscillator often is more inclined toward runaway than is an amplifier using the same transistor and dc voltages.

In audio oscillators, examine the waveform of the output signal, using an oscilloscope or harmonic distortion meter. The distortion percentage tends to run somewhat higher than in comparable tube oscillators because of semiconductor nonlinearity. Usually, one or more components in the circuit can be adjusted (with an individual transistor) to reduce distortion significantly.

Because the internal capacitances of audio transistors are high, tank loading can be severe in oscillator circuits. For this reason, it is a good idea to tap the transistor down the tuned circuit (when one is employed), whenever possible, instead of connecting it across the full tank.

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- ² Rufus P. Turner, (Guy Dexter), *Transistorized Code-Practice Oscillator*, Radio & Television News, April, 1956; p. 126.
- ³ Rufus P. Turner, *Transistorized Phase-Shift Oscillator*, Radio & Television News, April, 1956; p. 108.
- ⁴ Rufus P. Turner, *Transistor Oscillator Is Powered by Light*, RADIO-ELECTRONICS Magazine, August, 1953; p. 66.
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power supplies

THE transistor has four important roles as a power supply component: (1) controlled rectifier similar to the grid-controlled vacuum-tube rectifier, (2) voltage regulator for the electronic stabilization of dc voltage, (3) constant-current converter or adaptor for converting a constant-voltage output into a constant-current output and (4) oscillator in inverter-type supplies. Transistors make possible for the first time fully electronically-regulated low-voltage dc supplies.

Because the transistor has no filament, the power supply circuits in which it can be used are considerably simplified and their power requirements markedly reduced. Its small size and freedom from heat-up time are added advantages.

Variable low-current dc power supply

The transistor serves as a controlled half-wave rectifier in the circuit shown in Fig. 501. Adjustment of potentiometer R2 changes the bias applied to the base of the n-p-n transistor (Sylvania 2N35) and this varies the dc output from zero to 10.5 volts (no-load) in response to 115 volts ac input to the filament-type transformer, T.

The curve in Fig. 502 shows the voltage regulation obtained when R2 is set for maximum dc output voltage. While this plot indicates only 50.5% regulation, this poor value imposes no difficulty since at any current level within the range the voltage may be reset to a desired level (if it has fallen off) by adjustment of potentiometer R2.

Output current drains higher than 10 ma are not recommended since this is the maximum rated collector current of the 2N35 tran-

sistor. If difficulty is experienced in covering the full output voltage range from zero to maximum, the value of the limiting resistor R1 may be changed.

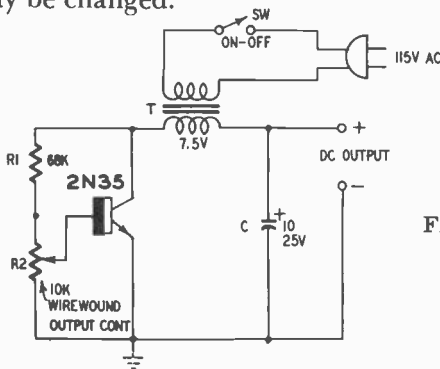


Fig. 501. Variable low-current dc power supply.

Variable high-current dc power supply

A power transistor may be substituted in the simple variable power supply to deliver higher output currents. Fig. 503 shows the circuit after the necessary changes have been made.

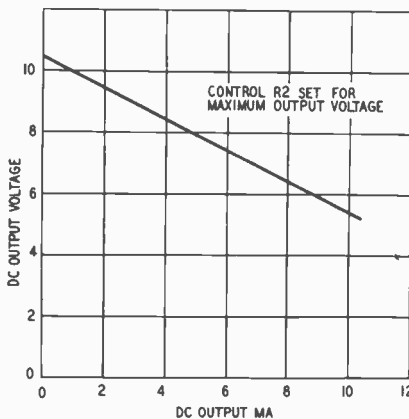


Fig. 502. Voltage regulation of low-current supply.

A Sylvania 2N95 power transistor in this circuit delivers a dc output voltage which is continuously variable between zero and 10 volts, with maximum output current up to 1.5 amperes.

At each output voltage level, the maximum current which may be drawn with safety to the transistor is limited by the maximum power dissipation rating of the transistor. For the 2N95, the maximum dissipation is 1.5 watts if the transistor is operated in free air; 4 watts when a heat sink is provided. Fig. 504 illustrates the performance limits. For example, curve A shows that at 4-volt output 0.5 ampere may be drawn safely when the transistor is operated in

free air but curve B shows that 1 ampere could be drawn at this same voltage if a heat sink were provided.

A heat sink may be obtained simply by bolting the transistor solidly to an aluminum chassis (based upon chassis dimensions of 25 -square-inch area and 1/16-inch thickness). However, the transistor must be insulated electrically from the chassis by a mica separating washer (not more than 0.0005-inch thick) since the transistor case is connected internally to the collector electrode.

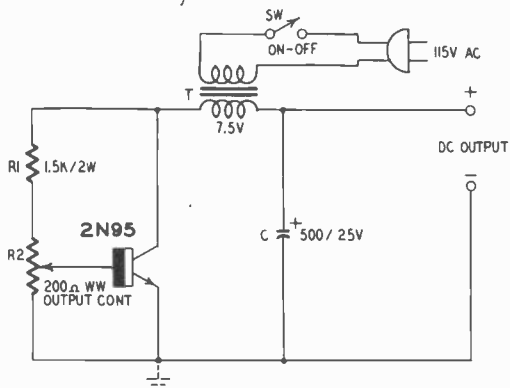


Fig. 503. Variable high-current dc power supply.

The resistance of the OUTPUT CONTROL potentiometer R2 has been changed to 200 ohms since this is the maximum external base-emitter resistance permissible for the type 2N95 transistor.

If difficulty is experienced in covering the full output voltage range from zero to maximum, the resistance of the limiting resistor R1 may be altered.

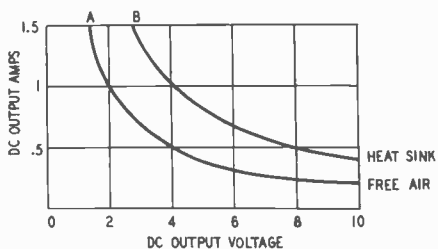


Fig. 504. Performance limits of high-current supply.

Full-wave variable dc power supply

Full-wave rectification often is preferred in a power supply because of its higher output current than that supplied by a half-wave circuit and the increased ease of filtering.

Fig. 505 shows the circuit of a full-wave, high-current, variable power supply which delivers twice the output current given by the

half-wave circuit described previously. Up to 3 amperes output may be obtained. The secondary of the transformer (T) may be chosen to supply up to 21 volts rms each side of center tap. This is the

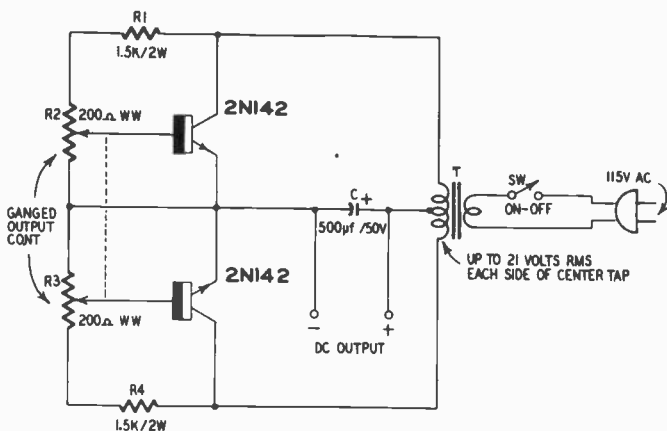


Fig. 505. Full-wave variable dc power supply.

maximum recommended voltage since the peak value of a higher voltage will exceed the maximum permissible collector-to-emitter voltage rating of the Sylvania 2N142 power transistor employed in this circuit. When a 21-0-21-volt secondary is employed, the circuit delivers 28 volts dc output.

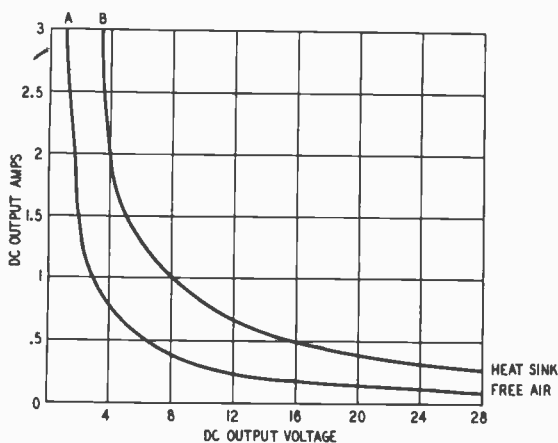


Fig. 506. Performance limits of full-wave supply.

The maximum output current which may be drawn at each dc output voltage level without exceeding the maximum power dissipation rating of the 2N142 transistor is shown by the curves in Fig.

506. Curve A shows the limits when the transistors are operated in free air and curve B the limits with heat sink.

For most efficient operation, the two transistors should be matched with respect to collector current and alpha. The base voltages are varied simultaneously by adjustment of the ganged 200-ohm dual potentiometer R2-R3. If difficulty is experienced in covering the full output voltage range from zero to maximum, the resistance values of the two limiting resistors R1 and R4 may be altered.

Light-duty dc voltage regulator

Fig. 507 shows a shunt-type voltage regulator which may be added to the output of a 10-volt rectifier-filter circuit to deliver 6 volts dc regulated within 2-5% at output currents up to 10 ma. The dc output voltage is developed across the n-p-n transistor (Texas Instruments 202).

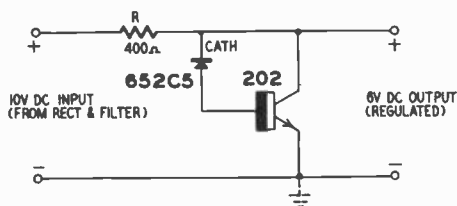


Fig. 507. Light-duty dc voltage regulator.

The control device in the circuit is the silicon junction diode (Texas Instruments 652C5) which is reverse-connected (i. e., cathode positive). Silicon junction diodes have very high back resistance (often several hundred megohms) up to a critical reverse voltage (the Zener voltage) at which the reverse current increases suddenly to a higher value at a steep slope. This is somewhat similar to the breakdown of a gaseous diode tube. The Zener point is fixed in a given diode and provides a convenient voltage reference in applications such as voltage regulators.

In Fig. 507, the 652C5 diode has a Zener voltage of 6.0 ($\pm 5\%$). Below this point, any current flowing through the diode is negligible due to the high back resistance. When the dc output voltage rises above the Zener point, however, the diode conducts, passing current through the base of the transistor. Because of the base-collector current amplification factor (beta), a larger current flows in the collector circuit of the transistor and causes a corresponding voltage drop across resistor R. This voltage drop then reduces the output voltage. Output voltage regulation thus is obtained.

The output impedance of this shunt type voltage regulator circuit is approximately 25 ohms.

Heavy-duty dc voltage regulator

The preceding voltage regulator is limited to low currents and voltages by the ratings of the conventional transistor.

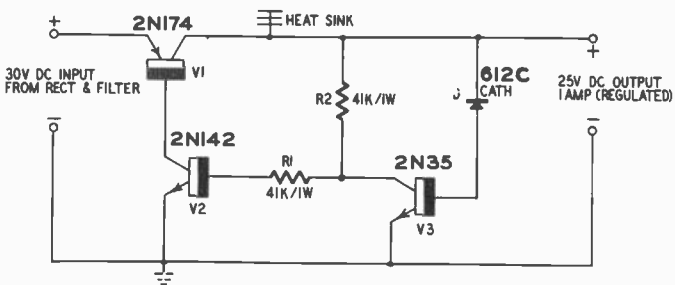


Fig. 508. Heavy-duty dc voltage regulator.

Fig. 508 shows the circuit of a voltage regulator which may be added to the output of a 30-volt rectifier-filter unit to deliver a dc output of 25 volts up to 1 ampere. This is a series-type circuit providing better than 1% regulation at an output impedance of approximately 1 ohm.

Two power transistors V1 (Delco 2N174) and V2 (Sylvania 2N142), one conventional transistor V3 (Sylvania 2N35) and one silicon junction Zener reference diode D (Texas Instruments 612C) are employed. A heat sink must be provided for transistor V1 but none is required for V2.

In this circuit, transistor V1 acts as an electronic variable resistor, introducing a controllable voltage drop in the line between the positive DC INPUT and DC OUTPUT terminals. Its resistance is controlled by varying its dc base bias. The variable bias needed for this purpose is obtained from transistor V2 which, in turn, is actuated by transistor V3 and Zener reference diode D from output-voltage changes. This is the chain through which voltage regulation is obtained. The system is sensitized by the two-stage dc feedback amplifier resulting from the cascade connection of transistors V2 and V3. A small change in dc output voltage thus is amplified sufficiently to shift the base bias of transistor V1 far enough to restore the output voltage to its original level. The 612C diode (D) is chosen for a Zener point of 25 volts.

When the 30-volt dc input is obtained from a semiconductor rectifier-filter unit (selenium, germanium or silicon), a completely tubeless heavy-duty regulated power supply results.

Light-duty constant-current adaptor

The testing of some modern electronic components (such as diodes, transistors, thermistors, varistors, tungsten filaments and other nonlinear resistors) requires a constant-current power supply. The reason for this is that in the testing (and often the operation as well) of such devices, current is the independent variable. This is the opposite condition to that of testing vacuum tubes and linear resistors where voltage is the independent variable and a constant-voltage supply is needed.

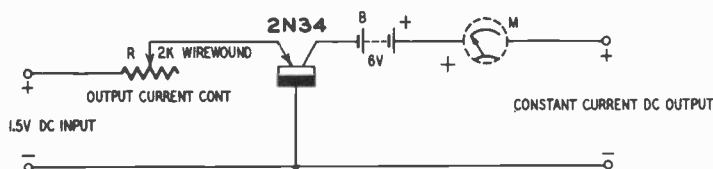


Fig. 509. *Light-duty constant-current adaptor.*

When using a constant-current supply, the load resistance can be varied over a wide range without materially disturbing the output-current level. The plate-cathode circuit of a pentode tube often is connected in series with a dc supply and load to obtain a constant current. The current regulation results from the constant-current plate characteristic of the pentode. (The E_p - I_p curve of a pentode is flat between widely separated limits, plate voltage changes having negligible effect upon plate current.) Fortunately, the collector

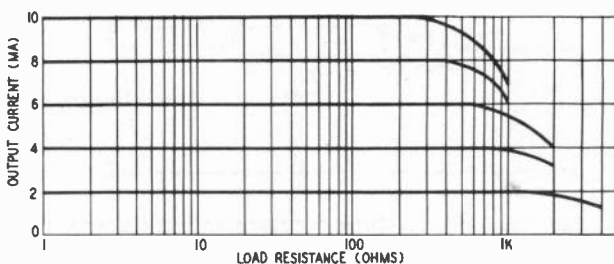


Fig. 510. *Current regulation of light-duty adaptor.*

characteristic of a common-base transistor also has the pentode-like constant-current shape, a fact which suits the transistor to use as a simple tubeless constant-current adaptor.

Fig. 509 shows how a p-n-p transistor (Sylvania 2N34) is employed to obtain constant current output from a 6-volt dc source (battery B). The output current may be set to any desired level

between zero and 10 ma (the maximum rated collector current of the 2N34) by adjusting the emitter-bias current. Emitter bias is supplied by a separate 1.5-volt source and the current level is adjusted by means of rheostat R. To read the output current value, a 0–10 dc milliammeter may be inserted in one output lead, as indicated by the symbol, M. Both the 1.5- and 6-volt potentials may be obtained from ac-operated rectifier-filter circuits, if desired.

The curves in Fig. 510 depict the current regulation afforded by the adaptor at output current levels of 2, 4, 6, 8 and 10 ma. From these plots, it is seen that the current is constant over more than three decades of load resistance values (10, 100, and 1,000 ohms) when it is 2 ma and more than two decades each (10 and 100 ohms) for higher values up to 10 ma.

To use the adaptor, first short-circuit the dc output terminals and set the current to the desired level by adjusting rheostat R. Then remove the short circuit and connect the load device to the output terminals. Refer to Fig. 510 to determine the maximum permissible value of load resistance which may be used before current regulation falls off. When testing a device requiring several test-current levels, as when plotting a family of I–E curves, vary the constant current throughout the desired range by adjustment of rheostat R.

Heavy-duty constant-current adaptor

Fig. 511 is the circuit of a constant-current adaptor operating on the same principle as the unit previously described but supplying dc output currents up to 50 ma.

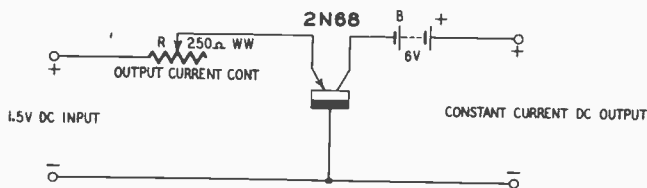


Fig. 511. Heavy-duty constant-current adaptor.

This circuit employs a p-n-p power transistor (Sylvania 2N68) with the same 1.5- and 6-volt sources as in the preceding example. These voltages may be obtained either from batteries, as shown, or from ac-operated rectifier-filter units. The OUTPUT CURRENT CONTROL rheostat (R) has been changed to 250 ohms in this circuit.

The curves in Fig. 512 depict the current regulation afforded by the heavy-duty adaptor at output current levels of 10, 20, 30, 40 and 50 ma.

The adaptor is used in the same manner as the light-duty unit and the operating instructions given for the latter may be followed.

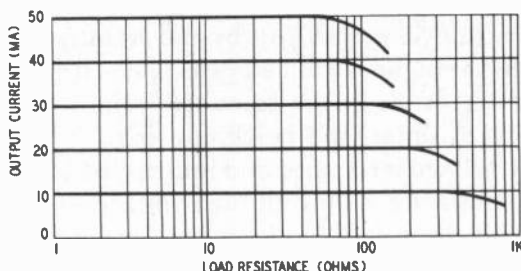


Fig. 512. Current regulation of heavy-duty adaptor.

High-voltage dc-to-dc power supply

Fig. 513 is the circuit of a compact supply operated from a small 6-volt battery and delivering from 900 to 1,000 volts dc at 5 to 10 μ a. This unit is convenient for operating a GM tube, such as Victoreen Type 1B85, in Geiger counters. It is useful also in other low-current-high-voltage applications such as insulation testing, electrostatic charging, photoflash operation, oscilloscope anode supply and photomultiplier tube supply. The low operating current requirement (not exceeding 2 ma at 6 volts) permits the use of four series-connected penlight cells when extremely small size and light weight are demanded.

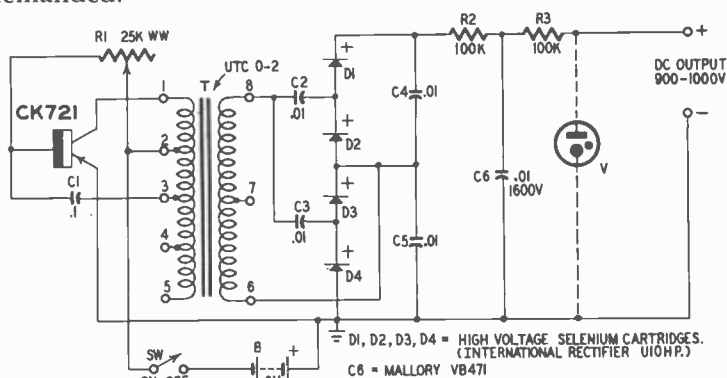


Fig. 513. High-voltage dc-to-dc power supply.

The basis of the circuit is a Hartley type oscillator employing a high-alpha p-n-p transistor (Raytheon CK721). A portion of the primary winding (taps 1, 2 and 3) of a subminiature "ouncer" transformer (UTC type O-2) forms the split tank coil of this oscil-

ator. (The use of this particular transformer and the oscillator circuit configuration have been suggested by Thomas G. Knight.) The ac voltage developed across the primary (between taps 1 and 3) by the oscillator action is stepped up by the transformer and a high voltage (somewhat higher than 300 volts open-circuit) consequently is available across the entire secondary winding (available between taps 6 and 8, center tap 7 being unused).

The high ac voltage is rectified and multiplied by a voltage-quadrupler circuit consisting of the four high-voltage selenium cartridges (D1, D2, D3, D4: International Rectifier Corp. type U10HP) and the four 0.01- μ f capacitors (C2, C3, C4, C5). Filtering is provided by resistors R2 and R3 and capacitor C6. This capacitor must have a working voltage of 1,600. A satisfactory small component is the Mallory VB471 suppressor capacitor. Capacitors C2, C3, C4 and C5 must be of the mica type.

When regulation of the dc output voltage is desired, a simple method is to connect, a miniature, filamentless, corona-type regulator tube (V) such as Raytheon CK103B (900 volts), across the output terminals, as shown by the dotted symbol. A series string of small neon lamps, such as type NE2, also can be used.

Fig. 513 shows the negative output terminal grounded. When a positive ground is desired, reverse each of the selenium cartridges (D1 to D4) and regulator V.

The initial adjustment of the unit is simple: Connect a dc vacuum-tube voltmeter (set to its 1,000- or 1,500-volt range) to the DC OUTPUT terminals. Close switch SW and adjust rheostat R1 for a dc output (as indicated by the vtvm) between 900 and 1,000 volts. Open the switch and close it again. If the output voltage does not reappear immediately after closing the switch, readjust R1 for quick starting of the oscillator (quick buildup of dc output voltage).

radio receivers

IN radio receivers, transistors have been a boon to miniaturization, making possible for the first time, for example, a practical superheterodyne receiver small enough to fit into a coat pocket. In large receivers, as well as in small ones, the low current requirements of transistors have made battery operation both economical and feasible. Less expensive batteries and long battery life characterize transistor radios as contrasted to tube-type portable sets.

The chief interest of many experimenters with transistors is building and operating receivers. To pick up broadcasts at loud-speaker volume with a miniature set containing only filamentless crystal devices has a constant exhilaration.

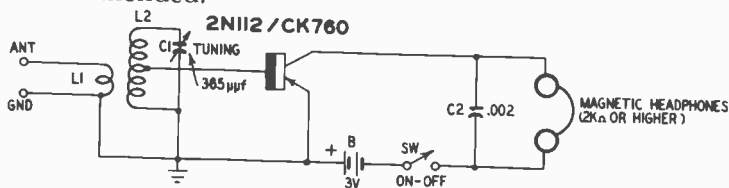
During the early years of the transistor, engineers and experimenters alike tested the transistor in all types of receivers. Substantial progress was delayed until high-frequency commercial transistors and suitable miniature *if* and class-B *af* transformers to match transistor impedances appeared on the market. The first factory-built transistorized superhet receiver was placed on sale during the transistor's sixth year of existence. A commercial shortwave converter did not appear until its eighth year.

This chapter gives the practical details of a number of receiver circuits. A cross-section has been made of the supply of good circuits calculated to fill the need of experimenters, hobbyists, and others.

Single-transistor broadcast receiver

When high sensitivity is not required, as in picking up strong

local stations, a simple one-transistor receiver circuit will give good headphone operation. For maximum performance, an rf transistor is recommended.



L1 = 25 TURNS NO. 32 ENAMELED WIRE CLOSEWOUND OVER GROUND END OF L2.

L2 = 144 TURNS NO. 32 ENAMELED WIRE CLOSEWOUND ON 1" DIAMETER FORM, TAP BOTH TURN FROM GROUND END.

Fig. 601. Single-transistor broadcast receiver.

Fig. 601 shows a single-transistor receiver circuit employing a Raytheon 2N112/CK760 transistor and 3-volt battery. The base connection for the transistor is tapped down the secondary coil L2 of the rf input transformer for impedance matching and improved selectivity. The collector output circuit feeds the headphones directly. Because the headphones are connected in series with battery B and the collector electrode, only the magnetic type may be used. Since crystal headphones block the passage of dc, they are unsatisfactory for use in this circuit.

The rf input transformer consists of a 144-turn tapped secondary winding and a 25-turn primary. The coils are wound on a 1-inch diameter plastic or fiber tube 1.5 inches long. Begin the construction by closewinding 144 turns of No. 32 enameled wire on the tube. Tap the 80th turn from the end of the coil to be grounded. This completes the secondary L2. Next, wind one turn of $\frac{3}{4}$ -inch (width) Scotch tape tightly around the ground end of the secondary coil. This provides insulation for the primary winding. Finally, closewind 25 turns of No. 32 enameled wire on top of the tape in the same direction as the secondary coil. This completes winding L1.

The single-section 365- μ f variable capacitor C1 provides tuning throughout the standard broadcast band. When the receiver is to be made very compact, a miniature flat-type tuning capacitor, such as Lafayette MS-215 or MS-274, may be used.

The receiver gives best results with a good outside antenna and earth ground. However, it will operate with various emergency antennas, either indoor or outdoor types (with or without a ground connection) but the headphone volume will depend directly upon the nearness and power of the station received.

The transistor detector is superior to a diode alone since the tran-

sistor provides amplification as well as detection of the signal. The low drain of $4 \mu\text{a}$ permits use of the smallest-size penlight cells.

Two-transistor broadcast receiver

In the receiver circuit shown in Fig. 602, a single-stage transformer-coupled af amplifier employing an inexpensive p-n-p transistor (Raytheon CK722) has been added to the simple detector circuit of Fig. 601. The tuner (L1-L2-C1) has the same specifications as in Fig. 601.

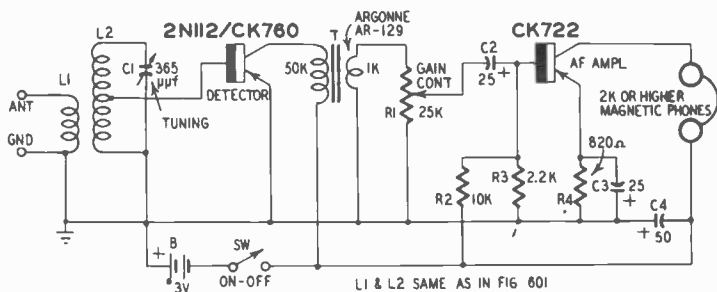


Fig. 602. Two-transistor broadcast receiver.

This receiver gives loud headphone signals when tuned to strong local broadcast stations. Because of the increased gain provided by the audio amplifier, performance with poor antennas is better than when using the one-transistor receiver.

An audio gain control R1 is included since the headphone volume can be excessive on strong nearby stations. Dc base-bias stabilization is provided by the voltage-divider network R2-R3 and the bypassed emitter resistor R4.

The total dc drain from the 3-volt battery is approximately 1 ma.

Miniaturization of simple broadcast tuners

A vestpocket receiver requires a tuner (rf input transformer and tuning capacitor) smaller in size than the L1-L2-C1 combination shown in Figs. 601 and 602.

A miniature unit may be constructed, as shown in Fig. 603, by substituting a ferrite-tuned antenna coil (L) for the original L1-L2 transformer and a flat-type miniature $365\text{-}\mu\text{f}$ variable capacitor C, such as Lafayette MS-215 or MS-274, for the standard-size variable. The rest of the receiver circuit remains the same as in Fig. 601 or 602.

The tuning slug of the coil may be set for full coverage of the broadcast band with the antenna in use. While this tuner arrange-

ment is not as selective as those shown in Figs. 601 and 602, due to poor impedance match to the transistor and absence of the primary

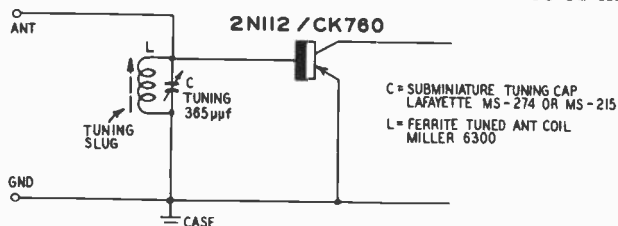


Fig. 603. Modification of detector for miniaturization.

coil, it permits considerable size reduction and fair performance.

The length of the antenna will affect the tuning range provided by capacitor C, but this can be compensated in most instances by a readjustment of the screw-tuned slug of coil L.

Regenerative broadcast receiver

Fig. 604 shows the circuit of a simple single-circuit regenerative broadcast receiver employing a p-n-p rf transistor (Raytheon 2N-112/CK760). The sensitivity of this detector circuit is increased by the rf amplifying properties of the regeneration.

A tickler-type feedback circuit is used, output energy being fed back by the tickler coil L2 to the input tuned circuit L1-C2 by inductive coupling. Adjustment of rheostat R controls the amount of regeneration.

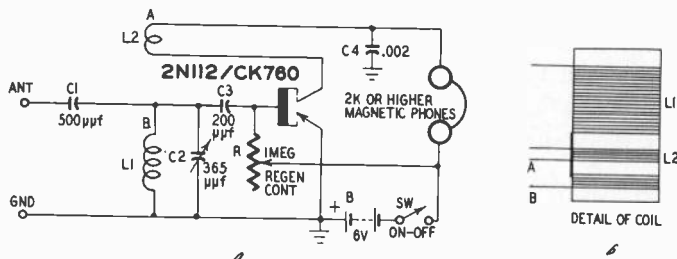


Fig. 604-a, -b. Regenerative broadcast receiver.

The tuner coil assembly L1-L2 is wound on a 1-inch diameter plastic or fiber tube, 1.5 inches long. The input coil L1 is wound on the tube first and consists of 144 closewound turns of No. 32 enameled wire. Then a single turn of $\frac{3}{4}$ -inch-wide Scotch tape is wound tightly around one end of the finished coil for insulation of the tickler. The tickler coil L2 next is wound on top of the tape in the same direction as L1. It consists of 40 closewound turns of No. 32 enameled wire.

For the feedback to be of the proper polarity for regeneration, the coils must be phased correctly. Fig. 604-b shows the top of winding L1 labeled B and the top of L2 labeled A. These ends of the coils are connected in the circuit to the headphones and C4 and to the C1-L1-C2-C3 junction, respectively, as shown in Fig. 604-a.

As in a tube-type regenerative detector, the regeneration level should be adjusted to the highest point short of oscillation. The receiver has maximum sensitivity to weak signals in this condition but slightly beyond this point the circuit will break into annoying oscillation.

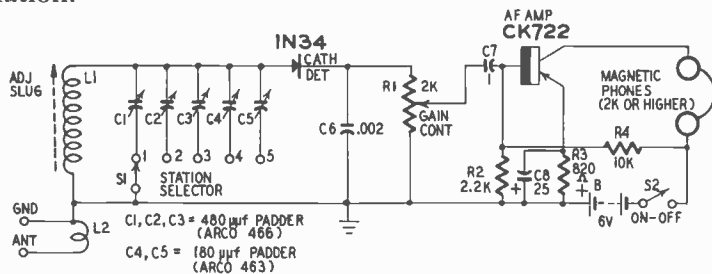


Fig. 605. Step-tuned diode-transistor broadcast receiver.

Step-tuned diode-transistor broadcast receiver

When only a few stations are located in the service area of a receiver, continuously variable tuning is not needed. Instead, it is convenient for the receiver circuit to be pretuned to the various station frequencies and the operator then can select these stations at will by means of a switch or pushbuttons.

Fig. 605 shows a simple step-type circuit employing a diode detector (1N34) and transistor audio amplifier (Raytheon CK722). The input rf transformer (L1-L2) is pretuned to five carrier frequencies by trimmer capacitors C1, C2, C3, C4 and C5, respectively. Tuning to either desired station then is accomplished by setting switch S1 to connect the corresponding trimmer capacitor across coil L1. The reader may provide as many trimmer capacitors as needed for the number of stations to be tuned in, as long as the selector switch (S1) has sufficient positions.

The rf transformer consists of a ferrite-tuned antenna coil L1 (Miller 6300) around the lower end of which are wound five turns of No. 24 dcc wire (L2). Trimmer capacitors C1, C2, and C3 are used for frequencies lower than 1,000 kc and each is a 480- μ f screw-adjusted padder (Arco 466). Trimmers C4 and C5 are for frequencies higher than 1,000 kc and each is 180 μ f (Arco 463). This

switched-tuner diode detector is based upon an earlier design by the author.¹

The receiver may be adjusted and aligned initially from received stations or with the aid of an amplitude-modulated signal generator. The signal is applied to the ANTENNA and GROUND terminals, switch S1 set to position 1 and the lowest-frequency signal tuned in by adjusting C1 and the slug of L1. Next, switch S1 is set to position 5 and the highest-frequency signal tuned in by adjusting trimmer C5. If this frequency cannot be reached, readjust the slug of L1, then return switch S1 to position 1 and readjust C1. The intermediate frequencies then may be tuned in successively with switch S1 at its corresponding settings and by adjusting the corresponding trimmer capacitors. However, do not disturb the L1 slug adjustment once the initial point 1 and 5 alignments have been completed.

Push-pull transistor broadcast receiver

Fig. 606 shows the circuit of a push-pull transistor detector circuit based upon a similar diode detector circuit developed earlier by the author.² This detector has increased sensitivity and output over the single-ended type and provides a single-stage broadcast receiver for efficient operation of headphones.

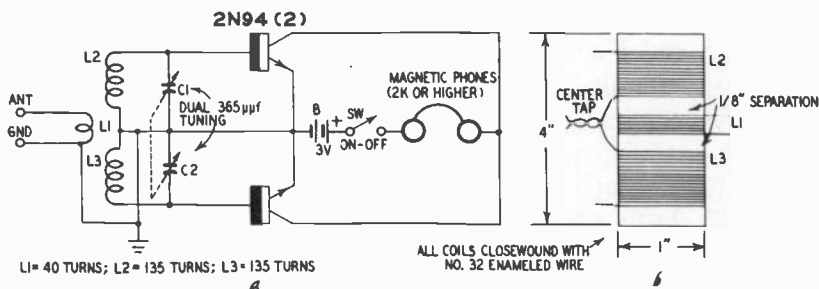


Fig. 606-a,-b. Push-pull transistor broadcast receiver.

Employing two rf transistors (Sylvania 2N94), this circuit has split-tuned input. The two halves of the secondary winding (L2-L3) of the input transformer are tuned by the two-section 365-μf variable capacitor C1-C2. The primary winding L1 is wound between the halves of the secondary. Fig. 606-b gives the constructional details of this transformer.

This receiver circuit is of special interest when two transistors are available but the components for an audio amplifier stage either are not available or cannot be accommodated in the layout.

Loop-operated diode-transistor broadcast receiver³

Although a diode detector is employed in the receiver circuit shown in Fig. 607, this set has high sensitivity due to the gain of the audio amplifier following the diode. It is high enough to permit use of a ferrite rod-type self-contained loop antenna and the audio output is sufficient to operate a loudspeaker. This arrangement makes an excellent, small-sized, portable receiver for local reception.

The circuit employs a diode detector (Raytheon CK705), high-alpha transistors (Raytheon CK721) in the af amplifier and driver stages and medium-alpha transistors (Raytheon CK722) in the push-pull class-B audio output stage. The audio power output is 110 milliwatts. Four size-D flashlight cells connected in series will supply the direct-current drain for more than 1 month of all-day operation.

The transistor-type ferrite-rod antenna (L) has a very high Q and is provided with a tap for matching the relatively low impedance of the CK705 diode. These features boost the sensitivity and selectivity of the detector circuit. The recommended Miller No. 2000 antenna is a long unit (approximately 9 inches). Shorter rods

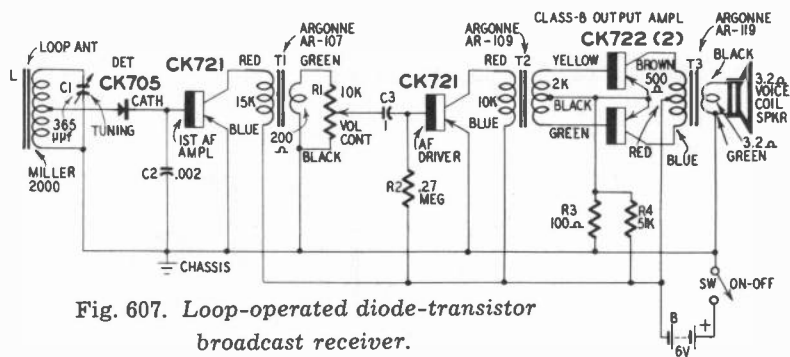


Fig. 607. Loop-operated diode-transistor broadcast receiver.

will not provide sufficient signal pickup. The antenna must be mounted horizontally and clear of metallic objects. The receiver therefore cannot be housed in a metal case, although it can be built on a metal chassis. The antenna is directional in operation, giving best pickup when it is rotated broadside to the station.

The transformers must be connected correctly according to the color coding indicated in Fig. 607. If this is not done, oscillation might occur as a result of feedback due to magnetic coupling between their cores. To minimize coupling, the transformers (when they must be mounted closely together in a compact layout) should be oriented so that their cores are at right angles.

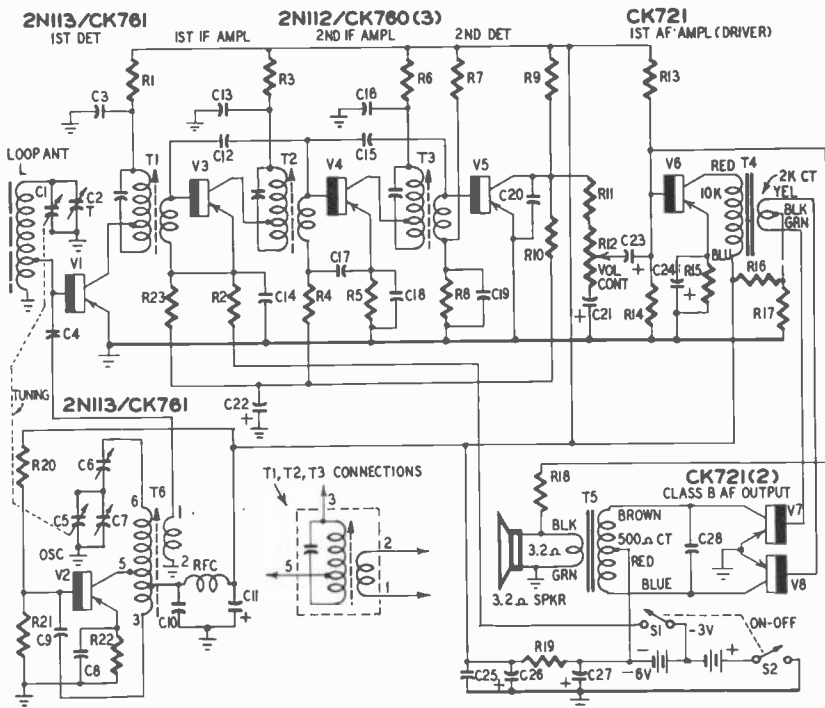


Fig. 609. Superheterodyne broadcast receiver.

Superheterodyne broadcast receiver

Fig. 609 shows the circuit of a conventional portable broadcast receiver transistorized throughout. This circuit, credited to Hazel

Parts list for portable broadcast receiver

Capacitors: C1-C5, dual 365- μmf variable; C2, built-in trimmer on section C1 of tuning capacitor; C3, C13, C14, C16, C17, C18, C19, .1- μf tubular; C4, 100- μmf ceramic; C5, see C1; C6, 1250- μmf padder (Arco 309-M); C7, built-in trimmer on section C5 of tuning capacitor; C8, C9, .001- μf ceramic; C10, C25, .01- μf tubular; C11, C21, C23, C24, C26, C27, 25- μmf electrolytic; C12, C15, 30- μmf ceramic; C20, C28, .05- μf tubular; C22, 10- μmf electrolytic.

Resistors: R1, R2, R3, R5, R6, R8, R9, 1000 ohms; R4, 150,000 ohms; R7, 39,000 ohms; R10, 56,000 ohms; R11, 220 ohms; R12, 25,000-ohm potentiometer; R13, 10,000 ohms; R14, 2200 ohms; R15, 820 ohms; R16, 5100 ohms; R17, 120 ohms; R18, 18,000 ohms; R19, 100 ohms, 1 watt; R20, 3900 ohms; R21, 27,000 ohms; R22, 1800 ohms; R23, 100,000 ohms.

Transformers and coils: L, transistor ferrite-loop antenna (Miller 2000); RFC, 2½-millihenry rf choke; T1, T2, T3, 455-kc if transformer (Miller 2041); T4, transistor class-B driver transformer (Argonne AR-109); T5, transistor class-B output transformer (Argonne AR-119); T6, transistor oscillator coil (Miller 2020).

Transistors: V1, V2, Rf transistors (Raytheon 2N113/CK761); V3, V4, V5, If transistors (Raytheon 2N112/CK760); V6, V7, V8, Af transistors (Raytheon CK721).

Miscellaneous: B, 6-volt battery tapped at 3 volts (see text); S1-S2, dpst switch.

tine, was used in the early Raytheon transistorized radio and is a

strong favorite among experimenters. The author has adapted it to readily available components.

Raytheon p-n-p transistors are employed in all channels. Transistors having 10-mc alpha cutoff (2N113/CK761) are used in the 1st detector and oscillator stages, V1 and V2, respectively. In the if amplifier and second detector (V3, V4 and V5) 5-mc units (2N112/CK760) are employed and high-alpha low-frequency units (CK721) in the first af amplifier (driver) and push-pull class-B af output stages (V6, V7 and V8).

The two-stage if amplifier is tuned to 455 kc. A tap on the primary winding of each if transformer (T1, T2, T3) is provided to match the 25,000-ohm output impedance of the transistor. The secondaries match the 600-ohm base-input impedance. To prevent oscillation in the if channel, the stages of this amplifier are neutralized automatically by fixed capacitors C12 and C15. The if channel has an overall power gain of 60 db. Negative dc bias (for the bases of the two if amplifier transistors) is obtained as avc voltage from the second detector V5 through the avc filter R10-C22-R23.

The second detector is a power type circuit providing a gain of approximately 10 db. This stage is coupled, through the volume-control potentiometer R12 and capacitor C23, to the af driver V6.

The push-pull class-B af output stage V7-V8 delivers 110 to 125 mw to a 3.2-ohm loudspeaker voice coil. Negative feedback for improving fidelity is supplied through resistor R18. It is imperative that the transformer connections follow the color coding indicated in Fig. 609, otherwise the feedback will not be of the proper phase for degeneration. Matching of transistors V7 and V8 with respect to collector current and alpha will reduce distortion. Fixed base-bias voltage supplied by the voltage divider R16-R17 minimizes cross-over distortion. Decoupling to prevent motorboating is provided by the filter C26-R19-C27.

Oscillator V2 is a Hartley-type circuit. The oscillator signal is injected through capacitor C4 into the base of the first detector transistor V1.

A high-Q ferrite-rod type loop antenna (L) is employed. Tuning is accomplished with a two-gang 365- μ f variable capacitor C1-C5, one section of which tunes the antenna while the other section tunes the primary of the oscillator transformer T6.

When the receiver will be used for private reception and loudspeaker operation is not desired, 2,000-ohm headphones or a 2,000-ohm hearing-aid type earpiece may be connected in place of the primary of transformer T4. In addition to the loudspeaker, the fol-

lowing components may then be omitted: C28, R16, R17, R18, T4, T5, V7 and V8.

The numbers appearing on the terminals of the if transformers (T1-T2-T3) and oscillator coil assembly (T6) in Fig. 609 correspond to the manufacturer's labeling and must be followed strictly.

The receiver is powered by a 6-volt battery consisting of four size-D flashlight cells connected in series. A tap is taken at the -3-volt point. The total zero-signal dc drain of the receiver is 10 ma and the total maximum-signal current 34 ma. This low drain means a battery life of several hundred hours.

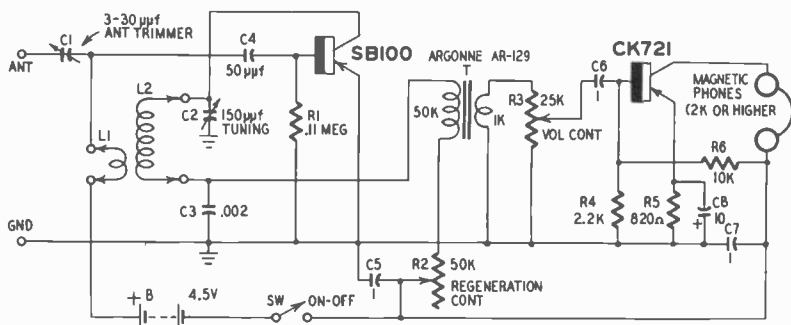


Fig. 610. All-wave regenerative receiver.

All-wave regenerative receiver

The high-frequency transistor makes a wide-range short-wave regenerative receiver entirely feasible. Fig. 610 shows the circuit of a set of this type.

Using a Philco SB100 surface-barrier transistor as the regenerative detector, this receiver has a general-coverage tuning range of

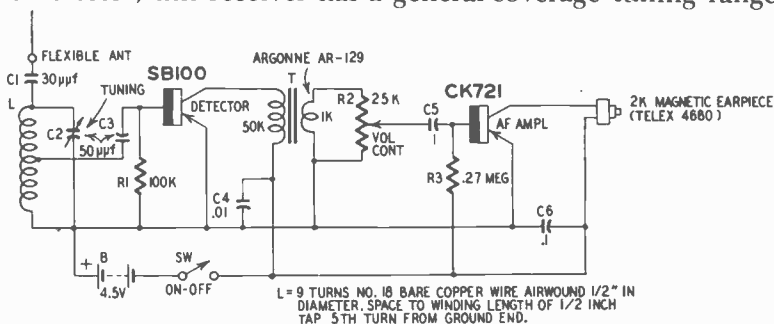


Fig. 611. High-frequency cueing receiver.

1,000 kc to 44 mc in four bands: 1-2.7 mc, 2.5-7 mc, 6-17 mc and 15-44 mc. The circuit in Fig. 610 accommodates plug-in coils; however, the coils may be mounted permanently in the set with a

two-gang rotary selector switch to provide a frequency-switching (band-switching) arrangement. Specifications for the coil combinations (L1-L2) may be found in the oscillator coil table (Table 4) in Chapter 4. The reader can alter the tuned-circuit constants for coverage only of restricted ranges such as the amateur bands.

A tickler type feedback circuit is employed. L1 is the low-impedance base tickler winding while L2 is the higher-impedance tuned-collector winding. Regeneration control is secured by varying the collector current with rheostat R2.

A single transformer-coupled af amplifier stage employing a p-n-p transistor (Raytheon CK721) gives headphone operation at approximately 2 mw. The operating point of this amplifier is stabilized by fixed dc bias (supplied by the voltage divider R4-R6) and the bypassed emitter resistor R5.

In operation, the 3-30- μ f compression type trimmer capacitor C1 must be set for maximum signal response with minimum disturbing effect upon the tuned circuit, with the antenna in use.

Total direct-current drain from the 4.5-volt battery B is approximately 2.5 ma.

High-frequency cueing receiver

Fig. 611 shows the circuit of a cueing receiver, which takes its name from the fact that it may be worn on the person of an actor to receive cues from the director or stage manager. The pickup antenna is stitched into the costume or pinned inside a sleeve. It consists of a length of flexible insulated wire.

Operating an earpiece of the hearing-aid type, this receiver is designed to pick up signals from a nearby crystal-controlled radio-phone transmitter operating on a carrier frequency of 27.255 mc. The transmitter can be quite simple and, because of its proximity, the receiver need not have high sensitivity nor the transmitter high power. The transmitter may be a small hand-carried unit placed in the wings of the stage or in the prompter's dugout.

The detector employs a Philco SB100 surface-barrier transistor; the af amplifier a Raytheon CK721. Total dc drain from the 4.5-volt battery B (which can consist of three of the smallest-size penlight cells connected in series) is approximately 1.2 ma. Small mercury cells may be used for stable voltage output throughout long life.

The shaft of the tuning capacitor C2, may be slotted for screw-driver adjustment since this unit does not need frequent retuning to the single transmitter frequency. A small knob should be attached to the volume control potentiometer shaft R2 however, since this control needs occasional readjustment.

triggers and switches

MANY of the modern applications of electronics utilize pulse techniques. In contradistinction to early communication-type circuits which handled signals having only sinusoidal or near-sinusoidal waveforms (although often of varying amplitude or frequency), pulse circuits are characterized by abrupt on-off (switching) operation and frequently depend upon the phase relation between two or more signal pulses to secure a desired result. These techniques are employed in computers, counters, radar, television, automation, industrial electronics and in some test instruments.

Some pulse-type equipment employs large numbers of vacuum tubes with the attendant problems of size, power demand, heat removal, circuit complexity and unreliability. Considerable attention accordingly has been directed toward the transistor—a small-sized, rugged and dependable component requiring only tiny amounts of operating power, generating negligible heat, endowed with long life and providing high power gain.

Much of the transistorized pulse circuitry which has been developed is highly specialized and without wide general interest, especially to the hobbyist. The few circuits which appear in this chapter have been selected as representative building blocks illustrating the transistorization of well-known pulse systems and having some application value to general experimenters.

Flip-flop

Fig. 701 shows the circuit of a transistorized flip-flop comparable to the Eccles-Jordan vacuum-tube circuit. Like its counterpart,

this is a bistable *on-off* circuit delivering one output pulse for each pair of successive input pulses. This feature suits the flip-flop as a basic scale-of-2 counter or frequency halver. For fast rise and fall times in the output pulse, rf transistors (Sylvania 2N94A) having 6-mc cutoff are employed.

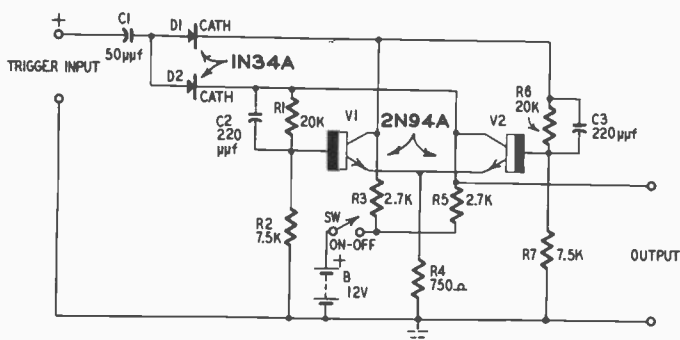


Fig. 701. Flip-flop circuit.

When the circuit is operating, either V1 or V2 is conducting collector current while the opposite transistor is cut off. Thus, a pulse is delivered to the OUTPUT terminals only when V2 conducts. A second, separate output terminal could, of course, be connected to the collector of V1. Output pulses then would be delivered alternately by the two pairs of output terminals as conduction shifted between V1 and V2.

Conduction is switched from one transistor to the other by means of a positive trigger pulse applied to the TRIGGER INPUT terminals. Application of this pulse causes the *off* transistor to switch on and the *on* transistor to switch off.

Operation of the circuit is explained in the following manner: When switch SW is closed, both transistors will start to draw collector current through collector load resistors R3 and R5, respectively. However, because of random transients (perhaps due to noise components in currents), the collector current of one transistor will increase, although this action may be only momentary. If this occurs in V1, for example, the increased collector current will produce a rise in the voltage drop across R3 and this will lower the V1 collector voltage. This change is coupled across to the base of V2 through the voltage divider R6-R7 and this action lowers the V2 base voltage. As a result, the collector current of V2 decreases and this causes the collector voltage of V2 to increase. This increase is coupled through voltage divider R1-R2 to the base of V1 and acts to increase the V1 collector current still more. The action now

continues rapidly in this direction until V1 is conducting heavily (low collector voltage) and V2 is cut off (high collector voltage). The action is completed in a rapid *flip*. This is one of the stable states of the circuit and will be preserved until switch SW is opened momentarily or conduction is shifted from V1 to V2 by an appropriate trigger pulse.

Now, a positive trigger pulse may be applied through capacitor C1 and the two "steering" diodes D1 and D2 to the collectors. The diodes are so poled that they allow easy passage of the positive trigger while preventing a short-circuit of the two collectors. The positive trigger voltage can have little or no effect upon V1 because the collector current of this transistor already is high and will undergo only a negligible change in response to a collector-voltage shift. But the trigger will lower the collector voltage of V2 momentarily. This voltage change is coupled through voltage divider R1-R2 to the base of V1. The effect of this action is to lower the V1 collector current. Once initiated, the action becomes cumulative, continuing rapidly until V1 is cut off and V2 conducting. This is the other stable state of the circuit.

The circuit derives its name from the fact that its conducting and nonconducting states are attained in rapid flips from conduction to cutoff, not by a smooth variation between zero and maximum current flow. This on-off action is, of course, true switching.

The function of the *commutating* (or "speedup") capacitors C2 and C3 is to transmit high-frequency trigger pulses directly from one collector to the opposite base to accelerate the initiation of switching.

At repetition rates up to 200 kc, the output-pulse waveform essentially is rectangular with a peak-to-peak amplitude of approximately 10 volts. A rise time of 1 microsecond (μsec) and fall time of 0.5 μsec may be achieved, with a pulse width of approximately 2.5 μsec . The trigger pulse must be positive-going with a peak amplitude of 10 volts maximum and rapid rise and fall times.

One-shot multivibrator

The one-shot multivibrator, also called a *univibrator*, differs from the conventional multivibrator (which is free-running) and the flip-flop (which is bistable) in that it is a monostable circuit. That is, the univibrator may be pulsed into operation but, after delivering one output pulse, it reverts to its zero-output resting state after the actuating pulse has passed.

Thus, the one-shot multivibrator delivers an output-signal pulse

each time it is triggered into operation by an input-signal pulse. The duration (width) of the output pulse is governed by the resistance and capacitance values in the circuit.

Fig. 702 shows the circuit of a one-shot multivibrator employing two rf transistors (Sylvania 2N94A). Like the flip-flop (Fig. 701), this circuit employs emitter feedback, obtained through the use of a common emitter resistor R4. An isolating diode (Sylvania 1N38A) is provided for the input trigger pulses.

When the circuit is in its quiescent state, transistor V2 conducts comparatively heavy collector current because of the connection of its base to the positive terminal of the dc supply through series resistor R5. This current flows through the common emitter resistor R4 and the resulting voltage drop developed across this resistor biases transistor V1 to cutoff. V1 therefore is *off* and V2 *on*. Capacitor C2 is in a charged state.

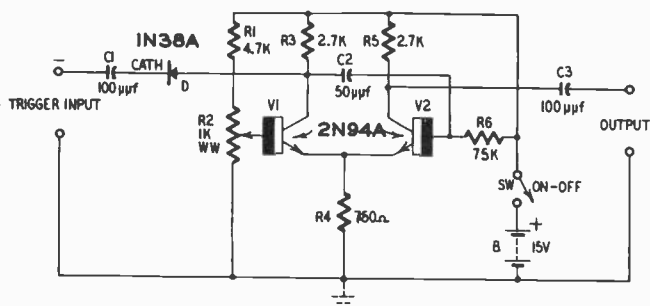


Fig. 702. One-shot multivibrator (univibrator).

A negative input pulse applied to the TRIGGER INPUT terminals reduces the positive potential on the base of transistor V2 and discharges capacitor C2. This reduces the collector current of V2 and lowers the voltage drop across resistor R4. Since the emitter bias of V1 accordingly is lowered, this transistor begins to pass collector current. The transition is rapid, V1 switching on and V2 off and the circuit delivering an output pulse.

Immediately after this switching operation, capacitor C2 begins to discharge and as it does so the positive voltage on the base of V2 begins to rise once more toward the potential of the supply, B. At the end of this discharge interval, V2 again is conducting (on) and V1 nonconducting (off), the quiescent condition of the circuit.

The limiting factor, which sets the length of the discharge time of C2 and therefore the duration of the output pulse, is the voltage drop across R4. Full conduction is not restored to V2 until the base

voltage of this transistor equals the emitter voltage; that is, the voltage drop across R4. When V1 is conducting and V2 is cut off, the voltage drop across R4 is governed by the setting of potentiometer R2 (which determines the V1 dc base bias). This variable resistor thus affords a means for setting the duration (width) of the output pulse.

Pulse delay circuit

The circuit in Fig. 703 delivers a negative-going output pulse, the occurrence of which is coincident to or delayed with respect to a negative input-trigger pulse. The delay interval is continuously variable. This circuit employs two rf transistors (Sylvania 2N94A), a germanium diode D1 (Sylvania 1N38A) and a bonded silicon diode D2 (Raytheon 1N300).

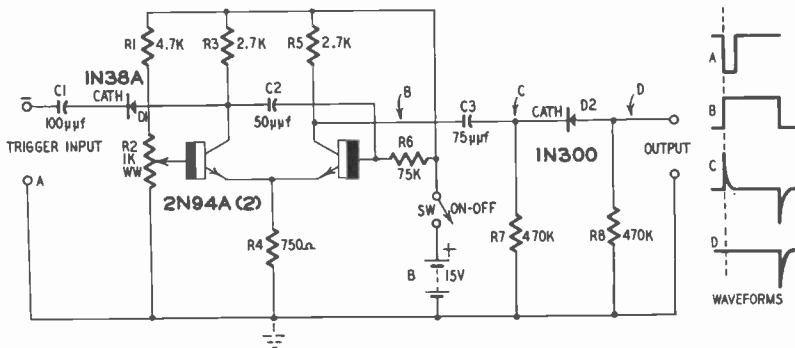


Fig. 703. Pulse delay circuit.

The first part of this circuit is the one-shot multivibrator previously described. Its purpose is to deliver an adjustable-width output pulse (waveform B) in response to a negative input-trigger pulse (waveform A). The second part is a resistance-capacitance differentiator (C3-R7) which produces sharp, narrow, positive and negative pulses (waveform C) corresponding in time spacing to the rise and fall of the one-shot output pulse (waveform B). The third part is a series-diode rectifier circuit (D2-R8) the purpose of which is to transmit the negative pulse while blocking the positive pulse of the differentiator output. The negative pulse is selected because the time position (delay) of this pulse may be varied by adjustment of the duration of the one-shot output pulse. The time position of the positive pulse, on the contrary, always coincides with the rise of the one-shot output.

The delay time is varied by adjusting potentiometer R2, the

primary function of which is to vary the duration of the output pulse (waveform B) of the one-shot multivibrator section.

It is necessary to employ a modern silicon diode D2 in the rectifier section of the circuit because of the high input and output resistances, R7 and R8 respectively. These high resistance values are necessary to minimize loading of the preceding transistor output circuit. The back resistance of the diode must be very high with respect to the total resistance of R7 and R8. In germanium and early point-contact silicon diodes, the back resistance is so low that rectification ceases when the load resistance is as high as that shown here. The 1N300 diode has a resistance of 10,000 megohms at -10 volts. This rating yields a diode-to-load resistance ratio of better than 10,000 to 1, with respect to R7 + R8.

Simple coincidence circuit

In the circuit shown in Fig. 704, the collector supply voltage of the transistor (Sylvania 2N35) is a rectangular or square pulse having a positive peak amplitude of 6 volts. This pulse voltage is applied to terminals C and D and is furnished by a source having adjustable pulse duration, amplitude, and repetition rate.

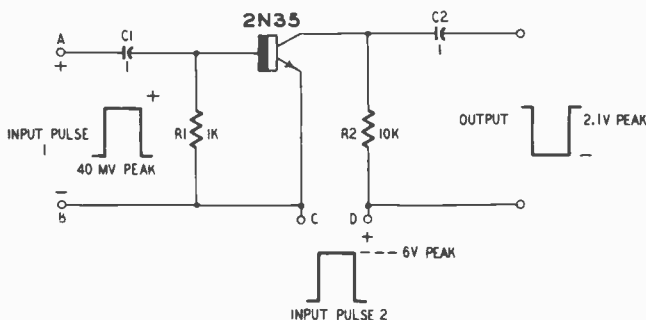


Fig. 704. Simple coincidence circuit.

The circuit may be said normally to have no output in the absence of an input-signal pulse. Actually, the output is never completely zero since a tiny amount of current (due to the i_{co} characteristic of the transistor) flows through the collector circuit when the supply voltage (input pulse 2) is at its peak value. But since the resulting voltage drop across the collector load resistor R2 is so minute (usually of the order of 50 db below maximum voltage output), the output may be regarded as zero.

When a positive pulse (input pulse 1) is applied to the base of

the transistor, through input terminals A and B, at the same time that the collector is positive, the circuit delivers an amplified output signal. If input pulse 2 is zero at the same time that input pulse 1 arrives, the amplifier delivers no output. (Actually, there is a tiny output, regarded as noise, due to R-C leakage through the transistor).

Input pulse 2 thus acts to switch on the circuit for transmission and amplification of input pulse 1. The number of input-1 pulses transmitted by the circuit is governed by the width of input pulse 2 which, figuratively, holds open the gate for input-1 pulses. When input pulse 1 is wider than input pulse 2, only that portion of the former is transmitted which occurs during the duration of the latter. Various desired time-delay actions may be worked out by delaying input pulse 1 with respect to a comparatively wide input pulse 2.

This circuit takes its name from the fact that each output pulse indicates a coincidence between the two input pulses. If the latter are out of step at all times, no output pulses are obtained. When each input pulse occurs at a random rate, the output pulses indicate the number of times they happen to fall into step. The circuit thus serves as the basis for a coincidence counter.

Pulse inverter

Switching-circuit applications sometimes require that a given pulse be reversed in polarity. Depending upon the end use, the operation goes under different names such as *inverting*, *priming*, *negating*, etc.

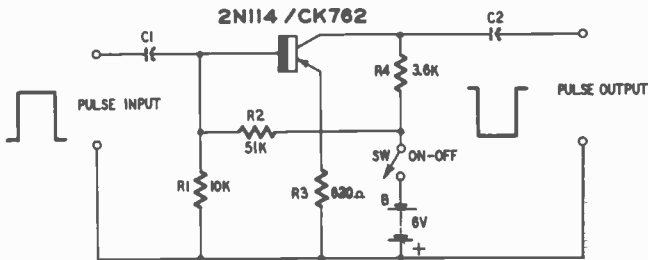


Fig. 705. Simple pulse inverter.

Fig. 705 shows a simple circuit in which inverting occurs as a result of the inherent phase-reversing property of the common-emitter transistor amplifier circuit. Employing a p-n-p type rf transistor (Raytheon 2N114/CK762) for fast rise and fall in the output waveform, the circuit is direct-coupled and is degenerative (unbypassed emitter resistor R3) for improved response. How-

ever, isolating capacitors (C1 and C2) will be required in those applications in which the dc biases of the transistor circuit must be protected from short circuit by external circuitry connected to the PULSE-INPUT and PULSE-OUTPUT terminals.

control devices

THE high sensitivity of transistor circuits to both ac and dc signals is an aid in the design of control equipment. This sensitivity provides high control efficiencies. The power economy of the transistor makes reliable and practical the unattended service of such equipment. Transistorized control devices may be operated from inexpensive batteries for long periods of time that often equal the shelf life of the battery. This ability divorces the equipment from power-line failures and allows compact, self-contained construction. Even when transistorized devices are operated from the power line, the current drain (and consequently the operating cost, heat generated, size, and complexity) is considerably lower than with tube-type equipment.

The light weight and small size of the transistor, together with its operation from miniature batteries and in conjunction with miniature circuit components, suit the transistor naturally to circuits for model control. The high collector current of power transistors permits these units to be used directly as nonmechanical relays in some applications.

Practical control circuits are described in this chapter. These circuits may be used by themselves or in combination with other devices. While particular applications are specified for each circuit, the alert reader will discover many other uses to which each circuit may be put.

Sensitive dc relay

Fig. 801 shows the circuit of a sensitive direct-current relay em-

ploying a single transistor (Sylvania 2N34). A signal input of 47.5 μA dc will close the 1.6-milliampere relay (RY) in the transistor collector circuit.

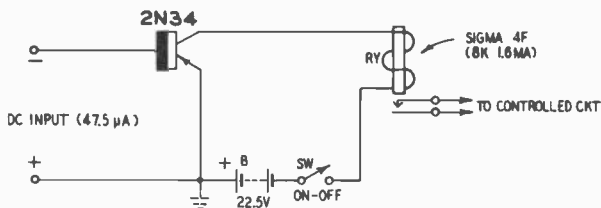


Fig. 801. Sensitive dc relay.

The transistor, connected as a common emitter, serves as a high-gain dc amplifier. It provides a current amplification of approximately 34 with the 8,000-ohm collector load presented by the relay coil. The input resistance of the circuit is 7,800 ohms, which means that the dc signal-input voltage is approximately 371 mv.

The dc bias is supplied by a 22.5-volt battery, B. The initial, steady collector current (i_{c0}) is very slight and does not cause significant battery drain nor does it create any difficulty in operation of the relay.

The dc input signal power required by this circuit is only 17.6 microwatts. The dc power required to operate relay RY without the transistor would be 20.4 milliwatts. The amplifier circuit thus provides a power gain of 116.

Ultra-sensitive dc relay

When it is desired to operate at lower dc signal levels than the sensitivity of the simple relay circuit of Fig. 801 will permit, additional amplification may be employed. Fig. 802 shows the circuit of an ul-

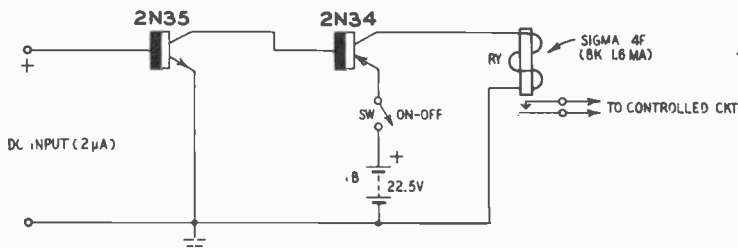


Fig. 802. Ultra-sensitive dc relay.

tra-sensitive circuit employing 2 stages of amplification. To use a single battery in the direct-coupled circuit, an n-p-n transistor (Syl-

vania 2N35) is employed in the input stage and a p-n-p type (Sylvania 2N34) in the output. This circuit has almost 25 times the sensitivity of the preceding one.

Direct current signal input is $2 \mu\text{a}$. The input resistance is 7,000 ohms and input voltage 14 mv. Signal-input power is 0.028 microwatt and the power gain is 735,000, or 58.7 db. Overall current gain is 800.

The single 22.5-volt battery B delivers 1.6 ma on closure of the relay. The current drain is negligible when the relay is not actuated (i. e., during absence of an input signal), being the small collector current due to the i_{co} characteristic of the transistor.

This is a complementary-symmetry type of dc amplifier circuit. The signal-input direct current flows into the base of the 2N35. This causes a larger current to flow in the 2N35 collector circuit, due to the beta characteristic. The 2N34 base-emitter circuit is the collector load of the 2N35, hence the amplified collector current of the latter flows into the 2N34 base and becomes amplified, in turn, in the 2N34 collector circuit which includes the coil of relay RY as its load.

Sensitive switching circuit for light-duty contacts

Switching operations often must be performed by light contacts which can carry only tiny amounts of current if sparking, pitting, and sticking are to be avoided. Currents often must be reduced also in full-sized contacts to prevent sparking in explosive or inflammable atmospheres. In either case, these light currents cannot be used directly to operate conventional relays.

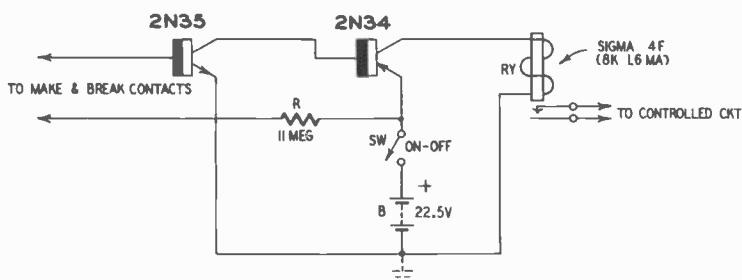


Fig. 803. Sensitive switching circuit for light-duty contacts.

Fig. 803 shows how the transistorized dc amplifier type of relay may be used to step up the small contact currents sufficiently to actuate a 1.6-ma relay, RY. The circuit also supplies approximately 2 microamperes for the contacts.

This circuit employs a 2-stage, direct-coupled amplifier similar to the one described in the preceding section and it has the same operating characteristics: overall current gain 800, signal-input power (handled by the contacts) 0.028 microwatt, and signal-input voltage 14 mv. The 22.5-volt battery B supplies 1.6 ma for relay closure.

The small signal current and power requirements of this relay circuit make it suitable for unconventional make-and-break contacts such as wires, electrolytes, foil or leaf strips, and light-duty thermostats. The circuit may be employed for temperature control by connecting an appropriate thermistor, thermocouple, or thermostat to its input leads. It may be employed also in electrochemical processes when the input leads are placed into an electrolyte, the relay sounding an alarm or stopping the process when the conductivity reaches the $2\text{-}\mu\text{a}$ level.

Sensitive ac relay

A bridge rectifier ahead of a transistorized dc amplifier and relay provides a sensitive ac relay. (See Fig. 804.)

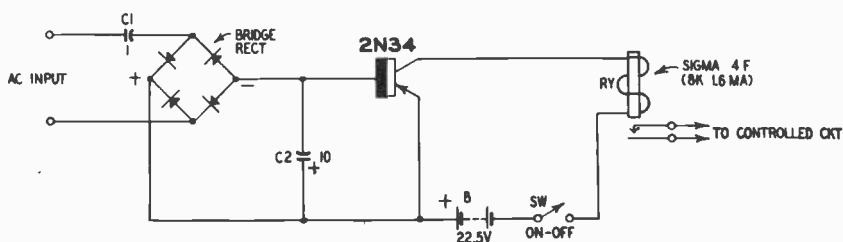


Fig. 804. Sensitive ac relay.

In this circuit, the ac is rectified by the bridge and the resulting dc applied to the input circuit of the transistor (Sylvania 2N34), base negative and emitter positive. Coupling capacitor C1 blocks any dc component which might be present in the input signal and would cause erroneous closure of the relay or damage to the rectifier. Capacitor C2 smooths the ripple in the rectifier output and thereby prevents chattering of the relay.

A signal-input voltage of 262 mv rms will cause closure of the 8,000-ohm relay in the collector circuit. The relay coil draws 1.6 ma dc for pickup. Current drain from the 22.5-volt battery (B) is negligible when there is no signal at the amplifier input terminals, being only the small component due to the i_{co} characteristic of the transistor.

For low-frequency and audio applications up to about 5 kc, the

bridge rectifier may be of the copper-oxide type. Selenium diodes may be used to 50 kc. Above 50 kc, germanium or silicon junction diodes will be required. Germanium and silicon are, of course, satisfactory also for the low frequencies.

Sensitive rf relay

In Fig. 805, a 1N34 germanium diode rectifier (D) has been connected ahead of a single-stage amplifier-type relay. A radio-frequency signal applied to the input terminals of the circuit is rectified by the diode and the resulting direct current is smoothed by the bypass capacitor C and passed into the base-input circuit of the transistor (Sylvania 2N34). The dc applied is negative to the base and positive to the emitter by poling the diode as shown in Fig. 805.

An rf input signal of 300 mv rms applied to the input terminals of the circuit will cause the relay (RY) to close. Somewhat more signal voltage will be required at the high frequencies since the rectification efficiency of the diode decreases as the operating frequency increases. It falls to 30% at about 40 mc.

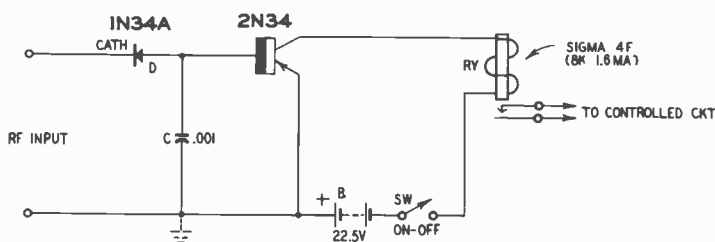


Fig. 805. Sensitive rf relay.

For the series-diode rectifier to operate correctly, a dc conductive path must be provided between the RF INPUT terminals by the output circuit of the radio-frequency source. If the latter cannot provide such a path, a 5,000-ohm resistor must be connected between the RF INPUT terminals. The sensitivity of the circuit will be approximately halved.

Dc drain from the 22.5-volt battery (B) is 1.6 ma when the relay closes and is negligible when there is no rf input signal.

Ultra-sensitive rf relay

In most instances, a single germanium diode and a sensitive transistorized dc amplifier-type of relay will be adequate for radio-frequency applications. However, when increased sensitivity is required, a higher-gain amplifier is needed. Fig. 806 shows a circuit

of this type in which a 1N34 series-diode rectifier (D) is operated at the input of a two-stage direct-coupled amplifier employing a p-n-p transistor (Sylvania 2N34) in the input stage and an n-p-n transistor (Sylvania 2N35) in the output stage to operate the 1.6-ma dc relay, RY.

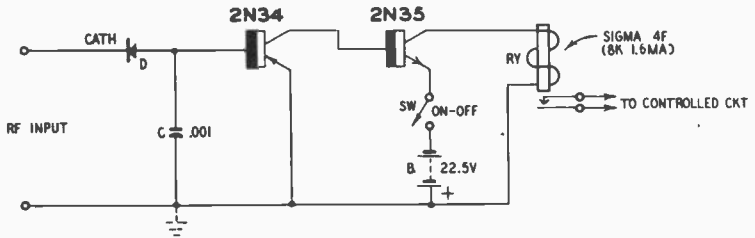


Fig. 806. Ultra-sensitive rf relay.

Diode D rectifies the rf signal applied to the input terminals of the circuit. Capacitor C serves to bypass the rf ripple component to ground. The anode of the diode is connected to the 2N34 base in order for this electrode to receive the negative dc output of the rectifier.

The rectified current flowing into the base is amplified in the 2N34 collector circuit by an amount governed by the beta of this transistor. Since the 2N35 base-emitter circuit constitutes the collector load of the 2N34, this amplified current flows into the base of the 2N35 and becomes amplified still further in the 2N35 collector circuit. Twice-amplified current then flows through the coil of relay RY.

An rf input of 10 mv is required for closure of the relay. Thus, this circuit is seen to be 30 times as sensitive as the one previously described. The required input voltage will increase as the operating frequency is raised, since the diode rectification frequency falls off as the frequency increases, decreasing to approximately 30% at 40 mc.

The output circuit of the radio-frequency source must provide a dc conductive path between the two RF INPUT terminals, otherwise the series-diode rectifier will not operate properly. If the rf source cannot supply such a path, a 5,000-ohm resistor must be connected between the RF INPUT terminals. The sensitivity of the circuit will be approximately halved. Shunt-diode rectification does not require the dc path but is notoriously insensitive.

Direct-current drain from the 22.5-volt battery (B) is 1.6 ma when the relay is closed and is negligible when there is no rf input signal.

Heavy-duty relay booster

In the preceding sections, a fairly sensitive dc relay has been shown in each circuit. This was done to keep safely within the maximum collector current and voltage ratings of the transistors. When a heavier-duty relay is preferred, a power transistor must be employed in the output amplifier. Fig. 807 shows a circuit of this type incorporating a 6-volt, 80-ohm relay (Potter & Brumfield SM5DJ).

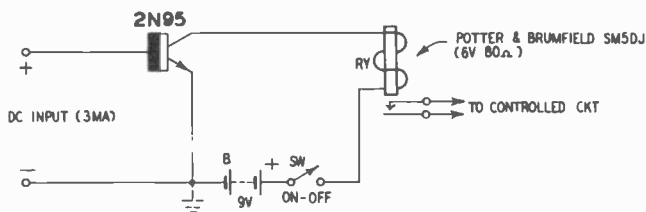


Fig. 807. Heavy-duty relay booster.

The relay constitutes the collector load of an n-p-n type power transistor (Sylvania 2N95) and is closed by 75 ma dc from the 9-volt battery B. A signal-input of 3 ma dc flowing in the base-emitter circuit of the transistor is amplified to 75 ma in the collector circuit and picks up the relay. This operation represents a current amplification of 25.

The input resistance of the circuit is approximately 50 ohms. From this factor and the input-current requirement, it follows that the dc signal-input power required for closure of relay RY is 0.45 mw and the overall power gain of the circuit is 45.5, or 16.6 db.

Battery B delivers 75 ma to close the relay. In the absence of an input signal, however, the steady collector current is 5 ma or less and will be lower still when the dc signal source has low resistance compared to the 50-ohm input resistance of the circuit.

Sensitized heavy-duty relay booster

Additional gain is provided in the circuit shown in Fig. 808 by an input amplifier stage employing a conventional n-p-n transistor (Sylvania 2N35) which is direct-coupled to the p-n-p power transistor (Sylvania 2N68). An input-signal current of only 110 μ a will cause closure of the 6-volt, 80-ohm, 75-ma relay, RY. This represents an overall current gain of 680.

The amplifier circuit is the direct-coupled complementary-symmetry type described in several of the preceding sections. The 2N68 base-emitter circuit is the collector load for the 2N35 while

the coil of relay RY constitutes the 2N68 collector load. The first amplified input current thus flows into the 2N68 base and the second amplified current through the relay coil.

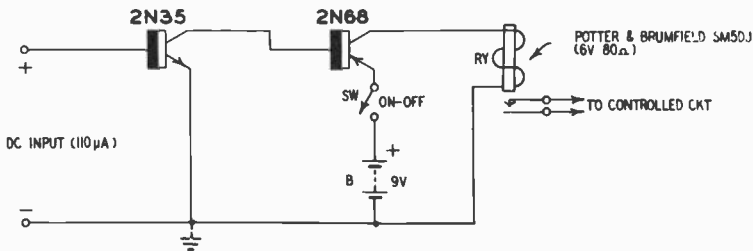


Fig. 808. Sensitized heavy-duty relay booster.

The dc drain from the 9-volt battery (B) is approximately 78 ma when the relay closes and is negligible in the absence of an input signal, being due at that time only to the i_{co} characteristics of the two transistors.

Sound-operated relay

Fig. 809 shows the circuit of a high-sensitivity, sound-actuated relay circuit in which the relay remains closed as long as a sustained sound is directed into the microphone. The circuit comprises a high-gain audio amplifier with class-B 100-mw output, a diode rectifier, and a 2.6-ma 13-ohm dc relay (Barber-Colman AYLZ-4436S Micropositioner).

The amplifier, operated from a crystal microphone, has an overall power gain of approximately 112 db. Transformer coupling is employed throughout for maximum gain per stage. The audio gain control potentiometer R4 serves as a sensitivity control. P-n-p type transistors (Raytheon CK721 and CK722) are employed.

The audio output of the amplifier, available at the 8-ohm secondary of transformer T4, is rectified by germanium diode D (Raytheon CK705) and smoothed by capacitor C5.

The high sensitivity of this circuit permits operation from low-intensity sounds when R4 is set for maximum gain. When operation is desired at a higher sound level, the SENSITIVITY CONTROL may be set accordingly by experiment to discriminate against lower-intensity ambient noise.

If sounds of only one pitch are to be used for control purposes and selectivity is desired with respect to all other pitches, suitable L-C type bandpass filters may be inserted in the base-input circuits of either or both of the first two stages of the amplifier.

Light-operated relay

Photoelectric relays have myriad uses in industrial and household electronics and in hobby pursuits. The transistorization of these light-controlled relay circuits has increased their effectiveness and widened their field of application.

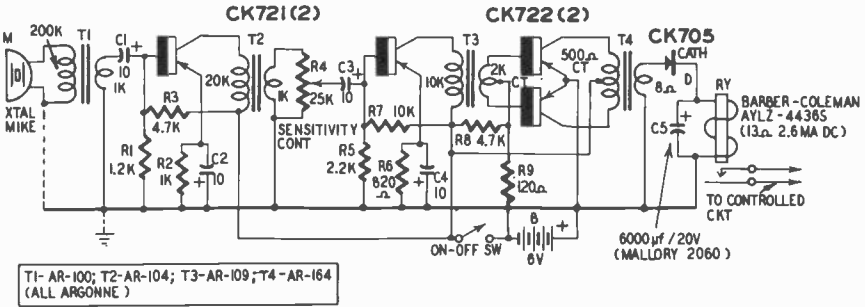


Fig. 809. Sound-operated relay.

Fig. 810 shows the circuit of a sensitive, light-operated relay employing a self-generating photocell and a two-stage direct-coupled transistor amplifier. This circuit may be operated from low-intensity light sources or from bright illumination, depending upon the setting of the sensitivity control, R.

The light-sensitive component in this circuit is the photocell, PC. A self-generating type is employed to obviate the necessity for a bias current. Light energy generates a voltage directly as a result of semiconductor action in the cell. A number of selenium and silicon photocells are available to the experimenter. Typical units

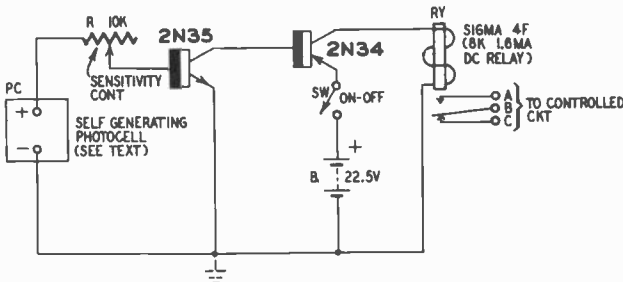


Fig. 810. Light-operated relay.

which will give excellent performance in this circuit are International Rectifier Corp. type DP-5 selenium and Natfab type P-100 silicon.

The photocell must be connected so that its positive dc output is applied to the base of the first transistor (Sylvania 2N35) through rheostat R. When the cell is illuminated, direct current proportional to the light intensity passes into the 2N35 base and becomes amplified in the collector circuit of this transistor. The amplified current flows into the base of the second transistor (Sylvania 2N34) and becomes amplified further in its collector circuit. This current then actuates relay RY.

The relay is provided with two stationary contacts, one of which is in contact with the movable armature when the latter is at rest. By making the proper connections to the three output terminals (A, B, C), the relay may be caused to open or close an external circuit, as desired. The following two examples illustrate use of the connections:

EXAMPLE 1: Illumination of photocell actuates external device

Make connections to output terminals A and C. As long as the photocell is darkened, the relay contacts are separated and the external circuit is open. When the cell is illuminated, the relay armature moves to the upper contact and the circuit through the external device is closed.

EXAMPLE 2: Interruption of light beam actuates external device

Make connections to output terminals B and C. As long as the photocell is illuminated, the relay armature is held against the upper contact and the external circuit is broken. When the light beam is interrupted, current ceases to flow in the relay coil, the armature drops to the lower contact and the circuit through the external device is closed.

During closure of the relay, the 22.5-volt battery (B) delivers approximately 1.65 ma. When the photocell is darkened, the current drain from the battery is negligible, being only the small value governed by the i_{co} characteristics of the transistors.

Photoelectric counter

In the circuit shown in Fig. 811, a photoelectric counter has been obtained by adding to the light-controlled relay circuit a five-digit electromechanical counter and an ac-operated power supply. The transistorized photoelectric amplifier is identical with the circuit given in Fig. 810.

The power supply is a voltage tripler consisting of the 6.3-volt filament-type transformer (T), three 1N34 germanium diodes and three 10- μ f 50-volt electrolytic capacitors (C1, C2, C3). The open-

circuit dc output of this tripler circuit is 26.7 volts which is three times the *peak* value of the 6.3 volts delivered by the transformer secondary. At full current load, when relay RY closes, the voltage drops to approximately 24.

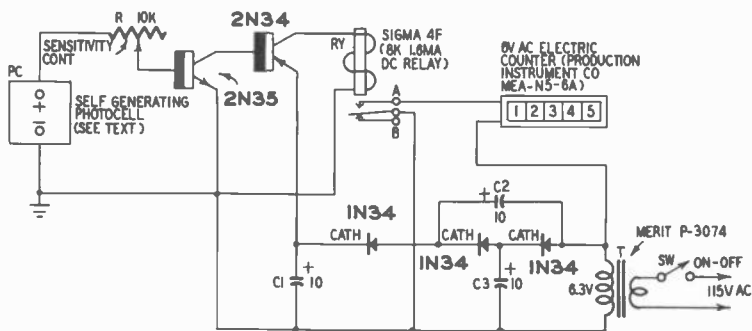


Fig. 811. Photoelectric counter.

The 6-volt ac counter is connected to the transformer secondary through the relay contacts. When wired as shown in Fig. 811, the counter will advance one digit each time the photocell is illuminated, since the relay armature then moves to the upper contact A. To operate from interruptions of a light beam, the counter must be connected instead to the lower relay contact B.

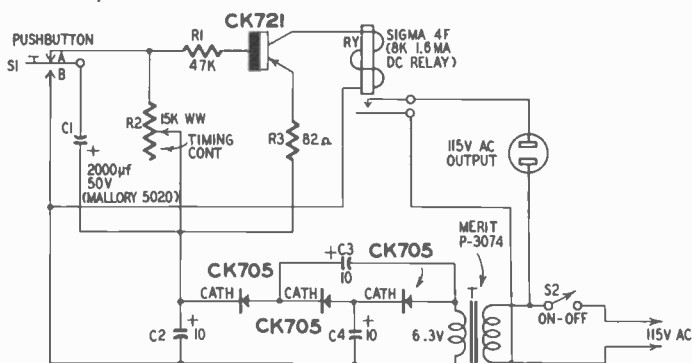


Fig. 812. Phototimer.

Phototimer

Fig. 812 shows the circuit of a phototimer which provides time intervals adjustable between 1 and 48 seconds. In addition to its intended use as a phototimer, this device may be used as a general-purpose timer.

The circuit is operated from the ac power line. The dc power supply is a voltage tripler consisting of the 6.3-volt filament-type transformer (T), three CK705 germanium diodes, and three 10- μ f 50-volt electrolytic capacitors (C2, C3, C4). The open-circuit dc output of this tripler circuit is 26.7 volts, which is three times the *peak* value of the 6.3-volt transformer secondary voltage. At full current load, when relay RY closes, the voltage falls to approximately 24. When the relay closes, 115 volts appear at the terminals of the female output receptacle into which may be plugged the photographic enlarger, printing box, or other ac-operated device which is to be timed.

Timing action is afforded by the 2,000- μ f capacitor (C1) and the 15,000-ohm rheostat (R2). Pushbutton switch S1 normally is in the position shown, with its arm against the upper contact A. To initiate operation S1 is depressed momentarily, causing C1 to be charged from the dc power supply through switch contact B. When S1 then is released, the voltage due to the capacitor charge is applied to the base-input circuit of the transistor (Raytheon CK721) through resistors R1 and R3. Current flowing into the CK721 base becomes amplified in the collector circuit of this transistor and picks up relay RY.

From this instant on, capacitor C1 discharges through R2 and the transistor input circuit at a rate determined mainly by the resistance setting of R2. When the capacitor voltage falls below the magnitude necessary to maintain the proper level of base current, the relay coil current drops below its hold-in value and the relay opens, de-energizing the 115-volt outlet. A dial attached to rheostat R2 accordingly may be calibrated to read directly in seconds.

Miniature-motor control

Fig. 813 shows a simple high-gain dc amplifier arrangement for starting, stopping and varying the speed of a miniature, model-type dc motor. The input-signal current for top speed of the motor (Lafayette F-177) is only 200 μ a and this signal may be derived from a number of convenient control sources, allowing this circuit to be used in model operation, remote-control systems, servomechanisms and instruments. The amplifier SENSITIVITY CONTROL rheostat (R) serves as a speed control for the motor.

The motor constitutes the collector load of the output transistor (Sylvania 2N68). A power transistor is required in this stage, since the motor current is 0.25 ampere. The input amplifier stage, which is direct-coupled to the power stage, employs an n-p-n transistor (Sylvania 2N35).

Dc drain from the 7.5-volt battery (B) is approximately 255 ma when the motor is running at full speed. When the DC INPUT signal is zero, this current drain drops to approximately 15 ma and is due to the i_{co} characteristics of the two transistors.

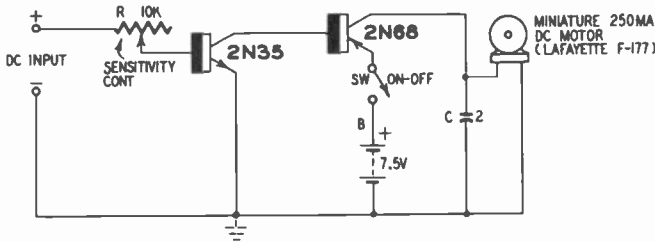


Fig. 813. Miniature-motor control.

General-purpose radio control relay

Radio-controlled relays find many commercial and amateur remote-control applications including model control, safety-alarm operation, intrusion (burglar) alarms, garage-door opening, electronic baby sitting, etc. Fig. 814 shows a receiver relay circuit for radio control involving an unmodulated carrier.

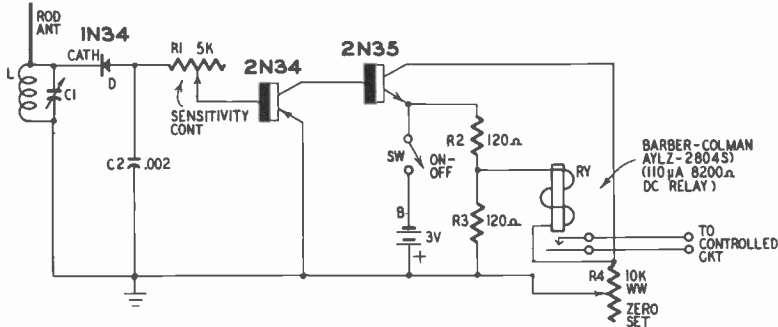


Fig. 814. General-purpose radio control relay.

The inductance of coil L and capacitance of variable capacitor C1 are chosen to resonate at the desired carrier frequency. A vertical rod or whip antenna will suffice in most instances, but other antennas are permissible.

The picked-up rf signals are rectified by the 1N34 germanium diode which is so poled in the circuit (anode output) that its negative dc output voltage is applied to the base of the p-n-p input transistor (Sylvania 2N34). Capacitor C2 serves as an rf bypass.

Rectified current passes into the 2N34 base after being limited

by the SENSITIVITY CONTROL rheostat R1. This current becomes amplified in the 2N34 collector circuit. The base-emitter input circuit of the n-p-n output transistor (Sylvania 2N35) constitutes the collector load of the 2N34 so that the amplified collector current of the 2N34 passes into the 2N35 base. This current then is amplified further in the 2N35 collector circuit and actuates the dc relay RY.

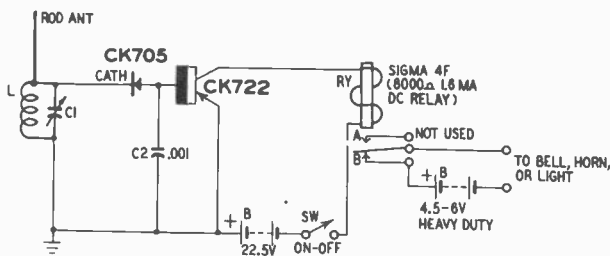


Fig. 815. Carrier failure alarm.

Because of the low current requirement of the relay coil, a four-arm resistance-bridge circuit is needed to balance the static zero-signal collector current out of this coil, otherwise the relay will be closed at all times. The process is the same as zero-setting a vacuum-tube voltmeter. The bridge consists of resistors R2, R3, R4 and the internal collector-to-emitter resistance of the 2N35 transistor. With zero-signal input to the circuit, the ZERO-SET rheostat R4 is adjusted to balance the bridge. This null point is indicated by the opening up of the relay contacts.

At the maximum-gain (minimum-resistance) setting of SENSITIVITY CONTROL rheostat R1 a 1-mv rms radio-frequency signal at the antenna will cause relay RY to close. Lower-intensity signals are ineffective, not because of lack of sensitivity in the amplifier, but because most semiconductor diodes cease to rectify at levels under 1 mv.

Current drain from the 3-volt battery (B) is approximately 13 ma. Most of this is current through the bleeder (voltage divider) R2-R3.

Carrier failure alarm

A sensitive device for signaling carrier interruption is invaluable around radio stations. Simple diode detectors driving sensitive dc relays have been used for this purpose but are so insensitive that they require tight coupling to the transmitter.

The carrier failure alarm circuit shown in Fig. 815 employs a

diode detector but follows it with a dc amplifier to build up the feeble dc output of the diode. This amplification permits the use of a small antenna (usually a vertical rod or whip) for signal pickup and a dc relay RY, the sensitivity of which does not exceed 1.6 ma.

Inductance of coil L and variable capacitor C1 are chosen to resonate at the carrier frequency. The diode (Raytheon CK705) is poled in the circuit for anode output so that the base of the transistor (Raytheon CK722) receives a negative voltage when the diode rectifies an rf signal. The resulting direct current flowing into the CK722 base becomes amplified in the collector circuit and picks up the relay.

The 22.5-volt battery (B) delivers 1.6 ma when the relay is actuated. In the absence of an rf input signal, the current drain is negligible, being the low value governed by the i_{co} characteristic of the transistor.

Impulse counter

Fig. 816 shows the circuit of a device for counting unipolarity pulses. These pulses—which may be as low as 14 mv or $2 \mu a$ (0.028

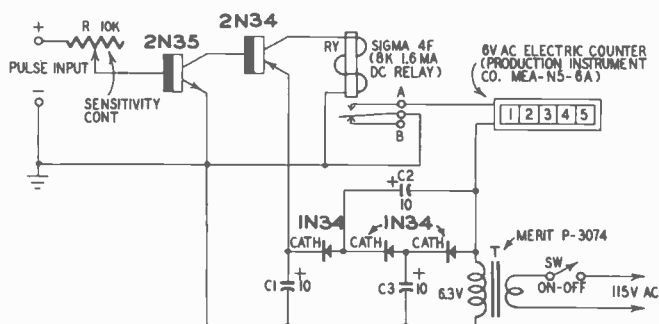


Fig. 816. Impulse counter.

microwatt) when the SENSITIVITY CONTROL rheostat (R) is set to its high-gain (low-resistance) position—may be derived from a photo-cell, make-and-break contactor, rectified radio signals, or any other source delivering either random or uniform pulses. The maximum counting speed is limited by the response of the five-digit electro-mechanical counter connected to the output and is 600 counts per minute. The total possible count is 99,999.

The signal pulses are presented to the PULSE INPUT terminals with the upper terminal positive. Pulse current flowing into the base of

the input transistor (Sylvania 2N35) through rheostat R becomes amplified in the collector circuit of this transistor. The base-emitter circuit of the second transistor (Sylvania 2N34) constitutes the collector load of the 2N35, hence the amplified current flows into the 2N34 base. Further amplification occurs in the output stage so that the 2N34 collector current flowing through the coil of the relay (RY) is sufficient to pick up the relay. On closure of the relay, 6.3 volts ac, derived from the secondary of the transformer (T) are applied to the counter.

The power-line-operated power supply is a voltage tripler consisting of the 6.3-volt filament-type transformer (T), three 1N34 germanium diodes and three 10- μ f 50-volt electrolytic capacitors (C1, C2, C3). The open-circuit dc output of this tripler circuit is 26.7 volts which is three times the *peak* value of the 6.3-volt transformer secondary voltage. At full current load, when relay RY closes, the voltage is approximately 24.

test instruments

BATTERY-OPERATED electronic test instruments offer the advantages of complete freedom from the power line, circuit simplicity and heat-free operation. Self-power minimizes the interaction and stray coupling which often occur between separate pieces of line-operated equipment and removes the instrument hum problem. Operation from batteries also makes completely portable operation practicable.

Heretofore, however, it has not been prudent to employ batteries in many instruments in which their use would have been desirable. The reason for this is that vacuum tubes impose a comparatively heavy load on small batteries, resulting in short battery life. If a battery-operated instrument were left switched on accidentally during idle periods, an expensive set of batteries would be destroyed quickly. Such operation of course is uneconomical. Coupled to this disadvantage is the necessity in tube circuits of both an A and B battery, a factor which increases circuit complexity as well as initial and replacement costs.

The low current requirements of the transistor and its ability to operate at dc voltages easily supplied by inexpensive flashlight cells fit the transistor for use in test instruments. Even when a standard B battery must be used in a transistorized instrument circuit, it may be expected to give up to several hundred times its normal service life in an equivalent tube circuit. The current drain in some transistorized instruments is so low that the instrument might be left running continuously for several months without harming the battery. Thus, the accidental on-time hazard is removed. For this

reason, some experimenters prefer to omit the ON-OFF switch, there being little point in switching off a battery when its service period will equal its shelf life.

The small size of the transistor and its circuit components permits miniaturization of instruments, an important consideration when the number of devices on the laboratory bench is continually growing.

At this writing, not all well-known instruments may be transistorized successfully. Further refinement is needed in some of the parameters of commercial transistors and further circuit study is necessary. The day is in sight, however, when all portable test instruments (and perhaps the stationary variety as well) will be transistorized. This chapter describes some of the circuits which are readily transistorized and may be duplicated with success by the experimenter.

Low-gain single-transistor dc microammeter

In Fig. 901, the base-to-collector current amplification factor of the Raytheon CK722 transistor is utilized to convert a 0-1 dc milliammeter into a 0-50 dc microammeter. The common-emitter amplifier circuit is employed with the milliammeter as the collector load.

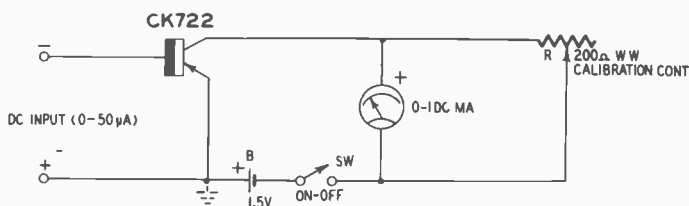


Fig. 901. Low-gain dc microammeter.

Full-scale deflection of the meter is obtained with a dc input of 50 microamperes. The 200-ohm wirewound rheostat R serves as a meter-shunting CALIBRATION control. With an accurately known 50- μ a input to the circuit, R is adjusted for exact full-scale deflection of the meter. Response of the circuit is linear; e. g., 0.5-ma deflection indicates an input of 25 μ a.

With no signal at the DC INPUT terminals, a meter deflection of 20 to 30 μ a will occur because of the i_{c0} characteristic of the transistor. This initial deflection may be suppressed, however, by setting the pointer to zero mechanically with the zero-set screw.

The input resistance of this circuit is approximately 8,000 ohms.

Zero-setting circuits for microammeter

The zero-signal collector current (i_{co}) increases with temperature. It very easily can deflect the meter so far upscale that suppression cannot be accomplished with the mechanical zero set of the instrument. This seriously restricts the range and accuracy of the simple microammeter shown in Fig. 901. Zero-setting circuits, similar to those employed in vacuum-tube voltmeters, therefore are desirable in all but the simplest transistorized dc microammeters. Figs. 902 and 903 show how zero-set circuits may be added to the microammeter. Except for this addition, the microammeter circuit and its operating characteristics remain the same as stated for Fig. 901.

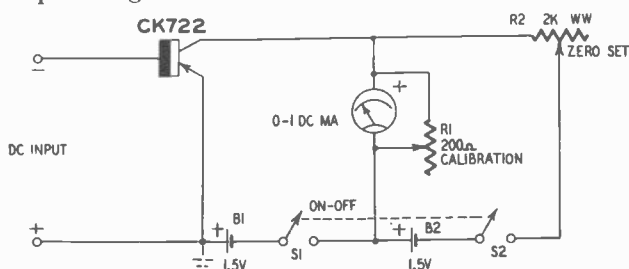


Fig. 902. Battery-type zero-set circuit for microammeter.

In Fig. 902, a second battery (B2) and a 2,000-ohm wirewound rheostat (R2) constitute the zero-set circuit. Current flows from B2 through the milliammeter in the opposite direction to that of i_{co} and is used to zero the meter by adjustment of R2. The range of the zero-set rheostat has been chosen such that the resistance setting

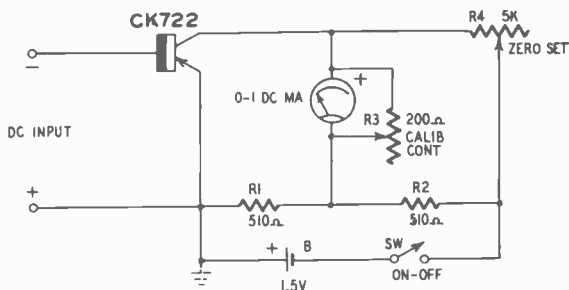


Fig. 903. Bridge-type zero-set circuit for microammeter.

when the meter is zeroed will be at least 10 times the resistance of the meter and calibration rheostat (R1) in parallel. This insures that the collector-current change, when the circuit is actuated by an input signal, will pass through the meter instead of the zero-set circuit.

The zero-set battery B2 may not be left in the circuit when the instrument is not in use, since it will discharge through R1, R2 and the meter. To disconnect it, a second pair of terminals (S2) has been added to the original ON-OFF switch S1. This resultant dpst switch operates both batteries.

The current drain of the second battery is quite small, not exceeding under any circumstances twice the i_{co} value encountered. Nevertheless, it is an extra power supply component and may be regarded as a nuisance by some instrument builders and users. To permit single-battery operation of the entire circuit, a bridge-type zero-set circuit (similar to the one employed in conventional vacuum-tube voltmeters) may be used. This arrangement is shown in Fig. 903.

The four arms of the bridge are R1, R2, R4 and the internal collector-to-emitter resistance R_t of the transistor. The bridge is balanced (meter zeroed) by adjustment of rheostat R4. At null, since R1 and R2 are equal arms: $R4 = R_t$.

This circuit has the convenience of single-battery supply. However, its current drain is higher since a "bleeder" current of 1.47 ma flows continuously through R1 and R2 in series as long as the instrument is switched on.

High-gain dc microammeter

There are three ways to increase the sensitivity of the simple low-gain microammeter circuit: (1) substitute a high-alpha transistor, (2) substitute a microammeter for the milliammeter or (3) both (1) and (2). Only the circuits shown in Figs. 902 and 903 are recommended for use with either of these schemes since the higher sensitivity increases the need for electrical zero setting.

Full-scale deflections as low as $20 \mu\text{a}$ may be obtained with the 0-1 milliammeter when a high-alpha transistor is used. Examples are Raytheon CK721 ($25 \mu\text{a}$), Sylvania 2N34 ($35 \mu\text{a}$) and Raytheon CK725 ($20 \mu\text{a}$).

When a dc microammeter is substituted for the milliammeter, rheostat R2 in Fig. 902 must be changed to one having a total resistance equal to at least 10 times the internal resistance of the meter. Also, R1 must have a total resistance equal to 2 times the meter resistance. When a microammeter is substituted in the circuit given in Fig. 903, rheostat R3 must have a total resistance equal to 2 times the internal resistance of the meter and the total resistance of R4 must equal 10 times the meter resistance. Fig. 904 shows a high-gain circuit in which a 0-100 dc microammeter has been sub-

stituted for the milliammeter. An n-p-n transistor (Sylvania 2N35) is employed. Full-scale deflection indicates a dc input of $10\ \mu\text{a}$. The steady bleeder current through R1 and R2 in series is $6.83\ \text{ma}$.

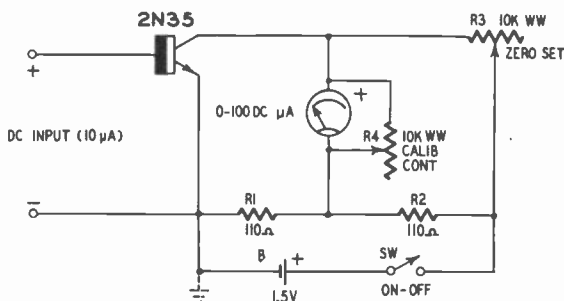


Fig. 904. High-gain dc microammeter.

A large increase in sensitivity is obtained when both the transistor and the indicating meter are changed. For example, a Raytheon CK725 may be used in conjunction with the 0-100 dc microammeter in Fig. 904, provided the polarities of DC INPUT terminals, microammeter and battery *all* are reversed. A dc input of only $2\ \mu\text{a}$ then will give full-scale deflection of the meter.

Supersensitive dc microammeter

Fig. 905 shows the circuit of a two-stage dc microammeter which gives full-scale deflection with $0.5\ \mu\text{a}$ input. This is a direct-coupled dc amplifier circuit in which collector current of the input transistor flows directly into the base-emitter input circuit of the output transistor.

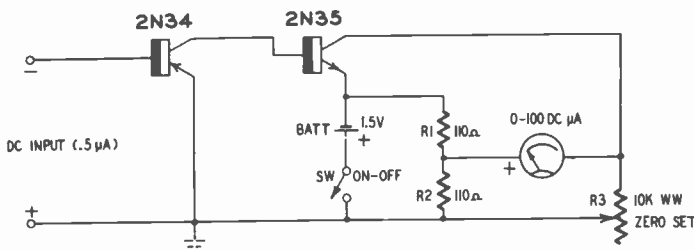


Fig. 905. Supersensitive dc microammeter.

To secure single-battery operation, a p-n-p transistor (Sylvania 2N34) is used in the first stage and an n-p-n (Sylvania 2N35) in the second stage. A bridge-type zero-set circuit (R1-R2-R3) is provided.

The steady bleeder current through R1 and R2 in series is 6.83 ma.

Electronic dc voltmeter ¹

A transistorized version of the familiar dc vacuum-tube voltmeter is shown in Fig. 906. Basically, this circuit is a two-stage amplifier-type dc microammeter (0–10 μ a) employing a 0–1 dc milliammeter and provided with suitable multiplier resistors to convert it into a multirange voltmeter.

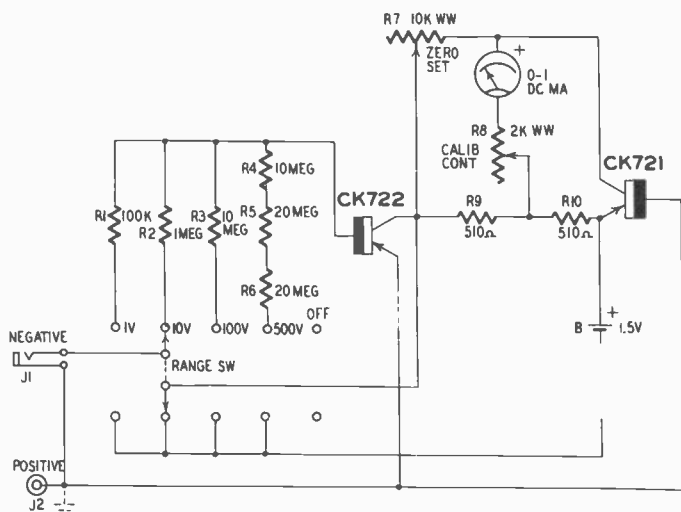


Fig. 906. *Electronic dc voltmeter.*

The high sensitivity of the microammeter circuit (10 μ a full scale) gives this voltmeter a resistance rating of 100,000 ohms-per-volt. On the 100- and 500-volt ranges, it has an input resistance equal to or higher than that of commercial units. The voltage ranges are 0–1, 0–10, 0–100 and 0–500.

The multiplier resistors R1 to R6, must be selected to have the exact specified values. The 50-megohm value required for the 500-volt range is obtained by series-connecting one 10- and two 20-megohm resistors. While the highest-voltage range (0–500) will be adequate in many applications, a 0–1,000-volt range may be desired as well as various intermediate ranges. Table 5 shows the values for multiplier resistors required for common voltage ranges other than those indicated in Fig. 906.

Table 5—Multiplier Resistor Values

Range (volts)	Multiplier Resistance (megohms)
0-2.5	0.25
0-3	0.30
0-5	0.5
0-7.5	0.75
0-15	1.5
0-25	2.5
0-50	5.0
0-250	25
0-300	30
0-750	75
0-1,000	100

One pole of the RANGE switch disconnects the battery. For protection of the instrument, the OFF position is placed after the highest voltage range.

Phone jack J1 is provided for the "high" dc input lead which in this circuit is connected to the negative terminal of the voltage source under test. This type of jack is employed because the "high" test lead usually is shielded. The "low" positive lead is connected to pin jack J2.

The microammeter portion of the circuit is similar to the circuit described in the sections immediately preceding this. A direct-coupled dc amplifier with two p-n-p transistors (Raytheon CK722 input and Raytheon CK721 output) is used. A bridge-type zero-set circuit is comprised of resistors R7, R9 and R10.

After setting the meter to zero by adjustment of rheostat R7, the instrument is calibrated initially by (1) setting the RANGE switch to its 1-volt position, (2) applying an accurately-known 1-volt potential to the input jacks and (3) adjusting rheostat R8 for exact full-scale deflection.

Audio voltmeter-millivoltmeter

An instrument for measuring audio voltages in the millivolt ranges is convenient for checking gain, small signal amplitudes, noise voltages, etc. Fig. 907 shows a transistorized millivoltmeter circuit.

Ranges provided are 0-10-30-100-300 millivolts and 0-1-3 volts. All are rms values. Higher voltage ranges have not been provided since they require impractically high resistance values in the range selector network. The indicating meter is a 0-100 dc microammeter.

The circuit consists of a five-stage amplifier driving a rectifier-type output voltmeter consisting of microammeter M and the four 1N34 germanium diodes. The input stage of the amplifier uses an emitter-follower circuit for high input impedance. This circuit ordinarily has an input impedance of 1 megohm but this is reduced to 0.5 megohm by the 1-megohm section (R8) of the input voltage divider. The input transistor is a Raytheon CK725 while all others in the circuit are Raytheon CK768's.

The instrument requires no zero adjustment since the meter receives no ac voltage until a signal is applied to the AC INPUT terminals. The only adjustment is the RANGE switch S1. The 25,000-ohm CALIBRATION CONTROL potentiometer R10 needs only occasional adjustment. It should be mounted safely inside the instrument case where its setting will not be disturbed accidentally.

Resistors R1 to R8 constitute the input-signal voltage-divider string and must have the exact specified values. Aerovox precision Carbofilm resistors are obtainable in the resistance values shown in Fig. 907, with an accuracy $\pm 1\%$.

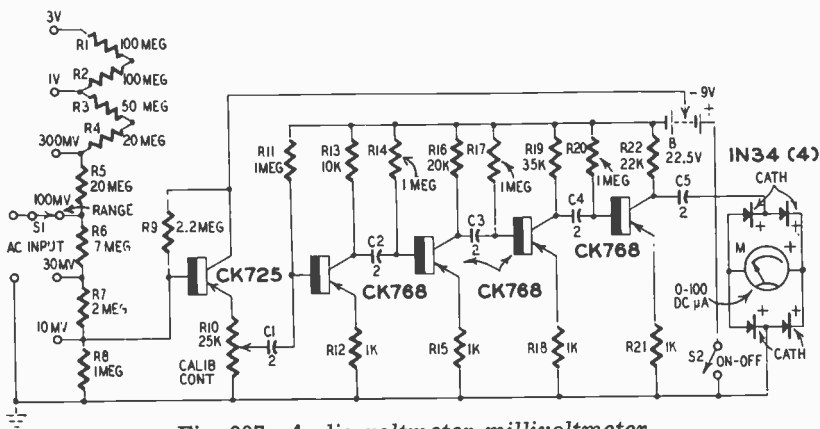


Fig. 907. Audio voltmeter-millivoltmeter.

To calibrate the instrument initially: (1) Close switch S2. (2) Set RANGE SWITCH S1 to its 1-volt range. (3) Apply an accurately known 1-volt rms, 1,000-cycle signal to the AC INPUT terminals. (4) Adjust CALIBRATION CONTROL R10 for exact full-scale deflection of the microammeter. (5) Check the deflection at 0.1, 0.2, 0.3, 0.4, 0.5, 0.6, 0.7, 0.8 and 0.9 volt input and at as many intermediate voltage points as possible. (6) Change RANGE switch S1 to its 3-volt position. A 3-volt input signal now must be applied to the input terminals and should deflect the meter to full scale. (7) Check the deflection at 0.5, 1, 1.5, 2 and 2.5 volts and at as many intermediate

voltage points as possible. (8) From the data obtained in steps 4, 5, 6 and 7 prepare a special card for the meter, with one scale graduated from zero to 1 and the other from zero to 3.

The instrument finally should be checked for accuracy throughout the frequency range from 50 cycles to 20 kc. Response falls off rapidly at frequencies higher than 20 kc.

Millivolt adaptor for ac vacuum-tube voltmeter

The lowest range of conventional ac vacuum-tube voltmeters usually is 1.5 volts. Many of the older instruments go no lower than 3 volts full scale. Fig. 908 shows the circuit of a simple adaptor which may be used as a probe ahead of an ac vacuum-tube voltmeter to convert it into a millivoltmeter. The adaptor increases the meter sensitivity 100 times, converting the 1.5-volt range to 0-15 mv and the 3-volt range to 0-30 mv.

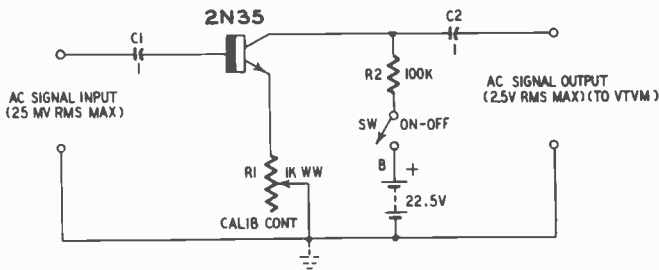


Fig. 908. Millivolt adaptor for ac vtvm.

The adaptor is a single-stage common-emitter amplifier employing a Sylvania 2N35 transistor. The high input impedance of 25,000 ohms is made possible by the large amount of degeneration introduced by the unbypassed emitter resistor R1 which serves also as the calibration control. The maximum input-signal voltage before severe output distortion is 25 mv rms. The corresponding maximum output is 2.5 volts rms. The adaptor output must be applied to a high-impedance load such as the input circuit of a vtvm or oscilloscope. The minimum permissible load resistance, for full utilization of the voltage gain of the adaptor, is 1 megohm.

To calibrate the adaptor initially: (1) Connect an ac vacuum-tube voltmeter to the AC SIGNAL OUTPUT terminals. (2) Apply an accurately known 10-mv rms, 1,000-cycle signal to the AC SIGNAL INPUT terminals. (3) Set the CALIBRATION CONTROL rheostat R1 for 1-volt deflection of the vtvm.

Sensitizer for high-range dc milliammeter

The base-to-collector current amplification of a power transistor (CBS 2N55) is utilized in the circuit shown in Fig. 909 to boost the sensitivity of a high-range dc milliammeter. The circuit is shown connected to a 0-10 milliammeter which it converts into a 0-0.5 dc milliammeter (0-500 microamperes). The input-signal voltage is 200 mv dc.

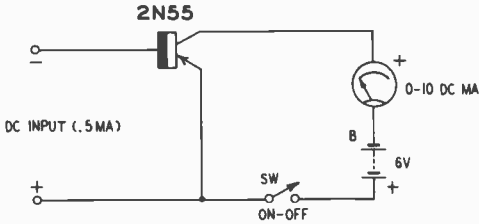


Fig. 909. Sensitizer for high-range dc milliammeter.

This same scheme may be used to convert a 0-50 dc milliammeter to a 0-2.5 ma, a 0-100 ma to a 0-5 ma, or a 0-1-ampere instrument to 0-50 ma. A heat radiator (based upon 6 x 6 x 1/8-inch aluminum base) must be used when 1 ampere of collector current is to be handled.

TV antenna compass²

The conventional compass used in the orientation of TV antennas for maximum signal pickup consists of a diode-type rf voltmeter (germanium diode plus current meter). The current meter

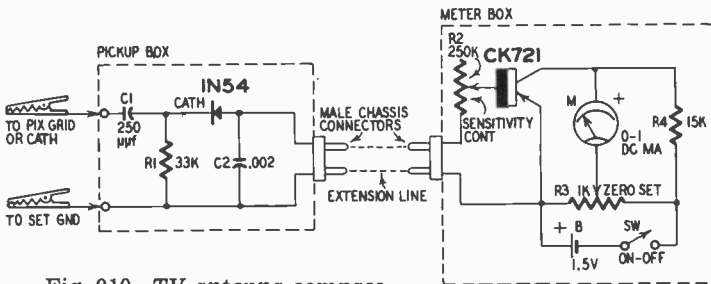


Fig. 910. TV antenna compass.

is placed on the roof in clear view of the antenna technician while the diode is connected to the receiver and feeds the meter through a long extension line. The diode is capacitance-coupled to the picture-tube grid or cathode, depending upon which element receives the output of the video amplifier.

The current meter usually is a sensitive dc microammeter. It can be both expensive and delicate. A single-stage transistorized dc amplifier makes it possible to use instead a 0-1 dc milliammeter in the compass and, at the same time, to obtain greater sensitivity, than is afforded by the microammeter. Fig. 910 shows the circuit of the transistorized compass.

The pickup box is placed at the receiver. Crocodile clips permit connection to the picture tube and to receiver ground. The detector diode (Sylvania 1N54) is capacitance-coupled through C1 to protect it from any dc component coming from the receiver. Dc output from the diode passes through a two-wire extension line (which may be ordinary rubber-covered electric cord) to the meter box on the roof.

The meter box contains, in addition to the 0-1 dc milliammeter M, the dc amplifier built around the p-n-p transistor (Raytheon CK721). The base rheostat R2 serves as the SENSITIVITY control, while potentiometer R3 is the ZERO-SET control in the collector bridge circuit.

The instrument requires no calibration since it is used only as a peak indicator. The antenna is rotated for maximum deflection of meter M. If the pointer is driven off scale by a strong signal, the sensitivity of the circuit must be reduced by resetting R2.

TV field-strength meter ³

Fig. 911 shows the circuit of a field-strength meter which may be used to measure the intensity of signals picked up by an antenna or as a highly sensitive TV antenna compass. The single tuning range of this instrument extends from 40 to 226 megacycles, covering all television channels and the various frequencies, amateur and commercial, between these frequency limits.

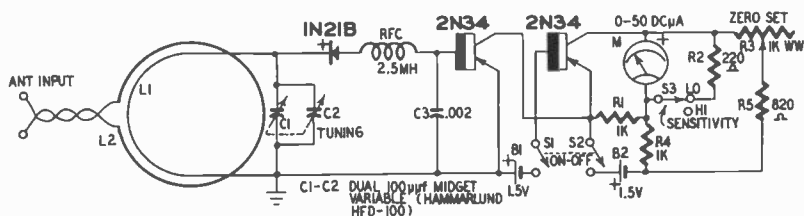


Fig. 911. TV field-strength meter.

The instrument circuit includes a tuner (L1-L2-C1-C2), a detector employing a silicon point-contact diode (Sylvania

1N21B) for improved high-frequency performance, a two-stage dc amplifier (Sylvania 2N34's) which builds up the feeble dc output of the diode sufficiently to deflect the 0–50 dc microammeter M, and a bridge-type zero-set circuit (R1–R3–R4–R5). The meter shunt resistor R2 reduces the sensitivity of the circuit when the SENSITIVITY switch S3 is thrown to its LO position. When S3 is in its HI position, an rf signal voltage of 1 mv will produce the smallest readable deflection of the meter.

The main coil of the tuner consists of a hairpin loop L1 made with No. 20 bare tinned copper wire. Fig. 912-a shows its dimensions. This coil is tuned by a 200- μ f midget variable capacitor. A satisfactory 200- μ f capacitor free from self-resonance within the tuning range of the field-strength meter could not be obtained. A dual 100- μ f unit (C1–C2) accordingly has been used with its two sections connected in parallel by a short *straight* piece of No. 20 bare wire. The antenna coupling coil L2 is a single 0.75-inch-diameter turn of insulated hookup wire, made as shown in Fig. 912-b and mounted 1/16 inch from L1 (Fig. 911).

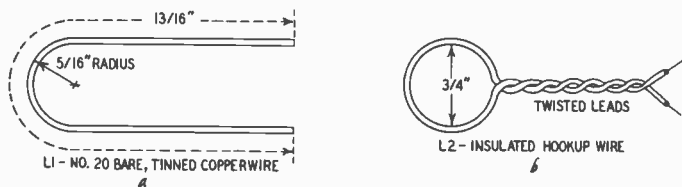


Fig. 912-a, -b. Details of field-strength meter coils.

The SENSITIVITY switch S3 shunts the microammeter with a 220-ohm resistor R2 to change the meter deflection by approximately 10. Cutting this resistor into the circuit disturbs the zero setting of the meter by less than 1% of full scale (one small scale division). When S3 is set to its HI position, any random fluctuation of the meter pointer due to transistor noise is less than one-quarter of a small scale division.

With switch S3 at its HI position, battery B1 delivers 18 μ a and B2 1.6 ma when an rf input signal deflects meter M to full scale. With S3 set to LO, B1 delivers 30 μ a and B2 2 ma. These low current drains make operation from flashlight cells entirely feasible.

The best frequency-calibrating source will be an accurate rf signal generator covering the range of 38 to 230 mc, preferably on fundamental frequencies all the way. This instrument should be

a standard oscillator (either modulated or unmodulated) but *not* a sweep generator. The generator output is connected to the ANTENNA INPUT terminals of the field-strength meter. If the generator has an output control (attenuator) calibrated in microvolts, it may be used to calibrate the scale of the microammeter in microvolts or millivolts.

In this field-strength meter, the microammeter response is approximately square-law due to the shape of the conduction curve of the crystal diode at the low signal amplitudes involved.

If the meter is to be used exclusively for television, the C1-C2 dial may be graduated conveniently in the TV channel numbers. The tuner responds separately to the video and sound frequencies. The former are recommended for the dial marking. For this purpose, the following table shows the video carrier frequencies corresponding to the various channel numbers.

Table 6—Video Carrier Frequencies

Channel Number	Frequency (mc) Video Carrier
2	55.25
3	61.25
4	67.25
5	77.25
6	83.25
7	175.25
8	181.25
9	187.25
10	193.25
11	199.25
12	205.25
13	211.25

Bridge null detector

An R-C-coupled, high-gain transistor amplifier followed by a rectifier type microammeter makes a compact null detector suitable for impedance bridges. Operated from 6 volts of flashlight battery, this self-contained instrument may be built small enough to be installed inside the bridge.

Fig. 913 shows the circuit of a null detector. By employing a common-collector input stage, the circuit is given an input impedance of 0.25 megohm. This stage is followed by two common-emitter stages. The output stage is capacitance-coupled, through C4, to the meter circuit consisting of a miniature bridge-type copper oxide meter rectifier RECT and a 0-50 dc microammeter M. Single-battery operation is made possible by employing an n-p-n

transistor (Sylvania 2N35) in the common-collector input stage and p-n-p's (Sylvania 2N34) in the two other stages.

When the GAIN CONTROL potentiometer R1 is set for maximum gain, an input signal of 5 millivolts rms will produce full-scale deflection of the meter.

The total drain from the 6-volt battery B is 1.2 ma dc.

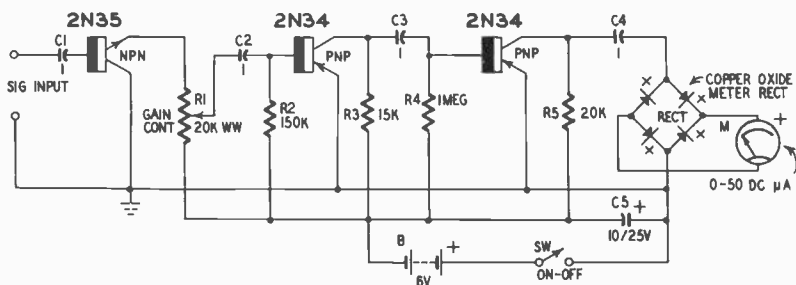


Fig. 913. Bridge null detector.

Sharply tuned null detector

Fig. 914-a shows the circuit of a bridge null detector having high selectivity and high input impedance. Like the circuit previously described, it has an input impedance of 0.25 megohm afforded by the common-collector input amplifier stage. Two n-p-n (Sylvania 2N35) and three p-n-p transistors (Sylvania 2N34) are employed.

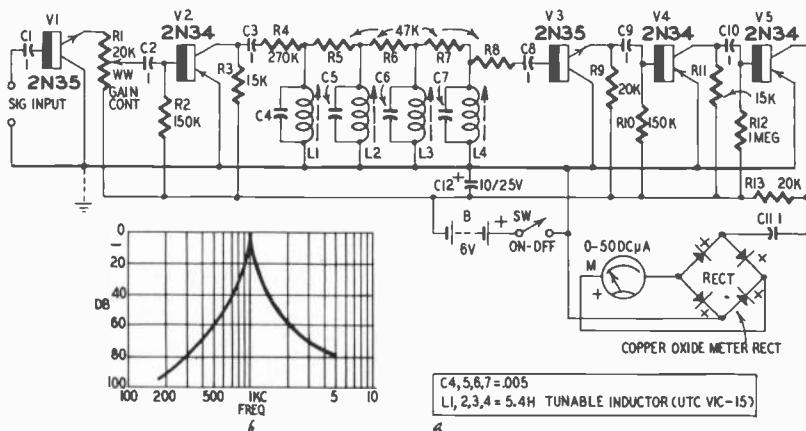


Fig. 914-a, -b. Sharply tuned null detector.

The common-collector stage V1 is followed by a high-gain common emitter V2. The output of the latter is presented to a four-

section tunable bandpass filter ⁴ comprising variable inductors L1, L2, L3 and L4; capacitors C4, C5, C6 and C7 and resistors R4, R5, R6, R7 and R8. The inductors (UTC type VIC-15) are tuned by means of Allen-screwed slugs. A second common-collector stage (V3) follows the filter, the high input impedance of this stage offering negligible loading to the filter. The remaining common-emitter stages (V4 and V5) build up the signal to compensate for the attenuation introduced by the filter and to drive the meter circuit (rectifier RECT and the 0–50 dc microammeter M).

The circuit may be pretuned to any bridge operating frequency between 500 and 1,000 cycles. By increasing or decreasing capacitances C4, C5, C6 and C7, other frequency ranges may be obtained. Fig. 914-b shows the response when the circuit is tuned to 1,000 cycles. Note from this plot that transmission is down 60 db at points one octave on each side of the center frequency. A 5-mv rms input signal at the center frequency will deflect meter M to full scale.

The null detector may be aligned in the following manner: Connect an audio signal generator to the SIGNAL INPUT terminals and set its output high enough to give quarter-scale deflection of meter M. With the GAIN control R1 set for maximum gain, start with L4 and work progressively back to L1, adjusting each inductor for peak deflection of the meter. If at any point in this adjustment the pointer is driven off scale, reduce the setting of the generator output control or of R1 or both. Readjust each inductor for sharper peaking.

If the selectivity indicated by Fig. 914-b is too high, the four inductors may be stagger-tuned for a flat-topped response curve.

Crystal-type sound marker generator ⁵

Crystal-type marker-pip generators are required for spotting the 4.5-mc sound-channel point in visual TV alignment since self-excited oscillators are not accurate enough for this purpose.

Many early-model TV sweep generators have no provision for crystal marker-circuit operation. In such instances, the crystal oscillator circuit shown in Fig. 915 may be connected externally. This circuit employs an n-p-n rf transistor (Sylvania 2N94A) and is resonated by a 9-to-16-mh slug-tuned inductor L (Miller 4506). Peak deflection of an rf vacuum-tube voltmeter connected temporarily to the RF OUTPUT terminals will indicate resonance.

Emitter dc bias is supplied by the voltage drop across resistor R2; collector dc bias by the drop across R3. Total current drain

from the 7.5-volt battery B is 13 ma dc. The open-circuit rf output of the oscillator is 5 volts rms.

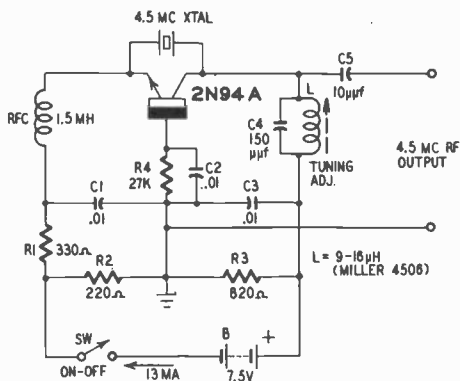


Fig. 915. Crystal-type sound marker generator.

Audio signal injector ⁶

Fig. 916 shows the circuit of a miniaturized 1,000-cycle oscillator which may be built into a probe handle and used to troubleshoot audio amplifiers by the signal-injection method. A p-n-p transistor (Sylvania 2N34) is operated from a 1.5-volt penlight cell in this device.

A simple tickler-feedback oscillator circuit is employed. The frequency is determined by the 2,000-ohm winding of transformer T and the capacitance of C1. The transformer must be phased correctly or oscillation will not be obtained. The simplest procedure is to follow the color coding shown in Fig. 916. The ground connection marked *chassis* in the wiring diagram is a connection

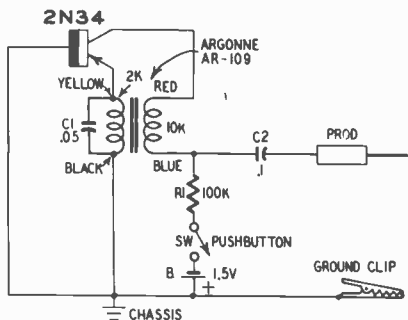


Fig. 916. Audio signal injector.

to the metal shell or shield of the probe into which the signal injector is built.

The signal output amplitude is 1 volt peak-to-peak across a load

the instrument is recalibrated. The coupling coil L2 is made by jumble-winding 20 turns of No. 24 dcc wire around one end of L1.

The circuit consists of the crystal rf oscillator employing an n-p-n rf transistor (General Electric 2N170), 1N34 diode modulator and a 1,000-cycle audio oscillator employing a p-n-p transistor (General Electric 2N107). When switch S4 is set to its UNMOD position, the rf output signal is unmodulated.

The polarity of the 1N34 diode must be followed or this unit will not function correctly as a modulator. The polarity of the feedback transformer T also is important. Follow the transformer color coding shown in Fig. 917, otherwise this circuit will not oscillate. If a modulating frequency lower than 1,000 cycles is desired, increase the capacitance of C6. Two or more modulating frequencies may be accommodated by arranging a switch for cutting appropriate capacitors into the circuit.

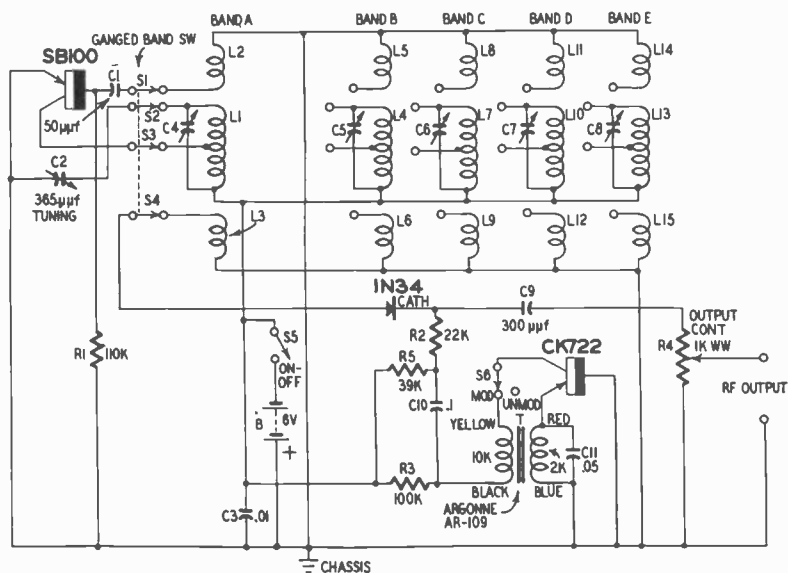
To adjust the instrument: (1) Connect an rf vacuum-tube voltmeter to the RF OUTPUT terminals. (2) Set OUTPUT control R6 for maximum output. (3) Set the FREQUENCY switch S2-S3 to its position A. (4) Screw the slug of inductor L1 about halfway into the coil. (5) Close switch S1. (6) Adjust trimmer C2 for peak deflection of the vtvm. (7) Throw switch S2-S3 to position B and adjust trimmer C3 for peak deflection of the vtvm. (8) If either adjustment (C2 with switch S2-S3 at position A or C3 with S2-S3 at position B) cannot be reached in the tuning range of the capacitor, reset the slug in L1 and repeat. However, this changes the inductance of the coil and both frequency adjustments must be remade.

The modulation may be checked by feeding the rf output signal into a receiver, throwing S4 to its MOD position, tuning the receiver to the aligner carrier frequency and monitoring by ear.

Rf signal generator

Fig. 918 shows the circuit of a general-purpose rf signal generator supplying both CW and amplitude-modulated signals in the range from 440 kc to 30 mc. The upper frequency is dependent upon the individual characteristics of the rf transistor (Philco SB100) employed in the rf oscillator stage. Some of these transistors will reach 40 or 45 mc. However, all should attain 30 mc without difficulty. The frequency range is covered in five switched bands: 440-1,200 kc, 1-3.5 mc, 3.4-9 mc, 8-20 mc and 18-30 mc. Harmonics of the last band are usable to much higher test frequencies.

A complete coil set (main coil, tickler and output coupling winding) is switched into the circuit each time the band is changed. These coils are selected by means of the four-pole, five-position nonshorting switch (S1-S2-S3-S4). The SB100 collector is tapped down the main coil for impedance-matching purposes. Each main coil is shunted by a 75- μf air trimmer capacitor (C4, C5, C6, C7, C8) for use in alignment. The coil table gives winding instructions for the five sets of coils.



C4, C5, C6, C7, C8 = 75 μf AIR TRIMMER

BAND A = 440-1200 MC; BAND B = 1-3.5 MC; BAND C = 3.4-9 MC; BAND D = 8-20MC; BAND E = 18-30 MC.

Fig. 918. Rf signal generator.

A 1N34 diode modulator is used and is driven by a 1,000-cycle audio oscillator employing a p-n-p transistor (Raytheon CK722). Lower modulating frequencies may be obtained by increasing the capacitance C11. Switch S6 in the af oscillator circuit is the modulation selector. When it is thrown to its UNMOD position, the rf output signal is unmodulated. The correct poling of transformer T is essential, otherwise the af stage will not oscillate. Follow the color coding shown for this transformer in Fig. 918. Mount the transformer so that it is not in the fields of the rf coils.

The signal generator may be aligned initially by the conventional method, using another accurately-calibrated rf signal generator or a frequency standard. Each coil set must be peaked (by

Table 7—Coil Table for Rf Signal Generator

Band A 440–1,200 kc	L1: 187 turns No. 32 enameled wire closewound on 1-inch-diameter form. Tap 90th turn from low end.
	L2: 45 turns No. 32 enameled wire closewound on same form as L1. Space 1/16 inch from top of L1.
	L3: 20 turns No. 32 enameled wire closewound on same form as L1. Space 1/16 inch from bottom of L1.
Band B 1–3.5 mc	L4: 65 turns No. 32 enameled wire closewound on 0.5-inch-diameter form. Tap 33rd turn from lower end.
	L5: 15 turns No. 32 enameled wire closewound on same form as L4. Space 1/16 inch from top of L4.
	L6: 8 turns No. 32 enameled wire closewound on same form as L4. Space 1/16 inch from bottom of L4.
Band C 3.4–9 mc	L7: 27 turns No. 26 enameled wire closewound on 1/2-inch-diameter form. Tap 13th turn from low end.
	L8: 8 turns No. 26 enameled wire closewound on same form as L7. Space 1/16 inch from top of L7.
	L9: 5 turns No. 26 enameled wire closewound on same form as L7. Space 1/16 inch from bottom of L7.
Band D 8–20 mc	L10: 10 turns No. 22 enameled wire on 0.5-inch-diameter form. Space to winding length of 0.5 inch. Tap 5th turn.
	L11: 4 turns No. 22 enameled wire closewound on same form as L10. Space 1/16 inch from top of L10.
	L12: 3 turns No. 22 enameled wire closewound on same form as L10. Space 1/16 inch from bottom of L10.
Band E 18–30 mc	L13: 51/2 turns No. 22 enameled wire airwound 1/2-inch in diameter. Space to winding length of 1/2 inch. Tap 3rd turn from low end.
	L14: 4 turns No. 22 enameled wire airwound, closewound 1/2 inch in diameter. Mount 1/16 inch from top of L13.
	L15: 3 turns No. 22 enameled wire airwound, closewound 1/2 inch in diameter. Mount 1/16 inch from bottom of L13.

means of C4, C5, C6, C7 or C8) with tuning capacitor C2 set to its top-of-band point.

Radio-frequency comparator

An untuned heterodyne detector is useful for comparing aurally two rf signals, one usually being of unknown frequency and the other known. A germanium diode is convenient for the purpose but its audio output is so feeble when the rf amplitude is low that that zero-beat identification becomes difficult.

Fig. 919 shows the circuit of a diode type heterodyne detector to which has been added a two-stage audio amplifier. Simplified direct coupling between amplifier stages has been made possible by the use of an n-p-n transistor (General Electric 2N170) in the first stage and a p-n-p (General Electric 2N107) in the second.

The two rf input signals are presented simultaneously (through RF INPUT 1 and RF INPUT 2) to the 1N34 detector. Capacitance coupling, through C1 and C2, blocks any dc component which might be present in the two signal sources. A radio-frequency filter (C3-RFC-C4) is provided in the output circuit of the diode.

A strong headphone signal is delivered by this circuit. A certain amount of volume control is obtained by adjustment of rheostat R5. The base resistor R3 of the first transistor may require some adjustment (usually an increase) for a smooth signal in the headphones.

Heterodyne frequency meter ⁷

A heterodyne frequency meter uses the beat-note method to measure an unknown radio frequency by comparing the unknown (by zero beating) with the frequency of a single-range variable-frequency rf oscillator. The two signals are mixed in an untuned detector and the beat note is amplified by an af amplifier and presented to headphones. The wide measurement range of the heterodyne frequency meter results from the fact that harmonics of the unknown signal may be used and so may harmonics of the oscillator. This extends the range far below and above the fundamental frequency band of the rf oscillator in the test unit.

Fig. 920 shows the circuit of a heterodyne frequency meter which may be used to measure frequencies from 50 kc to 30 mc. The tuning range of the oscillator in this instrument is 500 to 1,000 kc.

The rf oscillator is a tickler-feedback circuit employing a p-n-p

rf transistor (Raytheon 2N112). Details of the coil combination (L1-L2-L3) are given in Fig. 921. The windings of the coil must be phased correctly or the feedback will not be of the proper polarity

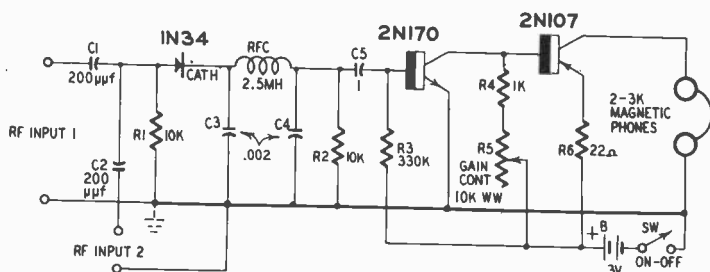


Fig. 919. Radio-frequency comparator.

for oscillation. The ends of the windings have been given corresponding letter symbols in Figs. 920 and 921.

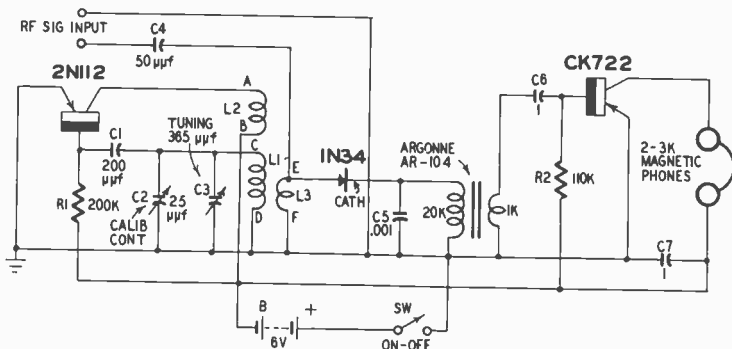


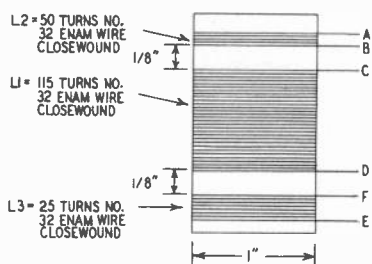
Fig. 920. Heterodyne frequency meter.

The detector (mixer) stage employs a 1N34 germanium diode. The oscillator output is coupled into this diode through winding L3 of the coil combination. The external signal is coupled to the diode through capacitor C4 from the RF SIGNAL INPUT terminals. The single-stage audio amplifier (Raytheon CK722) is transformer-coupled for maximum af power gain.

The initial calibration of the instrument may be made in the following manner: (1) Feed a 500-kc unmodulated signal into the instrument through the RF SIGNAL INPUT terminals. (2) Set the tuning capacitor C3 to its full-capacitance position. (3) Adjust trimmer C2 for zero beat. The C3 dial now may be marked 500 kc at this setting. (4) Substitute a 100-kc frequency standard for the

signal generator. (5) Reset C2, if necessary, to establish a more accurate zero beat with the standard. (6) Tune C3 slowly from this 500-kc setting until the next standard-frequency point is tuned in at zero beat. Mark this point 600 kc on the C3 dial. (7) Repeat at each standard spot frequency, marking the dial 700, 800, 900 and 1,000 kc accordingly. If the frequency standard is equipped also with a 10-kc multivibrator, 10-kc points may be located and inscribed between adjacent 100-kc graduations on the dial.

Fig. 921. Detail of coil for heterodyne frequency meter.



Periodically, a single-point frequency check of the heterodyne frequency meter may be made in the following manner: Set the dial to 1,000 kc (1 mc) and, with a 100-kc frequency standard delivering a signal to the RF SIGNAL INPUT terminals, adjust trimmer C2, if required, to re-establish exact zero beat. This single adjustment

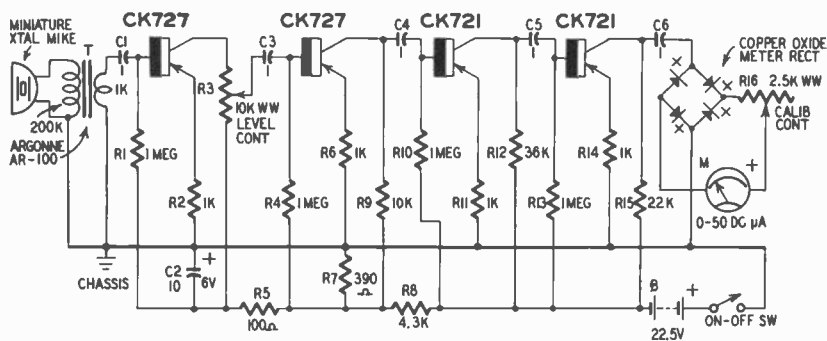


Fig. 922. Sound-level (noise) meter.

compensates for any frequency drift due to variations in circuit or battery characteristics that have occurred since the initial calibration.

Sound-level (noise) meter

Fig. 922 shows the circuit of a compact portable instrument which

may be used to make sound surveys and noise measurements. If a piezoelectric vibration pickup is substituted for the crystal microphone, this instrument becomes a vibration meter. It also may be employed without change as an electronic stethoscope.

The instrument consists of a crystal microphone, four-stage high-gain af amplifier and output meter. Low-noise transistors (Raytheon CK727) are used in the two input stages; high-alpha audio units (Raytheon CK721) in the last two stages. Voltage divider R7-R8 reduces the collector supply to 2 volts for the two CK727's.

Potentiometer R3 serves as the LEVEL control. A precision step-type potentiometer (attenuator) may be used in this position when accurate decibel steps are required. Rheostat R16, the CALIBRATION CONTROL, is set for full-scale deflection of meter M with a standard sound input and with R3 set for maximum transmission.

The response of the instrument, not including the microphone, is 7 db down at 10 kc and is -14.8 db at 20 kc. Flat response is obtained from 50 to 1000 cycles, dropping to -2 db at 5,000 cycles.

AF-Rf signal tracer

The signal tracer circuit shown in Fig. 923, unlike many early transistorized tracers, is competitive with tube circuits since it has high input impedance (0.5 megohm), adequate audio output for loudspeaker operation (100 mw) and provides aural (loudspeaker) and visual (meter) indications. Its low current drain permits operation from self-contained flashlight-type batteries, giving complete divorcement from the power line, compact construction, full portability and freedom from interaction with the device under test or with the signal generator.

An amplitude-modulated test signal is employed when tracing rf signals with this instrument. An af signal at any frequency between 50 and 20,000 cycles is satisfactory for tracing in audio systems. The rf probe contains a 1N34 germanium diode which acts as a demodulator. The af probe is a conventional meter-type unit containing no resistors, diodes or capacitors. Both probes have shielded cables terminated with shielded phone plugs for insertion into INPUT jack J.

The tracer circuit is a four-stage, high-gain, high-output af amplifier employing five p-n-p transistors. The high input impedance of the first stage (1 megohm at the common test frequencies of 400 and 1,000 cycles) is due to the connection of the first transistor (Raytheon CK725) as an emitter follower. This basic input impedance

is reduced approximately to 0.5 megohm by the shunting effect of the GAIN CONTROL potentiometer R1.

The high gain of the amplifier is contributed by the two intermediate stages (Raytheon CK721). The push-pull class-B output stage also employs two Raytheon CK721 p-n-p transistors.

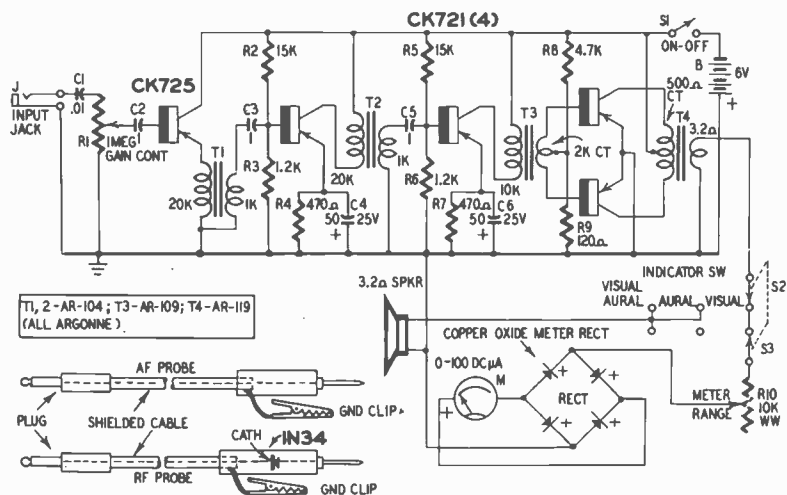


Fig. 923. Af-rf signal tracer.

The two-pole three-position indicator switch S2-S3 allows the operator to select visual or aural indication, or both. When this switch is set to its right position, the output-signal voltage is delivered only to the output meter circuit. When S2-S3 is at its center position, output is delivered only to the loudspeaker. And when S2-S3 is set to its left-hand position, output is delivered to both meter and speaker. The METER RANGE rheostat R10 allows the deflection to be restricted to the full-scale point of the meter scale when GAIN control R1 is operated wide open.

"Grid"-dip oscillator

Fig. 924-a shows the circuit of a conventional dip oscillator which covers the tuning range of 1 to 30 mc in four plug-in-coil bands: 1-3.5, 3.4-9, 8-20 and 18-30 mc. The word *grid* actually is a misnomer in the name of this transistorized instrument since no grid is present.

The oscillator circuit is designed around an rf transistor (Philco SB100). Individual transistors of this type may oscillate as high as 45 mc. However, all units should reach 30 mc with no difficulty.

The indicating meter is an rf voltmeter consisting of the 1N34A

diode, 0–100 dc microammeter M and the METER SET rheostat R1, capacitance-coupled across the tuned circuit through C3.

Table 8—Coil Table for "Grid"-Dip Oscillator

Coil A 1–3.5 mc	L1: 72 turns No. 32 enameled wire closewound on $\frac{3}{4}$ -inch-diameter form. L2: 18 turns No. 32 enameled wire closewound on same form as L1. Space $\frac{1}{16}$ inch from low end of L1.
Coil B 3.4–9 mc	L1: 19 turns No. 22 enameled wire closewound on $\frac{3}{4}$ -inch-diameter form. L2: 5 turns No. 22 enameled wire closewound on same form as L1. Space $\frac{1}{16}$ inch from low end of L1.
Coil C 8–20 mc	L1: 9 turns No. 22 enameled wire on $\frac{3}{4}$ -inch-diameter form. Space to winding length of $\frac{1}{2}$ inch. L2: 4 turns No. 22 enameled wire on same form as L1. Space to winding length of $\frac{3}{8}$ inch. Space $\frac{1}{16}$ inch from low end of L1.
Coil D 18–30 mc	L1: 3 turns No. 22 enameled wire on $\frac{3}{4}$ -inch-diameter form. Space to winding length of $\frac{1}{2}$ inch. L2: 3 turns No. 22 enameled wire on same form as L1. Space to winding length of $\frac{1}{4}$ inch. Space $\frac{1}{16}$ inch from low end of L1.

A tickler-feedback oscillator circuit is employed. Each plug-in coil consists of a tuned winding (L1) and tickler winding (L2) wound on a 0.75-inch-diameter plastic plug-in form. Amphenol

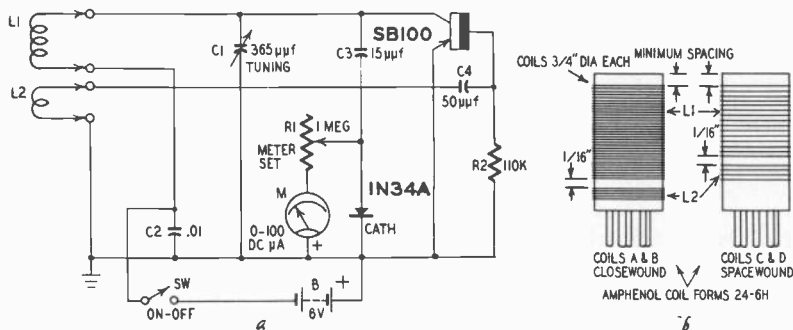


Fig. 924 -a, -b. "Grid"-dip oscillator.

24-6H forms have been specified for this instrument since they are obtainable in the required diameter. However, only four of the six prongs of these forms are used for coil connections.

Fig. 924-b shows constructional details of the coils. Note that each L1 winding is placed as close as possible to the top of the form

to facilitate inductive coupling into the external circuit under test. The coil table gives winding data.

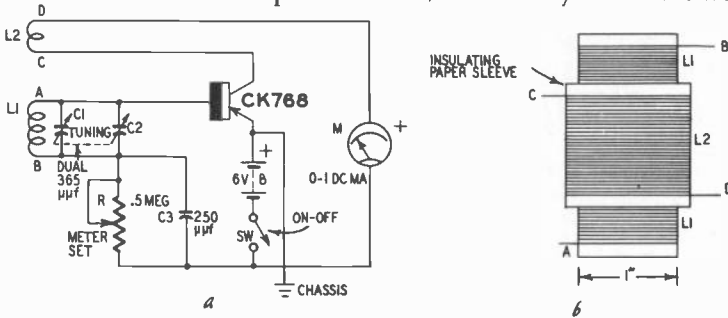
This dip oscillator is used in the regular manner. Potentiometer R1 is adjusted for full-scale deflection of the meter and must be reset as capacitor C1 is tuned through a band. A strong downward dip is obtained when resonance is established with the external circuit under test.

Low-frequency "grid"-dip oscillator ⁸

Broadcast and intermediate frequencies are covered by the single-range dip oscillator circuit shown in Fig. 925-a. This instrument has a tuning range of 350–1,700 kc without plug-in coils and in this range is convenient for the usual resonance, capacitance and inductance measurements afforded by "grid"-dip oscillators. An inexpensive rf transistor (Raytheon CK768) is used.

The tuning capacitor C1–C2 is a midget 365- μf dual variable with both sections connected in parallel to give a total capacitance of 730 μf . This unit must be insulated from the chassis or case since both the rotors and stators are above ground.

Fig. 925-b shows constructional details of the L1–L2 coil assembly. The tuned coil L1 is wound on a 1-inch-diameter form (1 $\frac{1}{4}$ inches long). An insulating sleeve of Scotch tape or insulating paper then is wound on top of this coil, and finally coil L2 is wound



L1: 113 turns No. 32 enameled wire closewound on 1-inch-diameter form.

L2: 42 turns No. 26 enameled wire closewound on insulating-paper or Scotch tape sleeve on top of L1. Wind in same direction as L1.

Fig. 925-a,-b. Low-frequency "grid"-dip oscillator.

on top of the sleeve. Both coils must be wound in the same direction. They must be connected in the circuit as shown by the letter symbols in Figs. 925-a,-b. Otherwise the feedback will not be of the correct polarity for oscillation.

"Grid"-dip adaptor

The circuit shown in Fig. 926-a permits the use of an rf test os-

cillator or signal generator as a "grid"-dip oscillator. The resonant frequency is read from the dial of the signal generator. Coil L in the adaptor is coupled to the external circuit under test, as when using a regular "grid"-dip oscillator, and dip is indicated by 0-1 dc milliammeter M. Rheostat R3 is the ZERO-SET control used to set the meter to zero in the absence of an input signal.

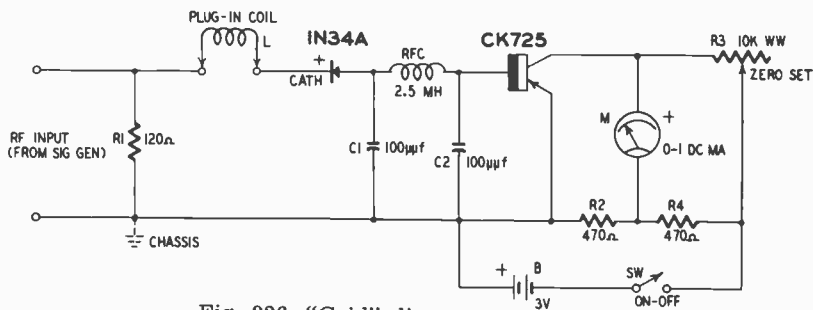


Fig. 926. "Grid"-dip adaptor.

The adaptor is connected to the rf output (modulated or unmodulated) of the signal generator through the RF INPUT terminals. The adaptor circuit may be built into a hand-type probe, with coil L inserted into its end and the rf connections made to the signal generator through a flexible, shielded cable.

The adaptor coil is untuned. All tuning is done with the signal generator. The generator output is adjusted for full-scale deflection of meter M. Three plug-in coils give the adaptor a frequency coverage of 100 kc to 250 mc in three ranges: 100 kc-6 mc, 5-35 mc and 30-250 mc. All coils are wound on 1-inch diameter plug-in forms. Table 9 gives coil-winding data for the adaptor.

Table 9—Coil-Winding Data for Adaptor

Coil A 100 kc to 6 mc	86 turns No. 32 enameled wire closewound on 1-inch diameter plug-in form.
Coil B 5 mc to 35 mc	11 turns No. 24 enameled wire closewound on 1-inch diameter plug-in form.
Coil C 30 mc to 250 mc	2½ turns No. 24 enameled wire on 1-inch diameter plug-in form. Space to winding length of 1/8 inch.

The adaptor circuit consists of the untuned coil L, a 1N34A germanium diode detector, rf filter C1-RFC-C2, high-gain dc amplifier and indicating microammeter M. High current amplification is obtained through the use of the super-alpha transistor (Ray-

theon CK725). The meter-zeroing circuit is a four-arm bridge.

Sensitive light meter

Most simple light meters (or exposure meters) composed of a self-generating photocell and dc microammeter have a full-scale

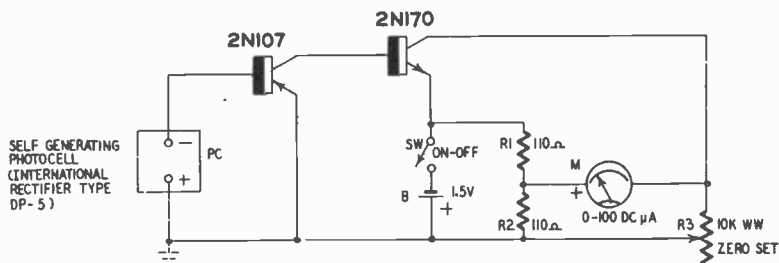


Fig. 927. Sensitive light meter.

sensitivity between 70 and 100 foot-candles. This sensitivity is inadequate for experimental low-intensity measurements.

Fig. 927 shows how a two-stage high-gain dc amplifier may be added to the simple light meter to increase its sensitivity. This circuit gives a full-scale meter deflection with 0.1-foot-candle light input.

Direct coupling is employed in the amplifier. To operate successfully from a single 1.5-volt cell B, transistors having opposite electrical characteristics are employed: one p-n-p (General Electric 2N107) in the input stage and one n-p-n (General Electric 2N170) in the output stage. The meter is connected in a bridge-type zero-set circuit. This circuit has good stability because of the drift of i_{co} in opposite directions in the two transistors.

In use, the meter is set to zero (by adjustment of rheostat R3) with the photocell darkened. The instrument then is ready for operation. The meter scale may be calibrated in foot-candles if a standard light source is available. The basic range 0.1 foot-candle may be multiplied by means of shunt resistors connected across the photocell (2N107 input circuit).

Instrument circuits in other chapters

The following instrument circuits have been described in other chapters under the headings of which they most logically fall:

Audio Oscillators. Chapter 4.

Variable-Frequency RF Oscillator. Chapter 4.

Frequency Standards. (100-kc Oscillators). Chapter 4.

Square-Wave Generators. (Multivibrator). Chapter 4. (Flip-Flop). Chapter 7.

Modulation Monitor. Chapter 10.

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- ³ *Transistorized TV Antenna Compass and Field-Strength Meter*, Radio & Television News, January, 1956; p. 43.
- ⁴ *Tunable 500-to-1000-Cycle Bandpass Filter*, Radio & Television News, August, 1956; p. 106.
- ⁵ (Guy Dexter). *Crystal Marker for TV Alignment*, Radio & Television News, July, 1956; p. 80.
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- ⁷ *Heterodyne Frequency Meter Uses Pair of Transistors*, RADIO-ELECTRONICS Magazine, October, 1953; p. 88.
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* Author is Rufus P. Turner.

amateur devices

IN the mass of circuits and data that appeared during the earliest years of the transistor's existence, the amateur radio operator found very little of direct use in his activities. This situation is changing, however, as new circuits are developed and as higher-frequency and higher-power transistors become available.

This chapter describes several circuits of interest chiefly to the radio amateur. Elsewhere in this book are other circuits also of interest to the ham but which fall logically into other chapters. Examples of these are 50-kc If Amplifier (Chapter 2); Code-Practice Oscillators, 100-kc Standard-Frequency Oscillators, High-Frequency Crystal Oscillator, Wide-Band Rf Oscillator (Chapter 4); All-Wave Regenerative Receiver (Chapter 6); Sensitive Rf Relays, Sound-Operated Relay, Carrier-Failure Alarm (Chapter 8) and many of the instrument circuits in Chapter 9.

Keying monitor

Fig. 1001 shows a simple monitor circuit for making CW signals audible to the transmitting operator. This monitor consists of an audio oscillator feeding a pair of headphones and operated from direct current obtained by rectifying the rf carrier picked up from the transmitter. No battery is needed. As long as the transmitting key is held down, rf is present, the transistor receives a dc operating voltage and a tone is heard in the headphones. In this way, the transmitted dots and dashes may be monitored.

The rf is picked up by the small pickup coil L which is placed near the final tank coil in the transmitter and is rectified by the

1N34A germanium diode. Rectified dc passes through the choke coil of a radio-frequency filter (C1-RFC-C2) and through the high-impedance winding of transformer T and the headphones in series to bias the collector of the p-n-p transistor (Raytheon CK722) negative.

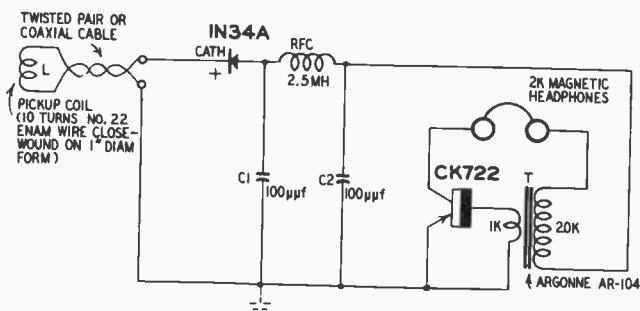


Fig. 1001. Keying monitor.

The oscillator circuit is a simple tickler-feedback arrangement with transformer T providing the regenerative feedback and also the impedance match between the high-impedance collector and low-impedance base. The transformer connections must be phased correctly for positive feedback, otherwise the circuit will not oscillate. If oscillation is not obtained readily, reverse the connections of either the primary or secondary winding of the transformer. The audio tone frequency will be approximately 2,000 cycles. It may be lowered, if desired, by connecting a capacitor in parallel with the 20,000-ohm winding.

Phone monitor

A simple diode detector connected to magnetic phones and rf pickup coil makes a convenient setup for the aural monitoring of radiophone signals. But the sensitivity of this combination is so low that the pickup coil must be coupled tightly to the transmitter, a practice that is dangerous both to the diode and the operator.

When an audio amplifier is added to the detector, the pickup coil may be removed to a safe distance from the transmitter circuit and, at the same time, the headphone signal is boosted. Fig. 1002 shows a phone monitor comprising a diode detector and two-stage transistor amplifier.

In this arrangement, the 1N34A diode acts as a demodulator. Its output develops an audio-frequency voltage across load resistor R1 which is bypassed for rf by capacitor C1. This af energy is

coupled through capacitor C2 to the first transistor. The first stage is direct-coupled to the second stage. Although direct coupling is employed, single-battery operation is made possible by the use of an n-p-n transistor (General Electric 2N170) in one stage and a p-n-p (General Electric 2N107) in the other.

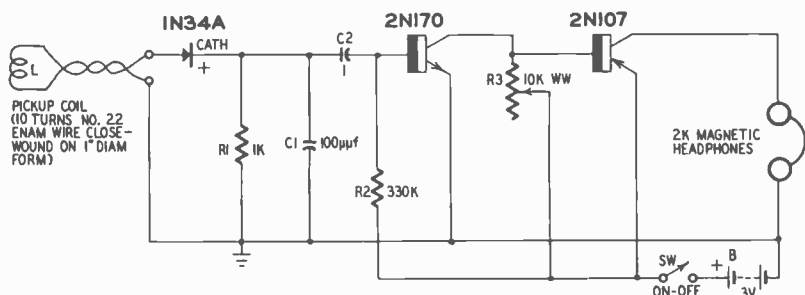


Fig. 1002. Phone monitor.

Speaker-type dynamic microphone

Fig. 1003 shows how a single-stage transistor amplifier may be used with a small PM dynamic loudspeaker to convert the latter into an inexpensive dynamic microphone with increased output. The complete amplifier may be built small enough to be mounted on the back of the speaker frame.

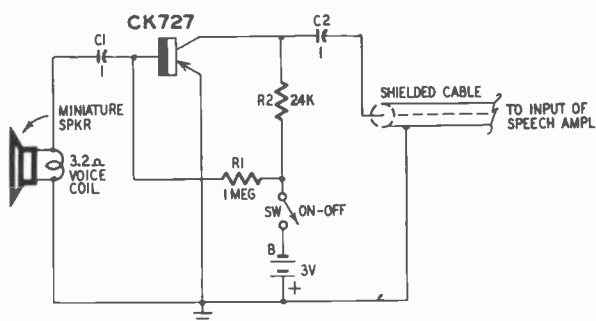


Fig. 1003. Speaker-type dynamic microphone.

Since this device is to be operated into the speech amplifier of a transmitter, its own noise level must be low. A low-noise type of transistor (Raytheon CK727) accordingly is used.

The battery drain is quite low ($75 \mu\text{a dc}$), so two 1.5-volt penlight cells connected in series as battery B will bias the transistor ade-

quately. Mercury cells provide slightly less than 3 volts but maintain their voltage throughout their service life.

Any size of loudspeaker is permissible in this device. The best voice quality will be obtained with the larger diameters. In most instances, however, size requirements will limit the choice to the 4- or 5-inch size.

Signal peaker (CW filter)

Fig. 1004 shows the circuit of a peaked (bandpass) audio amplifier for increased selectivity in cw reception. The center response frequency is 1,000 cycles but may be shifted several hundred cycles above or below this figure by adjustment of the tuning screw of inductor L (UTC type VIC-15).

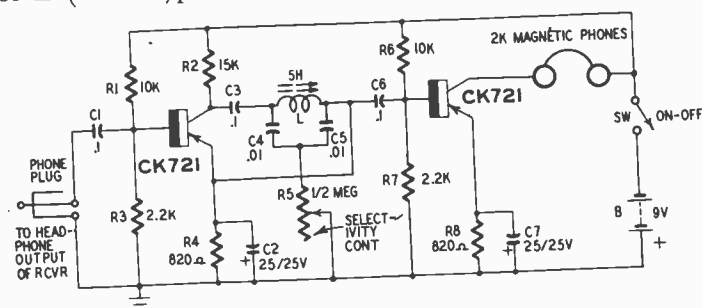


Fig. 1004. Signal peaker (CW filter).

The amplifier consists of two stages, each employing a Raytheon CK721 transistor. In the first stage, a bridged-T null network (C4-C5-L-R5) is connected as a negative feedback filter from the collector-output circuit back to the base-input circuit. This filter may be adjusted for sharp rejection of one frequency by varying the inductance of L and the resistance of R5. This suppresses one frequency from the feedback energy. The result is that the negative feedback cancels the first-stage gain on all frequencies but this one which is readily amplified.

The second stage serves the dual purpose of isolating the filter from the headphone output circuit and of providing power gain for the headphones.

To tune the amplifier initially: (1) Close switch SW. (2) Connect the input phone plug to the output of an audio oscillator set to 1,000 cycles. (3) Connect an ac vacuum-tube voltmeter in parallel with the headphones. (4) Using an Allen wrench, adjust the tuning screw of inductor L for peak deflection of the vtvm. (5) Adjust po-

tentiometer R5 for a sharper and higher-amplitude peak and read-just L. Reduce the output of the audio oscillator if the amplifier blocks (as evidenced by mushy headphone signals and sluggish operation of the meter).

When the circuit is being used, adjustment of R5 gives a degree of selectivity control, allowing the response to be broadened slightly to prevent ringing of the signal.

Field-strength meter

The field-strength meter circuit shown in Fig. 1005 covers the frequency range 1.8 to 68 mc in 5 bands: 1.8–4, 3.8–8.6, 8–18, 15–34 and 30–68. The first of these includes amateur 160 and 80; the second, amateur 40; the third, amateur 20; the fourth, amateur 10, 11 and 15; and the fifth, amateur 6 meters. Plug-in coils are used.

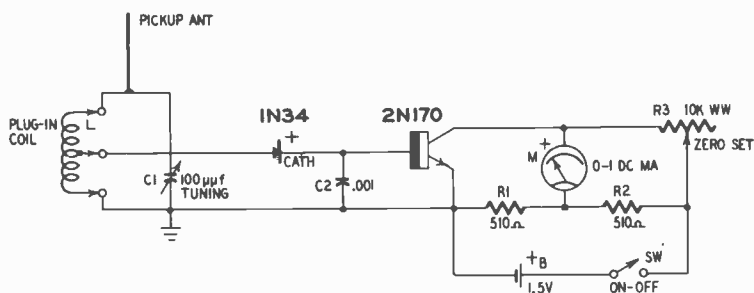


Fig. 1005. *Field-strength meter.*

General coverage tuning, rather than bandspread, has been employed, since this allows the instrument to be used for harmonic checking in transmitters, antennas and in interference tracing.

The instrument employs a single tuned circuit L-C1 operated from a pickup antenna which may be a short rod or whip. All coils are center-tapped for improved impedance match to the diode detector. They are wound on 1-inch-diameter plug-in forms. The coil table supplies winding instructions.

Dc output from the 1N34A diode detector is applied to the base-emitter input circuit of the transistor (General Electric 2N170) with the transistor base positive. The transistor provides dc amplification with a gain of approximately 15. The collector-current increase due to application of the input signal deflects the 0-1 dc milliammeter M. A bridge-type zero-set circuit is provided for the meter.

When using the field-strength meter do not close switch SW until

a coil is plugged in. The reason for this precaution is that the base input circuit is open when the coil is removed and this can cause violent deflection of the meter.

Table 10—Coil Table for Field-Strength Meter

Coil A 1.8—4 mc	57 turns No. 32 enameled wire closewound on 1-inch-diameter plug-in form. Tap 28th turn from ground end.
Coil B 3.8—8.6 mc	25 turns No. 26 enameled wire on 1-inch-diameter plug-in form. Space to winding length of $\frac{1}{2}$ inch. Tap 12th turn from ground end.
Coil C 8—18 mc	12 turns No. 22 enameled wire on 1-inch-diameter plug-in form. Space to winding length of $\frac{1}{2}$ inch. Tap 6th turn from ground end.
Coil D 15—34 mc	$5\frac{1}{2}$ turns No. 22 enameled wire on 1-inch-diameter plug-in form. Space to winding length of $\frac{1}{4}$ inch. Tap $1\frac{1}{2}$ turns from ground end.
Coil E 30—68 mc	$2\frac{1}{2}$ turns No. 22 enameled wire on 1-inch-diameter plug-in form. Space to winding length of $\frac{1}{4}$ inch. Tap $1\frac{1}{2}$ turns from ground end.

Modulation monitor

Fig. 1006 shows the circuit of an amplitude-modulation monitor possessing sufficient input-signal sensitivity to permit its operation at a distance from the transmitter. This removes the necessity for dangerous close coupling to the transmitter tank circuit. An important feature of this instrument is that it uses no transformers.

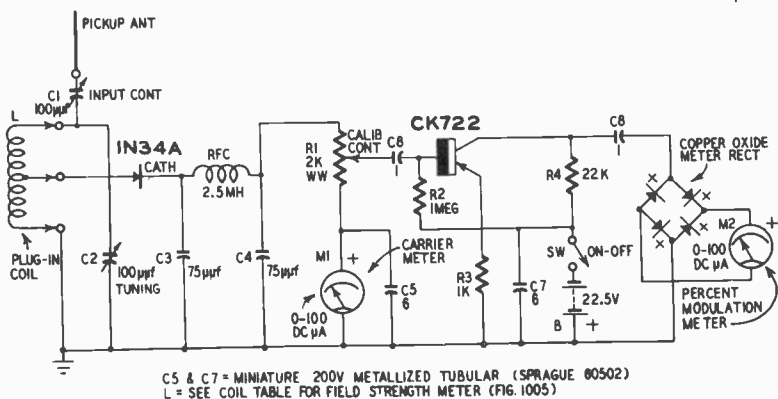


Fig. 1006. Modulation monitor.

The single tuned circuit (L—C2) is identical with the one employed in the field-strength meter previously described, and coil-

winding data may be obtained from the coil table for it. An additional variable capacitor C1 is provided for adjusting the amplitude of the rf input signal.

The dc output of the 1N34A diode rectifier passes through the choke coil of the radio-frequency filter (C3-RFC-C4) and load resistor R1 and 0-100 dc microammeter M1 in series. The deflection of M1 is proportional to the average amplitude of the carrier component of the AM signal. This deflection is adjusted exactly to full scale by means of variable capacitor C1.

The audio-frequency component (modulation envelope) of the AM signal develops an audio voltage across R1 and this voltage is presented to the transistor af amplifier (Raytheon CK722) through capacitor C6. The output signal of the amplifier deflects an audio voltmeter comprised of the copper oxide rectifier RECT and the 0-100 dc microammeter M2.

At any modulation percentage level, there is a corresponding ratio between the carrier and envelope amplitudes. At 100% modulation, for example, both amplitudes are equal. That is, when 100 μ a flows through R1 and is indicated by M1, an af drop of 0.2 volt peak is developed across R1. This is because the ac and dc components are equal. This af voltage is amplified by the transistor and deflects meter M2. The deflection of M2 may be set to full scale (100% modulation) by adjustment of R1. The scale of M2 is calibrated to read directly in PERCENT MODULATION.

The procedure in using the calibrated instrument is to tune in the signal by adjustment of C2 (with the proper plug-in coil L in place), then to adjust the deflection of the carrier meter M1 to full scale by adjustment of C1, and finally to read the modulation percentage from the percent modulation meter M2. Potentiometer R1 is set during the initial calibration and is not disturbed during normal use of the instrument.

Since the response of meter M2 is not linear at the low signal levels involved, the initial calibration of the instrument is best performed with the aid of an rf oscillator or signal generator having an adjustable amplitude-modulation percentage (0-100%) measured either with an oscilloscope or another calibrated modulation meter. The generator must be connected to the antenna and ground input terminals of the instrument. First set the generator for 100% modulation. Then tune in *on peak* the modulated signal by adjusting tuning capacitor C2. Next, adjust C1 for full-scale deflection of meter M1. Finally, adjust R1 for full-scale deflection of meter M2. From this point on, do not disturb the setting of R1. Make addi-

tional checks at 90%, 80%, 70%, 60%, 50%, 40%, 30%, 20%, 10% and at as many intermediate points as possible. At each of these modulation percentage levels, readjust C1, if necessary, for full-scale deflection of M1 and note the reading of M2. After a complete calibration run, a meter card reading in percent modulation may be prepared for meter M2 or a graph drawn for this instrument.

CW transmitter

The CW telegraph transmitter circuit shown in Fig. 1007-a has an rf output of only 27 mw but under favorable conditions it can give good account of itself.

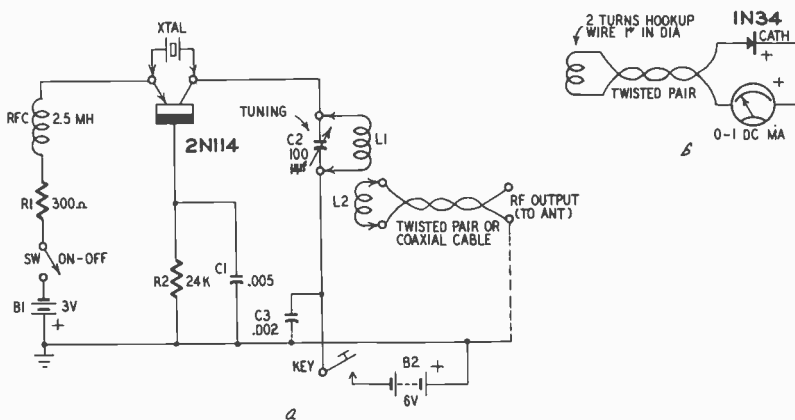


Fig. 1007-a,-b. CW transmitter.

The circuit is that of a simple 80- and 160-meter crystal oscillator employing an rf transistor (Raytheon 2N114). Two batteries are used: 3 volts (B1) for emitter bias and 6 volts (B2) for the collector. The current drain from each is approximately 10 ma.

Baby-type, commercial, end-link, plug-in transmitter coils (L1-L2) are employed. In each instance, L2 is the link coil. The 80-meter coil pair is a Barker & Williamson MEL-80 and the 160-meter pair a Barker & Williamson MEL-160. Low-impedance output (50 to 75 ohms) is provided for coupling to an antenna coupler or directly to the center of a dipole antenna.

Tuning in the conventional manner with a dc milliammeter in the oscillator circuit is not practicable with a transistorized oscillator. For tuning purposes, a crystal-type meter is shown in Fig. 1007-b. The small two-turn pickup coil is held close to coil L1 of the transmitter, and capacitor C2 turned for peak deflection of the 0-1 dc milliammeter.

Conelrad receiver

Fig. 1008 shows the circuit of a versatile receiver for Conelrad compliance. This receiver gives visual and aural indications and will operate a signal alarm. The circuit comprises a broadcast-band bandpass tuner, diode detector, two-stage dc amplifier, indicating meter and sensitive relay. Deflection of the 0–1 dc milliammeter M indicates the presence of the station carrier. Relay RY also is held in as long as the carrier is present. The station may be monitored aurally by means of headphones plugged into jack J. When the carrier is interrupted, the meter returns to zero and the relay drops out. The relay will operate a normally-closed relay in an external alarm device, such as a bell or horn circuit.

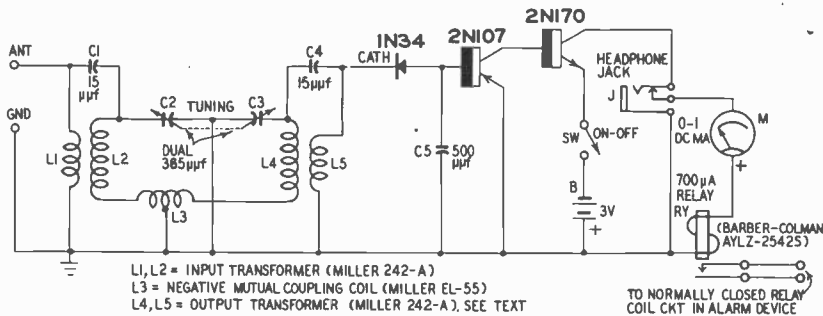


Fig. 1008. Conelrad receiver.

A bandpass tuner is employed since this gives maximum selectivity with diode detection. The tuner consists of the dual 365- μ F variable capacitor C2–C3, two radio-frequency transformers L1–L2 and L4–L5 and the negative mutual coupling coil L3. Transformer L1–L2 is connected in the normal manner with the low-impedance winding L1 as the primary. To match the low impedance of the 1N34 diode, however, the second transformer L4–L5 is connected into the circuit backward. That is, the low-impedance winding L5 is used as the secondary.

Dc output of the diode is amplified by the direct-coupled amplifier and actuates milliammeter M and relay RY in series. The first-stage transistor is a p-n-p unit (General Electric 2N107), while the second-stage unit is an n-p-n (General Electric 2N170).

Transistors in conventional amateur receivers

There are numerous points in standard, tube-type amateur re-

ceiver circuits where transistorized subcircuitry might be employed to advantage. Examples are: signal peakers, Q5ers, noise silencers, signal-strength indicators, received-signal modulation monitors, Q-multipliers, etc. which can be self-powered when transistorized and therefore place no added burden on the receiver power supply.

The amateur with experimental bent will find both education and diversion in transistorizing devices of this sort from their original tube circuits. Respectable power supply economies may be obtained by transistorizing as much as possible of the circuitry, especially in portable receivers.

miscellaneous circuits

THE circuits gathered together in this chapter are those which do not fit logically into the subject matter of the preceding chapters but are important enough to be included in this book. Their appearance at the end of the book certainly does not imply that they are superfluous.

Here, more than in examining the other chapters, a reader might discover that his peculiar needs have not been satisfied. It should be acknowledged, however, that many of the circuits we would like to transistorize do not yet lend themselves favorably to this modification. The author has tried to avoid those devices which in the present state of the art are mere laboratory curiosities. Such circuits might be reproduced *ad infinitum* but to the assured disservice of the reader.

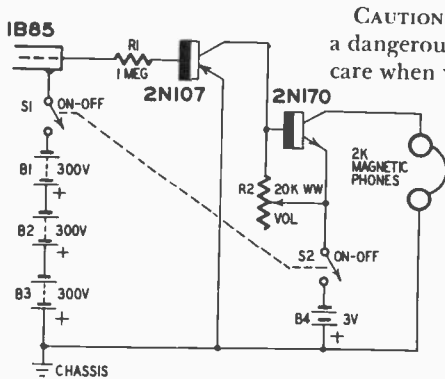
Additional practical applications of transistor circuitry may be expected to appear with regularity in the periodical literature and the best of these ideas undoubtedly will find their way into future editions of this book.

Simple Geiger counter

Fig. 1101 shows the circuit of a miniature Geiger counter employing a 1B85, 900-volt G-M tube and a two-stage transistor amplifier. This arrangement has excellent sensitivity.

The 900-volt dc polarizing potential for the counter tube is supplied by three miniature 300-volt batteries B1, B2, B3 (Burgess U200 or RCA VS093) connected in series. A 300-volt counter tube, such as type 1B86, also may be used in this circuit with one of the

300-volt batteries. Somewhat less sensitivity is obtained with this arrangement, however.



CAUTION: The 900-volt battery can deliver a dangerous electric shock. Use extraordinary care when working with this circuit.

Fig. 1101. Simple Geiger counter.

Output pulses from the counter tube are amplified by the direct-coupled amplifier employing a p-n-p transistor (General Electric 2N107) in the first stage and an n-p-n transistor (General Electric 2N170) in the second stage for single-battery (B4) operation.

Geiger counter with transistorized high-voltage supply

In the Geiger counter circuit shown in Fig. 1102 the counter-tube dc supply is a transistorized oscillator (employing a General Electric 2N107 transistor) with a miniature stepup transformer.

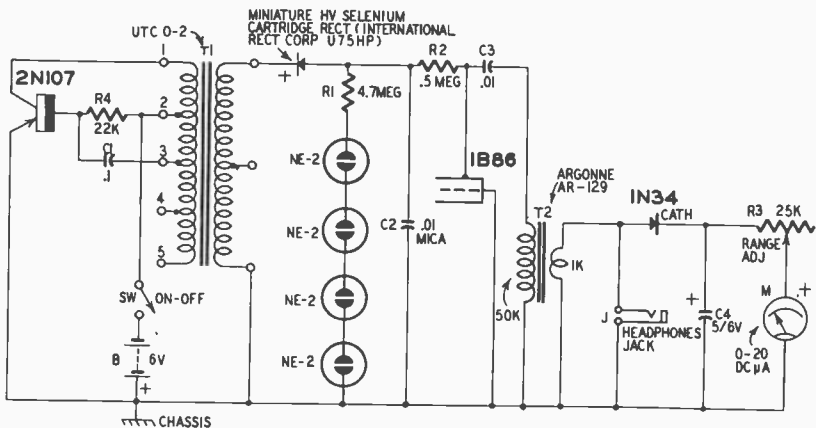


Fig. 1102. Geiger counter with transistorized high-voltage supply.

This dc supply, described originally by Thomas G. Knight (RADIO-ELECTRONICS Magazine, September, 1956) converts the 6-volt

battery potential to 300 volts of regulated dc for the 1B86 counter tube. It does this by using the 6 volts as collector bias for the transistor in a Hartley type oscillator in which the primary winding of the miniature transformer T1 is the tapped feedback coil required by this type of circuit. The oscillation sets up a high ac voltage across the secondary winding and this ac is rectified by a high-voltage cartridge type selenium rectifier RECT and filtered by capacitor C2. A compact voltage regulator is formed by the four series-connected NE-2 neon lamps and resistor R1.

Output pulses from the counter tube are delivered to the indicator circuit by the coupling transformer T2. Headphones may be plugged into jack J for aural indications. The 1N34 germanium diode rectifies the pulses and the resultant direct current deflects the 0-20 dc microammeter M for visual indications. By limiting the direct current through the meter, rheostat R3 serves as a RANGE control.

Phono oscillator

Fig. 1103 shows the circuit of a simple phonograph oscillator employing a single rf transistor (Raytheon 2N112). This circuit may be fix-tuned to any frequency in the broadcast band and is operated directly into the antenna and ground terminals of a radio receiver tuned to a clear channel.

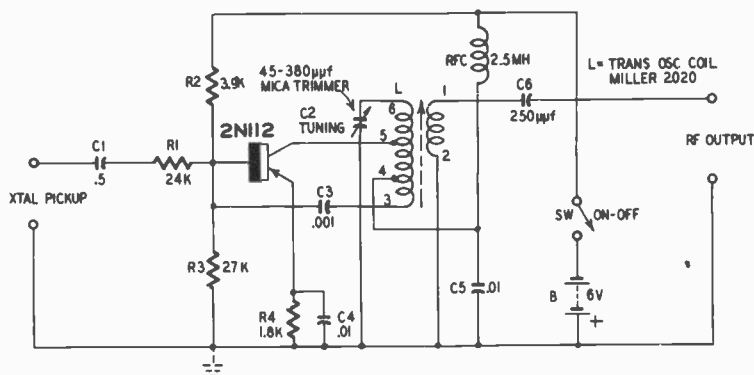


Fig. 1103. Phono oscillator.

A self-excited Hartley type oscillator circuit is used. The tapped oscillator coil assembly L is a conventional component (Miller 2020). The number symbols appearing on this coil in Fig 1103 correspond to the manufacturer's coding and the coil must be con-

nected as shown by these numbers, otherwise oscillation will not be obtained. Tuning is accomplished by means of a 45–380- μf compression-type mica trimmer capacitor C2 and the screw-adjusted slug in coil L.

The oscillator may be modulated directly by a crystal-type phonograph pickup connected to the CRYSTAL PICKUP terminals. A transformer-coupled preamplifier will be required for a reluctance-type pickup. Suitable amplifier circuits may be found in Chapter 1.

While maximum signal transfer will be obtained by connecting the RF OUTPUT terminals of the phono oscillator to the antenna and ground input terminals of the receiver, a sensitive receiver does not need this direct coupling. In this instance, a strong signal will be transmitted across a room from a short length of wire hung as an antenna from the top RF OUTPUT terminal.

Wireless microphone

This phono oscillator circuit has been modified (Fig. 1104) for use as a *wireless microphone*. This simple device may be used to “broadcast” into nearby radio receivers.

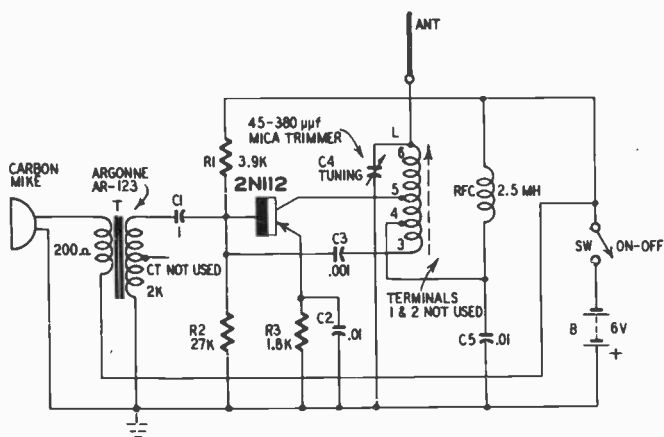


Fig. 1104. *Wireless microphone.*

Like the previous circuit, this arrangement uses a single rf transistor (Raytheon 2N112) in a Hartley type oscillator. Oscillator coil L must be connected exactly as shown in Fig. 1104 according to the manufacturer's number coding, otherwise oscillation will not be obtained. The output coupling coil (terminals 1 and 2) is not used in this circuit.

Tuning to any clear spot in the broadcast band is accomplished

by means of the compression type mica trimmer capacitor C4 and the screw-adjusted tuning slug in coil L.

A short vertical rod or whip antenna is sufficient for the short-range transmission permissible with this type of device. Good results have been obtained with a 2-foot length of No. 12 bare copper wire.

The carbon microphone MIKE is operated from the same 6-volt battery B that supplies the dc bias for the transistor. The battery may be four 1.5-volt flashlight cells connected in series.

Photoelectric intrusion alarm ¹

Fig. 1105 shows the circuit of a simple light-operated intrusion alarm employing semiconductor components throughout. The electric eye is a selenium photocell PC, the relay amplifier employs a transistor (Raytheon CK722) and the power supply uses two 1N34 germanium diodes.

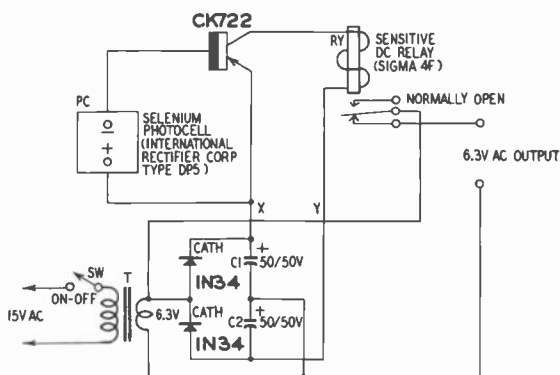


Fig. 1105. Photoelectric intrusion alarm.

As long as the photocell is illuminated by a light beam, the armature of relay RY is pulled away from the lower contact and the circuit to the output terminals is broken. When the cell is darkened, as when the light beam is interrupted by a person passing between the light source and the cell, the armature is released and returns to the lower contact, closing the output circuit. This applies 6.3 volts ac from transformer T to the output terminals, and this voltage may be used to operate a bell, horn or other alarm device. The circuit is fast-operating and will work when the cell is illuminated from an incandescent lamp or by daylight. An appropriate lens in front of the cell will concentrate the light rays for increased sensitivity.

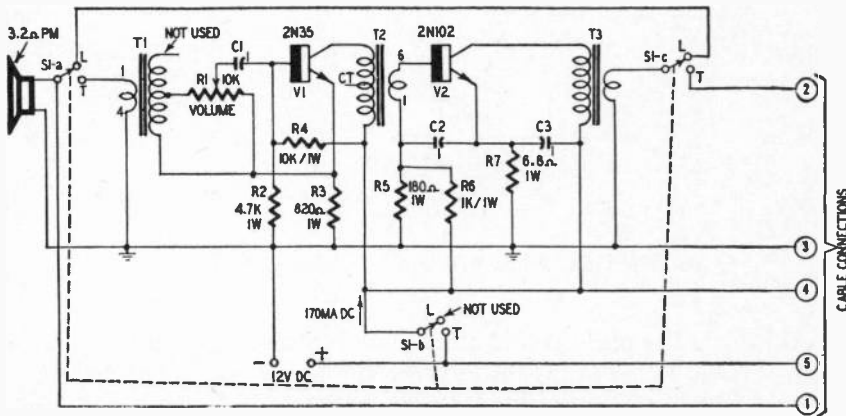
Dc operating voltage is furnished by a simple voltage-doubler type of power supply consisting of the 6.3-volt 1-ampere filament transformer T, two 1N34 germanium diodes and two 50- μ f 50-volt electrolytic capacitors C1 and C2. With 115 volts ac input, the dc output voltage of this power supply under full load is 17.5.

Battery operation may be employed if this is desired. Simply omit C1, C2, T and the 1N34's and connect a single 22.5-volt battery to leads X and Y. Connect the positive terminal of the battery to X and the negative to Y. Connect the ON-OFF switch SW in series with lead Y and the battery.

The dc relay is a Sigma type 4-F. This relay normally closes on 1.6 ma but can be adjusted to operate on currents as low as 1 ma by unscrewing its pivot screw a small amount.

Transistorized 0.6-watt intercom ²

Operated from a single 12-volt battery, the intercom unit of Fig. 1106 provides an audio power output of better than 0.5 watt. The high output is provided by a class-A amplifier employing a power transistor (Sylvania 2N102). The high-gain input amplifier stage uses a conventional transistor (Sylvania 2N35).



T1, primary impedance (as per diagram) 3 ohms, secondary (in circuit) 2,500 ohms (Merit A-2900, Stancor A-3823 universal output types, or equivalent); T2, primary impedance 12,000 ohms, secondary 32 ohms (same type as T1); T3, primary impedance 100 ohms at 150 ma, secondary 3.2 ohms (Acme T-24041).

Fig. 1106. 0.6-watt intercom.

While the class-A collector direct current of the 2N102 (150 ma) might appear high, this drain is present only during the relatively short intervals when the intercom is in operation. No current whatever is drawn during standby periods. In a tube-type intercom, 150 ma would be drawn *continuously* by the filaments alone. Eight 1.5-

volt size-D flashlight cells connected in series will power this circuit for over 1 month on a 2-hour-per-day *continuous* basis which is much more extreme than the actual intermittent operation demanded of an intercom. Economical battery operation makes the unit completely portable and divorces it from power-line complications.

The small PM dynamic speaker serves as a microphone when transmitting and as a loudspeaker when listening. A 3- or 4-inch speaker is adequate. The transmit-receive changeover is handled by the three-pole two-position lever switch S1 (Centralab 1457 with one section unused). This switch rests normally in its LISTEN (L) position where S1-a and S1-c connect the output transformer to the speaker. In the TALK position, S1-c disconnects the speaker from the amplifier output and S1-a connects it as a dynamic microphone to the amplifier input.

Fig. 1107 shows how two intercom units may be connected together with a five-wire cable. Changeover switch S1 at each station rests in its LISTEN position, rendering both intercom units dead. Section S1-b of either switch, however, closes the battery circuit to

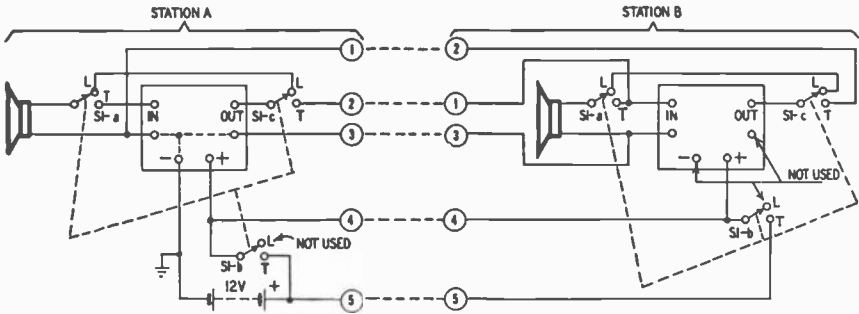


Fig. 1107. Typical two-station intercom installation.

both amplifiers when the switch is in its TALK position. When station A is transmitting, its loudspeaker is switched to the amplifier input as a microphone; at station B, the speaker is at the amplifier output and the output from Station A is applied to the input at station B. Battery voltage is applied to both units. Thus, A talks and B listens. When station B is transmitting, the opposite set of conditions results.

The circled numbers in Fig. 1107 refer to terminals of the cable plug and to cable leads. No. 1 at station A is connected to No. 2 of station B, 2 to 1, 3 to 3, 4 to 4 and 5 to 5. Additional stations may be

connected by running No. 1 of station A to No. 2 of the new station, 2 to 1, 3 to 3, 4 to 4 and 5 to 5. With this arrangement, any station can communicate with any other station.

Fig. 1108 shows a simpler though less effective system for short-distance operation. Here only one intercom amplifier is used and the remote installation (station B) consists only of a loudspeaker and changeover switch. The switches at each station in this setup are of the double-pole double-throw type. When station A is transmitting, speaker A becomes the microphone and speaker B is switched to the output circuit of the amplifier. When station B is transmitting, speaker B is switched to the amplifier input to serve as the microphone while speaker A is transferred to the amplifier output.

In either of the installations (Fig. 1107 or 1108), only one battery is needed and can be installed at any of the stations which is most convenient.

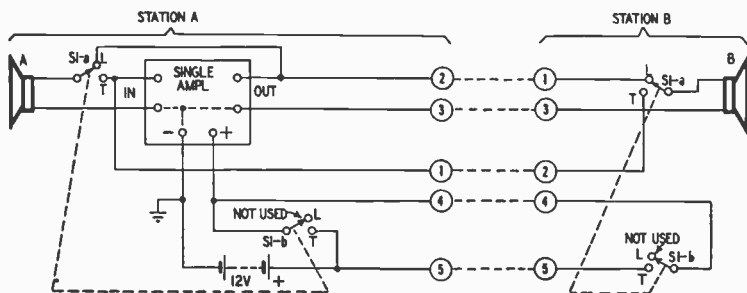


Fig. 1108. Economy two-station short-distance installation.

Hearing aid

Fig. 1109 shows the circuit of a hearing aid having good sensitivity and output. The three-stage amplifier in this unit employs a low-noise transistor (Raytheon CK727) in the input stage and high-alpha units (Raytheon CK721) in the intermediate and output stages. Transformer coupling is used for maximum overall power gain.

Readily obtainable subminiature components are employed throughout. Switch SW is integral with the specified VOLUME CONTROL potentiometer R4. The low-impedance microphone is connected directly into the base-emitter input circuit of the first transistor and, similarly, the magnetic earpiece is connected directly into the collector circuit of the output transistor.

The dc drain from the 1.5-volt cell B is approximately 2.35 ma. Bias stabilization is supplied by voltage-divider networks R2–R3

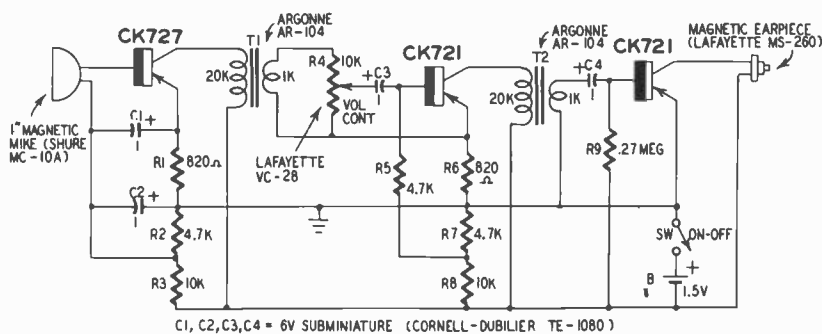


Fig. 1109. Hearing aid.

and R7–R8. Some adjustment of resistance R9 may be necessary for maximum output vs minimum distortion.

Diode-type transistor power supplies ³

Because of the low current drain of conventional transistors, germanium diodes may be employed as rectifiers in ac-line-operated power supplies used to furnish dc to these transistors when battery operation is not desired. Because of the low voltage requirements of transistors, such power supplies may employ small filament transformers. It is a simple matter to connect diodes and capacitors together to form voltage-multiplier circuits when the required dc voltage is higher than the rms voltage of the filament winding of the transformer.

Fig. 1110 shows several power supply circuits employing a filament transformer, diodes and capacitors. In each circuit, T is a 1-ampere 6.3-volt transformer. The dc output characteristics (for an ac input of 117 volts rms) are shown by the curves in Fig. 1111.

These circuits will be satisfactory as they appear in Fig. 1110 for most low-drain circuits using one or two transistors operated at fixed voltages. For better voltage regulation (less fall-off of voltage as the dc output current increases) and higher output current (such as will be required for multistage transistor circuits containing bias-resistor networks), General Electric type 1N91 miniature germanium power rectifiers may be substituted for the 1N34 diodes and the capacitor values increased. A 1N91 in the simple half-wave circuit (Fig. 1110-a) with C equal to 100 μ f, for example, gives 60

ma output at 6.3 volts dc contrasted to the 5 ma at 6.3 volts obtained with the 1N34 and 10 μ f.

Safety diode for transistor circuits ⁴

Accidentally reversing the battery in transistorized equipment often means quick and complete ruin of the transistors. Unless an

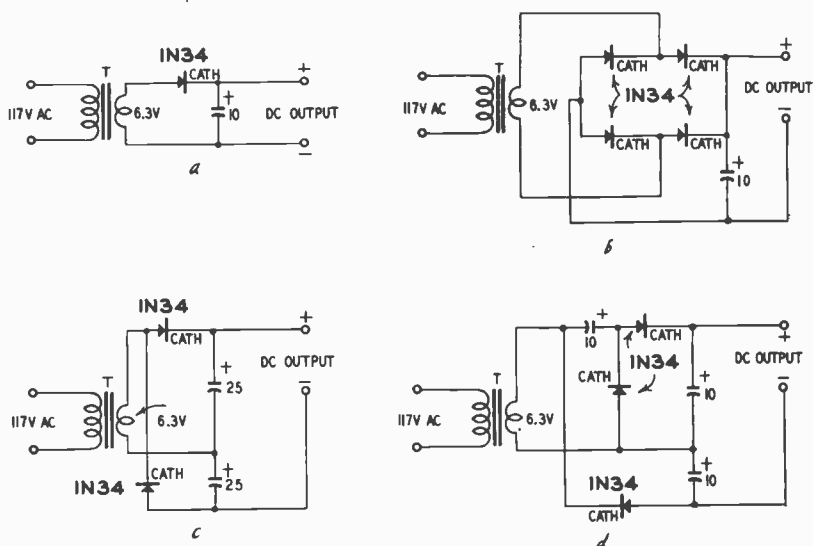
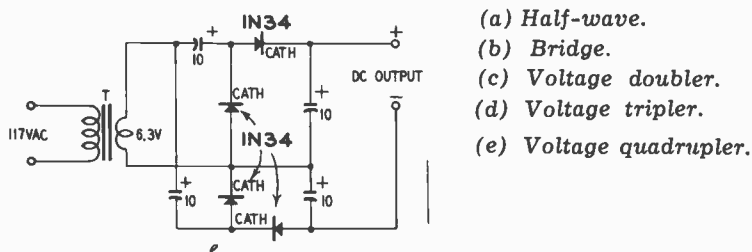


Fig. 1110 (a to e). Diode-type transistor power supplies.



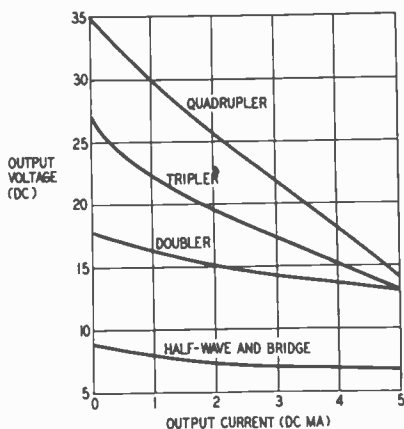
operator is on the alert constantly, this mistake is almost inevitable.

In most transistorized devices, such accidents can be prevented simply by connecting a germanium or silicon-junction diode in series with one of the dc input terminals of the device (Fig. 1112-a). The diode acts as a polarity-sensitive switch which acts instantaneously and requires neither attention nor adjustment.

The diode is poled so that only an extremely small current flows when the battery is reversed. This current is too minute to damage the transistors. As shown in Figure 1112-a, the anode of the diode is

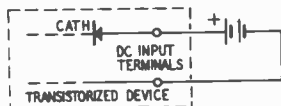
connected to the positive dc terminal. This polarity allows high forward current to flow through the diode to the circuit. When the battery is reversed, however, the anode becomes polarized negative

Fig. 1111. Performance of diode-type power supplies.



and the diode resistance increases—often several thousand times higher than the forward resistance. The tiny reverse current passed by the diode in this condition can cause no damage to the transistors.

Fig. 1112. Diode protector for transistor circuits.



Total Current Drain	Voltage Loss due to Diode
20 μ a	0.05
100 μ a	0.10
200 μ a	0.15
500 μ a	0.22
1 ma	0.30
2 ma	0.40
3 ma	0.50
4 ma	0.60
5 ma	0.65
10 ma	0.90
20 ma	2.0
30 ma	2.5
40 ma	3.1
50 ma	4.0

The natural internal resistance of the diode causes a voltage drop across the diode when the latter is carrying current. The voltage at the transistor circuit therefore is somewhat less than the battery voltage and is equal to the battery voltage minus the diode voltage drop. When this voltage drop is appreciable, the battery voltage

may be boosted by this amount to offset the loss. Figure 1112-b shows the approximate voltage lost across a 1N34 diode at various common transistor currents. The loss will be lower with low-resistance diodes, such as type 1N455, and with gold-bonded types.

The safety-diode scheme usually is unsatisfactory in class-B amplifiers because the collector current in these circuits swings (on signal excitation) from a few microamperes up to several milliamperes. The diode loss accordingly is both large and fluctuating and it will reduce power output and increase distortion of the class-B amplifier.

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* Author is Rufus P. Turner.

index

A	
Ac Relay, Sensitive	92
Ac Vacuum-Tube Voltmeter, Millivolt Adaptor for	113
Adaptor: for Ac Vacuum-Tube Voltmeter, Millivolt	113
"Grid"-Dip	131
Heavy-Duty, Constant-Current	66
Light-Duty, Constant-Current	65
Af-Rf Signal Tracer	128
Aid, Hearing	152
Alarm, Carrier Failure	102
Alarm, Photoelectric Intrusion	149
Aligner, Crystal-Type Set	121
All-Wave Regenerative Receiver	79
Alpha-Beta Conversion	43
Amateur Devices	135
Amplifier: Cascaded P-n-p Current	39
Cascaded P-n-p-N-p-n Current	40
Class-A	15
Class-B	16
Complementary-Symmetry Power with Conventional Transformers	21
Loudspeaker, Class-A with Power Transistor	20
Loudspeaker, Class-B with Conventional Transistors	18
Loudspeaker, Class-B with Power Transistors	19
Multistage R-C-Coupled	13
Multistage Transformer-Coupled	14
Power, Complementary-Symmetry with Power Transistors	22
R-C-Coupled, Multistage	13
Single-Stage Current	37
Single-Stage Dc Voltage	41
Single-Stage Power-Type Current	42
Single-Stage, R-C-Coupled, Common- Base	7
Single-Stage, R-C-Coupled, Common- Collector	11
Single-Stage, R-C-Coupled, Common- Emitter	9
Single-Stage, Single Tuned 455-Kc If	30
Single-Stage, Transformer-Coupled, Common-Base	8
Single-Stage, Transformer-Coupled, Common-Collector	12
Single-Stage, Transformer-Coupled, Common-Emitter	10
Two-Stage, 455-Kc If	31
Two-Stage, High Gain, Power-Type Current	42
Amplifiers: Audio	7
Dc	37
If	29
Rf	29
Rf and If	29
Antenna Compass for TV	114
Audio: Amplifiers	7
Oscillator, Phase-Shift	48
Oscillator, Single-Frequency	45
Oscillator, Step-Type Variable- Frequency	49
Oscillator, Tuned, Variable- Frequency	49
Signal Injector	120
Voltmeter-Millivoltmeter	111
B	
Blocking Oscillator	55
Booster, Heavy-Duty Relay	95
Boosting Transistor Input Impedance	25
Bridge Null Detector	117
Broadcast: Band Rf Amplifier and Preselector	29
Receiver, Loop-Operated, Diode- Transistor	75
Receiver, Push-Pull Transistor	74
Receiver, Regenerative	72
Receiver, Single-Transistor	69
Receiver, Step-Tuned, Diode- Transistor	73
Receiver, Sun-Powered	76
Receiver, Superheterodyne	77
Receiver, Two-Transistor	71
Tuners, Miniaturization of Simple	71
Two-Transistor Receiver	71
C	
Carrier Failure Alarm	102
Cascaded: Dc Voltage Amplifier	41
P-n-p Current Amplifier	39
P-n-p-N-p-n Current Amplifier	40
Circuit, Pulse Delay	85
Circuit, Simple Coincidence	86
Circuits, Miscellaneous	145
Circuits, Push-Pull Output	15
Class-A: Amplifier	15
Loudspeaker Amplifier with Power Transistor	20
Push-Pull Output Circuits	15
Class-B: Amplifier	16
Loudspeaker Amplifier with Conventional Transistors	18
Loudspeaker Amplifier with Power Transistors	19
Push-Pull Output Circuits	16
Code-Practice Oscillator	46
Code-Practice Oscillator, Loudspeaker- Operating	47
Coincidence Circuit, Simple	86
Common: Base Amplifier, Single-Stage, R-C- Coupled	7
Base Amplifier, Single-Stage, Trans- former-Coupled	8
Collector Amplifier, Single-Stage, R-C-Coupled	11
Collector Amplifier, Single-Stage, Transformer-Coupled	12

Emitter Amplifier, Single-Stage, R-C-Coupled	9
Emitter Amplifier, Single-Stage, Transformer-Coupled	10
Compass, TV Antenna	114
Complementary:	
Symmetry Power Amplifier with Conventional Transformers	21
Symmetry Power Amplifier with Power Transistors	22
Symmetry-Type Phase Inverter	16
Conelrad Receiver	143
Constant:	
Current Adaptor, Heavy-Duty	66
Current Adaptor, Light-Duty	65
Current, Light-Duty Adaptor	65
Control Devices	89
Control, Miniature-Motor	100
Conventional Transformers with Comple- mentary-Symmetry Power Amplifier	21
Conversion, Alpha-Beta	43
Counter:	
Impulse	103
Photoelectric	98
Simple Geiger	145
with Transistorized High-Voltage Supply, Geiger	146
Crystal:	
High-Frequency Oscillator	52
Oscillator, 100-Kc	50
Oscillator, High-Frequency	52
-Type Set Aligner	121
-Type Sound Marker Generator	119
Cueing Receiver, High-Frequency	80
Current Amplifier:	
Cascaded P-n-p	39
Cascaded P-n-p-N-p-n	40
Single-Stage	37
Single-Stage, Power-Type	42
Two-Stage, High-Gain, Power-Type with Fixed-Resistance Input, Single- Stage	38
CW Filter, Signal Peaker	138
CW Transmitter	142
D	
Dc:	
Amplifiers	37
Heavy-Duty Voltage Regulator	64
Light-Duty Voltage Regulator	63
Microammeter, Low-Gain Single Transistor	106
Microammeter, Supersensitive	109
Milliammeter, High-Range Sensitizer for	114
Power Supply, Full-Wave Variable ..	61
Power Supply, Variable High- Current	60
Power Supply, Variable Low- Current	59
Relay, Sensitive	89
Relay, Ultra-Sensitive	90
-to-Dc, High-Voltage Power Supply ..	67
Voltage Amplifier, Cascaded	41
Voltage Amplifier, Single-Stage	41
Voltage Regulator, Heavy-Duty	64
Voltage Regulator, Light-Duty	63
Voltmeter, Electronic	110
Detector, Bridge Null	117
Devices, Amateur	135
Devices, Control	89
Diode:	
Protector for Transistor Circuits ...	155
-Type Power Supplies, Performance of	155
-Type Transistor Power Supplies	153
Dynamic Microphone, Speaker-Type	137
E	
Electronic Dc Voltmeter	110

F	
Field-Strength Meter	139
Field-Strength Meter, TV	115
Fixed-Resistance Input, Single-Stage Current Amplifier with	38
Flip-Flop	81
Frequency Meter, Heterodyne	125
Full-Wave Variable Dc Power Supply ..	61
G	
Geiger Counter, Simple	145
Geiger Counter with Transistorized High-Voltage Supply	146
General-Purpose Radio Control Relay ...	101
Generator, Crystal-Type Sound Marker ..	119
Generator, Rf Signal	122
"Grid":	
-Dip Adaptor	131
-Dip Oscillator	129
-Dip-Oscillator, Low-Frequency	131
H	
Hearing Aid	152
Heavy Duty:	
Constant-Current Adaptor	66
Dc Voltage Regulator	64
Relay Booster	95
Relay Booster, Sensitized	95
Heterodyne Frequency Meter	125
High-Current Dc Power Supply, Variable High-Frequency:	60
Crystal Oscillator	52
Cueing Receiver	80
Rf Amplifiers, Transistors for	34
High-Gain Dc Microammeter	108
High-Gain, Power-Type, Two-Stage Cur- rent Amplifier	42
High-Impedance Input, Preamplifier with High-Range Dc Milliammeter, Sensitizer for	23
High-Voltage Dc-to-Dc Power Supply ...	67
High-Voltage Supply, Geiger Counter with Transistorized	146
I	
If Amplifier:	
50-Kc	33
Single-Stage, Single-Tuned, 455-Kc If	30
Two-Stage, 455-Kc	31
If Amplifiers	29
If and Rf Amplifiers	29
Impulse Counter	103
Injector, Audio Signal	120
Input:	
and Emitter Resistance vs. Voltage Gain	27
Impedance, Boosting Transistor	25
Preamplifier with High-Impedance ..	23
Instruments, Test	105
Intercom Installation, Typical Two- Station	151
Intercom, Transistorized 0.6-Watt	150
Intrusion Alarm, Photoelectric	149
Inverter, Pulse	87
Inverters, Phase	16
K	
Keying Monitor	135
L	
Light-Duty:	
Constant-Current Adaptor	65
Contacts, Sensitive Switching Circuit for	91
Dc Voltage Regulator	63
Light Meter, Sensitive	133
Light-Operated Relay	97
Light-Powered Oscillators	56

Loop-Operated Diode-Transistor Broadcast Receiver	75
Loudspeaker Amplifier:	
Class-A, with Power Transistor	20
Class-B, with Conventional Transistors	18
Class-B, with Power Transistors with Conventional Transistors, Class-B	19
with Power Transistor, Class-A	20
with Power Transistors, Class-B	19
Loudspeaker-Operating Code-Practice Oscillator	47
Low-Current Dc Power Supply, Variable	59
Low-Frequency "Grid"-Dip Oscillator	131
Low-Gain Single-Transistor Dc Microammeter	106
Low-Noise Preamplifier, Single-Stage	23

M

Meter:	
Field-Strength	139
Heterodyne-Frequency	125
Sensitive Light	133
Sound Level (Noise)	127
TV Field-Strength	115
Microammeter:	
High-Gain Dc	108
Low-Gain Single Transistor Dc	106
Supersensitive Dc	109
Zero-Setting Circuits for	107
Microphone, Speaker-Type Dynamic	137
Microphone, Wireless	148
Milliammeter, Sensitizer for High-Range Dc	114
Millivolt Adaptor for Ac Vacuum-Tube Voltmeter	113
Millivoltmeter, Audio Voltmeter	111
Miniature-Motor Control	100
Miniaturization of Simple Broadcast Tuners	71
Miscellaneous Circuits	145
Modulation Monitor	140
Monitor:	
Keying	135
Modulation	140
Phone	136
Motor Control, Miniature	100
Multistage R-C-Coupled Amplifier	13
Multistage Transformer-Coupled Amplifier	14
Multivibrator	54
Multivibrator, One-Shot	83

N

Noise Meter, Sound-Level	127
Null Detector, Sharply Tuned	*118

O

One-Shot Multivibrator	83
Oscillator:	
Audio, Phase-Shift	48
Audio, Step-Type, Variable-Frequency	49
Blocking	55
Code-Practice	46
Crystal, 100-Kc	50
"Grid"-Dip	129
High-Frequency Crystal	52
Loudspeaker-Operating Code-Practice	47
Low-Frequency "Grid"-Dip	131
Operating Notes	57
Phono	147
Self-Excited, 100-Kc	51
Self-Excited, Rf	53
Single-Frequency Audio	45
Wide-Range Rf	53
Oscillators	45
Oscillators, Audio, Single-Frequency	45
Oscillators, Light-Powered	56

Output Circuits, Push-Pull	15
----------------------------------	----

P

Peaker, Signal (CW Filter)	138
Performance of Diode-Type Power Supplies	155
Phase Inverters, Complementary-Symmetry Type	16
Phase Inverters, Single-Transistor Type	16
Phase-Shift Audio Oscillator	48
Phone Monitor	136
Phono Oscillator	147
Photoelectric Counter	98
Photoelectric Intrusion Alarm	149
Phototimer	99
P-n-p Current Amplifier, Cascaded Power Amplifier:	39
Complementary-Symmetry, with Power Transistors	22
with Conventional Transformers, Complementary-Symmetry	21
with Power Transistors, Complementary-Symmetry	22
Power Supplies	59
Power Supplies, Diode-Type Transistor	153
Power Supply, High-Voltage Dc-to-Dc	67
Power Transistor Amplifier, Loudspeaker, Class-A	20
Power Transistors, Class-B, Loudspeaker Amplifier	19
Power Transistors, Complementary-Symmetry Power Amplifier with	22
Power-Type Current Amplifier, Single-Stage	42
Power-Type Current Amplifier, Two-Stage, High-Gain	42
Preamplifier, Single-Stage, Low Noise	23
Preamplifier with High-Impedance Input	23
Preselector, Broadcast-Band Rf Amplifier and	29
Pulse Delay Circuit	85
Power Inverter	87
Push-Pull Output Circuits	15
Push-Pull Transistor Broadcast Receiver	74

R

Radio:	
Control Relay, General-Purpose	101
Frequency Comparator	125
Receivers	69
R-C-Coupled:	
Amplifier, Multistage	13
Common-Base Amplifier, Single-Stage	7
Common-Collector Amplifier, Single-Stage	11
Common-Emitter Amplifier, Single-Stage	9
Receiver:	
All-Wave Regenerative	79
Conelrad	143
High-Frequency Cueing	80
Loop-Operated Diode-Transistor Broadcast	75
Push-Pull Transistor Broadcast	74
Regenerative Broadcast	72
Step-Tuned, Diode-Transistor Broadcast	73
Superheterodyne Broadcast	77
Two-Transistor Broadcast	71
Receivers, Radio	69
Regenerative Broadcast Receiver	72
Regenerative Receiver, All-Wave	79
Regulator, Heavy-Duty, Dc Voltage	64
Regulator, Light-Duty, Dc Voltage	63
Relay:	
Booster, Sensitized, Heavy-Duty	95
General-Purpose Radio Control	101
Heavy-Duty Booster	95
Light-Operated	97
Sensitive Ac	92

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