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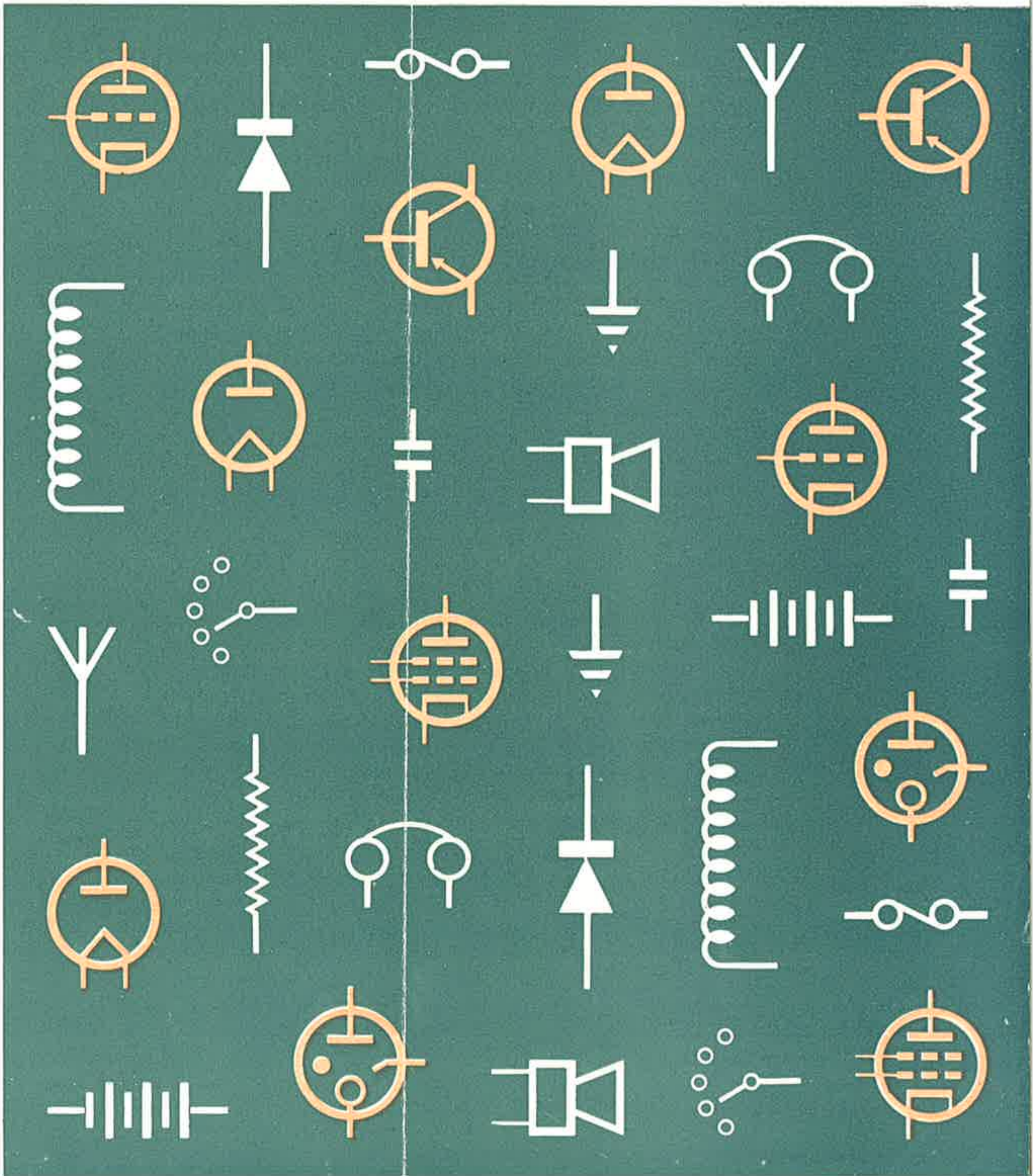

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RADIOTRONICS

Vol. 25, No. 2, 1960

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A Transistorized QSO-Getter

BY

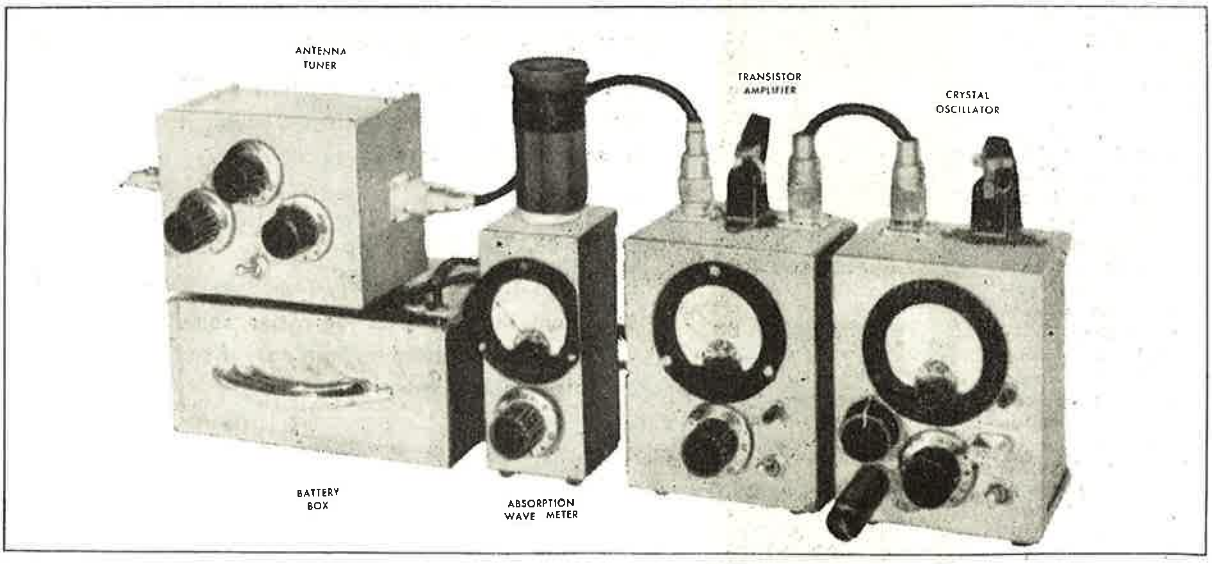
E. M. WASHBURN

W2RG

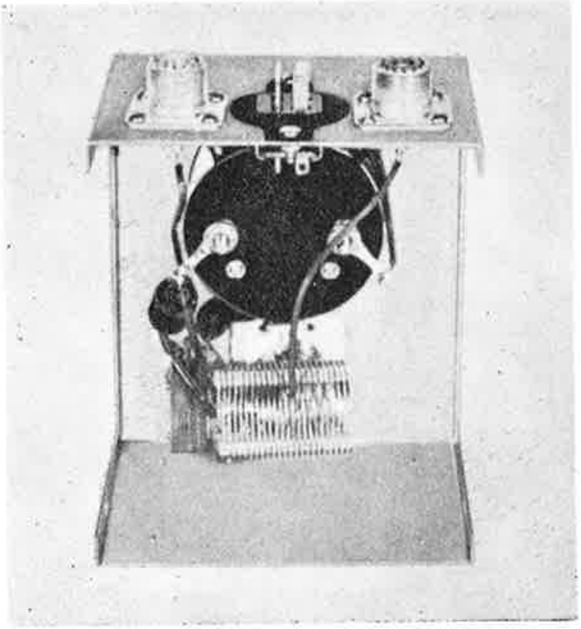
Many radio amateurs have expressed a keen interest in the amazing possibilities of low-power transistorized transmitters. This expressed interest has prompted the following description, so that others may join the growing ranks of QRP operators working hundreds — or even thousands — of miles on a fraction of one watt input.

The transistorized QRP transmitter illustrated and described in this article is essentially a 40-metre cw rig, using one 2N219 or 2N140 transistor in the crystal oscillator and another in the amplifier. The transmitter is adequately powered by two 6-volt, heavy-duty dry batteries, connected in series, which are provided with a switch to permit tuning up at 6 volts. When the transmitter is operating at full load, the crystal oscillator operates at 12 volts with a collector current of 15 milliamperes, while the amplifier operates at 12 volts, 18 milliamperes. Admittedly, these inputs are in excess of the manufacturer's ratings and some transistors may not operate satisfactorily under these overload conditions. A 1.5-volt dry cell is used in the oscillator emitter circuit as shown in the schematic diagram.

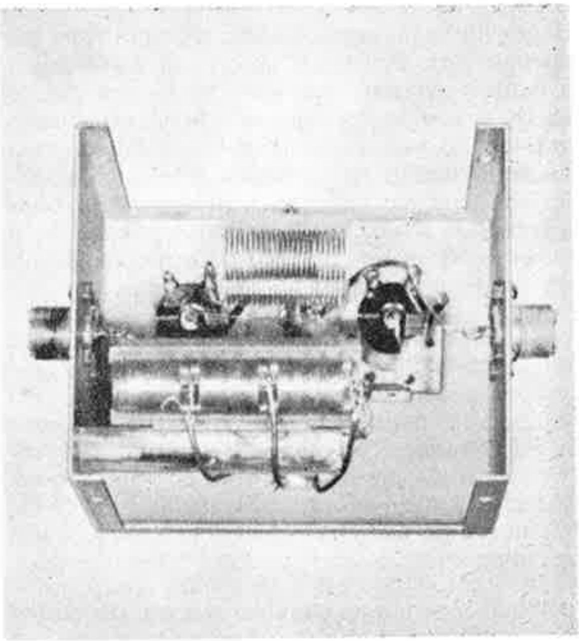
With this equipment, the author worked 18 states, Ontario, Quebec, Puerto Rico, Windward Islands and Transvaal, S. Afr.—all on 40 metres —with an antenna consisting of a single, 106-foot wire. The wire used with the author's transmitter is strung 28 feet across the basement rafters, then



Complete Transistor Rig. Maximum Input to the Final Amplifier: 216 mw.



Rear View of Transistor Amplifier.



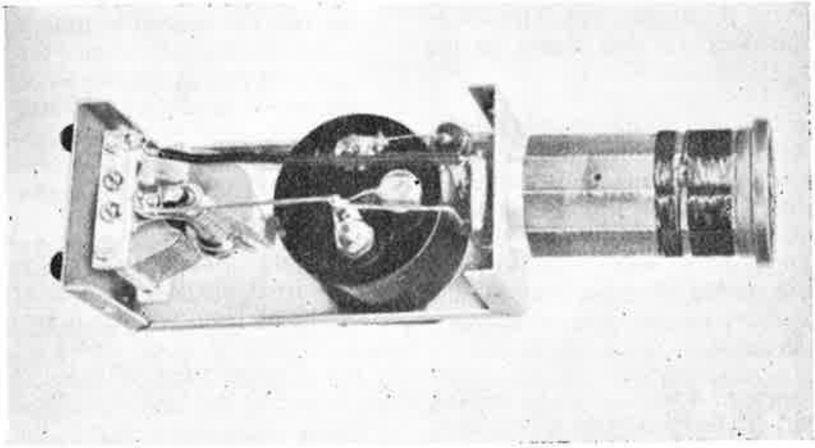
Interior View of Antenna Tuner.

leaves the confines of the shack and slopes upwards for a distance of 42 feet to a flat top which is 36 feet long and 28 feet above the ground. The antenna can be voltage-fed from the amplifier or it can be fed from the antenna tuner.

Over a long period of operations, the signal reports received by the author have varied from RST-339 to 589, depending upon band conditions, distance and the type of receiver used by the receiving station. The QSO with Transvaal, S. Afr., (ZS6TR) appears to be a world record for a 40-metre low power/transistor transmitter and the contact was made without any form of prearrangement and without any previous communications using a higher-power rig. At 216 milliwatts and covering a distance of 8,000 miles, this performance is comparable to 37,000 miles per watt at a frequency of 7002 Kc.

Several contacts have been made on the 80-metre band, but the most gratifying and successful results have been accomplished in the 40-metre band. To the present, no attempt has been made to put the QRP transmitter on the 20-metre or the higher frequency bands; however, this band could be worked by using the amplifier as a doubler stage and substituting 2N219 or 2N247 transistor for the 2N140.

The complete transistor transmitter station comprises five units as shown in the photograph. The wave meter would normally be placed several feet



Construction Details of Absorption Wavemeter Used to Indicate Maximum Radiation from Antenna.

away with its pickup coil about 2 inches from the antenna wire. Since this absorption wave meter is entirely conventional and the battery box is merely a housing for the two 6-volt dry batteries and the 1.5 volt dry cell, the circuit description will be limited to the crystal oscillator unit, amplifier and antenna tuner. Each of these three units is housed in a box which measures 5 inches by 4 inches by 3 inches. Interior construction details are shown in the photograph of each.

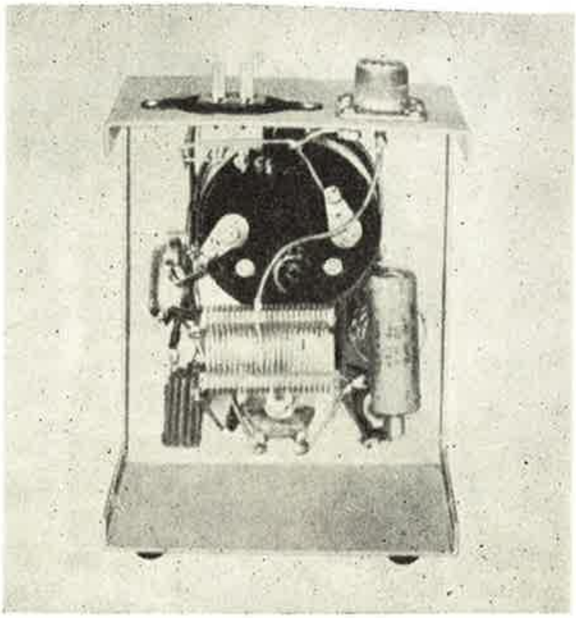
Although VFO circuits have been tried, the only successful operation has been with crystal control, and in this particular design the crystal unit is in the emitter circuit. The key is by-passed by a low-voltage 2μf capacitor to improve keying characteristics, particularly when a "bug" is used. The most critical adjustment is the location of the output tap on the base inductor to achieve stable performance, free from "birdies."

Whether or not an amplifier is used, the output tap on the tuned circuit inductor should be just far enough from the ground end for a stable signal, free from multi-vibrator type birdies when the key is first closed. In the unit described, the tap is almost at midpoint, 10½ turns from the ground end with a total of 23 turns in the coil. The optimum location for this tap must be obtained by "cut and try" method, keeping the collector voltage low and backing down on the emitter potentiometer to avoid exceeding 15 ma collector current.

The transistorized crystal oscillator is shown with the transistor just above the crystal unit. The lower central knob adjusts the variable tuning capacitor, while the left-hand knob is used to set the "bias" potentiometer at optimum for clean keying at full output. The switch at the lower right is the main battery on-off switch, while the jack at the lower left is for the key. On top of the unit, the coaxial connector is for the rf output and the four-prong male connector is for the 12-volt and 1.5-volt supplies from the battery box. The inside components of the oscillator are shown in the photograph.

In the amplifier circuit, the only critical adjustment is the location of the tap on the collector tank coil. The optimum position must be found by trial, but should be near the midpoint or slightly towards the ground end. Because one set of batteries is used as the 12-volt supply for both the oscillator and amplifier, the on-off switch may be omitted since the common ground is made through the coaxial cable.

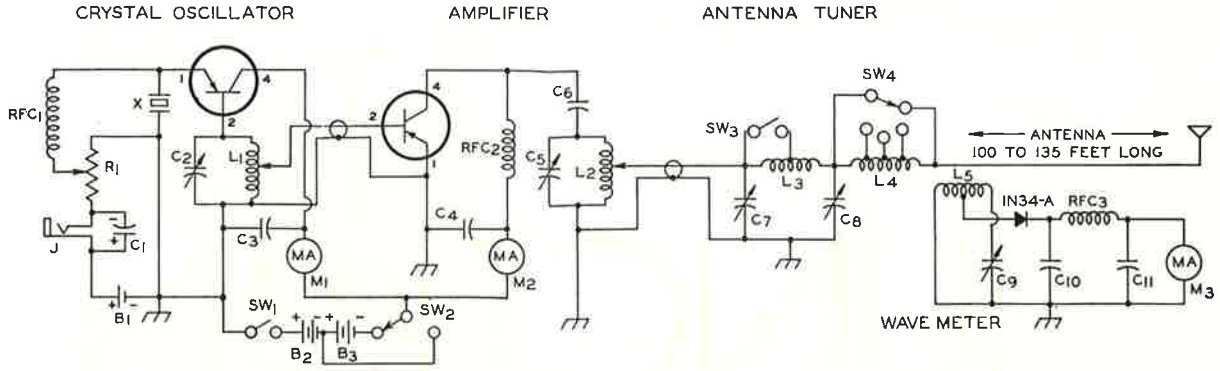
The use of an antenna tuner was found helpful, although not essential. The absorption wave meter, however, is considered an absolute necessity, since it gives a sensitive indication of the radiated energy. During tuning operations, this meter pickup coil may be located close enough to the antenna wire (about 6 feet from the tuner or transmitter) to give a meter reading at about half-scale, assuming that full scale is about 200 μa. Then it should be removed completely or decoupled until the needle movement is just visible. Although the tuner circuit contains more components than absolutely required, it does permit precision tuning for optimum radiated power at minimum collector current, and in low power work of this particular type every individual milliwatt must be utilized to produce maximum power for maximum contacts.



Rear View of Crystal Oscillator. Best dx on 40 Metres was ZS6TR, 8,000 miles, Transvaal, South Africa, without prearranged contact.

As in conventional transmitter tuning, increasing the load will also increase the collector (plate) current, but instead of tuning the tank circuit for a dip in collector current, the more positive indication of proper loading is maximum wave meter current at minimum collector current. Maximum radiation normally will not be at maximum current in the collector circuit. Adjustment of the emitter potentiometer in the oscillator is quite critical for optimum setting.

In tuning operations it is advisable to listen to the signal in the station receiver. As the voltage on the oscillator emitter is gradually increased and oscillation starts, the signal will sound very strong, even before there is any indication of collector current in the amplifier. As the emitter potentiometer is advanced slowly, the oscillator collector current will increase and the keyed signal will become clean, with a slight ringing which is



Grounds shown are to individual metal cases; no earth ground is used. Maximum collector input is 216 milliwatts to oscillator and also to amplifier. Batteries used should be heavy-duty dry cells.

characteristic of crystal oscillator keying. Unfortunately, if there is any indication of radiated power under this setting of the potentiometer, it will be very small, and the emitter voltage should be further increased. At about 10 ma collector current, there should be a definite amplifier collector current and a wave meter indication of radiated power, and all tuning controls must be adjusted carefully until peak radiation is reached. During this final tuning, birdies are very liable to be heard in the receiver all over the dial, and tuning must be readjusted until the only signal heard is at the crystal frequency. If tuning alone is not effective in eliminating these spurious oscillations with a "cold" transistor, the emitter voltage in the oscillator must be reduced or the tap on its base coil moved further from the ground end.

When the keyed signal is clean and free from birdies, with collector current between 12 and 15 ma in the oscillator and 15 to 18 ma in the amplifier, and with a good indication of radiation in the absorption wave meter, that meter should be removed or coupled very loosely. The rig is then all set for normal use.

In at least one respect, however, operation will not be normal, and that is in establishing contacts. The only successful method experienced by the writer has been in answering general calls and rarely by calling CQ, CQ-TR, CQ-QRP, or any other form inviting a QSO. Experience teaches that it is well to listen for a few seconds before answering a CQ, to see if others are answering the same call. If so, it is almost a waste of time to answer, even assuming your crystal frequency is close enough to be hopeful of establishing contact. The writer has had best success by having a fair selection of crystals, choosing one which is in the least occupied portion of the band, tuning

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PARTS LIST

B1—Battery, 1.5-volt.
B2—Battery, 6-volt.
B3—Battery, 6-volt.

C1—2 μf, electrolytic.
C2—0-100 μμf, variable.
C3—0.01 μf.
C4—0.01 μf.
C5—0-100 μμf, variable.
C6—0.001 μf.
C7—0-100 μμf, variable.
C8—0-100 μμf, variable.
C9—0-50 μμf, variable.
C10—0.001 μf.
C11—0.001 μf.

L1—23 turns, 18 gauge, 1¼" diameter, 2" long, tapped near centre.
L2—As L1.
L3—As L1.
L4—90 turns, 18 gauge, 1¼" diameter, 2¾" long, with 3 equal taps.
L5—Any size 40-metre pickup coil centre-tapped, which tunes through band, with C9.

M1—0-20 dc milliameter.
M2—0-20 dc milliameter.
M3—0-100 microammeter.

R1—100,000 ohms.
RFC1—RF choke, 1 mh.
RFC2—RF choke, 1 mh.
RFC3—RF choke, 1 mh.

SW1—SPST switch.
SW2—SPDT switch.
SW3—SPST switch.
SW4—Switch, 4 position.

X—Crystal, 3.5 or 7.0 Mc.
Transistors, 2N219 or 2N140.
Diode, 1N34A.

THE JUNCTION TRANSISTOR

By R. W. Hurst

Part 2 — The Common-Emitter and Common-Collector Amplifiers

This article, which is being published in three parts, is abstracted from a group of transistor lectures given jointly by the author and Mr. A. C. Luther. The first part of the article was presented last month, and dealt with the transistor as an extension of a junction diode, and showed the characteristics and behaviour of a common-base amplifier, which can give large voltage gains, but always gives a current gain less than one. Operating it as a voltage amplifier is very difficult, since the driving source must have such a low impedance, and operating it as a current amplifier has no advantage, since it can offer no current gain. Although certain intermediate types of operation can make use of the common-base amplifier in spite of its restrictions, it is more common to see a transistor connected in the common-emitter configuration. Therefore, this second article will deal with the common-emitter configuration. The common-collector configuration will also be discussed.

The common-emitter configuration for a transistor corresponds roughly to the grounded-cathode configuration of a valve. It is one of the most frequently used transistor configurations, chiefly because it offers the greatest gain capabilities of any of the three possible configurations. It also provides a higher input impedance.

The ability of the common-emitter configuration to provide large current gains lies in the fact that the signal current is applied to the base, where it adds to or subtracts from the very tiny base current. Since the base current (bias current plus the signal current) is a fixed percentage of the collector current, variations in the base current will cause proportional variations in the much larger collector current, so resulting in a current gain.

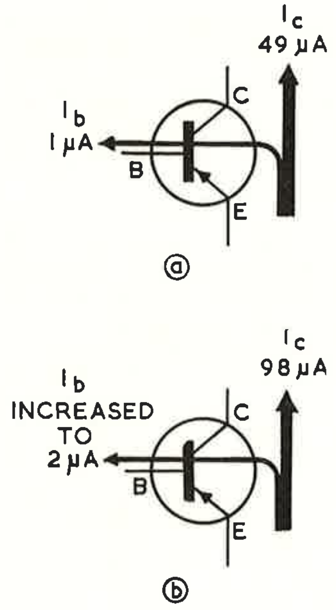

Fig. 17

This point is illustrated in Fig. 17. At Fig. 17a we have a common-emitter configuration in which, under no signal conditions, the base current is about 2% of the collector current. When a signal is applied, shown in Fig. 17b, a change of collector current of 49 μa takes place for a change in base current of only 1μa.

Common Emitter Characteristic Curves

In studying the common-base configuration in the first part of this article, a laboratory experiment was described (see "Characteristic Curves of Transistors") in which the transistor was connected to a tapped battery, and the emitter current/collector current curves were determined. They were determined firstly for zero emitter current, and then for several values of emitter current, thereby generating the emitter current versus collector current curves for the common-base configuration. With these curves the operation of the common-base amplifier was analysed.

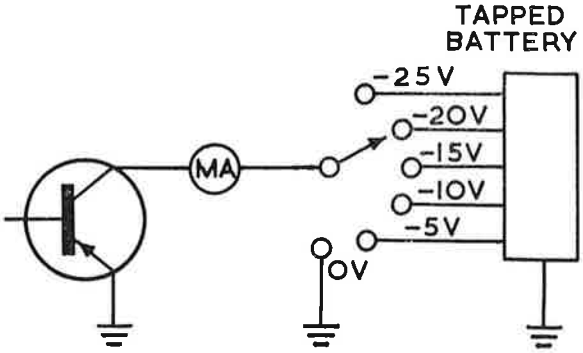

Fig. 18

In a similar manner a laboratory experiment can be set up with a transistor in common-emitter configuration, as shown in Fig. 18. The emitter current versus collector current curves can now be plotted, firstly for zero base current, and then for several values of base current. Finally sufficient data is available to construct a complete family of curves, as shown in Fig. 19.

It will be observed that these curves differ from the common-base curves in several respects. First, the curves have a steeper slope (1); second, the curves do not run all the way over to the left-hand axis (2); third, the running parameter, the base current, is a very small current, as compared to the collector current (3); and fourth, the collector current for zero input (base) current is much greater than the corresponding curve for the common-base configuration (zero emitter current) (4). These differences are indicated in Fig. 20, where the curves for the two configurations are compared. The significance of these differences will become apparent in the following discussion.

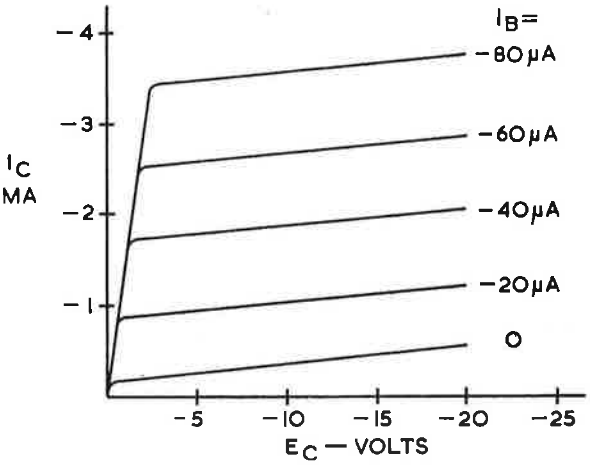

Fig. 19

Let it be supposed that a piece of equipment contains the amplifier shown in Fig. 21. It is desired to know how this amplifier is operating—what the bias is, how much collector current flows, how much power is dissipated in the collector, and how much gain it will offer. All these facts may be ascertained by a simple construction on the common-emitter characteristics.

First, a warning: the bias method shown in Fig. 21—a resistor R_b supplying bias current to the base from the collector power supply—is an extremely poor way to bias a transistor. It would cause the amplifier to be very sensitive to both transistor replacement and change in temperature. It is used in this example because of its simplicity, which gives it value as a means of explaining the common-emitter configuration. To make the following discussion valid, it must be assumed that the curves used are exact representations of the individual transistor in the circuit. Normally the published curves for a particular type of transistor are for a unit whose characteristics are in the centre of the allowable manufacturing tolerances. The fact that there are rather wide tolerances on transistors is one of the factors which makes necessary the use of more elaborate biasing techniques in practical circuits.

Analysis of this common-emitter amplifier follows basically the same pattern as the analysis of the common-base amplifier carried out in Part I. The analysis is begun by drawing the load line on the common-emitter characteristics; this has been done in Fig. 22.

Just as for the common-base circuit, this amplifier's operating point must lie somewhere on its load line. For the common-base amplifier, it was shown that the operating point lies at the intersection of the load line and the particular bias-current chosen. The operating point of this common-emitter amplifier may be found in a

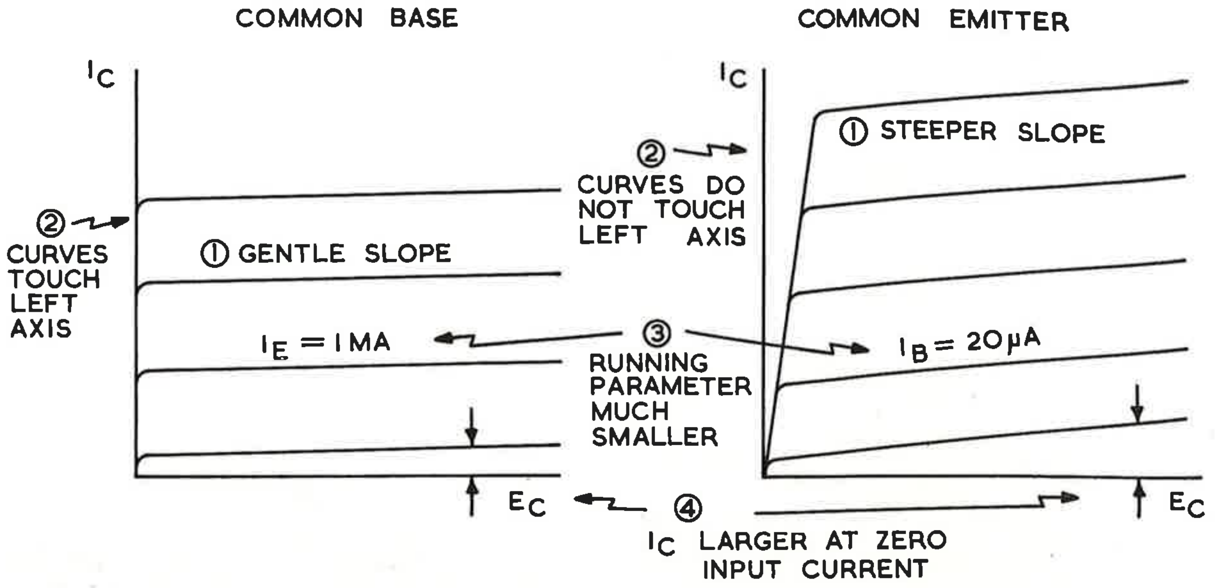

Fig. 20

similar manner, where

$$I_b = \frac{-20 \text{ volts}}{500{,}000 \text{ ohms}} = 40 \ \mu\text{a}.$$

This common-emitter amplifier therefore operates with a bias current of —40 μa, a collector current of —2 ma, and a collector voltage of —10 volts. The operating point has been included in Fig. 22.

The power dissipated at the collector is:

$$\begin{aligned} P = IE &= (-2 \text{ ma})\ (-10 \text{ volts}) \\ &= 20 \text{ milliwatts} \end{aligned}$$

The gain of this amplifier depends upon whether it is being used to produce voltage gain or current gain. In either case, it can produce useful gain. A simple amplifier such as this one has a current gain equal to beta — about 50 for a typical transistor — if the load is a short circuit or very low impedance.

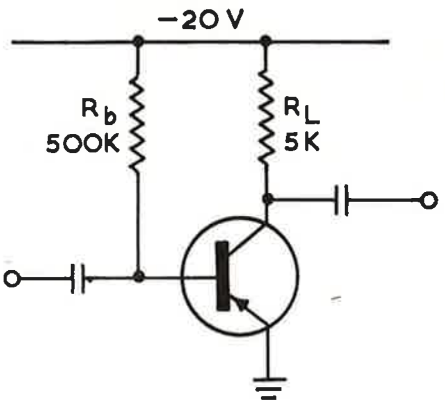

Fig. 21

However, this particular amplifier has a load of 5000 ohms, so its current gain will be slightly less than beta. The actual value can be determined from the characteristics in the manner shown in Fig. 23.

Since the output is only 1.6 ma (instead of 2 ma, as it would be for a short-circuit load), the current gain of this amplifier is:

$$G_e = \frac{1.6 \text{ ma}}{40 \ \mu\text{a}} = 40$$

The voltage gain of this particular common-emitter amplifier is somewhat less than the voltage gain of the common-base amplifier already analyzed. It was shown, in that analysis, that the voltage gain is given approximately by the expression:

$$Gv_{CB} = \frac{R_L}{R_e} = \frac{5{,}000}{12.8} = 391$$

A more nearly exact value for voltage gain is:

$$\begin{aligned} Gv_{CB} &= (\text{Current Gain}) \times \frac{R_L}{R_e} = \alpha \frac{R_L}{R_e} \\ &= \frac{(0.98)\ (5{,}000)}{12.8} = 382 \end{aligned}$$

but the current gain for the common-base case is so near unity that it can be omitted from the expression, without introducing appreciable error. In the common-emitter case, however, the current gain is so large that it must be included in the expression:

$$Gv_{CE} = G_e \frac{R_L}{R_b}$$

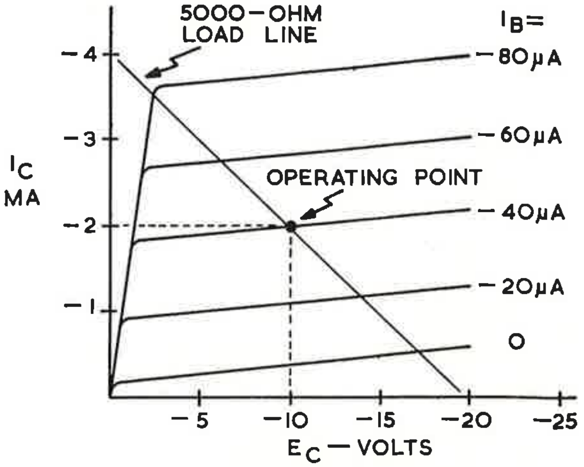

Fig. 22

Note also that in the expression for common-emitter voltage gain the input impedance is given as R_b, the base resistance, in place of R_e, the emitter resistance. The base resistance is larger than the emitter resistance by a factor of $\beta + 1$. Therefore, the voltage gain is given by:

$$Gv_{CE} = \frac{G_c R_L}{(\beta + 1)\ R_e} = \frac{(40)\ (5{,}000)}{(51)\ (12.8)} = 306$$

An important point contained in the foregoing paragraph is the fact that the impedance seen looking into the base is $\beta + 1$ times larger than that impedance seen looking into the emitter.

This base impedance can become very large—approaching that of a valve grid — if an external resistor is included in the emitter lead. Fig. 24 illustrates this point, where the use of a 3,000 ohm emitter lead resistor raises the base impedance from a little over 3,000 ohms to 150,000 ohms.

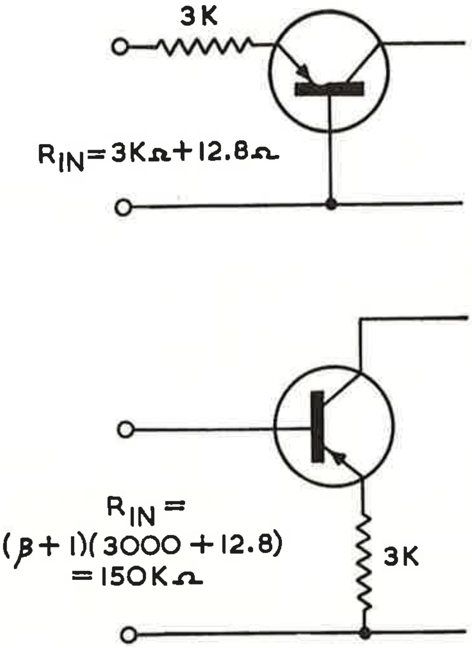

Fig. 24

This external impedance, however, behaves like an unbypassed cathode resistor in a valve circuit — it gives higher input impedance (and greater stability as well) at the expense of gain.

The common-emitter circuit analyzed thus far is admittedly an impractical circuit. Biasing a transistor through a large base resistor (in this case, 500,000 ohms) results in a circuit which may work well at room temperatures but becomes completely inoperative at higher temperatures. Or, it may work well with one transistor but work poorly or not at all with another transistor of the same type. Therefore, it is also necessary to analyze a common-emitter amplifier using more practical biasing techniques.

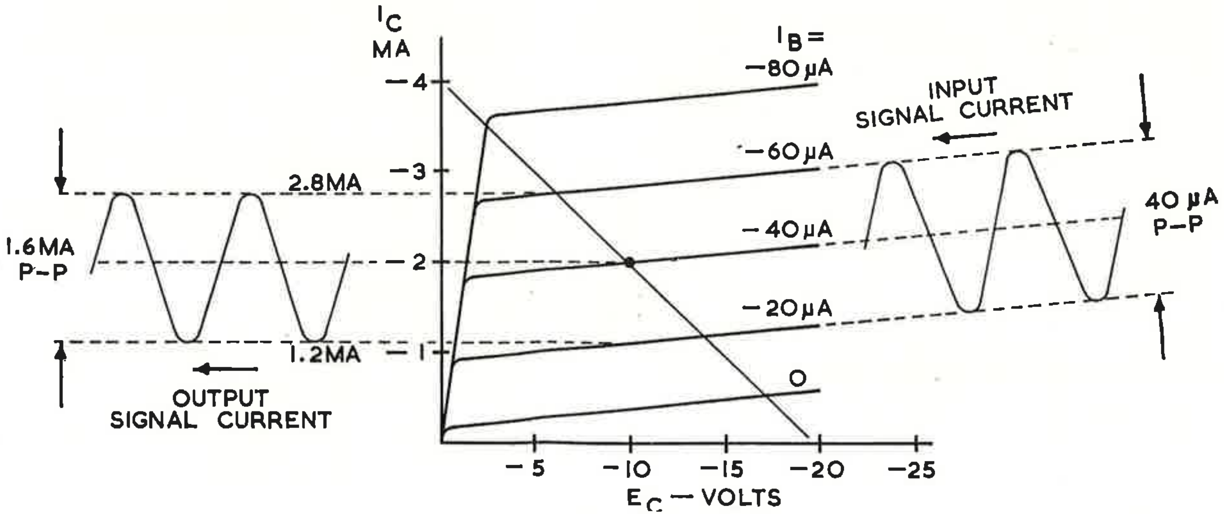

Fig. 23

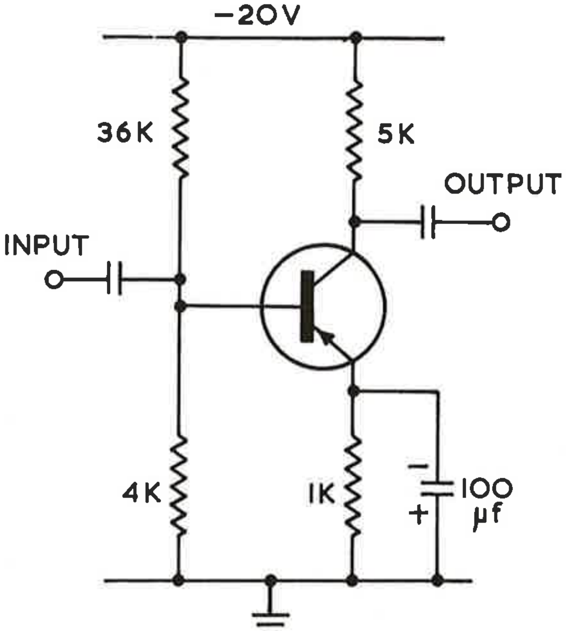

Fig. 25

Practical Common-Emitter Amplifier

Let us suppose that a piece of equipment contains the circuit shown in Fig. 25. It is desired to know how this amplifier is operating; that is, what its bias is, what its dissipation is, how much collector current flows, and how much gain it provides. Although a construction of the common-emitter characteristics could be made to yield this information, it is possible to make a fairly accurate analysis without the characteristics, since the 1000 ohm emitter resistor stabilizes the circuit and thereby makes its dc behaviour fairly independent of the transistor's characteristics.

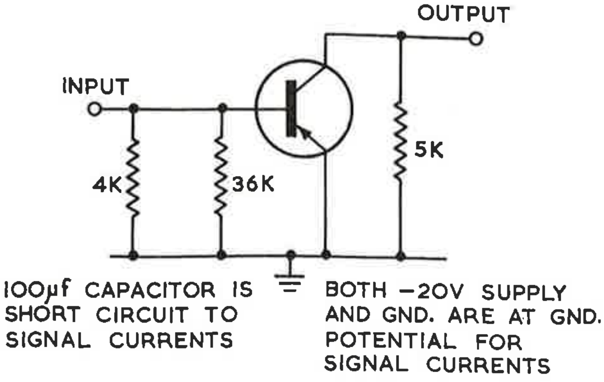

Fig. 27

The analysis is begun by observing that, with the transistor out of the circuit, a potential of +2 volts appears at the junction of the 36,000 ohm resistor and the 4,000 ohm resistor which form the bias network. The calculations are shown in Fig. 26.

When the transistor is connected, the base current, which flows out of the (p-n-p) transistor, joins the 0.5 ma current in the bias network and slightly alters this —2 volt potential. However, the base current is assumed to be much smaller than the bias-network current and may therefore be safely ignored in an approximate analysis such as this one. (This assumption is usually correct in the analysis of well-designed circuits. If it should happen to be incorrect, one of the succeeding steps will reveal the error.)

Therefore, even with the transistor in the circuit, the potential at the base is —2 volts. Since the emitter and base taken together form a forward-biased diode, the voltage drop from emitter to base is very small — about 0.2 volt; a neg-

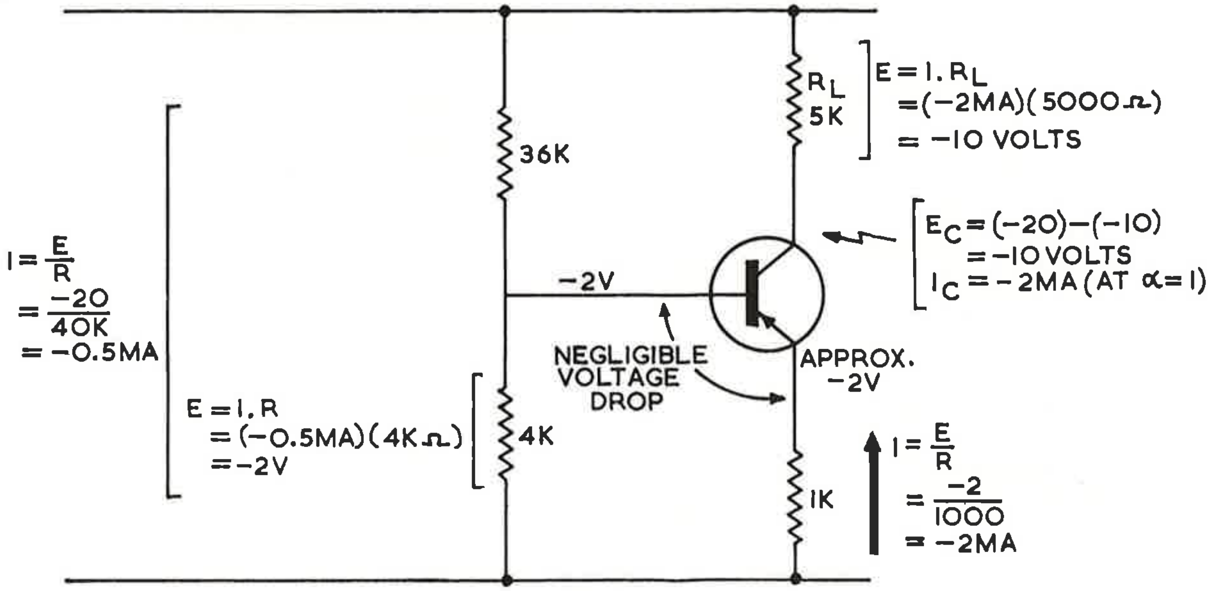

Fig. 26

ligible amount in this analysis. Therefore the emitter is also at approximately —2 volts. Since the emitter resistor is 1,000 ohms, the current causing this 2 volt drop must be —2 ma.

Assuming that $\alpha = 1$ (instead of 0.98) for this approximation, the current flowing from the collector is also —2 ma, and the potential at the collector is therefore —10 volts.

Since the emitter is at —2 volts, the voltage across the transistor is only —8 volts; therefore the power dissipation is:

$$P_c = IE = (-2 \text{ ma})(-8 \text{ v}) = 16 \text{ milliwatts}$$

Also, with the knowledge that $I_c = -2$ ma, (and assuming that $\beta_{DC} = 50$), then the base current can be calculated:

$$I_b = \frac{I_c}{\beta_{DC}} = \frac{-2 \text{ ma}}{50} = -40\ \mu a$$

The earlier assumption that the base current was much smaller than the 0.5 ma bias-network current is thereby substantiated.

In computing the gain of this amplifier, the circuit may be redrawn as shown in Fig. 27, where the circuit is shown as it appears to the signal current. If a signal current of 10 μa is supplied to this amplifier, part of it is lost in the 4,000 ohm and 36,000 ohm resistors. Only a portion of the signal current goes into the base, represented by a resistor of 650 ohms, the theoretical input impedance of the transistor. This is illustrated in Fig. 28.

Since 15 per cent of the input signal is lost before it ever gets into the base the overall current gain of this amplifier is 15 per cent lower than the gain of the simple amplifier already analyzed. (That amplifier lost a negligible amount of signal current in its 500,000 ohm biasing resistor.) Since that amplifier was shown to have a current gain of 40, this amplifier will have an overall current gain of:

$$G_c = 40 - (0.15)(40) = 34$$

The loss of gain is the price paid for the increased stability of this circuit.

The voltage gain of this amplifier may be approximated by observing that the 1.5 μa lost in the bias network causes a voltage swing of 5.4 millivolts to appear across the two resistors which are effectively in parallel across the input signal.

At the same time, 8.5 μa is being amplified by the transistor so that $(8.5) \times (40) = 340\ \mu a$ appears as a current swing in the load resistor, and an output of 1.7 volts appears across the load resistor.

Therefore, the voltage gain of the amplifier is:

$$G_v = \frac{1.7 \text{ v}}{5.4 \text{ mv}} = 315$$

The Common-Collector Configuration

So far, the common-base and common-emitter amplifiers have been considered and their curves and characteristics briefly indicated. The common-base amplifier is capable of large voltage gain but less-than-unity current gain. The common-emitter amplifier can provide both voltage and current gain. The final configuration, the common-collector, is the opposite of the

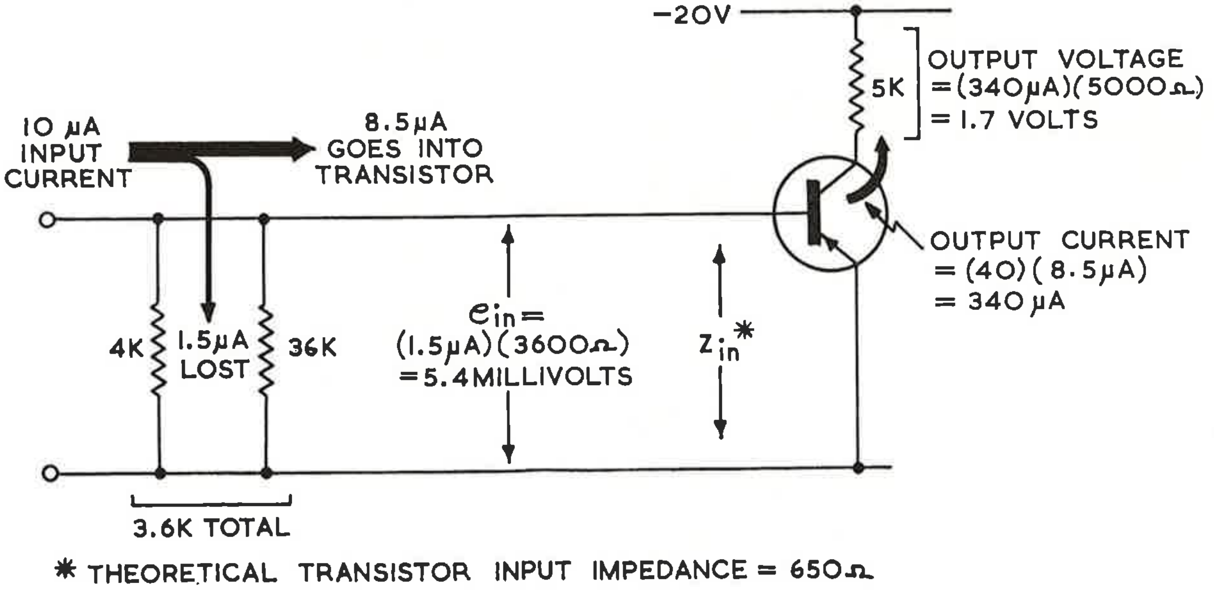

Fig. 28

common-base configuration in that it can produce a large current gain but less-than-unity voltage gain. In that respect, it resembles its valve counterpart, the cathode follower. Indeed, it is often called the emitter-follower configuration. In this article, however, the name common-collector will be employed, except where clarity may be served by the other name.

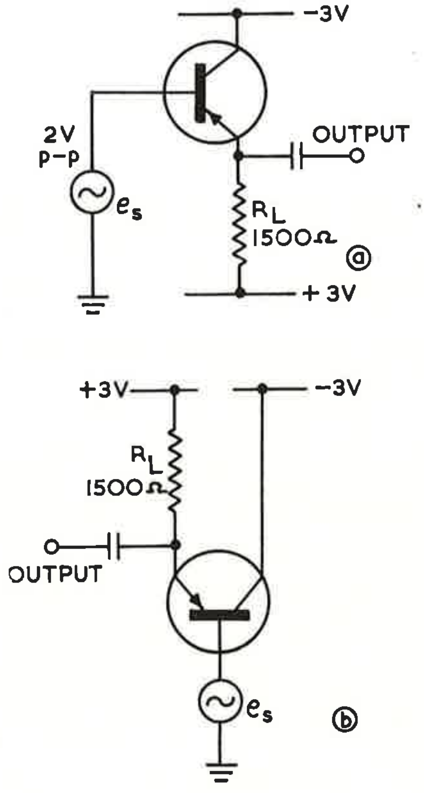

Fig. 29

The discussion of common-collector amplifiers could run parallel to the discussions of the other configurations, starting with a derivation of the common-collector characteristic curves, and using these curves to analyze a typical amplifier. In this case, however, such an approach is not practical, because common-collector curves are rarely given in data sheets. Any analysis based on common-collector curves could not be duplicated in a practical situation, without first deriving a set of common-collector curves from the data sheet's common-emitter curves. Fortunately, an approximate analysis can be performed using common-emitter curves directly. This is demonstrated in the following analysis of the common-collector amplifier shown in Fig. 29a.

This amplifier is driven from a very-low-impedance source which approximates a true voltage source. This amplifier's voltage gain and its dc operating conditions will be determined by a construction on the common-emitter curves.

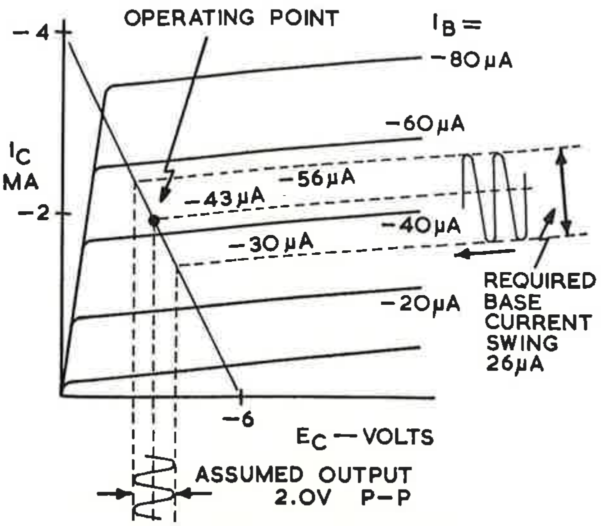

Fig. 30

The biasing arrangement is practically identical to the common-base biasing-method already discussed. To show this similarity, the circuit can be redrawn as shown in Fig. 29b.

Remembering that e_s, the signal source, is practically a short circuit to the bias currents, it can be seen that the biasing arrangement bears a strong resemblance to common-base biasing.

It has been shown that the input (emitter) resistance could usually be ignored in determining bias current for a common-base stage. Therefore, it can be assumed that the full + 3 volt supply appears across R_L, giving an approximate bias current of 2 ma flowing in the emitter. With this approximate bias, we may find the operating point by drawing a load line on the common-emitter curves. The terminal points of this load

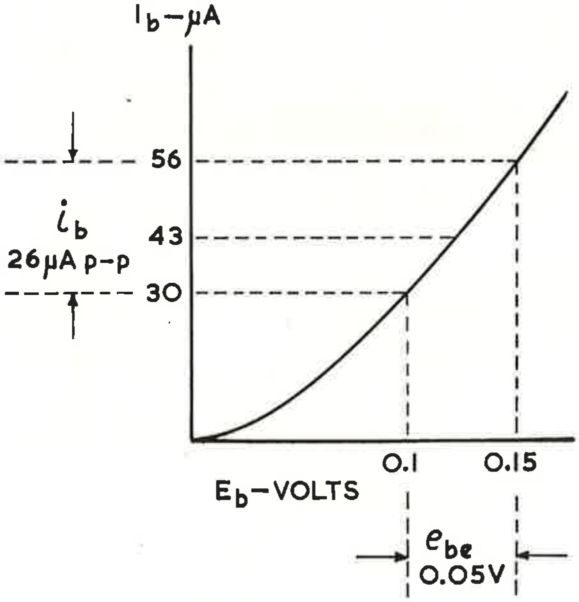

Fig. 31

line (or any load line) are the open-circuit voltage and the short-circuit current available from the external circuit at the E and C terminals of the transistor; in this case the terminal points will be —4 ma and —6 volts, and Fig. 30 shows the load line drawn in on a set of curves.

Actually, drawing a straight line between the two end points determined in the above manner is not entirely accurate because the current scale of the graph is I_c, whereas it is really I_e which flows through R_L. However, the error involved is small since I_c and I_e are nearly equal.

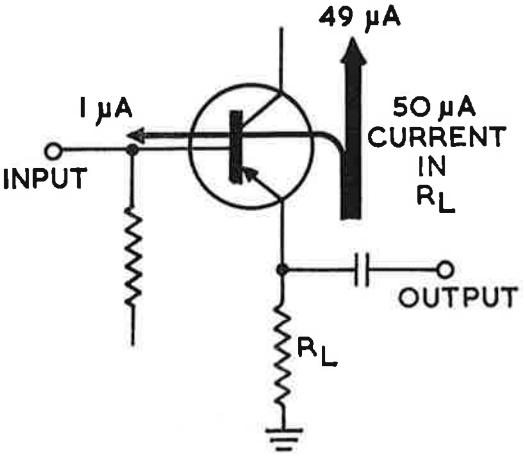

Fig. 32

An approximate operating point may then be found by entering the approximate bias current on the load line of Fig. 30. The operating point is the intersection of the —2 ma and —6 volts co-ordinates on the load line, shown by P on the diagram.

It can be seen in Fig. 30 that the operating point lies between the curves for $I_b = -40\ \mu a$ and $I_b = -60\ \mu a$. Judging the relative position of the point by simple visual inspection, one can approximate the no-signal base-current to be about $-43\ \mu a$.

To simplify the analysis at this point, a rather surprising assumption is made. It is now assumed that the voltage gain is unity; that is, it is assumed that the full 2 volt input signal appears at the output. Working backwards from this assumed output voltage, one can then proceed to show that more than 2 volts of signal is required at the input to produce a 2 volt output. The relationship between this new input voltage and the assumed 2 volt output voltage will give the actual gain, which is slightly less than unity.

Starting with the assumed 2 volt output signal, the common-emitter characteristics can be used to determine the base-current swing required to deliver this output. See Fig. 30.

The data sheet for this transistor is then consulted to find the curve giving the relationship between base current and base-to-emitter voltage. This curve is usually given for transistors designed to handle fairly large signal swings. Using this curve, a section of which is shown in Fig. 31, it can be seen that 0.05 volts of the input signal is lost in the base-to-emitter voltage drop. Therefore, in order to produce the assumed 2 volt output, the input voltage must swing 2.05 volts. The voltage gain is therefore:

$$G_v = \frac{e_{out}}{e_{in}} = \frac{2.0}{2.05} = 0.976$$

The common-collector configuration, driven from a current source, can be described in terms of current gain. In a simple configuration, in which negligible signal current is lost in the biasing arrangement, a transistor with a beta of 49 will give a current gain of 50, which, in general terms, is a curent gain of $\beta + 1$. See Fig. 32.

The input impedance of a common-collector amplifier is very large. This may be seen from the fact that, in the example, a 2-volt signal causes only 26 μa to flow into the amplifier. This gives an input impedance of:

$$R_{in} = \frac{e_{in}}{i_{in}} = \frac{2.0}{26 \times 10^{-6}} = 77{,}000 \text{ ohms}$$

which agrees closely with the value of input impedance calculated by multiplying R_L, (1500 ohms), by $\beta + 1$.

TABLE I

	COMMON BASE	COMMON EMITTER	COMMON COLLECTOR
CURRENT GAIN	0.98	49	50
VOLTAGE GAIN	382	306	0.976
INPUT R	12.8Ω	650Ω	77KΩ
FREQ. RESPONSE	I MC	20 KC	20KC – IMC*

*DEPENDING ON SOURCE AND LOAD

Thus far descriptions have been given for the characteristics of the three important transistor-amplifier configurations: the common-base, the common-emitter, and the common-collector. Their important characteristics may be summarized in Table 1 which gives the typical values calculated for the amplifiers used as examples.

The last row of information, frequency response, indicates the price paid to obtain the improved gain capabilities of the common-emitter configuration. Although the current gain of the

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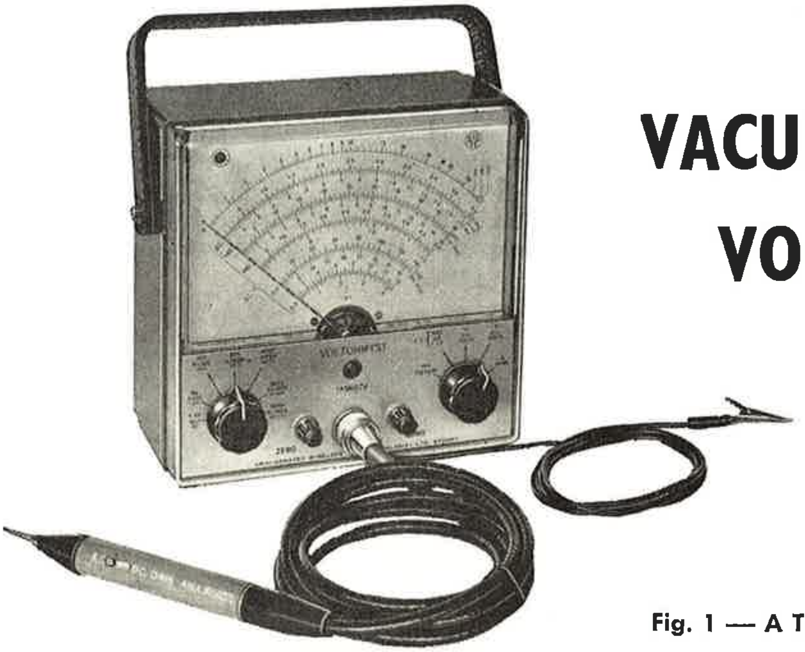


THE VACUUM – TUBE VOLTMETER

Fig. 1 — A Typical Vacuum-Tube Voltmeter.

A vacuum tube voltmeter is one of the most widely used pieces of test equipment in the television servicing industry. It is invaluable to a technician in the process of analyzing circuitry within a television receiver, since it provides for accurate measurement of voltages and resistance in circuitry wherein ordinary multimeters would register inaccurate information. The vacuum-tube voltmeter has a much higher input impedance than a multimeter; it has the ability to measure ac voltages at very high frequencies, and it has a greater sensitivity than ordinary multimeters. However, to attain maximum benefit from the use of a vacuum-tube voltmeter, it is essential that a technician understand something about its basic principle of operation, know just what it can measure, and understand how to use it properly.

In a multimeter the input resistance is limited by the sensitivity of the meter movement itself. In a vacuum-tube voltmeter the input resistance is not limited by the meter movement; the input resistance of the instrument is increased by the use of vacuum-tube circuitry.

DC Voltage Readings

Typical circuitry used in a vacuum-tube voltmeter consists of a balanced bridge type network, such as illustrated in Fig. 2. When the circuitry of the two tube sections are balanced, the electron flow through each of the sections is the same and the voltages at point A and B (plates of the tube sections) are at the same potential. A variable potentiometer (R3) is provided in the plate supply of the bridge circuitry so that points A and B can be set to the same potential each time measurements are to be made. This potentiometer is controlled by the knob on the front of the instrument labelled ZERO ADJ.

In operation, to measure voltages, the voltage to be measured is applied between the grid of a tube in one section of the bridge (V1), see Fig. 2, and ground. If the voltage is positive, the electron flow through the tube, V1, will increase; point A (plate of V1) will decrease in potential while point B (plate of V2) will remain unchanged since its grid is grounded. A meter connected from point A to point B will indicate the potential difference that exists between the two points. The meter dial is calibrated in volts. A negative voltage applied to the grid of V1 would normally make the pointer on the meter move in the opposite direction; so instruments that have the zero volts position at the extreme left of the scale employ a switch to reverse the meter connections. This switch is normally labelled + DC and —DC.

In order for the vacuum-tube voltmeter to measure an adequate range of voltages a voltage divider network is incorporated in the input circuit between the meter input terminals and ground. A typical voltage divider network is illustrated in Fig. 3. This network provides for an input resistance of 11 megohms and provides taps for several voltage ranges.

Whenever unknown voltages are measured, it is advisable to use the highest scale available on the instrument and work down to the scale that permits a reading without the meter going off scale. To insure accuracy, the meter zeroing adjustment should be checked each time the voltage range is changed. When making this adjustment the meter leads should be shorted to prevent any possibility of stray voltage pickup that may cause a misleading indication. This is particularly important when the lowest voltage scale is being used. Most vacuum-tube voltmeters provide enough range on the ZERO ADJ to allow for the pointer to be positioned at the centre of the scale. Zero centre-reading vacuum-tube voltmeters often have a separate small scale on the dial face of the meter marked with zero in the centre. The ZERO CENTRE position is not employed to indicate specific voltages but merely to show whether the circuit under test is balanced, such as is done during a discriminator adjustment.

To extend the range of dc voltage measurement up to 25 or 50 kilovolts, a high-voltage probe must be employed. A typical probe extends the range of the meter to 50,000 volts (a multiplying factor of 100); it increases the total input impedance to 1100 megohms, a valuable feature for making accurate high-voltage measurements in high-impedance circuits.

AC Voltage Readings

To provide for ac voltage measurement in vacuum-tube voltmeters, ac voltages that are to be measured are rectified to form a dc voltage, of approximately peak value, applied to a voltage divider network, and then measured in the same fashion as dc voltages. The meter scale for ac measurement is normally calibrated to indicate the rms value of sine wave voltages. Vacuum-tube voltmeters, such as that shown in Fig. 1, that are capable of measuring peak-to-peak values of complex waves as well as sine waves, use a voltage doubler in the ac rectifying circuit. Some meters having peak-to-peak markings on the meter are designed to read peak-to-peak values of sine waves only. A technician should always be certain of the capabilities of the meter he's employing before he uses it.

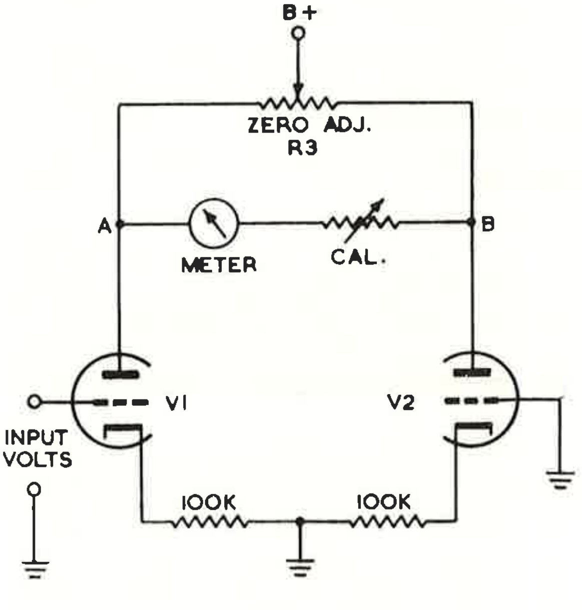


Fig. 2 — Balanced Bridge — Type Network used in a VTVM.

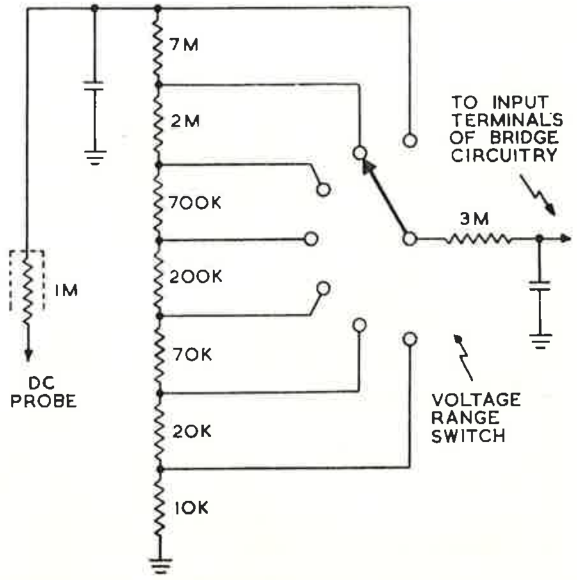


Fig. 3 — Typical Voltage Divider Network used in a VTVM.

Alternating voltages at frequencies up to approximately 50 kilocycles (in circuits of 1000 ohms impedance or less, depending on the particular meter) can usually be measured accurately through a direct probe, but for frequencies above this range an RF Probe (Crystal-Diode Probe) must be employed to provide for voltage measurements without disturbing the operating characteristics of the circuits under test. This is so because as frequencies increase, circuits become more sensitive to the capacitive and inductance effects of cable connections. The rf probe rectifies the ac voltages close to the source and thereby minimizes the effects of the capacitive and inductive characteristics of the cables. With the use of an RF Probe (Crystal-Diode Probe) a typical meter can measure rf voltages within 10% in frequency ranges of 50 Kc to 250 Mc.

Resistance Measurements

Resistance measurements of as high as 1000 megohms can be measured with a vacuum-tube voltmeter. This is far in excess of measurements that can be made with an ordinary multimeter.

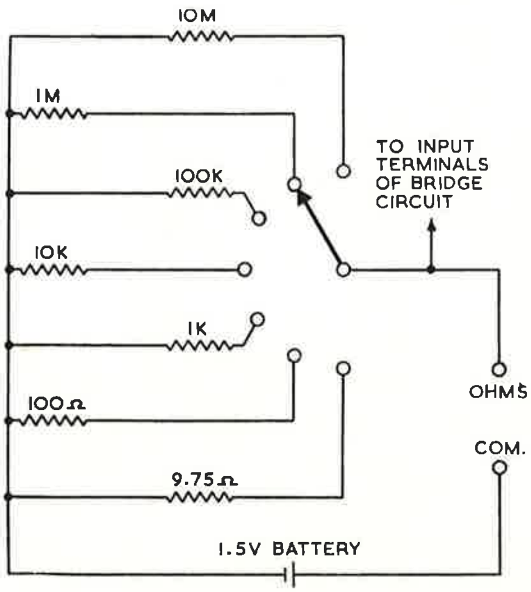


Fig. 4 — Divider Network Employed in a VTVM for Making Resistance Measurements.

The ohmmeter circuitry of a vacuum-tube voltmeter usually employs a 1.5 volt battery as a voltage source that is applied to a voltage divider network that can be switched into the grid circuitry of the bridge system when resistances are to be measured. See Fig. 4. The potential developed by the battery (assuming the battery is good) is sufficient to effect a full scale deflection of the pointer accurately at the upper end of the scale OHMS ADJ located on the panel of the meter allows for the meter pointer to be accurately positioned on the maximum ohms mark on the meter scale (final righthand marking on the meter dial). When the resistance under test is connected between the common and ohms test leads, a voltage divider is formed consisting of the battery in series with one of the meter circuit resistors and the resistor under test. The voltage across the unknown resistor is proportional to its resistance. This voltage, when applied to the bridge circuit network will cause the meter to deflect proportional to the resistance.

When measuring resistance it is important to remember to remove all voltage from the circuit under test. Always zero the meter at the low end (using the ZERO ADJ. control), and position the pointer accurately at the upper end of the scale (using the OHMS ADJ. control). At the low end the leads must be shorted together — at the high end the leads are kept apart. Always use the scale that places the pointer somewhere in the centre or left hand section of the scale whenever possible.

(With acknowledgements to RCA)

A TRANSISTORIZED QSO-GETTER

(Continued from page 25)

for optimum radiation at minimum power input, and waiting for someone to call CQ on that frequency. On 40 metres you don't have to wait long under normal conditions.

On 80 metres, a 2N218 or 2N139 may be used in place of the 2N140 or 2N219. Cutoff frequency of the 2N218 is approximately 5 Mc. The 2N219 should be used for 7 Mc operation, with its higher cutoff frequency at about 8 Mc. Future QRP rigs hold many possibilities of higher-frequency operation, voice modulation, an increased efficiency.

Your author wishes to emphasize the importance of selecting the proper location for the tap on the oscillator coil and also on the amplifier coil. Both are extremely critical for optimum performance. The antenna tuner described will load almost any kind of wire, but obviously the better antenna system employed, the better the results will be.

Your author has never used any form of beam and all contacts, nearly 200 at this writing, have been without previous arrangements and without previous contact with higher power equipment. In the author's opinion, such "piggy-back" contacts void the attraction of the adventure in transistorized QRP amateur communications.

(With acknowledgements to RCA)

Off The Beaten Track

A SERIES DESCRIBING SOME OF THE MORE UNCOMMON VALVES AND VALVE DESIGNS

No. 7--VERY-HIGH-POWER VALVES

Whilst the very-high-powered valves used in large transmitters and similar equipment are fundamentally only enlarged versions of the valves in your radio receiver, in many other respects they show little similarity. It is the purpose of this seventh article in the series to describe and explain some of the differences, not only in the construction of the valves, but in the problems associated with their operation.

One of the differences which first comes to notice is that high-powered valves are force-cooled, either by air or water. In the case of forced-air-cooled valves, the valve is generally provided with a finned outer jacket, and is mounted in a duct so that cooling air can be blown over the fins by means of a fan. The fins are provided of course so that a greater cooling area is available, as in the case of air-cooled petrol engines.

Where the valve is intended for water-cooling, it is either supplied with an integral water-jacket, or provision is made for the valve to be inserted into a separate water jacket. The integral cooling jacket in the case of air-cooled valves, and the water jacket in the case of water-cooled valves, are frequently made part of the anode assembly of the valve. This makes for efficient plate cooling but means that the cooling area of the valve is frequently at a high potential above ground.

This problem is not a large one when the valve is air-cooled, but when water is the cooling medium certain precautions have to be taken. The piping adjacent to the valve can be made of insulating material, whilst the use of deionized or distilled water means that conduction throughout the water is small, i.e., it has good insulating properties. The insulated section is made long

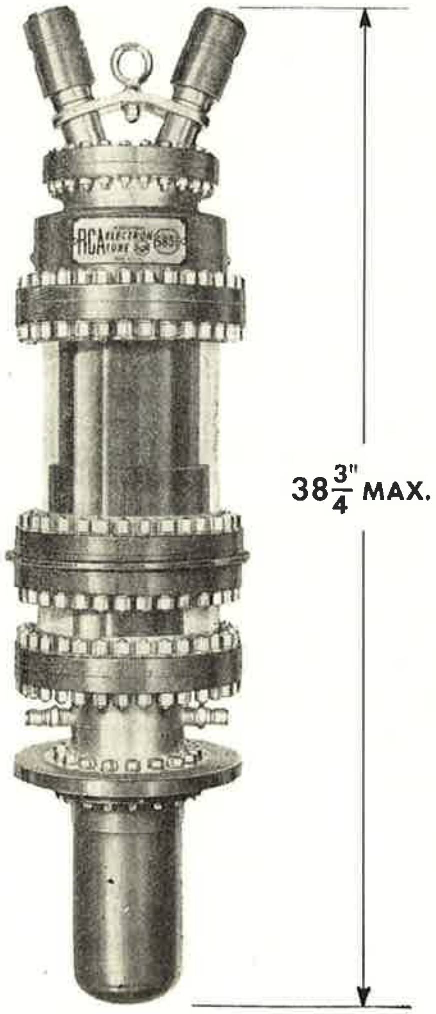


Fig. 1 — The 5831 Super-Power Beam Triode, Capable of 500 Kilowatts Output in Class C Telegraphy.

enough to reduce dc leakage, and is sometimes made into a helix to reduce rf leakage.

A variation of the water-cooling principle is vapour-cooling, in general applicable only to valves designed with this cooling method in mind. This system is particularly attractive where neither forced-air cooling nor water cooling of high-power valves is entirely satisfactory. Air may be so dirt-laden that the cleaning of filters and radiators calls for undue effort and cost. At the same time water may be expensive, scarce, or unsuitable for other reasons. In such locations vapour-cooled valves may produce the advantages of low water costs, built-in protection against water supply failure, quiet operation, and greater over-load capacity.

In the vapour-cooled valve a closed system, called the boiler, contains distilled or otherwise purified water in contact with the specially designed valve anode. When the valve is operating, heat from the anode rapidly converts some of the water into steam which passes via a condenser system back to the reservoir of water at the bottom of the boiler. Efficient circulation is brought about by the provision of specially designed circulation holes in the walls of the anode.

Two different condenser systems are employed and the choice between them is largely dependent upon local conditions. In the External Condenser system the steam is led outside the boiler to the condenser, where cooling may take place by natural convection or by any other convenient means. In the Boiler-Condenser system, the steam is condensed by being directed over the top turns of a cooling coil inside the boiler. Admittedly, a supply of cooling water is necessary for the coil, but the consumption is only a fraction of that for water-cooled valves of similar power capabilities. The Boiler-Condenser system makes a strong appeal by reason of its compactness.

The Boiler-Condenser may be fitted with a protective device for shutting off the power supply in the event of water failure or of serious overloading of the anode. Visual indication of water level is also generally provided.

Typical of some of the high-powered valves used today, and excluding highly specialized types like magnetrons, is the 5831, shown in Fig. 1. The 5831 is a water-cooled beam triode of unique design capable of generating several hundred kilowatts of power at high efficiency and with exceptionally low driving power. It is intended primarily for use as a class C rf power amplifier, either modulated or unmodulated, but is also useful as a class B af power amplifier and modulator. In unmodulated class C service, the 5831 has a maximum plate voltage rating of 16000 volts, a maximum plate input of 650 kilowatts, and a maximum plate dissipation of 150 kilowatts.

Fig. 2 — The 6166 VHF Tetrode, Capable of 12 Kilowatts Synchronizing Level Output in Television Service, with a Bandwidth of 8.5 Megacycles.

The 5831 has a multi-strand thoriated-tungsten filament. Each individual strand is recessed in a slot in a beam-forming cylinder through which cooling water is circulated.

The 5831 is unique in that it features a symmetrical array of unit electron-optical systems embodying a mechanical structure which permits close spacing and accurate alignment of the electrodes to a degree unusual in high-power valves. Ducts for water cooling the plate and the beam-forming cylinder are built in and have simplified hose connections. The grid-terminal flange requires a water-cooled connector. Because of the electron-optical principles incorporated in its design, the 5831 has low grid current and hence requires less than 2 kilowatts of driving power.

In order to gain a better appreciation of the construction techniques used in this valve, it may be interesting to trace the various parts of the valve as shown in Fig. 1. At the top of the photograph are shown the lifting eye-bolt and two connections for the plate cooling water supply. The plate connection is the second bolted flange just below the label, so that the whole of the top portion of the valve is at plate potential.

The water jacket is then continued through a glass tube to a further flange at about the mid-way point. This is the grid terminal. A second and shorter section of glass water jacket then leads to the cathode terminal flange, beneath which are two connections for the water supply to the beam-forming cylinder. Finally the lowest flange and the straight section of the body seen below it are the filament terminals.

Some of the problems in installing and operating this type of valve have now appeared from what has already been said. Two cooling water supplies (for plate and beam assembly) are required, and in addition the ring clamp carrying the grid connection must be water-cooled. But these are not the only problems the user must face.

First of all, the water supply used for cooling should be distilled water or deionized water to prevent the possibility of scale formation and corrosion, both of which can be expected with tap water. This calls for a closed circuit water system, including a sufficient supply of water, a heat exchanger, a pump and the necessary interconnecting pipes. Furthermore, interlocks must be used between the cooling system and the electrical supplies to the valve so that failure of the flow of cooling water immediately removes all applied voltages.

In this valve, as with many high-power valves, momentary failure of the water supply is sufficient to damage the valve. In fact, without cooling water, the heat of the filament alone is sufficient to cause damage. The filament draws 2,100 amperes at 6 volts, a power of a little over 12 kilowatts for the filament alone.

There are yet more protective devices essential to the safe operation of large valves of this type. A filament starter must be used to raise the applied voltage gradually to the normal value. This is to avoid a high initial surge through the cold filament. Then the plate voltage supply must incorporate a time delay so that plate voltage cannot be applied until the filament has reached normal operating temperature. Finally a high-speed electronic protective device must be used to remove the plate voltage in case of abnormal operation such as internal arcing, and the grid circuit must incorporate overload relays to remove all grid power in the event of excessive grid current flow.

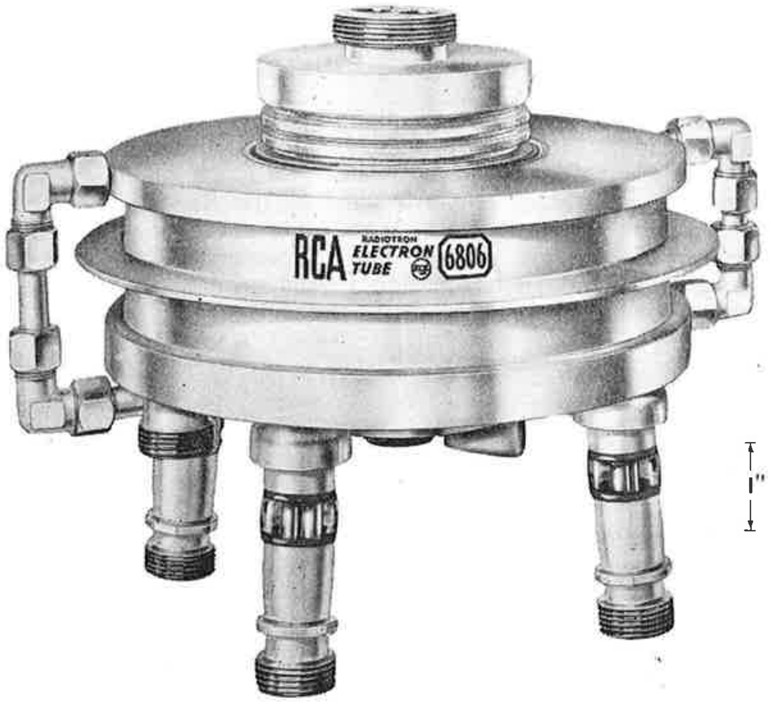


Fig. 3 — The 6806 Beam Power Valve for VHF Work. This Valve is Capable of Delivering a Synchronizing-Level Power Output of 28 Kilowatts at 550 Megacycles in Television Service.

It will be seen that the installation and operation of these high-power valves is a complex problem. This accounts for the fact that when a high-power transmitter is examined, less than half of it seems to be "active" circuitry, whilst the balance of the equipment is protective and ancillary equipment.

A further high-power valve, this time an air-cooled type, is shown in Fig. 2. This photograph shows a 6166, which has an integral radiator. This valve is a vhf tetrode with thoriated tungsten filament, designed for service in television and cw applications. The 6166 has a maximum plate dissipation of 10 kilowatts in those types of service, and its coaxial type construction facilitates operation at full input up to over 200 megacycles in class B and class C television service.

The 6166 shown in Fig. 2 is about 11½ inches overall height, and is intended to be mounted in a coaxial assembly which consists not only of connection rings for the plate and two grids, but also forms a duct through which air is driven up over the valve, and discharged at the top. Frequently, and especially where several such valves are used at one location, two sets of air ducts are provided. One introduces cool air through motor driven fans for cooling, the other extracts the heated air after is has passed over the valves and discharges it outside the building.

In cold climates the heated air from the transmitter is sometimes used to heat the station buildings. As in the case of water-cooled valves precautions must be taken by means of interlocks to prevent the valves being switched on or power applied in the event of failure of the cooling air supply. This is usually done by mounting a vane in the air duct, so arranged that when sufficient air pressure is available the movement of the vane actuates a switch to complete an interlock circuit.

Referring again to Fig. 2, the integral radiator, lifting handles and anode form one assembly. The filament terminals are seen at the bottom of the figure. The two grids are connected to rings fabricated into the glass envelope, grid No. 1 being at the bottom, and grid No. 2 about half-way between grid No. 1 and the anode assembly.

To complete our look at a few high-power valves, Fig. 3 shows a 6806 beam power valve. The 6806 is a water-cooled valve of unique design intended for use as a grid-driven power amplifier at frequencies up to 1000 Mc.

Unique in design, the 6806 features a coaxial-electrode structure in which the centrally located plate is surrounded by a symmetrical array of unit electron-optical systems. These embody a structural design which permits not only close spacing but also unusually accurate alignment of the electrodes. Furthermore, effective bypassing of grid No. 2 to cathode is provided by built-in capacitors. Ducts for water cooling the plate, the grid-No. 2 block, the grid-No. 1 block, the rf cathode terminals, and the filament-section blocks, are built in. This valve uses a multistrand thoriated-tungsten filament in a beam-forming assembly, similar to that described for the 5831. The filament is actually assembled in two discrete sections, which can be operated either in parallel, or in phase quadrature to reduce hum when ac filament heating is used.

The unusual construction of this valve requires some explanation. The anode connection is at the top of the valve, and includes two cooling water connections. The plate connection is insulated from the main body of the valve by means of a ribbed ceramic cylinder forming part of the valve envelope. The rf plate contact surface is the large outer diameter of the connector, whilst the threaded section is the dc plate connection.

The filament connections for the two sections are combined with the cooling water connections for them, and are shown in the photograph at the lower left. In between these two connectors is a shorter one, which is the common point for the two filament sections. The rf grid No. 1 connection is a large-diameter flange at the underside of the valve (partially obscured) whilst the dc connection and water connection to the grid is centrally disposed underneath. The layout of the grid No. 1 connections is similar to that for the plate. The bottom right hand connection shown in the photograph is a combined power and water connection for grid No. 2.

Whilst only three valves have been studied in this article, enough has been said it is felt to convey to the reader who is unfamiliar with this class of valve some of the design, installation and operating problems association with high power. It can readily be understood why both the manufacture and application of these valves is a highly-specialised part of the radio art.

THE JUNCTION TRANSISTOR

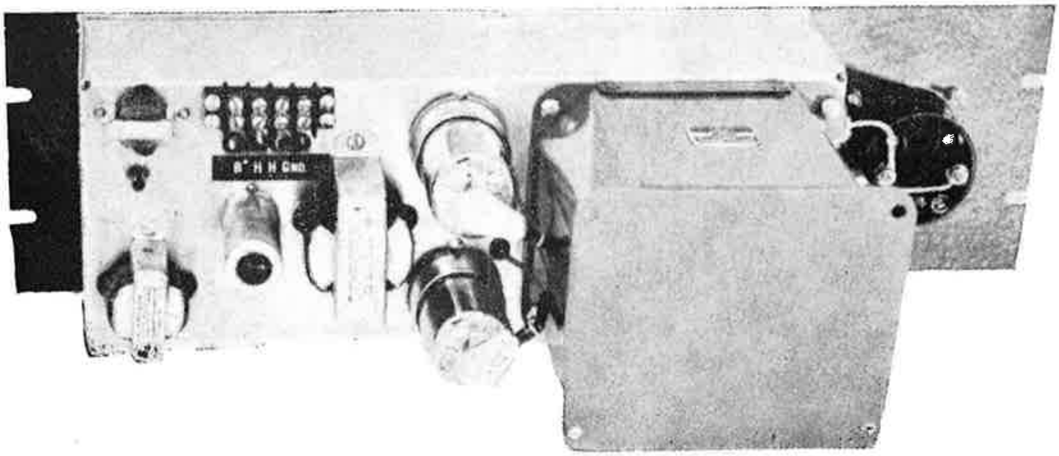
(Continued from page 33)

common-emitter amplifier is about 50 times greater than the current gain of the common-base amplifier, the available bandwidth of the common-emitter amplifier is about 50 times less than the available bandwidth of a common-base amplifier using the same transistor. The values chosen for the table represent the frequency response which could be expected from typical audio transistors. Much better frequency response can be obtained by using transistors specifically designed for high-frequency operation. These matters and others will be considered in the third part of this article.

(With acknowledgements to RCA)

VERSATILE MODULATOR

(January, 1960, p. 19)



Several readers have drawn our attention to the fact that the wrong illustration was used for Fig. 3 of this article. The correct illustration is shown here.


AWV