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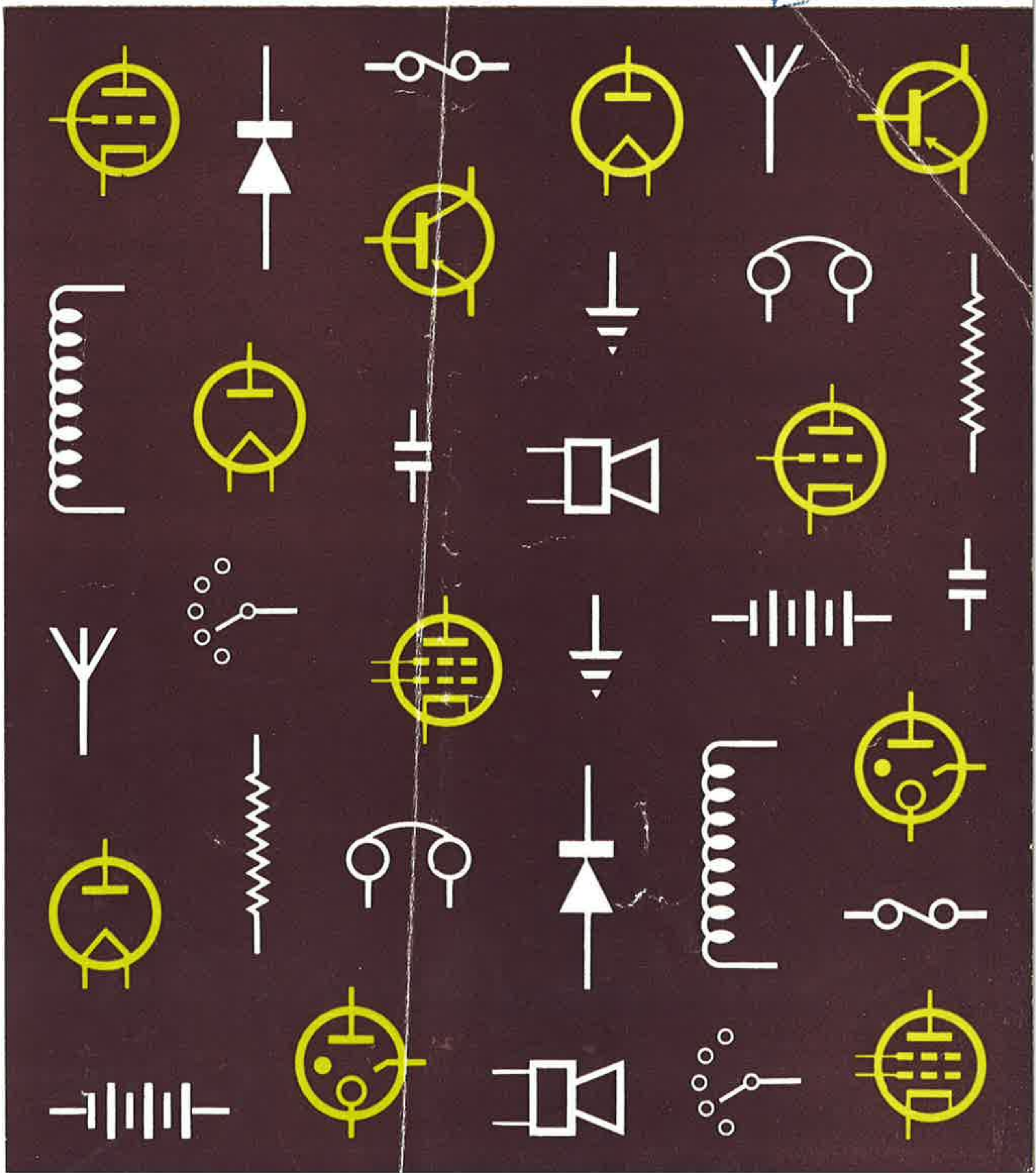

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IN THIS ISSUE

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A TV SOUND SECTION USING THE LOCKED-OSCILLATOR QUADRATURE-GRID DETECTOR

by R. A. Darnell, A.S.T.C., A.M.I.R.E. (Aust.)

Introduction

Efforts to produce a sound section for a television receiver which was simpler than the conventional ratio detector circuit, resulted in the introduction of the quadrature-grid detector circuit. The 6BN6, a "gated-beam-valve" was used widely in the United States, but suffered from the disadvantage that it required a minimum input signal of 1.25 volts rms. (3.5 volts peak to peak)*. In the Australian television system, the demands on the sound detector are greater because of the wider bandwidth required. The added bandwidth must be gained at the expense of sensitivity, and possibly suppression of unwanted amplitude modulation components.

The circuit which is described in this paper features the Radiotron 6DT6 and operates with input voltages to the driver stage down to 5 mv, with a bandwidth suitable for use in Australian television receivers.

Operation

The 6DT6 is a relatively low priced pentode in which the No. 3 grid is able to control the plate current. The turns in the No. 3 grid are more closely spaced than in other pentodes giving it a comparatively sharp cutoff characteristic. This is illustrated in Fig. 1, which shows the E_{c3}-I_b relationship for a typical 6DT6 compared with that of a typical 6AU6. The E_{c1}-I_b characteristics are also shown.

The ability of the No. 3 grid to control the plate current enables the two grids of the 6DT6 to form a gating action. Plate current will flow only when both grids are above their respective cutoff potentials relative to the cathode.

When a signal at the centre frequency (5.5 Mc) is applied to the No. 1 grid of the 6DT6 in the circuit illustrated in Fig. 2, space charge coupling within the valve causes a 5.5 Mc voltage to be developed across the tuned circuit (which is resonant at the centre frequency) at the No. 3 grid. This voltage is in quadrature with (i.e. 90° out of phase with) the voltage at the No. 1 grid. (See Appendix I) As the frequency of the incoming signal is varied the phase of the voltage at the No. 3 grid varies relative to that at the No. 1 grid. These variations in phase cause variations in the length of time in each cycle during which both grids are above their respective plate current cutoff potentials. The width of the pulses of plate current is thus varied. This is illustrated in Fig. 3 which shows the difference in width of the plate current pulses for input signals of slightly different frequency.

The voltage at the No. 1 grid is normally of sufficient amplitude to drive the grid to conduction. The resultant grid current flow has a damping effect on the driver transformer which, as will be discussed later, plays a part in the suppression of amplitude modulation. The RC bias network in the No. 3 grid circuit limits the flow of current in this grid and clamps the positive peaks of the No. 3 grid voltage at a potential which is dependent on the contact potential of this grid but approximately equal to the cathode dc potential.

The plate circuit consists of a 270,000 ohm resistor and a bypass capacitor. The effective

*Television Engineering Handbook. D.G. Fink — McGraw-Hill

Kriesler 16/2/61 Charlie 79/4

plate load resistance and the effective shunt capacitance from plate to ground form an integrating network which filters out the 5.5 Mc components. An audio voltage is produced across the capacitor. This audio voltage corresponds to the audio modulation on the 5.5. Mc carrier. It is produced from the plate current pulses by the integrating action of the RC circuit.

Due to the time constant of the plate circuit there is an attenuation of the higher audio frequencies relative to the amplitude of the output at mid-frequencies. This attenuation can be used to provide the 50 microsecond de-emphasis required in sound sections for use with the Australian television system.

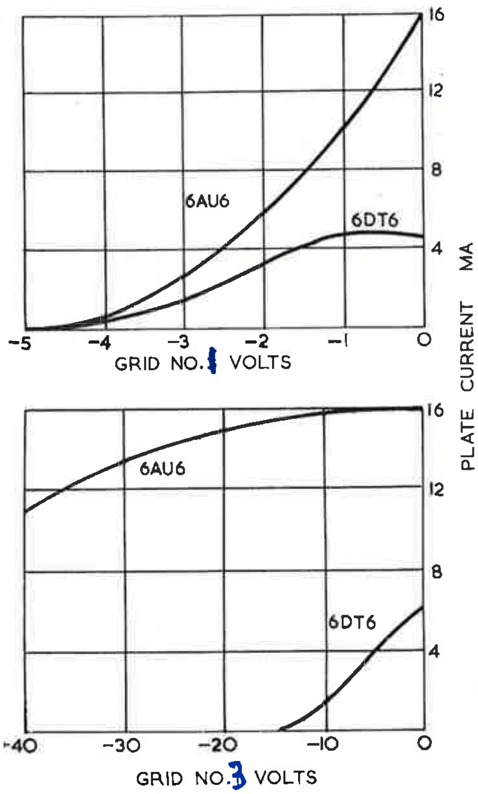


Fig. 1 — Plate Current Cutoff Characteristics.

The mode of operation described above is similar to that of other quadrature grid detectors, and exists for large input signals. As the input level falls, grid current flow in the No. 1 grid decreases causing less damping on the driver transformer. A level is reached below which the tuned circuit impedances and the coupling between them through the g_3-g_1 capacitance are such that the system functions as a locked oscillator detector. This mode of operation, in which the method of detection is similar to that of the high level mode, enables operation down to much lower input levels than is possible with other forms of quadrature grid detectors. The transition from one mode of operation to the other occurs gradually and is not accompanied by any noticeable changes in performance. If the input signal is removed the circuit continues to oscillate at approximately 5.5 Mc. There is a certain "minimum input level"* below which the oscillations will not lock to the incoming signal. This "minimum input level" increases as the amount of deviation on the input carrier is increased.

The sound take-off point in the receiver must provide a level well above this minimum during operation. The minimum level can be as low as approximately 0.5 millivolts in some circuits (using double tuned driver transformers). In the circuit described in this paper it varies from approximately 5 to 10 millivolts depending on certain variations in the circuit.

As long as the input level is above the "minimum locking level" the circuit will function with an output which is independent of input level but there is an increase in bandwidth as the level is increased. The bandwidth reaches a steady value at a level about six times greater than the "minimum locking level."

Alignment

The alignment of the locked oscillator quadrature grid detector circuit is quite simple when the correct procedure is used. It has been common practice to align with a frequency modulated input signal, observing the audio output waveform on an oscilloscope. If this method is used the procedure is as follows:—

1. With a large input — say 100 mv at 5.5 Mc, frequency modulated with a deviation of ± 15 Kc — to the grid of the driver stage, align the quadrature coil, L_1, for maximum undistorted output observed at the top of the volume control which must be set below maximum volume to avoid grid clipping by the output stage. This step eliminates the possibility of tuning to a wrong peak which will provide very poor performance.

2. Reduce the input to about 10-20 mv and carefully tune the driver transformer, T_1, to a point midway between the extremes at which the audio output waveform begins to "break up" on the peaks.

3. Readjust L_1 if necessary — this adjustment will be more sensitive at the lower input level.

This method will be referred to as the "dynamic" alignment procedure.

The disadvantage of this method on a production basis is the need for a frequency modulated signal source. Another, and much simpler

*Unless otherwise specified the term "input level" always refers to the input to the 6AU6 grid.

method is to use an unmodulated 5.5 Mc signal and use as an indicator the dc bias built up across the No. 3 grid bias resistor. As this resistor is normally of the order of 100,000 ohms to 560,000 ohms a high impedance meter is necessary. The procedure is then similar to the previous one:—

1. With 100 mv input signal tune L_1 for maximum negative dc voltage. (This should be of the order of -2 to -3 volts.)
2. Reduce input to approximately 10 mv and adjust T_1 for a maximum reading on the meter. There may be a small range over which the coil can be tuned at the peak without any noticeable change in output. This range increases with increased input level or with valves (especially driver) of higher transconductance. The core should be set for the middle of this range.
3. Readjust L_1 if necessary.

This method will be referred to as the 'static' alignment procedure.

Use of the different alignment techniques does not provide the same operating point and circuit performance. In general, the dynamic alignment

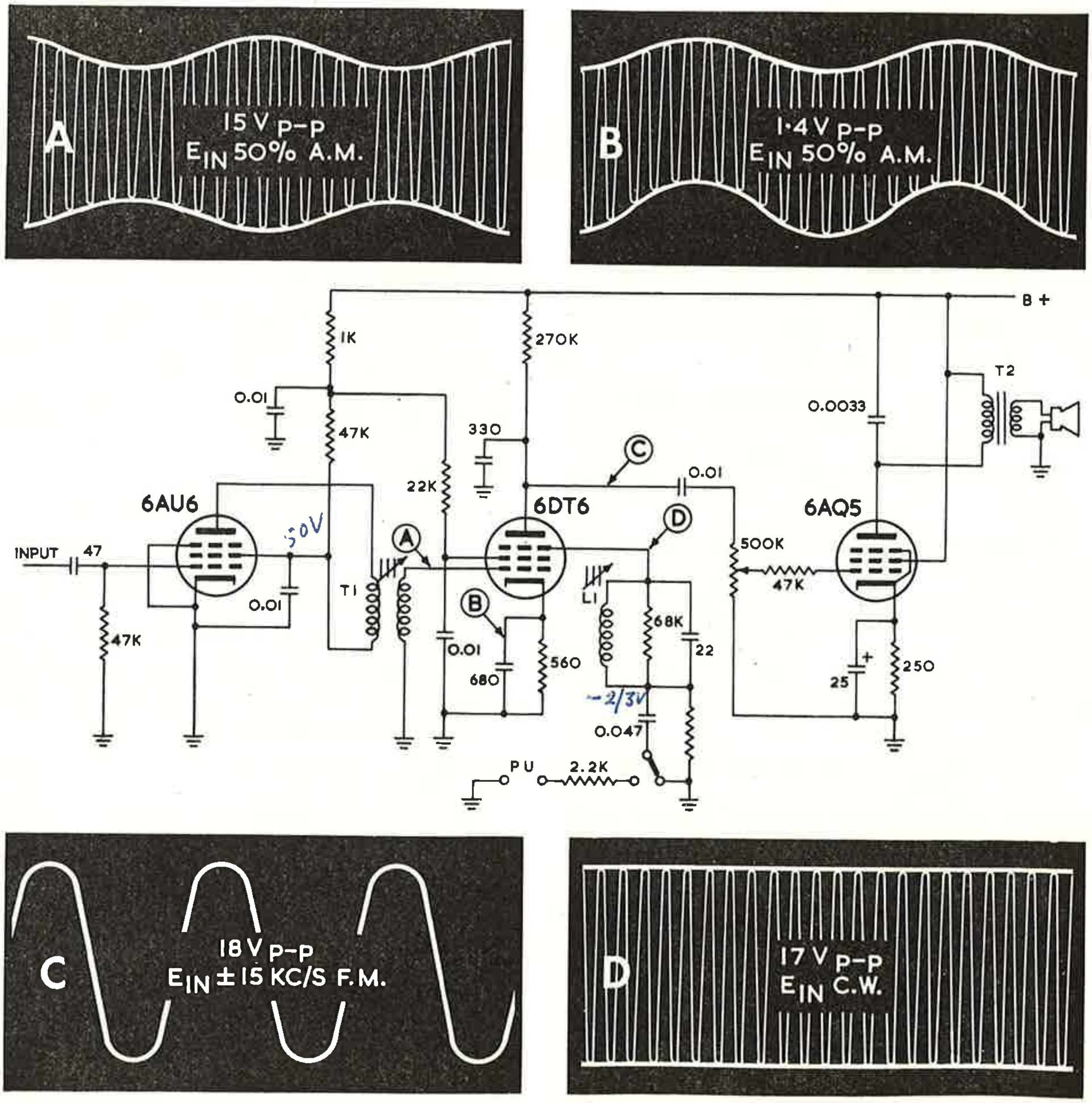


Fig. 2 — A Typical TV Receiver Sound Section Using the 6DT6.

gives a greater bandwidth, but poorer AM rejection and a higher minimum locking level than the static alignment. This will be illustrated later when the performance of the circuit is listed. The static alignment is more definite and because of this allows much better repeatability for measurements. This was found to be quite important in the measurement of AM rejection and distortion.

The differences in performance for the two alignment methods are due to a slightly different type of operation. This is discussed in Appendix I.

Driver Stage

The 5.5 Mc sound carrier can be obtained from either the video detector or the output of the video amplifier through suitable coupling circuits. When the output is taken from the video detector some form of voltage step up is required to ensure sufficient input to the driver stage to provide good quality sound under all conditions and with all signal strengths.

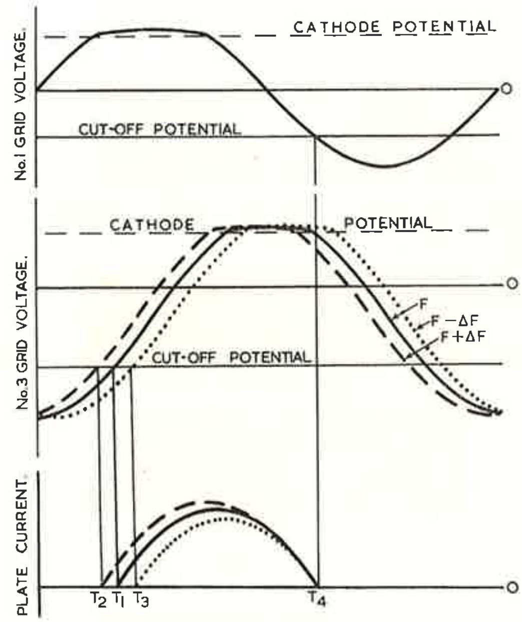
When the sound carrier is taken from the video-amplifier plate, in which case the 5.5 Mc trap is normally used as the plate load to provide a high gain at this frequency, an RC network with a time constant of the order of two to three microseconds is included to provide more effective limiting. The need for limiting to increase the rejection of amplitude modulation is greater when the video amplifier is used to amplify the sound carrier. This is due to the increased degree of amplitude modulation which results from non-linearity in this stage. The degree of modulation varies and depends on the operating point of the valve, but it can on occasions be several times higher than when the take-off point is before the video amplifier.

In Fig. 2, the input circuit is shown in a form suitable for use with a coupling network connected to the video amplifier plate. The screen of the 6AU6 is kept to about 50 volts by the 47,000 ohm dropping resistor to assist the limiting action. When the sound carrier is taken from the video detector through a step-up transformer, a 68 ohm cathode bias resistor bypassed by 0.01 microfarads must be included in the cathode lead. The screen voltage can also be raised to say, 100 volts.

The plate circuit can be either a double-tuned transformer or a tightly coupled single-tuned transformer. The former gives more sensitivity allowing a lower "minimum locking level." It is more critical to adjust, and, as it involves two tuning cores instead of one, necessitates more alignment time. The single-tuned transformer has been used because it is felt that the advantages in lower cost and alignment simplicity are important factors. Furthermore it is felt that there is no advantage in making the circuit function with input levels to the driver stage of less than say, 10 millivolts.

Suppression of AM

The variable damping on the driver transformer due to changing grid current as the amplitude of the incoming signal varies, tends to smooth out these amplitude variations. The effectiveness of this action in reducing the percentage AM is illustrated in Fig. 4.



T_1 — PLATE CURRENT FLOW COMMENCES FOR SIGNAL AT CENTRE FREQUENCY F

T_2 — PLATE CURRENT FLOW COMMENCES FOR SIGNAL AT FREQUENCY F + ΔF

T_3 — PLATE CURRENT FLOW COMMENCES FOR SIGNAL AT FREQUENCY F − ΔF

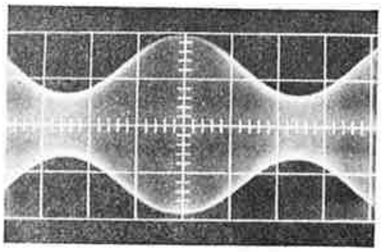
T_4 — PLATE CURRENT FLOW CUTOFF BY NO. I GRID

Fig. 3 — Gating Action of g1 and g3, Showing Variations in Plate Current Pulse Width for Signals of Different Frequency.

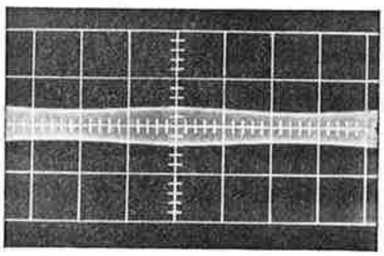
At high input levels, cathode degeneration augments the ability to suppress AM. As the cathode resistor is bypassed to rf, only an audio voltage is developed across the bypass capacitor which corresponds to the amplitude of the modulation envelope. This voltage affects the cathode to No. 3 grid potential in such a way as to reduce the effect on the plate current of changes cathode current. As the amplitude of the cathode current changes the plate current remains relatively constant while the screen current varies. This will be discussed in more detail in a later section.

The ability of the circuit to suppress amplitude modulation, or as it is more commonly called, the

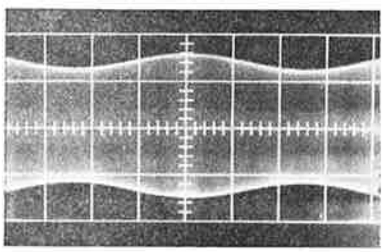
AM rejection, is affected by several factors. As the grid damping plays a major part, the source impedance looking back from the grid of the 6DT6 is important. The turns ratio of the driver transformer, which consists of two tightly coupled windings was found to be an optimum for a step down of 1.1:1. This ratio was not extremely critical, but did yield optimum performance for AM rejection, bandwith and sensitivity.



(a) Valve cold. Vertical scale 20 volts per division.



(b) Valve normal. Vertical scale as for (a).



(c) Valve normal. Vertical scale 5 volts per division. Input to 6AU6 grid, 100 mv, 50% AM at 1000 cps.

Fig. 4 — Effect of Grid Damping in Reducing Amplitude Modulation.

Others factors are the Q of the quadrature grid circuit, the effectiveness of the cathode degeneration which will be discussed later, and the screen voltage. The Q of the quadrature grid circuit is affected by several parameters but it may be said that as the Q becomes higher, AM rejection becomes slightly better, but the bandwidth narrows.

The choice of screen voltage is not easy as it will depend to some extent on the particular requirements in the integrated television receiver design. In the circuit in Fig. 2 the screen voltage varies between about 140 and 150 volts for different valves and different input levels. (It was the use of this high screen voltage which governed the choice of the cathode bias resistor. A 560 ohm resistor is needed to keep the screen dissipation at a safely low value.) It was found that a low screen voltage, say 100 volts, gave better AM rejection at low input levels, say 10-20 millivolts, but worse at high levels, say 100 millivolts and higher, than the higher screen voltage. As it is considered that the input will always be above, say 50 millivolts, the higher screen voltage was chosen. If a manufacturer found that under some extremes of fringe area reception the sound was weak and added AM rejection was necessary, and, if at the same time he felt that the AM rejection could be lowered under normal signal conditions, a lower screen voltage might be chosen. Such a decision could only be made after exhaustive fringe area tests and is outside the scope of this paper.

Effect of Parameters on Performance

In this section the discussion turns on the effect of several parameters on the performance of the circuit.

L/C Ratio of Quadrature Grid Circuit

The L/C ratio of the quadrature (No. 3) grid tuned circuit was varied, all other components remaining unaltered. As the L/C ratio was increased, C being made to vary from 39 $\mu\mu$f to 12 $\mu\mu$f, the audio output voltage decreased by about 25%, the distortion for large deviations decreased, the AM rejection at low input levels decreased slightly and the bandwidth increased.

If the tuning capacitor was decreased below about 10 $\mu\mu$f the circuit became very difficult to tune, there being quite a noticeable "pulling" as either driver or quadrature coil was tuned.

Quadrature Circuit Q

The Q of the quadrature grid circuit can be controlled to some extent by the damping resistor across the coil. This has more effect on the performance at low input levels. As the level of the incoming signal increases, grid current flow in No. 3 grid circuit increases thus increasing the damping on the tuned circuit.

Lowering the Q of the quadrature grid circuit will cause an increase in the bandwidth, a decrease in the audio output voltage and AM rejection, an improvement in linearity, and it will decrease the "minimum locking level."

Quadrature Grid Bias Resistor

Increasing the resistance of the quadrature grid bias resistor limits the grid current flow and thus increases the effective Q of the quadrature grid circuit. The effect on the circuit will be more noticeable at medium input levels where the grid current flow is sufficient for the dynamic impedance at the grid to lower the effective Q of the tuned circuit.

The reduction in bandwidth experienced when this resistor was changed from 100,000 ohms to 470,000 ohms was nearly two to one under most conditions. The increase in AM rejection was very slight, and occurred only at very low input levels. The choice of this resistor is thus governed almost solely by the bandwidth required. It must be chosen in association with the choice of Q of the quadrature grid coil, but is the more important factor, especially at medium and high input levels.

Cathode Bias Resistor

An increase in the resistance of the cathode bias resistor over the range from 330 ohms to 560 ohms does not appreciably affect the performance. At low input levels there is a slight reduction in both AM rejection and bandwidth. Below 330 ohms the performance is adversely affected. The bandwidth becomes broader, but the AM rejection drops.

A value of 390 ohms was originally chosen for optimum performance. The value was changed to 560 ohms later to keep the screen dissipation within the maximum rating when the screen voltage was increased. The reason for this increase in screen voltage has already been discussed.

Cathode Bypass Capacitance

Current practice overseas is to use an 0.01 microfarad capacitor to bypass the cathode resistor. It was found, during the development of this circuit, that the amount of AM rejection decreased with increasing amplitude modulating frequency at high input levels. This was occurring with input levels of the order of 0.5 volts rms and higher — levels which can quite easily be obtained at the plate of the video amplifier. This phenomenon would be particularly undesirable in a television receiver because the amplitude modulation impressed upon the 5.5 Mc sound carrier is commonly caused by the "vertical sync block" and line sync pulses. These are of the step function form and contain many high order harmonics. If the AM rejection for high audio modulating frequencies is reduced, the sharp edges at the beginning and end of the vertical block would not be sufficiently attenuated. A series of pulses would then appear in the output, being repeated at the field repetition rate, and thus cause buzz.

By reducing the cathode bypass capacitance to 680 $\mu\mu$f an improvement in AM rejection of up to 7 db was obtained at high input levels. Reference to the cathode voltage waveform in Fig. 2 will show a considerable amount of 5.5 Mc at the cathode. This is because the impedance of the 680 $\mu\mu$f bypass is approximately 40 ohms at 5.5 Mc thus not providing complete bypassing. The 680 $\mu\mu$f capacitor was chosen experimentally. The 5.5 Mc component at the cathode did not affect the performance in any way. The improvement in AM rejection with the smaller bypass capacitor is explained in the following manner:—

At high input levels the AM rejection is augmented by cathode degeneration as mentioned previously. An audio voltage appears across the cathode bias network which follows the amplitude variations of the incoming signal. The amplitude variations are already reduced considerably by the grid damping action. The audio voltage at the cathode affects the cathode to No. 3 grid voltage in such a way as to oppose the plate current variations introduced by the remaining amplitude variations between grid No. 1 and cathode.

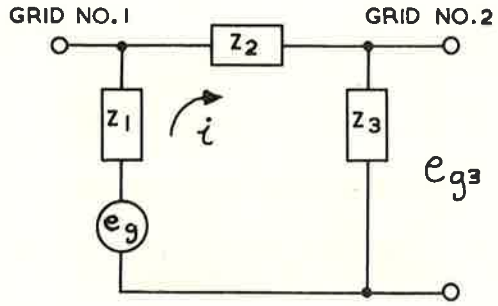


Fig. 5 — Simplified Equivalent Circuit for Calculation of g1 — g3 Phase Shift.

For this to be fully effective, the cathode must be unbypassed at audio frequencies. That is, the cathode impedance must appear purely resistive at audio frequencies. With a cathode resistor of 560 ohms, bypassed by 0.01 microfarads, the phase shift in the cathode impedance for a 1000 cycle per second amplitude modulating frequency is approximately 2 degrees. The phase of the voltage developed across the cathode resistor will lag the phase of the undesired plate current component by an angle which is dependent on, but less than, the phase angle of the cathode impedance. (This is briefly explained in Appendix II). In this case it will be less than 2 degrees. The plate current component which is due to the No. 3 grid to cathode voltage, is, for best cancellation and hence maximum AM rejection, required to be 180 degrees out of phase with the undesired plate current component, but is now shifted by an additional angle of "less than 2 degrees." The degree of cancellation of the two components will not be affected to any appreciable extent by this small additional phase shift and hence the effect on AM rejection will be negligible.

For an amplitude modulating frequency of 7,500 cycles per second, the phase angle of the cathode impedance would be 15 degrees. The additional phase shift between the two plate current components would now be about 10 degrees which becomes quite significant. (This figure is based upon calculation for 5 Kc in Appendix II) The 10 degrees additional phase lag between the two plate current components will result in an added and unwanted component of plate current at the modulating frequency, thus reducing the effectiveness of the cathode degeneration. For two components of equal amplitude, the amplitude of this unwanted component due to the 10 degrees phase shift would be about 20 per cent of the amplitude of each component. At higher audio frequencies the unwanted phase shift would be higher still, and the AM rejection would be reduced further still.

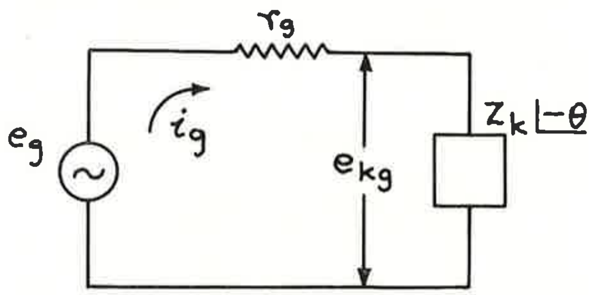


Fig. 6 — Simplified AF Equivalent Circuit of 6DT6 g1—k Impedance.

With the bypass capacitor reduced to 680 $\mu\mu$f the additional phase shift at 7,500 cycles per second is much less than one degree. This will have a negligible effect on the AM rejection.

Plate Circuit

It is quite common in American television receivers to see the plate load resistor fed from the B Boost point in the horizontal deflection circuit, and also to see a plate bypass capacitor of the order of 0.001 microfarads. The high supply voltage allows a higher load resistance to be used, and hence a larger audio output voltage to be obtained. In the circuit described in this paper, sufficient audio output voltage to drive the 6AQ5 output valve to the maximum output of which it is capable can be obtained with a deviation of little more than $\pm$15 Kc. As the maximum deviation used is $\pm$50 Kc, this allows a large safety factor and so no attempt has been made to use the higher supply voltage. Also, the additional drain on the horizontal deflection circuit is avoided.

It was mentioned earlier, in the section entitled "Operation," that the time constant of the plate circuit can be used to provide the necessary de-emphasis. This time constant is equal to the product of the total shunt capacitance from plate to ground and the effective plate load resistance. The shunt capacitance is made up of a fixed capacitor, and the parallel combination of the output capacitance of the 6DT6, the input capacitance of the 6AQ5 and the stray capacitance to ground. The latter three components total 40 to 50 $\mu\mu$f. The effective plate load resistor is the parallel combination of the load resistor, 270,000 ohms, the grid leak of the 6AQ5, 500,000 ohms, the plate resistance of the 6DT6 (about 750,000 ohms) and the input resistance of the 6AQ5 which is sufficiently high at audio frequencies to be neglected. This gives an effective resistance of approximately 140,000 ohms.

Hence, for a 50 microsecond time constant

$$C = \frac{50 \times 10^{-6}}{140 \times 10^{3}}$$
$$= 360 \times 10^{-12}$$

Hence the fixed capacitance is required to be about 310 to 320 $\mu\mu$f. A 330 $\mu\mu$f capacitor has been used.

With a 0.001 microfarad capacitor, the de-emphasis would be 150 microseconds with the same load resistor. This will cause quite a severe treble cut which may be considered undesirable. By using a 330 $\mu\mu$f capacitor the audio output voltage is increased but the performance is otherwise relatively unchanged. Having a de-emphasis of 50 microseconds just compensates for the pre-emphasis introduced at the transmitter and allows an effectively flat overall frequency response from the sound pick up at the television studio to the grid of the 6AQ5. The effective frequency response at the speaker is then solely dependent on the audio output stage and speaker, and can be as good as the manufacturer is prepared to make it.

A plate shunt fixed capacitor of 100 $\mu\mu$f can be used to allow a "lift" at high frequencies. A relatively low cost output stage with inherent attenuation of high frequencies can now be used, the fall off due to this stage providing part of required de-emphasis. The effect can be an overall de-emphasis of close to 50 microseconds, and hence good overall frequency response.

Use of a Pickup

If pick-up terminals for use with a record player are to be provided, the 6DT6 can be used as an audio frequency amplifier. The switching operation is very simple. The suppressor grid is used as the input grid. The 0.047 microfarad capacitor across the 100,000 ohm bias resistor is disconnected from earth and used as the audio coupling capacitor. A 2200 ohm resistor is necessary in series with this capacitor to ensure complete freedom from 5.5 Mc oscillations. This is illustrated in Fig. 2.

Performance

For all measurements below, the input quoted is the input to the grid of the 6AU6. The results for the two different alignment methods are given.

Minimum Locking Level. Measured for $\Delta f = \pm 15$ Kc, 8 mv (dynamic alignment), 5 mv (static alignment).

Audio Output Voltage. Measured at the top of volume control at various input levels for $\Delta f = \pm 15$ Kc, 24.5 v p-p (dynamic), 18 v p-p (static).

Distortion. Measured at 6DT6 output for input; 25 mv, $\Delta f = \pm 15$ Kc, 0.6% (dynamic), 2.2% (static); 100 mv, $\Delta f = \pm 15$ Kc, 0.65% (dynamic), 2.4% (static); 25 mv, $\Delta f = \pm 50$ Kc, 4.5% (dynamic), 6.7% (static); 100 mv, $\Delta f = \pm 50$ Kc, 5.7% (dynamic), 7.4% (static).

AM Rejection. The output for a 30% AM input compared with the output for a frequency—modulated input with $\Delta f = \pm 15$ Kc; 20 mv 29.5 db (dynamic), 29 db (static); 100 mv, 29 db (dynamic), 31 db (static).

Bandwidth. This is the maximum deviation at any specified input level that can be used before the output waveform (observed on an oscilloscope) shows signs of "breakup" or flattening on the peaks; 20 mv, ±40 Kc (dynamic), ±55 Kc (static); 25 mv, ±50 Kc (dynamic), ±72 Kc (static); 50 mv, ±100 Kc (dynamic), ±84 Kc (static); 100 mv, ±105 Kc (dynamic), ±84 Kc (static).

DC Operating Conditions. Refer to Table I.

TABLE I

INPUT	6AU6	6DT6			
	E_{c2}	E_{c2}	E_b	E_k	E_{c3}
0	56V	146V	87V	3.4V	−2.1V
25 mV	56V	142V	91V	3.5V	−2.1V
50 mV	57V	142V	92V	3.6V	−2.2V
100 mV	57V	137V	93V	3.8V	−2.2V

Coil Specifications

Driver Transformer T_1. Primary, 140 turns, single layer, close wound; secondary, 127 turns, close wound directly over primary; wire, 40 SWG, SSE; former, 5/16" OD; Core, Neosid grade 500.

Quadrature Coil L_1. L_1 must tune at 5.5. Mc with 22 $\mu\mu$f. The Q measured at 5.0 Mc with Q—Meter tuning capacitor of 40 $\mu\mu$f, with damping resistor and in can, should be 45. For a former with 5/16" OD, and a Neosid grade 500 core, L_1 has 70 turns of 40 SWG SSE wire close wound in a single layer.

APPENDIX I

Before reading this appendix it will be necessary to read the sections of this paper on "Operation" and "Alignment."

The phenomenon of space charge coupling has been discussed by Herold in relation to the operation of the pentagrid converter in AM broadcast receivers.* While Herold claims that "the use of an equivalent impedance to explain the behaviour of 'space charge coupling' is somewhat artificial," he does indicate that the effect is similar to that of a "negative capacitance" connected between the No. 1 and No. 3 grids, and can be cancelled by the addition of a small capacitor of the correct magnitude between the two grids.

With the 6DT6 operating at a centre frequency of 5.5 Mc the same analogy can be used except that the "negative capacitance" appears to have a fairly low Q. The addition of a small trimmer between the No. 1 and No. 3 grids will show a gradual reduction in the amplitude of the voltage at the No. 3 grid as the capacity of the trimmer is increased. A point of minimum amplitude will be reached after which the voltage begins to rise again, but with a phase shift of between 120 degrees and 150 degrees relative to that which exists with the trimmer at minimum capacity.

This indicates that the "negative capacitance" is shunted by a resistive component — without this resistive component one would expect a phase shift of 180 degrees.

To explain how the 90 degrees phase shift occurs betwen the voltages at the No. 1 and No. 3 grids a simplified equivalent circuit has been drawn — see Fig. 5. Several explanations are necessary before proceeding. Reference to either of the alignment procedures given previously will show that the final adjustment is made with a small input signal to the driver stage grid. The valve operates then in the locked oscillator mode in which case the circuit resembles a "tuned plate-tuned grid" oscillator, the No. 3 grid acting as the "plate" and the No. 1 grid as the "grid."

For the static alignment procedure the No. 3 grid current is observed by reading the voltage developed by it across the No. 3 grid bias resistor. The circuit is thus tuned for maximum amplitude of oscillation in which case the No. 3 grid tuned circuit will be tuned to a frequency slightly higher than the centre frequency of the input signal. At the centre frequency then, the impedance of this circuit is inductive.

*Frequency Converters for Superheterodyne Reception. E. W. Herold, Proc. I. R. E. Vol. 30 Feb., 1942.

For the dynamic alignment procedure the No. 3 grid tuned circuit is tuned to a lower frequency and at the centre frequency the impedance may be either less inductive than for the static alignment case, resistive, or even slightly capacitive.

If capacitive, there could be an effective series resonance occurring between the "negative capacitance" of the space charge coupling and the capacitive impedance of the tuned circuit. As in any of the three cases the phase angle of the impedance will be close to zero, for purposes of illustration the resistive case can be used without affecting the results to any appreciable extent.

The following considerations will be confined to medium and high input level operation in which case it will be assumed that $\bar{Z}_1$, the impedance of the No. 1 grid tuned circuit becomes small enough to neglect. It will be also assumed that $\bar{Z}_2$, which represents space charge coupling, is purely reactive. While this latter assumption is false, it will simplify the results, and the effect on them will be small.

Let us first consider that case when the circuit is aligned using the dynamic alignment procedure. Referring to Fig. 5:—

$\bar{Z}_1$ will be neglected.

$$\bar{Z}_2 = jXc = Z_2 \underline{|90^\circ}$$

$$\bar{Z}_3 = R = Z_3 \underline{|0}$$

If $|Z_2| >> |Z_3|$, $\bar{i}$ will have a phase angle relative to $\bar{e}_g$ of a little less than 90 degrees —

$$\bar{i} = i \underline{|-(90-\theta)} \text{ where } \theta \text{ is small}$$

$$\text{Then} \quad \bar{e}_{g3} = i \,.\, Z_3 \underline{|-(90-\theta)}$$

In practice, as far as could be determined accurately, the voltage at the No. 3 grid actually lagged that at the No. 1 grid by between 75 degrees and 85 degrees.

Consider now the case when the circuit is aligned using the static alignment procedure:—

$\bar{Z}_1$ will be neglected

$$\bar{Z}_2 = +jXc = Z_2 \underline{|90^\circ}$$

$$\bar{Z}_3 = Z_3 \underline{|\alpha} \quad \text{where } \alpha \text{ is small}$$

If $|Z_2| >> |Z_3|$, $\bar{i}$ will have a phase angle relative to $\bar{e}_g$ of a little less than 90 degrees —

$$\bar{i} = i \underline{|-(90-\beta)} \text{ where } \beta \text{ is small}$$

$$\text{Then} \quad \bar{e}_{g3} = i \,.\, Z_3 \underline{|-(90-\beta-\alpha)}$$

A measurement of the phase lag for this case showed an angle of approximately 70 degrees.

The effect of the resistive component of $\bar{Z}_2$ which, for the calculation was neglected, would be to reduce slightly the phase angle of $\bar{i}$ and hence $\bar{e}_{g3}$ in each case.

The output voltage $\bar{e}_{g3}$ would, for the equivalent circuit shown, be much smaller than $\bar{e}_g$. This is not so in practice. The equivalent circuit of Fig. 5 should contain a multiplying term which will be dependent on the grid No. 1 transconductance thus making $\bar{e}_{g3}$ much greater in magnitude. This term was omitted for simplicity.

APPENDIX II

It was stated that the phase angle between the audio voltage developed across the cathode impedance and the undesired plate current component is less than the phase angle of the cathode impedance. This is because this voltage is the vector sum of two voltages which will be designated $\bar{e}_{kp}$ and $\bar{e}_{kg}$.

Let us consider $\bar{e}_{kp}$ first. This is the voltage developed across the cathode impedance by the undesired component of plate current due to the amplitude variations at the No. 1 grid. This component of plate current will be in phase with the amplitude variations at the grid and will be used as the reference for all phase angles quoted. If the phase angle of the cathode impedance at a given audio frequency is $-\theta^\circ$, the voltage across this impedance due to the plate current component will be

$$\bar{e}_{kp} \qquad e = i_p Z_k \underline{|-\theta} = e_{kp} \underline{|-\theta}$$

The second voltage component, e_{kg}, is due to grid current flow. When the input signal is large, the magnitude of the grid current can be quite appreciable and can account for a voltage component at the cathode which may be as large or possibly greater than 0.5 e_{kp}. The flow of grid current is determined by the effective audio voltage at the grid due to amplitude variations on the incoming signal, and the impedance which is made up of a resistive component and the cathode impedance in series as illustrated in Fig. 6. The resistive component r_g, is the effective forward resistance of the diode formed by the grid and cathode and it can be quite low (say 500 to 1000 ohms) with grid currents of the magnitude being experienced in this circuit.

From Fig. 5 it can be seen that

$$\bar{i}_g = \frac{e_g}{r_g + Z_k \underline{|-\theta}} = i_g \underline{|\phi} \quad \text{where } \phi < \theta$$

$$\text{Then } \bar{e}_{kg} = i_g \underline{|\phi} \,.\, Z_k \underline{|-\theta} = e_{kg} \underline{|(\phi-\theta)}$$

Since $\phi < \theta$

$$\overline{e}_{kg} = e_{kg} \underline{|-\alpha} \quad \text{where } \alpha = \theta - \phi$$

Hence the component of the cathode voltage due to grid current flow, $\overline{e}_{kg}$, lags the plate current component, which we have established as our reference, by an angle which is less than θ, the phase angle of the cathode impedance.

The resultant cathode voltage $\overline{e}_k$ is the vector sum of $e_{kp} \underline{|-\theta}$ and $e_{kg} \underline{|-\alpha}$ and its phase angle will be between $-\theta$ and $-\alpha$.

By making several approximations, $\overline{e}_k$ was calculated for an audio frequency of 5000 cycles. The values obtained gave a resultant phase angle at the cathode of -6.5 degrees. The phase angle of the cathode impedance at 5000 cycles per second is $\theta = 10$ degrees. This phase shift is very difficult to measure with any real degree of accuracy because of the low amplitudes involved. The measurement was made as accurately as the conditions allowed and the phase shift was found to be approximately -6 degrees which agrees with the calculated value. Even so, because of the inaccuracy of the measurement and the approximations made in the calculation it is felt that it could vary from, say, 4 degrees to 8 degrees or even more under the conditions measured.

Also, while θ is fixed at any given frequency, ϕ will depend on the value of r_g and hence i_g which in turn depends on the input level and the degree of modulation.

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RECENT DEVELOPMENTS IN MAGNETRON DESIGN

Among the 1959 winners of the John Scott awards for developing inventions for the benefit of mankind are Dr. Henry Boot, of the Services Electronic Research Laboratory at Baldock U.K., and Professor John Randall, Wheatstone Professor of Physics at the University of London. A recent report by Reuter from America announced that the two scientists were cited jointly for their invention of the cavity magnetron, which was used extensively in narrow beam radar during the Second World War for the precise location and tracking of aircraft. During the post war years, Dr. Boot and his colleagues at S.E.R.L. carried out further work on this device and developed a tube of greatly improved performance giving high peak power outputs combined with extreme stability.

The English Electric Valve Company were quick to recognise the importance of these advances and in 1955 work was started on the new magnetron. Subsequent extensive development work has culminated in the wide range of reliable high performance magnetrons now offered by them, the only company in the world engaged in the quantity production of magnetrons of this type.

*

The remarkable performance afforded by this type of magnetron, which was described recently in a paper by Dr. Boot and other authors* to the Institution of Electrical Engineers, has resulted in its incorporation in the design of nearly all new high power S-band radars. Peak power outputs of well over 5 megawatts, with completely unprecedented stability, have been obtained from recent development samples. A further outstanding feature has been the remarkably high average life results attained. These have been as high as 10,000 hours and even these figures, which are exceptional at such high output levels, are likely to be exceeded as further life reports are received.

Extensive tests, carried out both in U.K. and America, have shown that a major break through in the technique of magnetron manufacture has been achieved by the English Electric Valve Company. This is exemplified by the 7182 (M543), a recent addition to the range of E.E.V. magnetrons, which offers longer life, greater pulse length, lower missed pulse count, and less frequency change (either during a pulse or during life) than any other magnetron hitherto produced.

* A New Design of High-Power S-Band Magnetron. H. A. H. Boot, B.Sc., Ph.D., H. Foster and S. A. Self, B.Sc.

5-BAND MOBILE TRANSMITTER

A 50-Watt Rig for Phone & CW Operation

by George D. Hanchett, W2YM

The 50-watt transmitter described in this article started out to be a simple single-band mobile job for use on a fishing trip. As the design progressed, however, more and more features seemed necessary or desirable, and it finally emerged as a five-band, crystal-controlled rig for phone and CW operation on 80, 40, 20, 15, and 10 metres.

This transmitter features a bandswitching system which automatically provides the proper drive for the final on each band, and remote control of practically all operating functions from a dashboard control unit. It was designed to operate from a 12.6-volt car battery and 450-volt, 250-milliampere dynamotor. With minor modifications, as described later, the rig can be operated from dynamotors or other plate-supply sources delivering as little as 300 volts.

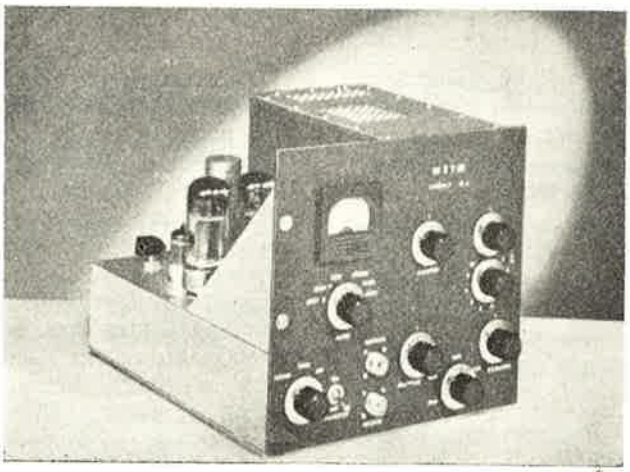

Fig. 1 — Front View of the 50-watt Mobile Transmitter.

The circuit of the transmitter is shown in Fig. 7. For its structural features and layout, see the photographs provided, Figs. 1, 4, 5, 6.

The rf section consists of a crystal-oscillator stage using a 7056, a buffer-frequency-multiplier stage using a 7054, and a final stage using a 6883. The modulator section includes a two-stage voltage amplifier using a 7058 twin triode, and a class AB1 output stage using two 7027-A's.

Recently introduced types, the 7054, 7056, and 7058 are similar, respectively, to the 12BY7A, 6CB6, and 12AX7, but specially designed for use in mobile communications equipment operating from 6-cell storage batteries. These types have heaters which operate dependably at any voltage between 12 and 15 volts and can withstand momentary excursions from 11 to 16 volts. The 6883 is the 12.6-volt equivalent of the 6146.

The 7027-A's used in the output stage of the modulator are beam power valves designed especially for use in high-fidelity applications. They have characteristics similar to those of the 6L6-GB but with substantially higher plate-voltage and grid-No. 2 voltage ratings (600 volts and 500 volts, respectively) and higher power-output capabilities in class AB1 service. They were selected for use in this transmitter because their high plate- and grid-No. 2 voltage ratings permitted them to be operated directly from the 450-volt supply, and because they easily provide the audio power required for 100% plate and grid-No. 2 modulation of the 6883.

Keying for CW operation is accomplished in the cathode circuit of the final amplifier.

The power-supply unit, shown schematically

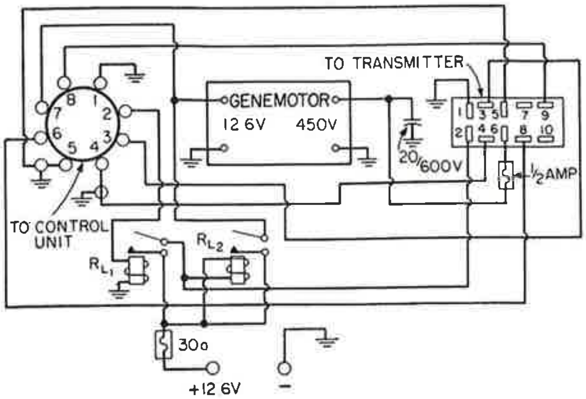

Fig. 2 — Power supply for the Mobile Transmitter.

in Fig. 2, contains the dynamotor, a filter capacitor for the 450-volt line, the relays used to open and close the main battery and dynamotor-input circuits, and fuses for these circuits.

The dashboard control unit, shown schematically in Fig. 3, contains the heater-and-plate-power on-off switches, a crystal-selector switch, the receiver-B+ switch and relay, and a switch used to transfer the car speaker from the broadcast receiver to the communications receiver and vice versa. It also contains a heater-circuit fuse, pilot lamps showing the conditions of the heater and plate-supply circuits, and the input connectors for the microphone and key.

Circuit Details

The transmitter is designed to use 3.5-Mc crystals for 80 metres, 3.5- or 7-Mc crystals for 40 metres, and 7-Mc crystals for all other bands. In the 80-, 40-, 20-, and 15-metre positions of the bandswitch (S1), the oscillator output is untuned. In the 10-metre position of the band-switch, the oscillator output is tuned to twice the crystal frequency — that is, to 14 Mc — by C6 and L1. The second stage, therefore, operates as an amplifier on 80 metres, as either an amplifier or a doubler on 40 metres (depending on the crystal used), as a doubler on 20 and 10 metres, and as a tripler on 15 metres.

Inductor L1, the tuned grid circuit inductor for the 7054 on 80 metres, consists of 12 turns of 18 gauge tinned copper wire, approximately $1\frac{1}{2}$ inches long, air-spaced; the coil can be stiffened by cementing strips of perspex or similar material along the outside.

L2, the grid-circuit inductor for the 6883, is a 57-turn air-spaced coil, 2 inches long and 1 inch in diameter. Construction may follow the suggestion put forward for L1. The inductor should be tapped at $5\frac{1}{2}$, $8\frac{1}{4}$, $11\frac{1}{2}$ and $26\frac{1}{2}$ turns from the grid end. L2 is tuned by means of C15.

In the 80-metre position of the bandswitch, the total capacitance across L2 is increased by the addition of C13. The 6883 is neutralized by a bridge circuit consisting of C14, RFC4, and C16 to assure good stability of the 6883 on all bands.

The plate tank for the 6883 is a conventional pi-network type using two tapped coils. L4, the coil for the 20-, 15-, and 10 metre bands, is a 10-turn winding of No. 10 enamelled copper wire, having an inside diameter of 1 inch and an overall length of $1\frac{3}{4}$ inches. L5, which is in series with L4 for the 80- and 40-metre bands, is an 18-turn air-spaced coil approximately 2 inches long, 1 inch in diameter, constructed as previously described. L4 should be tapped at $5\frac{1}{2}$ and $8\frac{1}{2}$ turns from the plate end. L5 should be tapped at 8 turns from the L4 end.

L3 is a parasitic-suppressor choke consisting of 6 turns of plastic-insulated hookup wire, about $\frac{1}{4}$ inch in diameter, and is installed directly between the plate-cap connection of the 6883 and the plate end of RFC5.

The adjustable loading-capacitance at the output end of the pi network consists of a 140-$\mu\mu$f variable capacitor (C26), and a group of fixed capacitors controlled by a band-switch and the "Coarse Loading" switch S3. A 500-$\mu\mu$f capacitor (C25) is connected in parallel with C26 in the 80-metre position of the bandswitch, and the nine 150-$\mu\mu$f capacitors (C27 through C35) are successively added in parallel with C26 when S3 is rotated counterclockwise.

The modest and noncritical drive requirements of the 6883 permitted the use of a simple step-type drive control ganged with the bandswitch. As shown in Fig. 7, section S1b, of the bandswitch is connected to taps on a resistive voltage-divider network across the 450-volt supply circuit, and automatically adjusts the grid-No. 2 voltage of the 7054 buffer/frequency multiplier so as to provide the proper drive for the 6883 on each band.

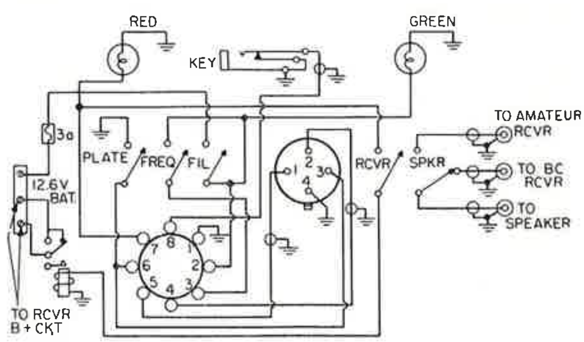

Fig. 3 — Transmitter Control Unit.

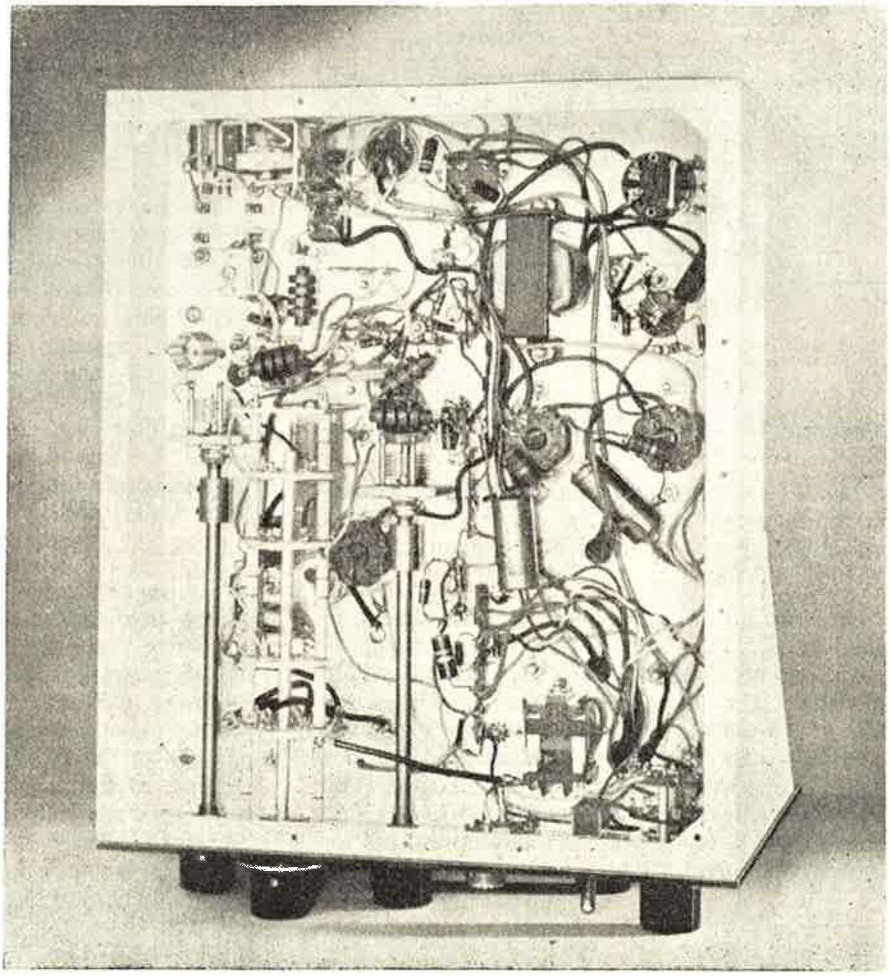

Fig. 4 — Underside View of the Mobile Transmitter.

The voltage-divider network shown at the input to the modulator circuit in Fig. 7 was designed for use with a transistorized microphone described elsewhere in this issue. Note however that for this application the three-wire cable mentioned in the description of the microphone should be changed to a four-wire cable. This change allows the ground connection for the audio input to be made right at the 7058 socket, thereby eliminating any possibility of ground-current pick-up. Alternative input connections for use with a carbon microphone are shown in the inset to Fig. 7.

To minimize the drain on the 450-volt supply under no-signal conditions, the 7027-A's are operated with somewhat higher bias than that required for true class AB1 operation. Although this method of operation might cause severe distortion of a steady-tone modulating signal, it has relatively little effect on the quality of speech modulation because of the very low average power of speech signals.

Changeover from phone to CW operation is accomplished by means of the "PHONE-TUNE-CW" switch (S6). In its "TUNE" position, this switch removes grid-No. 2 voltage from the 6883 and plate and grid-No. 2 voltage from the 7027-A's, so that the oscillator and buffer/multiplier stages can be tuned without danger of damage to the final amplifier.

The meter and associated switch (S2) are used to measure: the 7054 grid-No. 1 and plate current. the 6883 grid-No. 1, grid-No. 2, and plate currents; and the combined plate and screen currents of the 7027-A's.

Switch S5 is mounted on the modulator gain-control potentiometer (R29), and may be used to remove heater voltage from the modulator tubes when long periods of CW operation are contemplated.

Mechanical Features

The transmitter was built on a 10-inch by 12-inch by 3-inch aluminium chassis and a 10-inch by 10-inch panel. The valves and output network of the rf section are enclosed in a 5-inch by 7-inch by 9-inch aluminium utility box.

The bandswitch (S1), shown in Fig. 5, was assembled from steatite wafers and spacers to permit each section to be located as near as possible to the associated stage or components. L2, the grid coil for the final stage, is mounted on a small standoff insulator on the bandswitch support bracket.

The crystal-selector switch and relay permit change of the operating frequencies directly at the operating position. If the crystals are selected so that the resulting output frequencies are separated by not more than about 0.05%, it will not be necessary to readjust the transmitter when shifting from one crystal to the other.

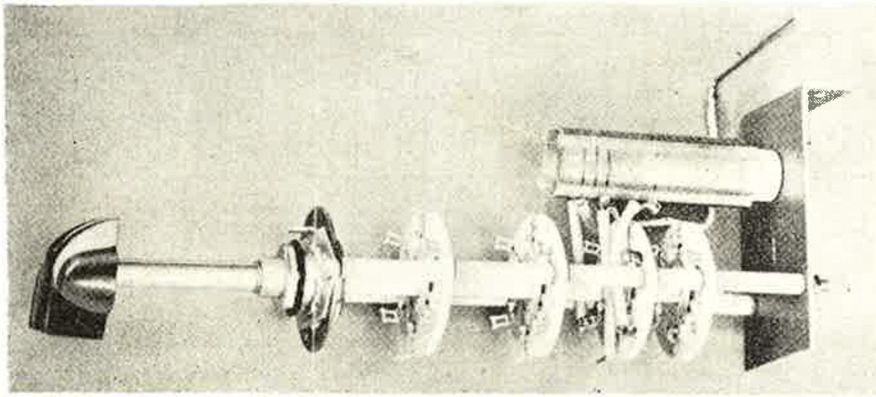

Fig. 5 — View of the Bandswitch S1 Showing Construction and Mounting of the 6883 Grid Coil.

The "DYNAMOTOR ON-OFF" switch on the transmitter panel and a local microphone connector (J1) permit the transmitter to be operated directly at its location in a car boot or elsewhere.

Modifications

If the transmitter is to be operated from a plate supply delivering less than 450 volts, it will be necessary to change the values of the series resistor in the plate-supply circuit for the oscillator and buffer/doubler stages (R22), the grid-No. 2 resistor for the 6883 (R12), and the cathode resistors for the 7027-A's (R32 and R33). The valves of these components for plate supply voltages of less than 450 volts are as follows:—

R12:	400 volts,	22,000 ohms.
	350 volts,	18,000 ohms.
	300 volts,	12,000 ohms.
R22:	400 volts,	3,300 ohms.
	350 volts,	1,200 ohms.
	300 volts,	not required.
R32:	400 volts,	1,200 ohms.
	350 volts,	1,000 ohms.
	300 volts,	860 ohms.
R33:	400 volts,	1,200 ohms.
	350 volts,	1,000 ohms.
	300 volts,	860 ohms.

This transmitter has been in use for about one year and has produced very rewarding signal reports as well as some excellent DX. The reports indicate that the quality of the phone signals provided by the transistorized microphone is greatly superior to that of most mobile transmitters using conventional carbon microphones.

PARTS LIST

C1—22 μμf, mica.
C2—80-400 μμf, compression mica.
C3, C4, C8, C9, C10, C11, C17, C18, C19, C20, C21—.001 μf, disc ceramic, 600 vw.
C5, C7, C12—.002 μf, mica.
C6, C15—50 μμf, variable.
C13—47 μμf, NPO Ceramicon or equiv.
C14—.001 μf (Feed-Thru Ceramicon or equiv.)
C16—3.5-12 μμf, tubular trimmer.
C22—.002 μf, mica, 1500 vw.
C23—.001 μf, 1500 vw, disc ceramic.
C24—325 μμf, variable.
C25—500 μμf, mica.
C26—140 μμf, variable.
C27, C28, C29, C30, C31, C32, C33, C34, C35—150 μμf, mica.
C36—.01 μf, 400 vw, paper.
C37—10 μf, 25 vw, electrolytic.
C38—.005 μf, 400 vw, paper.
C39, C40—20 μf, 450 vw, dual electrolytic.
C41, C42—50 μμf, 50 vw, electrolytic.
J1—4-point microphone connector.
J2—Coaxial antenna connector.

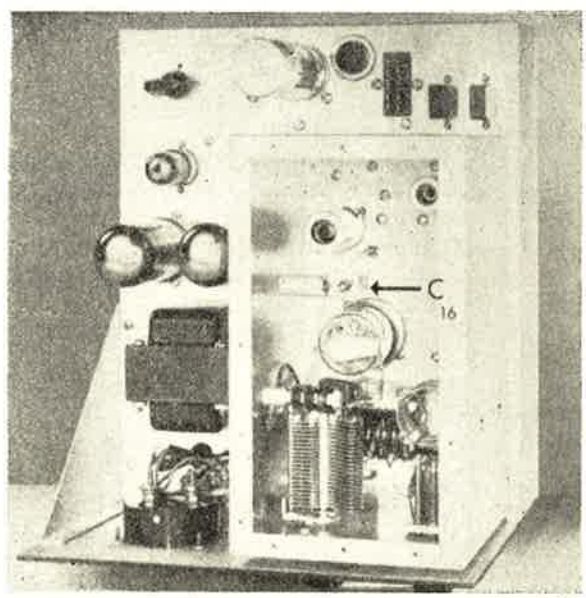

Fig. 6 — Top View of W2YM'S Mobile Transmitter. Note the Neutralizing Capacitor C16 Mounted on a Small Bracket and Stand-off Insulator Between the 6883 and 7054.

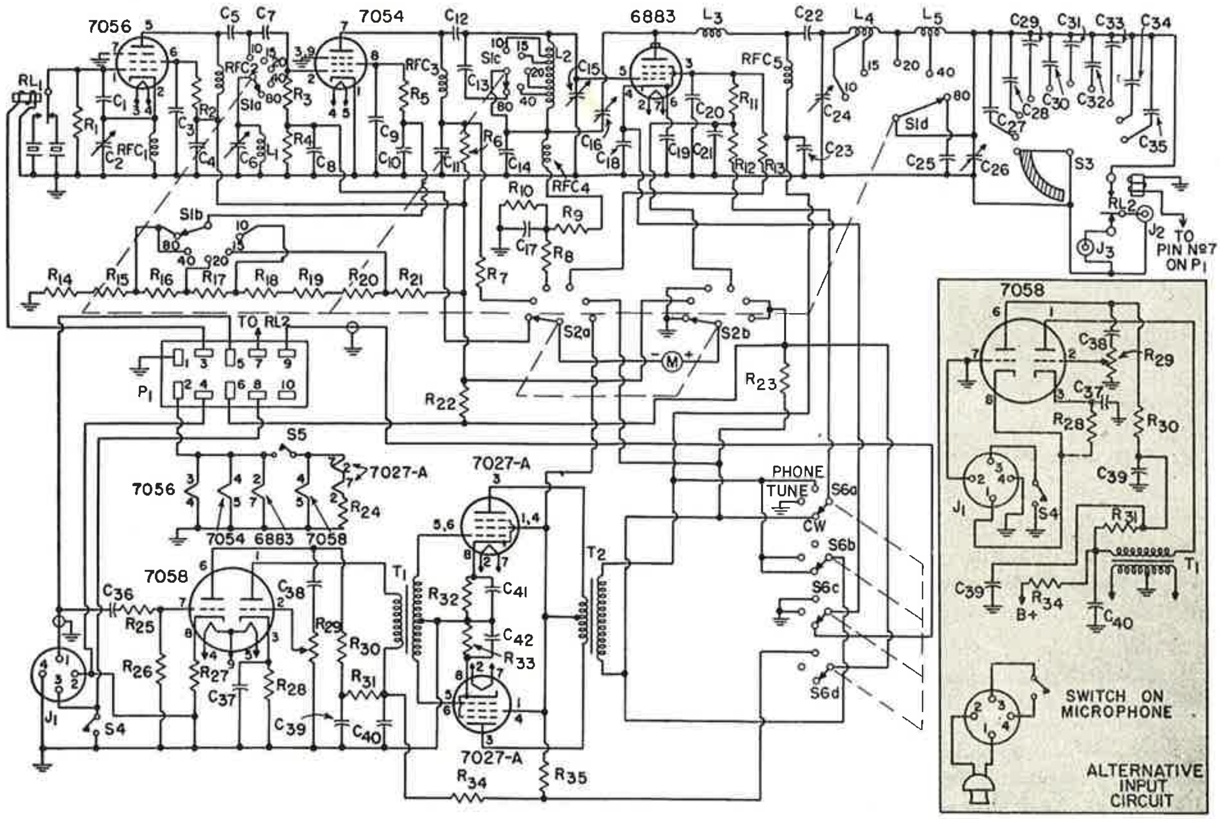

Fig. 7 — Schematic Diagram of the Five-band Mobile Transmitter. The Alternative Suggested Speech Amplifier Circuit for use with Carbon Microphone is Shown Inset.

J3—Coaxial receiver antenna connector.
L1, L2, L3, L4, L5 — See text.
M—Meter, 0-3 ma, 2".
P1—10-point "Jones" connector or equiv.
R1—100,000 ohms, ½ watt.
R2—33,000 ohms, ½ watt.
R3—68,000 ohms, ½ watt.
R4, R5, R6, R7, R8, R10, R13, R27, R28, 1,000 ohms, ½ watt.
R9—27,000 ohms, 1 watt.
R11—100 ohms, 1 watt (made by connecting two 220-ohm, ½-watt resistors in parallel).
R12—24,000 ohms, 2 watts.
R14, R15, R16, R17, R18, R19, R20—5,100 ohms, ½ watt.
R21—12,000 ohms, 1 watt.
R22—4,700 ohms, 1 watt.
R23, R35—10 ohms, ½ watt.
R24—1 ohm, ½ watt.
R25, R26, R30—47,000 ohms, ½ watt.
R29—½ megohm, ½ watt, volume control with switch (S5).
R31, R34—3,900 ohms, ½ watt.
R32, R33—1,500 ohms, 2 watts.
RFC1, RFC2, RFC4—2.5 mh, rf choke.
RFC3—1.0 mh, rf choke.
RFC5—1.0 mh, rf choke.
RL1, RL2—12-volt dc relays, SPDT.
S1—4 pole, 5 position.
S2—2 pole, 6 position.
S3—Single pole, 10 position, progressively opening.
S4—SPST, toggle.
S5—SPST (See R29).
S6—4 pole, 3 position.
T1—3:1 single plate to push-pull grids.
T2—10,000 ohms P to P to RF load.

(With acknowledgements to RCA)

EEV AT THE 1960 PHYSICAL SOCIETY EXHIBITION

The English Electric Valve Company Limited exhibited a selection of their latest developments, and tentative data are given in the following summary.

High Power Klystron

K352 is a three cavity magnetically focused klystron amplifier for high power pulsed operation at 3000 Mc. It is capable of delivering a peak power output of 6 Mw, at a mean level of 9 Kw, with a gain of 30 db and an efficiency in excess of 35%. The cavities are an integral part of the vacuum envelope and can be adjusted over a limited frequency range.

Magnetron

M565 is a water cooled, multi-resonator, pulse operated magnetron for use in the frequency range 1215 to 1365 Mc. It gives a peak output power of 5 Mw, at a mean output power of 15 Kw, with pulses of 15 μsec duration and a duty cycle of 0.003. The valve is designed for use with a separate electro-magnet and a launching section to couple it to waveguide No. 6 (6.5 inches x 3.25 inches internal dimensions).

Vacuum Variable Capacitors

An entirely new range of vacuum variable capacitors has been introduced and four types were exhibited. These are U30/15 (5 to 30μμf), U50/15 (8 to 50μμf), U80/15 (10 to 80μμf) and U200/10 (15 to 200μμf).

The design is such that the present range can be readily extended to meet future requirements for additional capacitance values. A special feature of these capacitors is their small physical size relative to their high voltage rating.

Cold Cathode Trigger Tubes

Three cold cathode trigger tubes types 5823, QT1250 and AT1251, have been developed for relay applications in low current circuits. QT1250 and QT1251 have identical electrical characteristics and are variants of the 5823, with closer tolerances on trigger breakdown and anode maintaining voltages. 5823 and QT1250 have B7G bases whereas QT1251 is fitted with flying leads.

Parametric Amplifier

N1036 is a 200 Mc transverse field parametric amplifier characterised by its extremely low noise factor. It will give a gain of more than 20 db with a bandwidth of about 25 Mc; the broadband noise figure is about 2 db. A magnetic focusing field of 70 gausses is required.

The tube operates at a very low beam current and voltage, 30 microamperes at 6 volts being typical. The cathode current is 5 milliamperes at the highest dc voltage in the tube is 100 volts. The pump power requirements are only a few milliwatts at about 400 Mc.

High Voltage Inverse Diode

AX295 is a hydrogen filled inverse diode for use primarily in radar modulators. It may also be used as a charging diode or a conventional power rectifier with a general rating similar to the AX228. It has a low forward impedance and a peak inverse voltage rating of about 15 Kv.

Industrial Heating Triode

BR1143 is a forced air cooled triode developed primarily for industrial rf heating equipments. This high perveance valve is designed to operate at low anode voltages (6 Kv is a typical value) and will deliver a power output of 50 Kw. The thoriated tungsten filament is rated at 11.0 volts, 240 amperes and the valve has a perveance of 8ma/v$^{3/2}$ with an amplification factor of 34.

Glow Modulator Tube

A glow modulator tube, type 1B59, has been added to the range of E.E.V. cold cathode devices. The hollow cathode contained in this tube provides a high ionisation density and forms a compact light source with an output of 0.13 candle power at 30 ma cathode current. A particular feature is the substantially linear relationship between the light output and the cathode current. This, together with the high frequency and directional characteristics of the tube, makes the 1B59 suitable for numerous industrial and communications applications including facsimile equipments, stroboscopes and photoelectric counters.

NEW RELEASES

1B59

The new EEV cold cathode glow modulator tube, type 1B59 employs a hollow cathode, providing a high ionisation density and forming a compact light source with an output of 0.13 candle power at a cathode current of 30 ma. The 1B59 has a maximum breakdown voltage of 250 volts and a maximum operating voltage of 150 volts at 30 ma dc. The average operating current ranges from 5 to 35 ma with a maximum peak operating current of 75 ma. Luminance is 42 candles per square inch with a red discharge colour. The substantially linear relationship between the light output and cathode current, together with the high frequency and directional characteristics make the 1B59 suitable for numerous industrial and communication applications.

2N1425, 2N1426

The 2N1425 and 2N1426 are new drift transsistors for use as if amplifiers and converters respectively in AM battery receivers. Features are exceptional stability, excellent uniformity of characteristics, good sensitivity and low feedback capacitance. The 2N1425 is capable of providing a useful power gain of 30.4 db in a common-emitter circuit without a neutralizing network. Where maximum power gain is desired with neutralization it can provide a useful power gain of 34 db. The 2N1426 provides a useful conversion power gain of 37.6 db in a common-emitter circuit with a dc collector-to-emitter voltage of — 11 volts and an emitter current of 0.65 ma.

6GH8

The 6GH8 is a medium-mu triode, sharp-cutoff pentode of the 9 pin miniature type. Intended primarily for use in multivibrator-type horizontal deflection oscillators, the 6GH8 has high peak cathode current ratings, high transconductance at low plate current for both units, and low interelectrode leakage.

2039

The ceramic-metal super-power valve 2039 is a water-cooled, shielded-grid beam triode of unique design intended for use as a plate-pulsed rf power amplifier. It is especially suited for use in long-range search-radar and in particle-accelerator service. In such applications with duty factor of 0.06 and pulse duration of 2500 microseconds, the 2039 can provide a useful peak power output of 1.5 megawatts at 200 Mc.

7539

The 7539 Scan-Conversion Tube provides a practical answer to the long-standing problem of large-screen radar display in brightly lighted rooms. The 7539 is designed to transform signal information continuously from one time base to another. For example, PPI information generated by a conventional radar system can be processed by this tube for display on a high-resolution, large-screen TV monitor or for transmission via TV circuits to remote locations.

Resolution capacity of the 7539 is 150 range rings per display radius with a response of 50% or better. To utilize fully the resolution capability of the 7539, the TV monitor system must be designed for resolution in excess of 1,000 TV lines. The 7539 employs two precisely aligned coaxial electron guns, one on each side of the central "target" section. One of the guns is used for writing; the other is used for reading.

7580

The 7580 is a new forced-air-cooled beam power valve for use linear amplifier service at frequencies up to 500 Mc. Very small and compact, it is constructed with ceramic-metal seals throughout. Its maximum plate dissipation is 250 watts. In single-sideband suppressed-carrier service with two-tone modulation, the 7580 can deliver a useful peak-envelope-power output of about 400 watts at 30 Mc or 360 watts at 500 Mc.

The ceramic-metal-seal construction employed in the 7580 permits operation at higher temperatures than a glass-seal construction and thus provides improved reliability. The specially designed, high-efficiency, plated radiator makes possible the maximum plate-dissipation rating of 250 watts with no sacrifice in tube reliability.

E704

The EEV E704 is of the Nesotron or "Island Storage" type with electrical input and output.

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SUPER-RADIOTRON 23MP4 PICTURE TUBE

The Super-Radiotron 23MP4 is an aluminized 23" 114° glass rectangular, directly viewed picture tube. The 23MP4 features a relatively flat, compound radius face plate and special internal contouring in the deflection yoke region to obtain 114° deflection with 110° deflection components. Featured also is the 19-1/4" x 15-3/16" screen with very slightly curved sides and relatively square corners having a minimum projected area of 280 square inches.

The 23MP4 which is of the low voltage electrostatic focus and magnetic deflection type, employs a small diameter (1-1/8") neck and a precision gun, specially designed to minimize deflection distortion, that does not require the use of an ion-trap magnet.

GENERAL

Heater Voltage .... 6.3 volts
Heater Current .... 0.6 amp

Direct Interelectrode Capacitances:
- Grid No. 1 to all other electrodes .... 6 $\mu\mu$f
- Cathode to all other electrodes .... 5 $\mu\mu$f

External conductive coating to
- ultor .... 2,500 max. $\mu\mu$f / 1,700 min. $\mu\mu$f
- Faceplate, Compound Radius .... Filterglass
- Light transmission (approx.) 76%

Phosphor .... P4 Sulphide Type
- Fluorescence .... White
- Phosphorescence .... White
- Persistance .... Short

Focusing Method .... Electrostatic
Deflection Method .... Magnetic

Deflection Angles (approx.):
- Diagonal .... 114°
- Horizontal .... 102°
- Vertical .... 84°

Electron Gun:
- Requires No Ion-Trap Magnet

Tube Dimensions:
- Overall Length .... 14-3/8" ± 1/4"
- Greatest Width .... 23-25/64"
- Greatest Height .... 20-1/2"
- Diagonal .... 16-1/2"
- Neck Length .... 5-1/8" ± 1/8"

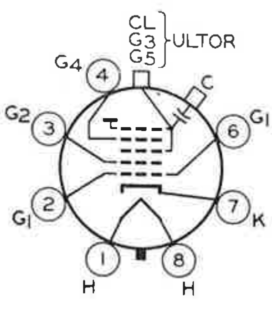
Screen Dimensions (minimum):
- Greatest width .... 19-1/4"
- Greatest Height .... 15-3/16"
- Diagonal .... 22-1/4"
- Projected area .... 280 sq. in.

Cap .... Recessed small cavity (JETEC No. J1-21)
Bulb .... J 187
Base .... Small Button Eightar 7-pin (JETEC No. B7-208)

Weight (approx.) 23-1/2lb
Mounting Position .... Any

SOCKET CONNECTIONS

(bottom view)



Pin 1 — Heater
Pin 2 — Grid No. 1
Pin 3 — Grid No. 2
Pin 4 — Grid No. 4
Pin 6 — Grid No. 1
Pin 7 — Cathode
Pin 8 — Heater
Cap — Ultor (Grid No. 3, Grid No. 5, Collector)
C — External Conductive Coating.

GRID-DRIVE SERVICE

Grid drive is the operating condition in which the video signal varies the Grid-No. 1 potential with respect to cathode.

(Unless otherwise specified, voltage values are positive with respect to cathode.)

MAXIMUM RATINGS, Design-Centre Values:

ULTOR† VOLTAGE	18,000 volts
GRID No. 4 VOLTAGE:	
Positive value	1,000 volts
Negative value	500 volts
GRID No. 2 VOLTAGE	500 volts
GRID No. 1 VOLTAGE	
Negative peak value	200 volts
Negative bias value	140 volts
Positive bias value	0 volts
Positive peak value	2 volts
PEAK HEATER-CATHODE VOLTAGE:	
Heater negative with respect to cathode	180 volts
Heater positive with respect to cathode	180 volts

EQUIPMENT DESIGN RANGES:

With any Ultor Voltage (E_{c5k}) between 12,000** and 18,000 volts and Grid No. 2 Voltage (E_{c2k}) between 200 and 500 volts

Grid No. 4 Voltage for Focus*	0 to 400	volts
Grid No. 1 Voltage for Visual Extinction of Focused Raster	-10% to -25% of E_{c2k}	volts
Grid No. 1 Video Drive from Raster Cutoff (Black Level):		
White Level Drive (Peak Positive)	10% to 25% of E_{c2k}	volts
Grid No. 4 Current	-25 to $+25$	μamp
Grid No. 2 Current	-15 to $+15$	μamp
Field Strength of Adjustable Centring Magnet†‡	0 to 8	gausses

EXAMPLES OF USE OF DESIGN RANGES:

With Ultor Voltage of	18000	volts
And Grid No. 2 Voltage of	400	volts
Grid No. 4 Voltage for Focus	0 to 400	volts
Grid No. 1 Voltage for Visual Extinction of Focused Raster	-44 to -94	volts
Grid No. 1 Video Drive from Raster Cutoff (Black Level):		
White-Level Drive (Peak Positive)	44 to 94	volts

MAXIMUM CIRCUIT VALUE:

Grid No. 1 Circuit Resistance	1.5 megohms

CATHODE-DRIVE-SERVICE

Cathode drive is the operating condition in which the video signal varies the cathode potential with respect to Grid No. 1 and the other electrodes.

(Unless otherwise specified, voltage values are positive with respect to Grid No. 1)

MAXIMUM RATINGS, Design-Centre Values:

ULTOR TO GRID No. 1 VOLTAGE	18,000 volts
GRID No. 4 TO GRID No. 1 VOLTAGE:	
Positive value	1,000 volts
Negative value	500 volts
GRID No. 2 TO GRID No. 1 VOLTAGE	640 volts
GRID No. 2 TO CATHODE VOLTAGE	500 volts
CATHODE TO GRID No. 1 VOLTAGE:	
Positive peak value	200 volts
Positive bias value	140 volts
Negative bias value	0 volts
Negative peak value	2 volts

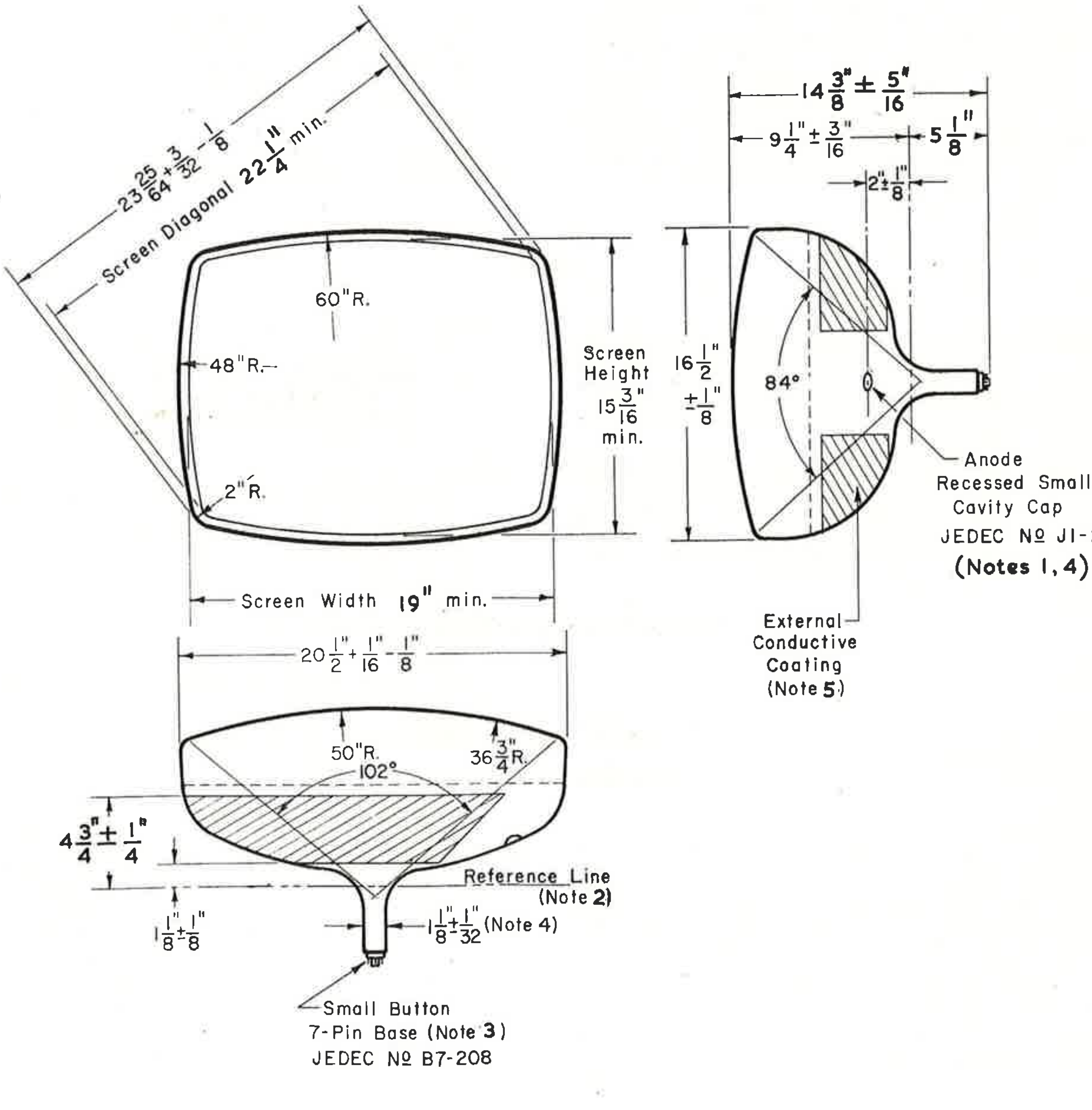
PEAK HEATER-CATHODE VOLTAGE:

Heater negative with respect to cathode	180 volts
Heater positive with respect to cathode	180 volts

EQUIPMENT DESIGN RANGES:

With any Ultor to Grid No. 1 Voltage (E_{c5g1}) between 12,000** and 18,000 volts and Grid No. 2 to Grid No. 1 Voltage (E_{c2g1}) between 225 and 640 volts

Grid No. 4 to Grid No. 1 Voltage for Focus*	0 to 400	volts
Cathode to Grid No. 1 Voltage for Visual Extinction of Focused Raster	10% to 21.5% of E_{c2g1}	volts
Cathode to Grid No. 1 Video Drive from Raster Cutoff (Black Level): White-Level Value (Peak Negative)	− 10% to − 21.5% of E_{c2g1}	volts
Grid No. 4 Current	− 25 to + 25	μamp
Grid No. 2 Current	− 15 to + 15	μamp
Field Strength of Adjustable Centring Magnet†‡	0 to 8	gausses



EXAMPLES OF USE OF DESIGN RANGES:

With Ultor to Grid No. 1 Voltage of	18000 volts
And Grid No. 2 to Grid No. 1 Voltage of	400 volts
Grid No. 4 to Grid No. 1 Voltage for Focus	0 to 400 volts
Cathode to Grid No. 1 Voltage for Visual Extinction of Focused Raster	42 to 78 volts
Cathode to Grid No. 1 Video Drive from Raster Cutoff (Black Level): White-Level Value (Peak Negative)	– 42 to – 78 volts

MAXIMUM CIRCUIT VALUE:

Grid No. 1 Circuit Resistance	1.5 megohms

† The ultor in a cathode ray tube is the electrode to which is applied the highest dc voltage for accelerating the electrons in the beam prior to its deflection In the 23MP4 the ultor function is performed by grid No. 5. Since grid No. 5, grid No. 3 and the collector are connected together within the 23MP4 they are collectively referred simply as ultor for convenience in presenting data and curves.

** This value is a working design-centre minimum. The equivalent absolute minimum ultor, or ultor to grid No. 1 voltage is 11,000 volts, below which the serviceability of the 23MP4 will be impaired. The equipment designer has the responsibility of determining a minimum design value such that under the worst probable operating conditions involving supply-voltage variation and equipment variation the absolute minimum ultor, or ultor to grid No. 1 voltage is never less than 11,000 volts.

* The grid No. 4 (or grid No. 4 to grid No. 1) voltage required for optimum focus of any individual tube will be a value between 0 to 400 volts independent of ultor current; and will remain essentially constant for values of ultor (or ultor to grid No. 1) voltage and grid No. 2 (or grid No. 2 to grid No. 1) voltage within design ranges shown for these items.

†‡ The maximum separation between a suitable PM centering magnet and the reference line is 2¼ inches. Excluding the effects of extraneous magnetic fields, the centre of the undeflected focused spot will fall within a circle having a ⅜ inch radius whose centre coincides with the geometric centre of the tube face. The earth's magnetic field can deflect the spot off the geometric centre by approximately ½ inch.

NOTES

NOTE 1: The plane through the tube axis and pin No. 4 may vary from the plane through the tube axis and ultor terminal by angular tolerance (measured about the tube axis) of 30°. Ultor terminal is on same side as pin No. 4.

NOTE 2: With the tube neck inserted through flared end of reference-line gauge JEDEC No. 126 and with tube seated in gauge, the reference line is determined by the intersection of the plane CC' of the gauge with the glass funnel.

NOTE 3: Socket for this base should not be rigidly mounted; it should have flexible leads and be allowed to move freely. Bottom circumference of base shell will fall within a circle concentric with bulb axis and having a diameter of 1¾ inches.

NOTE 4: To clean insulating coating around cavity contact, use only a soft dry lint-free cloth.

NOTE 5: External conductive coating must be grounded.

NEW RELEASES

(Continued from page 86)

The E704 has integrating properties and can be used to extend the range of radar systems by giving a marked improvement in signal to noise ratio. The tube employs two separate electron guns for writing and reading information on the target, both beams being electrostatically focused and electromagnetically deflected to scan the target in one direction along its length. Either gun has a heater voltage of 6.3 volts and a heater current of 600 ma. The target itself consists of an array of about 2,000 gold strips deposited on a thin glass plate which is backed by a continuous metallic film forming the signal output electrode. Cathode of the writing gun is – 7.5 Kv and – 1.2 Kv for the reading gun, both with respect to the collector which has a voltage of zero. Grid No. 1 cutoff voltage for either gun is – 35 volts, grid No. 2 voltage (either gun) 150 volts and grid No. 3 voltage being 25% of the respective cathode voltage of either gun. Cathode current, 0.5 ma and peak output current 5 μa.

A TRANSISTORIZED MICROPHONE

by G. D. Hanchett, W2YM

During the design of a mobile rig the author was recently confronted with the problem of finding a suitable microphone. A carbon microphone, although high in output, is noisy and of relatively uneven frequency response. A crystal microphone has good frequency response but is low in output and has the additional disadvantage — for mobile use especially — of temperature limitations.

An attempt was made, therefore, to construct a microphone that would have good audio quality, be fairly high in signal output, be rugged enough for mobile ham use, be insensitive to unwanted electrical pickup, and still be within the price range of the average ham. This article describes the result; a surprisingly simple build-it-yourself microphone that meets all the given requirements.

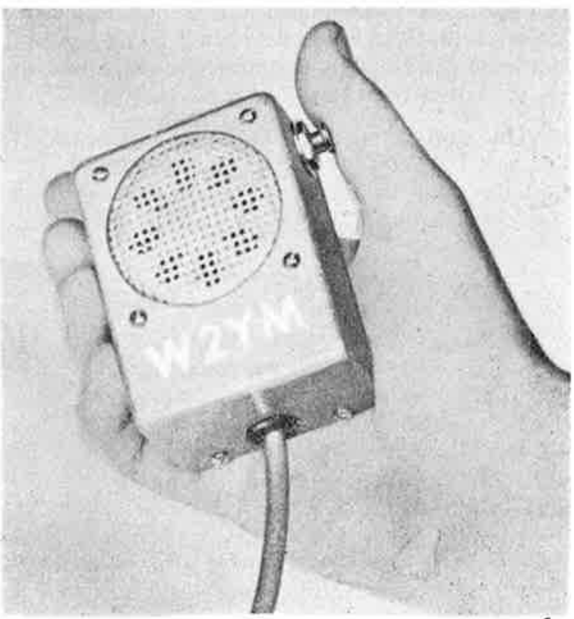


Fig. 1 — A view of W2YM's Microphone.

Because of the abovementioned limitations of carbon and crystal microphones, they were discarded in favour of a dynamic "mike". The active element chosen for the microphone is a 2-inch miniature PM speaker. The speaker works into a one-stage, transistorized amplifier. Both speaker and amplifier are contained in a small metal box that fits the hand comfortably.

The output of the microphone's built-in amplifier ranges between 0.75 volt and 1.0 volt measured across a load of 20,000 ohms or more. The audio quality of this microphone surpassed all expectations; in fact, for voice use it compared favourably with a so-called broadcast-quality crystal microphone.

Fig. 2 is a schematic diagram of the microphone A 2N217 p-n-p junction transistor is connected in a common-emitter, base-input amplifier circuit. Degeneration, provided through R2, stabilizes the transistor against the effects of temperature variations. Push-button switch S1 serves two functions: one pole (S1A) energizes the transistor amplifier; the other pole (S1B) can be used to control the transmitter.

Construction

The microphone is assembled in an aluminium box 3" long, 2¼" wide and 1½" high. At W2YM the box was made by folding a sheet of aluminium cut to the proper pattern, welding the corners, and applying a coat of paint to the outside. If no welding equipment is available, the aluminium may be bent to form lips that may be bolted together. The box may also be formed from brass or copper, in which case the edges can be soldered together.

A hole of appropriate size is cut in the front of the box to accommodate the speaker. A piece of perforated aluminium is placed over the hole, inside the box, to serve as a portective screen for the speaker.

The transistor and the associated small components for the amplifier were mounted on a strip of linenized bakelite. Any good insulating material, however, can be used for the mounting board. Even cardboard should be suitable, provided humid conditions are not present.

The leads of the components are passed through small holes drilled through the mounting board. Where the leads come through the holes at the back of the mounting board they are bent into small hooks with a pair of long-nosed pliers. These hooks in the leads hold the components to the mounting board and make for easy connection to the leads.

Caution: be sure to observe polarity when connecting the capacitors.

The mounting board is held in place inside the box by two of the screws that fasten the speaker. These two screws should be long enough so that the mounting board can be held above the speaker by two $\frac{5}{8}$"-long, $\frac{1}{4}$"-wide spacers.

Arrangements for mounting the battery will depend on the type chosen. It might be wise to stick a small piece of insulating tape on the inside of the box near where the terminals of the battery will be located. This precaution will remove any chance of the battery accidentally shorting to the box.

The bass response of the microphone can be adjusted by damping the miniature speaker. To obtain suitable damping, cover all but one of the holes in the rear housing of the speaker with felt. The felt may be held in place with ordinary household cement. The last hole in the rear of the housing should be covered with a piece of fibre or cardboard that has a 1/32" hole drilled through for pressure release. Another pressure-release hole, this one $\frac{1}{8}$" diameter, is drilled in the back cover plate of the microphone case. With this

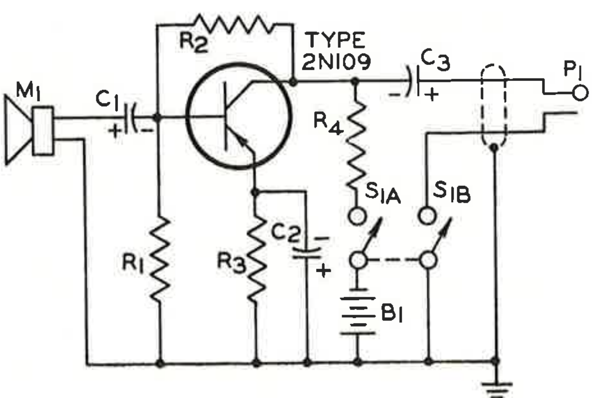


Fig. 2 — Schematic Diagram of the Transistorized Microphone.

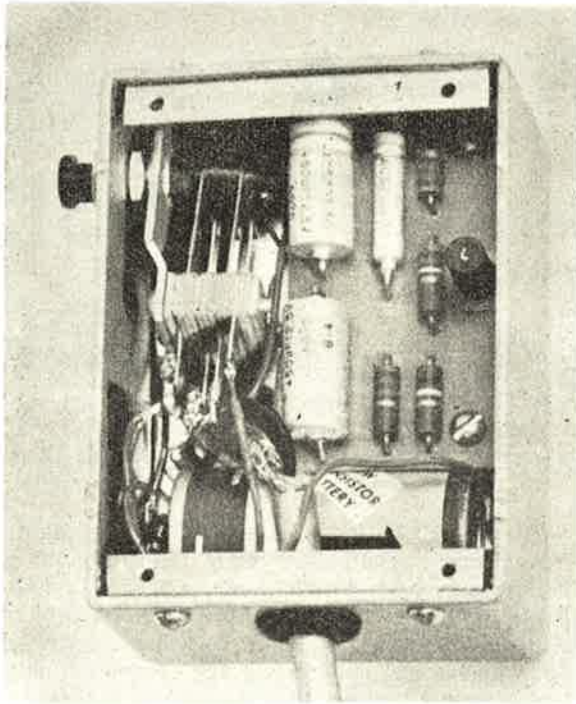


Fig. 3 — View of the Microphone with Rear Cover Removed.

construction the frequency response of the microphone will be smoothed out. Some experimenting may however be carried out to achieve the optimum results. Response, particularly at the low end, falls off rapidly beyond these limits — a desirable feature for a communications microphone.

Actual construction time for this microphone should be no more than a few hours. The author believes that the finished product is the best microphone for amateur communications now available. Try it and you'll agree.

The author wishes to thank J. Owens, W. Davies, F. Boryszewski, F. Wenzel, and J. Preston for their valuable aid or suggestions during the construction and testing of this microphone.

PARTS LIST

B1—Transistor Battery, 9-volt.
C1, C2—50 μf, 12 vw, electrolytic.
C3—2.5 μf, 25 vw, electrolytic.
M1—Microphone, 2" PM speaker.
P1—2-contact plug.
R1—10,000 ohms, $\frac{1}{2}$ watt.
R2—68,000 ohms, $\frac{1}{2}$ watt.
R3—1,200 ohms, $\frac{1}{2}$ watt.
R4—8,200 ohms, $\frac{1}{2}$ watt.
S1—Push-button switch, DPST, non-locking.
2N217 or 2N109 Transistor.

(With acknowledgements to RCA)

AWV