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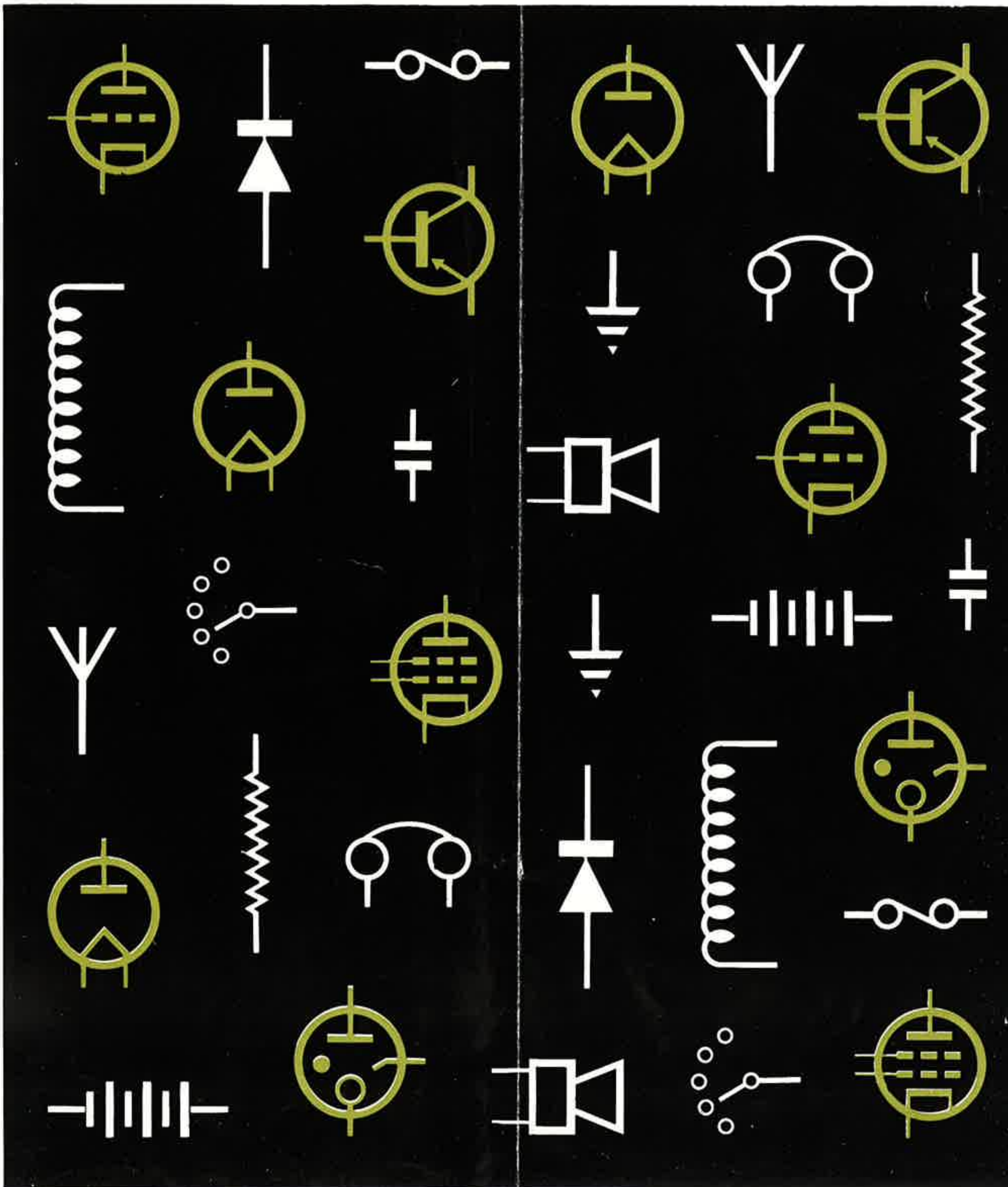

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Editor, Bernard J. Simpson

IN THIS ISSUE

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A TWO-TRANSISTOR REGENERATIVE RECIEVER

For 40 and 80 Metres

By E. M. WASHBURN, W2RG

The operating enjoyment on 80, 40, and 15 metres provided by the transistorized QSO-getter described in my article in the February, 1960 issue of "Radiotronics" prompted the construction of a companion transistor receiver.

It seemed desirable that the receiver be a superheterodyne, and literature was searched for a suitable circuit. Although several promising circuits were found, they all had the same shortcoming: they required more transistors and other components than would fit readily into a cabinet as small as the one used for the transmitter. The July, 1957, issue of "QST", however, contained a description of a transistorized regenerative "reflex" receiver built by W6WXU. It seemed to provide the answer. This receiver used only two transistors and two diodes, and was built in a case small enough to be held in the palm of one hand.

A breadboard receiver was constructed, using the basic circuit of W6WXU's receiver with modifications to suit the components on hand. Two 2N247 p-n-p drift field transistors were used, and the two diodes were type 1N34's. An air-spaced inductor was used as the antenna coil in place of the hand-wound ferrite-rod antenna used in the original. A small output transformer with primary and secondary connected in series was used instead of the 2-henry choke.

This breadboard model worked so well that it was rebuilt in a 7- by 5- by 3-inch metal box, as shown in the accompanying photographs. The receiver covers 3.3 to 8.5 Mc in three ranges, which can be adjusted by means of the bandsetting capacitor to provide continuous coverage or separate bandspread coverage for the upper and lower halves of the 80-metre band and for the 40-metre band.

Circuit Description

The circuit of the receiver is shown in Figure 2. The rf section includes the antenna coil L1; the padding, bandsetting, and tuning capacitors C2, C3, and C4, respectively; the first type 2N247 transistor; the two crystal diodes, and the re-

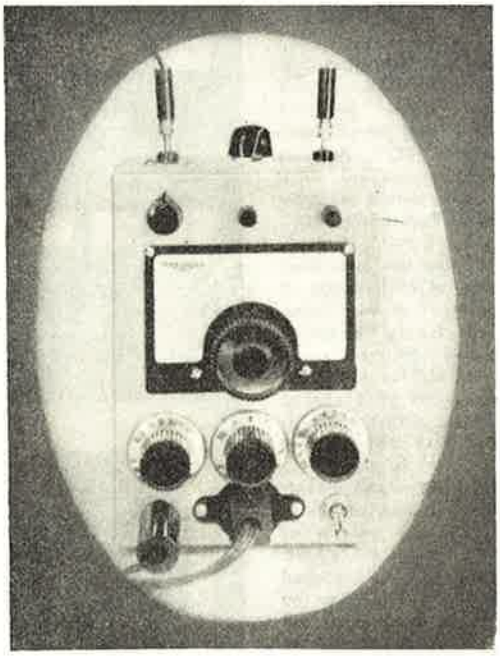

Fig. 1 — Front View of W2RG's Receiver. The two Transistors are Mounted Externally just Above the Bandspread Tuning Dial.

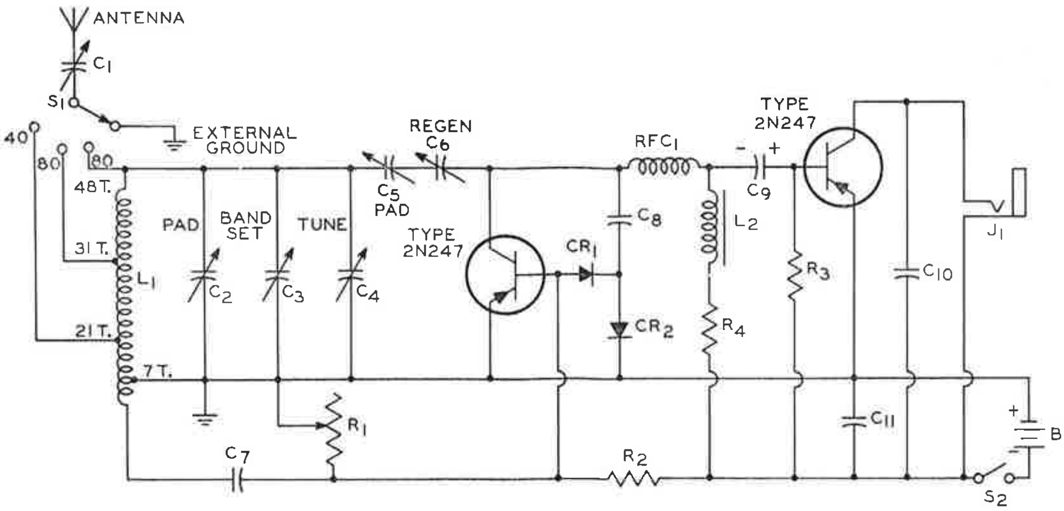

Fig. 2 — Schematic Diagram of W2RG's Two-Transistor 40- and 80-metre Receiver.

generation-control capacitors C5 in series with C6. The remaining components comprise the "reflex" and af-amplifier portions of the receiver. R1 controls the bias on both transistors. According to the description of W6WXU's receiver in "QST", the first transistor acts as a regenerative amplifier feeding the two series-connected crystal diodes. The af voltage appearing across the grounded diode is fed back to the first transistor (through the base), amplified, and reamplified in the second transistor. We have not questioned this very brief explanation because one thing is certain: the receiver works beautifully.

Construction Details

The arrangement of the major controls and components is shown in Fig. 1 in the front view of the completed receiver. From left to right across the top of the cabinet are the antenna jack, band-selector switch (S1), and ground-connection jack. Above the tuning dial are the antenna-trimmer capacitor C1, and the two 2N247 transistors. Just below the dial are the band-setting capacitor C3, the transistor-bias control R1, and the regeneration control C6, and across the bottom are the phone jack, power-supply connector, and ON-OFF switch S2.

The photographs of the inside of the receiver, Figs. 3 and 4, show the internal construction. Hand-capacitance effects in tuning are eliminated by the metal-skirted dial used for the band-spread capacitor, and by the use of flexible couplings and bakelite extension shafts for the antenna-trimmer, band-setting, and regeneration-control capacitors. To assure mechanical stability, the antenna-trimmer, band-setting, and regeneration-control capacitors are rigidly mounted on brackets and stand-off insulators, and smaller components are supported by heavy bus-bar wire.

The antenna coil L1 consists of 48 turns of 20 gauge tinned copper wire wound on an inside diameter of 1 inch, 3 inches long. Tappings should be provided at 7, 21 and 31 turns as shown in the circuit diagram. The coil should be stiffened by cementing strips of perspex along the outside, as shown in the photographs. Turns adjacent to the tap points and the ground connection are depressed to permit clean soldering of connections at these points without danger of shorted turns. The electrolytic capacitor, band switch, flexible couplings, and resistors shown in the photographs are larger than necessary, because I simply used whatever parts were at hand.

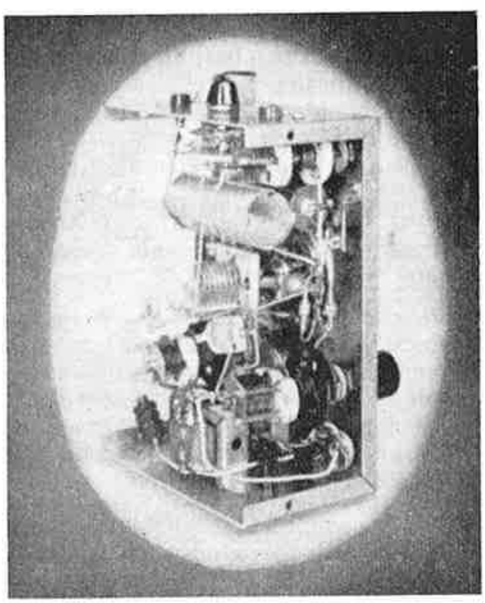

Fig. 3 — Inside View of the Receiver, Showing Placement of Major Components.

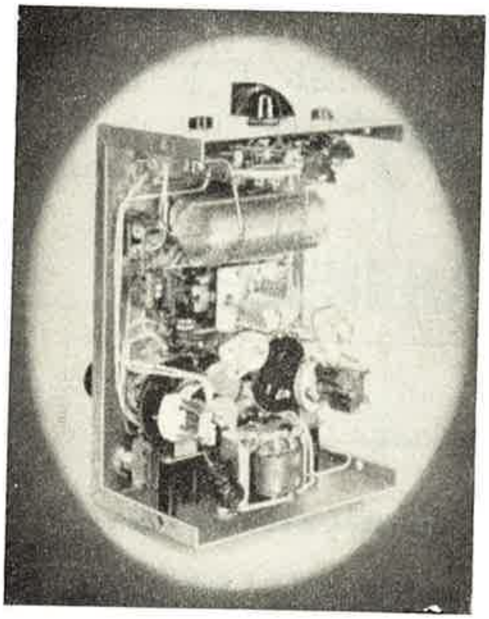

Fig. 4 — A Further Inside View of the Receiver, Showing Mounting of the Crystal Diodes.

Operation

The only power supply needed for the receiver is a 6-volt battery, which may be the same as that used for the transmitter. Although an antenna may not be necessary, a good external ground connection is essential to eliminate phone-cord body-capacitance effects. If an antenna is used, it should be only a very short length of wire—not the transmitting antenna—and should be very loosely coupled to the antenna coil.

The receiver performs best on CW signals. For most stable performance, the bias and regeneration controls should be advanced well beyond the threshold of oscillation. For very weak signals, it will be necessary to work closer to the oscillation point. For phone signals, as in any regenerative receiver, the regeneration control should be set just below the oscillation point. Weak phone signals, however, are difficult to copy.

High-resistance headphones should not be used with the receiver because they cause motorboating and make it difficult to maintain smooth control of oscillation. Phones having a dc resistance of about 135 ohms are satisfactory and provide plenty of volume on most ham signals, even with a 6-foot antenna.

To avoid burnout of the transistors, it is absolutely essential that the bandswitch be set in its "EXTERNAL GROUND" position before transmitting, and whenever the receiver is not being used. If even a moderately high-power transmitter is being used in the vicinity of the receiver, the receiving antenna should be disconnected.

When the receiver is first turned on, its tuning drifts slightly, and both the bias and regeneration controls may require readjustment. During transmitting periods, the receiver frequency tends to move upwards, so that when you turn it over to the other chap and throw the band-selector switch to either the 40- or 80-metre band, you have to retune slightly in the downward direction.

Nevertheless, the receiver has proved to be well worth the effort and expense that went into its construction. With its companion 300-milliwatt transistorized transmitter, it has provided many two-way QSO's over distances up to 1,000 miles, and many Europear ham stations on 40 metres have been copied during evening hours from the USA.

Modifications

If operation from a 12-volt battery is desired, the circuit should be modified as shown in Figure 5, and either a 2N408 or 2N109 transistor used in place of the 2N247 in the output stage of the receiver.

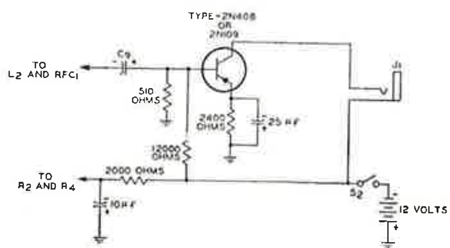

Fig. 5 — Modified Output Circuit for Operation of W2RG's Receiver from a 12-volt Supply.

PARTS LIST

B1 — Battery, 6 volt, heavy duty
C1 — 0-50 μμf, variable
C2 — 7-45 μμf, ceramic variable
C3 — 0-75 μμf, variable
C4 — 0-20 μμf, variable
C5 — 1.5-7 μμf, ceramic variable
C6 — 0-50 μμf, variable
C7 — 270 μμf, mica
C8 — 220 μμf, mica
C9 — 8 μf, electrolytic
C10 — 0.01 μf, ceramic disc
C11 — 0.01 μf, ceramic disc
CR1, CR2 — Crystal diode, type 1N34, 1N34A or equiv.
J1 — Phone jack, open circuit
L1 — See text
L2 — AF choke, 2 henrys
R1 — Bias control, 10,000 ohm potentiometer
R2 — 220,000 ohms, ¼ watt
R3 — 270,000 ohms, ¼ watt
R4 — 270 ohms, ½ watt
RFC1 — 2.5 millihenrys, rf choke
S1 — Wafer switch, single-pole, four position
S2 — On-off switch, SPST

SSB EXCITER CIRCUITS USING A NEW BEAM-DEFLECTION VALVE

Practical Circuit Data For Modulation,

Frequency Conversion and Detection

By. H. C. Vance, K2FF

The new beam deflection valve described here appears to be a bulbful of versatility, with more applications than were visualized when the valve was under development as an improved type of balanced modulator. This article tells how the 7360 can be put to work in a number of ways in the amateur field, particularly in SSB transmission and reception.

A new beam-deflection valve, the 7360, incorporates novel design features which, with suitable circuits, allow it to be used in many kinds of applications with improved performance at frequencies at least as high as 100 megacycles.

The valve was originally developed to provide a high degree of stable carrier suppression when used as a balanced modulator in single-sideband service. More than 60 db of carrier suppression has been obtained with it as a balanced modulator in SSB, exciters of both the filter and phasing types. It is, of course, equally valuable in double-sideband suppressed carrier service.

Frequency conversion, product detection, synchronous detection, single-ended to push-pull phase inversion switching circuits, fader circuits and compressor-expander-limiter circuits are among the many other interesting applications in which the unique properties of this new valve have been found to be valuable.

In this article circuits will be described which make use of the 7360 as a balanced modulator,

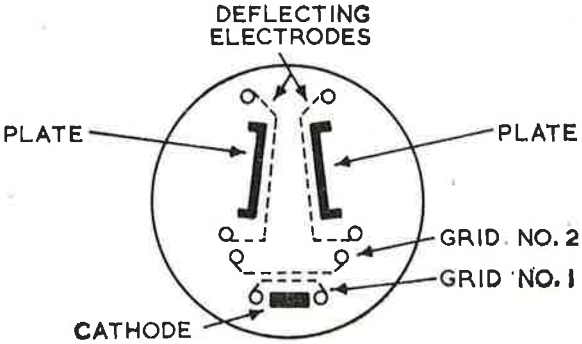

Fig. 1 — Arrangement of Electrodes in the Beam-Deflection Modulator 7360.

a frequency mixer and a product detector. Balanced modulator circuits will be shown for both the filter and phasing methods of SSB generation. The circuits are presented for the purpose of illustrating various methods of employing this new valve as a balanced modulator and as a frequency mixer rather than as a complete SSB transmitter design. The writer wishes to express his appreciation to M. B. Knigh and Ken Uhler of RCA for much of the information in this article.

First, let's examine Fig. 1, which is a cross-action sketch of the main elements of the valve. The single flat cathode, control grid and screen grid form an electron gun which generates, con-

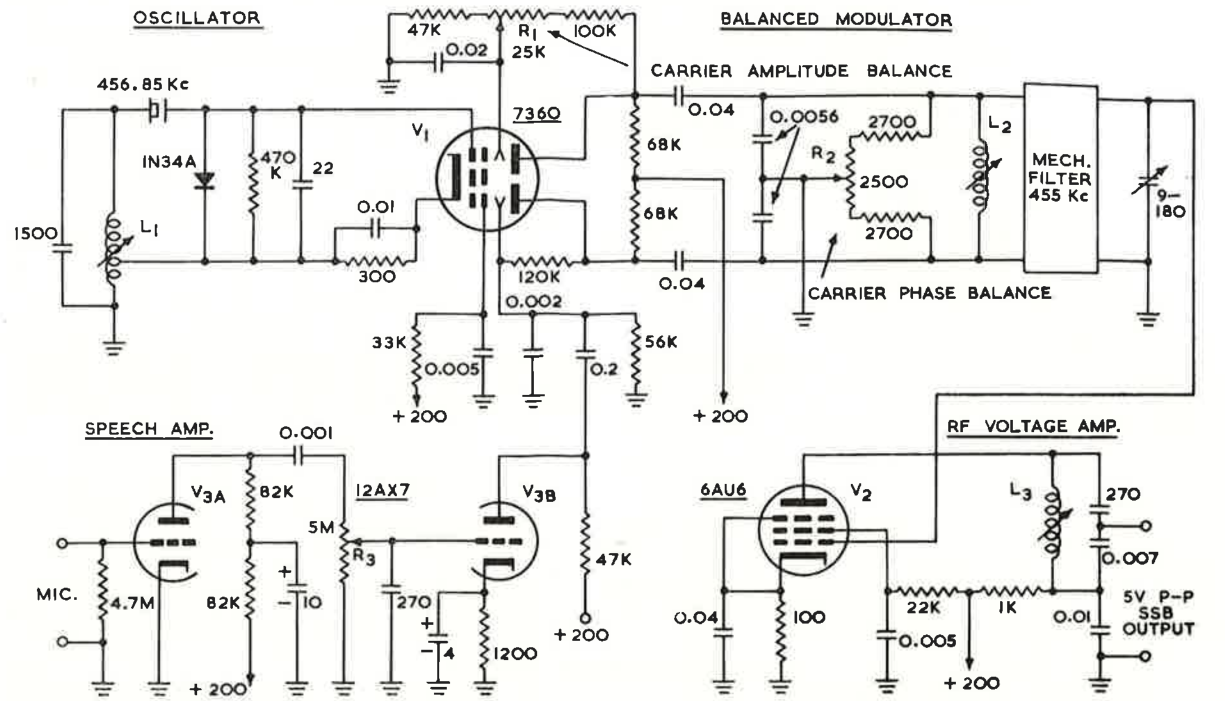

Fig. 2 — Filter-type SSB Generator using the 7360 as a Combined Self-excited Oscillator and Balanced Modulator. Fixed Resistors are ½ watt except as indicated.

L_1 — App. 88 μh, (see text).
L_2 — App. 50 μh, (see text).
L_3 — App. 450 μh, adjustable.

R_1, R_2 — Composition control, linear taper.
R_3 — Composition control, audio taper.

trols and accelerates a ribbon or sheet beam of electrons. The screen grid and the two deflecting electrodes act as a converging electron lens to focus this beam.

Varying the bias or signal voltage on the control grid varies the plate current as in a conventional valve. The total plate current to the two plates, at a given plate voltage, is determined by the voltages applied to the control grid and the screen grid. The division of the total plate current between the two plates is determined by the difference in voltage between the two deflecting electrodes.

Mechanical-Filter Type SSB Generator

Now, bearing these brief fundamentals in mind, let's see how they can be applied in a balanced modulator, using a mechanical filter to obtain a single-sideband signal. A 455-Kc circuit for this purpose is shown in Fig. 2.

The 7360 beam-deflection valve as used in this circuit combines two basic functions — it generates its own 456.85-Kc carrier as a crystal-controlled oscillator, and it also functions as a balanced modulator which delivers both sidebands without the carrier to the mechanical filter. The filter suppresses one side-band and delivers the other to its output circuit.

The control and screen grids of the 7360, together with its cathode, are used in a self-oscillating circuit. It is also entirely practical, of course, to supply the carrier to the control grid from a separate oscillator, if desired, as will be described later.

In the self-oscillatory circuit shown in Fig. 2 the 1N34A diode connected between the control grid and the ground side of the cathode-bias resistor acts as a clamp to prevent the voltage on the control grid from going positive or even to zero volts. As operated here the most positive excursion of the control grid is to -2.5 volts. This prevents excessive values of screen current from flowing and gives better modulaton linearity and carrier frequency stability.

We now have single-ended carrier input to our push-pull plate output. In order to suppress this carrier in the push-pull output circuit both ends of the output circuit must go equally positive and then equally negative at exactly the same times. That is, the amplitudes of the two voltages, one from each plate, must be exactly the same phase in order to balance out or cancel the carrier completely. Amplitude balance is obtained very simply by varying the dc voltage difference between the two deflecting electrodes by means of the amplitude-balance potentiometer, R_1.

Since a phase unbalance of only 1 degree makes it impossible to obtain more than about 40 db of carrier suppression, special pains were taken to provide good methods of obtaining stable phase balance. In the method used in Fig. 2 the phase angle of the load circuit is controlled by varying the resistance to ground from the two ends of the capacitance-centre tapped plate tank circuit by means of the phase-balance potentiometer, R_2. This method is best suited for relatively low-frequency operation. A small differential capacitor across the plate tank circuit is best suited for phase balancing in higher-frequency circuits.

These two balance controls allowed better than 60 db of carrier suppression to be obtained from the balanced modulator. An additional 20 db of suppression was obtained in this filter circuit because the carrier was located at a point 20 db down on the filter frequency-response curve. This gave a total of approximately 80 db suppression.

The audio signal is fed to this balanced modulator at a relatively high impedance — special transformers to obtain a low-impedance driving source are not required. Furthermore, a push-pull audio source is not required — a single-ended feed is used. With 10 volts rf on the control grid, 2.8 volts af deflecting voltage is required. Both values are peak-to-peak.

This arrangement is possible because of the fundamental characteristic of the valve mentioned initially — the voltage difference between the two deflecting electrodes controls the division of the total plate current between the two plates. This gives an intrinsically balanced push-pull output from a single-ended input.

Also, a 180-degree phase reversal can be obtained by simply switching the audio input from one deflecting electrode to the other.

In order not to overload the mechanical filter the gain of this circuit was reduced by

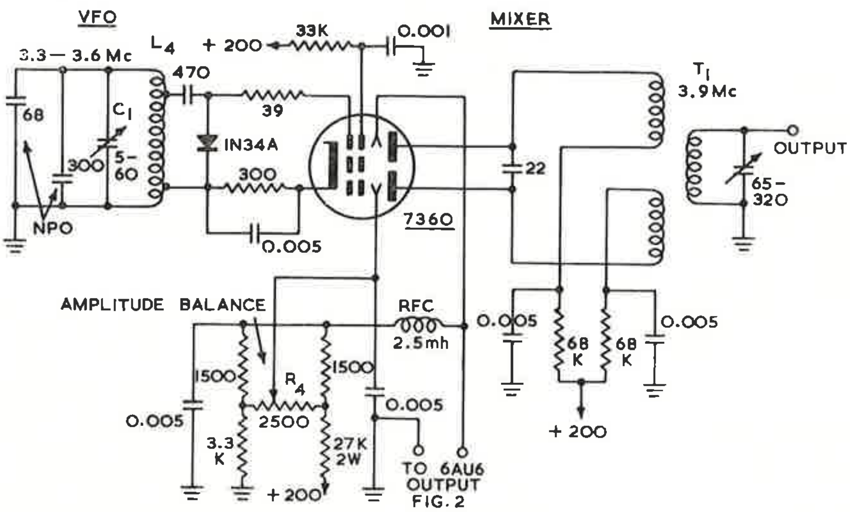

Fig. 3 — Combined VFO-Mixer for Frequency Conversion from 455 Kc to 3.8-4.0 Mc. Fixed Resistors are ½ watt except as indicated. Decimal Values of Capacitance are in μf, others are in $\mu\mu$f.

C_1 — Variable, 5-60 $\mu\mu$f.

L_4 — See text.

R_4 — Composition control, linear taper.

T_1 — See text.

Negative Feedback

You will notice that each defecting electrode receives its dc voltage (approximately 25 volts) from a tap on a resistance voltage divider from each plate to ground. A desirable by-product of this resistance coupling from plate to deflecting electrode is negative feedback. Additional negative feedback, which tends to correct any unbalance between the two plate currents, is obtained from the 68,000-ohm plate resistors. If desired, still more negative feedback can be obtained by reducing the capacitance of the two 0.04-μf plate blocking capacitors so that they offer a relatively high reactance to audio frequencies, but relatively low reactance to the carrier frequency being used. This negative feedback improves modulation linearity and reduces hum and microphonics originating in the balanced modulator.

shunting the mechanical filter input winding with a separate slug-tuned inductance, L_2 as shown in Fig. 2. This allowed about 1.5 volts input to the filter. Approximately 0.5 volt peak-to-peak output was obtained from the filter.

This shunting inductance, L_2, consisted of 63 turns of No. 36 enameled wire, close-wound in a single layer on a tube 9/32 inch in outside diameter. A ¼-inch-diameter iron slug in the tube allowed the plate tank circuit to be resonated at the crystal frequency.

The tank coil, L_1, for the 456.85-Kc crystal oscillator must be of the high-Q ferrite-core variety. One of the kind normally used as an oscillator coil with the 6BE6 converter for the standard AM broadcast band was used in our tests. The cathode excitation can be obtained by a tap on the coil. The cathode tap point should be located above ground 13 to 15 per cent of the total number of turns in the coil as a starting point. If possible, the rf voltage between grid and cathode should be measured with a high input-impedance vacuum-tube voltmeter equipped with an rf probe, and the tap point varied so as to obtain 10 volts rf peak-to-peak (3.5 volts rms) between the grid and cathode.

When using 1500 $\mu\mu$f total oscillator tank capacitance, provision should be made for varying the coil inductance above and below about 88.5 μh by a percentage a little larger than the capacitance tolerance percentage of the tank capacitor used, in order to resonate the coil-capacitor combination at the crystal frequency. Varying the tuning of the tank circuit around the resonance point will vary the oscillator frequency slightly.

Voltage Amplifier

A voltage amplifier suitable for raising the 0.5-volt output from the mechanical filter to a more usable level consists of a 6AU6 stage, Fig. 2, that has a capacitance voltage divider as a part of its plate tank circuit. With the constants shown, it can provide an SSB peak-to-peak output of about 5 volts. However, any output up to about 150 volts can be obtained by changing the capacitance ratio of the two voltage-divider capacitors that are connected in series across the 6AU6 plate tank circuit. When this ratio is changed the resultant capacitance of the two capacitors in series must remain constant so the L/C ratio of the tank circuit is not changed too much.

VFO — Mixer Circuit

Fig. 3 shows a schematic of a VFO mixer unit. Its resemblance to the balanced modulator circuit is quite evident. Here the front end again functions as an oscillator, except that it is of the variable-frequency type.

The modulating signal is the SSB output from the 6AU6 stage described above. Again the modulation is applied to only one deflecting electrode, the other being at rf ground due to the 0.005-μf bypass capacitor.

The mixer tank circuit employs a centre tapped, bifilar-wound inductor with the 68,000-ohm feedback resistors in its centre-tapped connections to the dc plate voltage supply. This mixer output transformer, T_1 in Fig. 3, was constructed as follows for our tests on 3.9 Mc:

Primary — Bifilar-wound on ½-inch diameter tube, tuned with a ¼-inch slug: winding length, 3/8-inch. Two wires wound parallel to each other on tube, 23½ turns of each wire (47 total). No. 34 wire, single Teflon insulation if possible (silk insulation can be used if necessary). The dielectric properties of the insulation on the wire are important because in a bifilar winding the distributed capacitance is relatively high and is a part of the tank capacitance. This accounts for the relatively low value of 22 $\mu\mu$f shown for the tank capacitor in this circuit.

Secondary — Twenty-six turns of No. 32 wire with Formex insulation, close-wound in a single layer. The spacing of this winding from one end of the primary winding should be adjusted so as to obtain satisfactory bandpass between 3.8 and 4.0 Mc. Approximately 190 $\mu\mu$f was required to resonate the secondary to 3.9 Mc.

Without the carrier amplitude-balance control, R_4, shown in Fig. 3, the balanced load circuit provides 20 to 25 db suppression of the VFO carrier. Including the carrier amplitude-balance control allows about 40 db total VFO carrier

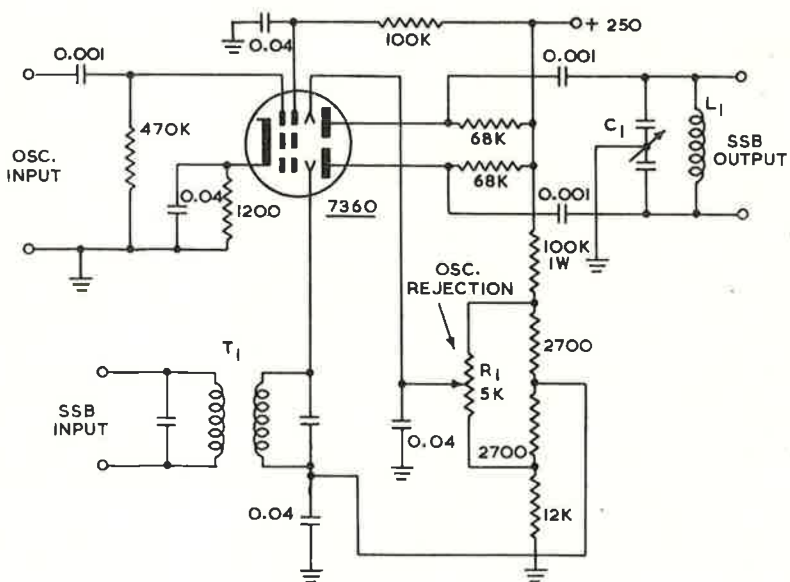

Fig. 4 — Balanced Mixer Circuit with Separate Excitation. Capacitances are in μf, Fixed Resistors are ½ watt except as Indicated. Tuned-circuit Constants Depend on Frequency. Ordinary L/C ratios may be used for L_1 C_1 and in the SSB input Transformer, T_1. R_1 is a linear-taper Composition Potentiometer.

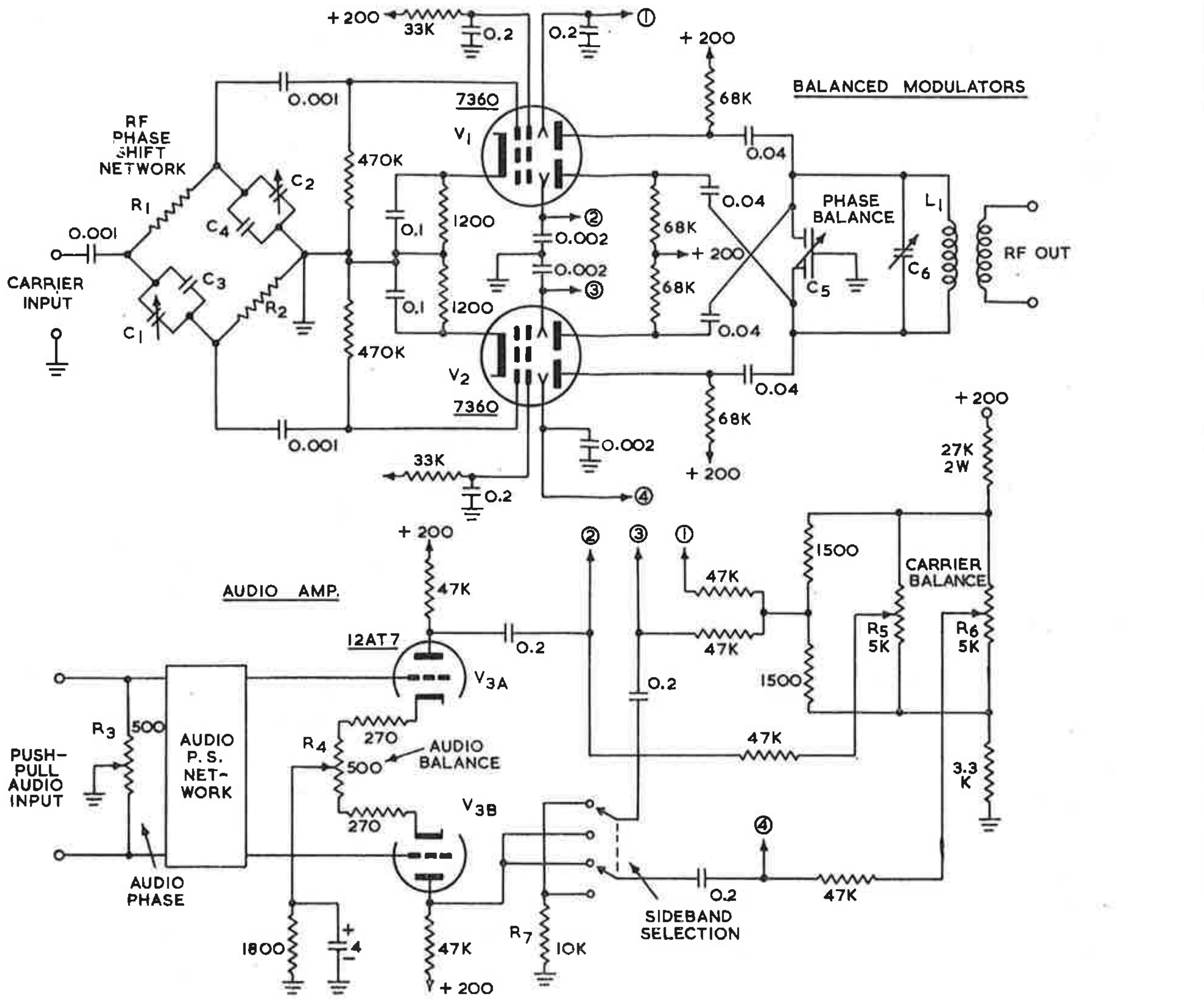

Fig. 5 — Phasing-type SSB Generator using two 7360s as Balanced Modulators. Capacitances are in $\mu\mu$f, Fixed Resistors are ½ watt Except as Indicated.

C_1 C_2	—	**Trimmers, ceramic or air, approximately 25 $\mu\mu$f.**
C_3, C_4	—	**Value dependent on frequency and R_1 R_2. Reactance at operating frequency should be approximately equal to the resistance of R_1 and R_2.**
C_5	—	**Differential capacitor, approximately 25 $\mu\mu$f per section.**
C_6	—	**Variable, to resonate with L_1 at output frequency.**
L_1	—	**To resonate with C_6 at output frequency. Conventional L/C ratio may be used.**
S_1	—	**DPDT toggle or rotary.**
R_1, R_2	—	**Non-inductive, ½ or 1 watt, values to be equal within close tolerances. Actual resistance not critical, but should be low to minimize effect to stray capacitances. Resistances of the order of 100 ohms are satisfactory.**
R_3, R_4	—	**500-ohm composition control, linear taper.**
R_5, R_6	—	**5000-ohm composition control, linear taper.**
R_7	—	**App. 10,000 ohms, ½ watt (see text).**

suppression, thus simplifying the selectivity requirements of the output circuit.

The grid and cathode connections of the VFO are tapped down on the inductor so as to reduce the coupling between the valve and the tank circuit, and thus improve stability and obtain the correct rf voltages on the valve elements.

The VFO tank coil, L_4 in Fig. 3, consists of 15 turns of No. 22 enameled wire spaced uniformly in a winding 0.6 inch long on a 1-inch diameter coil form. No slug was used. The grid tap should be located 7½ turns above the ground end of the coil. The cathode tap is 1-9/10 turns above the ground end of the coil. Actually, in building L_4, it was first determined that the coil specified above required a total length of wire measuring 121 cm. The taps were then soldered to this length of wire before winding it on the coil form, in order to avoid melting the polystyrene form with the hot soldering iron. The cathode tap was located 15.1 cm. (12½ per cent of the total wire length) from the ground end of the wire and the grid tap was located 60.5 cm. (50 per cent of the total wire length) from the ground end. At 3.5 Mc this coil had a Q of 150 and required 400 μμf to resonate it. The Q was measured with the coil shield in place.

7360 Mixer With Separate Oscillator

A generalized mixer circuit for use with an external rf oscillator is shown in Fig. 4. Here the SSB input is shown fed from a two-winding transformer instead of from a capacitive tap on the preceding tank circuit, as was used in Fig. 2.

Since the 7360 is not used as a self-oscillator the 1N34A diode clamp is not used and the cathode bias resistor is changed from 300 ohms to 1200 ohms. The rf oscillator input to the control grid must be adjusted to be between 5 and 10 volts peak-to-peak, measured between control grid and cathode, for best results. A 0.04-μf rf bypass capacitor effectively grounds one deflecting electrode (pin 9) so the SSB input is single-ended between the other deflecting electrode and ground.

In this mixer circuit the 68,000-ohm dc feed-back or plate-current equalizing resistors are connected as shunt feed resistors to the two plates, the same as was shown in the balanced modulator circuit in Fig. 2, instead of being in series with the centre-tapped connections to the dc plate voltage supply as was shown in the mixer circuit of Fig. 3. This difference allows L_1, the mixer plate tank coil, to be a simple untapped coil instead of requiring a bifilar-wound coil as is the case when the dc feed-back resistors are connected in series with the split, centre-tapped coil connections to the plate voltage supply. The capacitance and inductance values of T_1 and C_1 and L_1 in Fig. 4 will depend upon the input and output frequencies involved.

The combination of the two circuits shown in Fig. 2 and 3 will provide single-sideband output in the range between 3.8 and 4 Mc. Since these circuits were for the purpose of obtaining characteristics and specifications, as was previously stated, they do not include all of the facilities that might be required for actual amateur operation on the air, particularly as regards switching between upper and lower sideband.

Sideband switching can be obtained by any of the normal methods. For example, a simple method would be to employ two crystals in the carrier oscillator circuit of the balanced modulator, one for upper sideband and the other for lower sideband, with a switch for instant choice. Band switching would require an additional mixer stage to heterodyne the VFO mixer output to the various other bands.

RF Phasing-Type SSB Generator

Fig. 5 shows the schematic of an rf phasing exciter circuit for 455 Kc which gives a peak-to-peak output of about 4 volts single sideband. Here you see the usual rf and af 90-degree phase-shift networks, an audio amplifier, and two 7360 valves as balanced modulators.

The audio circuits include two sideband-balance potentiometers — one for adjusting the input voltage ratio to the audio phase shift net-

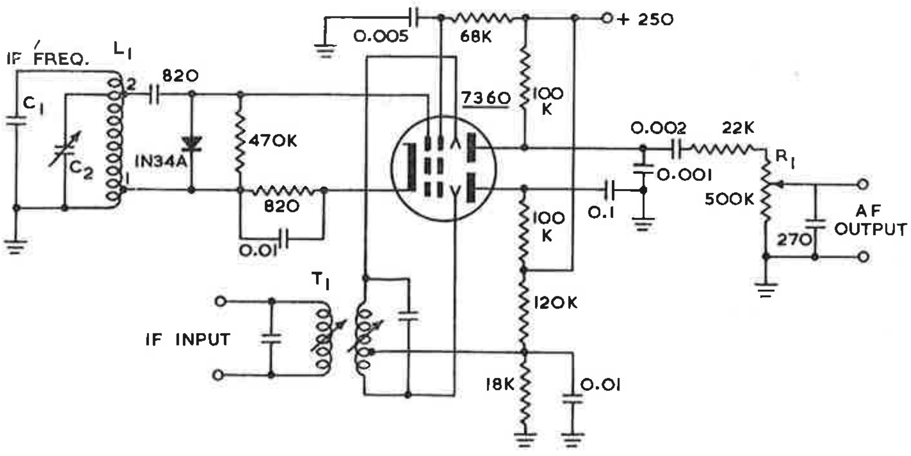

Fig. 6 — Product Detecor Circuit and Self-excitation. The beat-frequency oscillator tuned circuit, $C_1C_2L_1$, should be high-C at the intermediate frequency used; C_2 is for fine frequency adjustment. On L_1, tap 1 is approximately 5 per cent of the turns counted from ground, and tap 2 at approximately 25 per cent. T_1 is an if transformer with balanced secondary, such as is used for a push-pull diode detector, R_1 is the audio volume control.

work, R_3, and one for audio amplifier balance, R_4. Each balanced modulator has its own carrier amplitude-balance potentiometer which controls the dc bias voltage on one of its deflecting electrodes, as was done in the filter rig.

The outputs of the two balanced modulators are combined in a common push-pull tank circuit. Over-all rf phase balance is obtained in this tank circuit by the use of a differential capacitor connected across the tank circuit. This type of phase-balancing circuit is better suited to the higher carrier frequencies generally used in phasing-type exciters.

The rf phase-shift network used here is of the simple R/C bridge variety. Any of the other usual types of networks could be employed instead if desired.

As in the filter exciter, better than 60 db suppression of the carrier was obtained by proper adjustment of the rf phase and amplitude balancing controls.

Suppression of the unwanted sideband is limited by the degree of accuracy with which the audio phase-shift network maintains an exact difference of 90 degrees in phase between the two branches of the audio system, over the entire range of audio frequencies fed to the audio phase-shift network. With the audio phase-shift network used in our experimental setup carrier plus unwanted sideband measured about 40 db below the wanted sideband.

Switching the single-ended audio input to one of the balanced modulators from one deflecting electrode to the other allows a ready choice of upper or lower sideband output. This is done by switching the audio input to one balanced modulator, V_2, from one deflecting electrode to the other, through one arm of S_1. The other arm of S_1 connects a 10,000-ohm resistor, R_7, between the other deflecting electrode and ground through a dc blocking capacitor, in order to preserve better balance. The exact value of R_7 should be adjusted for best balance stability since various wiring layouts, and particularly various audio valve types, will require different values.

As in the case of the filter circuit, additional stages and functions would be required for a complete SSB exciter, including a VFO-mixer stage and a crystal oscillator-mixer stage to heterodyne the signal to various bands.

7360 as a Product Detector

A rather unique circuit for obtaining a single-ended audio output from the 7360 as a product dectector in a receiver is illustrated in Fig. 6. As in the VFO circuit of Fig. 3, the "front end" of the 7360 is used as a self-oscillating BFO and includes the diode clamp to prevent its control grid from going positive. A fairly high-Q, high-C BFO tank circuit is used, and with the coil tapped as in Fig. 6 about 8 volts peak-to-peak was obtained between control grid and cathode.

The deflecting electrodes were driven in push-pull primarily because it was felt that limiting of high-amplitude impulse noise would be better. Actually, tests with the same circuit connected for single-ended input to one deflecting electrode showed no substantial difference. Balanced operation has some second-order advantages, however, such as minimizing the contribution of stray capacitance to oscillator locking or pulling.

The if input to the deflecting electrodes should be held close to 10 volts peak-to-peak. At higher inputs the output "flattens" noticeably. If the input signal is held close to this maximum the flattening characteristic will limit impulse-noise peaks.

As indicated in Fig. 6, the BFO and audio output of one plate is bypassed to ground through a 0.1-μf capacitor. A 0.001-μf capacitor to ground from the other plate bypasses the BFO output to ground while the audio outut (more than 20 volts at maximum input signal) is used to drive the grid of a 6AQ5 output valve. This circuit was tested at an intermediate frequency of 910 Kc (the output from a BC-348M receiver). It is probable that more of an rf filter may be required at lower intermediate frequencies. For example, the 22,000-ohm rf filter resistor may need to be replaced with an if choke.

Maximum suppression of the if in the audio output of the 7360 plate circuit could be obtained by using a push-pull or balanced output circuit together with an amplitude-balance potentiometer on one of the deflecting electrodes, as is shown in Fig. 4.

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SUPER-RADIOTRON 23CP4 PICTURE TUBE

The 23CP4 is a directly viewed, bi-panel, rectangular, glass picture tube having an aluminized screen 19-5/16" x 15-1/4" with nearly straight sides and slightly rounded corners, and a minimum projected screen area of 282 square inches. It employs 110°-angle magnetic deflection and low-voltage electrostatic focus. The bi-panel construction used in the 23CP4 provides an integral Filterglass protective panel which is sealed to the faceplate of the tube. This construction eliminates the need for a separate safety-glass window and its companion dust seal in the receiver. It also eliminates the air separation between faceplate and protective window and thereby reduces reflections; consequently, picture contrast is improved.

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 / 2,000 min. $\mu\mu$f
- Faceplate and Protective Panel .... Filterglass
- Light transmission (approx.) 40%

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

Focusing Method .... Electrostatic

Deflection Method .... Magnetic

Deflection Angles (approx.):
- Diagonal .... 110°
- Horizontal .... 99°
- Vertical .... 82°

Electron Gun:
- Requires No Ion-Trap Magnet

Tube Dimensions:
- Overall Length .... 15-3/16" ± 3/8"
- Greatest Width .... 21-7/16"
- Greatest Height .... 17-7/16"
- Diagonal .... 24-23/32" ± 1/16"
- Neck Length .... 5-1/8" ± 1/8"

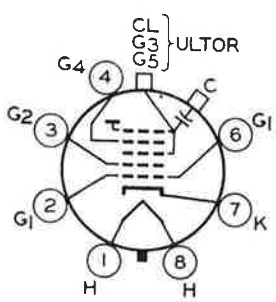
Screen Dimensions (minimum):
- Greatest width .... 19-5/16"
- Greatest Height .... 15-1/4"
- Diagonal .... 22-5/16"
- Projected area .... 282 sq. in.

Cap .... Recessed small cavity (JETEC No. J1-21)
Bulb .... J187 with Prot. Panel FP198
Base .... Small Button Eightar 7-pin (JETEC No. B7-208)

Weight (approx.) .... 33 lb.
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	20,000 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 20,000 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	18,000 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	20,000 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

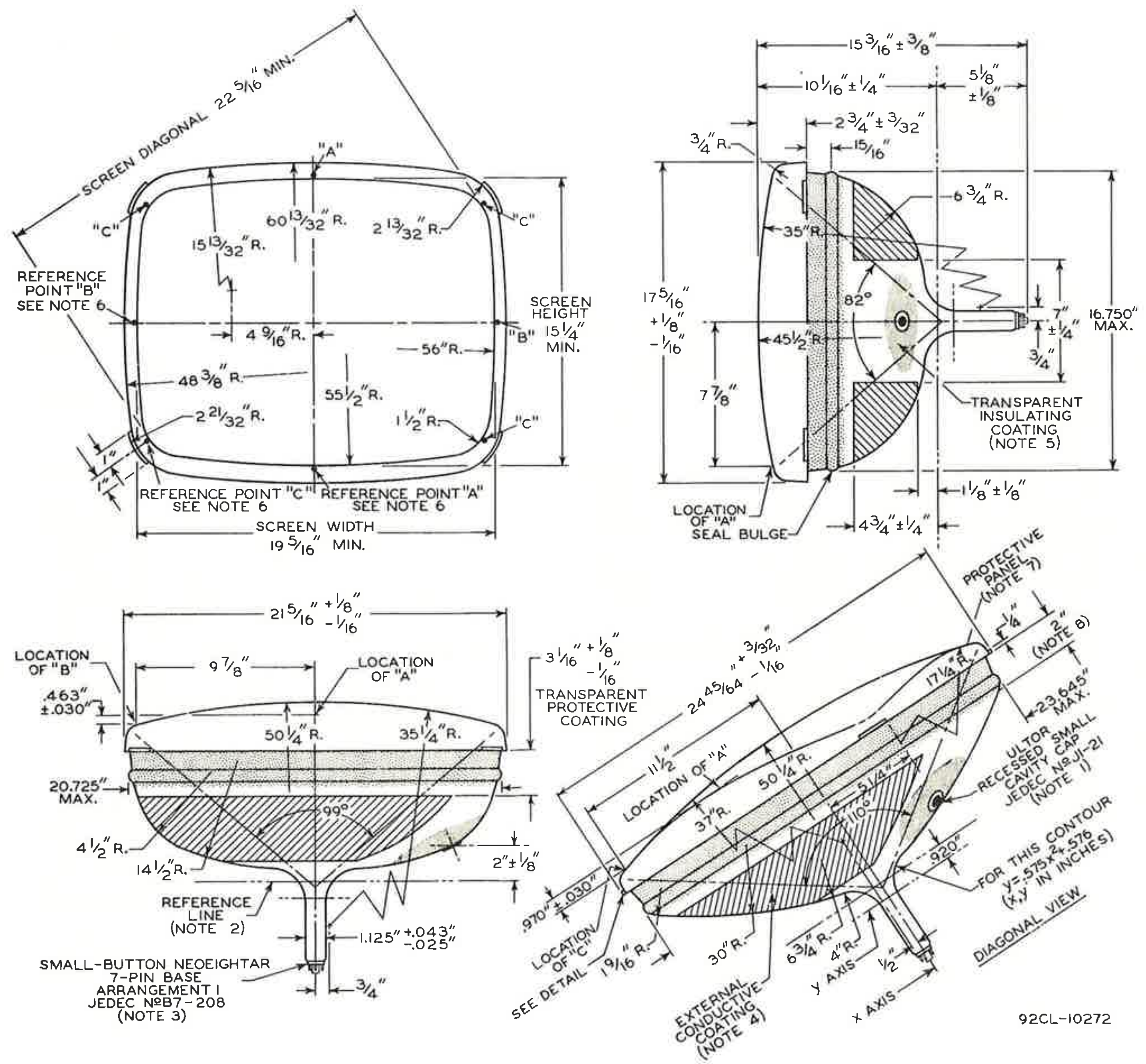
PEAK HEATER-CATHODE VOLTAGE:

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 20,000 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	18,000 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

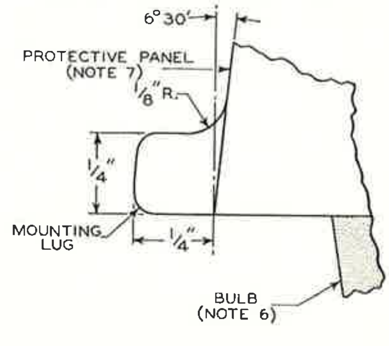
† The ultor in a cathode ray tube is the electrode to which is applied the highest dc voltage for accelerating the electrons in the beam prior to its deflection. In the 23CP4 the ultor function is performed by grid No. 5. Since grid No. 5, grid No. 3 and the collector are connected together within the 23CP4 they are collectively referred simply as ultor for convenience in presenting data and curves.

* 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 centring 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 3/8 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.

** 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 23CP4 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.

DETAIL OF MOUNTING LUG



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: External conductive coating must be grounded.

NOTE 5: To clean insulating coating around cavity contact, use only a soft dry lint-free cloth.

NOTE 6: Reference points A, B, and C are provided for use in design of a mask contoured for close fit to the protective panel.

NOTE 7: The centre of the protective panel may be eccentric with respect to the axis of the tube envelope. Associated shift of the protective panel along its minor and/or major axis will not exceed 1/16".

NOTE 8: Keep this circumferential area free of mounting hardware.

NOTE 9: Adequate tube support is obtained by clamping to the mounting lugs provided at each corner of the protective panel. Tube mounting and yoke support clamps must be spaced from the tube by use of cushioning pads made of material such as asphalt-impregnated felt, or equivalent.

Thermocouple Compensation Using A Thermistor

By J. ZIEGLER, B.Sc.

The thermocouple, although a cheap, reliable and potentially accurate means of temperature measurement, suffers from variations due to cold junction temperature fluctuation. To combat this, either the instrument scale zero may be shifted, or the cold junction maintained at a known temperature. In the case of a single low cost instrument installed at a small furnace, neither method is expedient. The use of compensating cable to extend the cold junction right to the instrument is not necessarily a solution, since often the instrument must itself be sited where ambient temperature change is large.

By reducing the circuit resistance as the ambient temperature rises, accurate compensation can be maintained at a single point on the instrument scale. Other points will be in error by some fraction of the difference between standard ambient and true ambient. The value of this fraction depends upon the scale reading, being unity when the furnace is at ambient temperature, and zero when the furnace is at the chosen point. Over a fairly wide range near the chosen point the compensating will be reasonably good. Such a scheme will be justified in most applications since the critical temperature is usually within a narrow range of the instrument scale. This object may be readily achieved by using a resistor having a negative temperature coefficient of resistance, such as a thermistor, in the thermocouple circuit.

In Fig. 1 is shown a general type of circuit for such compensation. RX is a suitable thermistor, having a constant coefficient over the range of ambient temperature intended. Fortunately, commercial thermistors fulfil this requirement sufficiently well for all practical purposes.

Procedure

The following method was found convenient. Suppose the lower ambient temperature for which compensation is required is T1, and the upper, T2. Let the values of RX at these points be RX1 and RX2.

Note that it will be an advantage to make R1 small compared with RX1, for then the setting of R1 has little effect on the reading at ambients near T1. This means that the meter may be calibrated near T1 by adjusting R3, and then near T2 by adjusting R1, no further adjustment being required. Otherwise, several successive adjustments are necessary.

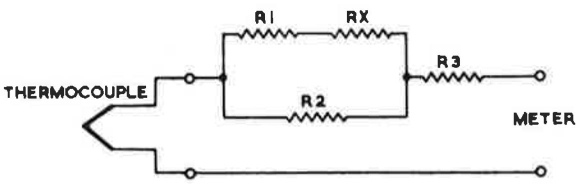


Fig. 1 — Circuit using Thermistor for Compensation.

Let the total circuit resistance required at T1 be R4, and that at T2 be R5. These are calculated from knowledge of the thermocouple output, the meter constants and the required operating point.

Writing:

$$K = R4 - R5$$
$$R' = R4 + RX1$$
$$R'' = R1 + RX2,$$

then:

$$\frac{R'R''}{R' + R2} - \frac{R''R2}{R'' + R2} = K$$

Solving for R2:

$$R2 = \frac{K(R' + R'')}{2(R' - R'' - K)} + \frac{\sqrt{K(R' - R'')\ [K(R' - R'') + 4R'R'']}}{1}$$

It remains to check whether a solution has been obtained, for:

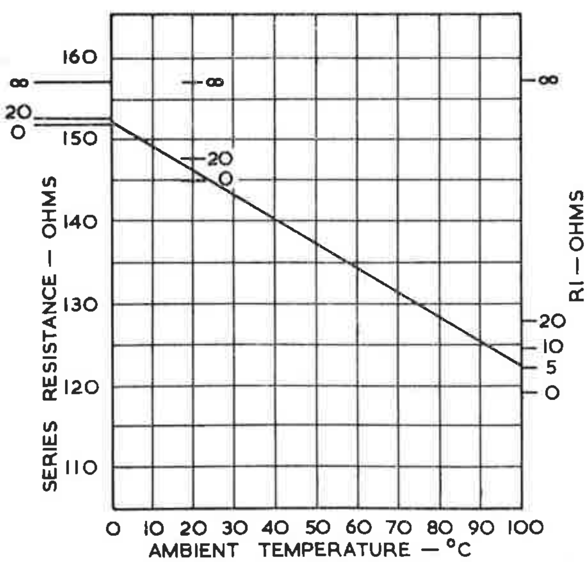
$$R5 = \frac{R''R2}{R'' + R2} + R3$$

+ meter resistance.

R3, for calibration purposes, must be at least a few per cent of R5. If this condition is not fulfilled, a second trial must be made. The difference T2 — T1 may be reduced, a more sensitive meter used, the thermocouple output increased by using series thermocouples, or another thermistor type substituted.

Example

The following circuit was designed to suit requirements of a small gas-fired glass annealing oven whose operating temperature is 520°C within 5°C over extended periods. The instrument cannot be prevented from traversing the range 20 to 40°C in the course of a day's run, and 10 to 50°C is the likely annual range. The "cold" ends of the thermocouple wires may reach 80°C.



NOTE: THE VALUES GIVEN AT 0°C, 20°C AND 100°C ARE VALUES OF R1. IT IS SEEN THAT R1 HAS LITTLE EFFECT ON THE SERIES RESISTANCE AT 20°C AND ALMOST NONE AT 0°C.

Fig. 2 — Chart showing Total Series Resistance Required.

The instrument used is a 100 ohm 17mV movement, calibrated 0/600°C in conjunction with the chromel-alumel couple. Following customary practice, zero is off-set to a "standard" ambient of 20°C.

Fig. 2 shows the total series resistance required in order that a true temperature of 520°C will indicate this figure on the meter, plotted as a function of ambient temperature.

Making T2 = 100°C and
T1 = 0°C,

we have R4 = 152 ohms and
R5 = 122 ohms.

The thermistor used is a STCK2211/120, for which

RX1 = 450 ohms approx. and
RX2 = 16 ohms approx.

Choosing R1 = 5 ohms,
then K = 30 ohms
R′ = 455 ohms.
R″ = 21 ohms.

Whence R2 = 49.5 ohms and
R3 = 7 ohms.

Performance

The above circuit was set up. R2 was set by ohmmeter to 2 per cent accuracy, and R1 set to approximately 5 ohms.

The furnace was run up to 520°C and at approximately 20°C ambient temperature the instrument was calibrated using R3. The cold junction was then heated by approximately 100°C, when it was recalibrated, using R1. No further adjustment was necessary.

No compensating cable is used, the thermistor being mounted at the cold ends of the thermocouple wires, connection being made otherwise with copper. It is possible to raise the cold junction temperature from 10 to approximately 120°C without an appreciable change in meter reading.

Note, finally, that it is convenient also to compensate for change in meter resistance with temperature by this method, if desired. The same calibration method is followed, except that the thermistor is mounted at the meter, compensating cable is used and the whole meter unit is heated to a suitable value of T2.

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HANDLING GLASS PICTURE TUBES

Warnings have been given before in these pages that care must be exercised in handling glass picture tubes, not only to protect the tube itself, but also the person handling it and others who may be nearby. With the current spread of television into States other than N.S.W. and Victoria, the time may be appropriate to repeat the warning, as there will be hundreds more people daily handling picture tubes.

It is important to get the danger inherent in an evacuated glass shape into proper perspective. The right way to approach the problem is to remember that implosions are rare, and that most people in the industry and handling these tubes have never witnessed an implosion. Remember when we take an air trip and buy an insurance certificate in the terminal hall whilst waiting to board the plane? Statistics show that you are much safer in the aircraft than crossing the road outside your office or workshop. Still we buy insurance against that more-than-a-million to one chance. Our precautions against implosion fall into the same category.

The second important thing to realise is just what happens when glass is knocked or bruised. Any glass ware, whether it be a picture tube envelope or a flower vase, will suffer damage when knocked. In many cases the damage is invisible and the glass is apparently unharmed. What possibly or even probably has happened is that the glass surface has been bruised. A tiny crack or flaw has opened up; steady atmospheric pressure and/or residual strains in the glass will do the rest with the aid of time.

Possibly several weeks or months later the glass will give way under the strain. This will often happen when the glass receives a further knock, which may in itself be very slight and much weaker than the original blow. We've all had experience of glass object shattering almost at a touch. Drinking glasses for example spend a total of some hours being clattered around in the washing-up water, along perhaps to meet an apparently ignominous end by shattering as they are placed on the storage shelf a little heavily. What has happened is that repeated rough handling has caused damage invisible to the naked eye and weakened the glass to the extent that it has finally given away.

This is why picture tubes must be handled AT ALL TIMES in such a way as to avoid this type of damage. Remember, even if you knock the tube accidentally with a screw-driver and it doesn't immediately implode, you haven't "got away with it". The chances are that the tube has been damaged. It is this fact as much as the general appearance of the tube that makes the manufacturer urge the maximum care in handling

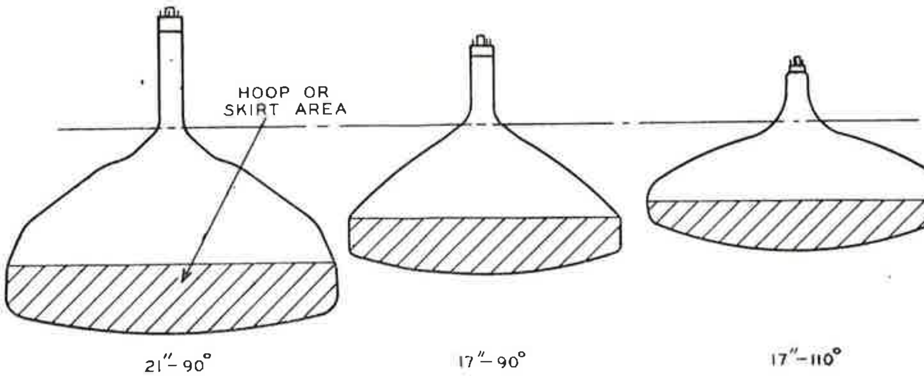

Fig. 1

tubes. Remember, all picture tubes are carefully inspected at the factory to ensure freedom from surface damage. There is very little chance that a tube which has a damage-free surface will implode. Provided the tube is handled carefully, both the handler and the final customer will be protected from the possibility of tube failure and flying glass.

Picture tubes should be kept in the carton or similar protective container until just prior to installation. In areas where unpacked and unprotected tubes exist, heavy protective clothing including gloves and eye shields should be worn. After the tube is removed from its carton, the "hoop" or "skirt" area must be protected from scratches, blows, or contact with metal or other glass surfaces. This critical area is shown in Fig. 1 for three types of glass picture tubes. The "hoop" area of tubes such as the 17-inch, 110-degree types is especially vulnerable because of the contour of the faceplate. The 21-inch, 110-degree contour is similar to that of the 17-inch, 110-degree tube.

Only a soft, dry cloth should be used to clean the "hoop" area. Wetting of the glass surface is not recommended. In the event that a tube is placed on its side, it should always be placed on material which will not scratch the glass surface, such as corrugated paper box liner or a rubber pad.

All picture tubes are subject to damage if they are placed side by side and allowed to knock together. Every precaution should be taken to eliminate the possibility of bruising the glass surface by bumping tubes together or accidentally striking them with tools during installation. The tube should never be hit deliberately with any instrument. Furthermore, the glass should be protected from direct contact with metal parts of the mounting assembly.

The foregoing remarks apply also of course to the new 19-inch and 23-inch picture tubes now coming into use. These tubes are of two types. The first type, such as the 23MP4, has the 23-inch contour but is otherwise similar to other all-glass tubes. Particular care should be taken with the corners of the faceplate.

The second type of 19-inch and 23-inch tube, such as the 23CP4, has a moulded glass implosion cap cemented to the front of the tube proper. Not only does this do away with the necessity for the safety glass window in the TV receiver, but renders the tube much safer in the event of the tube vacuum being broken. It is reported that even when the tube is forcibly destroyed, the implosion plate remains intact, and the cement holds the shattered faceplate, preventing it from flying. It will be obvious however, from the explanations already made that whether the tube is with or without the integral implosion cap, the same degree of care must be exercised in handling the tube.

LARGE AND SMALL

The rapid expansion of scientific endeavour in recent years, and the resulting need for greater precision in measuring and defining properties, has led to a need for both larger and smaller numbers. For the last decade the universally-used metric system has laboured along with double prefixes such as kilomegacycles, millimicroseconds, and so on; the use of micromicrofarads is even older.

When the metric system was first set up, the idea was to have a unique prefix for each order of magnitude. This idea no longer seems feasible except where the major orders are concerned, and is particularly necessary where the positive and negative ninth and twelfth orders of magnitude are concerned.

In 1958 the International Committee on Weights and Measurements adopted four new prefixes for denoting the positive and negative ninth and twelfth orders of magnitude. The National Bureau of Standards (U.S.A.) last year began using them, and other national committees seem likely to follow. Major publishers in the U.K. and U.S.A. have adopted some or all of the new prefixes.

"Radiotronics" will use the new prefixes on new material, so that they will be progressively introduced over the next few months. The prefixes and abbreviations to be used in these pages in the future are as follows:—

10^{12}	tera — T	10^{-12}	pico — p
10^{9}	giga — G	10^{-9}	nano — n
10^{6}	mega — M	10^{-6}	micro — μ
10^{3}	kilo — K	10^{-3}	milli — m
10^{2}	hecto — H	10^{-2}	centi — c
10^{1}	deka — D	10^{-1}	deci — d

NEW RELEASES

6L6-GC

The 6L6-GC is a high-perveance beam power valve of the glass-octal type designed for use in the output stage of audio amplifiers and radio receivers where relatively large power output is required. The 6L6-GC features high power output, high efficiency, and high sensitivity. In push-pull class AB_1 af power-amplifier service, two 6L6-GC's operating with a grid-No. 1 voltage of -37 volts, plate voltage of 450 volts, grid-No. 2 voltage of 400 volts, and peak af grid-No. 1-to-grid No. 1 voltage of 70 volts, can deliver a maximum-signal power output of 55 watts with a total harmonic distortion of only 1.8 per cent.

6GM6

The 6GM6 is a semiremote-cutoff pentode of the 7-pin miniature type especially designed for use in gain-controlled picture-if stages of television receivers operating at intermediate frequencies of the order of 40 megacycles. The 6GM6 has high transconductance 13,000 micromhos) and relatively low capacitances, features which contribute to a large gain-bandwidth product. The semiremote-cutoff characteristic of the 6GM6 minimizes envelope distortion and cross modulation enabling it to handle large signal voltages.

6CA4

The 6CA4 is a full-wave vacuum rectifier valve of the 9-pin miniature type intended for use in compact high-fidelity audio equipment. This type, which utilizes a unipotential cathode with a 6.3-volt heater, has high-voltage insulation between heater and cathode to eliminate the need for a separate winding on the power transformer. This feature permits the use of a smaller power transformer for compact equipment design. The 6CA4 also features low internal resistance to assure good power-supply regulation and a small internal voltage drop.

6FH5

The 6FH5 is a semiremote-cutoff tetrode of the 7-pin miniature type designed particularly for use as a grounded- cathode rf-amplifier valve in vhf tuners of television receivers. When connected as a triode, the 6FH5 features high transconductance (9000 micromhos) at a relatively low plate voltage. Grid No. 2, designed primarily for use as a shield, provides low grid-No. 1-to-plate capacitance to facilitate neutralization, reduce effective lead resistance and inductance, and facilitate isolation of the input and output circuits. The semiremote-cutoff characteristic of the 6FH5 minimizes intermodulation distortion.

6GH8

The 6GH8 is a multiunit valve of the 9-pin miniature type containing a medium-mu triode and a sharp-cutoff pentode in one envelope. It is intended primarily for use in multi-vibrator-type horizontal-deflection oscillator circuits in television receivers. It is also suitable for use as an agc-amplifier or sync-separator valve in such receivers. The 6GH8 features a high peak cathode-current rating and low inter-electrode leakage. These features assure dependable performance in horizontal-deflection-oscillator and sync-separator circuits. The high transconductance at low plate current of both units makes this type an efficient amplifier.

EEV BR1131

The BR1131, a high power transmitting valve, is a forced-air cooled triode. It has characteristics similar to those of the BR152B (CV28) but employs a thoriated-tungsten filament, rated at 8.5 volts, 22 amperes, and has a maximum anode dissipation of 3.5 Kw.

EEV BR1138

The BR1138 is a forced-air cooled triode identical with BR152B (CV28) apart from the filament which is of thoriated tungsten and rated at 8.5 volts, 22 amperes. This high power transmitting valve may be used as a direct replacement for the CV28 in existing equipments, the difference in filament voltage being readily accommodated by connecting the filaments of pairs of BR1138 in series.

EEV BW194

The BW194 is a water-cooled version of the BR194 with an anode dissipation of 50Kw. The filament rating is 13.0 volts, 240 amperes, and the plate dissipation 50 Kw maximum.

AWV