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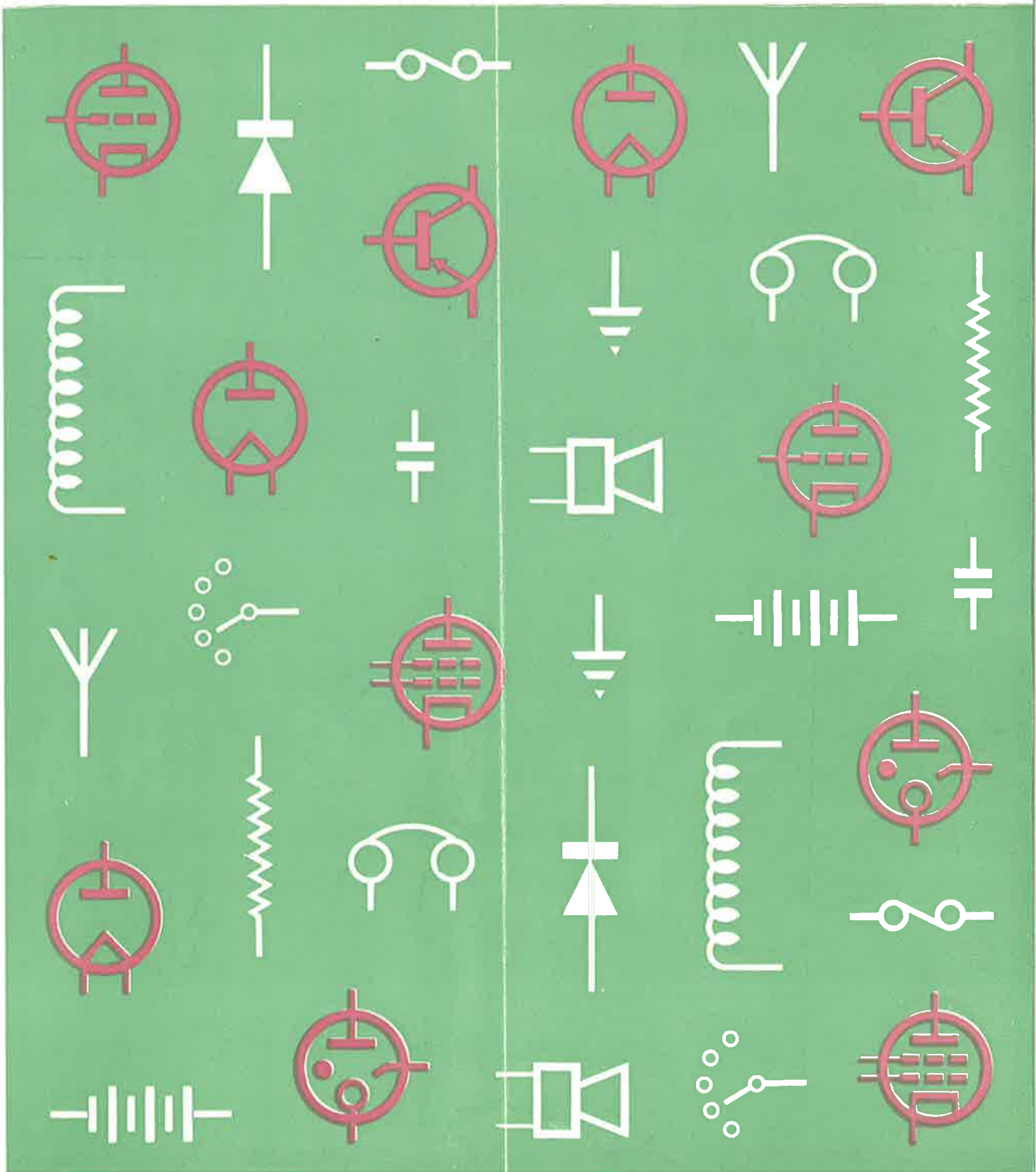

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Editor, Bernard J. Simpson

IN THIS ISSUE

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AUTOMATIC FREQUENCY CONTROL USING SEMICONDUCTOR DIODES

Introduction

The achievement of frequency stability of the local oscillator in superheterodyne receivers has exercised the ingenuity of circuit designers for a considerable time. Hitherto, the methods of providing automatic compensation for frequency drift have revolved round the reactance valve. These methods have very often proved unsatisfactory because, in the absence of elaborate and, therefore, costly precautions, the drift introduced by the reactance valve and its associated circuits has outweighed the correction of the original drift provided by the reactance valve.

The advent of the semiconductor diode has provided an alternative to the reactance valve that has many advantages. These include small size, simplicity of the associated circuits, low power consumption, and a capacitance variation that is a function solely of the variations in the bias applied to the diode. These factors combine to give a device that can be used to give a useful reduction in frequency drift in practical circuits.

This report discusses the principles involved in the provision of automatic frequency control, considering in turn the oscillator, the frequency sensitive detector and the feedback loop. The report is illustrated by a practical example.

Capacitance of a P-N Junction

It is found that the capacitance of a p-n junction is a function of voltage. Since the capacitance is substantially independent of temperature and since it has negligible shunt conductance when biased in the reverse direction, its variation with voltage can be exploited in applications which call for variable reactance elements. It is found that the relationship between capacitance (C) and voltage (V) takes the general form:

$$C = \frac{k}{\sqrt[n]{U-V}} \qquad \ldots\ldots\ldots\ldots (1)$$

where U is a constant determined by the materials and methods used in making the diode; n is a constant determined by the structure of the p-n junction; k is a constant of proportionality. Attention should be paid to the sign of V.

For the SVC1 series of variable capacitance diodes, U has the value of +0.6 volt and n has the value of 2. In addition, a stray capacitance of about 1 pf appears in parallel with the junction capacitance. This is produced by the leads and the case.

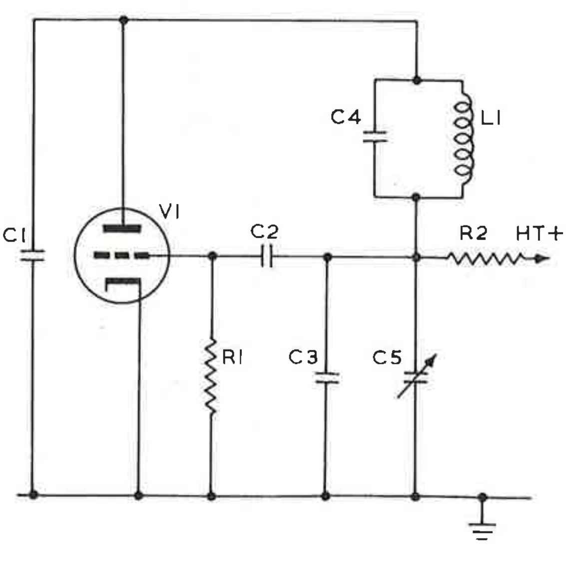


Fig. 1 — Basic Colpitts Oscillator Circuit.

Frequency Control of a Colpitts Oscillator

The basic circuit of a Colpitts oscillator is shown in Fig. 1. The capacitors C1, C3 and C4 include all the stray capacitances between their respective terminals and the variable capacitor C5 represents the variable capacitance diode. The effective total capacitance C_{tot} which resonates with the inductor L1, is, therefore, given by:

$$C_{tot} = C1 + \frac{(C3 + C5)\,C1}{C1 + C3 + C5} \quad \text{(2)}$$

The effect of variations in the value of C5 on the frequency of oscillation may be derived from equation (2) and the expression:

$$f = \frac{1}{2\pi\sqrt{LC_{tot}}} \quad \text{(3)}$$

Differentiating:

$$\frac{df}{dC_{tot}} = -\frac{f}{2C_{tot}}$$

$$\text{and} \quad \frac{dC_{tot}}{dC5} = \left(\frac{C1}{C1 + C3 + C5}\right)^2$$

$$\text{hence} \quad \frac{df}{dC5} = -\frac{f}{2C_{tot}}\left(\frac{C1}{C1 + C3 + C5}\right)^2 \quad \text{(4)}$$

This expression enables the change of frequency due to a small change in the value of C5 to be calculated, but it is of greater interest to consider the magnitude of the error signal that must be applied to the variable capacitance diode to produce a given frequency change. Equation (1) provides the additional information necessary to do this.

If the externally applied bias V is written as $V_0 + V_1$, where V_0 is the fixed dc bias on the diode and V_1 is the error signal, and if V_1 is assumed to be small compared with V_0, then it may be shown for the SVC1 series of diodes that:

$$\left.\frac{\delta C5}{\delta V_1}\right|_{V_0\ \text{const}} = \frac{C_0}{2(U - V_0)} \quad \text{(5)}$$

where $C5 = C_0$, when $V - V_0$.

The change in frequency due to the error signal V_1 is given by:

$$\delta f = \left.\frac{\delta f}{\delta V_1}\right|_{V_0\ \text{const.}} \times V_1$$

Hence, combining equations (4) and (5) yields the required differential coefficient as:

$$\left.\frac{\delta f}{\delta V_1}\right|_{V_0\ \text{const.}} = -\frac{f_0 C_0}{4C_{tot}\,(U - V_0)}\left(\frac{C1}{C1+C3+C_0}\right)^2 \quad \text{(6)}$$

where f_0 is the centre frequency, namely the value of f when $V = V_0$. The negative sign indicates that the frequency of the oscillator is shifted upwards when V_1 is negative (i.e. when it acts to increase the negative bias on the variable capacitance diode) and vice versa. This result is in accord with the capacitance change of the variable capacitance diode, which decreases with increasing negative bias.

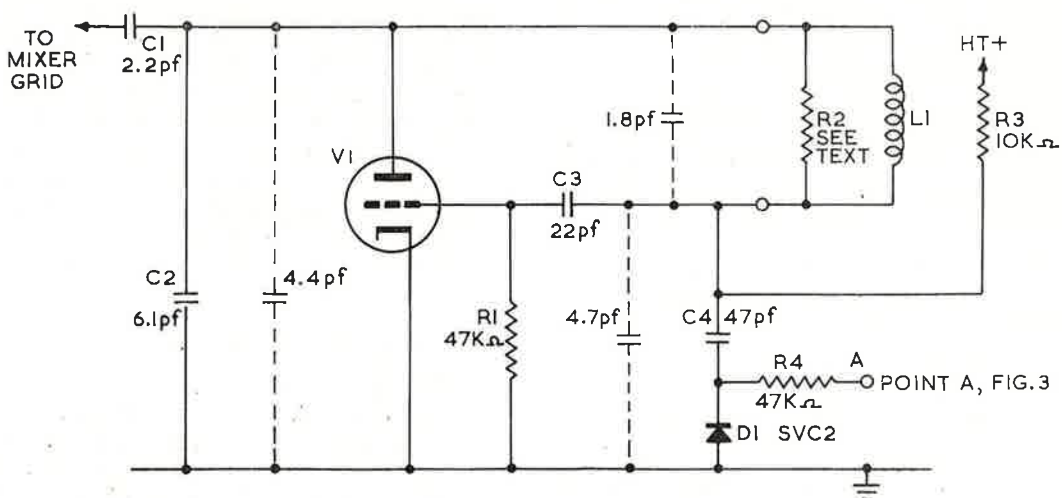


Fig. 2 — A Typical Colpitts Oscillator Circuit. Values shown are typical and do not represent any specific circuit.

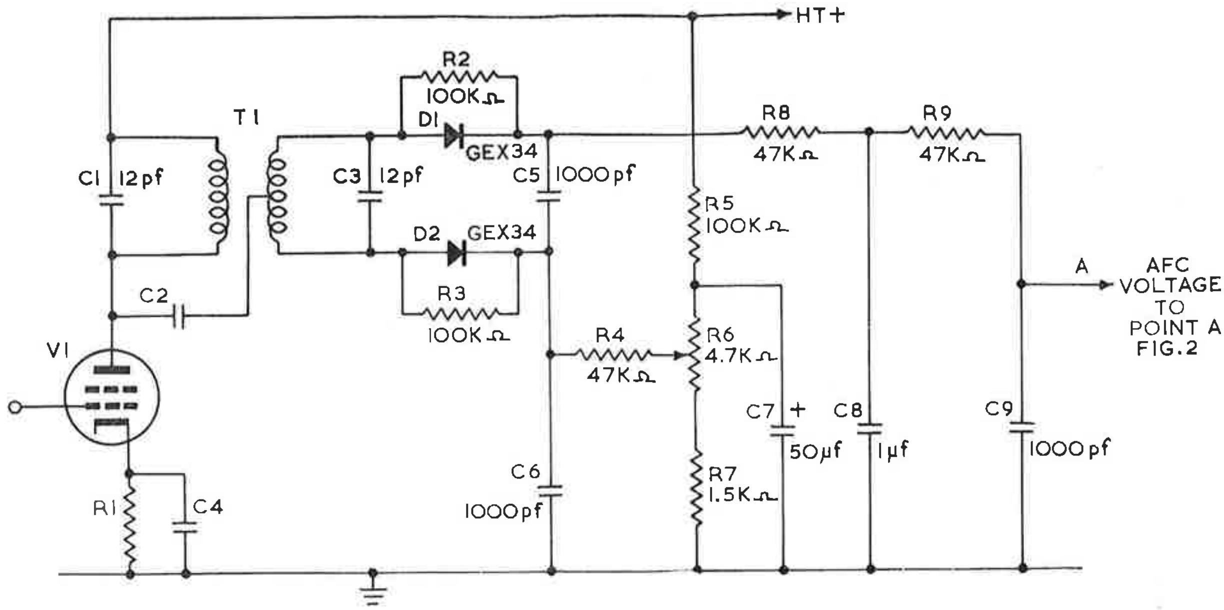


Fig. 3 — Foster-Seeley Discriminator and Filter Circuit. (Values of R1, C4 are determined by the valve used).

To achieve the optimum performance for the afc circuit, it is necessary to choose the operating point of the variable capacitance diode with some care. Three factors affect this choice. First, the dc bias must be such that the diode does not conduct during any part of the oscillator cycle; conduction may be caused by the rf voltage generated by the oscillator swinging the diode either into the forward conducting region or into the reverse breakdown region. If an excess in either direction occurs, the oscillator waveform will be seriously distorted and its efficiency will be impaired.

Secondly, sensitivity considerations make it desirable that, for any given value of the error signal V_1, the resultant capacitance change should be a maximum. An examination of equations (1) and (5) shows that $\delta C/\delta V_1$ is a maximum at low values of the standing bias V_o.

Thirdly, and perhaps less important since it can readily be accommodated by suitably designing the remainder of the oscillator circuit, the desired mean capacitance of the variable capacitance diode must be borne in mind.

Clearly a compromise must be made between the demands of these three requirements. As a general rule, the lowest value of V_o consistent with the first requirement is desirable in order to obtain maximum sensitivity.

The circuit diagram of a typical Colpitts oscillator suitable for use in Bands 1, 2 and 3* is shown in Fig. 2. The circuit has been modified by the addition of a variable capacitance diode; the 47 pf capacitor C4 in series with the diode serves to isolate it from the HT+ supply. The stray capacitance present in the circuit is also shown. Channel switching is accomplished by changing the inductors.

It was found that the fraction of the oscillator voltage developed across the diode at Band 3 frequencies had a peak amplitude of 3 to 4 volts. For this reason the nominal operating bias chosen for the variable capacitance diode was –6 volts. This value permits fine adjustments and variations due to the error signal without fear that the diode may swing into the forward conducting region. In Band 1, however, the oscillator amplitude was considerably greater than that in Band 3. It was therefore found necessary to damp the oscillations at Band 1 frequencies by the addition of the resistor R2 in parallel with the appropriate oscillator coils. A 2,700-ohm resistor was found to be suitable.

The total capacitance of a typical SVC1 diode at –6 volts is 5.3 pf. When allowance is made for the stray capacitance the junction capacitance is found to be 4.3 pf at –6 volts. This value, together with the typical values given in Fig. 2, yields the sensitivity as $7.1 f_o \times 10^{-3}$ Mc/V.

Frequency-Sensitive Detector

The discriminator used in the experimental system was of the Foster-Seeley type but a ratio detector could have been used equally well. Although the use of Foster-Seeley discriminator necessitates a limiting driver stage, this has the advantage that the loop gain of the afc system can be made independent of the gain of the if stages over a much wider range of frequency drift than is possible with the ratio detector.

* European TV Bands 1, 2 and 3 referred to here occupy 41 to 68 Mc, 87.5 to 100 Mc, and 174 to 216 Mc respectively.

The experimental circuit utilised shunt fed diodes in the detector, as shown in Fig. 3. This fact, combined with the need to provide a dc output only for afc purposes, has made possible a considerable simplification of the circuit.

Fig. 3 also shows the method of adding the discriminator output in series with the dc bias for the variable capacitance diode together with the filter circuit required for feeding the bias to the diode. These circuits are designed to ensure that the bias applied to the diode has no ac components. If this precaution is not taken, spurious frequency modulation of the oscillator will result, with a consequent deterioration in performance. It is important to ensure that the final decoupling resistor is adequately screened in order to prevent stray pick-up.

Performance

The performance of the afc system was assessed by measuring the oscillator frequency drift when the afc was functioning and comparing this with the oscillator drift with the afc inoperative. The result has been expressed as a ratio, termed the drift reduction factor.

Practical measurements made on a commercial television receiver modified by the inclusion of the afc circuits yielded the following drift reduction factors:

Band 1, Channel 1 : 2.9.
Band 3, Channel 9 : 6.5.

The improved performance in Channel 9 derives from the increased sensitivity of the oscillator to small capacitance changes at higher frequencies. This is shown theoretically in the equation (6).

(With acknowledgements to G.E.C., Application Report No. 15).

FEEDBACK WITH A DIFFERENT VOICE COIL IMPEDANCE

It is quite a simple matter for anyone with a slide rule to calculate the correct series feedback resistor to use with any voice coil, when the published value is not suitable. The formula is:

$$R_{f2} = R_{f1} \sqrt{\frac{Z_2}{Z_1}} \quad \ldots\ldots\ldots\ldots\ldots \quad (1)$$

where Z_1 = published voice coil impedance
Z_2 = required voice coil impedance
R_{f1} = published series feedback resistor
R_{f2} = required series feedback resistor.

Equation (1) is a close approximation only when the feedback resistor is considerably larger than the ac impedance between the point to which the feedback is applied, and earth.

A capacitor C_f shunted across R_f follows the inverse relationship:

$$C_{f2} = C_{f1} \sqrt{\frac{Z_1}{Z_2}} \quad \ldots\ldots\ldots\ldots\ldots \quad (2)$$

If a slide-rule is not available, the table may be used to determine the ratio $\sqrt{(Z_2/Z_1)}$ for use in equation (1). Its inverse may be used in equation (2).

TABLE OF $\sqrt{(Z_2/Z_1)}$

Z_1				Z_2			
	2.0	3.5	6.0	8.0	10	12.5	15
2.0	1.0	1.32	1.73	2.0	2.23	2.5	2.73
3.5	0.76	1.0	1.71	1.51	1.69	1.89	2.07
6.0	0.58	0.76	1.0	1.15	1.29	1.44	1.58
8.0	0.5	0.66	0.86	1.0	1.12	1.25	1.37
10	0.45	0.59	0.77	0.89	1.0	1.12	1.22
12.5	0.40	0.53	0.69	0.80	0.89	1.0	1.09
15	0.36	0.48	0.63	0.73	0.82	0.91	1.0

BANDPASS TRANSMITTER-EXCITER USING THE 6146

By Richard G. Talpey, W2PUD*

This article, the first part of which appeared last month, is concluded in this issue. The instalment last month dealt with the design of the unit, and contained the schematic diagram and parts list. This concluding part contains the constructional details and adjustment procedures.

CONSTRUCTION

The transmitter is built on an 8 by 17 by 3-inch aluminium chassis. The construction is somewhat unconventional inasmuch as the controls project out of the bottom of the chassis through a standard 8¾-inch relay-rack panel which forms the front of the shield enclosure.

The VFO is completely housed in the smaller aluminium box shown in Fig. 4. The larger box shields the 6146 final amplifier. The layout of the components of the VFO and the final-amplifier plate circuit are shown in Figures 5 and 7, respectively. Most of the other components are shown in Fig. 6 which is a close-up view of Fig. 2. This method of construction permits the bandswitches to be coupled with a single right-angle drive. This arrangement provides single-knob control of all bandswitches, thereby facilitating the layout.

The shield for the final-amplifier plate circuit was made from two aluminium cases. The unflanged portions were discarded, and the flanged sections were overlapped in the centre. A sheet of aluminium was cut to fit the top.

The bandswitch is mounted in the centre of the chassis so that the switch sections are located near the multiplier valves. The bandswitch is made from a standard index assembly and separate switch sections selected according to function. A standard two-section switch for the final tank is mounted above the chassis in line with the bandswitch knob: the switches for the multipliers are coupled to the right-angle drive located inside the chassis. The other controls are placed to provide a neat panel arrangement.

VFO

The VFO coil, L1, was wound by hand on a piece of mailing tube covered with a layer of wax paper. The wire was wound over the wax paper and spaced to occupy the required length. A few extra turns were included to allow for final trimming. A coat of household cement was applied in three longitudinal stripes 120° apart to secure the winding. A second coat was applied

* Tube Dept., Radio Corp. of America.

after the first coat hardened. The mailing tube was then collapsed and withdrawn along with the wax paper. Finally, each cement stripe was given one more coat of cement (inside and out) to make the coil rigid. After trimming, the whole coil was cemented to a ½ by 1/16-in. Polythene strip which was mounted on ceramic standoff insulators. This type of coil is very rugged and has the high Q required for the Clapp oscillator. The coil must be mounted as far from the sides of the shield as possible, because the shield acts like a shorted turn coupled to the coil and will reduce its Q materially if the spacing is made too small. Care should also be taken to see that the tuning capacitors and other parts cannot move with respect to the 'hot' end of the coil, which is the end connected to the capacitors.

The socket for the 12AU7 is mounted on metal spacers to permit the connections to be made easily. All of the VFO components (as well as connections) should be made as rigid as possible because the stability of a Clapp oscillator depends to a great extent on its mechanical construction. The grounds to the shield braids for the power leads should be made near the hole where they enter the compartment, and the by-pass capacitors should be grounded to this same point.

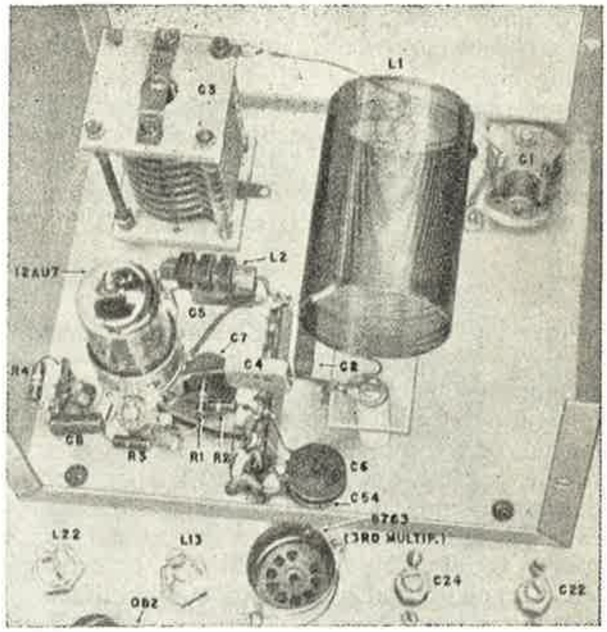


Fig. 4 — Inside view of the VFO (with cover removed). Note that the coupling capacitor C8 and R4, the grid resistor for the 6AU6 keyed amplifier, are located in the VFO compartment (see text).

General Layout

As many of the holes as possible should be drilled beforehand to eliminate difficulty later on. The paint should be scraped off the back of the panel where it butts against the flange of the chassis to insure a good rf connection and to prevent rf leakage. Careful layout of the panel is required to insure that the shafts for all controls line up properly. Be sure to use panel bearings where the shafts protrude through the panel to prevent the shafts from becoming antennas for TVI. It is helpful to drill the holes for the shield cans and make a trial assembly of the shields before mounting any of the major components. Trial fits for shaft line-up for the bandswitch and tuning capacitors are also recommended.

"First-Layer" Wiring

After making certain that everything will fit where intended, the valve sockets may be mounted and the heater and power wiring started. All grounds for each stage are made to lugs bolted under the valve-socket lugs. Components which are not mounted directly on the valve sockets are mounted on tie lugs bolted to the sides of the chassis. All of this "first-layer" wiring is best done before assembly of the bandswitch and coils.

Power is brought into the transmitter by means of an octal socket, and all leads are by-passed at the socket to a common ground point (to which are also tied the shield braids of the power wiring). Pins 5 and 6 on power plug P1 are connected by a jumper on socket J1. This arrangement serves as an interlock for the external power supply by preventing application of power to the primary winding of the plate-voltage transformer when plug P1 is removed from J1. Bypass capacitors for the 600-volt leads are made from two 0.01-μf, ceramic-disc capacitors connected in series to provide adequate voltage rating. The by-pass capacitor for the high-voltage lead to the final-amplifier plate is mounted through a hole in the chassis and supported by a small bracket as shown in Fig. 7.

Coils

After the preliminary wiring is completed, the bandswitch and the coils may be mounted. The coils should be wound according to the coil specifications given in the Parts List. Before they are mounted, the coils (with the exception of the link windings which will have to be adjusted later) should be given a coat of polystyrene coil dope. To obtain a high coefficient of coupling, the links are wound over a layer of cellophane tape on top of the main windings of the 3.5-Mc coils (L4, L5, L15, and L16).

All other links are wound at the "cold" end of the coils with provisions to move them slightly during line-up. Links L14 and L21 are made from a single length of No. 18 stiff, insulated wire and supported by cement on L13 and L22, respectively, after final adjustment. Link connections to the bandswitch and between various coils are made with 75-ohm twin lead.

Adding the VFO

After the VFO section has been constructed, it may be placed onto the main chassis at any convenient time. The output lead connects directly to the grid of the 6AU6; make certain that the portion projecting from the braid is as short as possible. Because this lead is in the low-impedance output circuit of the cathode follower, it's length is not critical. Grid capacitor C8 and resistor R4 are placed inside the VFO shield to preclude any possibility of radiation from exposed parts.

Connections to Panel

Initially, the leads to the switches on the panel should be longer than needed so that it will be convenient to allow the panel to rest on the bench while initial adjustments are made. After the adjustments are completed and the unit is ready for "buttoning up," these leads may be shortened and connected to the switches; they should be just long enough to allow the panel to be swung out.

TVI Precautions

The rear of the meter case is covered with a shield cut from a food can. It is easy to find the right size and cut it with a pair of tin snips. The meter chosen must be short enough not to interfere with the coils which are mounted inside the chassis. The meter shunts, R29 and R33, were wound with resistance wire to provide full-scale readings of 200 ma for the final plate current and 10 ma for the final grid current, respectively.

Final Amplifier

The coils for the pi network are constructed as noted in the Parts List, and no difficulty should be encountered if the taps are located as shown in the coil specifications. Coil L27 is mounted to the chassis by means of a small bracket which is attached to one of the plastic strips which was left a bit longer for this purpose. Coil L28 is supported by means of its leads, all of which are short. The output lead from L28 to the coaxial connector is shielded to reduce its inductance and to reduce stray pickup. Padding capacitors C50 and C53 are mounted between the bandswitch and ground lugs located directly underneath.

All under-the-chassis ground connections for the final amplifier are made to a lug which is mounted on top of the chassis and bent down through a clearance hole to receive the under-chassis leads. This arrangement keeps all rf paths on one side of the chassis and as short as possible.

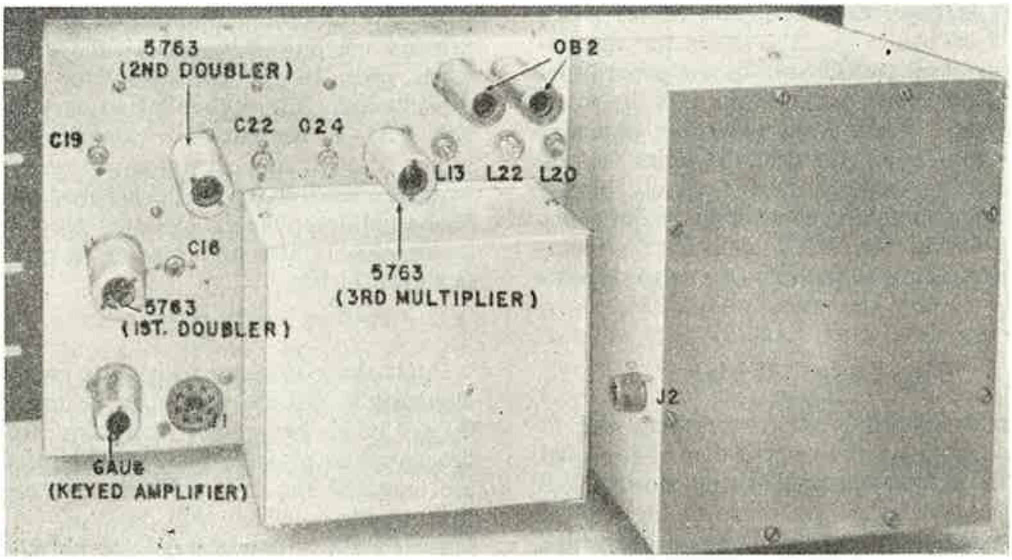


Fig. 5 — Rear view of the transmitter. Complete shielding plus the pi-L tank circuit for the 6146 make this unit a TVI-free transmitter. The small shield box contains the VFO; the final amplifier is housed in the larger box.

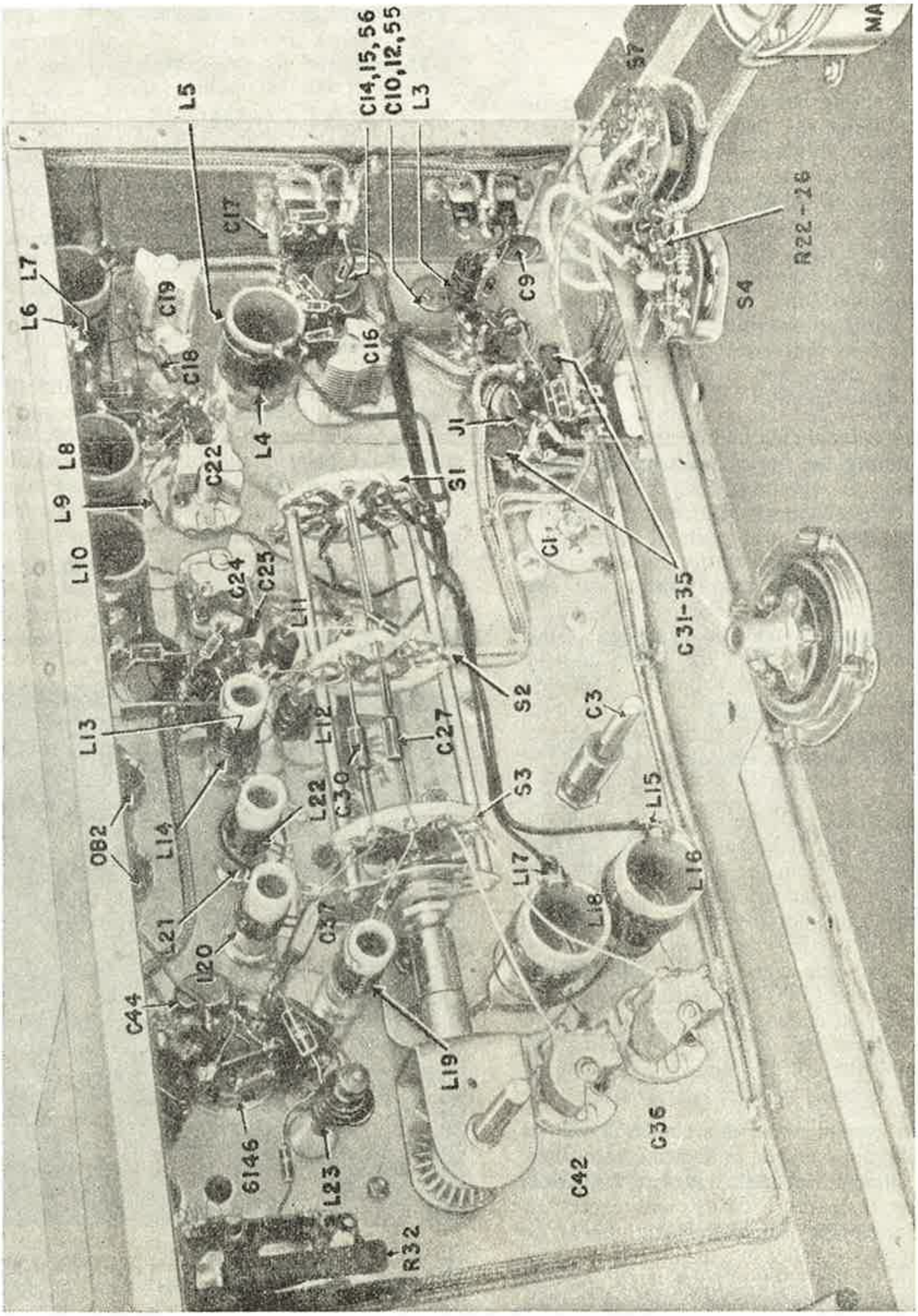


Fig. 6 — Inside view of the chassis. This is a close-up of the same view shown in Fig. 2 of this article. Note that all power and heater leads are wired with shielded wire which is clamped to the chassis at convenient points.

Copper strap is used for rf connections in the final amplifier to reduce inductance and keep spurious resonances at the highest frequency.

Adjustments

After the wiring has been completed, the rig is ready for lining up; the lineup may be done once and then forgotten. Remove all valves except the 12AU7 from their sockets and test the VFO with the shield off. Adjust C1 to set the band edge, and set C2 for minimum capacitance to make certain that the band is covered. Some cutting of L1 may be necessary to make the band fit the dial fully. Put the shield on the VFO, and check to determine whether the VFO can be heard in the receiver. If the shield is tight and the decoupling is done properly, the VFO will not be audible.

A milliammeter should be inserted in the 250-volt lead during the lineup procedure to check plate currents. A high-resistance, dc meter such as the AWA Voltohmyst will be found useful for reading the rectified grid voltage, although a milliammeter wired temporarily in series with the ground end of the grid resistor will also serve the purpose. The connection between the meter and resistor should be by-passed if this latter method is used. With the 6AU6 and the first 5763 in their sockets, about 2 ma of grid current will flow in R9 when the key is down. (Link L5 should have its coupling reduced, and the first doubler tank should be tuned to resonance.)

Insert the 6146; with the plate voltage off and the bandswitch to 7 Mc and set the VFO to current should flow in the 6146 when the grid tank is tuned to resonance. Connect a 1,000-ohm carbon resistor temporarily across L16 and set the VFO to about 3.7 Mc. Slide links L5 and L15 down over the coils slightly and resonate both circuits. The 1,000-ohm resistor reduces the Q of the coupled circuits to a low value, and in so doing, reduces the coefficient of coupling (dependent upon the Q). The undercoupled circuits can be peaked easily without interaction.

The grid current under this condition will be fairly small, but enough to indicate resonance. After the circuits have been tuned to resonance, remove the temporary resistor and check the grid drive over the band. It should have two peaks near the ends of the band and a valley in the centre. If necessary, readjust C16 and C36 slightly so that the drive is fairly uniform over the band. A couple of tries may be necessary to obtain the right coupling between the link and the tuned circuits for the best uniformity of drive. The above procedure should be repeated each time in tuning up.

Next, insert the second doubler valve and turn the bandswitch in the 3.5 Mc position, grid 7.2 Mc. Connect the grid-current meter to R13, and connect the 1,000-ohm resistor across L7. Couple L6 to L7 and resonate the circuit with C19 without touching the adjustment of C16. Removal of the 1,000-ohm resistor should now provide nearly uniform drive to the second doubler over the range of the VFO from 3.5 to 3.72 Mc. Again, the spacing of the links may have to be changed a couple of times to obtain the best results. The grid drive to the final amplifier through L9 may now be adjusted by the same technique, although it will usually be unnecessary to use the 1,000-ohm resistor for the bandwidth required to cover the 7-Mc band. (The bandwidth of the circuit containing L4 and L7 must be broad enough to cover the 28-Mc band, whereas the plate circuit of the second doubler when coupled to the final grid need only cover 7 to 7.3 Mc.) Adjust C22 and C42 to provide uniform drive over the 7-Mc band.

Now. connect the 1,000-ohm resistor across L10 and resonate this circuit with C24 (at 7.2 Mc) with the aid of the grid meter in series with R17. Do no readjust C22 unless it is necessary

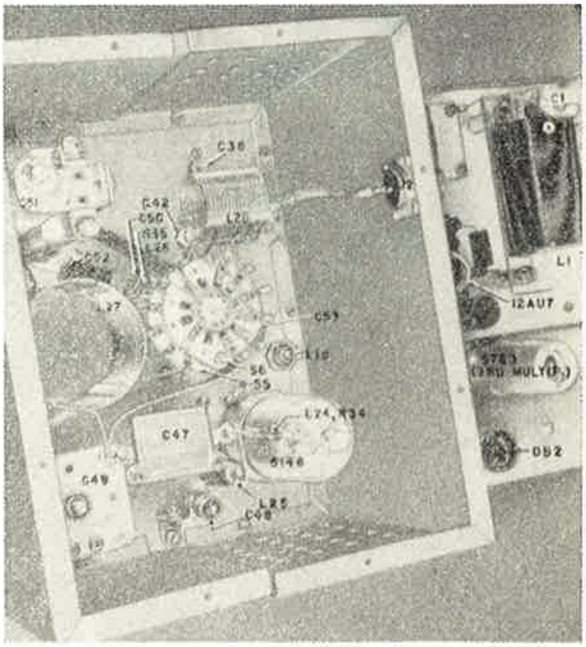


Fig. 7 — Beam-power final. Copper strap is used to reduce lead inductance between the bandswitch and the tuning and loading capacitors. The shield box is perforated above and below the 6146 for adequate ventilation.

Table of Voltages & Currents *(Typical at 7 Mc)

Tube	E_p (volts) Key Down	E_p (volts) Key Up	I_p (ma) Key Down	I_p (ma) Key Up	E_{g2} (volts) Key Down	E_{g2} (volts) Key Up	I_{g2} (ma) Key Down	I_{g2} (ma) Key Up	E_{g1} (volts) Key Down	E_{g1} (volts) Key Up	I_{g1} (ma) Key Down	I_{g1} (ma) Key Up
12AU7	45	45	6**	6**	—	—	—	—	—	—	—	—
6AU6	240	240	7	0	150	265	2.2	0	—	0	—	—
5763	240	240	20	17	145	180	1	1	—	−7	—	—
5763	240	240	18	19	130	170	1.5	1.5	—	−7	—	—
5763	240	220	14	19	160	180	1.0	1.2	—	−7	—	—
6146	600	650	150	10	200	210	15	—	−85	−45	3	0

* Heater voltage: 6.3 v. Supply voltages: 260 v and 600 v. ** Both sections.

in order to make the drive to the third multiplier uniform over the range of the VFO from 3.5 to 3.7 Mc. If C22 has to be readjusted, go back and check the final grid drive on 3.5 Mc to be sure it has not been altered. Remove the resistor again and check the drive to the third multiplier. The location of L10 with respect to L8, as given in the Parts List, should be about correct; however, this spacing may have to be changed slightly if the coils have not been wound exactly as described. L10 should not be closer to L8 than is necessary for the required bandwidth for 28-Mc operation.

The difficult part of the lineup is now over and you may relax. The slugs in L19 and L20 may be adjusted to peak the final grid drive in the centre of the 14- and 21-Mc bands, respectively. The 1,000-ohm resistor loading should be repeated on L22 and L13 to provide uniform drive across the 28-Mc band.

Go back and check the drive on each band and readjust wherever necessary. Then lock all capacitors and slugs. Apply a dab of cement to secure the links to the coils — you will not have to adjust these circuits until you rebuild!

Power may now be applied to the final amplifier. It is best to start at reduced plate voltage with a series resistor in the high-voltage lead. Connect a 50-ohm dummy load to the output jack with a pilot lamp across it. On any band, with C51 at maximum, C49 should be rotated to obtain a dip in plate current. The dip will be more pronounced on the higher frequency bands because the required capacitance for light loading will be less. Decreasing C51 will raise the plate current and the power output. Capacitor C49 should always be tuned for minimum plate current after C51 has been changed or the pi network will not behave correctly for best harmonic reduction. The presence of parasitics can be determined by reducing the fixed bias until the amplifier draws about 10 ma with the key up. Rotate C49 and note whether there are any changes in plate current. If there are, the amplifier is oscillating and the frequency of the parasitic oscillation should be determined with a grid-dip meter or wave-meter. During the design, the addition of L24 and L26 removed the last traces of parasitics and no tendency to oscillate was ever noted at the operating frequency.

TVI Check

With the panel and shields bolted securely, and a shielded dummy load connected to the output, no TVI was encountered with the transmitter on the bench beside a TV receiver protected with a high-pass filter. This test was made 30 miles from the TV transmitter. An inefficient TV antenna was used on the receiver which caused considerable snow on most channels. Removal of the shield from the dummy load produced crosshatching on some channels when the transmitter was operating on 14, 21, or 28 Mc. (The amplifier in this receiver was not in the 21-Mc band!) When the 6146 plate circuit was tuned off resonance, the weak channels were obliterated — dramatic proof that the final tank must always be tuned to resonance. In regions where TV signal strength is low, a low-pass filter may be required to reduce TVI to a minimum.

Antenna Matching

The pi-L network will accommodate slight variations from the 50-ohm antenna impedance it is designed for; if the coaxial line is not reasonably well matched, some trouble may be experienced in loading the final. A standing-wave bridge is invaluable for checking line match, either with direct feed or a line feeding an antenna tuner.

Keying

Very satisfactory keying was obtained without the use of a key-click filter. Because the multiplier stages and the final are not over-biased, no appreciable squaring of the wave shape results and the keying is clean, but not hard. If a softer note is desired, some filtering may be used provided that the cathode resistor of the 6AU6 is altered to take into account any resistance in the filter. The bias on the 6AU6 should be kept between 1 and 1.5 volts.

Modulation

The usual precautions in modulating any tetrode amplifier apply to this transmitter. The screen and plate are modulated together — about 40 watts of audio should be available. The use of a fixed screen supply for the 6146 is not recommended for phone operation.

A Few Afterthoughts

After the conclusion of such a projcet it is natural to wonder what possible improvements could have been made, given the benefit of hindsight. Among these afterthoughts might be included the following:

(1) Bandspreading of the VFO to make the narrow bands easier to tune.

(2) Substitution of slug-tuned coils and fixed capacitors for the tuning capacitors in the low-frequency stages.

(3) Several changes in mechanical layout to facilitate wiring and improve the appearance. But as one who enjoys rebuilding occasionally, these changes were left for another session.

Acknowledgment

The author wishes to thank Mr. George Grammer, W1DF, for his helpful correspondence on the matter of harmonic response and Q of the pi network, and Mr. George D. Hanchett, Jr., W2YM, for his encouragement and many helpful suggestions.

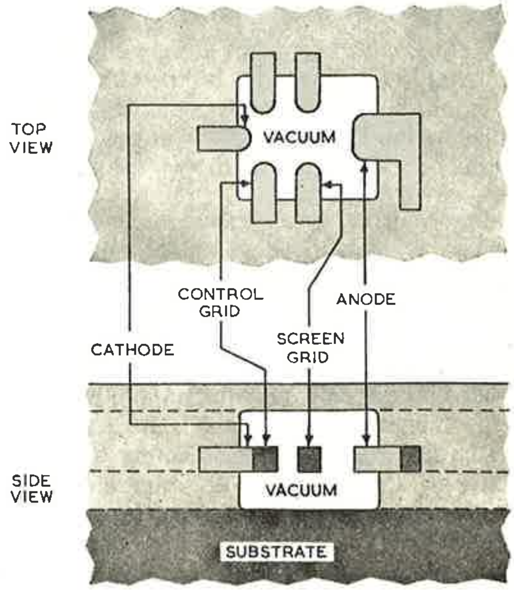
(With acknowledgements to RCA)

MICRON-SIZED VALVES

One of the editor's reference books defines a micron as 1×10^{-6} metre, or 3.937×10^{-5} inches. Or, translated again, it is about 4 one hundred thousandths of an inch.

Even after some of the astounding developments of recent years the reader may well wonder what this has to do with valves. Well it has this to do with them — K. R. Shoulders, of the Stanford Research Institute, Menlo Park, California has produced a paper entitled "On Microelectronic Components, Interconnections and System Fabrication," in which he proposes a construction method for valves which might be described as very small indeed.

The paper in question was delivered at a computer conference held in the U.S.A., and represents a feasible and original approach to the



problems of components for microminiature data processing systems. Mr. Shoulders has made a critical survey of the effects found in the microelectronic region, and has re-examined the electronic effects in ordinary valves. His proposal for valves visualises electron emission in a tiny evacuated bubble in a solid block of material; and consists of the formation of a micron-sized valve in a solid block of material by the addition of suitable electrodes to the evacuated bubble.

At this stage the proposed valve is a simulated device only, but details have been worked out for a new device to suit microelectronic system requirements and fabrication methods. The device is based on the quantum mechanical tunnelling of electrons into a vacuum. It has an estimated switching time of 10^{-10} seconds. The mechanical arrangement appears to be well suited to self-forming manufacturing methods, assuring uniformity. One of the most significant features outside its extremely small size is the immunity it promises to temperature variations over the range from a few degrees Kelvin up into the red-heat area. Other features are negligible transit time lag and the absence of troublesome space-charge effects caused by high fields.

The proposed valve is called the tunnel-effect vacuum tetrode. One proposed configuration is shown in the accompanying diagram, which shows a model intended for use in interconnected arrangements of many valves and associated components, having component densities of up to 10^{11} components per cubic inch.

In order to understand the tunnel-effect tetrode, two points have to be considered. They are firstly obtaining electron emission, and then using that emission. An investigation of cathode properties has shown that the current density can be very high with only very small space-charge effects. The current from this source can be varied over seven powers of 10 by a change in field of 2 to 1 at the cathode, implying high gain possibilities. Furthermore, tests carried out at 10,000 Mc have shown that at that frequency no deviation from the dc emission characteristics can be detected.

The intention with this valve is to form an array of small tips superimposed on the cathode shown in the diagram. These tips would be a few hundred angstrom units in diameter, and would provide emission below 100 volts when applied to the screen grid or plate. The cathode itself has a nominal width of 3,000 angstrom units. ($3{,}000 \times 3.937 \times 10^{-9}$ = approximately 12 millionths of an inch).

The method of using the emitted electrons is dictated by electron-optical considerations. Whilst a positive control grid would normally be used, this would not draw an appreciable current because the grid is located outside the electron path between cathode and plate. Negative grids are possible if they can be provided with a sufficiently smooth surface or have a sufficiently high work function to prevent emission.

If the electrons can be collected by a low-potential plate, this will avoid unnecessary heating. Potentials of just a few volts seem possible because the field strength would be sufficient (over such small distances) to allow the collection of electron current densities of more than 10,000 amperes per square centimetre without troublesome space-charge effects. In this mode of operation the screen grid would remain at about 100 volts positive, helping to cause field emission from the cathode but not collecting an appreciable current.

Because this device is visualized for application in computing equipment and the like, switching speeds are an important consideration. The upper limit in switching speed would be set by the permissible power dissipation. With a plate voltage of 10 volts, a plate current of 100 microamperes, and a capacitance of 10^{-15} farad, this valve would show a switching time constant of 10^{-10} seconds for a power density of 100 kilowatts per square centimetre.

Ionizing radiation does not affect the cathode properties or other electrodes until the dielectric block itself has been severely damaged. The choice of dielectric material is very wide, but materials such as aluminium oxide or beryllium oxide are likely to be chosen because of their stable characteristics under bombardment; they would be used in films to build up the assembly.

VIDEO WAVEFORMS

An oscilloscope can be of great value to a technician if he uses it to advantage in observing waveforms. The process of signal tracing by waveform is of great value in speeding up television receiver service; it provides a means of quickly determining the operational condition of the most complex circuitry.

The most informative waveform encountered in television service work is the composite video signal waveform consisting of the video informaton, the blanking pedestals and the sync pulses. If a technician learns to recognize discrepancies in this waveform, he can quickly check the performance of many of the circuits within the receiver.

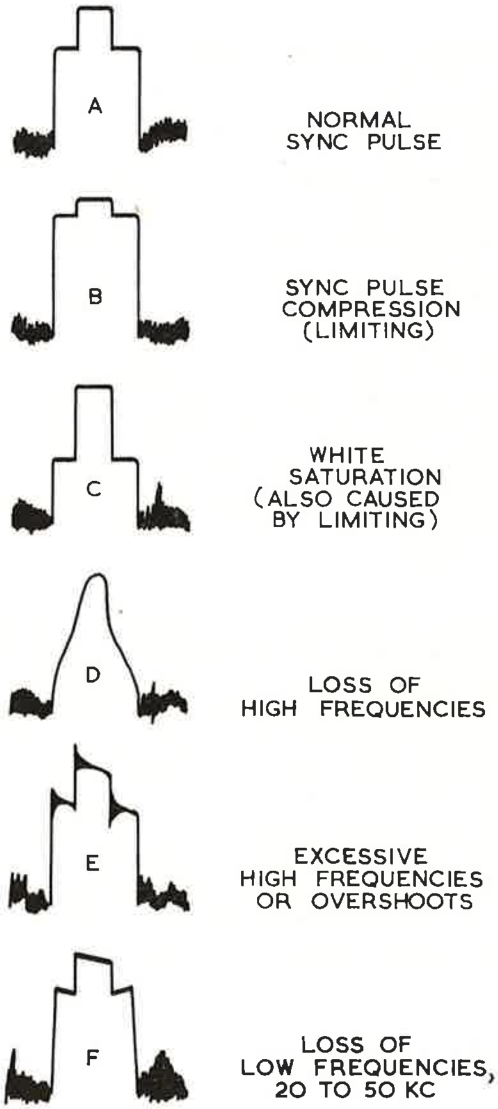
The measurement of the peak-to-peak voltages of these waveforms further reveals the condition of the circuitry; this information is normally supplied in related service information.

To attain reliable information from the observations of composite video waveforms, always use a low-capacity probe and a good oscilloscope. This will assure observation of waveforms as they actually exist in the circuitry. Learn to analyze video waveforms; they can often serve to reveal obscure troubles and save you a great deal of valuable time.

Observation of the horizontal blanking and sync pulse waveform alone can reveal a wealth of information:

Clipping or limiting caused by malfunctioning circuitry can be detected at a glance, as illustrated at B and C. Sync pulses should normally be approximately 25% of the overall signal level.

In many instances, distortion of the horizontal blanking and sync pulse can reveal malfunctioning circuitry affecting the frequency response within the video section of a receiver. Waveform D illustrates the rounding effects of the horizontal blanking and sync pulse caused by a loss of high frequencies and indicates a loss of picture detail. Waveform E shows the effect of excessive high-frequency response on the horizontal blanking and sync pulse which results in fine vertical black-and-white striations following a sharp change in picture shading. The horizontal blanking and sync pulse takes the form shown in F when subjected to a loss of low-frequencies. This would show up on the kinescope of the receiver as a change in shading of large picture areas and a general smear of picture detail.



A Note on NOISE IN AUDIO AMPLIFIERS

H. J. Woll and F. L. Putzrath

(RCA Victor Division)

There are a number of noise sources in audio amplifiers such as:

a. Thermal noise in the input coupling circuit
b. Shot noise
c. Partition noise in pentodes
d. Flicker noise
e. Ballistics and microphonics
f. Pops
g. Hum.

Hum will not be considered here although it is a serious problem and much can be said concerning its elimination. Ballistics, microphonics, and pops are matters of valve design and selection rather than circuit design and will not be considered either.

This note will be primarily concerned with flicker, shot, and partition noise and the effects of the input coupling network. An attempt will be made to outline the requirements of the input circuit and to discuss the magnitude of second stage noise in conventional configurations.

Noise is generated in the resistive component of any impedance by the thermal agitation of the electrons. The magnitude of the thermal noise voltage of a resistance in temperature equilibrium is:

$$e = \sqrt{4kTBR} \text{ volts rms}$$

where

k = Boltzmann's constant
T = temperature in degrees Kelvin
R = resistance in ohms
B = bandwidth in cycles per second.

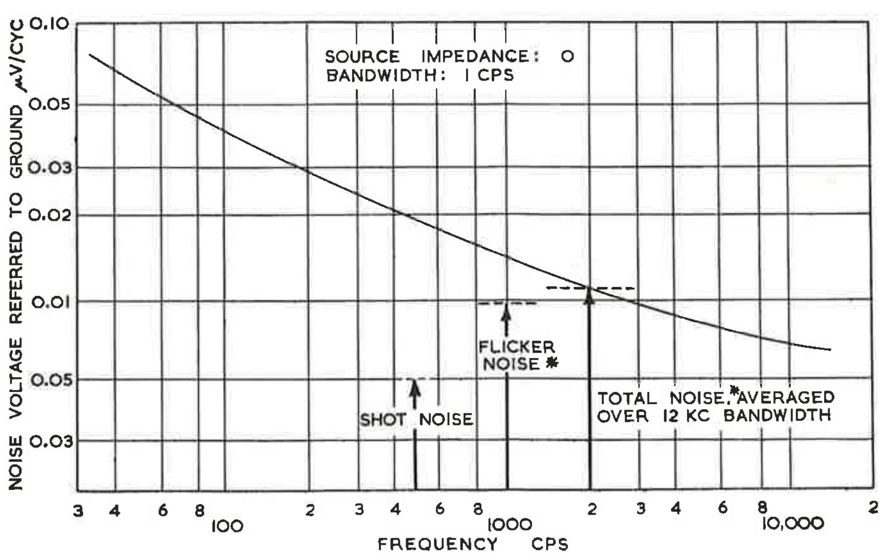


Fig. 1 — Triode Valve Noise Versus Frequency.

A certain signal-to-noise ratio is available from the source and is determined by the signal strength and the magnitude of thermal noise. Using an amplifier, it can be approached but never exceeded. Noise figure is a convenient way of expressing the amount by which an amplifier deteriorates the signal-to-noise ratio available from the source. Noise figure may be defined as the available signal-to-noise ratio at the source divided by the available signal-to-noise ratio at the amplifier output. An equivalent definition is that noise figure is the ratio of the total noise power at the output of the amplifier to component of the noise output power which is due to thermal noise in the source impedance. Noise figure is customarily expressed in db.

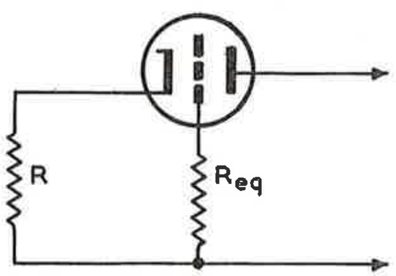


Fig. 2— Grounded-grid Equivalent Circuit.

Shot noise is generated because the plate current of a valve is not continuous, but consists of discrete charges which are numerous enough to approximate very closely a continuous current. Partition noise in pentodes is similar to shot noise but is caused by the division of current between the plate and the screen. Flicker noise is caused by local fluctuation of emissivity of the cathode. These sources of noise may be lumped and their effect duplicated by an equivalent voltage generator. This generator is commonly represented by a resistor, R_{eq}, in series with the grid of the valve and is chosen to be of such a value that the thermal noise voltage generated by it is equal to the sum of the shot, partition, and flicker noise voltages referred to the grid circuit. Thus:

$$R_{eq} = R_{shot} + R_{part} + R_{flicker}$$

Shot noise and partition noise are independent of frequency and their equivalent noise resistances are generally considered to be:

$$R_{shot} = \frac{2.5}{g_m}$$

$$R_{part} = \frac{20\ I_{screen}}{g_m\ I_{cathode}}$$

In the audio frequency band, flicker noise is much greater than either of the above two noise sources. It is a function of frequency and thus its equivalent noise resistance, $R_{flicker}$, is also. For any particular frequency characteristic an integrated $R_{flicker}$ can be found that is a constant.

The noise spectrum of a typical low noise triode with a coated cathode is shown in Fig. 1. In this valve, R_{shot} is 1500 ohms and $R_{flicker}$ integrated over a 12 Kc flat bandwidth is 6000 ohms.

Consider the problem of determining the noise figure of an audio amplifier stage. Identical grounded-grid and grounded-cathode triodes with the same source resistance, R, are shown in Figs. 2 and 3.

The valve noise is referred to the grid circuit and is represented by R_{eq} which is a fictitious generator of voltage $e = \sqrt{4kTBR_{eq}}$. The valves are then considered to be ideal amplifying devices. Since noise figure is:

$$F = \frac{\text{total noise power at the output}}{\text{that component of output noise due to thermal noise in the source}}$$

$$= \frac{4kTBR_s + 4kTBR_{eq}}{4kTBR_s} = 1 + \frac{R_{eq}}{R_s}$$

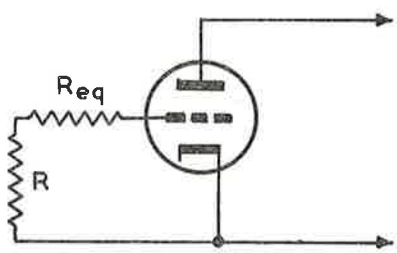


Fig. 3 — Grounded-cathode Equivalent Circuit.

The interesting fact is that the noise figures of the grounded-cathode and grounded-grid stages are identical. Thus, although the input resistance of a grounded-grid stage might be under 500 ohms, a high source resistance of perhaps 30,000 ohms is required to obtain a good noise figure just as in the case of the grounded-cathode connection.

It is to be noted that the above expressions represent **first** stage noise figures. The noise figure of an amplifier represents the deterioration of of signal to noise ratio by **all** the stages in the amplifier. If the first stage gain is high, the succeeding stages do not contribute appreciable noise, and the noise figure of the amplifier is about the

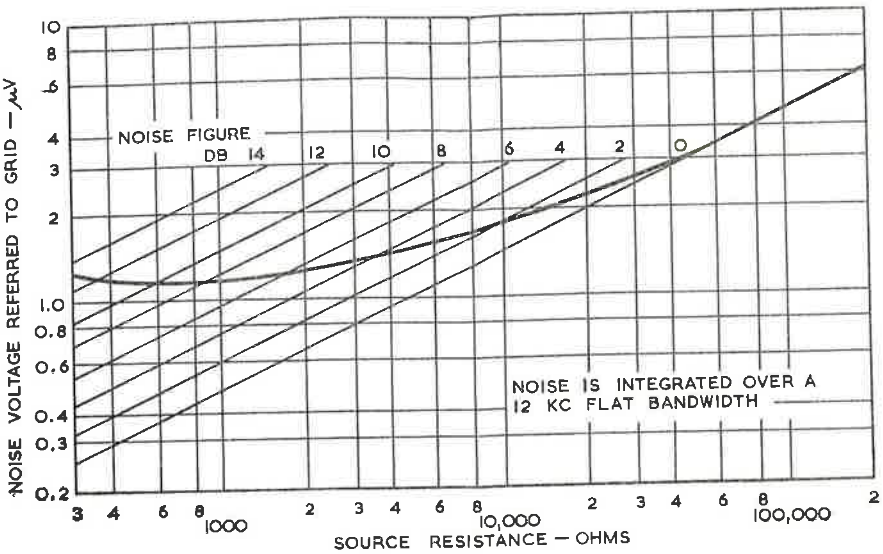


Fig. 4 — Noise Versus Source Resistance.

same as that of the first stage. This is generally the case in audio amplifiers.

On the other hand, an amplifier with a grounded-grid input valve must be operated from a low source resistance to obtain appreciable first stage gain. Hence this amplifier is either operated from a high resistance source and suffers from noise contributed by the second stage, or is operated from a low resistance source and has a poor first stage noise figure, or some combination of the two.

Excepting ballistics, microphonics, hum, and pops, the following general conclusions can be drawn:

Noise figure improves as the source impedance increases and zero db noise figure can be approached in practice. (A low-noise triode with 22,000 ohm source resistance has a 1.0 db noise figure. See Fig. 4.) As a result, efficient input transformers greatly improve the noise figure of a system with a low impedance source such as magnetic pickups and microphones.

If the source impedance is low and bandwidth requirements or other conditions permit, the noise figure can be improved by paralleling input valves, increasing the g_m and thereby lowering R_{eq}.

It is to be noted that noise figure, per se, is meaningless. To express the performance of an amplifier, one must specify the noise figure with a given source resistance.

The noise figure of a stage is independent of the configuration, i.e., whether the stage is grounded-cathode or grounded-grid. Generally the gain of the first stage is high enough so that second and later stage noise make only a small contribution to the total.

Thus it can be generalized that cascaded grounded cathode stages at audio frequencies will give as good or better noise performance than other possible circuit configurations. In addition this configuration is most advantageous from a practical point of view—such as heater and B + supplies.

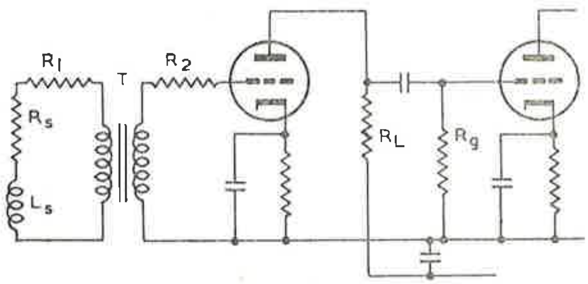


Fig. 5 — Typical Two-stage Amplifier.

APPENDIX

A typical amplifier employing two cascaded grounded cathode stages is shown in Fig. 5. R_s and L_s are respectively the series resistance and inductance of the source. T is an input transformer with a turns ration of **n,** a primary resistance of R_1, and a secondary resistance of R_2.

The equivalent circuit is shown in Fig. 6 where

$$L = n^2 L_s$$
$$R = n^2 R_s$$
$$R_t = n_2 R_1 + R_2$$
$$R_i = \frac{R_g + R_L}{R_g R_L}$$

The equivalent noise resistors of the two stages are represented by R_{eq1} and R_{eq2} respectively.

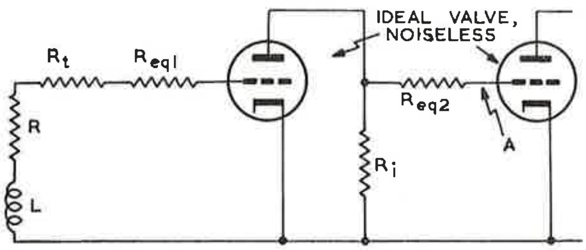


Fig. 6 — Equivalent Circuit of Two-stage Amplifier.

Examination of the noise figure of this amplifier will be made at point A. The noise figure at the input of the first ideal valve is:

$$F_1 = \frac{4kTB\,(R + R_t + R_{eq1})}{4kTBR} = 1 + \frac{R_t}{R} + \frac{R_{eq1}}{R}$$

The noise voltage due to the interstage coupling circuitry and the second valve is:

$$E_1 = \sqrt{4kTB}\sqrt{\left(\sqrt{R_i}\frac{R_p}{R_p + R_i}\right)^2 + \left(\sqrt{R_{eq2}}\right)^2}$$

$$= \sqrt{4kTB}\sqrt{\frac{R_i}{\left(1 + \frac{R_i}{R_p}\right)^2} + R_{eq2}}$$

where R_p is the plate resistance of the first valve. Dividing the above expression by the first stage voltage gain, G, this noise voltage is referred to the amplifier input so that the ratio of the ideal to actual noise powers due to the source and due to the interstage circuitry is:

$$F_2 = \frac{\dfrac{R_i}{\left(1 + \frac{R_i}{R_p}\right)^2} + R_{eq2}}{G^2 R_s}$$

The over-all noise figure of the two stages up to point A is then:

$$F = F_1 + F_2 = 1 + \frac{R_t}{R} + \frac{R_{eq1}}{R} + \frac{1}{G^2}\,\frac{\dfrac{R_i}{\left(1 + \frac{R_i}{R_p}\right)^2} + R_{eq2}}{R}$$

The above formula might be construed to mean that it would be desirable to let the source impedance have as high a resistive component as possible. However, an increase in source resistance must be accompanied by a corresponding increase in signal voltage, as is realized with an input transformer or with a "high impedance" magnetic playback head.

A typical amplifier using a 12AY7 twin triode might have the following constants:

$$n = 28.3$$
$$L_s = 2 \text{ mh}$$
$$R_s = 1 \text{ ohm}$$
$$R_1 = 8 \text{ ohms}$$
$$R_2 = 6000 \text{ ohms}$$

$$R_{shot_1} = R_{shot_2} = \frac{2.5}{1660 \times 10^{-6}} = 1500 \text{ ohms}$$

$$R_{flicker_1} = R_{flicker_2} = 6000 \text{ ohms}$$

$$\left.\begin{array}{l} R_L = 100{,}000 \\ R_g = 1{,}000{,}000 \end{array}\right\} R_i = 100{,}000 \text{ ohms.}$$

$R_p = 25{,}000$ ohms.

$G = 30$

then:

$$R_{eq1} = R_{eq2} = 7500 \text{ ohms}$$
$$n^2 = 800$$
$$L = 1.6 \text{ henry}$$
$$R = 800 \text{ ohms}$$
$$R_t = 12{,}400 \text{ ohms}$$

and

$$F = 1 + 15.5 + 9.4 + 0.02 = 25.9$$

or an equivalent ratio of 14.1 db.

From the above example it can be seen that the noise in this system is over 14 db worse than that which could have been obtained with an ideal amplifying device. A substantial noise contribution is made by the resistance in the input transformer, yet a negligible amount is contributed by the inter-stage coupling network and the second valve.

A preferred arrangement that would eliminate the noise contribution of the input transformer could be obtained by the use of a "high impedance" head.

Thus:

$$n = 1$$
$$L_s = L = 1.6 \text{ henry}$$
$$R_s = R = 800 \text{ ohms}$$
$$R_1 = R_2 = R_t = 0 \text{ ohms}$$

Other values as above

Under these conditions

$$F = 1 + 9.4 + 0.02 = 10.4$$

or an equivalent ratio of 10.2 db.

Even here, the second stage noise contribution is negligibly small.

(With acknowledgements to the Institution of Radio Engineers Inc. and RCA)

CALCULATING DECIBELS PER OCTAVE

The rate of attenuation, or of boosting, is usually given in the form of so many decibels per octave. In some cases the frequencies at which readings are taken do not conveniently cover an exact number of octaves. In such a case the procedure below may be adopted.

(a) To convert db/specified frequency ratio to db/octave.

Frequency ratio	Multiply db/specified ratio by factor to give db/octave
1.2:1	6.02
1.25:1	3.10
1.33:1	2.43
1.5:1	1.71
2:1	1.00
3:1	0.63
4:1	0.50
5:1	0.43
6:1	0.39
7:1	0.36
8:1	0.33
10:1	0.30

Example: A change of 0.7 db occurs with an increase of frequency from 1000 to 1250 cps. What is the rate of change in db/octave?

Rate of change $= 0.7 \times 3.10 = 2.17$ db/octave.

(b) To convert db/octave to db/specified frequency ratio.

Frequency ratio	Multiply db/octave by factor to give db/specified frequency ratio.
1.2:1	0.263
1.25:1	0.322
1.33:1	0.412
1.5:1	0.585
2:1	1.00
3:1	1.59
4:1	2.00
5:1	2.33
6:1	2.59
7:1	2.81
8:1	3.00
10:1	3.33

Example: What is the change in level for a frequency ratio of 1.5 to 1 when the rate of change is 6 db per octave?

Change in level $= 0.585 \times 6 = 3.51$ db.

INDEX, 1960

AWV