RADIOTRONICS

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By the way—

Our front cover illustrates a stage in the manufacture of an oil diffusion pump. The part being formed on the glass lathe is the condenser.

This type of pump makes it possible to achieve the high degree of vacuum necessary for the modern miniature valve.

———

Stocks of the new Radiotron Characteristics Chart are now available. For the convenience of readers an application form is enclosed with this issue. As completed forms are received at this office, charts will be despatched by return mail.

———

Presented in this issue is a four valve dual-wave receiver developed in the A.W.V. Circuit Design Laboratory. Several interesting features such as harmonic mixing make this circuit well worth study.

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Subscribers are informed that February, March and April, 1951, issues are now out of print.

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We would like to thank Mr. Marsland, of Camberwell, Victoria, and Mr. Riordan, of Glenelg, South Australia, for sending us some of the Radiotronics back issues we needed to complete our files. We still require numbers 52, 55, 57 and 62, together with the copies of Proc. I.R.E. referred to in our April issue, should any of our readers be able to help us.

CONTENTS

Circuit Laboratory Report   **Number 8**

Four-Valve Dual-Wave Battery Receiver

Four-valve battery-operated receivers, especially dual-wave types, are not commonly manufactured in large numbers, and one of the objects of this design is to show that performance adequate for the needs of the majority of country listeners can be obtained with a four-valve receiver. A feature of the design is the use of second-harmonic mixing with the oscillator operating on the low-frequency side of the signal on the short-wave band.

From the performance figures it will be seen that the sensitivity of the receiver is appreciably better than 10 μV (3.8 – 6.2 μV) all over the broadcast band and it is doubtful whether an increase in broadcast gain would result in better station-getting ability. On the short-wave band the sensitivity is less (33 – 65 μV), but still adequate for the reception of most stations that provide a signal strong enough to over-ride static and other noise. Few commercial r-f stage five-valve battery receivers give much more than twice the short-wave sensitivity of the RA41, a difference which it would not be easy to detect on the air.

Converter operation

The main advantage of harmonic mixing is that by its use it is possible to obtain a large difference between the resonant frequencies of the circuits connected to the oscillator grid and control grid of the converter, with a consequent reduction in the interference between these circuits. When the oscillator fundamental is used for mixing and the intermediate frequency is 455 Kc/s the separation between oscillator and signal frequencies is only $2\frac{1}{2}\%$ at 18 Mc/s so that the impedance of the signal-frequency circuit to oscillator-frequency voltages is high. For this reason, oscillator-frequency voltages appear at the control grid and with the 1R5 it is usually necessary to reduce them by applying an out-of-phase voltage through a neutralizing capacitor to obtain satisfactory converter operation.

With harmonic mixing the neutralizing capacitor is not needed and the stability of the oscillator is such that for small signal-circuit trimmer adjustments it is not necessary to rock the tuning control of the receiver when aligning at 18 Mc/s.

Considerable positive oscillator drive is needed for harmonic mixing and tests with a number of valves showed that maximum sensitivity at 6 Mc/s required the use of a 20,000 ohm grid leak and grid current as high as could be obtained without restricting the tuning range at the high-frequency end of the band. The oscillator coil described below gave a grid current of 250 μA in 20,000 ohms at 6 Mc/s, but at high frequencies the oscillator voltage

was excessive, causing a loss of sensitivity at 18 Mc/s. A 500 ohm grid stopper reduced the grid current at higher frequencies to a satisfactory value, the size of the stopper being a compromise between the requirements at 18 Mc/s and those in the middle of the tuning range.

Despite precautions taken to minimize coupling between oscillator and control-grid circuits, some oscillator voltage always appears on the control grid, at least on the short-wave range, and this voltage, depending on its phase, adds to or subtracts from the modulation of the electron stream by the oscillator grid voltage. With converters having oscillator voltage on the first grid, as is the case with the 1R5, the phase of the oscillator voltage is correct for increasing the degree of modulation, and thus improving the sensitivity, when the oscillator voltage on the signal grid is at a lower frequency than the signal circuit. In this receiver, the oscillator voltage which appears on the signal grid is at the frequency of the second harmonic of the oscillator.

Advantage has been taken of this phasing effect to improve the sensitivity, image ratio and signal-to-noise ratio of the RA41. Accordingly, the second harmonic of the oscillator on the short-wave band is maintained 455 Kc/s below the signal frequency. This arrangement necessitates padding the signal frequency circuit instead of the oscillator circuit, as shown in Fig. 1 and while it increases the difficulties of obtaining satisfactory oscillator grid current for a given short-wave signal frequency coverage, the oscillator coil described below provides adequate frequency coverage and grid current at the low frequency end of the 5.9 – 18.4 Mc/s tuning range.

The improvement in short-wave sensitivity obtained by operating the oscillator on the low-frequency side of the signal frequency is greatest at the high frequency end of the band (sensitivity is almost doubled at 18 Mc/s) and decreases at lower frequencies as the percentage of frequency difference between oscillator and signal frequencies increases.

Neutralized i-f amplifier

The i-f transformers used in this receiver are a high-impedance type tuned to 455 Kc/s by 85 $\mu\mu$F, and having an average unloaded Q of 91. The resulting gain from the i-f stage is sufficiently high to cause excessive regeneration due to feedback through the grid-plate capacitance of the 1T4 unless the stage is neutralized. A conventional neutralizing circuit is used, the comparatively large capacitor (9 $\mu\mu$F) being required because of the relatively high (0.01 $\mu\mu$F maximum) grid-plate capacitance of the 1T4. The value of the neutralizing capacitor is not critical, but will vary with different layouts. If the a.v.c. bypass is altered in size the neutralizing
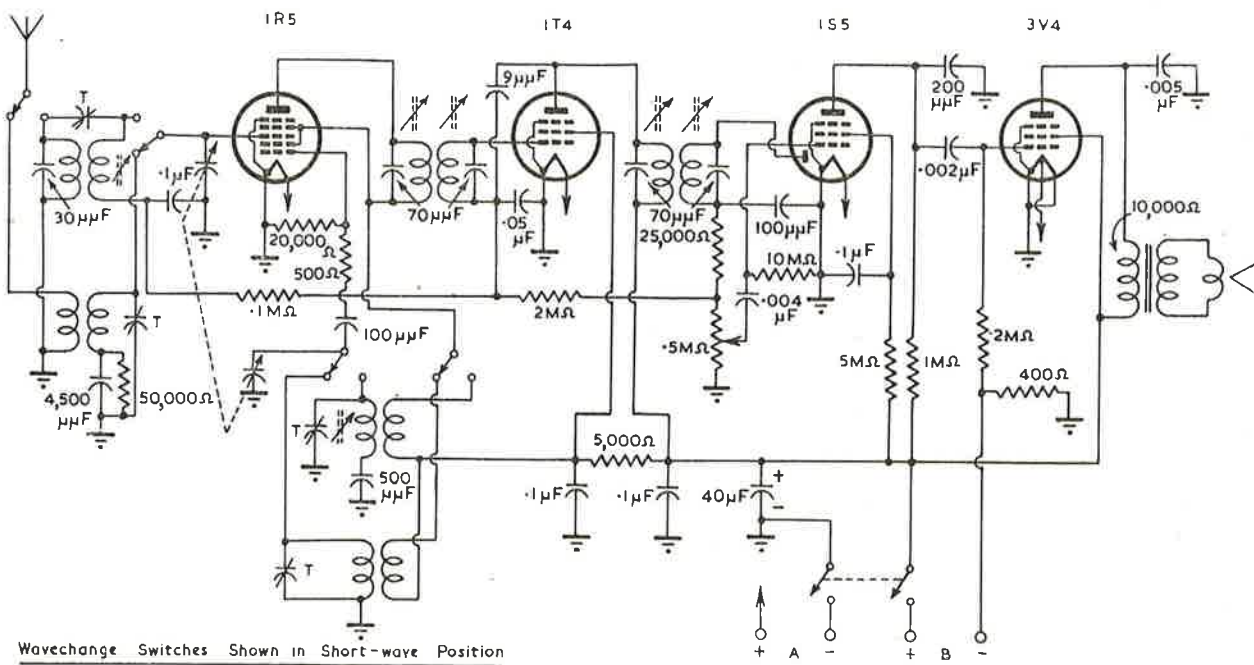
May, 1951

Fig. 1. *Circuit of Radiotron Receiver RA41.*

capacitor will need to be varied in the same ratio. However, it should not be made too large or difficulty may be experienced in peaking the plate trimmer of the second i-f transformer.

Circuit details

Although the RA41 receiver has been designed to obtain maximum sensitivity from the four valves used, it is comparatively economical in the use of components. The broadcast aerial trimmer is connected between aerial and control grid of the 1R5 which saves the usual small capacitor in this position and at the same time increases the aerial coil gain and signal-to-noise ratio, particularly at the high-frequency end of the band, although at the expense of signal-circuit selectivity and, consequently, image ratio.

Both the screen and the plate of the 1R5 are used as the oscillator plate in order to obtain maximum oscillator grid current for a given B supply drain. Although it was noticed that a decrease in oscillator grid current occurred at 10.3 Mc/s due to self-resonance of the i-f transformer winding this effect was not considered important as the grid current fell only to 250 μA and the sensitivity was unchanged. Past experience has shown that this effect does not vary greatly with different i-f transformer characteristics. However, the extent of the grid current reduction should be checked in any particular design.

It is not possible to dispense with the decoupling of the a.v.c. line between converter and i-f amplifier because of the i-f neutralizing circuit. In the absence of the decoupling components the voltage fed back

to the cold end of the first i-f secondary would also be applied to the converter grid and would then give rise to instability.

There is little a.v.c. control on the receiver output for some 15 db after the theoretical output passes the overload point of the output valve in Fig. 2. This might be considered to indicate excessive a-f gain. However, as the sensitivity on the short-wave band is much lower than that on the broadcast band the existing design is adopted as the best compromise.

The 50,000 ohm resistor in parallel with the short-wave padder is used to provide a d.c. return from grid to ground as a.v.c. is removed from the 1R5 on short waves. This is done to obtain maximum frequency stability with changes in signal strength and at 18 Mc/s a variation of signal input from 30 μV to 30 mV produces a frequency shift of less than 5 Kc/s. This shift decreases at lower frequencies. Although the i-f amplifier is the only stage controlled on short waves, the receiver will still handle signal inputs up to 0.1 volt satisfactorily.

The adequate frequency stability of the oscillator is also demonstrated by the fact that with any 30% modulated 18 Mc/s signal input up to 100 mV there is no sign of flutter until the volume control is turned up well past the a-f overload point. To obtain this degree of frequency stability it is necessary to cut the low-frequency response of the receiver sharply below the useful part of the a-f spectrum, which is the reason for the small a-f coupling capacitors, but it is not necessary to use a separate electrolytic capacitor to stabilize the oscillator plate supply voltage.

During the development of the receiver some i-f regeneration was noticed in the filament circuit. This could be removed with a large (0.5 μF) bypass from the filament string to ground, but it was also found that with the filament circuit wired as shown in Fig. 1 and with the positive filament lead from each valve returned to the point at which the positive lead from the A battery entered the chassis, no traces of regeneration from this cause could be detected. No other wiring in the chassis has unusual requirements, but wiring capacitance between grid and plate circuits of the 1T4 should be kept as low as possible.

The comparatively restricted power output of the receiver (200 mW in the primary of the transformer for 13% distortion) is due to the overbiased operating condition of the 3V4. Increased output can be obtained, at the expense of greater B battery drain, by reducing the value of the back-bias resistor so that the recommended bias of −5 volts is applied to the 3V4.

With new valves, the receiver's performance when the A & B battery voltages are reduced to 1.1 volts per cell (a B supply of 66 volts) indicates that good battery life from batteries having normal load voltage discharge characteristics should be obtained before the performance becomes unacceptable or the receiver fails to operate.

Components

The i-f transformers are permeability tuned, hav-ing characteristics as given above, and are representative of good commercial practice.

The broadcast aerial coil is also permeability tuned, the particular coil being a Vega type VC1 with a secondary Q of 125 at 600 Kc/s and 120 at 1400 Kc/s. The 30 $\mu\mu$F capacitor in parallel with the primary tunes it below the intermediate frequency with no aerial attached so that instability due to some particular size of aerial tuning the primary to the intermediate frequency is avoided.

The only requirements for the broadcast oscillator coil are that it should cover the tuning range and provide the required grid current with a minimum of oscillator voltage applied to the screen of the 1R5. For this reason small primaries should be used, coupled as closely as possible to the secondary. In this respect most commercial oscillator coils are satisfactory.

The short wave aerial coil is air-cored and uses an externally wound high-impedance primary of 20 turns of 36 B & S S.S.E. wire. This is a close-wound solenoid spaced 1/16th inch from the 9½ turn secondary. The secondary is wound with 22 B & S enamelled wire at 24 t.p.i. on a ¾″ former.

The short wave oscillator coil is also an air-cored solenoid and consists of 23 turns of 24 B & S wound at 32 t.p.i. on a ¾″ former. The primary is 9 turns of 36 B & S interwound with, and starting ¾ of a turn out from, the cold end of the secondary.
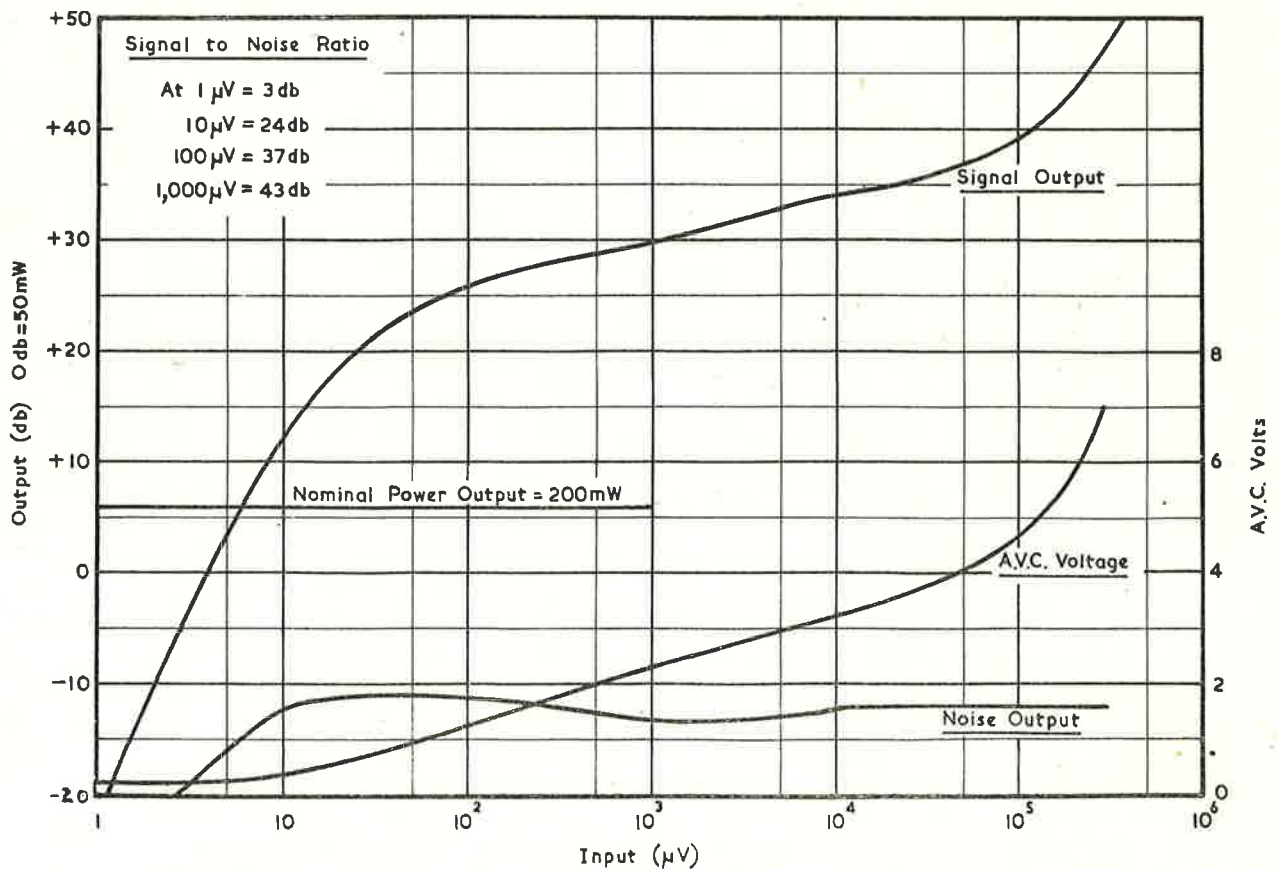

Fig. 2. *Curves of signal and noise output and a.v.c. voltage vs. signal input voltage at 1 Mc/s.*

TEST RESULTS

1. Sensitivity (for 50 mW output)

		c/s	I-F(Kc/s)	Broadcast(Kc/s)			Short Wave(Mc/s)		
Frequency		400	455	600	1000	1400	6	12	18
3V4 grid	(V)	1.45							
1S5 grid	(mV)	21							
1S5 diode	(mV)		240						
1T4 grid	(mV)		2.3						
1R5 grid	(μV)		40	52	51	51	92	95	92
Aerial	(μV)			6.2	4.2	3.8	65	45	33
Image ratio	(db)			43	33	25	16	8	8
Noise ratio	*(db)			22	24	25			
$I_{osc.}$ grid	(μA)			250	320	330	230	390	280

* Signal-to-noise ratio at 10 μV input. This figure is quoted because the high signal-to-noise ratio and moderate sensitivity lead to inaccuracy in the reading of the usual noise sensitivity.

2. Selectivity at 1000 Kc/s.

Times down		2 (6 db)		10 (20 db)		10^3 (60 db)
Bandwidth	(Kc/s)	7½		19		58

3. Distortion vs. output

(5 mV input at 1000 Kc/s, modulated 30% at 400 c/s)

Output	(mW)	20	50	100	150	200
Distortion	(%)	1.8	3.8	5.5	7.7	13

4. Distortion vs. signal input

(20 mW output; 1000 Kc/s input 30% modulated at 400 c/s)

Input	(mV)	0.1	1.0	10	30	100
Distortion	(%)	2.0	1.4	3.0	3.4	5.7

5. Distortion vs. modulation depth

(20 mW output; 5 mV input at 1000 Kc/s, modulated at 400 c/s)

Modulation	(%)	20	40	60	80	90	100
Distortion	(%)	1.6	2.0	2.5	3.5	4.6	7.2

6. Frequency response

(B.F.O. connected across diode load. 0 db = 25 mA in 3 ohm voice coil)

Frequency	(c/s)	50	75	100	200	400	1000	2000	3000
Output	(db)	−7	−4	−3	+1	0	−2	−5	−11

7. Slump performance

(A and B battery voltages reduced to 1.1 and 66 volts respectively)

Frequency	(Mc/s)	1.0	6.0
Aerial sensitivity	(μV)	9.0	250
Oscillator grid current	(μA)	175	115

A-F output:—100 mW for 13% distortion.

8. Voltage and current analysis

No signal input; all voltages measured to chassis with 500 ohm/volt meter (Avometer, model 7) on 400 volt range unless otherwise indicated.

1R5		Plate	Screen	Bias
Voltage	(V)	53	53	0
Current	(mA)	1.8	3.3	—
1T4				
Voltage	(V)	84	53	0
Current	(mA)	2.0	0.8	—
1S5				
Voltage	(V)	29*	9*	0
Current	(mA)	0.055	0.015	—
3V4				
Voltage	(V)	81	84	−6.0*
Current	(mA)	5.0	1.3	—

* Measured with V.T.V.M.   Total battery drain: A — 250 mA, B — 14½ mA.

The Operation of Phototubes

The behaviour of a phototube for given operating conditions can be predicted from the anode characteristics of the tube. These characteristics, which correspond to the plate family of an amplifier valve, show the relation between anode current and anode voltage for different values of light input. It is the purpose of this Note to show how the anode characteristics of a phototube can be used to predict performance under given operating conditions.

The solid-line curves of Fig. 1 are the anode characteristics of a typical vacuum phototube having a caesium-oxide coated cathode. When a small amount of inert gas, such as argon, is admitted to the tube, the anode characteristics of the phototube change to those shown by the dashed-line curves of Fig. 1. The dashed-line curves are typical of gas phototubes. These two sets of curves will be used in this Note to illustrate the performance of gas and vacuum phototubes.

Steady-light operation

Fig 2A is a typical phototube circuit in which the output voltage appears across resistor R_L. When light falls on the cathode of the phototube, a current flows through R_L and the phototube; at any instant, the sum of the voltage drops across R_L and the phototube equals E, the applied voltage. Hence, the voltage across R_L and the voltage across the phototube for any value of light input can be determined by the intersection of a load line and the anode characteristic of interest. For example, when R_L = 10 megohms, E = 90 volts, and F = 0.04 lumen, the voltage across R_L is 25 volts for the gas phototube and 8 volts for the vacuum phototube. When R_L = 45 megohms, the output voltage is 56 volts for the gas type and 37 volts for the vacuum type. Hence, as R_L is increased, the output voltage of the vacuum type approaches that of the gas type. The effect of load resistance on output voltage for a given value of light input is shown in Fig. 3.

The circuit of Fig 2A is suitable for applications in which the d.c. output voltage feeds a voltage amplifier which, in turn, actuates a relay. For this type of application, the gas phototube is more sensitive than the vacuum type with the same B-supply voltage and load. However, the sensitivity of gas phototubes changes with age, applied voltage, and values of light input. Because of these factors, circuits for gas phototubes should not be critical to reasonable changes in sensitivity. In some applica-

tions, these changes in sensitivity can be compensated by an adjustment of the gain following the phototube. In the event that readjustment of the gain is not desirable, a vacuum phototube used with sufficient amplification to give the desired overall sensitivity will prove more stable.


PHOTOTUBE ANODE CHARACTERISTICS

CURVE	DESCRIPTION
——	VACUUM TYPE
- - -	GAS TYPE

FIG.I
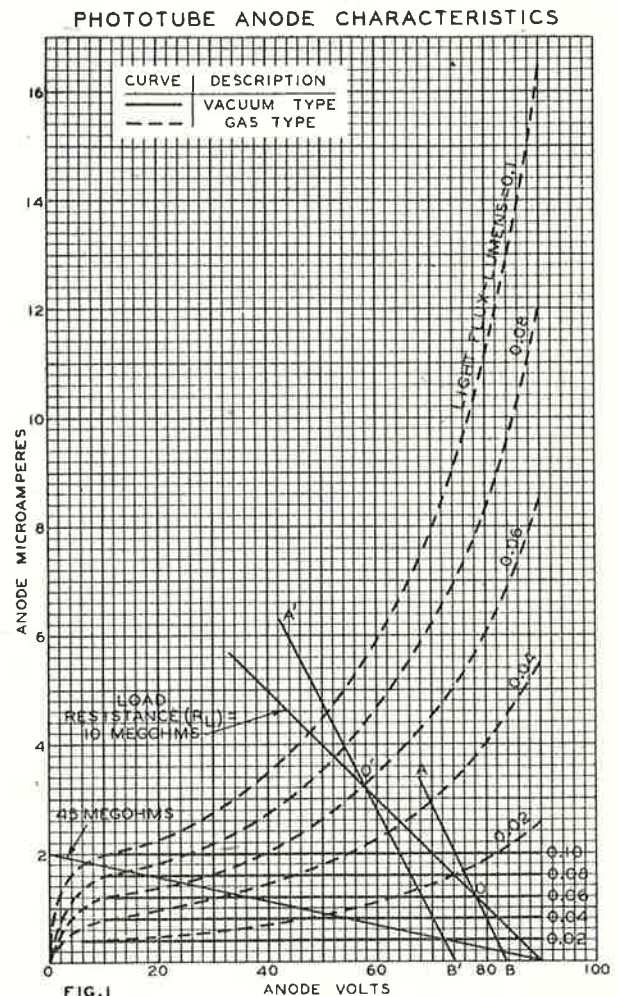
Modulated-light operation

Many phototube circuits depend on modulated light for their operation. In a sound-motion-picture projector, for example, the amplifier following the phototube responds only to the modulated component of light input. In such applications, the criteria for sensitivity are not the same as those for steady-light operation.

PHOTOTUBE OPERATING CHARACTERISTICS

FIG. 2A

FIG. 2B

FIG. 3
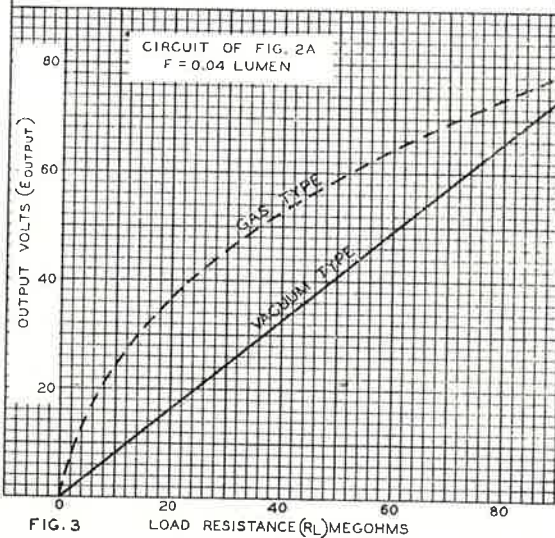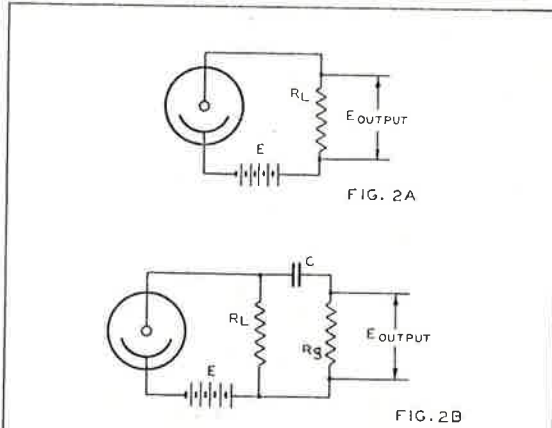
In circuits which depend on steady-light input for operation, phototube sensitivity is simply $S = I_a/F$, where I_a is the anode current in microamperes and F is the light flux in lumens received by the cathode. In circuits which depend on modulated-light input for operation, phototube sensitivity is defined as

$$S_d = \frac{dI_a}{dF}$$

S is the static sensitivity and S_d is the variational sensitivity of the phototube. Variational sensitivity is analogous to the transconductance (g_m) of an amplifier valve. Because the output of the phototube usually feeds a voltage-operated amplifier, it is important to know the voltage sensitivity of the phototube and its associated circuit. Voltage sensitivity (S_v) in this Note is defined as the ratio of the alternating voltage output to the alternating light-flux input. In symbols,

$$S_v = \frac{dE_a}{dF}$$

where E_a is the output voltage in volts. Now, the action of the circuit of Fig. 2A is analogous to that of an amplifier valve: the cathode is a source

of electrons and the anode collects these electrons; the varying light input on the cathode is analogous to an alternating voltage applied to the grid of an amplifier valve. Therefore, the alternating output voltage of a phototube is

$$dE_a = dF \quad . \quad S_d \frac{r_p R_L}{r_p + R_L}$$

where r_p is the variational resistance in ohms of the phototube and is equal to the slope (dE/dI) of the anode characteristic at the operating point. Since voltage sensitivity is the output voltage per unit of light-flux input,

$$S_v = S_d \frac{r_p R_L}{r_p + R_L}$$

The physical interpretation of this last equation is important. The internal resistance of the phototube shunts the load resistance R_L; the change in current due to a change in light flux causes a voltage drop across the parallel combination of r_p and R_L. Thus, the output voltage may be low, even though S_d is high, because the internal resistance of the phototube reduces the generated voltage.

The internal resistance of a vacuum phototube is very high, while that of a gas phototube is low over a large portion of its operating range. The value of r_p at a point on any anode characteristic can be determined by measuring the slope at the point of interest. S_d is constant for vacuum phototubes over their operating range, but is not constant for gas phototubes. For this reason, it is desirable to calculate the performance of phototubes by graphical methods. However, it should be noted that for vacuum phototubes the output voltage can be calculated with fair accuracy by the relation

$$E_a = F S_d R_L$$

where F is the alternating component of the light input in lumens and E_a is the alternating voltage output in volts.

Consider the typical phototube circuit of Fig. 2B. The load is R_L for steady-light input and is the parallel combination of R_L and R_g for alternating-light input, provided the reactance of C is negligible at the lowest frequency of interest. To predict the operation of the phototube when the light input is modulated, draw the load line R_L from the B-supply voltage, as shown in Fig. 1. For a steady-light input of 0.06 lumen, the operating point is 0 for the vacuum-type phototube and is 0' for the gas-type phototube.

A convenient value of $R_L = 10$ megohms is assumed. When $R_g = R_L$ and the amplitude of the sinusoidal light input is constant, load lines AB and A'B' represent the operating lines for the vacuum and gas phototubes, respectively. Now, when the steady-light input is modulated $66\frac{2}{3}\%$, for example, the peak value of light input is 0.1 lumen and the minimum value is 0.02 lumen; the corresponding changes in voltage across R_g are (67.5 − 51) = 16.5 volts for the gas phototube and

(82 − 74) == 8 volts for the vacuum type. These values are the total changes in output voltage; the peak values of the fundamental components are approximately half these values.

For a sinusoidal variation in light input, the distortion is negligible for the vacuum phototube and is appreciable for the gas phototube. The second-harmonic distortion may be calculated from the relation

Per cent. Second Harmonic =

$$\frac{(I_{max} + I_{min} - 2\,I_o)}{2(I_{max} - I_{min})} \times 100$$

where I_{max} and I_{min} are the maximum and minimum instantaneous values of current, respectively, and I_o is the current at the operating point. Substituting the values obtained from Fig. 1 for the gas phototube, we have

Per Cent. Second Harmonic =

$$\frac{4.6 + 1.3 - (2 \times 3.2)}{2(4.6 - 1.3)} \times 100$$

$$= -7.5\% \text{ (approx.)}.$$

The sign of a harmonic indicates its phase; the negative sign in this example signifies that the average value of the anode current is less with modulation than without modulation. This simple analysis shows that for chosen values of R_L, R_g, and E, and with the same B-supply voltage, the gas-type phototube furnishes about twice as much output as the vacuum type; however, the output of the gas type may contain too much distortion for some purposes.

A comparison of the anode characteristics of the two types of phototubes shows that for very large values of load resistances corresponding to the light flux on the cathode, the performance of both types is identical, because the sensitivity of the gas phototube approaches that of the vacuum phototube at low values of anode voltage. When R_g is infinite, the calculations for output voltage and distortion are made along the load line corresponding only to R_L.

The recommended maximum anode voltage for a gas phototube is 90 volts. When the anode voltage rises above 90 volts, a glow discharge takes place and the active emitting surface of the cathode sputters off. Thus, the peak value of the maximum alternating output voltage from a gas phototube is limited to a little less than 90/2, or 45 volts.

The recommended maximum d.c. anode-supply voltage for a vacuum phototube is 500 volts. In order to obtain the maximum voltage output from a vacuum phototube supplied with modulated light, it is necessary to adjust the anode voltage under static conditions to a value approximately one-half of the maximum d.c. supply voltage. This adjustment permits the modulated voltage output to have a peak value of nearly 250 volts.

From this discussion, it is seen that when the respective maximum anode voltages are applied to each type of phototube, and when the value of the load

on the vacuum phototube is increased until the minimum instantaneous anode voltage (E_{min}) is the same for both types, the voltage sensitivity of the vacuum phototube can be much higher than that of the gas type. In the limiting case when maximum output voltage is obtained from each type for the same value of light input, the voltage sensitivity of the vacuum phototube is approximately 250/90, or 2.7 times that of the gas phototube.

SIGNAL-TO-HISS RATIOS OF TYPICAL PHOTOTUBES

CIRCUIT OF FIG. 2A
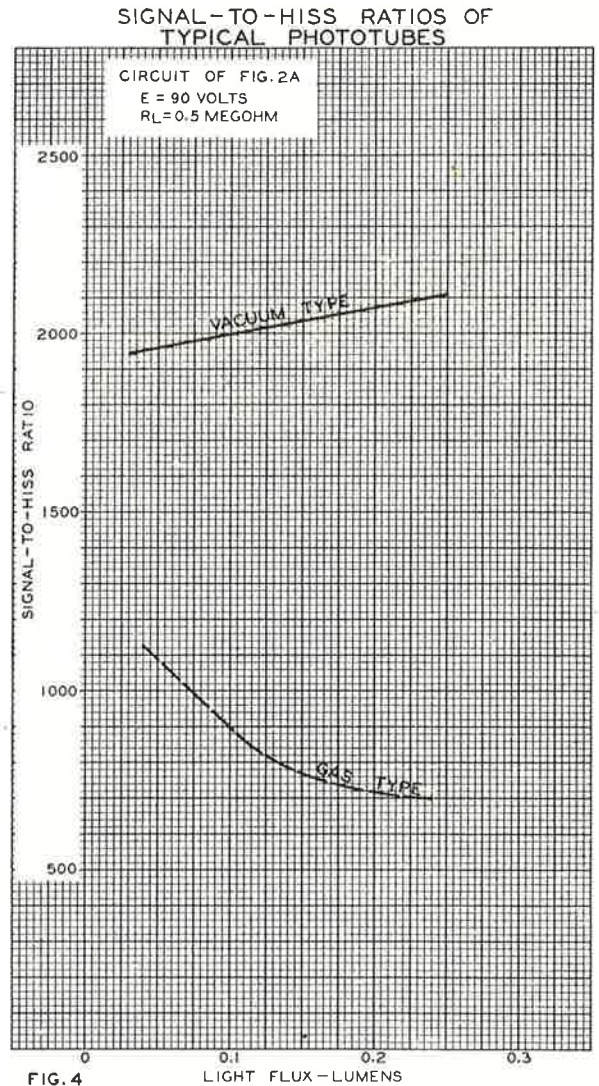E = 90 VOLTS
RL = 0.5 MEGOHM

FIG. 4

The static and variational sensitivities of gas phototubes vary with age, temperature, light-flux, and anode voltage. In applications where changes in sensitivity necessitate readjustments of circuit conditions, consideration should be given to the use of vacuum phototubes. It is easy to compensate for the comparatively low gain of vacuum phototubes under certain operating conditions by increasing the gain of the succeeding amplifier.

A good frequency characteristic is desirable in many cases. When a vacuum phototube is used,

the anode-cathode capacitance of the phototube and the equivalent shunt capacitance of the associated circuit determine the high-frequency response characteristic. When a gas phototube is used, the time necessary to deionize the gas is also a factor in determining high-frequency response. The relatitve magnitudes of the effects of capacitance and gas on high-frequency response depend on the physical placement of the components and their electrical characteristics.

Hiss output

The absolute value of hiss output is in general not as important as the signal-to-hiss ratio. The data in Fig. 4 show the relation between signal-to-hiss ratio and light flux for typical vacuum and gas phototubes. When it is desirable to have a large signal-to-hiss ratio, the use of a vacuum phototube may be preferable.

Determination of light flux

Radio engineers are frequently unfamiliar with the units used in connection with lamps and illumination, and are often hazy regarding the proper approach to the design of equipment incorporating a phototube. This article will assist in approaching the design with the minimum of wasted energy.

Illumination is the application of light for a particular purpose. The light may be sunlight, or more often artificial light usually from an electric lamp. It is necessary to measure light intensity and **the unit of light intensity** is the "candle". For instance, an electric lamp may be classified as a "50 candle-power lamp".

The unit of illumination is the "foot-candle" which is the illumination on a surface one foot distant from a lamp having a light intensity of one candle-power. This unit of illumination is used in connection with the determination of satisfactory lighting in offices and factories, and minimum values of illumination have been specified for particular purposes, such as general office work, general factory illumination, illumination for fine and delicate work, etc., all these being specified in terms of foot-candles.

The unit of light flux is the "lumen" which is the flux over the surface of 1 sq. foot when it is illuminated at the level of one foot-candle. Light flux can be regarded as being the light emitted by a lamp, and the total flux is constant irrespective of the distance from the lamp.

The total flux emitted by a lamp of one candle power is therefore 4π lumens, since the surface area of a sphere is $4\pi r^2$. We may therefore put down the following relationships for ready reference.

A point source of one candle power emits a flux of 4π lumens, 1 foot-candle is the illumination intensity of 1 lumen per square foot.

Lamps

The flux emitted by electric lamps depends on many factors including the type of lamp, the size and the voltage. Ordinary 240 volt lamps of the coiled-coil variety have approximately * the following characteristics:—

Watts	Lumens	Lumens/Watt
40	415	10.4
60	700	11.7
100	1340	13.4

The column "watts" indicates the electrical power consumed by the lamp. The column "lumens" indicates the total light flux emitted by the lamp, while the "lumens/watt" column gives an indication of the efficiency, which is shown to increase with the size of the lamp.

If a small object (such as the cathode of a phototube) is placed so as to intercept some of the light flux, the flux which will fall on it is given the expression.

$$\frac{A}{4\pi D^2} \times F$$

where A = area of object in square feet.
D = distance from lamp to object in feet,
and F = light flux in lumens emitted by the source.

As an example take as the "object" a phototube type 922 with a window area of 0.4 sq. inch, placed 2 feet from a 100 watt lamp.
A = 0.4/144 sq. ft.
D = 2 feet.
F = 1340 lumens.

The flux which will fall on the cathode is therefore

$$\frac{0.4 \times 1340}{4 \times \pi \times 4 \times 144} = 0.074 \text{ lumen.}$$

The following table will be found useful in connection with lamps and phototubes. It holds for 240-volt coiled-coil lamps, and the phototube is assumed to have a window area of 0.4 square inch. The flux on larger or smaller cathodes will be in proportion to the window area:—

Distance from Lamp (feet)	LAMP 40 / 415	60 / 700	100 / 1340	watts lumens
	Light flux intercepted by phototube (0.4 in².)			
1	0.092	0.156	0.296	
2	0.023	0.039	0.074	
4	0.0057	0.0097	0.018	
8	0.0014	0.0024	0.0046	
12	0.00064	0.0011	0.0021	
16	0.00036	0.0006	0.0012	
24	0.00016	0.00027	0.0005	
32	0.00009	0.00015	0.0003	

If a lens is used to focus the light on to the cathode of the phototube, the area of the lens should be used in the calculation instead of the area of the cathode (it being assumed that none of the light focused by the lens is lost).

*There are slight differences between the published figures of lamp manufacturers.

Restricted Range SPEECH AMPLIFIER

Audio Amplifier
Designed Expressly for
Speech Work
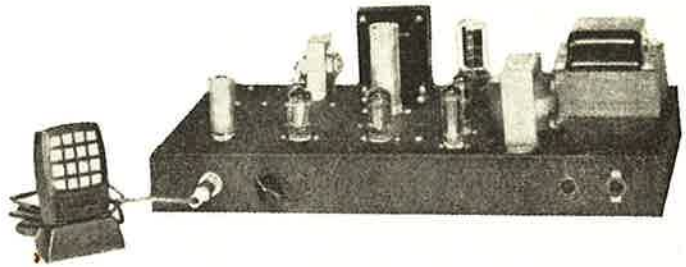
FEATURES —

"Speech" range from 500-2500 cycles   Six miniature tubes

Minimum distortion   Power output of 10 watts

General considerations

A speech amplifier for amateur radio service has the job of amplifying the human voice until the complex waveform which forms the human voice has sufficient power to drive the modulator tubes. The amplifier's job, then, is relatively simple. However, the frequency characteristic of the audio amplifier—that is, the amount of amplification which will be obtained at various audio frequencies—determines to a large degree the type of radio-frequency signal which is put on the air.

For example, if you are using a speech amplifier capable of amplifying frequencies beyond ten thousand cycles, and your voice (or extraneous background noise) contains energy at this frequency then the radio-frequency signal from your transmitter will extend out at least ten thousand cycles —10 kilocycles—on each side of your transmitted radio frequency. Stated another way, your signal has a minimum width of 20 kilocycles. Broad? Quite broad. Even aside from the fact that you have a broad signal, there is little point in transmitting a high fidelity signal. Primarily this is because the average communication receiver does not have an audio system capable of reproducing these high frequencies. In addition, a highly selective receiver will further restrict the audio frequency characteristic.

If you use another amplifier which has practically no gain at 10,000 cycles, but which drops rapidly in gain past 5,000 cycles, then this same voice, using this amplifier, will modulate the radio-frequency carrier so that energy exists out 5 kilocycles each side of the centre frequency. This gives

a signal with a width of 10 kilocycles. By using this second amplifier have you lost naturalness, does your voice sound exactly the same to the amateur receiving it over the air as it would if he heard you in person? No. Can you be understood? Yes.

How far can this process be carried? How much can we restrict the bandwidth of the speech amplifier, and still have voice modulation which is adequate for communication purposes? While it is impossible to give an answer to this question which will satisfy everyone, most engineers agree that a bandwidth, for understandable speech, of 500 to 2500 cycles is adequate. This is not as narrow a band as might be imagined. For example, the major radio networks send their programmes to their member stations on telephone lines. The best of these lines have a cutoff frequency of approximately 5000 cycles. Certainly we do not think of network broadcasts as having "poor quality", and yet 5000 cycles (approximately) is the highest audio tone which will be heard when listening to network programmes.

The primary advantage in using a speech amplifier which has a restricted high-frequency response is that the radio-frequency signal resulting will occupy less space in the spectrum. This is because the radio-frequency bandwidth of a properly operated transmitter is dependent only upon the range of the audio frequencies used to modulate the transmitter. This assumes that the transmitter is free of parasitics, is operating on only one frequency, and the modulation applied is within the modulation capability of the modulated stage, to cite a few of the effects which may give a broad signal, even though the modulating frequencies are within the proper range..

Thus far we have discussed primarily the higher-
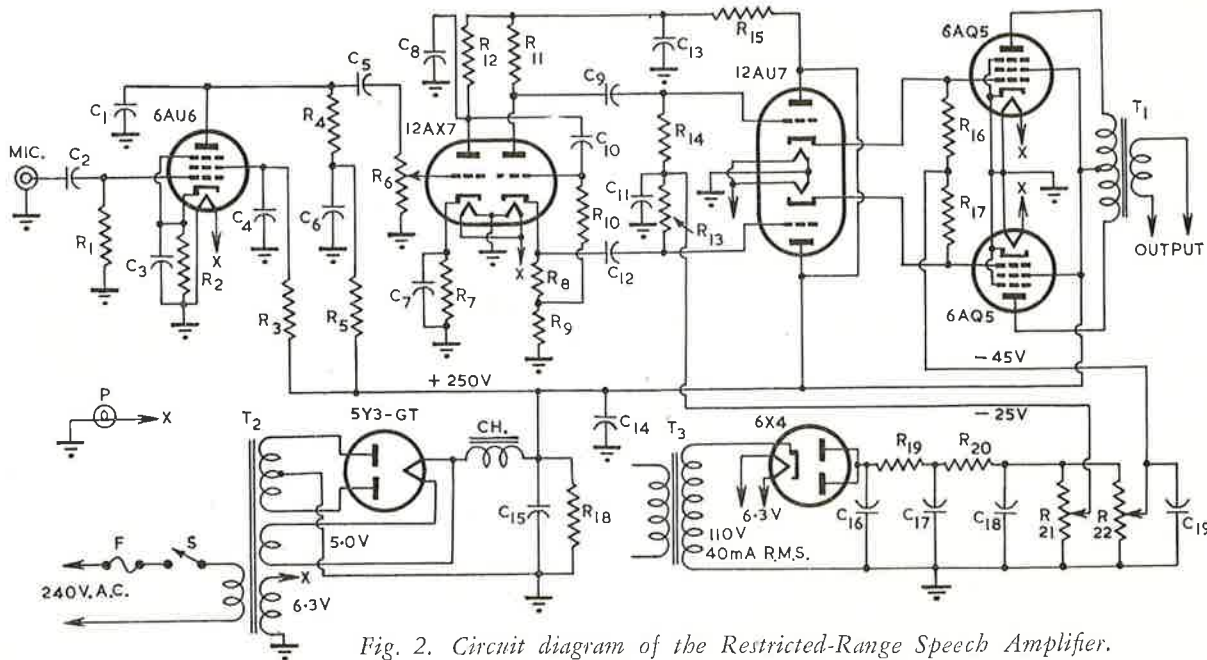
Fig. 2. Circuit diagram of the Restricted-Range Speech Amplifier.

CIRCUIT CONSTANTS

C_1 .... 1590 $\mu\mu$F mica (see text)
C_2 .... 0.005 μF 600 volt paper
C_3, C_7 .... 50 μF 50 volt electrolytic
C_4, C_6, C_{13}, C_{14} .... 8 μF 450 volt electrolytic
C_5 .... 1275 $\mu\mu$F mica (see text)
C_8 .... 1380 $\mu\mu$F mica (see text)
C_9, C_{12} .... 0.01 μF 600 volt paper
C_{10} .... 580 $\mu\mu$F mica (see text)
C_{11} .... 0.1 μF 600 volt paper
C_{15} .... 40 μF 450 volt electrolytic
C_{16}, C_{17}, C_{18}, C_{19} .... 40 μF 150 volt electrolytic
CH .... 8 henry smoothing choke, 150 mils
F .... 1 ampere fuse
P .... 6.3 volt pilot light
R_1 .... 1 megohm, $\frac{1}{2}$ watt
R_2 .... 5500 ohm, 1 watt
R_3 .... 0.47 megohm, 1 watt
R_4 .... 47,000 ohm, $\frac{1}{2}$ watt
R_5 .... 10,000 ohm, 1 watt

R_6 .... 0.25 megohm potentiometer
R_7 .... 3500 ohm, 1 watt
R_8 .... 2500 ohm, $\frac{1}{2}$ watt
R_9 .... 17,500 ohm, $\frac{1}{2}$ watt
R_{10} .... 0.1 megohm, 1 watt
R_{11} .... 20,000 ohm, 1 watt
R_{12} .... 0.33 megohm, 1 watt
R_{13}, R_{14} .... 0.5 megohm, $\frac{1}{2}$ watt
R_{15} .. 6600 ohm, 6 watt (three 20,000 ohm, 2 watt resistors in parallel)
R_{16}, R_{17} .... 6000 ohm, 1 watt
R_{18} .... 50,000 ohm, 10 watt
R_{19}, R_{20} .... 1000 ohm, 2 watt
R_{21}, R_{22} .... 10,000 ohm, 10 watt semi-adjustable
S .... SPST toggle switch
T_1 .... Output transformer (see text)
T_2 .. Power transformer, 350–0–350 at 150 mils, 6.3 volts at 4 amperes, 5.0 volts at 3 amperes

frequency audio tones. However, it is also desirable to eliminate, or attenuate, the low frequency audio response of speech amplifier. Elimination of all response below, say, 500 cycles, would have no effect on the width of our radio-frequency signal, but it would give us the effect of a stronger signal. It is difficult to put an actual number on the gain which could be achieved, but with relatively simple attenuation means used in the speech amplifier a gain of 5 to 6 decibels would be possible. This is the sort of gain which can be expected from a good two-element parasitically excited beam, or by increasing your power by a factor of four.

The energy output of the male voice is concentrated at the lower frequency end of the audio frequency spectrum. Unfortunately these low frequency components of the male voice contribute little to the intelligence in speech. However, being of high amplitude, a great deal of modulator power and modulation capability is required to transmit them. Obviously we can increase the effective transmitted power by reducing the number of low frequency components in the system. Paradoxically a system with restricted high-frequency response, such as discussed previously, sounds more natural if the low-frequency components are attenuated in a balanced manner.

There are many ways to accomplish the desired attenuation of the lower and higher frequency portions of the audio-frequency spectrum. All of these

methods use an audio network, either simple or complex, which will attenuate certain frequencies either more or less than other frequencies. The amount of attenuation achieved will depend upon the type of network used.

The attenuation achieved in the amplifier about to be described is shown in Fig. 3. This attenuation averages 12 db per octave. Stated another way, the power is down by a factor of sixteen for each octave considered. For example, the power output of the speech amplifier at 10,000 cycles is one-sixteenth of the power output at 5000 cycles.

Referring again to Fig. 3, the calculated operating range of the speech amplifier is from 500 to 2500 cycles. Note that the curve is not flat over this portion, but that the 500 and 2500 cycle points are approximately 6db down from the midpoint frequency, which is approximately 1000 cycles. For the first octave below 500 cycles and the first octave above 2500 cycles, the attenuation has not yet reached a slope of 12 db per octave. However, for further octave jumps the attenuation will be quite close to 12 db per octave, so that the 125 and 10,000 cycle points will be down by 26 db and the 62 and 20,000 cycle points down by 38 db.

Note this 62 cycle point. The attenuation at this point is theoretically 38 db or, as actually measured in the speech amplifier, 35 db. This means that the power output at 62 cycles will be only one four-thousandth of the power output at 1000 cycles. This means that normal precautions regarding sixty cycle hum need not be taken. As a result, the filament wires in this speech amplifier were neither paired and twisted nor carefully handled. One side of each filament connection was grounded and the other lead run as a single wire. While this may not seem startling, those of you who have had trouble with hum in high-gain amplifiers will appreciate this statement.

C_5 and C_{10} (see Fig. 2) have the job of attenuating the low frequency end of the audio spectrum, and C_1 and C_8 handle the attenuation of the higher audio frequencies. In other words, the entire job is handled by the proper choice of four condensers,

two of which would normally be employed in the amplifier even if a restricted bandwidth were not desired.

Electrical details

Referring to the circuit diagram, Fig. 2, the tube functions are as follows: The 6AU6 serves as a pentode voltage amplifier, giving a mid-band gain of well over 100. The first section of the high-mu double-triode 12AX7 serves as the second voltage amplifier, and gives a gain of approximately 50. The second section of the same tube acts as a phase inverter. The 12AU7 tube is a push-pull cathode follower which acts as a low-impedance driving source for the push-pull 6AQ5 output tubes. It is absolutely essential that distortion be held to as low a value as possible if full advantage is made of the restricted bandwidth of this speech amplifier. This is because distortion will cause the radio-frequency signal to become broad, and this is one of the effects that we wished to overcome by restricting the audio bandwidth.

One of the major causes of distortion in the audio systems of amateur transmitters is the use of driver stages with too high an internal impedance to properly drive class AB_1 or Class B stages. Distortion results, in this case, because of poor regulation in the driving voltage when the driver is called upon to supply the grid current drawn during voltage peaks. The 12AU7 cathode follower tube acts as a low-impedance driver. This permits more power output from the 6AQ5 tubes with less distortion than would be possible if the 6AQ5 tubes were driven directly from the phase inverter.

Essentially the 6AQ5 tubes are operated as class AB_1 amplifiers. Normally this means that no precautions need be taken with the driver stage to ensure minimum distortion provided that the grids are never driven positive. This condition is difficult to achieve unless the average level is kept quite low. By using a driver which presents a low source impedance, which the 12AU7 accomplishes, the average level may be pushed up quite high and
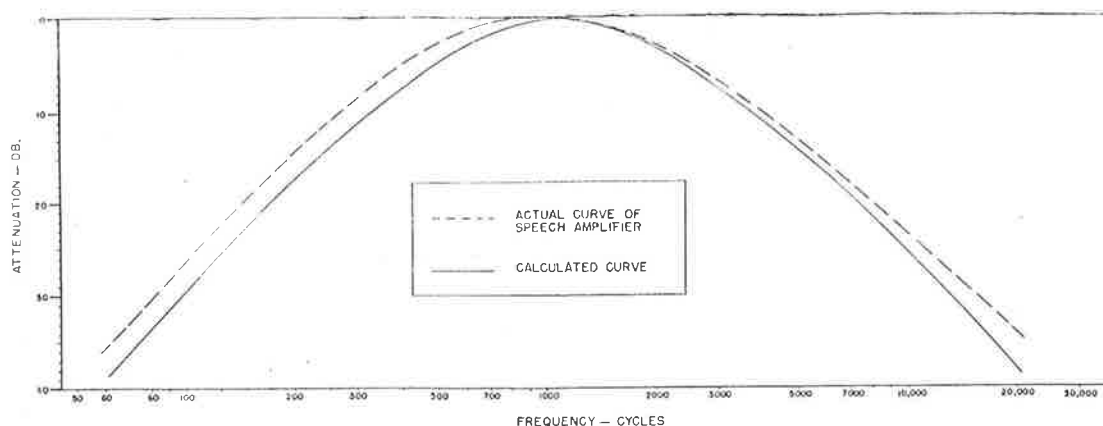

Fig. 3. Theoretical and Actual Frequency Response Curves for the Restricted-range Speech Amplifier.

the 6AQ5 tubes driven all the way up to the grid bias point. Even if an occasional voice peak causes this voltage point to be exceeded, no distortion will occur due to "folding-up" of the driver stage. The net result is high output, minimum distortion, and a "narrow" radio-frequency signal.

Condensers C_1, C_5, C_8 and C_{10} (the frequency controlling condensers) are listed in "Circuit Constants" with values which are not stock values. The values shown are those which calculations indicate to be correct. Try to obtain condensers moderately close to these values. It is not wise to trust the values marked on condensers, and it is recommended that a capacitance bridge be borrowed to check through your stock of mica condensers. It may be easier to parallel condensers in order to get the proper value. For example, C_1 could easily be made up with a 1000 $\mu\mu$F and a 600 $\mu\mu$F condenser in parallel.

One further point might be made, with reference again to the circuit diagram. Fixed bias is supplied to two stages, the 12AU7 stage and the 6AQ5 stage. The bias supply is unusual in one respect. Cathode current for the 12AU7 stage must flow through R_{22}. The total current for both 12AU7 sections is approximately 10 mils. In other words, this bias supply must be capable of supplying a voltage and a current, instead of just a voltage as in the usual case. If the circuit diagram is followed no difficulty will be encountered. However, if you attempt to use another source of bias, make certain that it can supply the required current.
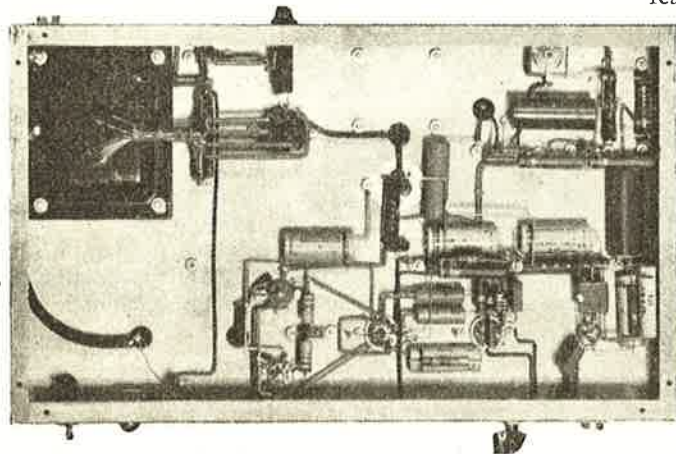

Fig. 4. Under-chassis of the Restricted-range Speech Amplifier.

Mechanical details

The amplifier was constructed on a 17 by 10 by 2 inch chassis. However, inasmuch as practically any layout scheme will work, the prospective builder can use any convenient size chassis and change the layout to suit. The entire speech amplifier and power supply could fit easily in a chassis of half the size of the one just mentioned.

The placement of parts can best be seen in Fig. 1. The tubes are, from left to right, 6AU6, 12AX7, 12AU7, 6AQ5's, and the 5Y3-GT rectifier

tube in the rear. Note that the 6AU6 uses a shield. On the rear of the chassis, at the left, is the bias transformer, T_3, with the choke and C_{15} to the right. The power transformer occupies the rear corner and the output transformer is directly ahead of it.

Only two controls are employed—the on-off switch and the gain control. The microphone input jack and the pilot light are mounted on the front of the chassis and the fuse on the rear of the chassis.

The underchassis view of the amplifier, Fig. 4, indicates the placement of the remainder of the components. No shielded wire was used, mainly because all leads to the first two stages were short. If the layout is altered from that indicated, it might be advisable to shield any long leads in the first two or three stages.

Operating adjustments

Once the amplifier has been completed, and it has been established that voltage can be supplied without any smoking, the 12AU7 bias voltage and cathode-return voltage should be adjusted. R_{21} should be adjusted so that the bias, as read from the arm of R_{21} to ground, is 25 volts. Adjust R_{22} until the voltage from the arm of R_{22} to ground is 45 volts. Next check the bias on the 6AQ5 tubes by reading the voltage from pin 1 of either tube to ground. This voltage should be 15 volts. If this is not true, change the tap on R_{21} slightly until the 6AQ5 bias (pin 1 to ground) reads 15 volts. The 45 volt cathode-return voltage should remain unchanged during this adjustment. It will not be necessary to have any input signal to the speech amplifier during the foregoing tests.

The last check to be made, assuming that the amplifier has been correctly wired and is operable, is to match the 6AQ5 tubes to their load. The amount of power that these tubes can deliver will depend to a great degree upon the output transformer. The selection of an output transformer will be governed primarily by the proposed application. It is recommended, however, that a transformer with a number of various impedances be used, so that minor changes in matching may be made.

It is further recommended that a transformer be purchased which has a generous power rating. For example, a 10 watt transformer will serve, but a 20 watt output transformer will permit more output to be achieved without distortion. The amplifier pictured uses a 10 watt output transformer. The highest output power which could be achieved without discernible distortion on an oscilloscope was 7.2 watts (measured output from the transformer). A second amplifier with an 18 watt output transformer permitted an output, under the same conditions, of 11.2 watts. In both cases the trans-

former was matched to the output load impedance, which took the form of a resistor.

Therefore, procure a transformer which is capable of matching from an appropriate impedance of 10,000 ohms (the plate-to-plate effective load resistance of the 6AQ5 tubes) to whatever class B grids you wish to drive. Or, you may wish to match 10,000 ohms to a 500 ohm line. In the latter case another transformer is required to match from the 500 ohm line to the modulator grids. If this system is used, approximately twice as much

power is lost between the driver plates and the modulator grids as compared to the case where only one transformer is used. You may expect to get losses up to 3 db in each transformer. Three db is two-to-one in power.

Once the transformer is procured and the amplifier tested while driving the required load, it may be advisable to make small changes in the impedance ratio between driver and modulator to ensure that you have an impedance match which will give maximum power transfer with minimum distortion.

New RCA Releases

The sensitivity of the **Radiotron type 1P28** — the multiplier phototube which responds to radiant energy in the ultra-violet region down to about 2000 angstroms — has been increased 4 times over that previously shown for this type.

Additional improvements include a reduction in ultra-violet equivalent noise input at 25°C to 6×10^{-15} watt — a value showing 2.5 times improvement, and a 2-to-1 reduction in d.c. anode dark current.

With its improved features, the 1P28 is especially suited for use in scientific research and specialized applications involving very low ultra-violet radiation levels. Such applications include use of the 1P28 in spectrophotometers and scintillation counters.

Radiotron type 1X2-A is a double-ended, 9-pin miniature type of half-wave vacuum rectifier tube used in high-voltage, low-current applications such as the rectifier in a high-voltage, r-f-operated power supply, or as a rectifier in television receivers to supply the high d.c. voltage for the anode of the picture tube. The 1X2-A has a maximum peak inverse plate voltage of 18,000 volts as compared with 15,000 volts for the older 1X2 which is superseded by the new A-version.

Radiotron type 4E27A/5-125B is a filament type of power pentode suitable for use in r-f power amplifier and oscillator service as well as in class B a-f amplifier and modulator service. It has a maximum plate dissipation of 125 watts in unmodulated class C service. Operation at maximum rated plate voltage and plate input is permissible at frequencies up to 75 megacycles per second.

Having high power gain, the 4E27A is capable of delivering large power output with low driving power. The low value of grid No. 1-to-plate capa-

citance—0.08 $\mu\mu$f—permits stable operation without neutralization in most applications.

The 4E27A does not require forced air cooling, but unrestricted air circulation through the ventilated shell of the 7-pin base and around the glass envelope is required. At maximum plate dissipation, the plate shows a red color.

The 4E27A, although having higher power capability, is unilaterally interchangeable with any of the following tubes: 4E27, 257, 257B, and 8001.

Radiotron type 50C6-G is a beam power amplifier of the glass-octal type used in the output stage of a.c./d.c. receivers. It is identical with the 6Y6-G except for its 50-volt, 0.15-ampere heater.

Radiotron type 50Y7-GT is a rectifier-doubler valve used particularly in "transformerless" receivers of either the a.c./d.c. type or the voltage-doubler type. The heater is provided with a tap for operation of a panel lamp. Except for this difference the 50Y7-GT is like the 50Y6-GT.

Frosted face plates—the latest improvement in television picture tubes — reduce specular reflection by diffusing reflections of bright objects which might otherwise be objectionable. This improvement supplements the use of Filterglass face plates to give improved picture contrast.

Frosted face plates have been incorporated in two most widely used kinescope types for new equipment design — the 16GP4 and 19AP4-A. These kinescopes when supplied with "frosted" Filterglass faces, at no increase in cost, are designated as 16GP4-B and 19AP4-B, respectively.

The data and curves for the "frosted" versions of the 16G- and 19A- are identical with those for the "unfrosted" versions.

NBFM MOBILE TRANSMITTER

By H. W. Brown, Jr.

10 Metre Rig

Phone
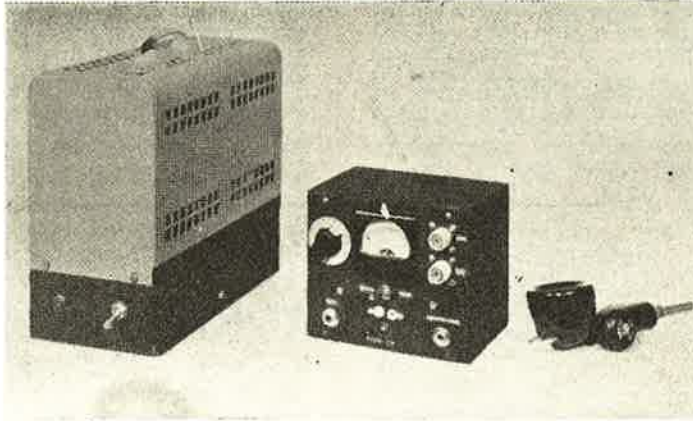
2E26 Final

C W

Here is a little rig, a complete 28-Mc/s transmitter, which can afford the operator much pleasure in either a mobile or fixed station location. This rig will especially hit the spot for those who travel by automobile and are away from home a lot. It is small, light, and versatile enough so that it can be carried about and set up at any location with not much more effort than hooking on the antenna and plugging into an appropriate power source. The transmitter features push-to-talk nbfm, or cw operation in either the 10- or 11-meter band. The power supply may be operated from either 6 V d.c. or normal a.c. supply.

Crystal reactance f-m

The basic device used for obtaining frequency modulation is something new and is known as the Gerber Crystal Reactance System. Its principle of operation briefly is as follows: An inductor, which is tuned slightly higher than the crystal frequency, is inserted in series with the crystal. The total inductance in the oscillator grid circuit is, therefore, increased. In order to meet the conditions for oscillation, the crystal will then assume a lower value of equivalent inductance and a lowering of the crystal operating frequency will result. In Figure 1, a typical reactance curve for a crystal, f_1 and f_2 designate points of series and parallel resonance, respectively. Point f_a is an operating point which meets the necessary conditions for operation. When an inductive reactance is added to the circuit, the crystal is required to adjust itself to a new frequency marked f_b for the total reactance to remain constant. If the magnitude of the added inductance is then varied electronically, as with a reactance-tube modulator, the crystal frequency will then swing up and down along the crystal reactance characteristic. Extremely wide deviations are possible with this

system; deviations in the neighbourhood of 20 kc/s are not uncommon at even 3 Mc/s. In this particular application, however, no such deviation is necessary.

R-F section

The crystal oscillator operates with a 7-Mc/s crystal and L_1 is the series inductor which produces the frequency deviation. The 6AG7, an excellent valve for oscillator-multiplier service because of its high control-grid screen-grid Mu factor, quadruples
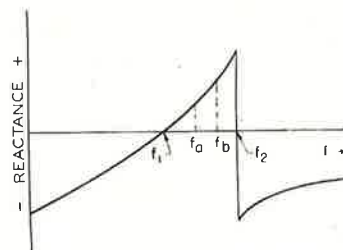
Figure 1. Typical reactance curve for a crystal.

to 28 Mc/s. The output of the 6AG7 drives a 2E26 final of conventional design to about 10 watts using the described power supply. The oscillator is basically a standard "Hot-Cathode Colpitts" circuit with a few modifications.
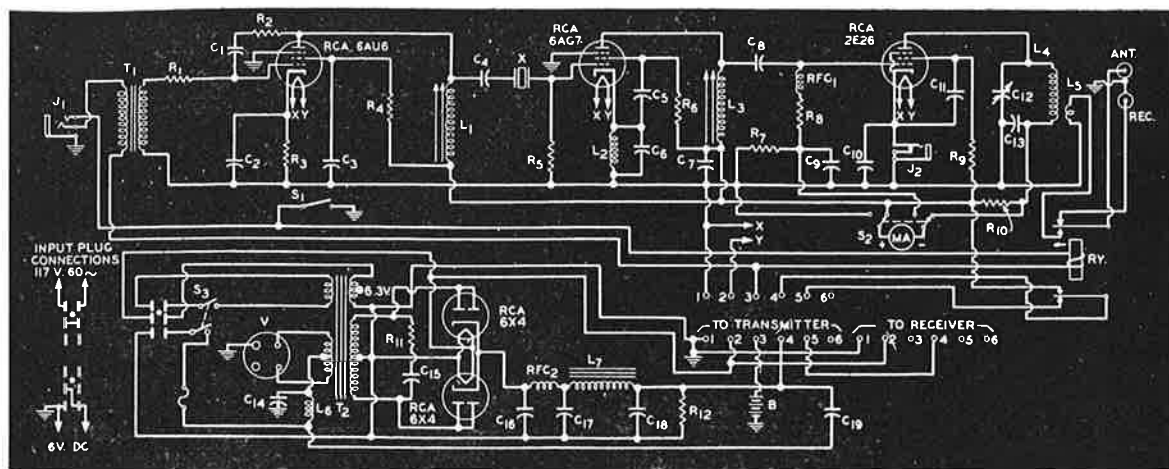
Fig. 2. Schematic diagram of the 10-metre transmitter.

Because it is desirable in this application to have good harmonic output consistent with low crystal current, the screen of the 6AG7 is bypassed to the cathode, and instead of a conventional 2.5-mH choke in the cathode tank circuit, a 6.2 μH choke (L_2) is used. This choke resonates closer to the crystal frequency, increasing the harmonic output considerably. Because the series inductor and the reactance valve introduce some losses into the oscillator, these two special design features are necessary so that ample drive is available for the final.

The reactance-valve modulator is conventional and uses a 6AU6 miniature pentode, which operates very satisfactorily. Adequate drive to obtain sufficient deviation is available directly from the microphone transformer.

The table below gives voltage and current measurements for the r-f section.

	6AU6	6AG7	2E26
Plate volts	290	290	290
Plate milliamperes	3	30	60
Input watts	0.87	8.7	17.4
Screen volts	150	240	180
Screen milliamperes	2	6	10
Control-grid milliamperes	—	—	2.2
Power-output watts	—	—	10

A single-button carbon microphone was used with this rig, a push switch on the mike performing the dual function of energizing the send-receive relay which switches the antenna and B+ from the receiver to the transmitter and controlling the mike current. Because the transmitter was designed to work from either a d.c. or a.c. input, it was necessary to include a microphone battery. If only d.c. operation is contemplated the relay and mike voltage may be obtained from the main storage battery.

It may be observed that there is no front panel control for the oscillator tuning. It was found that if the slug on L_3 is set for middle-of-the-band operation, there is no appreciable difference in power output when crystals of other frequencies are used. It proved necessary, however, in the interest of efficient operation to provide a tank tuning control on the 2E26 final amplifier.

L_1 should be adjusted to the point where a frequency shift of about 5 Kc/s is observed between the conditions when it is shorted out and when it is in the circuit. It is comparatively simple to adjust the deviation merely by listening to the modulation frequency in a receiver and setting L_1 accordingly; however, it is desirable to check the actual deviation by any of the usual methods.

It may also be noted that provision (J_2) has been made for keying the 2E26 cathode circuit. So that one would not have to keep the mike button depressed to hold in the relay, a separate phone-cw switch S_1, has been provided. This switch closes the circuit to the relay.

Metering is accomplished by switching a meter (0-10 mA) from the grid to the plate of the 2E26. When plate current is measured, a shunt is automatically connected into the circuit which multiplies the meter readings by 10. The exact value of this shunt (R_{10}) may be calculated or it may be determined by cut-and-try methods.

Construction and layout details

The entire r-f section is built into a 4" x 5" x 6" standard cabinet. The chassis layout is illustrated in the photo on page 106. From left to right may be seen the relay and the 6AU6, the 6AG7, and the 2E26 with its tank circuit. The deviation control L_1 is under the meter while the oscillator-tank tuning slug is between the 6AG7 and the 2E26. The r-f output is brought out to standard co-ax connectors on the front panel.
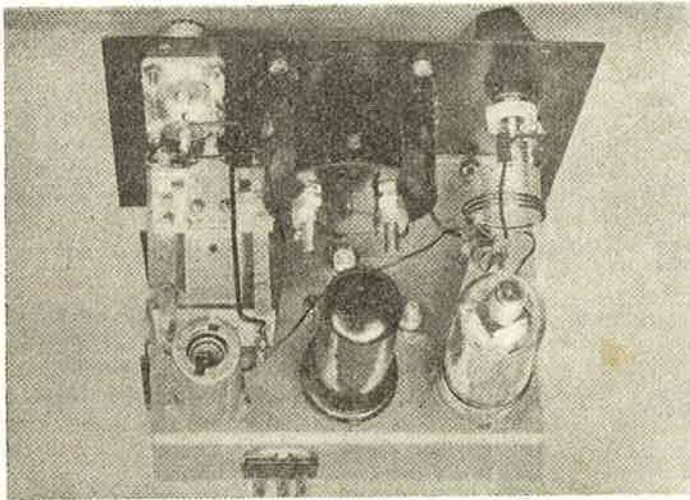
Power supply

As mentioned above, the power supply is designed to work from either a 6-volt battery or from a.c. mains. To accomplish this a transformer with a dual input is used in conjunction with a vibrator. For battery operation, of course, it is desirable to run the filaments from the d.c. source, so a switching arrangement is included with the power input plugs. A

heavy-duty male 6-connector Jones plug on the chassis is used for input and two matching female receptacles are used on the power cables. Connecting the appropriate plug to the unit automatically makes the proper filament connections. Two miniature 6-connector female chassis-mounting-type Jones receptacles are used for output power and control circuits: one feeds the transmitter and one the associated receiver. A pair of 6X4 rectifier valves, connected in parallel, will furnish more than enough current to operate the transmitter. Approximately 300 volts is available from the supply.
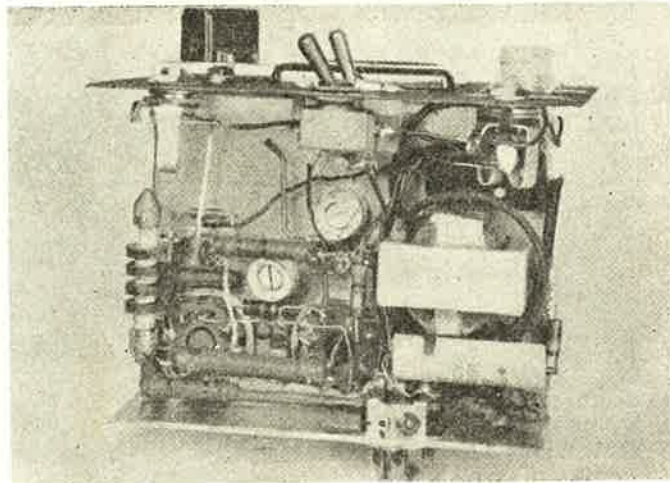
An underchassis view of the transmitter reveals simplicity of construction and the compact manner in which wiring and components are arranged.

C_{12} = 15 $\mu\mu$F variable
C_{14} = 5μF, 200 WV.
C_{15} = .01 μF, 1,600 W.V.
C_{16} = .01 μF, 600 W.V.
C_{17} = 8 μF, 450 W.V.
C_{18} = 30 μF, 450 W.V.
C_{19} = 100 $\mu\mu$F, mica
V = Vibrator
Ry = Relay dpdt, 6 volt d.c.
F = Fuse, 25 amp.
RFC_1, RFC_2 = 2.5mH
B = 6 V. Battery
T_1 = Microphone transformer.
T_2 = Dual Power transformer, 6 volts and a.c. mains input.
J_1 = 3 conductor microphone jack
J_2 = Key jack, closed circuit
S_1 = SPST toggle switch
S_2 = DPDT toggle switch
S_3 = DPST toggle switch, 15 amp.
Ma = 0-10 ma d.c. meter 3″
L_1 = 37 turns B & S #30 enamelled, 2 layers, $\frac{13}{32}$″ dia.
L_2 = 6.2 μH choke
L_3 = 10 turns B & S #24 enamelled, $\frac{13}{32}$″ dia.
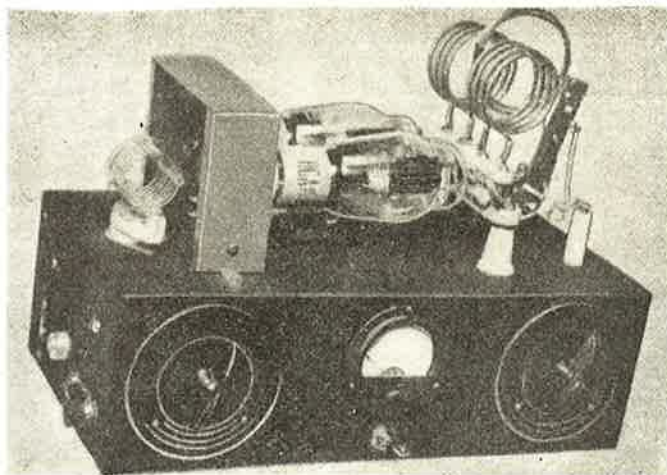L_4 = 10 turns B & S #16, $\frac{3}{4}$″ diameter, $\frac{1}{4}$″ long (air wound)
L_5 = Antenna link — 3 turns B & S #16 over cold end of L_4
L_6 = Hash choke
L_7 = Filter choke
X = Appropriate 7 Mc/s crystal

NOTE: All resistors $\frac{1}{2}$ watt unless otherwise noted.

Valves and relay are mounted with room to spare on small 4″ x 5″ x 6″ chassis. The 2E26 valve and tank circuit are shown on the upper right portion of the photo.

PARTS LIST

R_1, R_2, R_4 = 68,000 ohms
R_3 = 680 ohms
R_5 = 47,000 ohms
R_6 = 8,200 ohms, 1 watt
R_7 = 47 ohms
R_8 = 10,000 ohms
R_9 = 10,000 ohms, 5 watts
R_{10} = Meter Shunt
R_{11} = 5,000 ohms, 1 watt
R_{12} = 220,000 ohms
C_1, C_6, C_8 = 100 $\mu\mu$F, 600 W.V.
C_2 = 10 μF, 25 W.V.
C_3 = 1 μF, 400 W.V.
C_4 = 1,500 $\mu\mu$F.
C_5, C_7, C_9, C_{10}, C_{11}, C_{13} = 5,000 $\mu\mu$F, 600 W.V.

Stable 807 R-F Amplifiers

By J. H. Owens

One-hundred-and-fifty watts input to a cw final with a plate supply of only 750 volts! All-band coverage in the h-f region from 3 to 30 Mc/s, with plug-in coils! Complete freedom from parasitics, without neutralization! And less than two watts of grid-driving power easily obtainable from a 6V6-GT doubler! It's readily possible with a pair of Radiotron 807's.

Fig. 1a illustrates the usual layout for 807's. Little wonder that it causes so much difficulty when the three prominent feed-back paths are recognized and understood. As the arrows show, direct electrostatic coupling exists between the grid and plate circuits, (1) from the plate tank coil to the grid lead inside of the valve stem, (2) from the plate electrode to the grid tank coil, and (3) from the plate tank condenser to the grid tank condenser. The heavy dashed lines indicate the shielding required to eliminate these sources of stray coupling.

The torpedo attack

Fig. 1b shows a preferred mechanical layout for the 807's. Because the valve is mounted horizontally. "torpedo" fashion, it is naturally more stable than it would be in a conventional arrangement because of space isolation alone. The valve socket is mounted through a metal plate which acts as a shield against electrostatic and electromagnetic forces above the chassis. The plate coil is mounted as illustrated at right angles to the grid coil. A valve shield is not necessary.

Under the chassis, the grid and plate tank condensers are mounted with their rotor sides face to face. This arrangement, plus the greater distance of separation, usually provides sufficient isolation. For additional isolation, however, a shield, transformer, capacitor, or some other metal-cased component can be installed in the space indicated by the dashed line. External feedback is thus reduced to a minimum.

Beam valves versus triodes

Usually unwanted oscillations in electron valve apparatus result from interaction between the input and output circuits. The tendency toward instability depends on the degree of coupling and the grid-plate power gain. If there is zero coupling, there can be no feedback oscillation. Likewise, if there is zero

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Fig. 1a. This customary layout for the 807 valve may result in feedback difficulties.

Fig. 1b. A preferred mechanical layout for the same valve which reduces feedback to a minimum.

power gain, there can be no oscillation. In a practical circuit, the degree of coupling and the grid-plate power gain are positive factors, and oscillations are therefore possible whenever the feedback path is capable of transferring driving power to the grid.

In comparison with a beam valve, a triode having a power gain of only twenty is quite easy to stabilize. Since the power gain is relatively low, satisfactory operation may be expected even though neutralization and grid-plate isolation are imperfect.

However, a beam valve such as the 807, having a power gain of approximately 250, requires more careful handling. The plate has only to "breathe" on the grid to make the circuits oscillate. However, if the power gain of the beam valve is lowered to the level of the triode, it is more easily stabilized than the latter because of its internal shielding. It follows, then, that beam valves are more stable than triodes when operated under identical performance conditions.
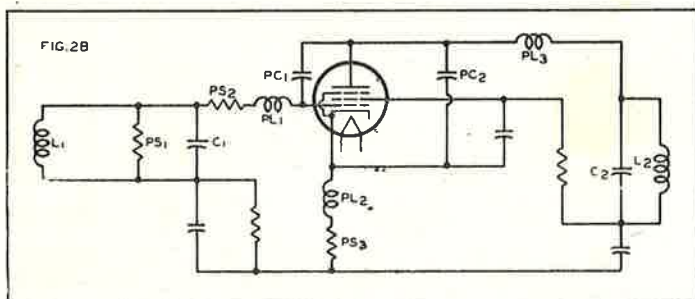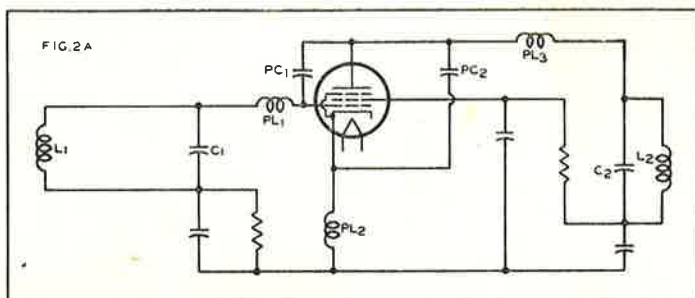

Fig. 2a (above). Parasitics are encouraged in this r-f amplifier circuit arrangement.
Fig. 2b (below). The same circuit after it has been stabilized by degeneration.

Parasitics

Parasitics are less likely to occur in equipment using old-type low-gain triodes; that is, valves that have a power gain of less than ten or valves that have wide electrode spacings and heavy electron transit-time loading. Parasitics will be found, however, in modern equipment using modern valves, triodes included.

Although parasitics are invisible, they furnish plenty of evidence of their presence. They are the commonest cause of plate tank condenser flash-overs. They heat plate and grid terminal caps. They prevent a pronounced dip in plate current when the unloaded tank circuit is tuned through resonance. They keep the plate efficiency low, and they are responsible for much modulation splatter and BCI.

If a circuit is free of parasitics, the tube will act like a pure resistance when excitation is removed. With full plate (and screen) voltage applied and with the plate tank unloaded, the grid current should drop to zero. Under such conditions, it should not be possible to light a neon bulb at the plate. In making such a test, it is essential to drop the plate voltage so that the rated plate dissipation is not exceeded.

LFP's, NFP's, HFP's?

There are three common forms of parasitics, LFP's, NFP's, and HFP's. LFP's are parasitics lower in frequency than the operating frequency. NFP's are parasitics that are simply self-oscillations at the normal operating frequency. HFP's are parasitics higher than the operating frequency.

LFP's are encountered in electron valve amplifiers having r-f chokes in both the grid and plate circuits. When these chokes resonate, low-frequency parasitics can be generated in tuned-grid, tuned-plate fashion. They can be suppressed by removal of either choke or by the use of a plate choke having a different resonant frequency than the grid choke. NFP's, commonly spoken of as "regeneration", and HFP's, also a common type of parasitic, are not so easily eliminated.

R-F degeneration

Although degeneration is known to benefit audio systems, little consideration has been given to it for r-f work. Yet its benefits can be essentially the same. It helps reduce the percentage of undesired harmonics and the trouble they cause when radiated. Moreover, when properly employed, it positively eliminates feedback parasitics.

Parasitic suppression through degeneration should be regarded as a desirable design practice rather than as an expedient for parasitic correction. It costs so little—just a slight increase in driving-power requirements. And the mechanics are so simple—just a few ohms of resistance in the proper places.

Fig. 2a shows a few of the inductive and capacitive elements that may cause parasitics in a typical r-f amplifier. HFP's are quite likely to be found in such an amplifier because at the frequency where the parasitic elements resonate, the regular tuning condensers C_1 and C_2 act like bypass condensers and provide a return path for the parasitic currents.

Fig. 2b shows the same circuit after it has been stabilized by degeneration. A parasitic suppressor, PS_2 is located in the grid circuit where d.c. and normal-frequency r-f currents are small, but where HFP currents would be large. It reduces the circuit power gain just a little but kills HFP's and NFP's which result from capacitive feedback from the plate.

PS_3 is connected to provide simple cathode circuit

degeneration. It lowers the power gain slightly but compensates by reducing harmonic generation. Parasitic element PL_3 has been left undisturbed because without a co-operating element in the grid or cathode circuit it can do little harm.

Non-inductive carbon resistors are favoured over parasitic chokes because the chokes simply shift the resonant frequency of the parasitic circuits. While

Fig. 3. Schematic of the 150-watt final amplifier using a pair of 807's.

this expedient may be very effective for a fixed frequency transmitter, it is not the answer to a multiband amateur unit.

Other stabilizing stunts

A very effective way of suppressing NFP's is to place a small load across the grid tank circuit. A carbon resistor (PS_1) having a value of something between 5000 and 50,000 ohms will really get results with an insertion loss of only a fraction of a watt. The resistor simply limits the impedance of the grid tank so that minute currents fed back from the plate cannot develop excessive grid voltages which react to cause greater plate-current fluctuation and eventually self-oscillations.

If the 807 is loaded to less than the maximum rated plate current of 100 milliamperes, the screen-grid voltage can be reduced proportionally. The effect is an increase in stability resulting from the reduction in transconductance. The difference in power gain is so small that it is negligible.

Still another way to make 807 meek and obedient if trouble is experienced is to let the driver valve give the grid all of the driving power it can handle. Then, the feedback power becomes small and ineffective by comparison. The procedure is to increase the bias over typical operating values and run the grid current up near maximum ratings.

A further recommended design practice is the use of a small amount of cathode bias or fixed bias.

This precaution keeps the transconductance within reasonable limits and protects the valves and other components during periods of momentary overload when excitation is lost or the plate circuit detuned.

The Torpedo Twin

The photograph and circuit diagram in Fig. 3 illustrate how a pair of 807's might be used in a 150-watt final amplifier. The same mechanical arrangement could be used on a larger chassis of a relay rack panel to provide space for a crystal oscillator and power-supply components. The circuit works all bands from 75 to 10 metres. The coils can be purchased units or they can be home-wound according to Amateur Handbook directions.

The meter arrangement is made by first removing the internal shunt from a 0-300 milliammeter. The shunt is then installed externally across one side of the DPDT toggle switch. Next, another shunt is made from a small piece of resistance wire which will make the meter read 0-30 mA. This shunt is placed across the other side of the toggle switch. A second meter, of course, is needed to calibrate the new shunt. In one position, the meter reads grid current, and in the other position it reads total cathode space current.
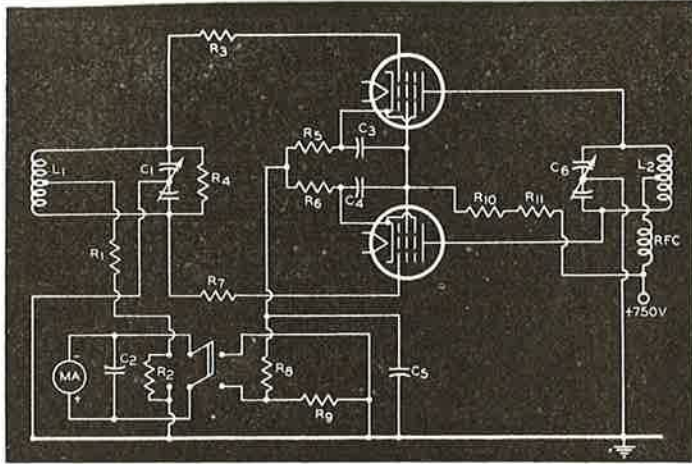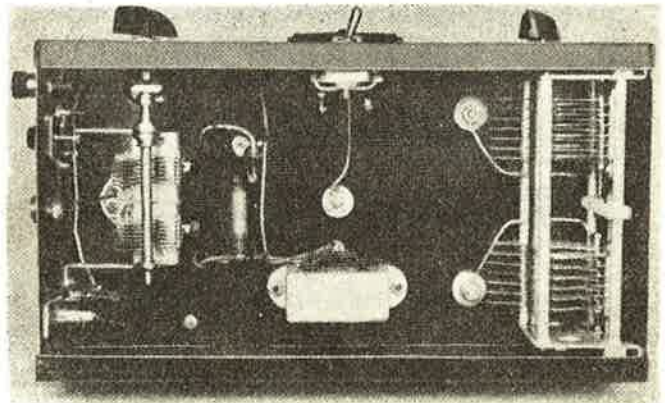

The mechanical arrangement of components contributes to its highly stable operation.

PARTS LIST

R1	5000 ohms ½ watt. carbon.
R2	Home-made shunt (see text).
R3, R7	50 ohms (or less) ½ watt. carbon.
R4	10,000 to 100,000 ohms 1 watt carbon (see text).
R5, R6	25 ohms (or less) ½ watt. carbon.
R8	200 ohms. 10 watt. wire wound.
R9	Resistor shunt taken from meter.
R10, R11	~~2-6,000 ohms~~ 10 watt. wire wound in series.
C1	100 μμF variable. each section.
C2	500 μμF midget mica.
C3, C4, C5	0.002 μF postage stamp. mica.
C6	100 μμF each section. variable. 0.077" spacing.
RFC	2 mH r-f choke.
L1, L2	See text.

(handwritten: (R10 R11) X Total res. of 45k, eg. 20 +25, each of 5W, in series.)

Germanium Crystal Rectifiers

The following gives an explanation of the terms used in specifying diode characteristics and the maximum ratings for the various crystal diodes.

Maximum inverse (reverse) current

This is the greatest amount of current which may be expected to flow through the diode when a given d.c. potential is applied across it with the negative polarity connected to the whisker or anode lead. The corresponding back resistance may be calculated by dividing the applied voltage by the current flow, and since the diode is a non-linear device the resistance will vary with the applied voltage. Radiotron diodes are normally rated for inverse current at − 10 volts.

Minimum forward current

This is the smallest amount of current which may be expected to flow through the diode when a given d.c. potential is applied across it with the positive polarity connected to the whisker or anode end. The corresponding forward resistance may be calculated by dividing the applied voltage by the current flow. The forward characteristic is also non-linear with voltage and measurements are normally made at + 1 volt.

Shunt capacitance

Capacitance measurements are made only as design checks and not as a manufacturing test. Such measurements can be made on a Q-meter, using a low value of capacitance in series with the diode. Since variations in back resistance will affect the "Q" of the test circuit at low frequencies and give capacitance readings which are not indicative of the true capacitance of the diode, all checks should be made at v.h.f., say 75 Mc/s.

Peak inverse (turnover) voltage

This rating is the maximum transient inverse voltage which may be applied to the diode without danger of voltage breakdown of the unit. Voltage breakdown is the point at which the dynamic back resistance of the diode becomes zero.

When the applied voltage is raised slightly beyond this point, the inverse current will increase rapidly to a high value.

Maximum continuous operating inverse voltage

This rating is the maximum continuous inverse voltage which may be applied under normal operating conditions to assure continuously reliable operation, and is 20% lower than the peak inverse voltage.

Average rectified current

This rating is the maximum average current which can be carried by the diode without appreciable heating of the unit or change in its character-

istics. For Radiotron crystal rectifiers this is 30 mA. Higher current up to 100 mA may be applied intermittently.

Peak rectified current

This refers to the peaks to which the rectified current may be allowed to rise provided the duty cycle and wave shape are such that the r.m.s. current does not exceed the average rating and the frequency of application is at least 25 cycles per second.

Maximum surge current

This represents the maximum current which may flow for one second without damage to the unit.

Ambient temperature range

Diode characteristics are a function of the diode temperature. The range of temperature 15°C - 50°C is that range over which the diode exhibits reversible changes, in that the characteristics return to their original room temperature values after being subjected to other temperatures in the rated range. Temperatures in excess of those shown (up to 100°C) will in some cases tend to cause permanent changes in characteristics.

SERVICE NOTES

I. Installation

A. An abnormal amount of tension applied to the leads when attaching them to terminals is undesirable and should not be necessary. It is preferable to leave some slack or an expansion "elbow" wherever possible.

B. The application of excessive heat in soldering the leads is not advised. Excessively high heat as encountered with a soldering iron, when applied longer than necessary, may permanently change the diode characteristics. The leads should be left as long as possible and only enough heat applied to melt the solder properly. It is common practice throughout the industry when short leads are necessary or several wires, in addition to those of the diode, are being soldered to a terminal, to grip the lead, between the body and the soldered point, with a pair of pliers to absorb some of the heat that would normally flow into the diode housing.

II. Check and inspection

A. General purpose types.

The effect of a germanium diode failure in a circuit is obviously peculiar to that particular circuit and cannot be covered by a general statement. However, if it is felt that operational difficulties are caused by the diode, it is suggested that one lead be disconnected and the diode characteristics checked. In order to determine whether or not the unit is open, an ohmeter can be employed. A relatively high resistance (many thousand ohms) should be
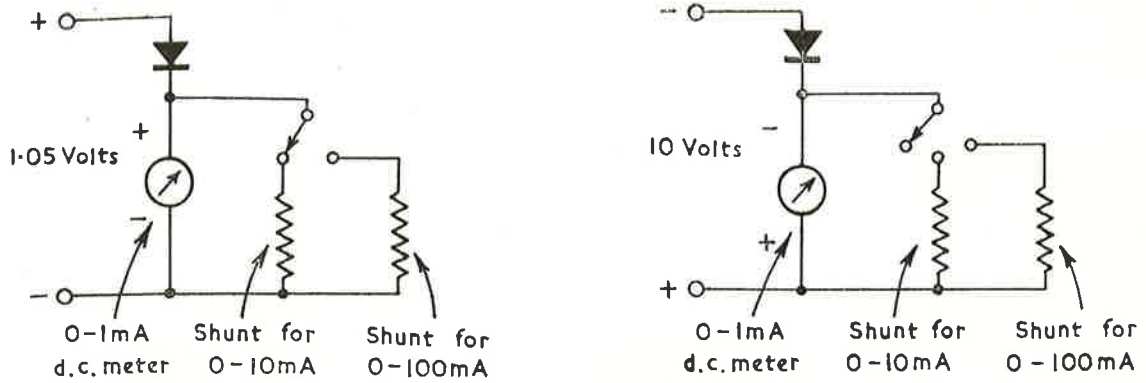
Test circuits.

noted with the leads reversed.

However, it may be desirable to determine whether or not the unit is within ratings. In this case an ohmeter cannot be used.

Some concern is occasionally encountered because diodes apparently fail to meet published limits of backward and forward resistance. It usually develops that the resistances in question have been checked with an ohmeter. The measurements made with an ohmeter will seldom agree with those made at the factory. Resistance ratings are based on definite applied d.c. potentials such as + 1 and − 10 volts. The resistance of a diode varies with the applied voltage and must, therefore, be checked under standard conditions at all times if the readings are to be used as criteria of grade.

Most ohmeters will apply approximately 1.5 volts to the diode on the Rx 1 scale and hence give low forward resistance readings; moreover, the reading generally occurs at the crowded upper end of the scale where the accuracy is poor and the scale difficult to read. Other scales give voltages much less than 1 volt and show correspondingly higher forward resistances. A back resistance reading made on an Rx 1000 scale will measure the resistance with applied voltages in the range of − 2 to − 12 volts d.c. This is a range as shown on the previous resistance-voltage curves. These readings, therefore, would have very little correlation with factory readings. In addition, the actual voltage will be different between diodes since the voltage impressed on the specimen will depend upon the current drawn by the particular unit under test.

For this reason, the ohmeter is not even a suitable comparator. Even an attempt to compare diode back resistances with an ohmeter will not give a true picture of the relative resistances since a small difference in actual resistance will result in a different applied voltage to the diode which in turn may show a relatively great difference in resistance.

It is therefore recommended that an ohmeter be resorted to only for very rough checks on diodes, and the above circuits used for accurately verifying the published characteristics. Batteries are generally most convenient for obtaining the required voltages unless a permanent test is to be used.

MOOD MUSIC

As a guide to readers who are interested in recorded music, we append a list of titles suitable for various occasions. No record numbers have been quoted, as in the majority of instances, a choice of orchestras and conductors is available.

1. Peaceful — Pastoral

Afternoon of a Faun .... Debussy
Khowantchina—Introduction .... Moussorgsky
Moonlight Sonata —1st Movement .... Beethoven
Peer Gynt Suite No. 1 (Morning) .... Grieg
Cloude .... Debussy
Air for G String .... Bach
Clair de Lune .... Debussy
Grand Canyon Suite .... Grofe
The Old Refrain .... Mattulath
Abendlied .... Schumann
Berceuse .... Godard
Dream Pantomime .... Humperdinck

2. Gay — Light

Caprice Viennois .... Kreisler
Meistersinger—Dance of the Apprentices, Wagner
Entrance of the little Fauns .... Pierne
Shepherds Hey .... Grainger
Country Gardens .... Grainger
Reminiscences of Vienna .... Strauss
Etude in G Flat Major .... Chopin
Canadian Capers
Gossips .... Dubensky
Dance Des Mirlitons
(Nutcracker Suite) .... Tschaikowsky

3. Fast — Exciting

Perpetuum Mobile .... Strauss
Snow Maiden—
     Dance of the Tumblers .. Rimsky-Korsakow
Scherzo .... Mendelssohn
Capriccio Italien .... Tschaikowsky
Eugen Onegin—Polonaise .... Tschaikowsky
An Irish Washerwoman .... Sowerby
Perpetual motion .... Paganini
La Playera .... Granados
Hora Staccato .... Dinicu-Heifetz

4. Slow — Solemn Religious

The Angelus .... Massenet
Finlandia .... Sibelius
Xerxes—Largo .... Handel
Festival Te Deum .... Buck
Parsifal—Prelude and Good Friday Spell
.... Wagner

5. Majestic — Grand Marches

Marche Militaire .... Schubert
Pomp and Circumstance No. 1 .... Elgar
Siegfried's Funeral March .... Wagner
Marche Slave .... Tschaikowsky

6. Wild — Tempestuous

Ivan the Terrible—Storm Music, Rimsky-Korsakow
Die Walkuere—Ride of the Valkyries ... Wagner
Flying Dutchman—Overture .... Wagner
Hungarian Rhapsody No. 2 .... Liszt

7. Weird — Mysterious

Isle of the Dead .... Rachmaninoff
Night on Bare Mountain .... Moussorgsky
Danse Macabre .... Saint-Saens
Coq d'Or—Hymn to the Sun ... Rimsky-Korsakow
La Mer .... Debussy

New Wide Range R-C Oscillator

A resistance-capacitance oscillator with many immediate applications in radio and electrical work has been developed by Peter G. Sulzer at the National Bureau of Standards. The new oscillator covers, in five steps, the frequency range from 20 cycles to 2 megacycles. Combining simplicity and compactness with excellent frequency stability over a wide tuning range, it has several advantages over previous R-C oscillators. In older models the top frequency is about 200 kilocycles; in the new R-C oscillator, a single amplifier driving a cathode follower provides wide-band operation with small phase shift, low output impedance, and constant output voltage.
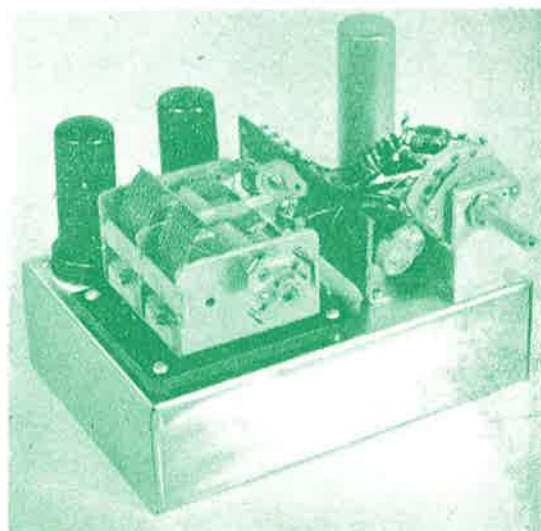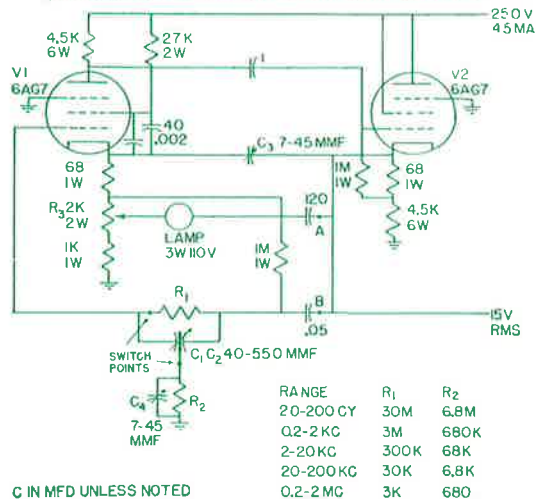
The oscillator circuit has two feedback paths: a regenerative cathode-to-cathode loop and a degenera-tive cathode-to-grid loop which includes a bridged-T network. Oscillation occurs at the frequency of minimum degeneration. The 15 volt output remains constant to within one decibel at all frequencies, and the output waveform is essentially undistorted. By proper shielding to prevent synchronization with the power-line frequency, the R-C oscillator may be mounted in the same small cabinet with a power supply and an output amplifier.

Note: For further technical details, see "Wide-range R-C oscillator," by Peter G. Sulzer, "Electronics" (1950), and also an article on the same subject in "QST" for January, 1951.

A new resistance-capacitance oscillator that covers, in five steps, the frequency range from 20 cycles to 2 megacycles, extends the frequency of R-C oscillators from a previous maximum of about 200 kilocycles.

RANGE	R_1	R_2
20-200 CY	30M	6.8M
0.2-2 KC	3M	680K
2-20 KC	300K	68K
20-200 KC	30K	6.8K
0.2-2 MC	3K	680

C IN MFD UNLESS NOTED

By proper shielding, to prevent synchronization with the power-line frequency, the R-C oscillator may be mounted in the same small cabinet with a power supply and an output amplifier.