RADIOTRONICS

ISSUE 120 JULY-AUGUST 1946.

LABORATORY RATING TEST SET

The highest standards of accuracy are maintained by the use of this equipment, which checks the 100% testing in the manufacture of Australian-made Radiotron Valves.

RADIOTRONICS

Editorial - -

The two outstanding items in this issue are the description of a four valve reflex a.c. mantel model receiver, and the reprinting of two reports on the Ratio Detector.

There is an increasing percentage of the market which calls for small broadcast-band mantel sets. The majority of these now on the market have very limited performance, and are more or less local station receivers, limited to regions of fairly high signal level for really satisfactory performance. The four valve mantel receiver described in this issue has been designed in our Applications Laboratory to make the best possible use of four of the single-ended a.c. valves. Its sensitivity is ample for the reproduction of the stronger interstate stations, or for the country reception of metropolitan stations, at full loudspeaker strength even with a small indoor aerial, at night time. Special attention has been paid to the a.v.c. characteristic and to the prevention of overloading by strong local stations, with very satisfactory results. The design is very conservative, and incorporates a neutralized i-f amplifier and negative feedback on the audio amplifier.

As a service to our readers, we reprint in this issue, through the kind permission of the Radio Corporation of America, two reports by the R.C.A. Laboratories on the Ratio Detector. This is an invention which looks as though it will play an important part in the design of F-M receivers, since its use may permit a reduction in the number of valves. The Ratio Detector is being investigated in our Applications Laboratory, and we hope to deal with its practical application in a future issue.

Radiotronics is issued six times a year and the annual mailing fee is 2/6. All communications, subscriptions and change of address notices should be directed to the Valve Company—except in New Zealand, where correspondence should be addressed to the National Electrical and Engineering Co. Ltd., 286-288 Wakefield Street, Wellington.

Technical Editor
F. Langford-Smith, B.Sc., B.E.

Published by the Wireless Press
for
Amalgamated Wireless Valve Company Pty. Ltd.
47 York Street, Sydney, N.S.W.

Radiotronics is printed in Australia by the Cloister Press (D. & W. Short), Redfern, N.S.W.

Number 120 ★ July-August, 1946

IN THIS ISSUE.

DESIGN SECTION:

THEORY SECTION:

VALVE DATA SECTION:

Radiotron Receiver RD31

4 VALVE A.C. REFLEX CIRCUIT USING SINGLE-ENDED G.T. VALVES

A 4 valve receiver is described which uses reflexing for increasing sensitivity and yet minimizes the disadvantages usually inherent in reflexed amplifiers. The sensitivity is more than adequate for the performance required from a small mantel receiver.

There can be no doubt there is a well established demand for a small, economical, but sensitive mantel receiver. The technical problem involved is not an easy one, because the requirements of small size and rigid economy are quite opposed to the other requirement of reasonable performance. The usual design is therefore rather more of a compromise than is desirable. In order to achieve sufficient selectivity and a sensitivity better than 300 microvolts, it is generally accepted that a superheterodyne receiver is necessary. Thus, with ordinary valves, a minimum of two are required for frequency conversion, intermediate frequency amplification and detection. With one valve for rectification, and one for power amplification, a total of four appears to be the practical minimum.

The omission of the audio voltage amplifier is the concession usually made to economy. A receiver designed along these lines will normally have a sensitivity of not much less than 300μV, and many situations will occur where this sensitivity is inadequate. The idea of using one of the valves to simultaneously amplify inputs of two different frequencies is not by any means new. Some of the very early receivers of the triode r-f amplifier era quite commonly employed this device. In more recent years, considerable thought has been given to eliminating the fundamental failings of the reflexed amplifier, one of which is the so-called "minimum volume effect", which causes a very distorted output from the speaker when the volume control is turned nearly off, and a substantial output when the control is turned right off. It results from the inevitable curvature of the valve mutual characteristic partially rectifying a large intermediate frequency voltage. Then, because of the audio frequency load and capacitance in the plate circuit, the valve acts as an anode-bend detector, which passes audio output directly to the power valve. This output is quite independent of the diode, and is unaffected by the volume control. Though normally masked by a much larger output from the diode detector, it can cause severe distortion when the diode-rectified-signal present in the reflexed amplifier plate circuit is approximately equal to the anode-bend-rectified signal. This is because the fundamental components of the two waves are in phase opposition and tend to cancel, leaving a smaller signal with a much larger proportion of harmonics. This is the point of minimum volume and maximum distortion. At still lower settings of the volume control, the diode output is attenuated, and the anode-bend signal is the principal source of output.

A further serious disadvantage of the usual reflexed amplifier has been the tendency to overload on moderate and strong input signals. This is because the presence of an audio load of the order of 0.1 megohm in the plate circuit makes its necessary to reduce the plate current, by means of lower screen voltages and/or increased negative grid voltages, to avoid excessive drop in the plate load. The best compromise considerably reduces the transconductance of the valve. As the gain between the grid of the converter valve and the diode is not high enough to provide a good a.v.c. characteristic, the i-f input to the reflexed valve is therefore quite high. Furthermore, the peak a.c. potential at the plate is the sum of the i-f and a-f signals, while the d.c. potential is only of the order of half the B+ supply. Thus, a valve which is not capable of handling large signals without distortion has, by virtue of the circuit arrangement, large signal voltages applied to it, thereby resulting in overloading.

The arrangement used in this receiver employs a very small a-f load in the plate circuit. This has two important advantages: the full i-f gain can be realized, and the minimum volume effect is reduced to negligible proportions for all but the most severe listening conditions. With an a-f plate load of 15,000 ohms, the high mutual conductance of type 6SF7-GT results in an audio gain of about 12.5 times for very little sacrifice of i-f gain. This extra amplification brings the sensitivity figure down to the region of 40 microvolts input for 50mW output. The receiver is completely stable on signal inputs from 1 microvolt to 1 volt. For high volume levels, overloading takes place first in the speaker and then in the power valve, which is designed to deliver a little more than 1 watt to minimize heat dissipation.

General Description of the Circuit

Type 6SA7-GT is used as converter in a conventional arrangement with a tapped coil cathode feedback oscillator. It is interesting to note that with the padding condenser placed as it is, the oscillator voltage between cathode and earth is higher at the low frequency end of the band—quite the reverse of the usual case. When the receiver is designed to cover only the broadcast band there is the useful possibility of making the grid condenser small enough in relation to the grid leak to minimize the potential rise for low frequencies. In this way it is feasible to preserve almost optimum oscillator voltage over the band.

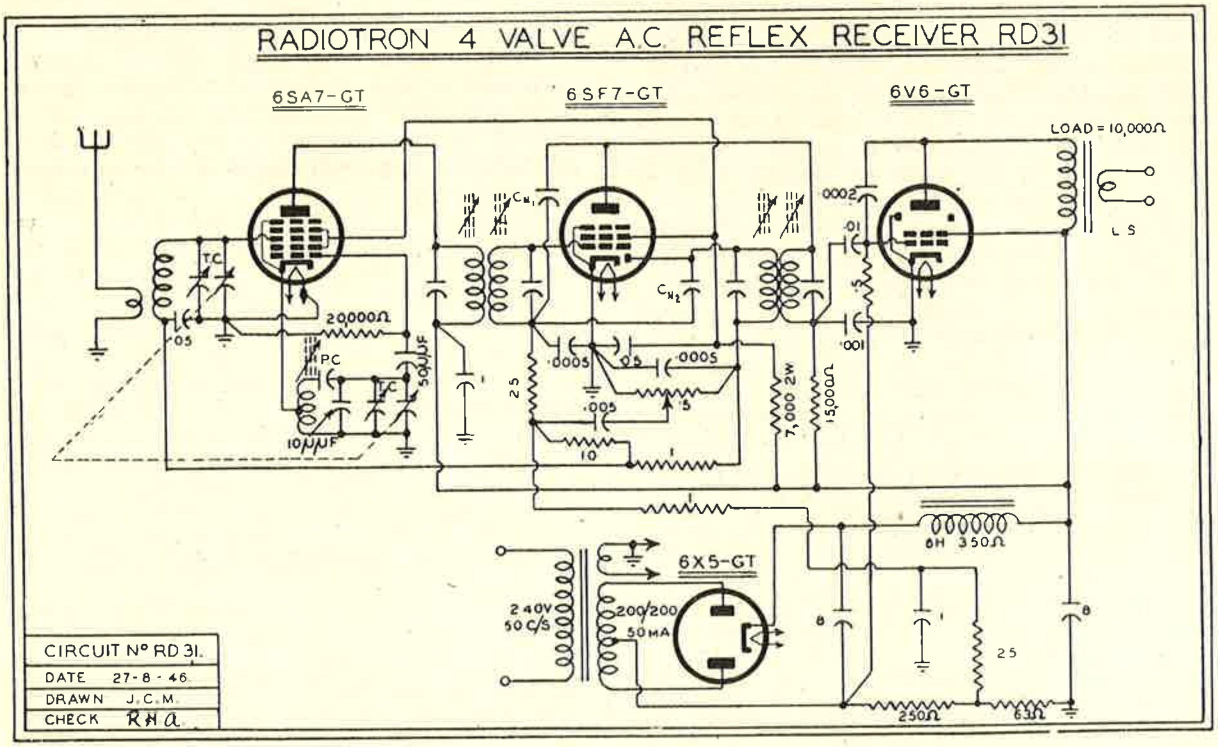
The intermediate frequency transformers are of medium quality and small dimensions, as is usual for a small mantel receiver. Improved selectivity, with some increase in sensitivity, could readily be achieved with higher Q i-f transformers. The aerial coil has the usual high impedance primary, and is unshielded. The oscillator coil is also unshielded but placed about 3 inches away and at right angles to the aerial coil. Several sample oscillator coils were tested in this receiver, including some from R.C.S. and some from Aegis. Both of these manufacturers can supply a quite satisfactory oscillator coil for the 6SA7-GT. The simplest test for optimum oscillator amplitude is to connect a valve voltmeter between cathode and ground. When this shows a potential of the order of 1.4 volts rms, the maximum conversion conductance will be obtained.*

The most interesting stage is the reflexed intermediate and audio frequency amplifier. Feedback is neutralized by the method used in our previous receivers employing 6SF7-GT amplifiers† The values of the capacitors in this case are much smaller because of the need to make the capacitor between the lower end of the i-f transformer and ground a high impedance to audio frequencies. The neutralizing capacitors C_{n1} and C_{n2} are very small indeed, and with a well designed layout they might even consist of stray wiring capacity. In our model a 3 inch length of shielded wire was run from the lower end of the i-f transformer, the shielding being earthed, with a $\frac{3}{4}''$ unshielded section at the end placed about $\frac{1}{8}''$ from the plate lug of the 6SF7-GT valve socket. Five turns of 26 B&S enamelled wire were wound over the rubber insulation of the unshielded portion and connected to the diode socket lug to form a small capacitor. In quantity production it should be found that a fixed neutralizing capacitor can be used, although individual receivers may be only partially neutralized, but sufficiently so to provide stability.

In practice the methods of adjustment previously described are inapplicable since the i-f valve is also an a-f amplifier; hence the capacitances of the pilot model should be adjusted during the i-f lining up process to give minimum sideband hiss as the signal generator is tuned through resonance. This results in minimum gain, through reduction of regeneration. The 0.25 megohm resistor connected to the lower end of the i-f transformer secondary prevents the volume control affecting the neutralizing, as well as providing i-f filtering of the diode output. While the maximum possible a.v.c. is applied to the converter valve, only about one ninth of this voltage

* R.C.A. Application Note "Operation of Radiotron 6SA7", reprinted in Radiotronics 95, pages 8 to 14.

† Radiotronics 118: Circuits RC41, RC42, and RC53; Article "Neutralization in circuits employing a valve as a combined intermediate frequency voltage amplifier and diode detector", page 33.



can safely be used for the reflex stage. This is because any bias applied to the 6SF7-GT is doubly effective in that it reduces both i-f and a-f gain. Furthermore, the reduction in a-f gain does not reduce the a.v.c. voltage, so that with large proportions of a.v.c. applied to this stage, the output can actually become smaller for larger inputs; this is the well-known effect with excessive audio a.v.c. The result, then, of tuning to a powerful signal is to produce less output when the receiver is tuned directly to the carrier than when it is tuned to one side, so that there are two adjacent tuning positions of maximum volume. This effect is avoided by applying just sufficient a.v.c. to minimize the normal rise of the a.v.c. curve, but not allowing the curve to fall. Quite good automatic volume control can be obtained in this way.

The 6SF7-GT is given initial negative bias of approximately 2.5 volts. This is the bias for maximum mutual conductance with a resistance plate load of the order of 15,000 ohms. Although a bias of –1 volt is recommended on the data sheet for the 6SF7-GT, this is for zero plate load. While the mutual conductance curve rises to a maximum at zero control grid voltage. The dynamic curve with a resistive plate load flattens off near the top so that maximum slope exists at some negative bias greater than zero. It is important to bias this stage with a potential of not less than that providing maximum gain, otherwise a small increase of a.v.c potential will result in increasing gain—which is equivalent to positive feedback and leads to instability. If the alternative of operating this stage with fixed bias and no a.v.c. is followed, there is danger of overloading on strong signals—which also results in instability. The arrangement of fixed bias and fractional a.v.c. used in this receiver is quite stable.

The value of 15,000 ohms for the a-f plate load was chosen to take advantage of the neutralization of hum output possible when the two audio amplifiers have the correct gain relationship. It so happens that this value satisfies the other requirements of the a-f load. The method used for selection was to visually indicate the hum output from the plate of the 6V6-GT on an oscillograph and to vary the 6SF7-GT a-f plate load to show minimum hum output. A substantial reduction in hum output is possible in this manner.

The power amplifier valve is operated with low plate and screen voltages, and increased negative bias on the grid to reduce unnecessary heat dissipation. The power output for reasonable distortion is slightly greater than one watt. This exeeds the power handling capabilities of the usual small speaker generally fitted to mantel receivers, so there is no sacrifice of performance for the saving in power consumption and heat dissipation.

TEST RESULTS

(1) VOLTAGE MEASUREMENTS

Valve.	Zero Signal Input. Plate.	Screen.	Grid.	100mV Signal Input. Plate.	Screen.	Grid.
6SA7-GT	185V	96V	0 + a.v.c.	186V	95V	0 + a.v.c.
6SF7-GT	104V	96V	— 2.6V + a.v.c.	145V	95V	— 2.55V + a.v.c.
6V6-GT	180V	185V	—13.0V	181V	186V	—13.0 V
6X5-GT	200Vrms					

Total B Current = 41mA. for zero signal input and 40.5mA. for 100mV. signal input and 1W output.

(2) OSCILLATOR

Frequency.	e_k	e_o (total coil)	Ic_1
540 Kc/s	1.45 Vrms	19.0 Vrms	285 μA.
1600 Kc/s	1.23 Vrms	16.8 Vrms	535 μA.

(3) OVERALL PERFORMANCE

For Output of 50 Milliwatts (Absolute).

Input to	Frequency.	Input.	Ratio.	E.N.S.I.
6V6-GT Control Grid	400 c/s.	1.0 Vrms		
6SF7-GT Control Grid	400 c/s.	0.08 Vrms	12.5	
6SA7-GT Control Grid	455 Kc/s.	123μV.		
" "	600 Kc/s.	165μV.		
" "	1000 Kc/s.	145μV.		
" "	1400 Kc/s.	145μV.		
Aerial	600 Kc/s.	51μV.		0.3
"	1000 Kc/s.	41μV.		0.25
"	1400 Kc/s.	39μV.		0.24

A small condenser of 0.0002 μF. connected between plate and grid of the 6V6-GT provides negative voltage feedback which is greater for high frequencies than for low, thus serving as a "top limiting" device. This has the added advantage of negative feedback over the alternative system using a much larger condenser between plate and screen of the 6V6-GT.

(4) SELECTIVITY

Times down.		Bandwidth.
3	..	14.3 Kc/s.
10	..	24.5 Kc/s.
30	..	37.8 Kc/s.
100	..	47.4 Kc/s.
300	..	57.4 Kc/s.
1,000	..	72.5 Kc/s.
10,000	..	117 Kc/s.

N.B.—The i-f transformers used in the developmental model were of the type commonly adopted for use in small table model receivers, having small dimensions and fairly low Q. This, and the absence of regeneration, account for the broad selectivity characteristic.

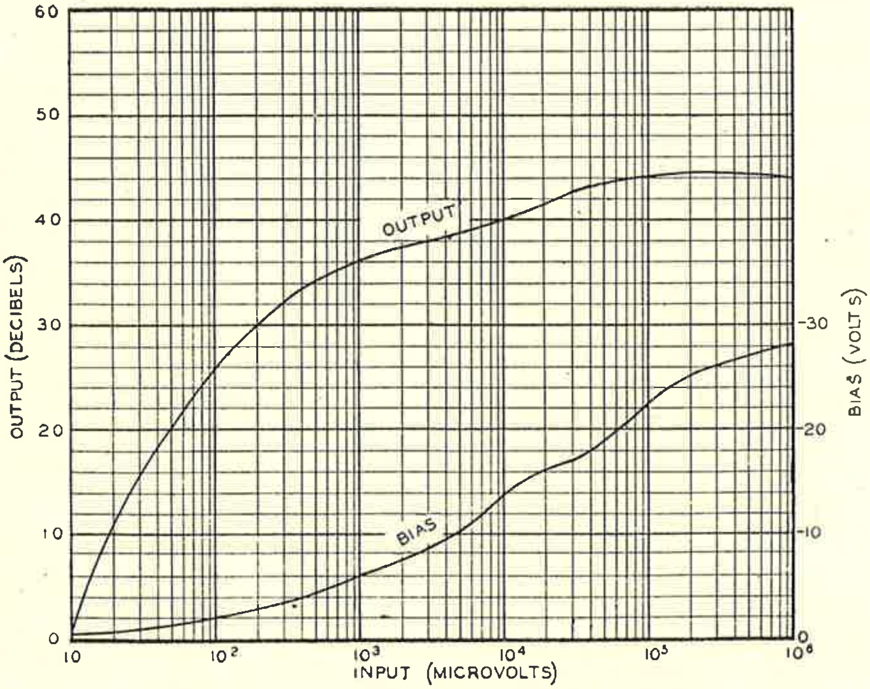
(5) A.V.C. CHARACTERISTIC*

Input.	a.v.c. volts.	Output.
10 μV.	— 0.6	— 29 db
30 μV.	— 0.8	— 14 db
100 μV.	— 1.8	— 4 db
300 μV.	— 3.8	+ 2.0db
1000 μV.	— 6.0	+ 6.0db
3 mV.	— 8.5	+ 8.0db
10 mV.	— 14.0	+ 10.0db
30 mV.	— 17.0	+ 12.5db
100 mV.	— 26.0	+ 14.5db
1 V.	— 28.0	+ 14.0db

This is a sensitive 4 valve mantel-model receiver using the single-ended GT valves. It has ample sensitivity for use in most country areas, while it is free from "overloading" in regions of high signal level. The circuit provides for i-f neutralization, hum neutralization, and negative feedback on the power stage.

The transformer design is based on the use of a permagnetic speaker and a 380 ohm filter choke. An alternative arrangement could be an electrodynamic speaker with a 1,000 ohm field used as a filter choke. The power transformer secondary voltage should then be 225 volts r.m.s.

* With retuning for maximum output.



A.V.C. CHARACTERISTIC—CIRCUIT RD31

A remarkably good a.v.c. characteristic for a 4 valve receiver is achieved by the use of fractional a.v.c. on the i-f (reflex) stage which simultaneously provides audio a.v.c.

Extension Loudspeakers

Provision is made in some radio receivers for the subsequent addition of an extension loudspeaker, but the subject is one well worthy of careful consideration.

This article analyzes various possibilities in connection with the use of one, two, or both loudspeakers with various arrangements of control.

The most common provision for an extension loudspeaker is shown in circuit form in figure 1, in which the extension loudspeaker (marked "EXT") is fed through a blocking condenser connected to the plate of the power valve. It is necessary for the set user to arrange for flexible leads to be taken from two terminals on the chassis, across the primary of the stepdown transformer, to the permagnetic extension speaker. If the extension speaker has the same reflected load impedance as the transformer of the original speaker, the power output will be shared equally between the two, but the impedance of the two speaker primaries in parallel will be only half what it should be for maximum power output. If no provision is made to provide correct matching, the arrangement will be quite practical, except that power output of both speakers together will be less than with correct matching.

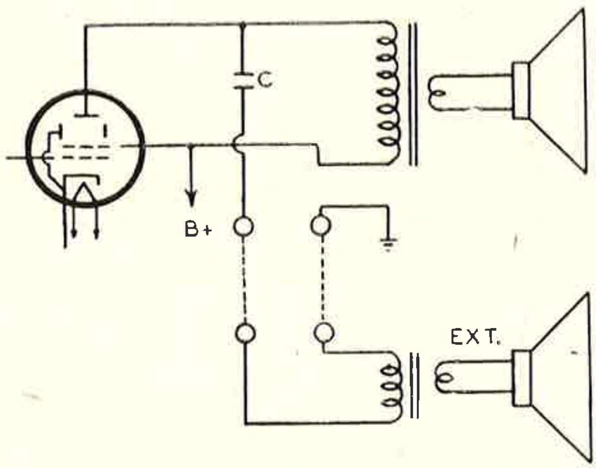


FIG. 1.

An alternative arrangement, which has much to commend it, is the use of an extension speaker with an impedance of about twice that of the speaker in the set. This will mean that the extension speaker will operate at a lower sound level than the one in the set, but the mis-matching will be less severe and the maximum volume obtainable from the set will not be seriously affected.

If negative feedback is used in the receiver, the effect of the connection of the external speaker on the volume level of the loudspeaker in the set will not be very noticable at low or medium levels, although the same problem arises in regard to the overload level.

Changeover Switch

It is possible to modify the previous arrangement by means of switches which can open-circuit the voice coil of the first speaker at the same time as closing a switch in the primary circuit of the second speaker. This is illustrated in figure 2, from which it will be seen that only one speaker will be operating at the one time and there is therefore no problem with correct matching. With this arrangement, the extension speaker should have the same impedance as the one in the set and the power input to both speakers will then be equal. The two switches S1 and S2 would, of course, be combined into a single wafer switch.

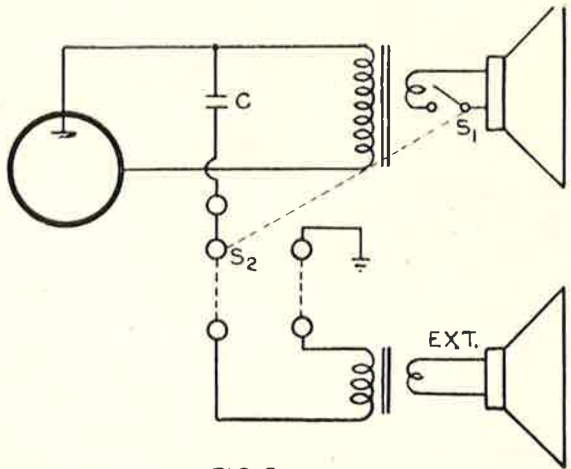


FIG. 2.

Voice Coil Extension Circuit

The previous arrangements have all adopted an extension from the primary of the speaker transformer and therefore at a high impedance. There are advantages to be gained in using the voice-coil circuit for the extension, as illustrated in figure 3, which is the equivalent of figure 1 except that it has a much lower impedance. This avoids the necessity for a step-down transformer on the extension speaker and for a blocking condenser. To obtain the same power from both speakers, it is necessary for the voice-coil impedances to be

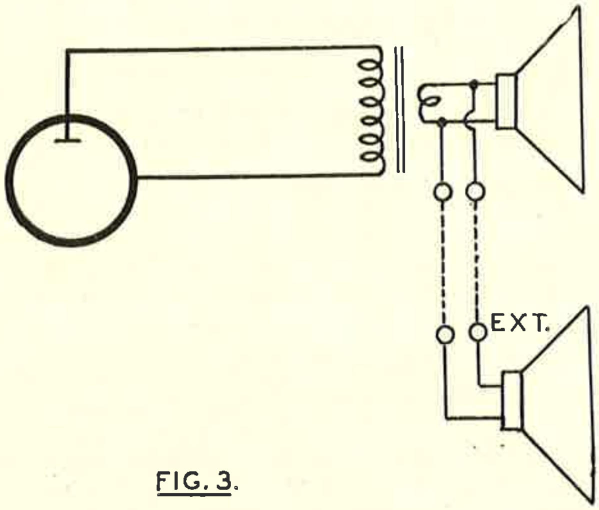


FIG. 3.

equal. If it is required to have one speaker operating at a higher level of sound than the other, the impedance of the second speaker should be made higher than that of the first speaker. In such a case it is possible to obtain correct matching by calculating the impedance of both voice-coils in parallel and selecting a step-down transformer to suit.

Very few voice-coils have an impedance less than about 2.2 ohms, so that it is possible to use quite ordinary wiring in the connections of the extension speaker. If the second speaker is to be situated more than say 10 feet from the first speaker, it may be desirable to use heavy wire, such as power flex (twin plastic is very convenient) or heavy V.I.R. cable. In some cases it may be desired to operate the extension speaker at a rather lower level than the first speaker, in which case losses in the extension line may be desirable.

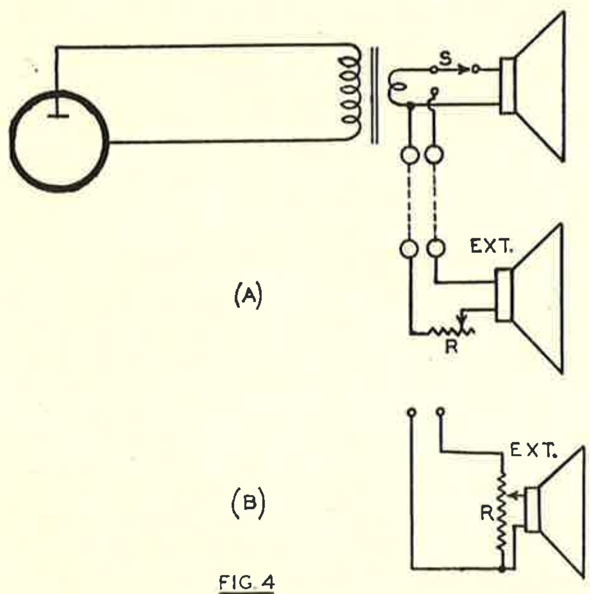


FIG. 4

Changeover Switch in Voice Coil Circuit

A single-pole double-throw switch may be used to change over from one to the other voice-coil as shown in figure 4 (a). Here switch S is used to open-circuit the first voice-coil and at the same time to close the circuit to the second voice-coil. The further refinement of a series volume control R is shown in the extension speaker circuit, so that the volume may be reduced below that of the speaker in the receiver. The resistance R should have a maximum value about 10 times that of the voice-coil impedance, but even so this arrangement cannot be used to reduce the volume to zero.

In order to have a complete control over the volume from the extension loudspeaker, the arrangement of figure 4 (b) may be used in which R is a potentiometer with the moving contact taken to the extension speaker.

The series resistor volume control shown in figure 4 (a) increases the effective impedance of the extension loudspeaker circuit at low volumes, but this is not a serious detriment since the volume will be low and the mis-matching of only minor importance, particularly if negative feedback is used. The potentiometer method of volume control in figure 4 (b) has to be a compromise, and is incapable of giving satisfactory matching under a wide range of conditions. A reasonable compromise for the resistance R would be about five times the impedance of the voice-coil, but this will result in appreciable loss of power even at maximum volume. For perfect matching, resistance R should be taken into account, but for many purposes the effect may be neglected provided that R is not less than five times the voice-coil impedance. At low levels this arrangement has a high impedance but here again the effect will not be serious.

Choice of Three Arrangements

All the preceding arrangements have involved the use of either of two speakers or given a choice between one and both. The ideal arrangement is the one permitting the use of one or other or both. This is particularly helpful when the extension speaker is used in a different room and one may wish to operate the extension speaker alone. Tuning in may be done by switching over to both speakers, adjusting the volume level to suit the extension speaker, and then switching over so that only the extension is in operation. Then, if at any time the speaker in the receiver is required to operate, this may be done simply by moving the switch and without any interference to the extension speaker.

The same arrangement is also desirable if the extension speaker is placed in the same room. Those who have had experience of this arrangement with two speakers in the one room have commented very favourably about the apparently improved tonal qualities. The use of two speakers gives a semblance of depth in the reproduction, while it also enables good hearing conditions to be maintained over the whole of a large room. A possible arrangement of the speakers is with one close to one wall of the room and the other close to the opposite wall but situated further along, so that the two speakers are not opposite one another.

Circuits for Three Alternative Switching Arrangements

If it is desired to use a single secondary winding on the transformer it is possible to arrange two speakers as shown in figure 5 so that either or both may be operated. With this arrangement the resistor R (which should have a resistance equal to the voice-coil impedance of one of the speakers) has to be used to provide correct matching when one speaker only is operating. This means that only half the full power output is available when one speaker is operating alone; the full power output is, however, available when both speakers are operating together (switch position 1). This has the result that the switching in of the second speaker does not affect the volume level of the first speaker, the resistor R really being a dummy resistance to take the power which would otherwise be supplied to the second speaker. The two switches S1 and S2 may be made from a single wafer wave-change switch.

A preferred arrangement, which does not result in any power loss under any circumstances, is the use of a tapped secondary winding on the transformer. As shown in figure 6, the full winding on the secondary is used for one or other loudspeaker, while the tap (switch S1 position 1) is used for both operating together; this involves a more elaborate switching arrangement but one which may be justified in certain circumstances.

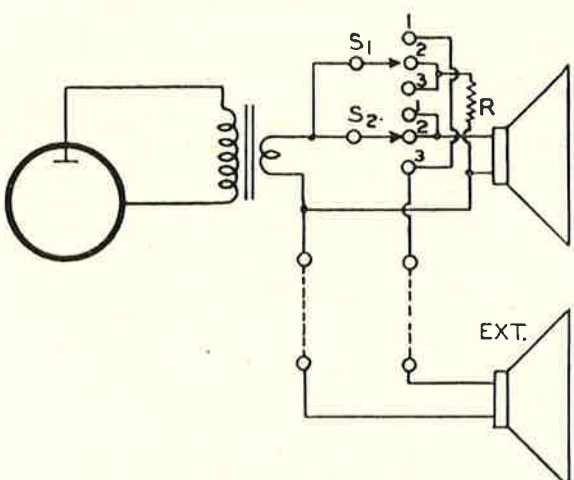


FIG. 5.

500 Ohm Line

If one or more loudspeakers are required to be operating at some considerable distance from the amplifier the preferred arrangement is to have the transformer on the amplifier stepping-down from the correct primary impedance to a secondary impedance of 500 ohms, which is connected to a line which may be of any reasonable length. At the far end of the line the input is taken to the primary of a second transformer (T2) which has a primary impedance of 500 ohms, stepping-down to a suitable secondary impedance supplying the voice-coils of the loudspeakers. A 500 ohm line is preferable, for long distances, to either a high impedance line (such as figures 1 or 2) or a very low impedance line (such as figures 3, 4, 5 or 6).

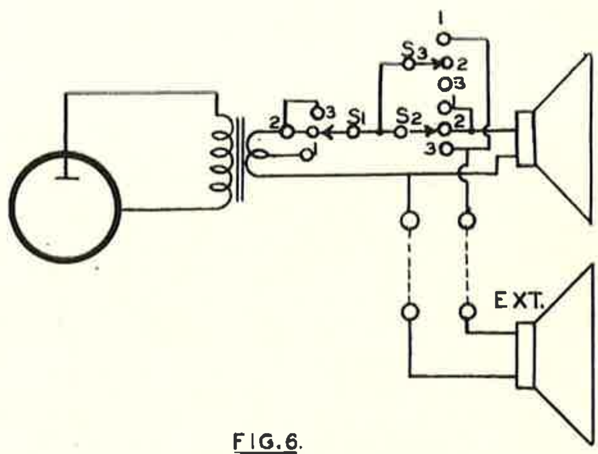


FIG. 6.

A Suggestion for Speaker Manufacturers

In order to facilitate the use of an extension loudspeaker, it is quite practicable to provide two suitable terminals on the speaker housing, connected to the voice-coil. These can be used with the arrangement shown in figure 3 for a simple form of extension speaker which does not require any additional

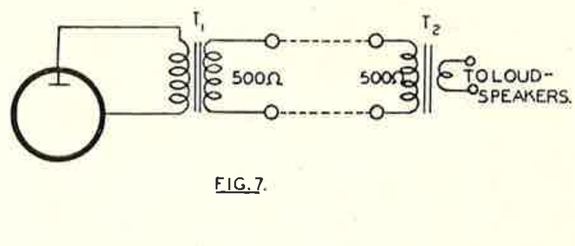


FIG. 7.

transformer. The additional cost of fitting these terminals would be quite small and the arrangement has much to commend it as an alternative to the more common method already in use, as shown in figure 1.

Radiotron 50 Watt Transmitter T202

Some years ago we designed a 50 watt 'phone transmitter using one type 809 triode in the final stage, operating on a plate input of 50 watts, with a 25 watt modulator.* The original design has been modified, the buffer (type 6P6 is no longer available) being replaced by type 807, and the 6L6-G valves in the final stage of the modulator replaced by 807's with the addition of screen "stopper" resistances and bypass condensers. The cathode keying has been changed to screen keying in order to reduce the current through the key, and to avoid excessive voltages between heater and cathode.

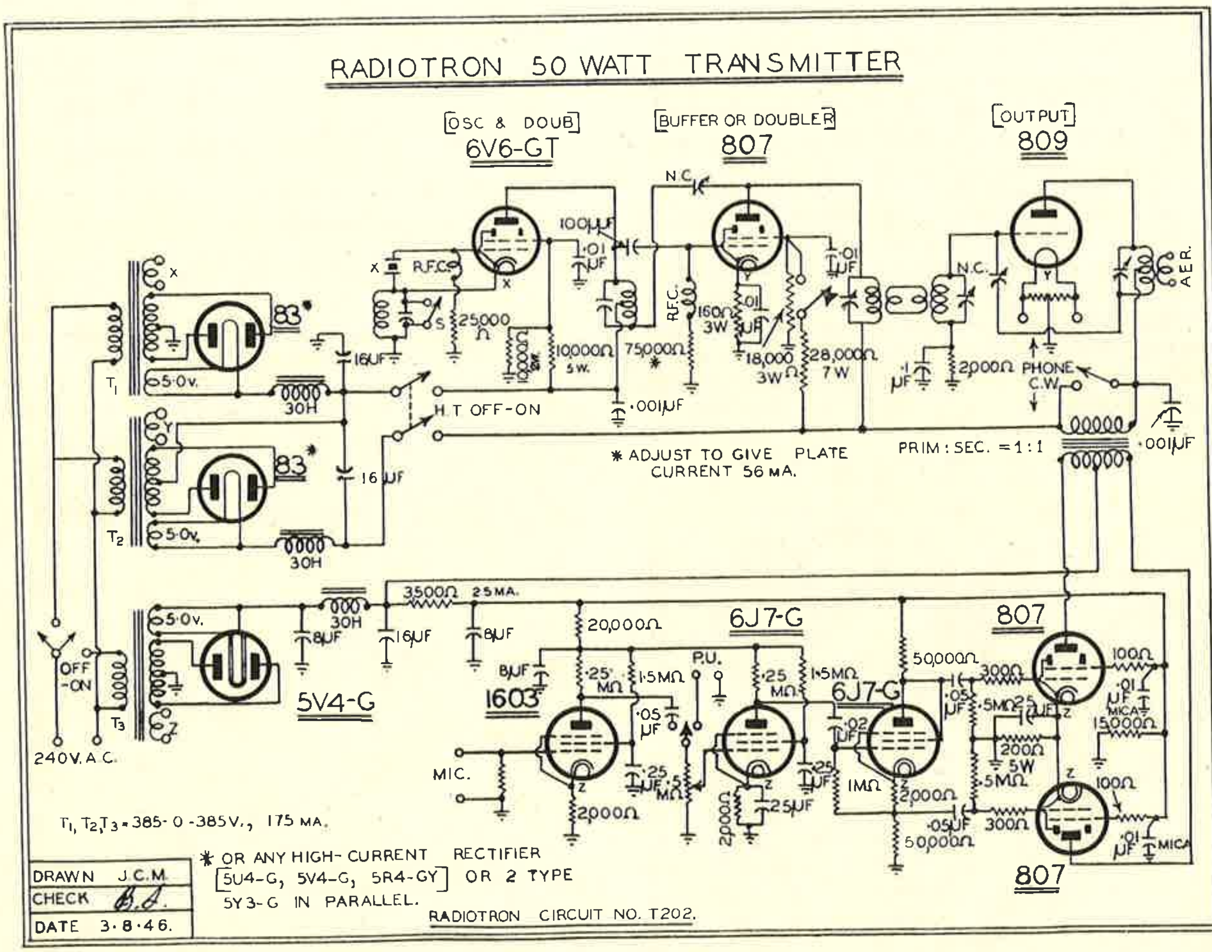
The crystal oscillator valve is type 6V6-GT, which may be used either as a straight crystal oscillator at crystal frequency, or as a tritet oscillator on the second harmonic.

The buffer valve is type 807, neutralized to allow stable operation under these conditions, but is also capable of operating as a doubler with sufficient power output to drive the final stage.

The **final power amplifier** is type 809 operating on a plate voltage of 600 volts, plate current 83mA. and power input 49.8 watts. Under these conditions the typical power output is 38 watts, which is capable of being modulated 100% without exceeding the valve ratings (CCS).

Two receiving type full wave rectifiers are used with the output voltages in series, thus providing 300 volts for the 6V6-GT oscillator, and 600 volts for the buffer and power amplifier.

The modulator is a conventional 25 watt amplifier using type 807 valves in class AB_1 with a plate voltage of 400 volts supplied by a separate 5V4-G rectifier. The 807 valves are fitted with both grid and screen "stoppers" to ensure complete stability under all conditions.



* Circuit T125, described in Radiotronics 84, page 110.

The Ratio Detector

One of the most interesting of recent developments in the field of F-M receivers is the Ratio Detector. It has called forth much enquiry and comment, since its use appears likely to reduce the minimum number of valves in F.M receivers. In order to make available all authoritative published data on the Ratio Detector, we are pleased to reprint below, by the kind permission of the Radio Corporation of America, two reports by the R.C.A. Laboratories Industry Service Division.

Ratio Detectors for F-M Receivers

R.C.A. LABORATORIES REPORT LB-645

Foreword

The underlying principle of most circuits for F-M detection has been the peak rectification of two intermediate frequency potentials whose relative amplitudes are a function of the applied frequency, together with means for combining the rectified voltages in reversed polarity. This produces a difference potential which is proportional to the difference of the two applied i-f potentials and thus is a function of the instantaneous applied frequency.

The disadvantage of this type of F-M detection lies in the fact that either an increase or a decrease in the applied i-f signal from which the two reference voltages are derived results in a corresponding change in the two rectified potentials and thus in a change in the difference (output) potential. That is simply another way of saying that all F-M detectors utilizing those principles are also detectors of any amplitude modulation which may exist simultaneously with the frequency modulation. This has, of course, been recognised and has resulted in the use of limiter circuits which tend to denude the applied signal of extraneous amplitude variations.

If a given fixed d-c voltage (or current) could, by some circuit means, be split into two parts such that the ratio of the amplitudes of those parts was always equal to the ratio of the two developed i-f voltages (in a present type discriminator) an ideal frequency modulation detector would result.

Such a system would be immune to all amplitude variations. However, a little thought given to one case wherein the mean signal level drops nearly to zero will show that either a very small sum voltage would have to be used or else considerable transconductance (gain) supplied between the two reference i-f potentials and the varying portions of that voltage. The circuits shown herewith act as ratio detectors and at the same time supply a good answer to this problem by causing the sum voltage to be a function of the mean applied signal level.

The descriptions given are general rather than specific since possible variations, both in connections and constants, are so numerous that only a small start has been made toward an examination of all of them at this writing.

General Description

Consider the circuit of Fig. 1 which shows a battery in series with two diodes, each of which can be supplied with signal energy through one of a pair of like transformers. The meter "I" will indicate the magnitude of any direct current which flows and the two voltmeters "E_1" and "E_2" will show the direct current potentials across each diode. Notice that the sum of E_1 and E_2 must always be

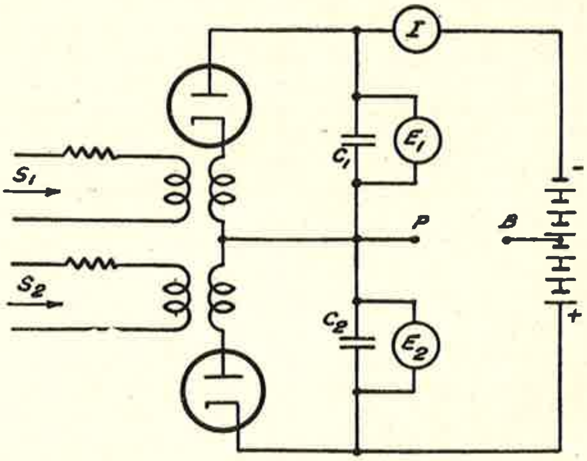


FIG. 1

equal to the battery voltage. Notice also that the polarity of the battery is such that no current will flow in the absence of applied signals if the battery voltage exceeds the sum of the contact potentials of the two diodes.

If, in this circuit, signals of equal energy are supplied to the two diodes through their respective transformers in sufficient amounts to cause current to flow, the two voltages E_1 and E_2 will be equal and the potential of point "P" will be the same as that of the center tap on the battery, point "B". Then if the two applied signals are both increased, but still kept equal in amplitude, the current I will increase by an amount greater than the percent increase in the applied signals. Also, current will flow through the diodes throughout a greater phase angle of the applied voltages, the peak conducting drop across the diodes will be greater and there will be greater potential drops in the transformer circuit impedances. However, E_1 and E_2 will not have changed and point "P" will still be at the same potential as "B".

If now, the values of the applied signals S_1 and S_2 are altered so that $S_1/S_2 = \frac{1}{2}$, $E_1 + E_2$ will still be equal to the battery voltage, but E_1 and E_2 will be in the ratio of 1:2, or in other words, E_1/E_2 will equal S_1/S_2 and that ratio will hold over rather wide limits in the amplitudes of the applied signals.

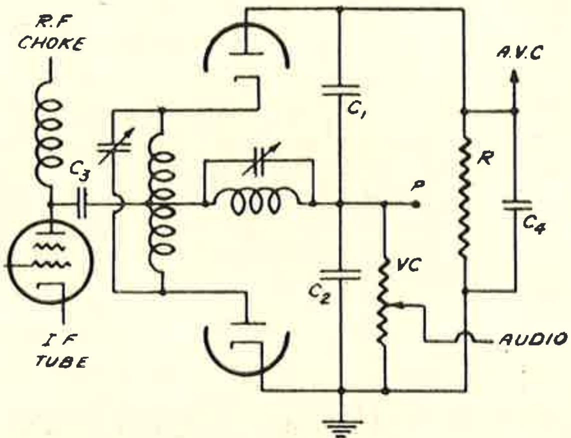
Assume now for the moment that

$$\frac{E_1}{E_2} = \frac{S_1}{S_2} = \frac{1}{2} \text{ and that}$$

by some means S_1 and S_2 are instantaneously removed and then re-applied. All conduction through the diodes would stop during that period, but the condensers C_1 and C_2 would maintain the potential of point "P" over that short interval and no variation in the output potential P-B would result.

Refer now to Fig. 2. Here the pair of input transformers has been replaced by a familiar discriminator network and the battery by a resistance and audio by-pass condenser. An output volume control is connected from point "P" to ground. The by-pass condenser C_4 may be a low voltage electrolytic of 5-10 mfd. The total capacity of $C_1 + C_2 + C_3$ can be sufficiently large to produce normal de-emphasis with the network output impedance. The latter is apt to be quite low and will be found to be a function, among other things, of the value of the resistor R. The volume control resistance is preferably sufficiently high to produce a time constant with condensers $C_1 + C_2 + C_3$ of the order of .002 seconds or more, in the absence of any conduction in the diodes.

As noted from Fig. 2 negative AVC voltages can be derived from the ungrounded end of the R — C_4 combination. That same voltage, which is stabilized by C_4 against audio frequency variations due to spurious amplitude modulation, is automatically adjusted to the correct value for any strength of applied



APPROXIMATE VALUES

C_1, C_2 = .004 μf — C_4 = 8 μf ELECTROLYTIC

C_3 = .0001 μf — R = 30,000 Ω

VC = 1 MEG.

FIG. 2

signal within the limits allowed by the application of the AVC voltage to all controllable tubes. The tube immediately preceding the discriminator is a normal i-f amplifier giving full output except as it is controlled by the AVC action.

Since a circuit of this type is relatively immune to amplitude modulation, it is unnecessary to precede it by a limiter stage. Also since its immunity is not a direct function of the signal strength, there is no threshold action of the type encountered when limiters are employed.

The stronger stations produce more output than the weaker ones for a given volume control setting exactly as in a conventional A-M set using AVC. However, even the very weakest signals have the same degree of freedom from corresponding degrees of amplitude interference.

As a consequence, a receiver utilizing this circuit need not have the extremely high i-f gain which is necessary for satisfactory limiter action. As little as ten to twenty millivolts at the grid of the discriminator driver tube is sufficient if adequate audio gain follows the detector. In this connection, it should be pointed out that while the circuit of Fig. 2 is, in effect, a voltage doubler as far as AVC voltage is concerned, the circuit does not produce as much audio output for the same signal applied to a given i-f driver tube as would be obtained from the conventional type discriminator. However, this sacrifice in output is not greater than the sacrifice in gain on the usual limiter as its threshold in comparison with a full gain i-f tube.

The value of the resistor R will affect the sensitivity of the circuit. Too low a value causes too great a loss of sensitivity. On the other hand, too high a value allows the peaks of interference or of spurious amplitude modulations to determine the value of the developed d-c voltage and will result in noisy, distorted reception of some signals.

One method for determining the desired size of the resistance R is to measure the open circuit voltage across the condenser C_4 with the resistor removed and a given signal applied to the discriminator. This can be compared with the voltage developed when a given resistance is inserted. If the ratio of these two voltages is designated by K, an incoming signal could drop instantaneously to a value of $^1/_K$ before the diodes would cease to conduct and distortion or noise result. Design considerations and tests of the complete receiver under varying noise conditions will dictate the proper value of K.

Exact mathematical analysis of the operation of a circuit of this type is extremely difficult if all factors are taken into account and is meaningless unless they are. For instance, it has been found that amplitude variations of relatively small amounts of even harmonics of the i-f signal produced by the preceding amplifier tube may, under some conditions, cause more disturbance than corresponding variations in the fundamental frequency. Incidentally, discovery of this fact led to an investigation of the effects of such harmonics in a normal "difference voltage" discriminator, and while the effect was less for a given amplitude, it was found to be more troublesome due to the relatively large amounts of such harmonics developed by the preceding limiter.

While many sample receivers employing variations of this circuit have been built and tested in comparison with conventional sets, the difficulty of exact circuit analysis by mathematical means led to the construction of "scaled down" models to operate at much lower frequencies than those used for the i-f of F-M receivers.

The first samples were built to operate at 8.25 Mc so a scale model was bread-boarded duplicating all reactive and resistive impedances at one twenty-fifth the frequency, or 330 Kc. In this circuit a deviation of 3 Kc corresponded to 75 Kc in the 8.25 Mc case. Care was taken to augment all tube and circuit element capacities, match the coefficients of magnetic and capacitive couplings, make the fundamental Q's of the circuits the same, etc.

With this set-up vacuum tube voltmeter and oscilloscope probes could be disposed rather freely about the circuit without disturbing its characteristics and a great deal of valuable information was obtained.

For instance, it was found, peculiarly enough, that amplitude variation accentuation of the primary voltage was occurring with incremental changes in the signal applied to the driver tube over part of the amplitude characteristic. This was due to the fact that the optimum coupling for the mean amplitude was such that the primary impedance actually increased as signal increments caused increased secondary damping. This, of course, means that some readjustment of the primary-secondary phase relations is bound to take place coincidentally with signal amplitude variations and accounts for some of the peculiarities of the output characteristics shown in Fig. 3.

Curves of this type can be taken point-by-point by disconnecting the volume control and placing a voltmeter between the point "P" and a center tap on R and shunting the R-C_4 combination with a battery to fix the voltage at the value reproduced when the

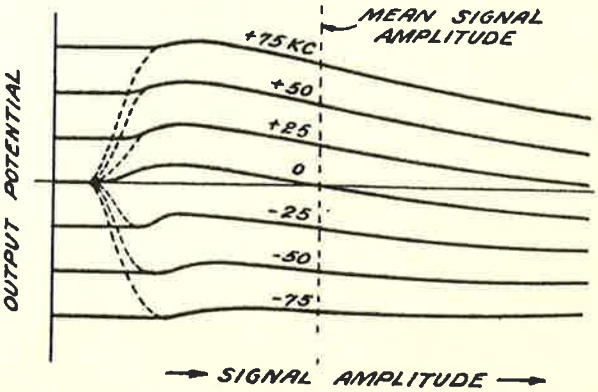


FIG. 3

mean value of the center frequency signal (indicated by the vertical dashed line of Fig. 3) is applied to the driver grid. Under these conditions the curves will converge to the mean output potential as shown by the dotted lines. If, however, the applied signals are 100% amplitude modulated, an oscilloscope* will reproduce the curve for each frequency setting as shown by the solid lines. This is more indicative of actual operation. The difference is due to the fact that the i-f by-pass and de-emphasis condensers C_1, C_2, and C_3 of Fig. 2 maintain the output potential during the short interval when the signal is below the diode conduction point. On the other hand, when taking static characteristics, the resistance of the voltmeter will be sufficient to discharge those condensers to their average potential in the absence of signal.

Two characteristics of the family of curves of Fig. 3 are worthy of note. The zero deviation curve is not straight and there is convergence for all deviations as the amplitudes are increased. As a point of interest, however, one sample receiver using a ratio detector, whose characteristics were even more distorted than those shown in this figure, was tested side by side with several cascade limiter receivers and found to equal or better the average noise reducing properties of each of them.

The variations of the zero deviation curve with amplitude were found to be due largely to the development of i-f harmonics in the driver tube and to dissimilar impedances to such harmonics developed in the two diodes.

* The horizontal deflection for the oscilloscope should be obtained from the signal generator modulating voltage.

Since only even harmonics can cause dissimilarities in the positive and negative halves of an applied wave and even the fourth harmonic is too high to produce much voltage across the lowered impedances of the tuned circuits, degeneration of the second harmonic in the cathode of the driver tube was tried to determine its effect.

The result was a decided straightening out of not only the zero deviation curve but of all others as well. Peculiarly, however, listening tests did not seem to indicate that the required additional components and added complexity of maintaining the low L-C circuit in the cathode properly aligned to the second harmonic were compensated by improved results. It was felt that added coupling means between the driver plate and the relatively low primary impedance (even with high L-C circuits) could raise the overall gain of the system and at the same time operate to further reject harmonic energy. Variations of all the possibilities in this direction have not as yet been tried.

Another circuit, almost completely immune to the effects of i-f harmonic energy, was examined in the "scaled down" set-up. This is shown in Fig. 4. Each of the two vector voltages is rectified in both phases in this arrangement so that dissimilarities of the positive and negative halves of the wave cannot alter the output potential. The resistor R of this circuit must be four times as large as in the two diode case to produce an equal effect on the tuned circuits. The audio output and A.V.C. voltages are both doubled over those derived from the circuit of Fig. 2. As a matter of fact, the contact potentials of the four diodes in series in this circuit may be too great for a zero signal A.V.C. voltage under some conditions. This might be remedied by grounding the center tap of R, by-passing each part separately and using only the negative half of the developed voltage for the A.V.C. string. With such a connection the resistance of the volume control which connects from "P" to "B" must be many times greater than one-half of R. Otherwise the A.V.C. voltage will maximize at one side of center.

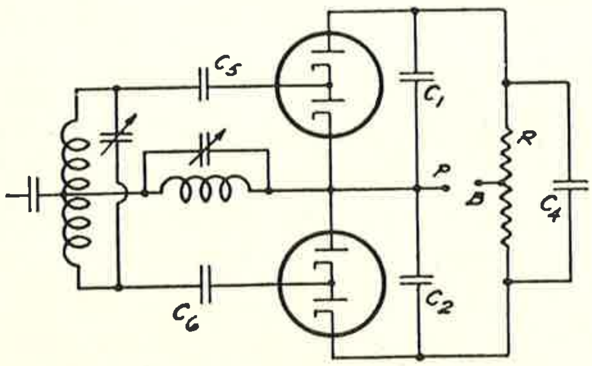


FIG. 4

The family of curves for this circuit are straight but still tend to converge with increasing amplitude. An interesting means for eliminating this convergence and making them a group of almost exactly parallel straight lines is to insert a predetermined resistance in series with the audio by-pass condenser (or battery for test purposes). This can allow just enough variation in the sum voltage to make up for variations of ratio caused by the action of differing amounts of damping on the secondary or phase shift circuit. If the center tap of the resistor R is grounded and two by-pass condensers are used it will, of course, be necessary to include an extra resistance in series with each.

The practical value of this expedient has not been fully determined. That evaluation must be made for each design case since so many other factors in a receiver can mask improvements in detector performance.

Operational Characteristics

An F-M receiver using a detector of the type described need not have the high i-f gain ahead of the tube preceding the discriminator since there is no fixed threshold level to which signals must be amplified for quieting. By the same token the standard practice of rating an F-M receiver in terms of the signal required to produce 20 db of quieting becomes meaningless.

It is obvious that the A.V.C. action requires careful design and use of expedients to prevent variations in the transit time input capacity of the amplifier tubes from causing excessive detuning action. What has not been as obvious nor as well recognized in the past is the fact that these same expedients will also prevent a burst of interference from being phase modulated by tube input capacity variations. An interfering impulse that has been subjected to a dicerent integrated reactance over each cycle of its envelope is bound to cause response in an F-M detector regardless of how its amplitude variations are subdued. A receiver which does not amplify those pulses to an i-f level at which excessive input capacity variations are encountered over the cycle is thus bound to be quieter than one with more i-f gain. This fact plus the increased freedom from regeneration difficulties both made for improved performance in a receiver which needs less high frequency amplification.

The use of A.V.C. action not only aids in keeping interference below the critical level at each tube grid, but also materially lessens the possibilities of cross modulation of a signal by noise or another signal in early stages of the receiver before it has been subjected to the full selectivity of all circuits.

The laboratory receivers using ratio detectors have generally used but one i-f stage preceding the discriminator driver although they have, without exception, included one r-f stage ahead of the converter. The fact that the discriminator drivers have more gain than a conventional limiter, plus the fact that more than the usual amount of audio gain was available, made these sets comparable in sensitivity to any of the commercial receivers with which they were compared. Actually, their most outstanding performance was with signals of less than 50 microvolts. Often these were badly submerged in noise on most of the receivers.

Another important characteristic of these detectors is the manner in which they behave when tuning across an incoming signal. Since the ratio of the discriminator voltages decreases after passing a peak on one side of resonance the side responses are present. However, the high degree of amplitude modulation with which an off-resonant F-M signal becomes endowed as it is modulated up and down the side of a sharp selectivity curve, produces little or no effect on a ratio detector. As a consequence the two side responses are very much subdued in comparison with the true peak reception and would seldom, if ever, be mistaken for the correct tuning adjustment. This factor causes such a receiver to appear to have more selectivity than an equivalent receiver using the limited-discriminator combination.

Design Considerations

The following generalized principles will usually apply in the design of ratio detectors.

The primary-secondary coupling of the discriminator will need to be closer than in the conventional type. If desired, for experimental work, the two can be completely uncoupled magnetically and a small adjustable condenser connected from one secondary terminal to ground to give capacity coupling.

Complete balance of both halves of the network is essential for best results. Slug tuning of the secondary should not be adopted unless it is arranged so that it cannot unbalance the secondary inductance.

Generally speaking, discriminator linearity is not as easily achieved with ratio detectors as with the conventional type, and variation of the transformer coupling cannot be used as readily to affect the linearity as was formerly the case.

For purposes of r-f and i-f alignment the audio by-pass condenser can be disconnected from across the resistor. A sweep signal generator will then reproduce the over-all selectivity curve at the ungrounded end of the load resistance for oscilloscope observation.

Balanced Phase Shift Discriminators

R.C.A. LABORATORIES REPORT LB-666

Foreword

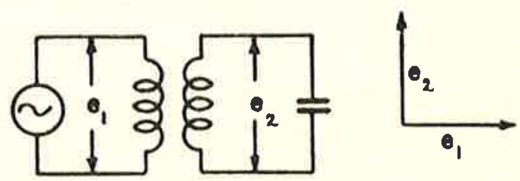
This bulletin provides a simplified means for predicting the relative behavior of the phase shift networks of balanced discriminators. The first portion is general while the latter paragraphs deal specifically with Ratio Detectors and means for maximizing their linearity and sensitivity.

Coupled Circuits—Secondary Resonant

In the circuit of Fig. 1 it may be assumed (for the sake of simplicity) that both e_1 and the induced secondary voltage are in quadrature to the primary current and that the voltage e_2 is in quadrature to the induced secondary current. These assumptions will cause negligible error in making vector analysis of balanced phase shift discriminators to determine their **relative** behavior, if the couplings and Q's are in line with those generally used in such circuits.

In the following analysis the voltage across the primary is used as the reference vector. Therefore, it is immaterial whether that voltage is derived from a zero impedance generator (as shown in Fig. 1) in shunt to the primary (in which case a primary tuning condenser is of no importance) or whether it is obtained by resonating the primary and placing it in the plate circuit of a high impedance vacuum tube.

From the equations of Fig. 1, it can be seen that the ratio of secondary to primary potentials, at resonance, is a function of the ratio of the mutual impedance and the primary reactance, and of Q_2. If these parameters were held constant, the ratio of primary to secondary voltage would be independent



$$i_1 = e_1 / j\omega L_1; \quad e_1 = i_1 j\omega L_1$$

$$e_2 \text{ (At Resonance)} = \frac{e_1 j\omega M j\omega L_2}{j\omega L_1 r_2}$$

$$\text{or } \frac{e_2}{e_1} = \frac{jQ_2 M}{L_1}$$

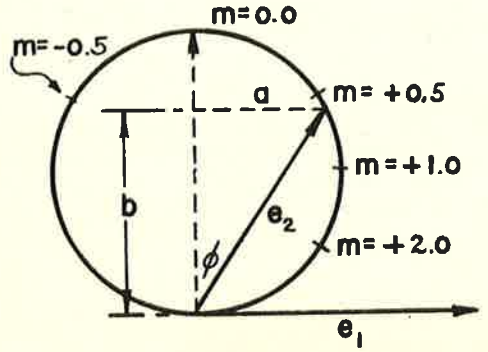
FIG. 1 — COUPLED CIRCUITS — SECONDARY TUNED.

of the secondary L/C ratio, although normally, changes in secondary inductance will affect the magnitude of the mutual impedance.

In Fig. 2, the locus of the tip of the secondary vector, for different applied frequencies, is shown to be a perfect circle. The ratio of the components, a, b, of the secondary vector is equal to the ratio of the series reactance to the series resistance of the secondary circuit and is thus equal to the tangent of ϕ, the secondary phase angle. The quantity a/b is assigned the notation m.

Discriminator Action—Primary Voltage Constant

In Fig. 3 the secondary circuit is center tapped and the tap connected to the voltage end of the primary. This is the basic diagram of balanced phase shift discriminators. The vector diagram represents the conditions when the secondary is resonant to the applied frequency. The potentials at points A and B (with respect to ground) are each made up of the total primary plus 1/2 secondary voltages.

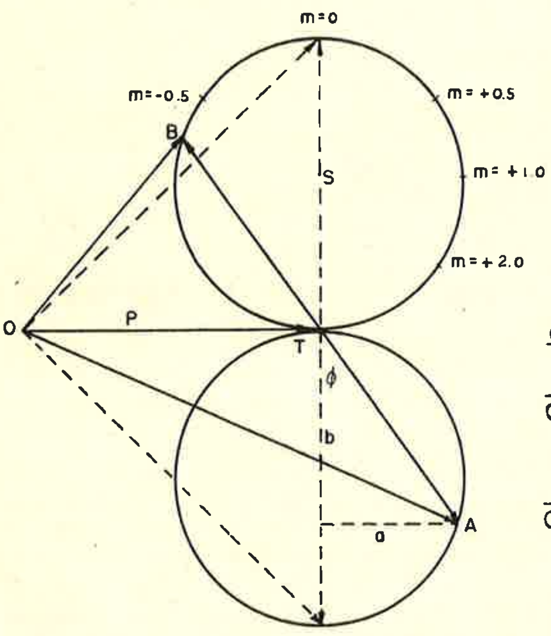


$$\frac{a}{b} = \tan\phi = m$$

$$m = Q_2\left(1 - \frac{f^2}{f_o^2}\right)$$

FIG. 2 — COUPLED CIRCUITS — APPLIED FREQUENCY VARIABLE.

Fig. 4 is the basic vector diagram for all balanced phase shift discriminators. The loci of the ends of the secondary vector are shown as two circles tangent



$$\frac{a}{b} = \tan\phi = m = Q_2\left(\frac{f^2}{f_o^2} - 1\right)$$

$$\overline{OA} = \sqrt{\frac{P^2 + P^2m^2 + 2PSm + S^2}{1+m^2}}$$

$$\overline{OB} = \sqrt{\frac{P^2 + P^2m^2 - 2PSm + S^2}{1+m^2}}$$

FIG. 4

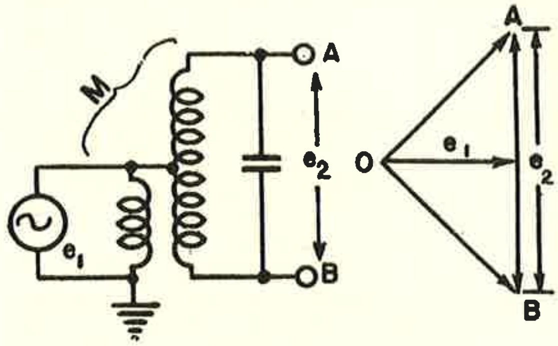


FIG. 3 — FUNDAMENTAL DISCRIMINATOR CIRCUIT.

to the tip of the primary vector P. The resonant secondary vectors are shown dotted. When the applied frequency departs from resonance, the secondary vector assumes some position such as that shown by AB.

The two discriminator potentials OA and OB are then of unequal length and the difference can be seen to be a function of m as it varies between its maximum positive and negative operating values.

The first derivative with respect to m, of the expression for $\overline{OA} - \overline{OB}$ at m = O, can be maximized to determine the optimum ratio of primary to 1/2 secondary voltage for a given length of resultant. This operation indicates that if a discriminator sensitivity is dependent only on the rate of change of the difference between $\overline{OA}$ and $\overline{OB}$, the resultants should be made up of equal primary and secondary components. In other words, the total secondary voltage should be equal to twice the primary voltage. The diagram of Fig. 4 is drawn to represent this condition.

Again it should be emphasized that the manner in which the primary potential is obtained has **no** effect except on the size of the diagram of Fig. 4. If the primary were tuned by a variable condenser and placed in the plate circuit of a high R_p tube, variations of the primary tuning capacitor would cause the entire diagram to be enlarged or reduced exactly as though by photographic processes. No changes in the **relative** lengths or phases of any of the vectors would occur as the primary was tuned through resonance. However, note that the quantity $\overline{OA}$ minus $\overline{OB}$ is proportional to the "size" of this diagram and if the output of a discriminator is a function of that difference, the audio output voltage will be altered correspondingly.

If the primary voltage P is maintained constant at P = 1 as the frequency is varied, and the 1/2 resonant secondary voltage also equals unity, the difference between $\overline{OA}$ and $\overline{OB}$ will be given by the expression

$$OA - OB = \sqrt{\frac{m^2 + 2m + 2}{1 + m^2}} - \sqrt{\frac{m^2 - 2m + 2}{1 + m^2}}$$

This is plotted in the curve of Fig. 5, together with the curve of its first derivative, or slope. The dotted line has the slope of $\overline{OA} - \overline{OB}$ at m = O.

Obviously, the lack of linearity of $\overline{OA} - \overline{OB}$ would restrict its use as a discriminator characteristic to the portion of the curve lying beween m = + 0.4 and m = — 0.4 for approximately 25% **variation** in slope. Then from the expression

$$m = Q_2 \left[\frac{f^2}{f^2_0} - 1 \right]$$

it can be determined that the secondary Q could not be over two times the center frequency in megacycles to provide a 200Kc. band between those limits. An operating Q_2 of 21.4 for a 10.7 megacycles discriminator is, of course, abnormally low.

This example might be indicative of the operation of a circuit which provided primary **voltage** limiting (or one with a perfectly flat topped primary resonance curve) and in which the primary and 1/2 secondary resonant voltages were made equal.

Discriminator Action—Primary Voltage Not Constant

As stated above an increase or decrease in the magnitude of the primary vector (the applied voltage) causes the whole vector diagram of Fig. 4 to be enlarged or reduced accordingly. Therefore, any variation (amplitude modulation) of the primary voltage causes the magnitude of the difference between $\overline{OA}$ and $\overline{OB}$ to be altered proportionally.

In Fig. 5 the ordinate of the curve of $\overline{OA} - \overline{OB}$ at m = + 1.0 is 0.873. The ordinate of the dashed line at m = + 1.0 is 1.414. Therefore, if the primary voltage were made a function of m such that it increased as m departed from zero in either a positive or negative direction and the increase amounted to 1.414 ÷ 0.873 at m = + 1, the curve for $\overline{OA} - \overline{OB}$ would coincide with the dashed line at those points. Proper shaping of the amplitude characteristic between those values of m, could cause the curve $\overline{OA} - \overline{OB}$ to become perfectly linear throughout the range. The relative primary voltage (as a function of m) to produce this result is shown in Fig. 6. A typical double-peak primary resonance curve is shown superimposed for comparison.

If the primary is resonated with a condenser and both Q_1 and the coupling are properly adjusted, the primary impedance characteristic can be made to match the desired primary voltage curve. Then, if the network is fed by a constant current source, such as a high R_p tube, the difference voltage ($\overline{OA} - \overline{OB}$) will be almost perfectly linear between m = + 1.0 and m = — 1.0 as shown in Fig. 7. This is usable as a discriminator characteristic over a range of m which is 2½ times as great as the fundamental curve of Fig. 5. The secondary Q for this condition should be 5 times the operating frequency in megacycles instead of 2 times and since m is a function of Q_2 the sensitivity would be 2½ times as great as in the constant primary voltage case.

Thus the introduction of the proper amount of amplitude variation (as a function of m) into the primary voltage of a discriminator network can serve to improve its linearity (and allow greater sensitivity) **if its output is dependent only on the difference in the magnitude of the two derived voltages.**

In both of the examples given above the output audio potentials can be derived by rectifying the two resultant voltages and adding the unidirectional potentials in series opposition. This is the normal circuit connection when the current limiters are used.

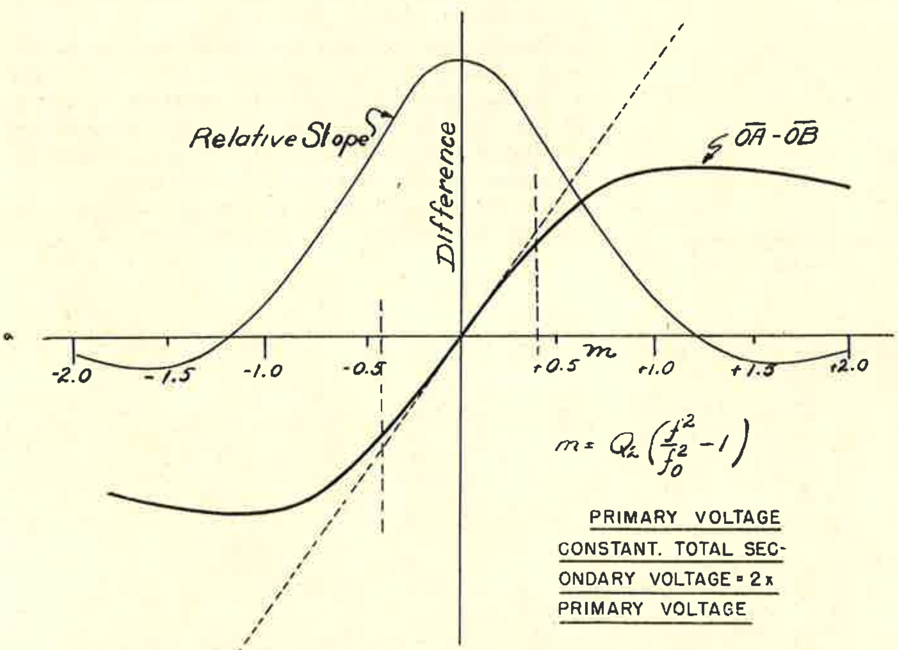


FIG. 5

Ratio Detector—Primary Voltage Equal to Half Secondary Voltage.

In Ratio Detector circuits (such as described in LB-645), the output potential is a function of the

ratio $\overline{OA}$ and $\overline{OB}$. In such circuits the output is given by the expression:

$$\text{Output} = K\,\frac{\dfrac{\overline{OA}}{\overline{OB}} - 1}{\dfrac{\overline{OA}}{\overline{OB}} + 1} = K\,\frac{\overline{OA} - \overline{OB}}{\overline{OA} + \overline{OB}}$$

where K is an arbitrary constant.

If the amplitude variations cause proportionate changes in both the **sum** and **difference** of $\overline{OA}$ and $\overline{OB}$ the output remains constant.

Substituting the values for $\overline{OA}$ and $\overline{OB}$ (when P = S) in the above expression gives

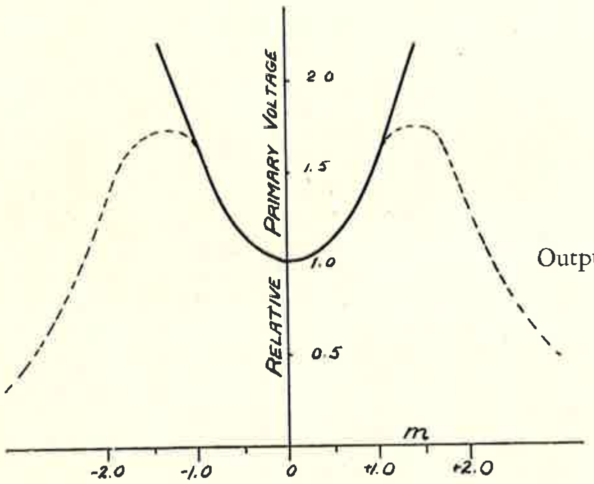
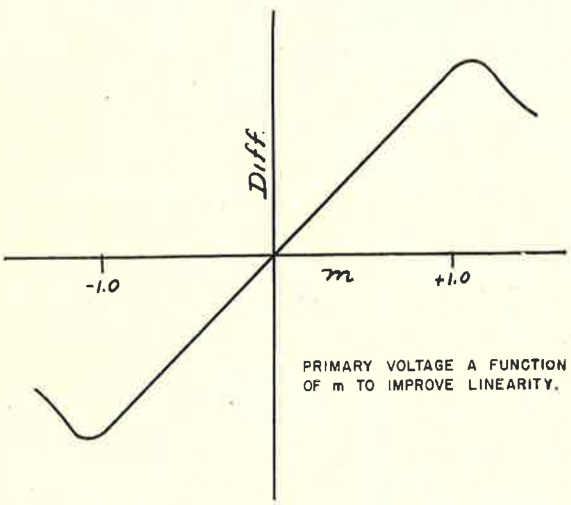


FIG. 6



PRIMARY VOLTAGE A FUNCTION OF m TO IMPROVE LINEARITY.

FIG. 7

$$\text{Output} = K\,\frac{\sqrt{2 + 2m + m^2} - \sqrt{2 - 2m + m^2}}{\sqrt{2 + 2m + m^2} + \sqrt{2 - 2m + m^2}}$$

A plot of this characteristic, together with the derivative with respect to m, is shown in Fig. 8.

If the useful portion of this curve is assumed to be that over whch the slope varies 25%, the limits are m = + 0.6 and m = — 0.6 which would provide a 200 Kc. band in a 10.7 Mc. discriminator if the secondary Q were 32.

This is better than the constant voltage primary case but it does not provide either the linearity or the deviation sensitivity of the curve of Fig. 7. Furthermore, the series additive connection of the Ratio Detector diodes produces an inherent 6db loss in gain compared to the series opposition connection of the diode load resistors.

Since Ratio Detectors are relatively immune to amplitude variations of the applied signals the introduction of such variations as a function of m will not produce the same effect as in the current limited case. However, one other parameter can be altered in the design of Ratio Detectors to provide the desired linearity.

Ratio Detectors—Primary Voltage Less than Half Secondary Voltage

So far all examples have assumed the primary voltage to be equal to 1/2 the secondary voltage at resonance. If the ratio of those components is designated by P, the Ratio Detector output becomes:

$$\text{Output} = K\,\frac{\sqrt{P^2 + P^2m^2 + 2Pm + 1} - \sqrt{P^2 + P^2m^2 - 2Pm + 1}}{\sqrt{P^2 + P^2m^2 + 2Pm + 1} + \sqrt{P^2 + P^2m^2 - 2Pm + 1}}$$

Fig. 9 is the first derivative with respect to m of this expression for m = 0 and m = 1 with the value of P as the independent variable. It is evident that the output curve can be made much more linear if P is made less than unity without undue sacrifice in the slope at resonance.

For instance, a decrease in the P/S ratio from 1.0 to 0.5 would decrease the slope at resonance to 82% of its values when P/S = 1. However, the discriminator output characteristic would now be sufficiently linear so that the usable range would extend between m = + 1.0 and m = — 1.0. For a 200 Kc. band at 10.7Mc. the secondary operating Q could then be 53 instead of 32 with a net **increase** in discriminator sensitivity of 32% or 2.5db. Furthermore, if the decreased primary voltage is obtained by tapping down on the primary, the increased impedance of the primary tuned circuit can

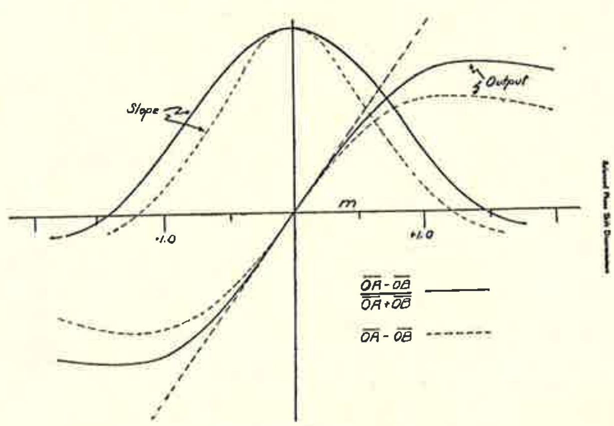


FIG. 8

result in an additional 8db to 12db gain in sensitivity. This connection is shown in Fig. 10.

If the primary voltage for the discriminator is derived from a few turns of wire tightly coupled to the primary tuned circuit, the plate choke and blocking condenser are unnecessary. This connection is shown in Fig. 11.

Fig. 12 is a tracing of the pattern on an oscilloscope screen when reproducing the discriminator output characteristics of a ratio detector with a primary voltage of approximately 1/5 the total secondary resonant voltage. The developed AVC voltage with 100 millivolts applied to the grid of a Gm-4000 driver tube was 8.2 volts and the peak to peak audio output was 3.2 volts for 75kc. deviation.

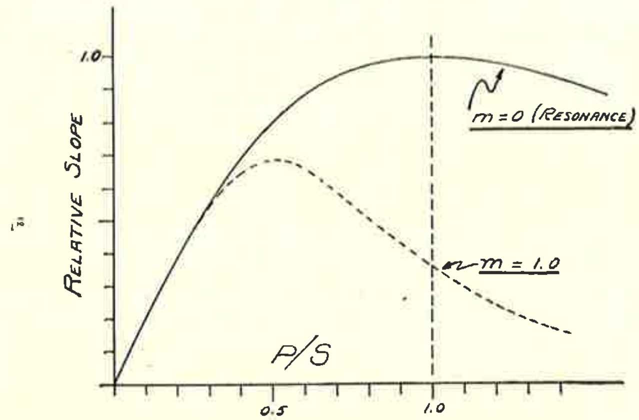


FIG. 9

The loops at the ends of the characteristic shown in Fig. 12 are caused by the sevenfold increase in the RC time constant of the output circuits when the i-f sensitivity causes the signal to drop below the diode conduction level at the edges of the pass band.

The circuit of Fig. 13 is the same as that of Fig. 11 except that the center of the d-c load resistor is grounded and the two ends bypassed with small condensers (approximately 100 $\mu\mu$fd. each). This arrangement makes for better symmetry and thus better balance of the discriminator. Also an AFC voltage which is zero on resonance is made available if desired. The ratio of AVC voltage to the developed audio output voltage is more nearly the same as that in AM receivers. The use of a coupling condenser between the audio output point and the volume control is recommended both to remove the d-c potentials and to allow the AVC voltage to maximize correctly on center frequency.

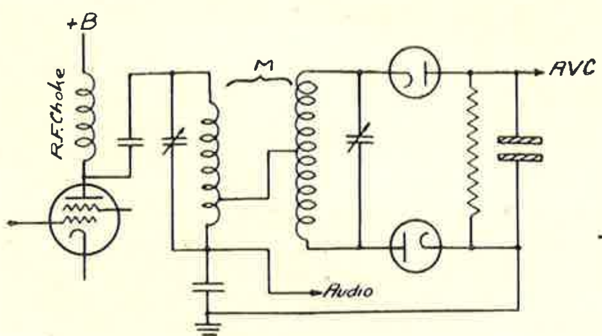


FIG. 10

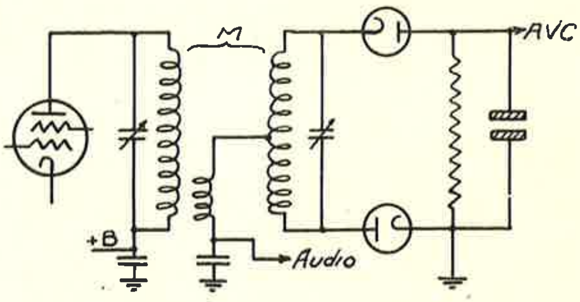
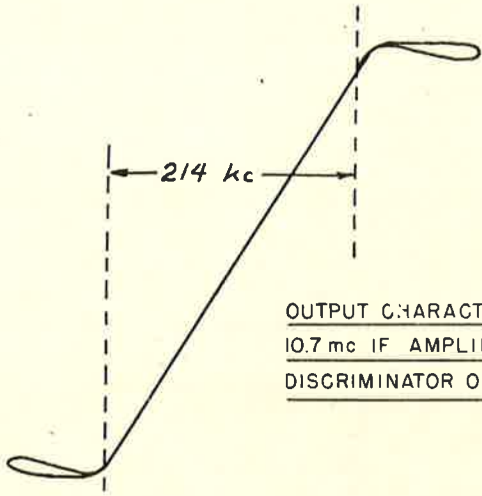


FIG. 11



OUTPUT CHARACTERISTIC
10.7 mc IF AMPLIFIER &
DISCRIMINATOR OF FIG 11

FIG. 12

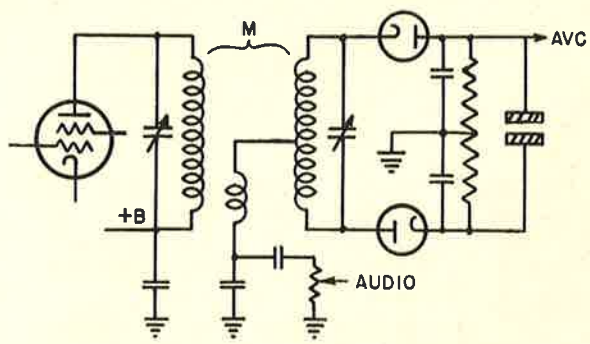


FIG. 13

The Plotting of Valve Characteristics

Special precautions must be taken to the measurement and plotting of valve characteristics if it is desired to achieve accuracy. Some of the problems involved, together with the methods used, are described in this article.

Most radio engineers have at times plotted valve characteristics for one purpose or another, but the precautions necessary for such a procedure are not usually understood to their fullest extent. The first problem is that the valve characteristics themselves are usually variable, that is to say the measured plate current under certain specified conditions does not remain constant. This is caused partly by the emission characteristics of oxide-coated filaments or cathodes, and this variation does not occur in the case of pure tungsten filament valves ("bright-emitters"), since the emission in these is purely a function of temperature. Thoriated tungsten filaments are also fairly stable and any changes which occur are comparatively slow.

Stabilisation of Emission

Most work on receiving valves at the present time is confined to those with oxide-coated emitting surfaces, so that it is necessary to take special precautions to obtain readings which are a true indication of the characteristics. To this end, every valve should be operated under its normal operating conditions, until the characteristics are stable, with the filament or heater maintained accurately at the rated voltage; the plate and screen supplies should be from a d.c. source and the grid bias should be maintained at the correct value. It is desirable to insert milliammeters into the plate and screen circuits so as to keep a check on the currents that are drawn.

At the end of this period, the emission should have become comparatively stable, provided that the valve continues under the same operating conditions. Unfortunately, it tends to be affected by operation at a higher cathode current. For this reason, it is desirable to carry out tests on cathode currents less than the normal value before proceeding to higher cathode currents.

It is usually safe to continue the measurement of characteristics up to twice the normal plate voltage, provided that the plate or screen dissipation is not exceeded.

Difficulties begin to arise as soon as the cathode current is appreciably higher than the normal operating current, and a valve which is taken up to two or three times its normal cathode current usually suffers a change in its emission characteristics so that the earlier readings taken at lower currents may not be repeated exactly. This problem is serious with both triodes and pentodes, since the characteristic curves are usually required to go up to more than twice the normal operating plate current.

Dynamic Tests

The only really satisfactory way of carrying out such tests in excess of the maximum dissipation of the valve is by the use of an dynamic test method, using pulse waves and measuring the current by means of cathode ray oscillograph. A square wave voltage pulse may be applied to the grid circuit and the resultant current measured to a fair degree of accuracy. It is necessary to use voltage-regulated plate and screen current supplies so as to avoid errors caused through the drop in plate voltage which would occur during the pulse of plate current.

At some future date we hope to go into further detail regarding equipment for the dynamic testing of valve characteristics, but this article is intended to deal with the static methods which may be used, within certain limits, with reasonable accuracy.

Equipment for Characteristic Testing

The most commonly used arrangement provides for variable adjustment of grid, screen and plate voltages, with indicating voltmeters on filament (or heater), grid, screen and plate. The grid, screen and plate should all be supplied from entirely independent sources, so that a variation of current drawn by one will not affect the others. The grid voltmeter should always be adjusted first, the screen (if any) second, and the plate voltage last. Above the knee of the pentode current curves in a pentode valve, the plate voltage has only a slight effect on the plate current, but the screen voltage is very critical. Below the knee of the curve the plate voltage also becomes critical, and additional care is required. In all cases, after the original setting of a particular point, it is good practice to recheck the grid voltage, screen voltage and then the plate voltage, before reading the plate and/or screen currents. If the screen current is required to be read at all, it should be read at the same setting as for the plate current and not at some other time, since any error in the setting will then affect both plate and screen currents and there will be less inconsistence; there will also be less time taken for the test.

Fluctuating supply voltages are very troublesome, and it is highly desirable for a regulated a.c. supply to be used for the rectifiers and filters supplying the various electrodes, unless regulated d.c. supplies are available. If the fluctuating supply is unavoidable, it is helpful to have two observers, one to read the voltages and the other the currents, the former giving the word as soon as the voltages are correct.

Instrument Errors

Errors in the measuring instruments are of the greatest importance, owing to the magnification of these errors which occur through the valves. For example, an error of 0.1 volt in the grid voltage reading will cause an error of 0.4 mA. with a valve having a mutual conductance of 4,000 micromhos. There is a further fact to be borne in mind, that instruments are accurate to a certain percentage of full scale rating; for example a 0-10 voltmeter with an accuracy of 1% of full scale, may have an error of 0.1 volt at any part of the scale. If it is used to measure a voltage of 1 volt, the possible error in the reading is 10%. For this reason instruments should always be used as closely as possible to full scale. For the highest accuracy, instruments better than 1% of full scale are necessary; this result may be achieved even with ordinary instruments by plotting a curve showing the error on the instrument for each point on the scale. These error curves should preferably be measured in an Instrument Measurement Laboratory where standard instruments are available for checking. Alternatively, if an instrument of sub-standard accuracy is available, this may be used as an approximate check on the less accurate instrument.

The procedure for dealing with instrument errors is set out below in the form of an example, treating first the triode case and later the pentode.

Triode Example

In this case the only readings are grid voltage, plate voltage, and plate current. It is helpful to keep one of these constant for a succession of readings, so as to achieve greater accuracy and avoid unnecessary work. In the case of the triode plate characteristic, the grid bias may be maintained constant for the whole of one characteristic curve, and the grid voltmeter should be set so as to give the desired grid voltage. For example, if the desired negative grid voltage is 2 volts, and the instrument error at this point is +0.1 volt, the instrument should be set to read 2.1 volts so as to give the desired 2 volts (which is the true voltage). The procedure is therefore to maintain the grid voltage precisely at this setting while the plate current is read with each increment of plate voltage. Slightly additional accuracy may be obtained by checking the readings with the plate voltmeter exactly on the scale divisions, even where there is known to be an error in the calibrations, and reading the plate current in each case. Thus the error in estimating the readings between scale divisions will only occur in one instrument, in this case the plate milliammeter.

The table below shows one suggested method of recording the readings.

Grid Voltage		Plate Voltage		Plate Current	
Rdg.	True	Rdg.	True	Rdg.	True
scale 0–4		scale 0–100		scale 0–5	
–2.10	–2.00	20	20.04	0.1	0.12
		30	30.02	0.22	0.24
		40	40.00	0.42	0.45
		50	49.95	0.65	0.68
		60	59.98	1.12	1.16
		80	79.95	2.15	2.16
		90	89.92	2.75	2.78
		100	99.0	3.52	3.55
		scale 0–300		scale 0–25	
		120	121.7	5.3	5.25
		140	142.2	7.4	7.38
		160	162.0	10.15	10.12
		180	182.2	12.75	12.71
		200	200.5	15.25	15.20

Grid Volage. Rdg.	True.	Screen Voltage. Rdg.	True.	Plate Voltage. Rdg.	True.	Plate. Current. Rdg.	True.	Screen Current. Rdg.	True.
Scale 0-4.		Scale 0-100.		Scale 0-50.		Scale 0-5.		Scale 0-5.	
—2.10	—2.00	99.9	100	5	5.2	2.3	2.32	2.3	2.27
				10	10.1	2.7	2.73	1.9	1.89
				15	15.05	2.9	2.91	1.7	1.68
				20	20.0	3.12	3.15	1.48	1.45
				30	30.1	3.3	3.61	1.3	1.27
				40	40.2	3.4	3.77	1.2	1.18
				Scale 0-500.					
				60	62	3.45	3.99	1.15	1.14
				100	101	3.52	3.56	1.08	1.06
				200	200	3.61	3.66	0.99	0.97
				300	301	3.68	3.72	0.92	0.88
				400	402	3.70	3.74	0.90	0.86

Pentode Example

In this case the necessary readings are grid voltage, screen voltage, plate current and screen current. As with the triode, the grid voltage should be maintained constant while taking readings for the plate characteristics, and the screen voltage likewise. It is then only necessary to set the plate voltage to the desired values and to read the plate and screen currents. Unless the screen supply has exceptionally good regulation, it will be found necessary to make adjustments to the screen voltage control for each reading. The table above shows a suggested method of recording the readings.

Mutual Characteristics

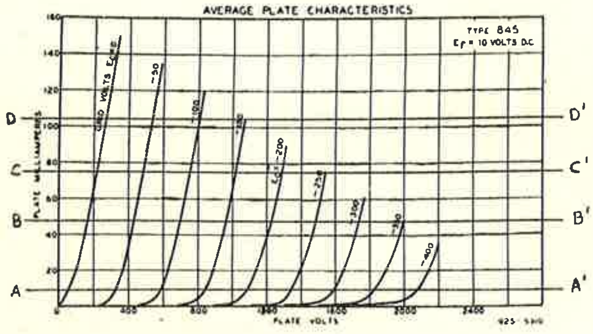
In the case of mutual characteristics, both triode and pentode, it is necessary to maintain the plate and screen voltages at constant values and to vary the grid voltage, noting the plate and screen currents in each case. As these characteristics are required to be drawn for precise plate and screen voltages, the plate and screen voltmeters should be set so that the true voltage is the desired voltage. Here again, as the variation in currents is quite large, it will probably be found necessary to make adjustments to the plate and screen voltage controls for the various readings.

Plotting

The values determined by the methods outlined above may then be plotted on squared paper. The most accurate method is to use a fine needle "pricker" for each point, and its position may be more readily identified by surrounding it with a circle in pencil. For rougher work it is permissible to use a well sharpened 4H pencil; it should be rolled between the thumb and fingers so as to give a distinct circular dot. When all the points on a particular curve have been completed, it is desirable to run a carefully drawn free-hand curve linking them before proceeding to the use of French curves, which should not be used until all the curves have been completed in free-hand. When this latter stage has been reached a careful visual examination should be made to pick out any points which seem out of alignment and any curves which appear out of position when compared with the other curves. An error in the setting or rating of one value of grid voltage will cause the whole of one plate characteristic to be moved sideways. If any curve or important point is in doubt, it should be repeated and the error corrected before proceeding to the next step.

Drawing and Checking Curves

Most French curves are made for general use and are not the most generally satisfactory shapes for



drawing valve characteristics. If many characteristics are to be drawn, it is desirable to prepare a celluloid curve specially for the purpose, but failing this the ordinary French curves may be used with suitable care. After the first curve has been drawn in this manner, the drawing should be lifted until it is almost edge-on to the eye, so that the draftsman can see any discontinuities or kinks. These are particularly likely when linking up two sections, each drawn by a French curve. When one curve has been drawn in this way, and found to be satisfactory by visual inspection, the other curves should be drawn in a similar manner, using the same portions of the same French curves for each. With only slight modifications, resulting from the slightly increased changes of curvature, a high degree of consistency can be obtained by this means between the various curves. It is now necessary to check the intercepts between the curves to see that they have been accurately drawn. It is fortunately true that the distance between two curves is very nearly constant over a fairly wide range. This applies particularly to mutual characteristics and also to triode plate characteristics. An example of the method applied to a triode plate characteristic is shown in figure 1 and it will be seen that the curves diverge very gradually as the plate current and/or plate voltage are increased.

For figure 1, the intercepts may be checked at four or five different plate currents as indicated by the quite arbitrary choice aa', bb', etc. In the case of each of these selected positions for checking, a piece of paper may be held so that it just touches the line aa', etc., and the intercepts of the curves marked with a hard pencil on a clean sheet of paper. The paper may then be moved one space further to the right, and the position of the curves noted in relation to the marks on the paper. In a correctly drawn set of curves, the curves should cut the paper very close to the intercepts marked in pencil, there being a slight tendency for the intercepts to become shorter at the high voltage end of the characteristics at low plate currents. The piece of paper is then folded over so as to expose a new clear edge for the next higher checking line bb', and the same general procedure adopted, and so on for cc', dd'.

An additional check may be made by using another portion of the clean paper which should be placed roughly at right angles to the two extreme left-hand curves, and about half-way up the curves. The intercepts of these two curves should be marked on the paper and the paper then moved both upwards and downwards to the extremeties of the curves to see whether the distance apart of the two curves is nearly constant. It will be found that the curves diverge slightly more at the high end but converge rapidly at the lower end. The same treatment could be given to other pairs of curves.

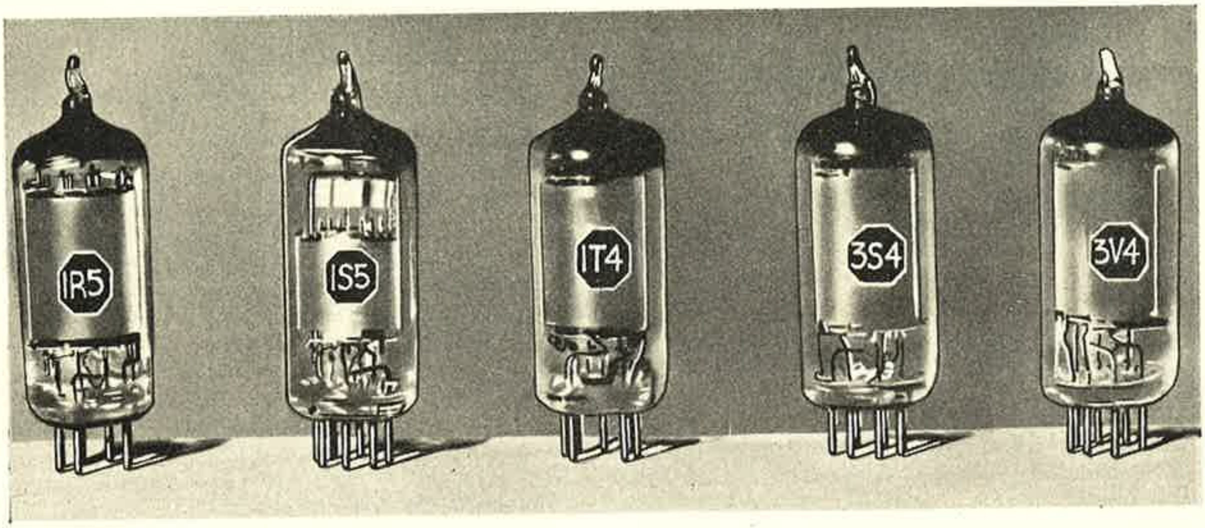
A still further check may be given by using two celluloid squares. One of these should be set so as to be a tangent to the extreme left-hand curve and the other placed so as to permit the square to be moved so as to be parallel to its original position. The plate current corresponding to the point of tangency should be noted mentally and the square run across the curve, the plate current for the tangent point being noted in each case. It will be noticed that, with correctly drawn curves, the plate current for equal slope increases as the negative grid voltage is increased, but the change is quite gradual and should be constant between all the curves. Some similar method may be used on the mutual characteristics and on the dynamic characteristics of resistance coupled triodes or pentodes.

The plate characteristics of pentodes require a different technique. In this case the curves above the knee are nearly straight, although their slopes become more pronounced as the grid voltage approaches zero.

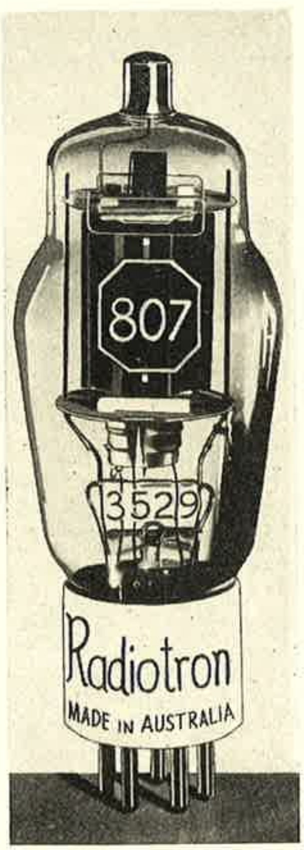
The Choice of Average Valves

If it is desired to publish the curves of an average valve, it is first necessary to record all the significant characteristics of some hundreds of normal valves. From these recorded ratings it is necessary to take the average of plate current, screen current, mutual conductance, amplification factor (or triode amplification factor in the case of pentode valves) and emission, and then to select the valve from among all those tested which has characteristics nearest to the average. If such a valve could be found, it is then only a matter of drawing its characteristics in detail. It frequently happens, however, that no such valve can be found, in which case it is necessary to select one having the correct average values of the more important characteristics, such as plate current and mutual conductance. Suppose, for example, that the valve is very close to average with the exception in the screen current; in this case all the characteristics which do not involve the screen current may be drawn as usual, while the screen current readings will have to be multiplied by a factor to bring the screen current, under normal working conditions, to the same value as the average.

RADIOTRON 1.4 VOLT MINIATURE VALVES



1R5	1S5	1T4	3S4	3V4
Converter	Diode-pentode	R.F. Pentode	Power Pentode	Power Pentode



Radiotron type 807 is the most commonly used and the most generally flexible transmitting valve in use to-day. In transmitters it may be used as an oscillator, buffer, doubler, tripler, quadrupler, intermediate amplifier or power amplifier. In modulators it may be used as the driver and power amplifier.

Its use is not restricted to radio transmitters, being a most generally useful type. It is widely used in electronic voltage regulators and other miscellaneous laboratory, industrial and control applications. It is also used in audio amplifiers giving output powers from a few watts up to more than 100 watts. In radio receivers it may be used as a replacement for type 6L6 or anywhere where a high output is required.

Radiotron type 807 is at present available from stock in unlimited quantities.

Interchangeability List

Other Mfrs. Type	Equivalent Radiotron Type	Other Mfrs. Type	Equivalent Radiotron Type
GL-3C23	3C23	GL-415	5550
PJ-8	5556	GL-502-A	502-A
FG-17	5557	WL-579-B	579-B
FG-32	5558	KU-627	627
FG-57	5559	WL-629	629
FG-95	5560	WL-631	5559
FG-104	5561	WL-632-A	5560
FG-105	105	WL-651/656	5552
FG-172	172	WL-652/657	5551
CE-220	*8020	WL-653-B	5555
FG-235-A	5552	WL-655/658	5553
FG-238-B	5555	WL-672	672
FG-258-A	5553	KU-676	676
FG-259-B	5554	WL-677	677
FG-271	5551	WL-678	678
WT-272	5557	WL-679	5554
WE-287-A	*5557	WL-681/686	5550
CE-302 (722-A)	*3C23	WE-967	5557
CE-303	*3C23	GL-8020	8020
CE-305	*3C23	EL-C1A	*3C23
CE-306	*105	EL-C6C	* 105
CE-309	*627	EL-C6J	* 105
WE-393-A	*3C23		

* Similar but not directly interchangeable.

Recent R.C.A. Releases

Radiotron type 1B3-GT/8016 is a half-wave high-vacuum rectifier designed for use in high voltage, low current applications. An important feature is its low wattage filament (0.25 watt). It supersedes the earlier type 8016.

Radiotron type 3C23 is a negative-control triode type thyratron, the equivalent of the GE type GL-3C23. It is a three-electrode, gas and mercury-vapour filled thyratron intended for service in regulated-rectifier applications. It has a maximum peak inverse and forward anode voltage of 1,250 volts, a maximum peak anode current of 6A. above 25 c/s, and a maximum average anode current of 1.5A. up to 210 c/s. Filament voltage is 2.5 volts and filament current 7 amperes.

Radiotron type 4-125A/4D21 is a v-h-f- power tetrode with a maximum plate dissipation of 125 watts which can be operated with a maximum plate voltage of 3,000 volts at frequencies as high as 120 Mc/s. With reduced plate voltage, it can be operated at frequencies up to 250 Mc/s. Its typical power output is 375 watts with class C telegraphy and 300 watts with plate-modulated class C telephony, using the maximum plate voltage in each case. Forced-air cooling of the bulb and plate is required when this valve is operated near maximum ratings at frequencies above 30 Mc/s. In addition, circulation of air through the perforated shell of the 5-pin base is required to cool the stem. This type has a thoriated tungsten filament operating at 5 volts, 6.2 amperes.

Radiotron type 5FP4-A is a 5″ diameter kinescope (cathode-ray tube) intended primarily for use as an electrostatic view finder on television cameras It has a high efficiency white fluorescent screen on a particularly flat face, and utilizes magnetic focus and magnetic deflection to produce a $2\frac{7}{8}$″ by $3\frac{7}{8}$″ view of the television screen. A feature of the design is a limiting aperture at the end of the electron gun to produce a sharper, rounder spot at the screen, especally when the tube is operated on a high beam current. Other desirable features include high light output, large useful screen surface, short overall length and a sturdy structure.

Radiotron type 6AQ5 is a miniature beam power amplifier with a T$5\frac{1}{2}$ bulb and miniature button 7-pin base. It has a heater operating at 6.3 volts 0.45A., and is the miniature equivalent of type 6V6-GT, except that it is limited to maximum plate and screen voltages of 250 volts.

Radiotron type 6BF6 is a miniature duplex-diode triode with performance equivalent to type 6SR7. Since the triode unit has a medium mu, this may be used as a driver in transformer coupled output stages. The heater rating is 6.3 volts, 0.3A., and the bulb is T$5\frac{1}{2}$ with a 7-pin button stem.

Radiotron type 7CP4 is a 7″ diameter kinescope (cathode ray tube) having electrostatic focus and magnetic deflection. It has a white fluorescent screen of medium persistence and is fitted with an octal base.

Radiotron type 12SW7 is a metal duplex-diode triode with a 12.6 volt, 0.15A. heater, for use with a 12 cell accumulator. Although capable of being operated on plate voltages up to 250 volts, it may be used with a plate voltage of only 26.5 volts, and under these conditions it has an amplification factor of 17, plate resistance 15,500 ohms, and transconductance 1,100 micromhos, with a plate current of 1.1 mA. With maximum plate voltage, the plate resistance is 8,500 ohms, transconductance 1,900 micromhos, amplification factor 16, and plate current 9.5 mA. It is fitted with an octal base.

Radiotron type 12SX7-GT is a metal twin-triode with a 12.6 volt, 0.3A. heater, for use with a 12 volt accumulator. Although capable of being operated on plate voltages up to 300 volts, it may be used with a plate voltage of only 26.5 volts, and under these conditions it has an amplification factor of 21, plate resistance 11,500 ohms, transconductance 1,800 micromhos, and plate current 1.8 mA. With maximum plate voltage, the amplification factor is 20, plate resistance 7,700 ohms, transconductance 2,600 micromhos, and plate current 9 mA. It is fitted with an octal base.

Selected 12SN7GT

Radiotron type 12SY7 is a single-ended metal pentagrid converter with a 12.6 volt, 0.15A. heater, intended for use with a 12 cell accumulator. Although it is capable of being operated with plate voltages up to 300 volts, it may be used with a plate voltage of only 26.5 volts, and under these conditions has a conversion transconductance of 250 micromhos, with a cathode current of 2.25 mA. With 250 volts on the plate and 100 on the screen it has a conversion transconductance of 450 micromhos, and cathode current 12.5 mA., thus resembling the characteristics of type 12SA7. It is fitted with an octal base.

Selected

Radiotron type 26A6 is a miniature remote-cutoff pentode intended for use with a 12 cell accumulator with 26.5 volt, 0.07A. heater, $T5\frac{1}{2}$ bulb, and miniature button 7-pin base. Although it is capable of being operated on plate voltages up to 250 volts it may be used with a plate voltage of only 26.5 volts, and under these conditions has a transconductance of 2,000 micromhos with a plate current of 1.7 mA. With the full plate voltage and 100 volts on the screen, it has a transconductance of 4,000 micromhos with a plate current of 10.5 mA. and cutoff at a grid voltage of —25 volts.

Radiotron type 26C6 is a miniature duplex-diode triode for use with a 12 cell accumulator, having a 26.5 volt, 0.07A. heater, $T5\frac{1}{2}$ bulb, and miniature button 7-pin base. Although capable of being operated on plate voltages up to 250 volts, it may be used on a plate voltage of only 26.5 volts, and under these conditions has an amplification factor of 17, plate resistance 15,500 ohms, transconductance 1,100 micromhos, and plate current 1.1 mA. Under full voltage conditions the amplification factor is 16, plate resistance 8,500 ohms, transconductance 1,900 micromhos, and plate current 9.5 mA.

Radiotron type 26D6 is a miniature pentagrid converter for use with a 12 cell accumulator, having a 26.5 volt, 0.07A. heater, $T5\frac{1}{2}$ bulb, and a miniature button 7-pin base. Although capable of operating on plate voltages up to 300 volts it may be used with plate and screen voltages of only 26.5 volts, and under these conditions it has a conversion transconductance of 270 micromhos, with a cathode current of 2.15 mA. With 250 volts on the plate the conversion transconductance rises to 475 micromhos, and cathode current 11.3 mA. The oscillator transconductance at zero grid voltage is 4,500 micromhos, with 26.5 volts on plate and screen; 7,200 micromhos, with 100 volts.

Radiotron type 105 is a tetrode type thyratron, the equivalent of the GE type FG-105. It is intended for service in applications where a high impedance supply actuates the control grid and the actuating grid power is small. In continuous service, this type has a maximum peak inverse and forward anode voltage of 2,500 volts, a maximum peak anode current of 40A. above 25 c/s., a maximum average anode current of 6.4A., and the indirectly heated cathode has a 5 volt, 10A. heater. In intermittent service, this type can be operated as high as 10,000 volts (peak inverse and forward) to deliver a maximum peak anode current of 16A. above 25 c/s., with a maximum average anode current of 4 amperes.

Radiotron type 172 is a tetrode type thyratron, the equivalent of the GE type FG-172. It is intended for service in applications where a high-impedance supply actuates the control grid and the actuating grid power is small. In continuous service, it has a maximum peak inverse and forward anode voltage of 2,000 volts, a maximum peak anode current of 40A. above 25 c/s., a maximum average anode current of 6.4A., and the indirectly heated cathode has a 5 volt, 10A. heater.

Radiotron type 304TH is a medium-mu multi-unit triode having a plate dissipation of 300 watts which may be operated with maximum ratings up to 40 Mc/s. Under class C telegraphy conditions it gives a typical power output of 1,200 watts with the maximum voltage of 3,000 volts on the plate, while two valves used in a class B a-f amplifier are capable of giving a typical power output 1,400 watts with 3,000 volts on the plate. It has a filament which may be operated with two sections either in series or parallel, giving filament ratings of 10 volts, 12.5A., or 5 volts 25 amperes. The nominal overall length is $7\frac{7}{16}''$ and maximum radius is $1\frac{13}{16}''$. The amplification factor is 20.

Radiotron type 5553 is an ignitron, size D, the equivalent of the Westinghouse type WL-655/658 and the GE type FG-258-A. It is recommended for welder-control service, but may also be used for conversion in low-power circuits. It is capable of handling 2,400 KVA demand in welder-control service, and is equivalent to a 1,200A. contactor.

Radiotron type 502A is a negative-control tetrode type thyratron, the equivalent of the GE type GL-502-A, intended for use in compact regulated-rectifier equipment. It has a maximum peak inverse anode voltage of 1,300 volts, a maximum peak forward anode voltage of 650 volts, a maximum peak anode current of 500 mA., and a maximum average anode current of 100 mA. It has an indirectly heated cathode with a heater requiring 6.3 volts, and drawing 0.6 ampere.

Radiotron type 579-B is a half-way high-vacuum rectifier, the equivalent of the Westinghouse type WL-579-B. It is a filament type valve for use in applications involving high d.c. voltage and relatively low d.c. current. It has a maximum peak inverse voltage of 20,000 volts, a maximum peak plate current of 270 mA., and a maximum average plate current of 25 mA. Its thoriated-tungsten filament operates at 2.5 volts and 6 amperes.

Radiotron type 627 is a negative-control triode type thyratron, the equivalent of the Westinghouse type KU-627. It is a mercury-vapour tube used in regulated rectifier applications and having a maximum peak inverse anode voltage of 2,500 volts, a maximum peak forward anode voltage of 1,250 volts, a maximum peak anode current of 2.5A., a maximum average cathode current of 0.64A., and the filament operates at 2.5 volts and 6 amperes.

Radiotron type 629 is a thyratron triode of the negative control type, the equivalent of the Westinghouse type KU-629. It is gas-filled and intended for use as a sweep-circuit oscillator or as a grid-controlled rectifier. It has a maximum peak inverse and forward anode voltage of 350 volts, a maximum peak anode current of 200 mA., a maximum average anode current of 40 mA., and its indirectly heated cathode has a heater which operates at 2.5 volts and 2.6 amperes.

Radiotron type 672 is a thyratron of the negative-control tetrode type, the equivalent of the Westinghouse type WL-672. It is intended for use in controlling the firing of ignitrons. It has a maximum peak inverse and forward anode voltage of 1,500 volts, a maximum peak anode current of 30A., a maximum average anode current of 2.5A., and its indirectly heated cathode has a 5 volt, 6A. heater.

Radiotron type 676 is a thyratron of the negative-control triode type, the equivalent of the Westinghouse type KU-676, in tended for use in relay and grid-controlled rectifier applications. In continuous service, it has a maximum peak inverse and forward anode voltage of 2,500 volts, a maximum peak anode current of 40A., a maximum average anode current of 6.4 amperes. In welder-control service, the tube is rated at a maximum peak inverse and forward anode voltage of 750 volts, a maximum peak anode current of 77A., and a maximum average anode current of 2.5 amperes. It has an indirectly heated cathode with a heater requiring 5 volts and 10 amperes.

Radiotron type 677 is a thyratron of the negative-control triode type, the equivalent of the Westinghouse type WL-677, intended for grid-controlled rectifier applications involving high voltage. It has a maximum peak inverse and forward anode voltage of 10,000 volts, a maximum peak anode current of 15A., a maximum average anode current of 4A., and an indirectly heated cathode with a 5 volts, 10A. heater.

Radiotron type 678 is a negative-control triode type thyratron, the equivalent of the Westinghouse type WL-678. It is intended for use in relay and grid-conrolled rectifier applications. It is rated at a maximum peak inverse and forward anode current of 15,000 volts, a maximum peak anode current of 6A., a maximum average anode current of 1.6A., and its directly heated filament operates at 5 volts and 7.5 amperes.

Radiotron type 5550 is an ignitron, size A. It is a small compact steel-jacketed ignitron designed for welder-control service and capable of handling 300 KVA demand. It is equivalent to a 150A. magnetic contactor. When the demand does not exceed 105 KVA, air cooling is used, but above this value, water cooling is used. It is the equivalent of the Westinghouse type WL-681/686 and the GE type GL-415.

Radiotron type 5551 is an ignitron, size B, recommended for welder-control service, but also useful for conversion in low-power circuits and for intermittent rectifier service. It is capable of handling 600 KVA demand in welder-control service and is equivalent to a 300A. magnetic contactor. The equivalent of this type in the Westinghouse line is WL-652/657, and in the General Electric line FG-271.

Radiotron type 5552 is an ignitron, size C, the equivalent of the Westinghouse type WL-651/656 and the GE type FG-235-A. It is recommended for welder-control service, but is also useful for conversion in low-power circuits and for intermittent rectifier service. It is capable of handling 1,200 KVA demand in welder-control service and is equivalent to a 600A. magnetic contactor.

Radiotron type 5554 is an ignitron, the equivalent of the Westinghouse type WL-679 and the GE type FG-259-B. It is intended primarily for rectifier service in the 125, 250, 600, and 900 volt d.c. power fields, but it is also useful in 2,400 volt resistance-welder-control service and is capable of handling 1,200 KVA demand in this service. In rectfier service at frequencies from 25 to 60 c/s., this tube is rated for operation with a maximum peak inverse and forward anode voltage of 900 volts, a maximum peak anode current of 900A., and a maximum average anode current of 100 amperes. It may also be used under conditions calling for maximum peak inverse and forward anode voltage of 2,100 volts, a maximum peak anode current of 600A., and a maximum anode current of 75 amperes.

Radiotron type 5555 is an ignitron, the equivalent of the Westinghouse type WL-643-B and the GE type FG-238-B. It is intended primarily for rectifier service in the 125, 250, 600, and 900 volt d.c. power fields, but it is also suitable for 2,400 volt resistance-welder-control service and is capable of handling 2,400 KVA demand in this service. In rectifier service at frequencies from 25 to 60 c/s., it is rated for operation with a maximum peak inverse and forward anode voltage of 900 volts, a maximum peak anode current of 1,800A., and a maximum average anode current of 200 amperes. It is also rated at a maximum peak inverse and forward anode voltage of 2,100 volts, maximum peak anode current of 1,200A., and a maximum average anode current of 150 amperes.

Radiotron type 5556 is an amplifier triode, the equivalent of the GE type PJ-8. It is a filament type of triode with a mu of 8.5 for use in amplifier, oscillator, and control service. In class A service it has a maximum plate voltage of 350 volts, and a maximum plate dissipation of 7.5 watts. Its power output is 0.6 watt. In class C telegraph service, it has a maximum plate dissipation of 10 watts, and a power output of 6 watts at frequencies up to 6 Mc/s. The filament operates at 4.5 volts and 1.1 amperes.

Radiotron type 5557 is a thyratron of the negative-control triode type, the equivalent of the GE type FG-17. It is for use in applications where a change in negative grid voltage actuates the tube. It has a maximum peak inverse anode voltage of 5,000 volts, a maximum peak forward anode voltage of 2,500 volts, a maximum peak anode current of 2A. above 25 c/s., and a maximum average anode current of 0.5 ampere. The 2.5 volt filament draws 5 amperes.

Radiotron type 5558 is a half-wave mercury-vapour rectifier, the equivalent of the GE type FG-32. It has a maximum peak inverse anode voltage rating of 1,000 volts, a maximum peak anode current of 15A. at 25 c/s. and above, a maximum average anode current of 2.5A., while the indirectly heated cathode has a heater operating at 5 volts and 4.5 amperes.

Radiotron type 5559 is a thyratron of the negative-control triode type, the equivalent of the Westinghouse type WL-631 and the GE type FG-57. It is intended for use in applications where a change in negative grid voltage actuates the tube. It is rated at a maximum peak inverse and forward anode voltage of 1,000 volts, a maximum peak anode current of 15A. above 25 c/s., a maximum average anode current of 2.5A., and the indirectly heated cathode has a heater which operates at 5 volts and 4.5 amperes.

Radiotron type 5560 is a thyratron of the tetrode type, the equivalent of the Westinghouse type WL-632-A and the GE type FG-95. It is useful in applications where a high-impedance supply actuates the control grid and the actuating power is small. It has a maximum peak inverse and forward anode voltage of 1,000 volts, a maximum peak anode current of 15A. above 25 c/s., a maximum average anode current of 2.5A., and the indirectly heated cathode has a heater operating at 5 volts and 4.5A.

Radiotron type 5561 is a half-wave mercury-vapour rectifier, the equivalent of the GE type FG-104. For continuous service, it is rated at a maximum peak inverse anode voltage of 3,000 volts, a maximum peak anode current of 40A. at 25 c/s. and above, and a maximum average anode current of 6.4 amperes. For welder-control-service, the rating is a maximum peak inverse anode voltage of 10,000 volts, a maximum peak anode current of 16A. at 25 c/s. and above, a maximum average anode current of 4A., and the indirectly heated cathode has a heater operating at 5 volts and 10 amperes.

Radiotron type 8020 is a half-wave high-vacuum rectifier, the equivalent of the GE type GL-8020. It is intended for use in applications requiring high d.c. voltage and relatively low d.c. current. It has a maximum peak inverse plate voltage of 40,000 volts, a maximum peak plate current of 750 mA., and maximum average plate current of 100 mA. Its thoriated tungsten filament operates at 5 volts and 6 amperes.