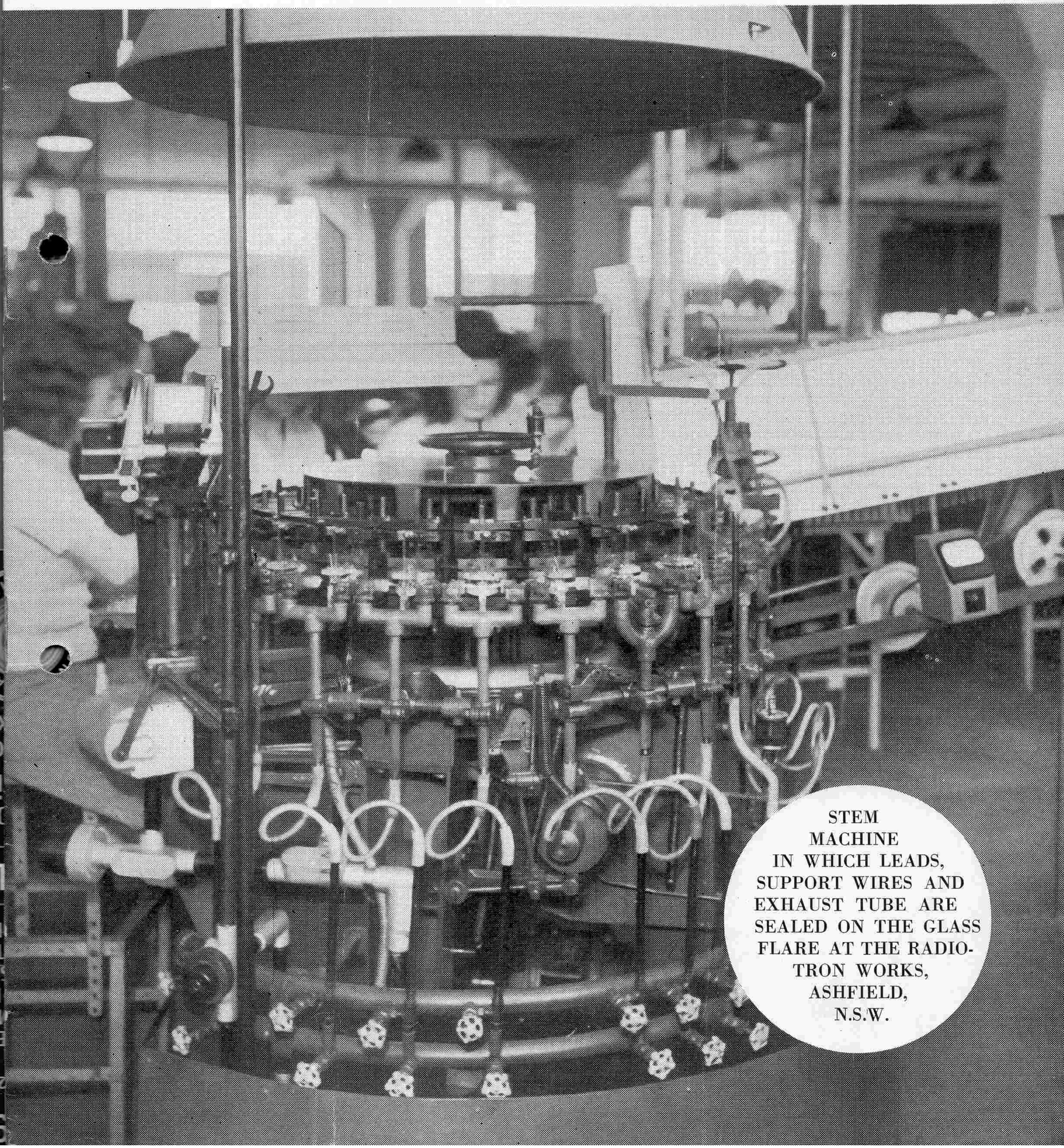
Radiotronics

Number 125 MAY — JUNE 1947



STEM MACHINE IN WHICH LEADS, SUPPORT WIRES AND EXHAUST TUBE ARE SEALED ON THE GLASS FLARE AT THE RADIOTRON WORKS, ASHFIELD, N.S.W.

RADIOTRONICS

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VALVE DATA SECTION:

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

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OUR COVER

Modern automatic stem making machine at the Radiotron Valve Works, which turns out complete annealed stems at the rate of 500 per hour.

Specially shaped gas fires quickly heat the glass parts to red heat, at which stage they are plastic enough to be moulded together. At the same time these fires form vacuum tight seals between leads and glass.

Stems are inloaded and passed through the tunnel annealer (at right), the four feet journey taking about 10 minutes, removing strains in the glass so that cracks will not occur on subsequent use of the stem in a valve.

An F-M Receiver for the 88-108 Mc/s Band

By B. SANDEL A.S.T.C.

This is the first of two articles dealing with some of the more important points in the design and construction of a complete receiver for the reception of frequency modulated carrier waves. The frequency range has been taken as 88 to 108 Mc/s, in conformity with the standard range used in the U.S.A., and it is assumed that the local transmissions will be confined to this same frequency band.

The receiver to be described is a compromise between cost and quality, and for this reason the audio system is a comparatively simple one. It does not seem logical to design an audio amplifier having practically flat response from 50 c/s to 15,000 c/s if a comparatively low priced speaker and output transformer are to be used. For this reason the audio frequency response has been limited to about 10,000 c/s as the upper limit. This allows standard components and speakers to be used, thereby helping to keep cost to a reasonable level. The high fidelity enthusiast can easily substitute the best amplifier and speaker available to obtain superior results, but for a commercial product this type of audio system is rather impractical.

This first article will give the methods used for the preliminary design of the receiver, together with some of the points to be observed when design details are not fully described, such as where the techniques are so well known as not to warrant further description.

General

Fig. 1 shows a block schematic diagram of the receiver. The valve types shown have been previously discussed in Radiotronics. The tuning range is 88-108 Mc/s and the i-f is 10.7 Mc/s; the oscillator frequency is higher than the signal frequency by the i-f. An advantage of this arrangement is immediately apparent as the image frequency (signal frequency + 2× intermediate frequency) falls outside the receiver tuning range, thereby materially assisting in the reduction of spurious responses and whistle interference. These remarks would also apply to the case where the oscillator frequency is lower than the signal frequency, but the oscillator frequency higher than the signal frequency has been adopted fairly generally in the U.S.A, because of the possibility of reduced inter-

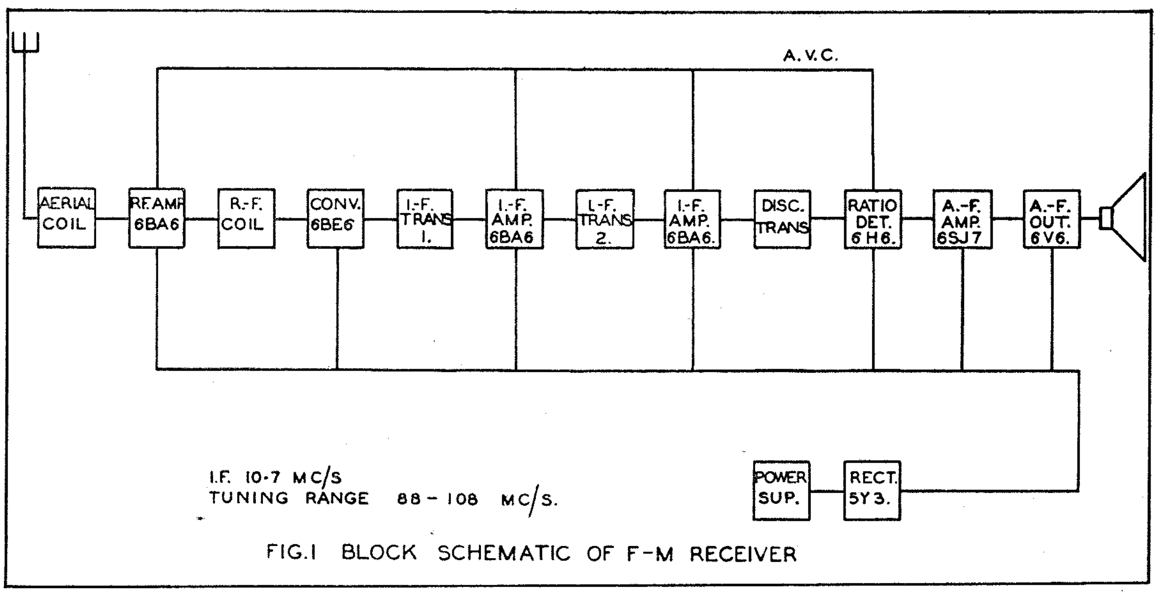

FIG. I BLOCK SCHEMATIC OF F-M RECEIVER

ference with existing equipment, such as mobile f-m transmissions on the 72-76 Mc/s band. An r-f stage has been included as this greatly improves image rejection. Further, the increase in signal-to-noise ratio, a particularly important factor at very high frequencies, and the additional gain, should be helpful in areas where reception is poor. Automatic volume control has been included as this is advantageous when using a ratio detector because the signal output depends on the strength of the received carrier.

As the i-f amplifier is of major importance in determining selectivity and gain it will be as well to consider first the design of the i-f transformers and the i-f amplifier.

I-F AMPLIFIER

(a) Design of I-F Transformers.

The i-f amplifier will largely determine the performance of the receiver and so considerable care is called for in its design. It is necessary to have as high a gain per stage as possible, consistent with stability, but at the same time a wide band of frequencies must be passed as otherwise severe distortion will be introduced. It is to effect this compromise that valve types such as the 6BA6 are used since they have a high g_m and provide fairly low damping on the tuned circuits thereby assisting in achieving good stage gains. That this is true can be seen from the expression for the gain of a stage using a critically coupled i-f transformer

$$\text{Gain} = \frac{g_m Q \omega L}{2}$$

where g_m = mutual conductance of the i-f amplifier valve

Q = the magnification factor of the primary and secondary windings (each in the absence of the other) of the i-f transformer (Pri. & sec. assumed identical).

ω = $2\pi \times$ intermediate frequency.

L = inductance of pri. & sec. windings. (Pri. & sec. assumed identical.)

Because of alignment difficulties, both from a manufacturing and servicing point of view, critically coupled transformers have been chosen as preferable to over-coupled transformers, or a combination of a critically coupled and an over-coupled stage. The latter alternative would offer some advantages in a very high quality receiver as regards band width and "skirt" selectivity.

The pass band required will determine the Q and k (coefficient of coupling) of the transformer; and the minimum capacitance permissible to give stable operation will determine the value of L. Considering the pass band requirements, these will be determined by the number of significant sidebands involved in the frequency modulated carrier wave. It is usual to consider only those sidebands whose value is greater than 1% of the magnitude of the unmodulated carrier as being significant. The maximum frequency deviation of the carrier is ±75Kc/s about the central carrier frequency and the highest audio frequency transmitted is 15Kc/s. It should be noted that the distribution of carrier and sidebands theoretically covers an infinite range of frequencies and not just ±75Kc/s about the carrier frequency. It is for this reason that some practical limit must be set and because of this the value of 1% is chosen, as above, for the smallest sideband amplitude, since over 99% of the total power contained in the carrier and all the sidebands (an infinite number) is then available. (Refs. 2 & 14.) It should also be noted that the carrier amplitude is not constant and can become zero under certain conditions of modulation.

It can easily be demonstrated that the required bandwidth, for sidebands greater than 1% of the unmodulated carrier amplitude, is greatest at the highest modulating frequency when a fixed deviation frequency is used. Since the spacing between the sidebands is equal to the audio frequency and the number of significant sidebands in this case is 16, the total bandwidth required is $15{,}000 \times 16 = 240$ Kc/s. This is the most severe condition normally encountered.

The next step to consider is that as the critically coupled transformers obviously haven't a flat top of 240 Kc/s width, how much attenuation can be allowed for this band of frequencies without introducing severe distortion, but at the same time giving a reasonable degree of selectivity? A practical answer to this question appears to correspond to an introduced value of amplitude modulation of about 50% for 240Kc/s bandwidth (or 30% a-m for the more commonly used value of 200Kc/s) due to the selectivity of the i-f transformers. This amount of a-m is fairly readily removed by amplitude limiters, and further all significant sidebands operate on a substantially linear portion of the phase angle-frequency characteristic (see Ref. 4). This figure approximates very closely to an attenuation due to the i-f transformers, of 3 times (9.54db) at a bandwidth of 240Kc/s, and as this figure is convenient for calculation it will serve as a practical design basis. It should be noted here that the frequency deviation of ±75Kc/s only enters into the calculations for the required bandwidth as one of several factors in determining the value of 240Kc/s.

So far then it has been found that we require a bandwidth of 240Kc/s for 9.54db attenuation. Also two i-f stages are necessary to obtain a reasonable amount of gain, as it will be seen later that the stage gains are considerably lower than those which can be obtained with the usual 455Kc/s i-f amplifier. This then means that we have available two i-f transformers and a discriminator transformer which will all affect the bandwidth. As will be seen later, the effect of the discriminator transformer on bandwidth is fairly small because of its low Q value, and it will be sufficient to design two i-f transformers which together will give the required bandwidth and attenuation.

The formulae required are

$$X = \sqrt{2}\left(K^{\frac{4}{n}} - 1\right)^{\frac{1}{4}}$$

and $$X = \frac{2\Delta f}{f} Q$$

where K = gain reduction factor
= 3 (corresponding to 9.54db.)
n = number of similar circuits coupled together in pairs.
= 4 for two i-f transformers.
f = i-f centre frequency
= 10.7 Mc/s
$2\Delta f$ = bandwidth
= 240Kc/s.

Then

$$X = \sqrt{2}\left(3^{\frac{4}{4}} - 1\right)^{\frac{1}{4}}$$
$$= 1.682$$

$$Q = \frac{fX}{2\Delta f}$$

$$= \frac{10.7 \times 10^6 \times 1.682}{240 \times 10^3}$$

$$= 75.2$$

For a critically coupled transformer

$$k_{CRIT.} = \frac{1}{Q}$$

$$= \frac{1}{75.2}$$

$$= 0.0133$$

Next, to determine the inductance of the transformer windings, it is necessary to determine the tuning capacitance (C) across each winding. If it is assumed that the total capacitance across each circuit should not be less than about $60\mu\mu F$ for stable operation, and for reducing effects such as detuning due to a.v c. etc., (the $60\mu\mu F$ being made up from say $10\mu\mu F$ strays and $50\mu\mu F$ fixed capacitance) the value of the inductance required is

$$L = \frac{25330}{10.7^2 \times 60}$$

$$= 3.69\mu H.$$

The transformer details are then

$$Q = Q_1 = Q_2 = 75.2$$
$$k_{CRIT.} = 0.0133$$
$$L = L_p = L_s = 3.69\mu H$$
and $C = C_p = C_s = 60\mu\mu F$
($50\mu\mu F. + 10\mu\mu F$ strays).

The stage gain using a type 6BA6 is

$$\text{Gain} = \frac{g_m Q \omega L}{2}$$

$$= \frac{4{,}400 \times 10^{-6} \times 75.2 \times 2\pi \times 10.7 \times 10^6 \times 3.69 \times 10^{-6}}{2}$$

$$= 41.1 \text{ times.}$$

This is a reasonable compromise as stage gains higher than about 50 often present difficulties due to regeneration at these frequencies.

To determine whether the transformers are completely satisfactory the overall response should be calculated over a range of at least $\pm$ 400Kc/s from the central intermediate frequency to ensure that if

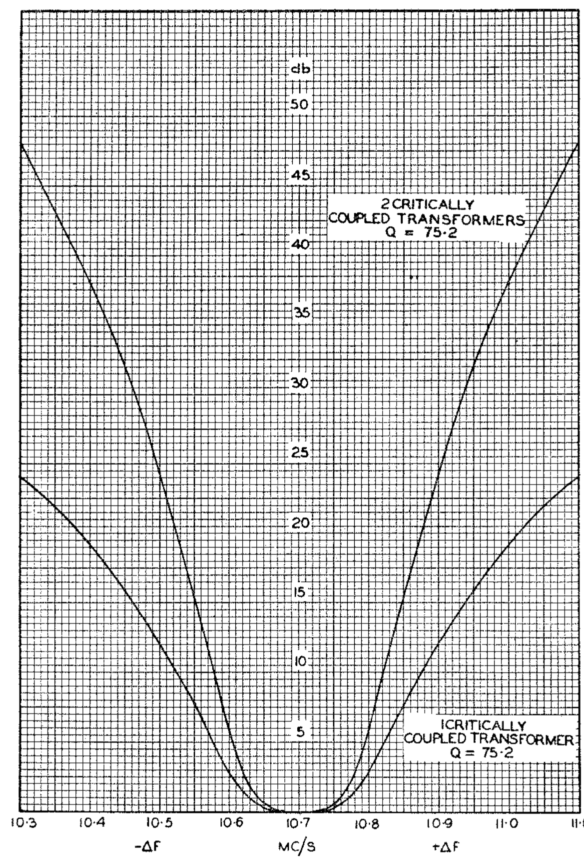

FIG.2

two stations are separated by, say, 400Kc/s severe interference will not result. The calculated curve is shown in Fig. 2. Also a response curve for a single transformer should be drawn as this assists in finding the actual spacing required between the primary and secondary windings since the coupling can be altered until approximately the required response curve is obtained.

From the F.C.C. recommendations adjacent channel interference is not considered to exist when the ratio of desired to undesired signal is at least 2:1, and for signals on the same channel the ratio is 10:1. This condition appears at first sight to be fulfilled quite easily, if it is taken that stations in the same city have their central carrier frequencies separated by 400Kc/s, but it is necessary to consider that a weak station may be tuned while the interfering station is an extremely strong one. If the maximum field intensity from the desired station is taken as 50μV/metre and that from the undesired station is say 1,000μV/metre it can be readily seen that the conditions are not so easily fulfilled. If the station central frequencies are separated by 400Kc/s, the undesired signal response at the central frequency of the desired station is (from the i-f selectivity curve) 48db. below 1,000μV/metre. So the ratio of the desired to the undesired signal is

$$\frac{50}{3.99} = 12.5$$

which fulfils the required condition; also some improvement on this figure will be obtained due to the other tuned circuits in the receiver. Of course, if the interfering signal has a greater intensity, interference could easily become a serious problem. This indicates that the sharpest selectivity curve possible, consistent with bandwidth requirements is desirable. This also indicates that as the bandwidth requirements determine the shape of the selectivity curve, increased "skirt" selectivity can only be obtained with extra i-f stages or combinations of transformers having various types of coupling.

Interference can be largely minimized by suitably locating transmitters, and doubtless these effects will be considered when sites are being chosen for new F-M broadcasting stations, as is done in U.S.A. where it is necessary to submit the predicted coverage of the transmissions, the areas of interference, and the basic data employed in computing the interference.

The details of the i-f transformers have now been determined. Winding details will be given when the actual receiver is described. The windings are solenoids and can be readily calculated from the formula due to Hayman quoted on page 145 of the Radiotron Designer's Handbook. Allowance must, however, be made for the presence of the iron "slugs" (if these are used) in the value of inductance used in the calculations; about 10% allowance is usual depending on the "slug" size. Our transformers are wound with 22 B&S enamelled wire on ¾" former, and use ⅜" dia. ½" long "slugs" for setting the inductance values. The powdered iron "slugs" must be of high quality, otherwise a serious change in Q is experienced as they are moved through the windings.

The Q of the transformer windings will normally be somewhat in excess of the value required and the correct value is obtained by using parallel damping resistors, after allowing for the effects of valve and circuit damping. The total parallel damping resistance required can be found from

$$R = \frac{Q_A Q_B \omega L}{Q_A - Q_B}$$

where Q_A is the undamped Q, Q_B is the required Q,
$\omega = 2\pi \times$ i-f,
$L = L_p = L_s$ (as previously determined).

The actual resistance values used in parallel with the windings are then found after allowing for the plate resistance and input resistance of the valves in the i-f amplifier. With the circuit used this effect is small and it is close enough to use the calculated values for the damping resistors.

(b) Design Of Discriminator Transformer.

The method used for the design of the transformer for the phase discriminator is quite general and applies directly to any practical case whether the receiver uses limiters, a ratio detector, or some other similar form of amplitude limiting. The actual transformer used with a ratio detector requires some slight modifications and these will be discussed as the design proceeds. A typical arrangement for a phase discriminator transformer is shown in Fig. 3.

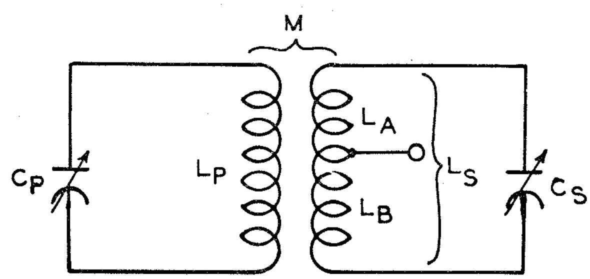

FIG. 3 PHASE DISCRIMINATOR.

It is possible to show (see Ref. 1, Part 1) that the criterion for a linear discriminator characteristic, i.e., the relation between output voltage versus off-tune frequency, is that the ratio of secondary to primary voltage (E_s/E_p) must be greater than 2. As the sensitivity of the discriminator (the frequency to amplitude conversion efficiency) is important, it is necessary to arrive at a compromise between linearity and sensitivity. As a suitable design basis the following values most nearly fulfil the two requirements,

$$\frac{E_s}{E_p} = 2$$

and $Qk = 1.5$

where

k is the coefficient of coupling.
Q is the magnification factor of the primary and secondary ($Q = Q_s = Q_p$).
(Unequal pri. & sec. Q's may offer some slight advantages, but normally this is small.)

From these constants, and since the intermediate frequency (f) is 10.7Mc/s, and the frequency deviation (Δf) is $\pm$75Kc/s, we have

$$\frac{L_s}{L_p} = \frac{\left(\frac{E_s}{E_p}\right)^2}{(Q_s k)^2}$$

(this is derived in Ref. 1, Part 2.)

$$= \frac{2^2}{1.5^2}$$

$$= 1.77$$

where

L_s and L_p are the secondary and primary inductances respectively.

also

$$Q = \frac{f}{2\Delta f} = \frac{10.7}{0.3}$$

$$= 35.7$$

where we have extended the range of linear response to a total of 300Kc/s, to fulfil the bandwidth requirements of 240Kc/s and to make allowance for possible slight detuning of the receiver due to oscillator frequency drift and mechanical errors in tuning the required frequency. In many practical cases this response is limited to 200Kc/s and the Q then becomes 53.5, but these figures do not seem particularly satisfactory in view of the previous discussion on bandwidth requirements.

As $Qk = 1.5$

$$\therefore\ k = \frac{1.5}{35.7}$$

$$= 0.042$$

Maximum sensitivity is obtained when the dynamic resistance of the transformer primary circuit has the highest possible value. This means that, since Q is fixed, L_p should be as large as possible. The value of L_p is limited, however, by the maximum value of L_s which in turn depends on the secondary tuning capacitance. If this value of secondary capacitance is taken as 60$\mu\mu$F, including strays, we have

$$L_s = \frac{25330}{10.7^2 \times 60}$$

$$= 3.68\mu H.$$

$$L_p = \frac{3.68}{1.77}$$

$$= 2.08\mu H.$$

$$C_p = \frac{25330}{10.7^2 \times 2.08}$$

$$= 106\mu\mu F.$$

where C_p is the primary circuit tuning capacitance.

Summarizing for the discriminator transformer,

$L_p = 2.08\mu H.$
$C_p = 106\mu\mu F.$
$L_s = 3.68\mu H.$
$C_s = 60\mu\mu F.$
$Q_p = Q_s = 35.7.$
$k = 0.042.$

A calculated response curve is again helpful in the actual design of the transformer, just as with the design of the i-f transformers, and this is very simply carried out using the curves given in Ref. 1, Part 1.

Because it is desirable to obtain balance between the two halves of the centre tapped discriminator transformer secondary, "slug" tuning of the windings is undesirable and capacitance tuning is preferable. In the practical arrangement it was found desirable to wind one half of the transformer secondary with less turns than the other half and use an iron "slug" to set the inductance of the two halves to the same values. The complete secondary is then tuned with a capacitance trimmer as is also the primary. Also the secondary winding was split so that each section was arranged symmetrically with respect to the primary. This minimizes lack of symmetry due to stray capacitance coupling.

(c) Modifications Required To Transformer When Using Ratio Detector.

A ratio detector has been chosen for several reasons. It eliminates the necessity for a limiter stage and hence an additional valve, i-f transformer, and other components. Further, very high overall amplification is not as necessary as when limiters are used, as the ratio detector will respond satisfactorily to inputs of the order of 10 to 20 millivolts, at the grid of the discriminator driver valve, whereas several volts are required to saturate a limiter stage. In this regard the R.C.A. report states that for 100 millivolts input to the 6BA6 driver stage the peak to peak audio output is 3.2 volts for $\pm$ 75Kc/s deviation. So that the discriminator driver stage has more gain than conventional limiters and if good audio gain is available the receiver can be made comparable in sensitivity with a receiver using limiter stages. For the audio amplifier an input of 0.142 volts gives 50mW output, or an overall gain of 110 times. Higher audio gain may be desirable under some circumstances.

In addition, the possibility of tuning errors is reduced, as discussed in the R.C.A. report quoted below, an important factor with an f-m receiver. In this regard a tuning indicator of a suitable type offers obvious advantages.

Because of these considerations it was decided to utilize the arrangement described in the R.C.A. Laboratories Reports LB-645 and LB-666 reprinted in Radiotronics 120. The actual circuit used was that of Fig. 4, and the reduced primary voltage obtained by tightly coupling a small tertiary winding to the primary and connecting it as shown. This then allows the primary inductance, and consequently the primary impedance, to be increased; so that although our secondary inductance value of L_s remains unchanged as does Q and k, the value of L_p can be increased and C_p decreased. The actual increase in L_p depends on the tertiary winding.

In our case the tertiary winding was arranged so as to give the secondary resonant voltage twice the value of the primary resonant voltage. This means that the primary inductance can be proportionately increased.

As a complete analysis of the ratio detector has not been made up to the present time, experimental methods must be resorted to. In order to keep the experimental work to a minimum the component values selected were based on those quoted in the two previously mentioned R.C.A. reports. Work carried out, however, indicates that these values give a sufficiently good compromise for an initial design. Final values will be quoted in the write-up on the receiver, but initial experimental work can be satisfactorily carried out with the values given, i.e. R = 30,000 ohms, C = 8 μF.

Experimental work has also been carried out and a method evolved for measuring the limiting action, or more correctly, perhaps, the a-m rejection capabilities, of the ratio detector. It is hoped to publish this information at some later date, together with a direct comparison between the limiting action of the ratio detector and single and double limiter stages.

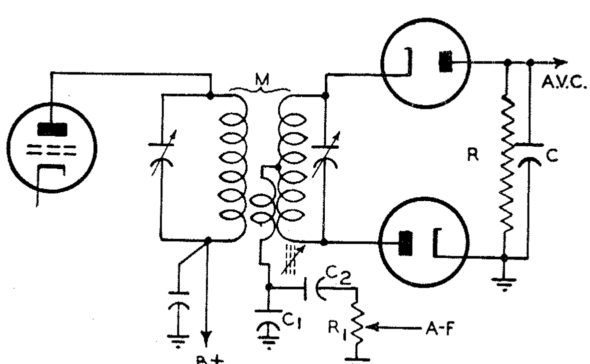

FIG. 4 RATIO DETECTOR.

Finally for the ratio detector circuit suitable values must be chosen for C_1, C_2 and R_1. C_1 connects the tertiary transformer winding to ground at i-f but its value must be such that it does not attenuate audio frequencies any more than that required to achieve normal de-emphasis of the audio voltages applied to the audio voltage amplifier. The value of C_1 is best selected experimentally by applying to the i-f amplifier a signal which is frequency modulated by a variable frequency audio oscillator, the modulation following the normal pre-emphasis curve given by a circuit having a time constant of 75 micro-seconds. The correct value of C_1 will of course depend on the value of R_1 and the other components in the circuit but it is chosen to give a substantially flat response to 15,000c/s. Whether this response is required will depend on the audio system, and it is worth noting that some compensation could probably be allowed to assist the higher audio frequency response if so desired. C_2 is merely a blocking condenser and its value is determined by its reactance, which must be small at 50 c/s (the lowest desired audio frequency), compared with the resistance of R_1. A suitable value is 0.1μF if R_1 is made 0.5MΩ. The value of 0.5MΩ is chosen for R_1 as its value should be as high as possible, but not so high that if a value of say 100$\mu\mu$F is chosen for C_1 the higher audio frequencies are affected by the comparatively low reactance of C_1; also C_1 cannot be made too small as otherwise the reactance between the transformer winding and ground at i-f would be appreciable.

AUDIO AMPLIFIER AND POWER SUPPLY

The audio amplifier used in this receiver is almost identical with that described in Radiotronics 117, and used with the RC52 receiver. The amplifier has an audio response which does not vary by more than about $\pm$1db from 50 to 10,000c/s, and the total harmonic distortion is less than 1% for 1 watt output at 400c/s. This allows a reasonable reserve of power on output peaks as the amplifier is capable of up to 4½ watts output, although the harmonic distortion will increase with larger outputs. The power reserve is essential, if reasonably good results are required, because of the greater range of dynamic sound intensity possible with F-M.

The quality of the output transformers and speakers available appear to offer the most severe barrier to really high quality reproduction, and it is to be hoped that improvements will be made in this direction particularly with standard types of speakers and transformers which will undoubtedly be used in medium priced receivers. For more expensive receivers good audio systems can be attached to the basic receiver to give improved results.

It does not seem worthwhile to go through the design of the audio amplifier in detail as this procedure is so well known. Further, a good deal of general description has already been devoted to the particular circuit used.

The inclusion of a bass boost circuit has been considered desirable as, even with normal receivers, in which the high frequency response is severely attenuated, bass boost is desirable at low volume levels. It appears that with the improved high frequency response, that bass boost will be even more desirable than previously; although normal f-m transmitters have better low frequency response than is usual with a-m transmitters because of the modulation methods employed. A control for treble attenuation may also be desirable from the user's point of view. A perfectly flat a-f response is then available if required, and the user may adjust the bass and treble response to suit his individual requirements. Further practical experience will be helpful, however, in determining the types of controls to be fitted to commercial receivers.

The power supply will be of the conventional type except that some additional care is necessary to keep the hum to extremely low levels, and the regulation of the H.T. voltages must be good. These features arise particularly because any rapid variations of the oscillator frequency due to hum or poor power supply regulation will be changed to amplitude variations by the discriminator circuits converting frequency changes to amplitude changes, and undesired hum or noise will be heard in the receiver output.

THE SIGNAL AND OSCILLATOR CIRCUITS

The procedure for the design of the signal and oscillator circuits is similar to that previously described in Radiotronics 101 under the heading of Radio Receiver Design (Part 3)—Tracking. Some details differ, however, so the complete design, and calculation of the receiver performance, will be set out. It will be seen that in general, the design does not differ very appreciably from that for an a-m receiver operating over the same frequency range.

Considerable care is necessary with the oscillator circuit as regards long and short period stability. As mentioned previously good power supply regulation and low hum level are necessary, as the discriminator responds to the effects on the oscillator frequency. This problem does not arise with a-m receivers as rapid variations of the oscillator frequency do not affect the detector which responds only to amplitude changes. If the oscillator frequency drifts by an appreciable amount over, say, a fairly long period in an a-m receiver, the result is similar to detuning and gives rise to an accentuation of the high frequency audio frequency components, and "screechy" reproduction. With frequency modulation the effect is to limit the maximum permissible frequency deviation of the i-f carrier, as the centre frequency is not set to the same frequency as that of the incoming signal. The result is to limit the output and to cause distortion of the reproduced audio frequencies.

(a) Signal Circuits.

The capacitance range of the condenser available is nominally 5 to 25$\mu\mu$F. The required tuning range is 88-108Mc/s and as some overlap is desirable at the ends of the band we will take the complete range as 87.5 to 108.5Mc/s.

Then using the same notation as in the article mentioned above

$$T = \frac{G_{max.}}{\alpha^2-1}$$

$$= \frac{20}{1.54-1}$$

$$= 37\mu\mu F.$$

$$L = \frac{1}{T\omega_2{}^2}$$

$$= \frac{25330}{37 \times 108.5^2}$$

$$= 0.058\mu H.$$

For the **AERIAL COIL** we will assume that the aerial is connected to the receiver through an 80 ohm transmission line, as cable having this characteristic impedance, or a near value, is fairly readily available.

It should be noted that the R.M.A. (U.S.A.) have standardized on 300 ohms, and using the same methods as those which follow, the aerial coil constants in this latter case would be $L_p = 0.49\mu H$, $L_s = 0.058\mu H$ and $k = 0.24$. A sketch is shown in Fig. 8 for a folded dipole aerial using the 300 ohm. cable available in U.S.A. The simplicity of the arrangement is apparent.

Following the method due to Rudd (1944) we have Geometric mean angular frequency.

$$= \omega_o = \sqrt{\omega_1 \ \omega_2}$$

$$= 2\pi \times 10^6 \sqrt{87.5 \times 108.5}$$

$$= 612 \times 10^6$$

$$Z_o = \omega_o L_p$$

$$\text{and } L_p = \frac{80 \times 10^6}{612 \times 10^6}$$

$$= 0.13\mu H.$$

where L_p is the primary inductance of the aerial coil.

To determine the required coefficient of coupling it is necessary to assume a value for Q_s, the secondary circuit Q in the absence of the primary. As the input resistance of the type 6BA6 at 100Mc/s is approximately 1600 ohms, this will seriously affect the value chosen. An undamped Q of 150 should be readily obtainable, and it is necessary to determine the value of Q_s when an equivalent resistance of 1600 ohms is added in parallel. A simple calculation shows this to be

$$Q_s = \frac{QR}{Q\omega L+R}$$

$$= \frac{150 \times 1600}{5480+1600}$$

$$= 33.9$$

The coefficient of coupling required is

$$k = \sqrt{\frac{1}{Q_s}\left[\frac{1+\alpha}{\alpha^{\frac{1}{2}}}\right]}$$

$$\text{where } \alpha = \frac{\omega_2}{\omega_1}$$

$$= \frac{108.5}{87.5}$$

$$= 1.24$$

$$\therefore k = \sqrt{\frac{1}{33.9}\left[\frac{1+1.24}{1.115}\right]}$$

$$= 0.24$$

The aerial coil details are then

L_p = 0.13μH (centre tapped)

L_s = 0.058μH (If allowance is made for the presence of the primary winding, this value is modified by approx. 0.35%.)

$k = 0.24$

$$M = k\sqrt{L_pL_s}$$
$$= 0.0212\mu\text{H}$$

Total trimming capacitance including strays and gang minimum capacitance

$= 37\mu\mu$F.

The circuit arrangement is shown in Fig. 5.

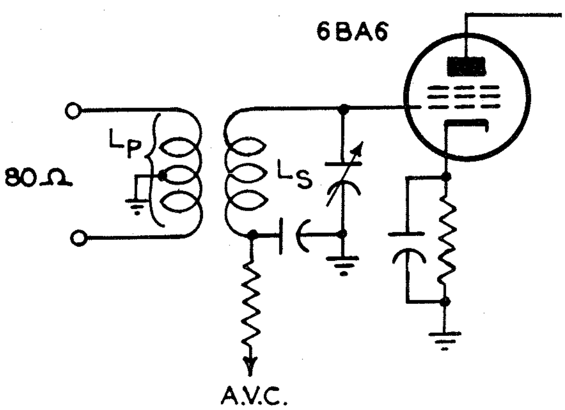

FIG. 5 AERIAL COIL ARRANGEMENT.

For the R-F COIL, as the high tension for the plate of the 6BA6 is supplied through a choke which is not coupled by mutual inductance to tuned circuit, the value for L and T found above apply directly. So that the inductance of the r-f coil is 0.058μH and the total trimming capacitance including gang minimum capacitance and strays is $37\mu\mu$F. The circuit arrangement is shown in Fig. 6.

The design of the r-f choke must now be considered as it will have an important effect on the tuned circuit, and it will be shown that the effect will be equivalent to a capacitance in series with the coupling capacitance C, the resultant capacitance being shunted across the tuned circuit. This resultant capacitance is not constant, but alters with frequency and so has an effect on the receiver tracking, as it

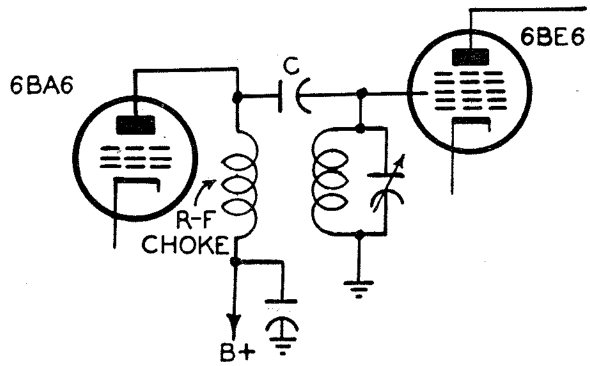

FIG. 6 R-F COIL ARRANGEMENT.

will alter the setting of the tuning condenser required for resonance across the r-f coil when receiving some particular signal frequency.

If we follow the procedure given in the previous tracking article the r-f choke will be resonated below the lowest signal frequency. As we require maximum stage gain and our stray capacitances across the tuning condenser must not exceed the total value of $37\mu\mu$F, for the given frequency coverage, some additional care is required. If reasonable precautions are taken it should be possible to keep the stray capacitance, due to 6BA6 output capacitance + coil and wiring capacitance, across the choke down to say $10\mu\mu$F. This is in the absence of the tuned circuit, etc.

Taking our previous method, the lowest signal frequency is 87.5Mc/s, then the resonant frequency for the choke is taken as $\frac{87.5}{\sqrt{2}}$ or 62Mc/s. The inductance required is then

$$L = \frac{25330}{62^2 \times 10}$$

$$= 0.66\mu\text{H}$$

To see what the effect will be at 88 and 108 Mc/s let us proceed to calculate the effective capacitance, and then choose a value for C to determine whether the circuit is practicable. From the circuit made up from the capacitance in parallel with the choke, the effective reactance is

$$X = \frac{j\omega L\left[\dfrac{-j}{\omega C}\right]}{j\omega L - \dfrac{j}{\omega C}}$$

$$= \frac{-j\dfrac{L}{C}}{\omega L - \dfrac{1}{\omega C}}$$

At 88Mc/s,

$$\omega L = 2\pi \times 88 \times 10^6 \times .66 \times 10^{-6}$$
$$= 365\text{ohms.}$$
$$\frac{1}{\omega C} = \frac{10^{12}}{2\pi \times 88 \times 10^6 \times 10}$$
$$= 181\text{ohms.}$$
$$\frac{L}{C} = \frac{0.66 \times 10^{-6}}{10 \times 10^{-12}}$$
$$= 0.66 \times 10^5$$
$$\therefore X = -j\frac{0.66 \times 10^5}{184}$$
$$= -j359\text{ohms.}$$

Hence the equivalent capacitance is 5.05$\mu\mu$F.

At 108Mc/s,

$$\omega L = \frac{365 \times 108}{88}$$
$$= 448\text{ohms.}$$
$$\frac{1}{\omega C} = \frac{181 \times 88}{108}$$
$$= 147\text{ohms.}$$
$$\therefore X = -j\frac{0.66 \times 10^5}{301}$$
$$= -j219\text{ohms.}$$

Hence the equivalent capacitance is 6.74$\mu\mu$F.

From these figures it is seen that the equivalent capacitance due to the r-f choke and strays across it can be included in the total capacitance across the r-f tuned circuit. Further, depending on the value chosen for the coupling condenser C, this value will be somewhat reduced because it is effectively in series with C, and its effect on the tuning should not be serious.

The value for C is important as it can be chosen so that the variation in the signal voltage applied to the grid of the 6BE6 can be reduced. This follows since the plate is effectively applied to a tap on the tuned circuit. If the full signal voltage were to be applied to the 6BE6 signal grid, zero reactance would be required for C, and it would appear advantageous to use a simple tuned-plate circuit, thereby eliminating the additional choke. The disadvantages of a tuned-plate circuit, however, are that its dynamic impedance increases with frequency (provided Q remains constant) and the gain across the band would be greater at the higher frequencies, and also the input capacitance of the 6BE6 and the output capacitance of the 6BA6 are both directly across the tuned circuit which may make it difficult to keep stray capacitance sufficiently low so as not to affect the tuning range available with the gang condenser used. A value for C is best determined either experimentally or, when the values of stray capacitance, etc., can be accurately measured and the variation of Q across the band of both the choke and the r-f coil have been ascertained, by calculation. Too low a value should not be chosen for C, however, otherwise there will be an excessive reduction in gain. A value of about 20$\mu\mu$F will be found to give an approach to the desired conditions. This value of condenser will also help slightly in reducing the stray capacitance placed across the r-f tuned circuit.

GAIN DUE TO AERIAL AND R-F STAGES

It is now of interest to calculate the aerial and r-f stage gains, and to determine the image rejection at 108 Mc/s, where rejection will be least. The stage gains are calculated at 100 Mc/s.

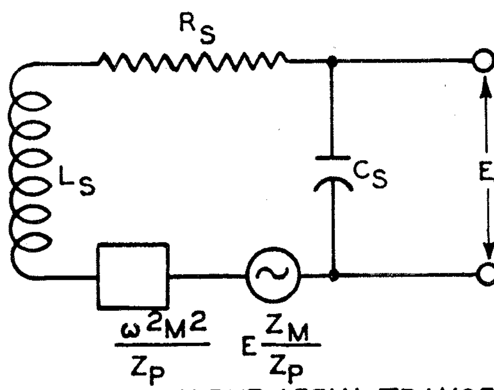

FIG. 7 EQUIVALENT AERIAL TRANSFORMER CIRCUIT.

Proceeding first with the aerial stage and applying Thévenin's theorem, the equivalent circuit is as shown in Fig. 7. The total secondary resistance R_s includes the effect of the damping due to the r-f amplifier valve. From our previous calculations since $Q_s = 33.9$ and $L_s = 0.058\mu H$

$$\therefore R_s = \frac{\omega L_s}{Q_s}$$
$$= \frac{2\pi \times 100 \times 10^6 \times .058 \times 10^{-6}}{33.9}$$
$$= 1.07\text{ohms.}$$

From the equivalent circuit

$$\frac{E_g}{E} = \frac{\dfrac{Z_m}{Z_p}\left[\dfrac{1}{j\omega C_s}\right]}{\dfrac{\omega^2 M^2}{Z_p} + R_s + j\omega L_s + \dfrac{1}{j\omega C_s}}$$

where

Z_p = the total primary impedance.
Z_m = mutual impedance between primary and secondary.
E = primary voltage.

At partial resonance "S"

$$\frac{E_g}{E} = \frac{-M}{C_s[\omega^2(M^2-L_pL_s)+R_oR_s]+L_p}$$

$$= \frac{-M}{C_s[\omega^2L_pL_s(k^2-1)+R_oR_s]+L_p}$$

Then since

$M = 0.0212\mu H.$
$L_p = 0.13\mu H.$
$L_s = 0.058\mu H.$
$k = 0.24.$
$R_o = 80$ ohms.
$R_s = 1.07$ ohms.

$$C_s = \frac{1}{\omega^2 L_s} = 43.6\mu\mu F.$$

$$\omega = 2\pi \times 10^8$$

On substitution in the above equation, or using a simplified expression for the aerial coil gain viz., $\frac{1}{2}\sqrt{\frac{Q_s\omega L_s}{R_o}}$ (derived in Ref. 1), we find that

$$\left|\frac{E_g}{E_s}\right| = 1.96.$$

This is the approximate aerial coil gain at 100Mc/s.

The R-F STAGE GAIN will be given approximately by $g_mQ\omega LA$ where A is a constant depending on the value of C in Fig. 6. It is assumed that the choke, feeding the plate of the 6BA6 r-f amplifier valve, has a Q of 200. The input resistance of the type 6BE6 will be higher than that of the type 6BA6, because there is a negative component of input conductance with this converter, which falls within the group of inner grid modulated types.

Because of the lack of practical data on the input resistance of the type 6BE6 an estimate only can be formed of the r-f gain. Approaching the problem from a somewhat different angle to overcome this difficulty, for negligible attenuation of the f-m sidebands the Q of the tuned circuit should not exceed about 70. Working on this assumption to obtain an idea of the magnitude of the gain and image rejection to be expected, we have

$$\text{Gain} = g_mQ\omega LA \text{ (approx.)}$$
$$= 4{,}400\text{x}10^{-6}\text{x}70\text{x}2\pi\text{x}100\text{x}10^6\text{x}.058\text{x}10^{-6}\text{x}.8$$
$$= 9 \text{ times.}$$

taking A = 0.8 at 100Mc/s. (This figure is near enough if C = $20\mu\mu F$ as suggested.)

Practical experience has shown that stage gains greater than about 10 to 20 times at these frequencies give rise to difficulties with regeneration, so that this gain is satisfactory.

To obtain the desired Q value of 70 it will be necessary to damp the circuit with a parallel resistor, the value depending on the loading already existing across the r-f tuned circuit. The value of the damping resistor will be of the order of 5,000 ohms for a circuit with an unloaded Q of 150.

The CONVERSION GAIN is given approximately by

$$\text{Gain} = \frac{S_cr_pR_L}{r_p+R_L}$$

where

S_c = conversion conductance
= 475μmhos
$r_p = 1M\Omega$

$R_L = \frac{Q\omega L}{2}$ for a critically coupled i-f transformer

$$= \frac{75.2\times 2\pi\times 10.7\times 10^6\times 3.69\times 10^{-6}}{2}$$

= 9,350ohms.

Since $r_p \gg R_L$

Conversion gain = $475\times 10^{-6}\times 9{,}350$
= 4.45 times (approx.)

From these figures the OVERALL GAIN from the aerial coil, using a transmission line having a characteristic impedance of 80 ohms, to the grid of the first i-f amplifier valve is 1.96×9×4.45 or 79 times approximately. This figure will be modified by regeneration, variation in circuit Q's from the assumed values, and the values of the input resistance for the types 6BA6 and 6BE6. Improvement might be effected by tapping down on the r-f and aerial coils, and the possibility of increasing the input resistance for the 6BA6 by only partially by-passing the cathode resistor is worth investigation.

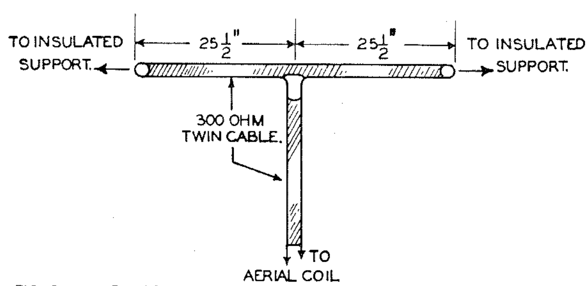

FIG. 8 FOLDED DIPOLE USING 300 OHM FEEDER.

IMAGE REJECTION

For the complete r-f amplifier the image rejection at 108Mc/s is found as follows:

Image rejection due to the aerial circuit alone is given by

$$\sqrt{1+Q^2\left(\frac{f}{f_o}-\frac{f_o}{f}\right)^2}$$

where

f_o = signal frequency = 108Mc/s.
f = image frequency = $108+2\times10.7$.
= 129.4Mc/s.

The nominal Q is 33.9, but this value is reduced to $\frac{33.9}{2}$ by a resistance of 1.07ohms (approx.) reflected from the primary circuit into the secondary, as would be expected from the conditions for matching.

So that

$$Q = 16.95 \text{ say } 17.$$

$$\therefore \text{Rejection} = \sqrt{1+17^2\left(\frac{129.4}{108}-\frac{108}{129.4}\right)^2}$$

= 6.26 times.
or 16db. approx.

The rejection due to the r-f stage

$$= \sqrt{1+70^2\left(\frac{129.4}{108}-\frac{108}{129.4}\right)^2}$$

= 25.6 times

or approximately an additional 28 db, so that the total image rejection is approximately 44 db. If no r-f stage is used the image rejection is approximately 28 db, so that the improvement with the additional stage is of the order of 18 db. This assumes a loaded Q for the aerial coil of about 70, when connected directly to the input of the 6BE6 converter. An improvement in gain of roughly 5 times and an improvement in signal to noise ratio of about 3 times can be expected by using the r-f stage. For those interested in investigating the noise problem the equivalent noise resistance of the 6BA6 can be taken as 3250 ohms, and of the 6BE6 converter 190,000 ohms (see Ref. 11). Tapping the secondary of the aerial coil would appreciably help the image rejection, and the loss in gain would not necessarily be serious, but it is for the individual designer to decide whether the loss in gain is more than offset by the increase in image rejection. By using an unbalanced feeder line and an aerial coil tapped down for both feeder and grid connections it is possible to improve the image rejection by more than 3 times, but the aerial coil gain is reduced to approximately unity.

From the comparatively low Q values of the aerial and r-f stages and because of the high signal frequencies, it should be clear that there will be negligible attenuation of the modulated carrier sidebands. This can easily be checked by calculation if so desired.

In winding the actual aerial and r-f coils $\frac{9}{16}''$ former is suitable, and the windings should be approximately equal in length to the diameter, for high Q values. 18 B & S enamelled wire is suitable, but the aerial coil primary can be wound with say 36 B & S D.S.C. as its r-f resistance is not very important.

OSCILLATOR CIRCUIT

Although many circuits for f-m receivers have not included the usual padding condenser in the oscillator circuit because of the small ratio of maximum to minimum frequencies and their signal and oscillator circuits use two point instead of three point tracking, the additional component appears to be justified.

If the padding condenser is not required, the design proceeds exactly as for a single tuned circuit covering the range 98.2 to 119.2Mc/s.

In this case

$$T = \frac{G_{max}}{\alpha^2-1}$$

$$= \frac{20}{\left(\frac{119.2}{98.2}\right)^2-1}$$

$$= 43\mu\mu F$$

$$L = \frac{25330}{119.2^2\times43}$$

$$= 0.0414\mu H$$

If three point tracking is used we proceed as follows, using the methods set out in detail in Radiotronics 121, taking the centre tracking frequency (98Mc/s) as the arithmetic mean of the band limits.

$$P_{max.} = \frac{G_{max.}}{p-1}$$

$$p = \frac{\alpha^2}{\beta^2}\times\frac{3+\alpha}{3+\beta}\times\frac{1+3\beta}{1+3\alpha}$$

$$= \frac{1.54}{1.465}\times\frac{3+1.24}{3+1.21}\times\frac{1+3\times1.21}{1+3\times1.24}$$

= 1.039 (use logs., slide rule not sufficiently accurate).

$$P_{max.} = \frac{20}{1.039-1}$$

$$= 513\mu\mu F.$$

$$T_{c\,max.} = \frac{G_{max.}}{p\beta^2-1}$$

$$= \frac{20}{1.039\times1.465-1}$$

$$= 38.5\mu\mu F.$$

$$P_{min.} = P_{max.} - T_{c\,max.}$$
$$= 513-38.5$$
$$= 474.5\mu\mu F.$$

$T_L \ll P$ and assume $T_L = 3\mu\mu F$.

$$P = P_{max.} + T_L$$
$$= 474.5+3$$
$$= 477.5\mu\mu F.$$

$P = 475\mu\mu F$ would be near enough as the small changes in L_o, T_c and T_L can be found by the method outlined in the previous article on tracking.

$$\begin{aligned} T_c &= T_{cmax.} - T_L \\ &= 38.5-3 \\ &= 35.5\mu\mu F. \end{aligned}$$

$$L_o = \frac{P_{min.} \; P_{max.}}{T_{cmax.} \; P^2 \; (\omega_2 + \omega_i)^2}$$

$$= \frac{25330 \times 474.5 \times 513}{38.5 \times 477.5^2 \times 119.2^2}$$

$$= 0.0495\mu H.$$

The oscillator circuit components are then

$P = 477.5\mu\mu F.$
$T_L = 3\mu\mu F.$
$T_c = 35.5\mu\mu F.$ (this includes the gang min. cap. of $5\mu\mu F.$ and strays).
$L_o = 0.0495\mu H.$

If the circuit uses two point tracking the oscillator components are

$T = 43\mu\mu F.$ (including gang min. cap. etc.)
$L_o = 0.0414\mu H.$

The oscillator coil is wound on $\frac{9}{16}''$ former using 18 B & S enamelled wire. The circuit arrangement is shown in Fig. 9. The tap on the oscillator coil can be taken as about a third of the total turns for initial adjustment, and then finally set to give an oscillator grid current of about $300\mu A$ with a lower limit of approximately $160\mu A$.

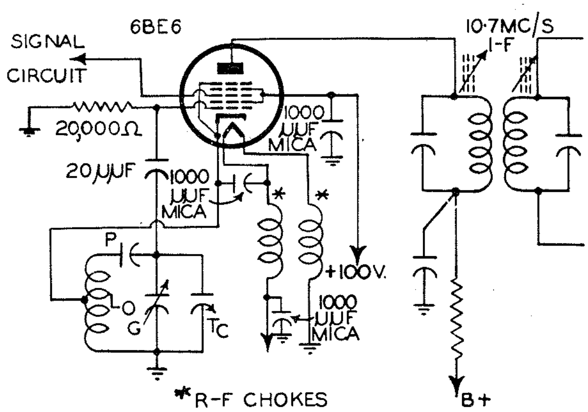

FIG. 9 CONVERTER CIRCUIT.

SUMMARY

Some of the more important points concerning the design of an f-m receiver have been discussed. It is seen that for the most part the design, in many respects, is not very different from that of an a-m receiver covering the same frequency range. Because of space limitations it is not proposed to carry the design further, and such matters as the a.v.c. system, more elaborate audio systems, and so on, are left to the individual designer who can obtain full details from the references quoted below. A number of the finer points in the design have not been treated because of space limitations, and the reader is again referred to the references, particularly in regard to the aerial and r-f stages where many alternative circuit arrangements suggest the possibility of improved performance. A complete receiver circuit with constructional details and performance figures will be given in the next article.

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Radiotron Receiver RD32

By R. H. ASTON, A.M.I.R.E. (Aust.)

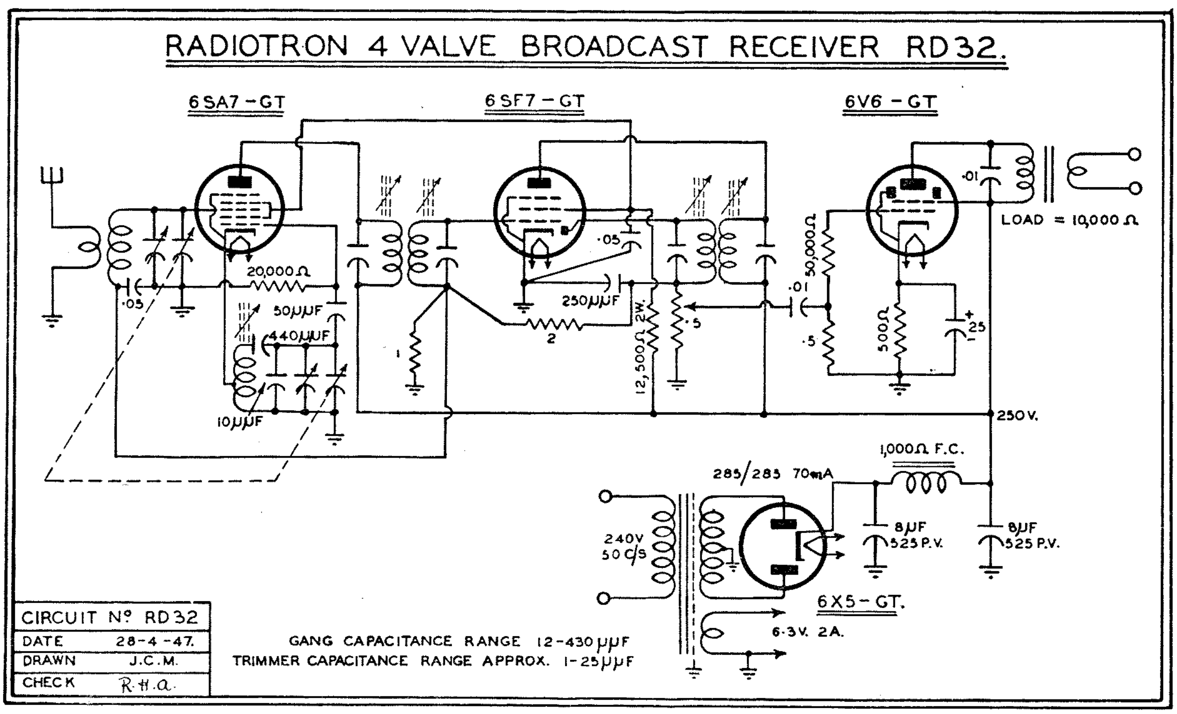
A four valve broadcast receiver design is described which is suitable for production as a small mantel set.

The usual difficulty in the design of a four valve receiver is to achieve sufficient sensitivity to provide reliable reception in average suburban and country areas. It is generally considered that an input sensitivity of not more than 100 microvolts for a 50 milliwatt output is necessary. In a previous issue of Radiotronics (Number 120) a four valve receiver was described which used a reflexed amplifier to achieve the necessary sensitivity. However, the use of reflexing involves some complication and it was thought a simpler design in the same class would be of general interest.

The high transconductance of the single-ended valves makes it feasible to employ them in a quite straightforward arrangement which provides a reasonable sensitivity.

A further increase in gain has been achieved in this receiver by using high gain intermediate frequency transformers which have a Q of the order of 125, and yet have dimensions small enough to fit into the usual mantel set chassis.

A departure was made from our usual practice of neutralizing the intermediate amplifier, as in a receiver of this type the additional gain resulting from a small amount of regeneration is quite useful. It is, of course, dangerous to allow more than a very small degree of regeneration, and, in general, no attempt should be made to introduce positive feedback, but rather slightly relax the precautions to prevent it. A very simple test for the presence of regeneration can be made during the normal alignment procedure. With the signal generator connected to the input of the i.f. amplifier valve, the second i.f. transformer is tuned for maximum output in the normal manner. There can be no regeneration in this condition because the generator is an effective short circuit across the input of the valve. Then, with the generator connected to the input of the converter valve and the first i.f. transformer tuned for maximum output, the second i.f. transformer is retuned. If there is any increase in output above that resulting from the initial setting the presence



of regeneration is clearly indicated, and its magnitude can be judged by the increase in output resulting from this retuning.

In a receiver which feeds the output valve directly from the diode, it is necessary to limit the a.v.c. action so that a sufficiently large audio output is available from the detector. The application to the converter and intermediate frequency amplifier valves of one-third of the total a.v.c. voltage developed satisfies this requirement, and yet provides a reasonable a.v.c. characteristic.

It was considered necessary to make the high tension 250 volts to take advantage of the higher plate resistance of the 6SF7-GT for this condition.

The 6V6-GT output valve is overbiased to reduce general power consumption and heat dissipation. This arrangement will provide quite adequate output voltage for the type of speaker normally used in a mantel cabinet.

TEST RESULTS

(1) Voltage Measurements (AVO Model 7)

	Zero Signal Input			100 mV Signal Input		
Valve	Plate	Screen	Grid	Plate	Screen	Grid
6SA7-GT	250V	94V	0	270V	100V	–14.0
6SF7-GT	250V	94V	0	270V	100V	–14.0
6V6-GT	215	234V	–16	234V	253V	–17
6X5-GT	285Vrms			287Vrms		

Total B Current = 62 mA zero signal and 51 mA for 100 mV signal.

(2) Oscillator

Frequency	e_k	e_o (total coil)	I_{c1}
540 Kc/s	1.45 Vrms	19.4 Vrms	305μA
1600 Kc/s	1.35 Vrms	15.7 Vrms	490μA

(3) Overall Performance

For output of 50 milliwatts (absolute)

Input to	Frequency	Input	Ratio
6V6-GT Control Grid	400 c/s	0.9v	—
6SF7-GT Control Grid	455 Kc/s	32mV	—
6SA7-GT Control Grid	455 Kc/s	390μV	82
6SA7-GT Control Grid	600 Kc/s	520μV	61.6
6SA7-GT Control Grid	1000 Kc/s	500μV	64.0
6SA7-GT Control Grid	1400 Kc/s	480μV	66.7
Aerial	600 Kc/s	75μV	6.9
Aerial	1000 Kc/s	85μV	5.9
Aerial	1400 Kc/s	78μV	6.2

(4) Selectivity

Times Down	Bandwidth
3	5.9 Kc/s
10	10.0 Kc/s
30	14.25 Kc/s
100	19.25 Kc/s
300	24.8 Kc/s
1,000	32.2 Kc/s
10,000	51.7 Kc/s

(5) A.V.C. Characteristic

Input	A.V.C. Volts	Output
10μV	0.3	–24db
30μV	0.9	–9db
100μV	2.2	+0.5db
300μV	4.2	5.0db
1mV	6.2	8.5db
3mV	7.6	11.0db
10mV	9.4	12.5db
30mV	11.2	14.5db
100mV	14.0	17.0db
1V	27.0	26.5db

The Design of a High Fidelity Amplifier

By F. LANGFORD-SMITH, B.Sc., B.E.

(2) Negative Feedback Beam Power Amplifiers and the Loudspeaker

In the first article of this series* it was shown that a loudspeaker, instead of presenting a constant load on the output valve, has an impedance varying in the ratio of about 10 to 1 for a typical loudspeaker on a baffle, and that this impedance ratio could be reduced to something less than 3 to 1 with a rather elaborate loudspeaker system. For general purposes, it seems desirable to design on the basis of an impedance ratio over the whole frequency range of 6 to 1 for a single vented-baffle speaker and 3 to 1 for two such baffles arranged with the impedance peaks staggered. If a flat baffle is used, it will be necessary to adopt the higher figure of about 10 to 1. The earlier article demonstrated the severe distortion which occurs with a pentode or beam power amplifier operated with full grid excitation but without negative feed back. Such an amplifier valve is extremely critical regarding load impedance under conditions of full grid excitation, and in this regard differs markedly from a power triode which, although it has a minimum load impedance for limited distortion, is unlimited in the upward direction.

The obvious question to be asked is, therefore, whether the application of negative feedback to a beam tetrode or pentode would not produce the same characteristics, and similar tolerances as regards the load impedance, as a triode. The answer is given by a glance at Figure 1, which shows the plate characteristics of a typical beam power amplifier with a negative feedback factor $\beta = 0.1$. The normal loadline is marked R_L, and extends through the working point to the knee of the curve as for operation without feedback. It will be seen that, over the greater part of the family of plate characteristics, these are of the general form of triode characteristics giving approximately uniform intercepts along the loadline.

The curve marked "grid current curve" is the equivalent of the zero grid voltage when the valve is used without negative feedback, but in this case it corresponds to the commencement of grid current flow. No loadline for Class A1 or Class AB1 operation should, therefore, extend above this "grid current curve". It will be seen at once that this is the major difference between the characteristics of a triode and those of a beam power amplifier with a negative feedback. As a result of the knee in the "grid current curve", there is a more or less triangular area enclosed between this curve on the one side and the –5 volt input curve on the other hand. So long as the loadline is in its normal position, the intercept between the "grid current curve" and the –5 volt curve is approximately equal to the other 5 volt increments along the loadline, but as the impedance is decreased this intercept becomes smaller until it finally becomes zero at a point where the load resistance is 0.72 R_L. As the load is made still smaller it will be seen that this effect extends to the –7.5 volt and then to the –10 volt curves in succession for load resistances of 0.50 R_L and 0.31 R_L respectively. If the valves were operated under these conditions with maximum grid excitation, the output wave-form would show a flat top in the high plate-current region closely resembling that which occurs under ordinary operation without negative feed-back. Only by reducing the signal input voltage can this flattening be avoided; the normal peak input voltage of 23 volts must be reduced to 15.5 volts for 0.50 R_L and 13 volts for 0.31 R_L.

As the load resistance is increased above R_L, the effect is not quite so sudden and the loadline shown for 2.46 R_L has only a slight amount of distortion provided that the input voltage is restricted to a peak value of 18 volts. If the input voltage is further reduced to a peak of 15.5 volts, the load resistance may then be increased to infinity without showing a flat top.

In the general examination of these curves, no mention was made of the necessity for shifting the loadlines to allow for the rectification effect. This is of no great importance with the higher load resistances since the second harmonic distortion is very small, but it becomes important as the load is decreased below R_L.

The second harmonic distortion calculated from the loadlines without adjusting for rectifications is

Load	Peak Input Volts	H_2
Infinity	15.5	1.5% approx.
2.46 R_L	18	0 ,,
R_L	23	5% ,,
0.72 R_L	18	11.5% ,,

and increases rapidly thereafter.

A tentative conclusion may, therefore, be reached that for a negative feedback factor $\beta = 0.1$, the input voltage should be reduced to about 70% of its

*Radiotronics 124, page 25.

maximum value without feedback for reasonably low distortion with wide variations of load resistance. This is equivalent to about 50% of the power output obtainable without feedback.

If the valve is operated under Class A1 conditions and the load resistance is only permitted to vary over the ratio 3 to 1, the input voltage may be reduced to 80% of the maximum value without feedback resulting in a power output of about 64% of that obtainable without feedback.

Increased Negative Feedback

An increase in negative feedback causes the characteristic curves to approach more nearly to the vertical, thus reducing the distortion which occurs as the load resistance is increased above R_L.

The bad effects which occur with load resistances below R_L are unchanged, and grid current will flow before reaching full grid excitation. Cathode loading is merely a special case of $\beta = 1.0$, and the same remarks apply; it may be used with full excitation for load resistances from R_L to infinity, but must have reduced grid excitation for low load resistances, if grid current is to be avoided.

Thus the reduction in grid excitation, to permit load resistance variation from R_L to some specified maximum value, may be less severe as the feedback factor (β) is increased above 0.1. No reduction is necessary when the slope of the characteristic curves is equal to the slope of the "grid current curve" below the knee. A slight increase in grid excitation is possible for still higher values of β, although this is only of academic interest.

The Effect on Amplifier Design

If it is desired to use normal factors of negative feedback with beam power amplifiers operating on a loudspeaker load, it appears to be essential to allow for reduced grid excitation which is a function of the variation in load impedance presented by the loudspeaker. Two beam power amplifier valves giving a power output of 30 watts without negative feedback would, therefore, be required to operate with reduced grid excitation so as to give a power output of only 15 or 20 watts when operated with a negative feedback factor $\beta = 0.1$., on a loudspeaker of normal characteristics, if the distortion is to be kept low. Higher power output for a limited distortion may be obtained by increasing the feedback factor, although this may involve a more elaborate circuit to avoid instability at very low audio or very high (supersonic) frequencies.

CORRECTION TO RADIOTRON AMPLIFIER A514

In the circuit diagram of this amplifier which was published in Radiotronics No. 124, the wattage ratings of the resistors comprising the screen voltage divider were incorrectly shown. They should be: B+ to screen 40W, screen to cathode 40W, cathode to earth 8W.

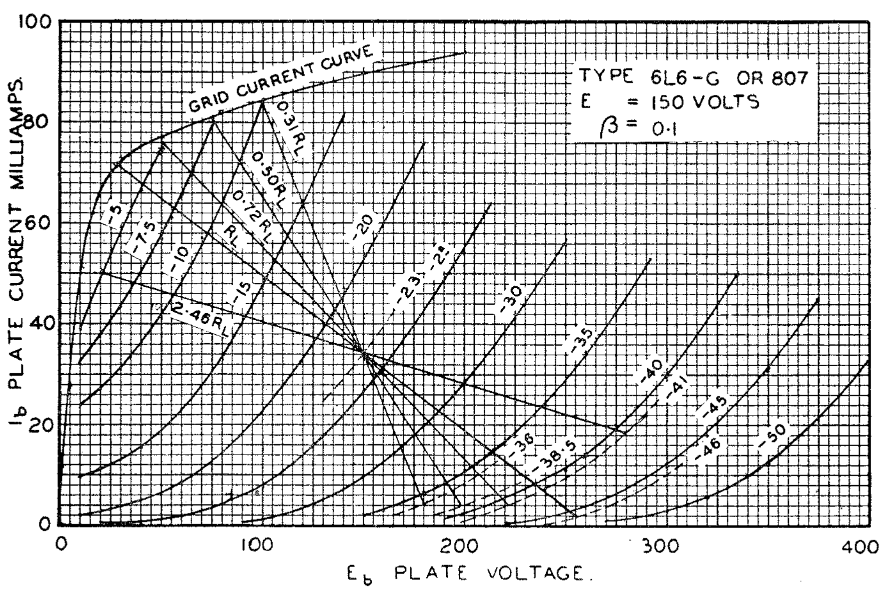


Fig. 1. Plate Characteristics of a typical beam power amplifier with negative feedback (6L6 or 807 with $\beta = 0.1$) showing effect of varying load resistance.

Adapting 115 Volt A.C./D.C. Receivers for 240 Volt Operation

An article published in Wireless World for July, 1946, discusses an interesting point in connection with the adaption of 115 volt a.c./d.c. receivers to 240 volt operation. The usual voltage dropping arrangement is shown in figure 1 with part of the circuit to the right of the dotted line representing the

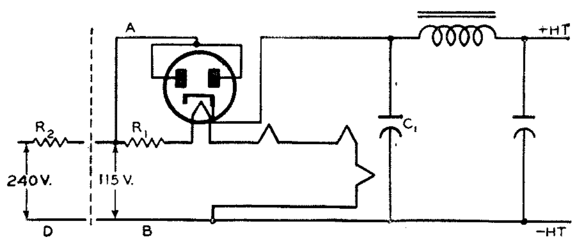


FIG. 1

115 volt condition and to the left of the dotted line the additional resistor generally added for a 240 volt supply. While this arrangement is quite satisfactory for a d.c. supply, it has serious disadvantages when used with an a.c. supply. The reason for this is as follows:—The half wave rectifier only passes current when the anode is positive with respect to the cathode, so that even during the positive half cycles, current only flows when the supply voltage exceeds the potential to which the filter capacitor C_1 has dropped during the non-conducting period of the diode. During this brief period when the rectifier conducts, sufficient current has to flow into C_1 to maintain its charge over the whole cycle. Consequently the peak current flow through the rectifier may be of the order of ten times the average high tension drain. As the actual value of the pulsed current drawn from the line is generally unknown, the most convenient method of determining the correct voltage dropping resistor is by experiment.

In the case of the current supplied to the valve heaters the drain is proportional to the instantaneous supply potential, and for a sine waveform has the usual relations between peak and rms, etc. Thus the voltage dropping resistor for this branch of current can be easily calculated on an rms basis.

Thus, while the dropping resistors necessary for the two current branches can be easily determined individually, it is impossible to combine them because of the different natures of the current waveforms involved. The only satisfactory approach to the problem is to use separate voltage dropping resistors

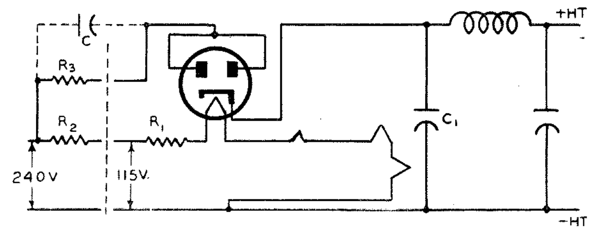


FIG 2

for the heaters and the high tension as shown in figure 2. However, if the simple series resistor is used to reduce the high tension with an a.c. supply, its value would be far too small if the set were connected to a d.c. supply of the same voltage.

To make the receiver useable on a.c. or d.c. mains, R_3 should be experimentally determined for the d.c. case and shunted by a capacitor of a value—also experimentally determined—which will result in correct high tension voltage from an a.c. supply. The capacitor used should not be an electrolytic type. In a practical case, R_3 might be of the order of 2000 ohms and C of the order of 5 microfarads.

The arrangement of figure 2 utilizes the reactance of the capacitor C. It might be argued that this voltage drop would not be constant for different supply frequencies. To avoid such discrepencies the

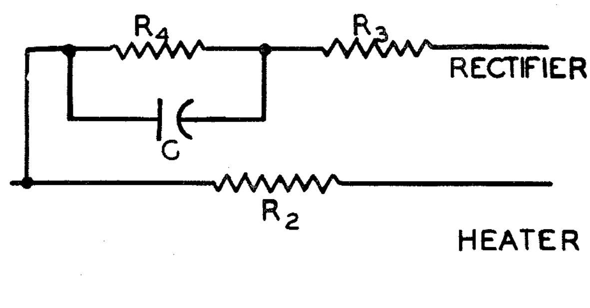


FIG. 3

alternative arrangement, shown in figure 3, can be used. In this case, typical values for 50 c/s would be $R_3 = 500$ ohms, $R_4 = 1{,}400$ ohms and $C = 22.7\ \mu F$.

From the foregoing discussion it can be seen that the "linecord" resistor, often used with American receivers on high voltages is quite satisfactory when the receiver is to be used on d.c. mains only, but the practice is open to serious technical objections when the receiver is connected to an a.c. supply.

Type 807 as a High-mu Triode

We were recently approached for an opinion on the performance of a type 807 as a high-mu triode with control grid and screen tied together. Tests carried out on a few valves gave the following results at a plate current of 25mA:—

Amplification factor	195
Mutual conductance	4,500 μmhos
Plate resistance	43,000 ohms.

The plate current at zero bias was approximately as under:—

Plate Votage	Plate Current
200 volts	1.9 mA.
300	2.9
400	3.7
500	4.8
600	5.6
700	6.7

The tests at positive grid voltage were very disappointing, owing to the heavy grid current. It can therefore be taken that type 807 as a high-mu triode, is not suitable for use as a zero-bias Class B Amplifier. It may, possibly, find a limited application as a high-gain resistance-coupled amplifier operating in the negative grid region.

BARRETTERS

Type 302 barretters should be mounted base downwards with the filaments approximately vertical. Horizontal operation should be avoided as barretters have comparatively weak filaments, considerably weaker than those of the tungsten filaments used in lamps, and care should be taken to avoid unnecessary stress on the filament. Vertical mounting also permits better ventilation and the minimum of heat on the base.

Everything possible should be done to protect barretters from excessive vibration, both in transit and in operation.

ERROR IN RADIOTRONICS 123

On page 14 of Radiotronics 123 are shown the average constant-current characteristics of a valve. These curves apply to the Radiotron 4—125A/4D21, data for which was printed in Radiotronics 121. Obviously the curves have no connection with the Radiotron type 8020, and were inserted by error.

New R.C.A. Releases

Radiotron type 3JP7—is a short cathode-ray tube having unusually high spot intensity, high grid-modulation sensitivity, and high deflection sensitivity. Type 3JP7 is designed with a high-voltage accelerator anode to increase the intensity of the fluorescent spot. The tube in intended for use in oscillographic applications where a temporary record of electrical phenomena is desired.

Radiotron types 5CP1-A, 5CP7-A, 5CP11-A—are a series of five inch cathode ray tubes utilizing electrostatic focus and deflection. They differ from one another only in the spectral—energy emission and persistence of their respective phosphors P_1, P_7, P_{11}.

The types in this series are designed with a high-voltage accelerator electrode (anode No. 3). This electrode permits the use of a high intensity, fluorescent spot with minimum sacrifice in deflection sensitivity, and with a slight increase in spot size.

Radiotron type 12AW6—is an r-f pentode of the miniature type with a sharp cut-off characteristic and a 12.6 volt, 150 milli-ampere heater rating. Its high transconductance in combination with low values of input and output capacitance makes the type 12AW6 particularly useful as an r-f or i-f amplifier in ac/dc FM receivers.

Radiotron type 5527—is a small, television camera tube employing electrostatic deflection, intended for use in industrial applications, for television experimentation in laboratories, and for demonstrating television principles in schools. It is designed so that it can be operated with equipment which is simple and relatively inexpensive. The resolution capability is approximately 250 lines.

Radiotron type 5581, 5582, 5583, 5584—are gas phototubes featuring very high sensitivity to light sources predominating in blue radiation, and no response to infra-red radiation. They are particularly suitable for use in sound reproduction involving a dye-image sound track in conjunction with an incandescent light source. They may also be used in measurement and colour control applications.

With minor circuit changes the type 5581 can be used in place of the type 930, the type 5582 in place of type 921, type 5583 in place of the type 927, and type 5584 can replace type 920.

Radiotron type 5588—is a very compact, forced-air-cooled power triode designed for ultra-high frequency service. It has a maximum rated plate dissipation of 200 watts, and can be operated with full plate voltage and plate input at frequencies as high as 1200 Mc/s. Operation at higher frequencies is permissible with reduced ratings,