Radiotronics

Number 127 SEPTEMBER—OCTOBER 1947

A GENERAL VIEW OF GRID WINDING MACHINES FABRICATING DELICATE "GRID" COILS AT THE RADIOTRON VALVE WORKS, ASHFIELD, N.S.W.

RADIOTRONICS

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ERRATUM. Page 87 Column 1. Receiver Alignment. Paragraph 2 seventh line, word "advantage" to read "disadvantage"

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RADIOTRON CHART

The new "Radiotron Characteristic Chart", incorporating classification tables, valve socket connections and substitution directory is now available. A copy will be posted to any Radiotronics subscriber upon application to our Sales Promotion Department.

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Our Cover . . .

Shows a battery of grid-winding machines operating within tolerances of 1/1000 part of an inch to fabricate the "grid" coils for use in Australian-made Radiotron Valves.

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

(Part 2).

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

In the previous article general design features were discussed and preliminary details given for a typical f-m receiver using a ratio detector. This article describes the actual receiver constructed and gives sufficient data to allow a similar experimental model to be made.

Introduction

So as to permit cross reference between this and the previous article the description will be given in a similar order as far as possible.

It will be noticed that there are several minor differences in the r-f section of the receiver such as the elimination of filament chokes, and the padding condenser in the oscillator circuit. The reasons for these changes will be discussed as the description proceeds.

I-F Amplifier

The i-f amplifier was made to conform as closely as possible to the design conditions. Each transformer is wound on $\frac{3}{4}''$ former, the primary and secondary each being close wound solenoids consisting of $11\frac{1}{4}$ turns of 22 B & S enamelled wire. The spacing between the primary and secondary windings is 23/32″. Kingsley slugs are used for tuning and these are $\frac{1}{2}''$ long and $\frac{3}{8}''$ diameter. Fixed condensers each of 50 $\mu\mu$F are shunted across each winding. The i-f cans are $1\frac{5}{8}''$ diameter and $3\frac{1}{2}''$ long.

The Q's of the primary and secondary were individually adjusted to give the desired Q of 75.2, and this necessitated a damping resistor of 80,000 ohms in each case. The coefficient of coupling was adjusted to 0.95 of critical on the Q meter by altering the spacing between the primary and secondary windings. All measurements were made with the transformer in the can.

The method used was to tune the secondary for maximum absorption when the primary is connected across the Q meter and tuned by a capacitance of 60 $\mu\mu$F. The change in Q is noted and the coefficient of coupling found from the relationship

$$\frac{k}{k_{crit.}} = \sqrt{\frac{Q_1 - Q_2}{Q_2}}$$

where:

Q_1 is the Q of the primary when the secondary is untuned.

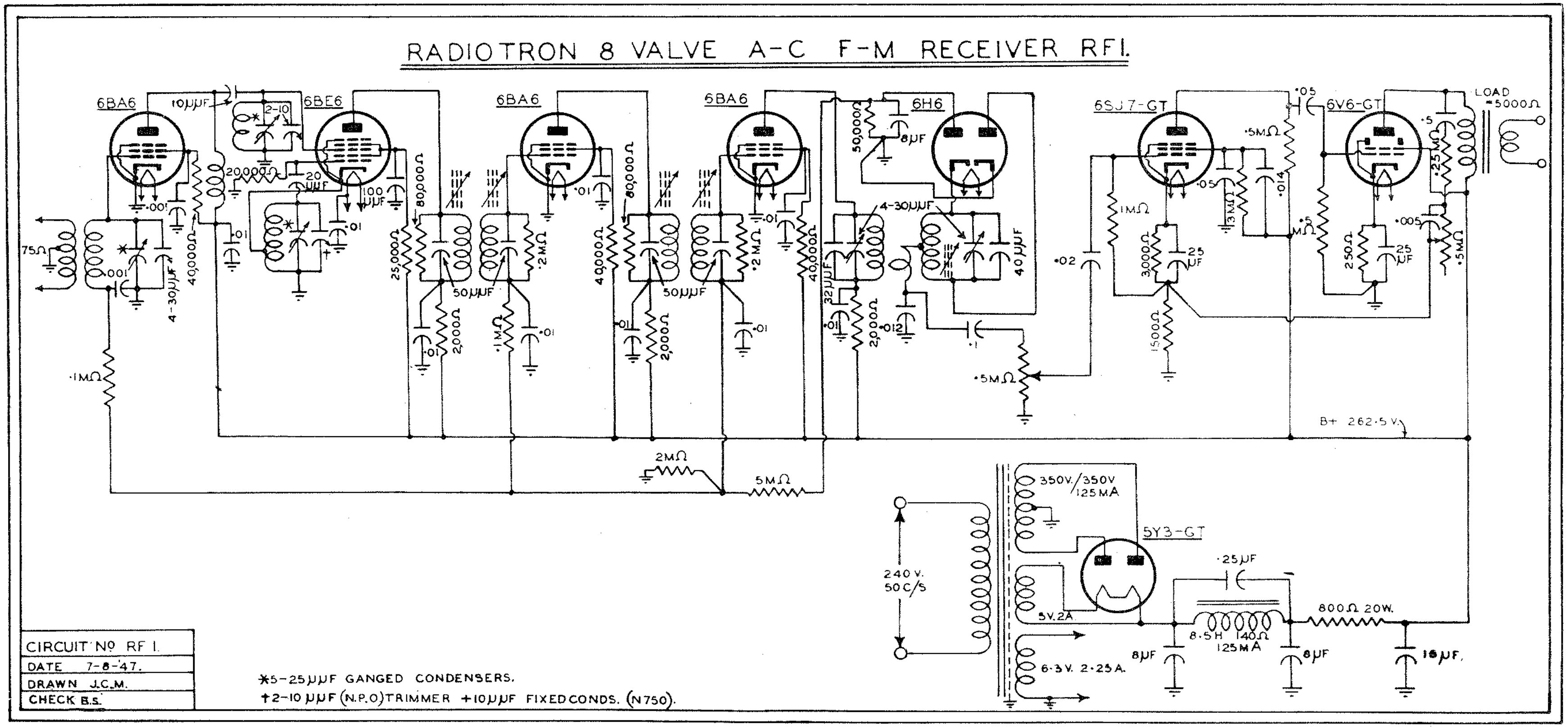
Q_2 is the resultant Q of the primary when the secondary is tuned for maximum absorption (60 $\mu\mu$F condenser across secondary).

From the relationship given it is seen that critical coupling occurs when the primary Q is halved due to the presence of the tuned secondary. A value less than critical is always chosen with this arrangement because of the additional coupling which occurs when the transformer is wired into a receiver. Lower values than 0.95 of critical are preferable in receivers where stray coupling is likely to be high, and it is common practice to use as low as 0.8 of critical in ordinary 455 Kc/s i-f transformers. The effect of using slightly less than critical coupling is to increase selectivity near the "nose" of the selectivity curve and to decrease selectivity at frequencies well off resonance; the change in gain is very small and can usually be neglected.

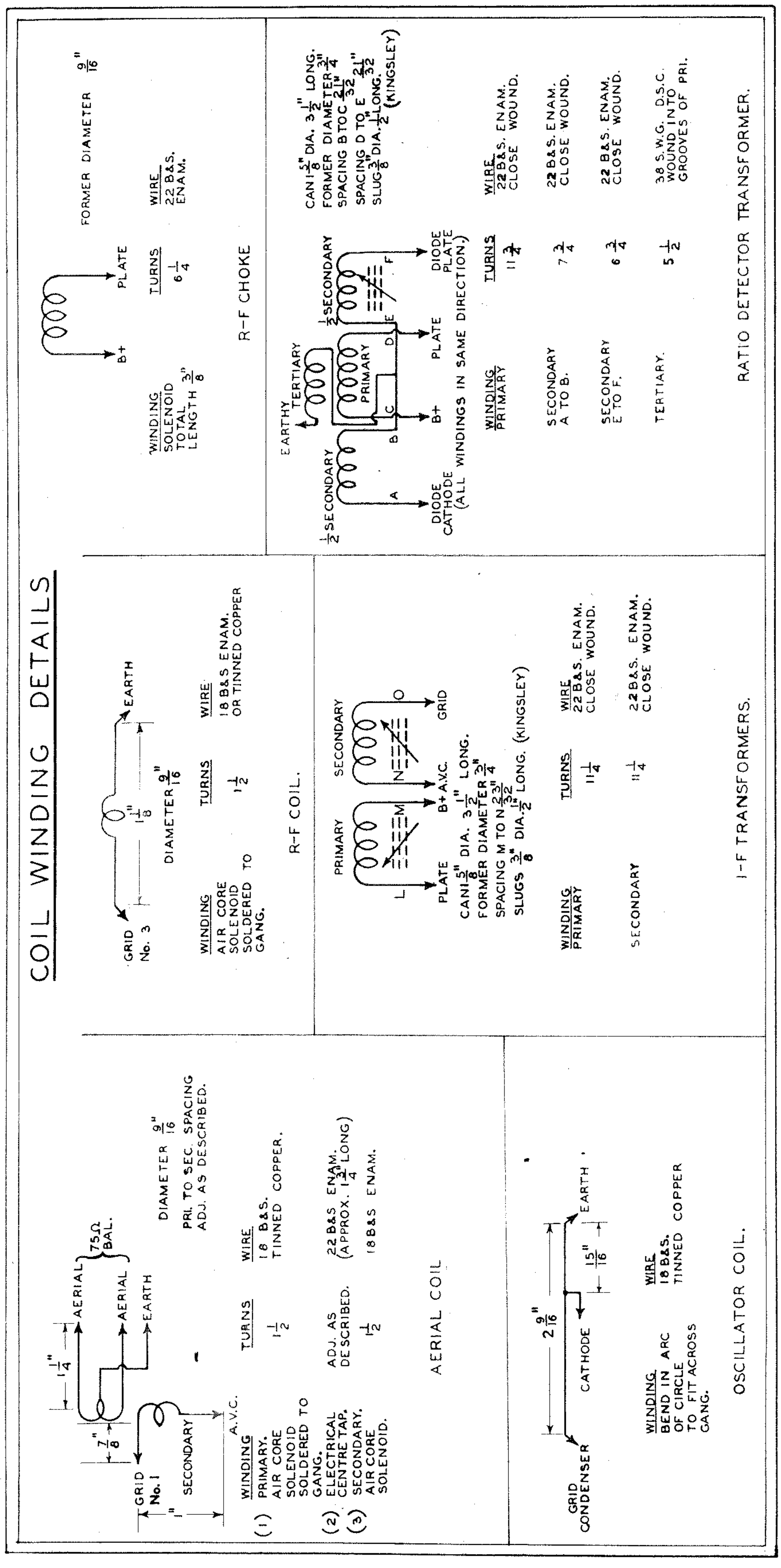
If the additional coupling does not give exactly critical coupling, or very near it, this will show up when the measured selectivity curve is compared with the calculated curve. This can be checked initially by the bandwith at the half power points, and then making any necessary adjustment in the primary to secondary spacing. It is preferable, of course, to effect any alterations to the transformer before wiring it into the receiver and in most cases a suitable relationship connecting the measured coefficient of coupling to critical coupling can be found, making unnecessary the adjustment of spacing between windings, in the receiver.

On the circuit diagram it will be seen that 80,000 Ω has been connected across the i-f transformer primary and 0.2 M Ω across the secondary. The value of 0.2 M Ω comes about due to the input resistance of the 6BA6 at 10.7 Mc/s. A reasonably close approximation to the input resistance can be found by using a square law extrapolation to the figures of input resistance quoted in RCA Application Note 118 (reprinted in Radiotronics 126). At 100 Mc/s the input resistance is given as 1600 Ω. Then the input resistance at 10.7 Mc/s is approximately 0.14 M Ω. This value in conjunction with 0.2 M Ω gives a figure reasonably close to 80,000 Ω and is near enough in a practical circuit, since it avoids the necessity for using odd resistance values.

Since the i-f selectivity curve was made to conform fairly closely to the calculated curve it has not been repeated here.



Circuit Diagram of F-M Receiver for the signal frequency range 88-108 Mc/s. The intermediate frequency is 10.7 Mc/s and the oscillator tuning range is 98.7—118.7 Mc/s. An additional 0.5 Mc/s is allowed at each end of the band; making the maximum coverage 87.5—108.5 Mc/s for the signal circuits and 98.2—119.2 Mc/s for the oscillator tuning range.



Coil winding details for Radiotron 8 Valve A-C F-M receiver RF1. All the signal and oscillator coils require adjustment in the receiver, at the operating frequencies, as they form only part of the tuned circuit inductances.

The gain obtained from a single complete i-f stage, i.e. from the grid of the first 6BA6 to the grid of the second 6BA6, was 38 times. This is lower than the value of 41.1 times previously calculated and the loss in gain was found to be partly due to the g_m of the 6BA6 being somewhat lower than 4,400 micromhos. Complete i-f performance figures are quoted in the summary at the end of the article.

It is considered that capacitance tuning of the i-f's would be somewhat preferable to "slug" tuning as then greater control of the circuit components can be exercised. For example the stray capacitances across the i-f transformer windings may not be exactly 10 $\mu\mu$F, as used in the calculations, and so slight alterations are made in the inductance values, by moving the "slug" positions, which alters the circuit conditions somewhat. This alteration need only be small, and in most cases is unimportant, but it does introduce a possibility of error that can be reduced if desired. The final choice between the two methods will undoubtedly depend on practical convenience.

Smaller coil cans than those specified could be used as there is no difficulty in obtaining the required Q's. With a correct choice of wire gauges, and a smaller can, probably the damping resistors could be eliminated, since small variations from the specified Q are not particularly important.

Ratio Detector

The details for the ratio detector transformer used are as follows:

Former $\frac{3}{4}''$ diameter.

Primary winding: $11\frac{3}{4}$ turns 22 B & S enamelled wire close wound. Placed at centre of former.

Secondary winding: wound in two halves symmetrically placed in regard to the primary. The half winding which is "slug" tuned has $6\frac{3}{4}$ turns of 22 B & S enamelled wire close wound; the slug being $\frac{3}{8}''$ diameter and $\frac{1}{2}''$ long. The other half of the secondary winding has $7\frac{3}{4}$ turns of 22 B & S enamelled wire close wound. The spacing in each case between primary and secondaries is $\frac{21}{32}''$.

Tertiary winding: $5\frac{1}{2}$ turns of 38 S.W.G. D.S.C. wire wound into the first five grooves formed by the turns of the primary winding; starting from the B+ end of the primary winding. All windings on the transformer are in the same direction.

The can dimensions are $1\frac{5}{8}''$ diameter and $3\frac{1}{2}''$ long.

It will be noticed on the circuit diagram that a value of 50,000 Ω has been chosen as the load resistor for the diode circuits. Lower values of load resistance give decreased sensitivity, but have the desirable effect of decreasing the time constant of the a.v.c. circuit, and further, give very slightly reduced distortion at audio frequencies above about 10 Kc/s. No undesirable effects were noted in the tuning of the receiver because of the long a.v.c time constant.

Tests made with lower values of load resistance showed that the sensitivity of the ratio detector, from the driver stage, was almost directly proportional to the value of load resistance used. One such test showed that for 18 millivolts at i-f (deviated ± 75 Kc/s) applied to the 6BA6 driver stage grid gave 50 milliwatts of output power using a 50,000 Ω diode load. For a 30,000 Ω load the input required for the same output was 30 millivolts. Somewhat similar effects were found with higher and lower values of load resistors provided they were not more than about two to one away from 30,000 Ω as a reference point. Distortion became rather severe for very high load resistance values and decreased somewhat for lower values. Lower values than 50,000 Ω may be advantageous because of these factors, and the desirability of a more rapid time constant for the a.v.c. circuit, but this is left to the individual to decide experimentally for himself.

The bandwidth of the complete detector circuit was measured statically using a normal a-m generator with 50% modulation. The peaks of output under this condition were at 10.86 Mc/s and 10.54 Mc/s, giving a total bandwith of 320 Kc/s. It was not considered worthwhile to alter this value to bring it down to 300 Kc/s. A point which was observed was that the peaks were not exactly the same distance apart when the signal strength was altered. Stronger signals increased the apparent bandwidths. The figure of 320 Kc/s was obtained with 70 millivolts input to the 6BA6 driver stage grid.

It should also be appreciated that the bandwidth under dynamic conditions may not be identical with that obtained using static conditions because of the variable load presented across the discriminator tuned circuits by the ratio detector, and a preferable method would be to use a f-m generator having a very wide sweep and observing the output on a C.R.O. Since this type of generator was not available the static test was used.

In regard to setting the bandwidth it was found that the "slug" in the half secondary could be adjusted so as to have some effect on the high frequency peak and the primary tuning condenser could be made to affect the position of the low frequency peak. These two adjustments also affected the relative amplitudes of the two peaks and a compromise setting, after coil spacing was set, allowed symmetrical spacing of the peaks about 10.7 Mc/s, giving relative peak values which differed by less than half a decibel with 70 millivolts input. Greater unbalance occurred with larger inputs.

A direct check on the discrimination against amplitude modulation due to the ratio detector did not allow exact figures to be obtained, but the following method was used to indicate that amplitude discrimination occurred, and the reader may form his own conclusions. First it is necessary to realise that an ordinary discriminator circuit, if exactly balanced, would not pass through a-m on an f-m i-f carrier centred at the intermediate frequency. If the central reference frequency of the i-f carrier is not exactly say 10.7 Mc/s but is, for example, 10.75 Mc/s, then some of the undesired a-m on the f-m signal will pass through the detector. From this, it is necessary to separate out of the insensitivity to a-m, due to a discriminator circuit alone, from that obtainable with the ratio detector. Whatever improvement results can then be compared with that obtainable from conventional limiter circuits, locked oscillators, etc. This does not consider any other advantages that may accrue due to the use of the ratio detector, but merely considers its properties as an amplitude limiter.

A carrier simultaneously frequency and amplitude modulated was applied to the input to the ratio detector. The audio frequency causing f-m was 400 c/s and the audio frequency causing a-m was 3000 c/s. The f-m input gave ± 75 Kc/s deviation and the a-m input 50% modulation. The audio output from the ratio detector was applied to a wave analyser and the relative amplitudes of the desired and undesired signals measured. The figures obtained were:

At centre frequency

$$\frac{\text{amplitude of 400 c/s voltage}}{\text{amplitude of 3000 c/s voltage}} = \frac{272.5}{1}$$

At 50 Kc/s off centre frequency

$$\frac{\text{amplitude of 400 c/s voltage}}{\text{amplitude of 3000 c/s voltage}} = \frac{150}{1}$$

At 100 Kc/s off centre frequency

$$\frac{\text{amplitude of 400 c/s voltage}}{\text{amplitude of 3000 c/s voltage}} = \frac{60}{1}$$

The same procedure was then carried out on a straight-out balanced phase discriminator. This discriminator was one specially designed for use in an f-m station monitor and is kept in a temperature controlled oven. Its balance is probably much better than could be obtained with discriminators for use in radio receivers. The results obtained were:

At centre frequency

$$\frac{\text{amplitude of 400 c/s voltage}}{\text{amplitude of 3000 c/s voltage}} = \frac{666}{1}$$

At 50 Kc/s off centre frequency

$$\frac{\text{amplitude of 400 c/s voltage}}{\text{amplitude of 3000 c/s voltage}} = \frac{62.5}{1}$$

At 100 Kc/s off centre frequency

$$\frac{\text{amplitude of 400 c/s voltage}}{\text{amplitude of 3000 c/s voltage}} = \frac{33.3}{1}$$

It does not appear quite fair to directly compare these two circuits so that if we take the relative improvement a somewhat better idea of the results is gained. Taking the case of 50 Kc/s off centre frequency, for the discriminator alone the deterioration in a-m rejection is $\frac{666}{62.5}$ or 10.64 times worse. For the ratio detector the deterioration is $\frac{272.5}{150}$ or 1.82 times. If the assumption is made that the two circuits can be made equally good at the centre frequency, and that the other results are directly proportional to the degree of rejection at the centre frequency, we can say that the ratio detector appears to offer an improvement of at least 5.85 times at 50 Kc/s off centre frequency and that possibly this figure would be improved upon in a very well balanced circuit. It is clear from the discussion why the figures cannot be taken as exact, but serve as a guide only. Other modulating frequencies and different amplitudes for the input voltage would further modify the results obtained. Harmonic and intermodulation effects are not considered.

The alignment of the ratio detector is as simple as an ordinary discriminator, once the correct "slug" position is obtained, and will be described in the section on receiver alignment.

High perveance diodes, such as type 6AL5, and elaborate balancing circuits have been used to obtain better ratio detector performance, but as this information was not available when the receiver being described was developed, these newer circuits have not been tried out. In any case a type 6H6 is the only diode at present readily available in Australia, of the medium to high perveance type, which could be satisfactorily incorporated in the circuit.

Audio Amplifier and Power Supply

As stated previously the audio amplifier is almost identical with that used in the Radiotron RC52 receiver described in Radiotronics 117. A reduction in feedback was made, however, as improved audio gain was desirable. The overall performance was not noticeably impaired and the only real change, as would be expected, was the reduction in the maximum obtainable bass boost to approximately 6 db at 60 c/s, which seemed quite adequate on listening tests.

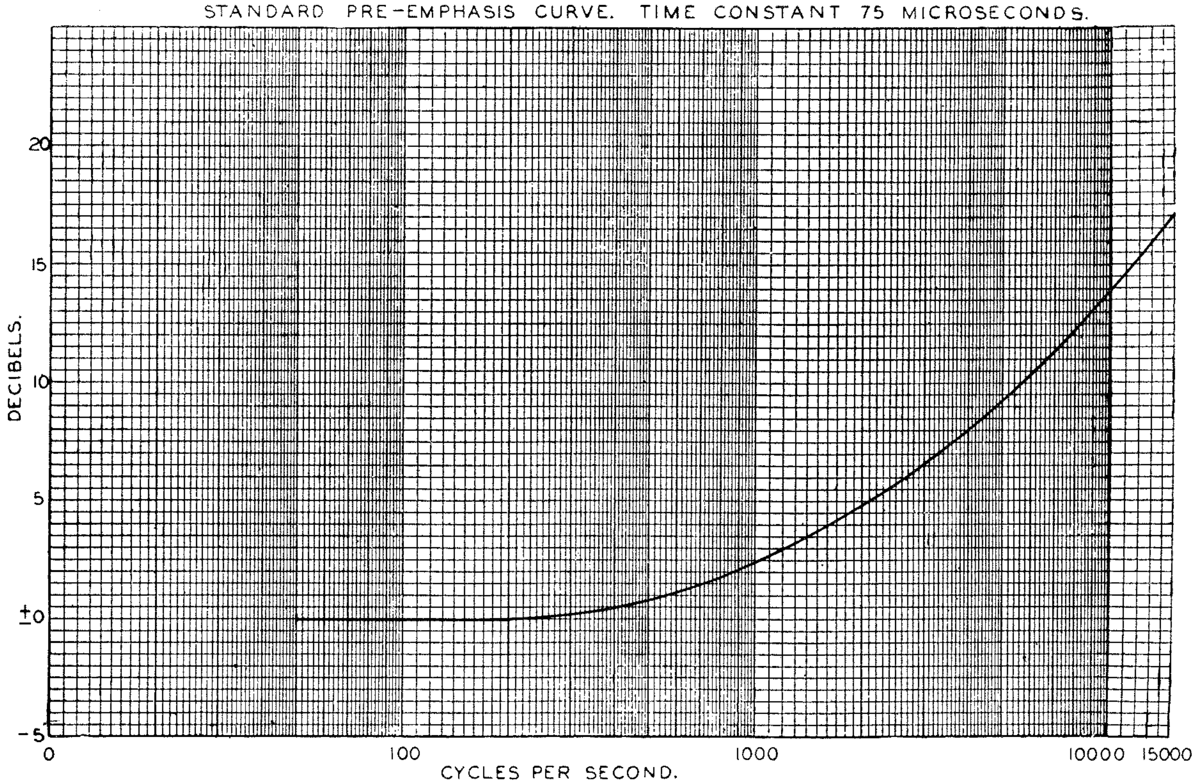
Bass boost seemed particularly desirable because of the improved high frequency response, and this was adequately borne out on a number of listening tests particularly on recorded music. Treble attenuation would also appear desirable when listening to some ordinary types of recordings as the record hiss is rather objectionable.

Tests made on the complete audio amplifier gave results, particularly in regard to overall frequency response, which were much better than those quoted in Radiotronics 117. For this reason the results are given below. In this respect it is interesting to note that the total variation in audio power output from 50 c/s to 15 Kc/s, when measured on a standard 2M8832 A.W.A. output meter, was only $1\frac{1}{4}$ db. The power output level was $\frac{1}{2}$ watt. A voltage input to the amplifier of approximately 0.26 volts at 400 c/s gives an output of ½ watt. The gain reduction due to feedback is 7.36 times.

The tests that could be made for harmonic distortion were limited by lack of equipment to measure distortion at fundamental audio frequencies higher than about 7,500 c/s. Tests were made at several frequencies for ½ watt output, and these indicated that the total r.m.s. distortion was less than 1%. Oscillographic observation up to 15 Kc/s was made, and as no noticeable distortion could be observed, it was concluded that the distortion was under 3%, since past experience has shown this to be about the least observable.

the required curve was less than 1 db at any point in the range from 100 c/s to 15 Kc/s. It should be clear that the output from the variable frequency oscillator should not be applied through a standard 75 microsecond pre-emphasis network, as the deviation frequency of the f-m generator would vary with the applied voltage, and there are no standards available to determine how much the deviation should alter at a particular frequency, as the deviation may be set at ± 75 Kc/s at 400 c/s, but would be much greater than this at say 10 Kc/s. The required deviation at the higher frequencies could be calculated, but this does not seem to be worthwhile, particularly as most f-m generators could not give the extra deviation required.

STANDARD PRE-EMPHASIS CURVE. TIME CONSTANT 75 MICROSECONDS.



All measurements were made on the audio amplifier with the input voltage applied across the volume control, and the distortion figures do not include distortion introduced by the ratio detector.

To obtain the required 75 micro-second de-emphasis an f-m signal generator, set at 10.7 Mc/s and connected to the 6BA6 in the driver stage, was modulated with a variable frequency oscillator, the deviation held constant at ± 75 Kc/s, and the audio output from the amplifier made to fall off in exactly the opposite way to the normal pre-emphasis curve shown. This result was achieved by setting the value of the condenser at the earthy end of the tertiary winding in the ratio detector transformer so as to give the required response curve. The variation from

The *power supply* is a conventional one except that the choke is tuned to reduce hum voltages in the output to a minimum. Because of the good low frequency response of the amplifier the three electrolytic condensers shown are necessary, and in some cases even better filtering may be desirable. Increasing the 16 μF condenser to 32 μF is helpful if the audio amplifier shows any tendency to "motor-boat".

Hum neutralization in the audio amplifier is utilized by adjusting the 0.014 μF condenser, across the 3M Ω resistor in the screen circuit of the type 6SJ7-GT for minimum hum output. Incorrect values for this condenser can lead to instability, particularly when the values are too high. The hum level in the final circuit is so low as to be almost completely inaudible on normal listening tests, even with the tone control in the bass boost position.

Decoupling is shown in all the i-f plate circuits, but if the layout is reasonably good, and the length of B+ lead from the nearest by-pass condenser is small, it is possible to eliminate the decoupling circuit to the plate of the first i-f amplifier. With due care other decoupling circuits could doubtless be eliminated, but some precautions are necessary because of the high intermediate frequency.

The Signal and Oscillator Circuits

Oscillator Circuit

The first change to be noted in this circuit is the elimination of the padding condenser. This condenser was not used because of the greater practical convenience afforded by connecting the oscillator coil straight across the tuning condenser, thereby eliminating the need for a coil former, or, alternatively, a mounting post for the coil and padder junction. Deterioration in tracking between the signal and oscillator circuits was remarkably small. Two point tracking was obtained at 88 and 108 Mc/s, and the error near the centre of the band, where the tracking error would tend to be greatest, was only about ½ db. The complete tuning range is 87.5 to 108.5 Mc/s, and the oscillator tunes from 98.2 to 119.2 Mc/s.

The next change is the elimination of all r-f filament chokes. These were found to give no noticeable improvement and so were removed. One of the arguments advanced for using chokes in the oscillator heater leads is that the heater-cathode capacitance alters as the valve heats up, and as this capacitance is directly across portion of the oscillator tuned circuit some change in frequency occurs. To overcome this it is suggested that the heater be connected to the cathode for r-f, by means of a large condenser. Chokes are then required to prevent the heater-cathode circuit from being by-passed to ground by the heater leads. It appears, however, that the chokes could introduce appreciable frequency drift in themselves, and although some improvement may possibly be obtained, it was decided to keep the number of components to a minimum unless their introduction showed some definite improvement.

By-passing the heater leads to earth with a 0.01 μF condenser is desirable, as it reduces coupling between stages through the heater leads.

Complete temperature runs were not made on the oscillator circuit, but some checks were made (at room temperature) of oscillator frequency change with mains frequency. The receiver was tuned to a signal of 108 Mc/s generated by a crystal calibrator. At the same time a 10.7 Mc/s signal from another crystal calibrator was fed into the i-f amplifier. Any variation in the receiver oscillator frequency from that required to give exactly 10.7 Mc/s with the incoming signal then shows up as a beat note in the receiver output. The frequency can then be measured with a frequency meter or some similar arrangement.

With mains variations from 230V to 270V the initial frequency change was found to be of the order of $\pm$ 5 Kc/s but, after approximately 30 seconds when the heater temperature had stabilised, the frequency change dropped to less than $\pm$ 2 Kc/s in every case. The frequency change does not remain constant, however, but varies about the reference point within the limits stated.

Any possible changes in oscillation frequency due to variations caused by changes in valve temperature were minimized, as far as possible, by mounting the tuning condenser and coils under the chassis. This arrangement is also helpful in preventing microphonic effects due to vibration of the tuning condenser plates.

The change in frequency during the warm-up period when the receiver is first switched on, or over a long period of operation, is sufficiently small to allow the receiver to be tuned initially to a required carrier; retuning is unnecessary when the receiver is put into operation on subsequent occasions. The effects of mechanical vibration are small, and the receiver can be bounced fairly severely on the bench without appreciable alteration in oscillator frequency.

A.V.C. is not applied to the converter, as this could cause undesirable oscillator frequency changes because of interaction between the signal and oscillator circuits.

Two other points are of interest in the oscillator circuit. The first is the desirability of keeping the length of the cathode-tap lead as short as possible. Even a lead $\frac{3}{8}''$ long will cause undesirable variations in oscillator grid current. This difficulty was overcome by bending the oscillator coil so that it could be soldered directly on to the valve socket. A short lead may be helpful, initially, in determining the best tapping point on the oscillator coil. Our oscillator coil is a piece of 18 B & S tinned copper wire $2\frac{9}{16}''$ long, bent into an arc of a circle, and soldered directly across the tuning condenser. The tapping point can readily be found by moving the cathode lead along the tinned copper wire until the best point is found to give optimum signal to noise ratio, or the desired oscillator grid current. Also the oscillator coil inductance is readily altered merely by changing the length of wire soldered across the gang. In our case, the tap was located at $\frac{15}{16}''$ from the bottom end of the coil. The other point of interest is the value of by-pass condenser used in the screen circuit. If the value of this condenser is greater than about 500 $\mu\mu$F, large variations in oscillator grid current occur when tuning across the required frequency range. This is possibly due to the effects of inductance, inherent in the condenser and its leads, not allowing adequate by-passing of the screen circuit. In the final circuit the maximum variation in oscillator grid current was from 180 μA to 225 μA.

R-F Stage

The r-f stage requires little comment except that the value of the coupling condenser between the r-f choke and the tuned circuit is 10 $\mu\mu$F, as this offers the most satisfactory compromise between gain variation and stray capacitance loaded across the tuned circuit. A damping resistor was not included as the combined Q's of the r-f choke and r-f coil were of the order of 75 in the circuit. The r-f choke was wound on a $\frac{9}{16}''$ former, and consists of $6\frac{1}{4}$ turns of 22 B & S enamelled wire wound over a length of $\frac{3}{8}''$. The length of the plate lead is $1\frac{1}{2}''$, and the length of lead to the nearest by-pass condenser $1''$. The r-f coil does not use a former, but is $\frac{9}{16}''$ in diameter, and consists of $1\frac{1}{2}$ turns of 18 B & S enamelled wire; the total length, including the leads, which are soldered directly to the gang condenser, is $1\frac{1}{8}''$. Inductance variation is obtained by merely adjusting the distance between adjacent turns.

Aerial Stage

The aerial coil secondary consists of $1\frac{1}{2}$ turns of 18 B & S enamelled wire $\frac{9}{16}''$ diameter, soldered to a junction point with the a.v.c. line and to the fixed plates of the tuning condenser, no former being used. The half turn is opened out slightly for connection to the a.v.c. junction. The overall length of the coil is $1''$. The primary winding is $1\frac{1}{2}$ turns of 18 B & S tinned copper wire, the turns being $\frac{9}{16}''$ in diameter. The leads are each $1\frac{1}{4}''$ long for connection to the aerial terminals. An electrical centre tap on the aerial coil primary is found by connecting a lead from the earth terminal, located next to the aerial terminals, and moving the position of this lead along the primary until equal signal outputs are obtained when the input signal is applied to either aerial terminal. Normally this tapping point will not be at the centre of the winding.

Coupling between the primary and secondary is obtained by arranging the two windings along a common axis, and then adjusting the spacing by bending the primary leads until the receiver gives approximately maximum output. More elaborate methods could not be employed to obtain optimum signal to noise ratios, as no suitable signal generator was available. In any case, the slight improvement obtainable in aerial circuit performance is hardly worth striving for in a receiver of this type.

From this description of the high frequency tuning circuits it can be seen that the inductances do not merely consist of the coil windings but are formed as well by leads and gang inductance. For this reason adjustments will have to be made experimentally to any coils used to obtain the desired results. These adjustments naturally depend on the type of tuning condenser used and the receiver lay-out employed.

A suitable aerial and lead-in were made from 75 Ω cable supplied by A.W.A. Telcon. A piece of the cable was cut into two sections each $28\frac{1}{2}''$ long to form a half wave dipole. The two conductors are soldered together in each half section. These halves were then soldered directly to the lead-in cable. 300 Ω cable would be preferable, for use with a folded dipole antenna, but this cable will not be available in Australia for about two months. Quite good results were obtained using the dipole aerial indoors in the usual picture rail style, but in areas of poor reception outdoor installations offer obvious advantages. The aerial is very directional and should be rotated so as to give maximum signal strength, as almost a complete null point can be obtained when the aerial points in the direction of a transmitter using polarization with the electric vector horizontal.

It may be worth noting that tuning of the aerial stage is not absolutely essential, as an untuned aerial circuit, resonated at the geometric mean frequency in the tuning range, would cause a variation in gain across the band of less than 2:1 using the circuit given for the receiver. This change in gain could be further offset by choosing a suitable point in the tuning range at which to resonate the aerial coil, so as to minimize other gain variations, for example in the r-f and converter stages. The overall receiver sensitivity could still be made about 10 μV at the worst point in the band and the saving in cost between a two and a three gang condenser would be appreciable.

A.V.C. Circuit

The a.v.c. circuit presents some difficulty because there is no delay voltage. If the full a.v.c. voltage is applied to the controlled stages the gain of the receiver drops so rapidly that even for strong signals there is insufficient audio voltage available to give maximum output. For this reason it is necessary to seek a compromise between audio output and a suitable a.v.c. characteristic. In this receiver partial a.v.c. in the ratio $\frac{2}{7}$ was applied. The results can be examined from the a.v.c. characteristics shown in the table of performance figures. For an input variation from 10 μV to 0.1 volt the change in audio voltage input to the amplifier is from 0.2 volt to 5.7 volts. This characteristic could be improved by alteration of the proportion of a.v.c. used, but less than 0.2 volt (r.m.s.) to drive the audio amplifier was thought to be undesirable, since 0.26 volt is required to give 0.5 watt output. A more sensitive audio amplifier would allow improved a.v.c. characteristics, but this introduces the undesirable condition of smaller voltages being applied to the diodes.

A method which was tried out to overcome the effect of no delay on the a.v.c. system may be of interest. Instead of using a type 6BA6 as a driver for the ratio detector a sharp cut-off valve was substituted, and arranged to act as a partial limiter for signals exceeding about 100 microvolts. The a.v.c. was then picked up from the limiter circuit instead of the ratio detector, and applied to the r-f and i-f valves. There are several advantages with this arrangement.

Additional limiting is obtained on strong signals and a long delay period for the a.v.c. circuit is readily obtained, and peak clipping of strong bursts of noise occurs when receiving weak signals; furthermore the a.v.c. time constant can be made very rapid. This overcomes what appears to be two of the inherent disadvantages of the ratio detector in which time constants for the a.v.c. system of less than about 0.1 to 0.2 seconds are not readily obtainable, concurrently with correct performance of the detector, and a delay voltage cannot be easily applied to the a.v.c. circuit. It is thought that this circuit would warrant further investigation and a type 6AC7 would appear to be very suitable for the driver stage. A type 6AU6 was tried in the circuit with satisfactory results. Sufficient a.v.c. action should be obtainable as the high g_m amplifier valves allow the gain of the controlled stages to be altered very rapidly with change in a.v.c. bias, and the limiting action of the driver stage further assists in maintaining constant audio output. Further, the receiver need not have a sensitivity of more than about 10 microvolts, as the ratio detector operates satisfactorily with about 10 millivolts applied to the grid of the driver stage, and there is no threshold effect as with conventional limiters. The only apparent disadvantage is that cathode bias would be required for the r-f, i-f and driver stages, but the additional cost of bias resistors and by-pass capacitors would probably be more than justified by the improved performance. The bias in the present arrangement is due to a combination of grid leak and a.v.c. voltages. The actual bias is 0.9 volts with the aerial terminals shorted, 1 volt with the terminals open, and 1.15 volts for a signal input of 50μV. These voltages were measured directly on the grids with a d.c. vacuum tube voltmeter.

Receiver Alignment

The complete alignment of the receiver can be carried out with ordinary a-m generators covering the required frequency ranges. Sensitivity measurements can only be made directly if an f-m generator is available. In our case an f-m i-f generator was available, and so overall sensitivity measurements could be made indirectly. Since f-m signal generators are not generally available the alignment using an a-m generator will be described.

If a sensitive microammeter is available, say 0-25 μA, this is shunted with a 3000 Ω variable resistor (or some other suitable value) to allow variations in sensitivity, and placed in series with the earth end of the 50,000 Ω resistor in the diode load circuit of the ratio detector. A 0-1 mA meter, unshunted, can be used, the only advantage being that small variations in trimmer settings are more difficult to observe. A standard output meter, or suitable a.c. voltmeter, is connected to the receiver output in the usual manner.

I-F Amplifier

Assuming the audio amplifier is operating satisfactorily a signal at 10.7 Mc/s is applied through a blocking condenser to the signal grid of the 6BA6 driver stage. The i-f signal is amplitude modulated (say 50% or any convenient value). First adjust the trimmer condenser on the primary of the discriminator transformer to give maximum current through the diode load resistance, next adjust the secondary trimmer until two peaks of audio output are found. The secondary trimmer is then set so that the output falls to its minimum value between these two peaks. This completes the initial adjustments.

Next, tune the signal generator above and below the centre frequency and note the peak amplitudes of the audio output that occur at about 10.55 Mc/s and 10.85 Mc/s. (Note that there should be practically zero audio output at 10.7 Mc/s.) The magnitudes of the peaks should be approximately equal and symmetrically spaced from the centre frequency. If the high frequency peak is too large and further out than the low frequency peak, set the generator at the desired high frequency peak and alter the setting of the "slug" to give a definite increase in output at this new frequency. The generator must now be returned to 10.7 Mc/s and the secondary trimmer re-adjusted to give minimum audio output. Recheck the peaks again; if they are approximately symmetrically spaced, but with the high frequency peak larger than the low frequency peak, tune the generator to the low frequency peak and retune the primary trimmer to peak at this frequency. This should only require very slight resetting of the trimmer. Also, if the low frequency peak does not fall where desired the same procedure is adopted. The whole process may seem rather tedious, but when carried out a few times the desired results are obtained very quickly. Once a setting is fixed for the "slug" there is no difficulty, since the secondary trimmer is always adjusted for zero output at the centre frequency and the primary trimmer adjusted on the low frequency peak to give approximately equal high and low frequency output peaks.

The resetting of the primary trimmer should have only a very small effect on the maximum value of diode circuit current. Larger diode currents can be obtained by detuning the secondary trimmer but this is not the desired condition. The secondary trimmer may be considered as balancing a bridge circuit to give zero output. Smaller values of trimming condensers than those shown on the circuit diagram are helpful, if they are obtainable, and the differences in capacitance can be incorporated in the fixed condensers (3-20 $\mu\mu$F would be suitable).

Having completed the alignment of the discriminator transformer, apply a signal at 10.7 Mc/s to the grid of the 6BA6 i-f amplifier. The signal should not be too large, but sufficient to give a good indication on the current meter, as the a.v.c. circuit is operating and is not disconnected because of the bias arrangements. The "slugs" in the i-f transformer are now peaked to give maximum current through the diode resistor. Next the i-f signal is applied to the grid of the converter, but it is preferable to remove the grid lead to the tuned circuit as this will short circuit the generator output. A blocking condenser is not used in the generator lead for this test so as to provide a d.c. path to ground from the converter signal grid. The "slugs" are now peaked in the first i-f transformer to again give maximum diode current. The complete alignment of the two i-f transformers can be rechecked if desired, as is usually done in an ordinary a-m receiver. The zero output setting of the secondary trimmer of the discriminator transformer can also be checked if desired. Only a small audio output signal will be heard during alignment. The smaller this residual audio output the better the balance of the ratio detector circuit.

As an alternative method of alignment, using only an output meter and an a-m signal generator, the following could be adopted. Align the discriminator circuit as described, setting the secondary trimmer for zero output and the primary trimmer for maximum (approximately) audio output at the low frequency output peak. Set the generator to 10.7 Mc/s and then detune the secondary trimmer to give maximum audio output, but do not alter the primary trimmer setting. The complete alignment of the i-f amplifier is then exactly as for an a-m receiver. On completion of the alignment of the i-f circuits the discriminator secondary trimmer is reset to give minimum a-m output at 10.7 Mc/s. The centre point is between the two audio output peaks as previously described.

From this it is seen that with receivers using ratio detectors the alignment is not much more difficult than for an a-m receiver, particularly if the i-f transformers are not overcoupled. In any case staggered tuning would probably offer less production difficulties than overcoupling of the i-f's, and the difference in gain need not be large enough to be important.

R-F Circuits

An amplitude modulated signal generator tunable over a range of about 80-130 Mc/s is required to align these circuits. The procedure is almost exactly the same as for an a-m receiver. It will be found on tuning the generator that two audio output peaks occur just as with the i-f alignment. However, it is convenient to adjust the receiver to one of these peaks and align the r-f stages for maximum output in the usual manner. Oscillator coverage is set as in an a-m receiver and image points should be checked.

The error in frequency setting using this procedure is only about 150 Kc/s if, say, the low-frequency peak is chosen for each new carrier frequency, and is unimportant when the tuning range at each end of the band is about 500 Kc/s more than required. The exact coverage can be determined, after the initial alignment is complete, by setting the generator carrier to give minimum output between the two audio peaks. Again, signal levels should not be too high during alignment because of a.v.c. action.

The output meter indications are easier to read than the current meter changes when adjusting the r-f and i-f circuits, and will usually lead to a more accurate alignment.

Sensitivity Measurements

If a f-m generator is available which operates at i-f, then the complete i-f sensitivity can be checked. To check overall sensitivity the diode current is noted when making the i-f sensitivity reading required to give 50 milliwatts output. This current reading can then be repeated for the carrier input on the signal circuit range.

In making overall sensitivity checks the leakage from the only two generators available covering the 88-108 Mc/s band was so bad that measurements of sensitivities under about 10 μV was virtually impossible. To find the overall sensitivity the best that could be done was to measure the aerial coil gain at a fairly high level and then to divide the sensitivity at the grid of the r-f amplifier by the gain obtained. This was fairly satisfactory as the sensitivity of the receiver from the r-f valve signal grid is approximately 10 μV. The limit to the sensitivity was not due to receiver input noise but mainly to lack of audio gain. It was considered, however, that a sensitivity of approximately 6 μV for 50 milliwatts output was adequate for all ordinary purposes.

Image Rejection

The measured image rejection is higher than the calculated value, and the change is mainly due to the somewhat higher resultant Q of the r-f stage tuned circuit. The rejection is 46 db compared to the calculated figure of 44 db.

General

Actual field tests with the receiver indicated several interesting points. Firstly, tuning was found to be little or no more difficult than for an ordinary broadcast receiver. The tuning action is somewhat different, however, as three response points can be found by careful dial setting. This does not lead to any ambiguity, as the correct centre frequency gives very much greater output than the side responses, which are so distorted as to make reception completely unsatisfactory. Correct tuning is indicated by lack of distortion and noise, and this is sufficiently sharply defined as not to lead to any serious error. A tuning indicator did not appear to be any more necessary than with ordinary receivers.

With the receiver located in a very bad area as regards electrical noise, reception was still satisfactory using an indoor aerial. Tram noises offered no difficulty, and it was only a very occasional car ignition system that gave any severe interference. It was noticed, however, that the output from the receiver is somewhat distorted when strong interference is present, and the input to the receiver by the desired carrier is about 100 μV or less. Because of the low deviation frequencies being used in the present broadcasts it is thought that even these results will be improved upon. Most of the noise heard was high pitched, and a tone control giving treble cut would be advantageous, as the decrease in high frequency response was less noticeable than the decrease in background noise. Positioning of the speaker was important in this regard, and if placed near the floor, as is usual in ordinary receivers, both the desired high frequencies and the noise present were almost inaudible. This indicates that if good high frequency response is required positioning of the loudspeaker is an important factor, and is a requirement not fulfilled by ordinary radio cabinets.

A further point of interest was the lack of noise when detuned from a carrier. The general noise level appeared far lower than with a-m broadcast receivers having similar sensitivity, and was very much less than with f-m receivers using ordinary limiter-discriminator combinations having lower overall sensitivities. This is probably due to the ratio detector, as this circuit provides appreciable amplitude rejection with low signal inputs.

Good oscillator stability is of great assistance in any f-m receiver using a balanced discriminator, whether of the ratio detector type or when preceded by limiters. Very appreciable improvement in amplitude rejection is secured when the i-f is correctly centred at the null point of the discriminator circuit, and the less the variation from this centre point the less will be the interference from amplitude variations. Further, a very well-balanced discriminator circuit greatly improves amplitude rejection quite independently of the other amplitude rejection circuits used in the receiver. Good discriminator and i-f tuned circuit stability offers obvious advantages.

As a final point for those designers who feel that an r-f stage is not essential in an f-m receiver, not only is the image rejection so low as to be unsatisfactory, but cases have been reported in U.S.A. where re-radiation at the oscillator frequency of the receiver has been received at distances up to four miles from the receiver location. This undesirable interference could be a serious problem when using receivers in densely populated areas. The explanation of the effect is simple. There is appreciable voltage at the oscillator frequency developed across the converter input circuit and this, combined with an efficient aerial system, allows undesirable radiation to occur.

TEST RESULTS

(1) VOLTAGE MEASUREMENTS.

All measurements are for zero signal input and with AVO model 7 meter.

Mains 240V r.m.s. 0.36A r.m.s. 50 c/s
Transformer 346V r.m.s./346V r.m.s.
Rectifier Output 360V d.c. 103 mA
Filter Output 262.5 d.c. 103 mA
Heaters 6.3V r.m.s. 2.25A r.m.s. (nominal).

(2) OSCILLATOR.

Signal Frequency	I_{c1}
88 Mc/s	205 μA
98 Mc/s	225 μA
108 Mc/s	180 μA

(3) OVERALL PERFORMANCE.

Output 50 milliwatts.
Frequency deviation ± 75 Kc/s.
Modulating frequency 400 c/s.

Input to	Frequency	Input	Ratio
6SJ7-G control grid	400 c/s	.082 V	—
6BA6 control grid (driver)	10.7 Mc/s	18 mV	4.56
6BA6 control grid (i-f)	10.7 Mc/s	474 μV	38
6BE6 control grid (gang closed)	10.7 Mc/s	184 μV	2.57
6BE6 control grid (osc. shorted)	10.7 Mc/s	92 μV	5.14
6BE6 control grid	88 Mc/s	101 μV	4.7
6BE6 control grid	98 Mc/s	96.5 μV	4.9
6BE6 control grid	108 Mc/s	145.5 μV	3.25
6BA6 control grid	88 Mc/s	9.7 μV	10.4
6BA6 control grid	98 Mc/s	8.8 μV	11.0
6BA6 control grid	108 Mc/s	10 μV	14.6
Aerial } Indirect meas-	88 Mc/s	6 μV	1.6
Aerial } urements only	98 Mc/s	7 μV	1.54
Aerial } (75 Ω input)	108 Mc/s	5 μV	2.0

(4) SELECTIVITY

Input	Frequency	Input	Ratio
Aerial	108 Mc/s	50 μV	—
Aerial	129.4 Mc/s	10.5 mV	210

(5) A.V.C. CHARACTERISTIC

(All measurements made indirectly)

Input	Grid Bias Volts	Audio Input Volts (r.m.s.)
10 μV	1.0 V	0.2 V
50 μV	1.15 V	0.38 V
100 μV	1.8 V	1.2 V
150 μV	2.0 V	1.3 V
200 μV	2.2 V	1.59 V
300 μV	2.4 V	1.62 V
400 μV	2.65 V	1.81 V
500 μV	2.9 V	2.08 V
600 μV	3.0 V	2.10 V
1 mV	3.8 V	2.71 V
2 mV	4.1 V	3.0 V
5 mV	4.75 V	3.4 V
10 mV	6.5 V	4.4 V
50 mV	8.0 V	5.0 V
0.1 V	9.2 V	5.7 V

(6) DISCRIMINATOR CHARACTERISTIC (RATIO DETECTOR)

Static measurements 70 millivolts input to driver stage

Ratio	Frequency	Output Change Decibels
7.58	10.3 Mc/s	17.6
8.74	10.4 Mc/s	18.8
13.15	10.5 Mc/s	22.4
13.9	10.54 Mc/s	22.9
12.2	10.6 Mc/s	21.7
0	10.7 Mc/s	0
12.0	10.8 Mc/s	21.6
14.1	10.86 Mc/s	22.9
13.58	10.9 Mc/s	22.6
9.9	11.0 Mc/s	19.9
6.84	11.1 Mc/s	16.7

(7) AUDIO FREQUENCY RESPONSE

Power Output 0.5 watt.
Input constant at 0.26 volt r.m.s.
Gain reduction due to feedback 7.36 times.
Reference frequency 400 c/s.

(A) Tone Control in normal position.

Frequency c/s	Output change db
50	+ 0.5
100	+ 0.25
400	0
1,000	0
3,000	0
5,000	0
10,000	— 0.5
13,000	— 0.5
15,000	— 0.75

(B) Tone Control bass boost position.

Frequency c/s	Output change db
50	+ 5.9
60	+ 6.0
100	+ 4.5
200	+ 1.75
300	+ 0.5
400	0

Continued on page 95.

R.C.A. Application Note AN-119

USE OF THE 2E24 AND 2E26 AT 162 MEGACYCLES

RCA-2E24 and RCA-2E26 are beam power transmitting tubes for use at full input up to 125 megacycles and reduced input at considerably higher frequencies. This Note gives circuit and performance data on the use of these tubes as rf power amplifiers and frequency multipliers at 162 megacycles, the upper frequency of the FM band (152-162 Mc) designated for railroad, police, and other telephone communications.

The 2E24, because it has a quick-heating, low-drain filament and low plate voltage requirements is particularly suited for portable operation. Its filament requires 4 watts of power at 6.3 volts. The filament temperature 1.8 seconds after the application of filament voltage is 80% of normal. The filament is designed for intermittent operation and, therefore, should not be used under standby conditions because short tube life would result.

In a 162-megacycle amplifier, a single 2E24 at a plate voltage of 350 volts can deliver 13.5 watts of useful power. As a 162-megacycle doubler, a 2E24 can deliver 6 watts of useful power; and as a tripler, 3 watts. The 2E24 can be operated under ICAS ratings at an ambient temperature of 60° Centigrade provided the maximum bulb temperature does not exceed 210° Centigrade.

The 2E26, because it has an indirectly heated cathode, is especially useful for standby operation. Operating within ICAS ratings as a single-ended, 162-megacycle amplifier, it can deliver a useful power output of 13.5 watts. When operated within CCS ratings, it can deliver up to 9.5 watts of useful power output.

The maximum plate-to-grid capacitance of the 2E24 is 0.11 $\mu\mu$f; that of the 2E26 is 0.2 $\mu\mu$f. The plate-to-grid capacitance together with the inductance due to the screen lead may cause a 2E24 or 2E26 amplifier operating in the 152-to-162-megacycle band to oscillate. Some precautions, therefore, must be taken to prevent such oscillations. For this purpose, screen (grid-No. 2) tuning has been used with good results in single-ended and push-pull amplifiers employing either 2E24's or 2E26's. Such amplifiers operate over the entire 152-to-162-megacycle band without readjustment of the screen tuning and are illustrated in Figs. 1 and 2. It has been found that the value of the screen-tuning capacitor necessary to make the amplifier stable is to a large part dependent upon the physical layout of the amplifier as well as upon the type of capacitor used. Consequently, the value of the screen-tuning capacitor needed to make the amplifier stable is best found by experiment. Its value is usually within the range of 25 to 100 $\mu\mu$f. The test for oscillation should be made at reduced plate and screen voltages and without fixed bias. The capacitance value should be chosen so that the amplifier without grid drive and with unloaded plate circuit does not oscillate with any combination of plate and grid circuit tuning. A particular value of capacitance can also be determined to decrease the feedback to a minimum. The proper value is indicated by minimum reaction of plate-circuit tuning on the dc grid current when plate and screen voltages are zero.

An rf choke, identified as L in Figs. 1 and 2, should be placed in the dc screen-voltage lead at the tube-socket terminal. A choke of 20 turns wound for a length of 1 inch on a form $\frac{1}{4}$ inch in diameter should suffice.

The rf grounding of the filament and filament mid-tap or the cathode socket terminals is also important because of its effect upon the stability of the amplifier. The most satisfactory method of grounding these terminals is to ground directly or to bypass to ground by the shortest possible path as shown in Fig. 3. Uncased mica bypass capacitors of 100 $\mu\mu$f are suitable for this purpose.

The details of the plate tank circuits used with the single-ended and push-pull amplifiers are shown in Figs. 4 and 5. Since the output capacitances of the 2E24 and 2E26 are the same, the same plate tank circuits can be used with either tube type. The input capacitance of the 2E26, however, is almost twice that of the 2E24 and, therefore, the same grid circuits cannot be used. Circuit elements with lumped constants of suitable values are used in the grid circuits of both tubes. Power output and tube driving-power data for the 2E24 and 2E26 operating as single-ended amplifiers at 162 megacycles are shown in the curves of Figs. 6, 7, and 8. The power output was measured into a lamp load and is the tube output minus the tank-circuit losses. The tube driving power does not include any losses in the grid circuit external to the tube except the power absorbed by the grid-bias resistor.

The 2E24 performance curves of Fig. 5 show conditions necessary for operation of the tube under ICAS ratings of 13.5-watts maximum plate dissipation and 85 milliamperes maximum dc plate current. The values of useful power output and tube driving power are for a single-ended, 162-megacycle amplifier. The plate tank-circuit loss was found to be 3 watts, so that with an input of 30 watts the plate dissipation rating of 13.5 watts is exceeded when the useful power output drops below 13.5 watts. As can be seen from the curves of Fig. 6, a tube driving power of at least 2 watts is necessary to obtain power out-

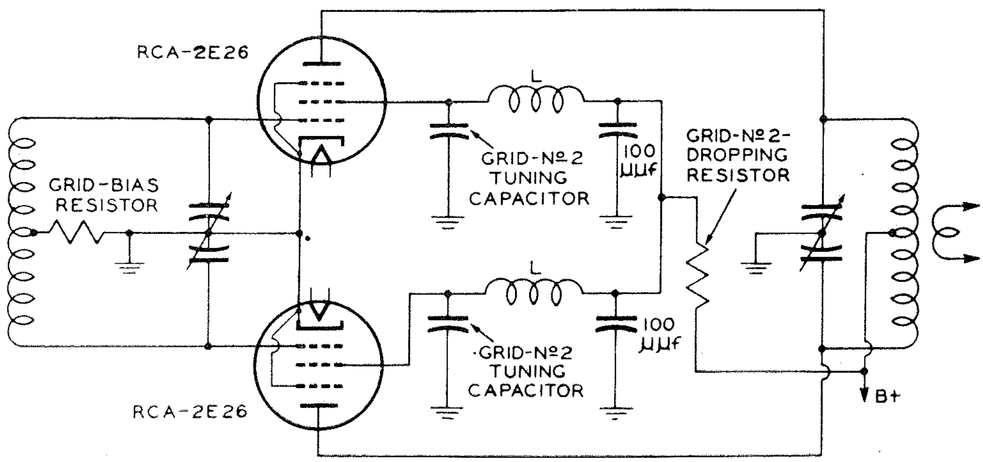


Fig. 1 - Schematic Diagram of 2E26 Push-Pull Amplifier Circuit (Identical Circuit Except for Filament and Cathode Connections Used with 2E24).

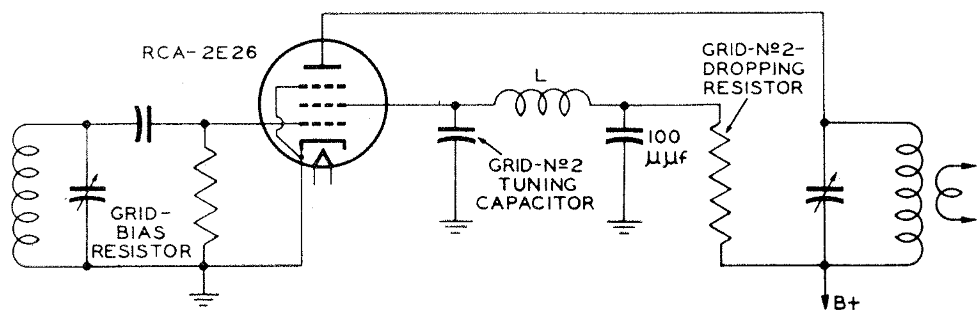
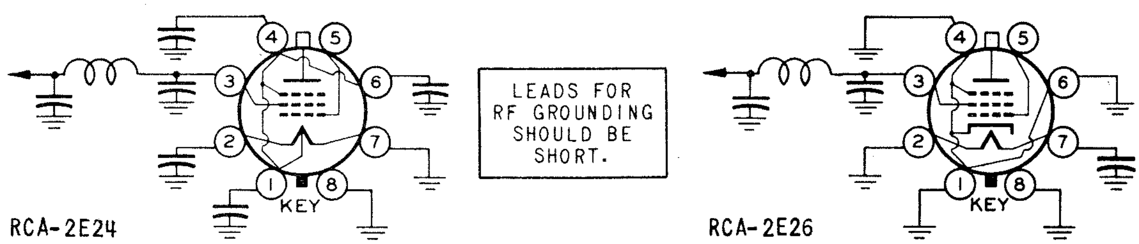


Fig. 2 - Schematic Diagram of 2E26 Single-Ended Amplifier Circuit (Identical Circuit Except for Filament and Cathode Connections Used with 2E24).



RCA-2E24	
PIN 1:	FILAMENT MID-TAP, GRID No. 3, INTERNAL SHIELD
PIN 2:	FILAMENT
PIN 3:	GRID No. 2
PIN 4:	SAME AS PIN No. 1
PIN 5:	GRID No. 1
PIN 6:	SAME AS PIN No. 1
PIN 7:	FILAMENT
PIN 8:	BASE SLEEVE
CAP:	PLATE

RCA-2E26	
PIN 1:	CATHODE, GRID No. 3, INTERNAL SHIELD
PIN 2:	HEATER
PIN 3:	GRID No. 2
PIN 4:	SAME AS PIN No. 1
PIN 5:	GRID No. 1
PIN 6:	SAME AS PIN No. 1
PIN 7:	HEATER
PIN 8:	BASE SLEEVE
CAP:	PLATE

Fig. 3 - Socket Connections for RF Grounding.

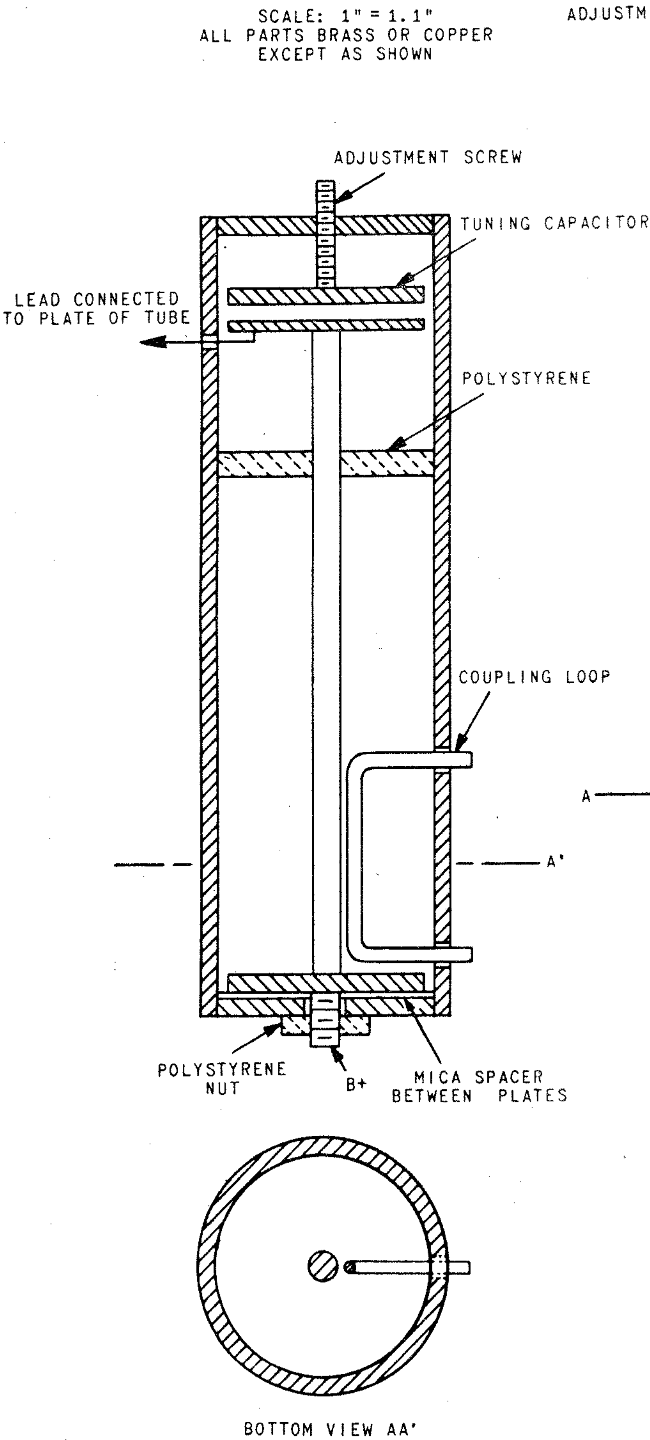


Fig. 4 - Single-Ended 162-Megacycle Tank Circuit

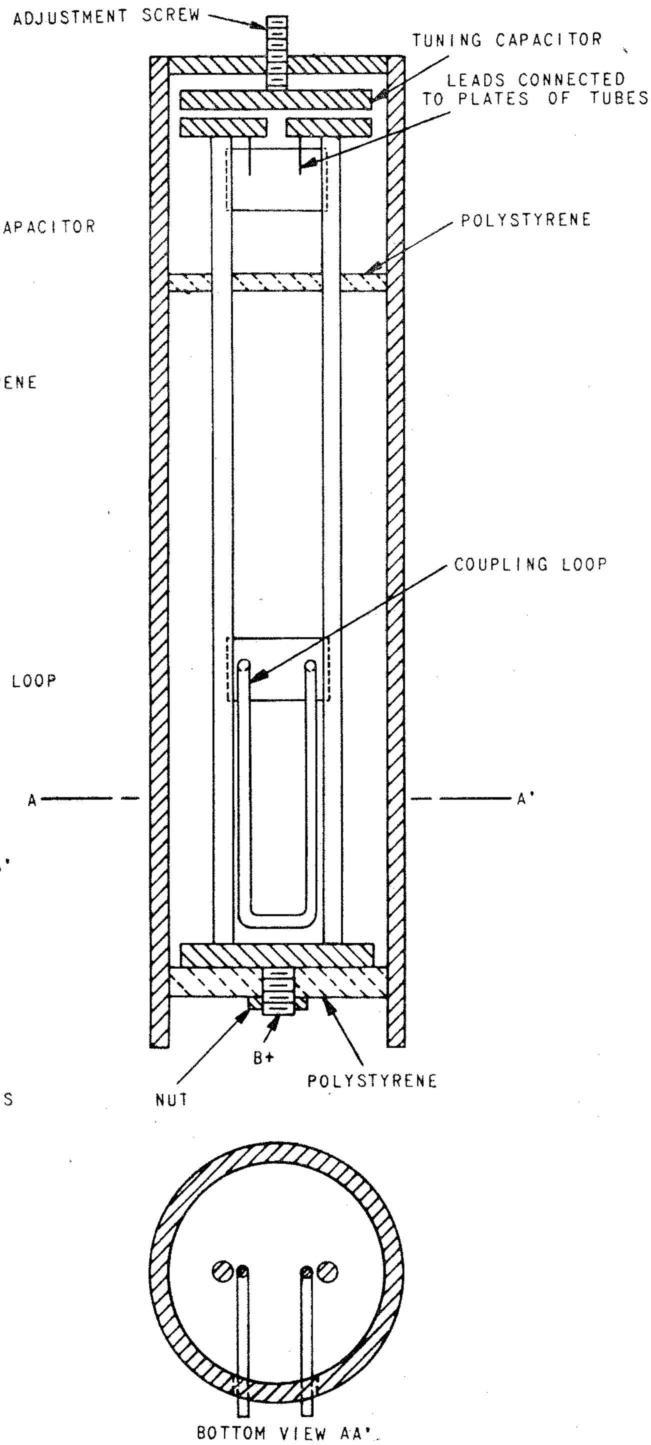


Fig. 5 - Push-Pull 162-Megacycle Tank Circuit

put of 13.5 watts. This driving power, which is much larger than that required at low frequencies, is needed because of I^2R losses in the grid-cathode structure of the tube and transit-time losses between the grid and cathode. A push-pull 2E24 amplifier will deliver approximately twice the power of a single-ended amplifier.

The performance curves for the 2E26 in Figs. 7 and 8 show conditions necessary for operation of the tube under ICAS and CCS ratings, respectively. The maximum plate dissipation of the tube under ICAS ratings is 13.5 watts, and under CCS ratings is 10 watts.

The 2E24 operating as a doubler at 162 megacycles will give a useful power output of 6 watts under the operating conditions shown below.

DC Plate Voltage	300	Volts
DC Plate Current	66	Milliamperes
DC Grid-No. 1 (Control-Grid) Current	3.5	Milliamperes
DC Grid-No. 1 Voltage*	–175	Volts
DC Grid-No. 2 (Screen) Voltage#	170	Volts
DC Power Input	19.5	Watts
Useful Power Output	6	Watts

* Obtained with 30,000-ohm grid resistor and 70 volts of fixed bias.

Obtained from the plate supply through a 15,000-ohm series resistor.

The 2E24 operating as a tripler at 162 megacycles will give a useful power output of 3 watts under the operating conditions shown below.

DC Plate Voltage	200	Volts
DC Plate Current	80	Milliamperes
DC Grid-No. 1 Current	3.5	Milliamperes
DC Grid-No. 1 Voltage*	–175	Volts
DC Grid-No. 2 Voltage	200	Volts
DC Power Input	16	Watts
Useful Power Output	3	Watts

* Obtained with 30,000-ohm grid resistor and 70 volts of fixed bias.

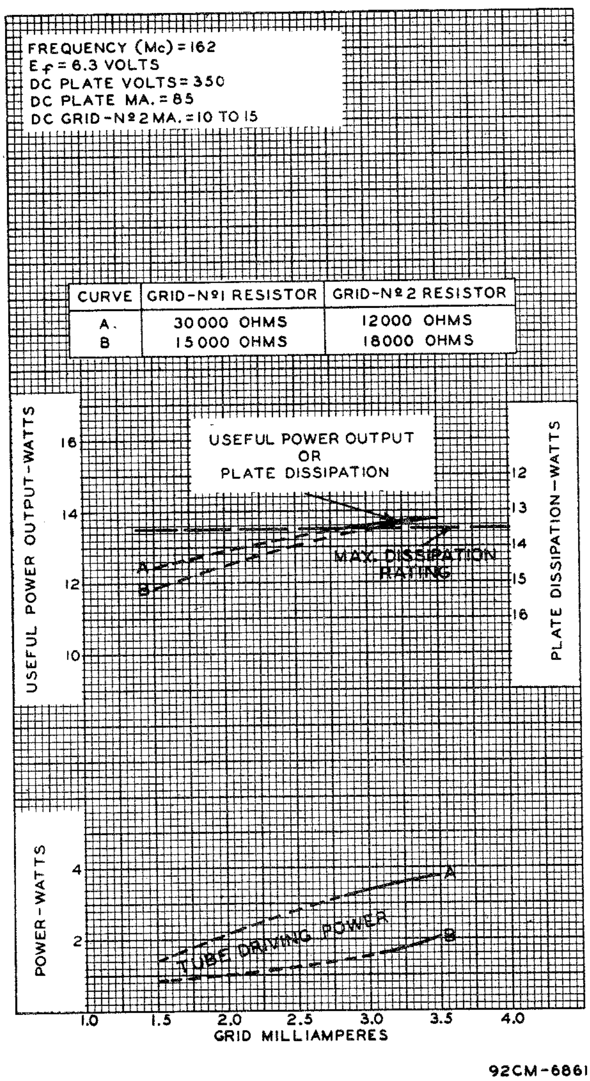


Fig. 6 - ICAS Operation Characteristics for Single-Ended 2E24

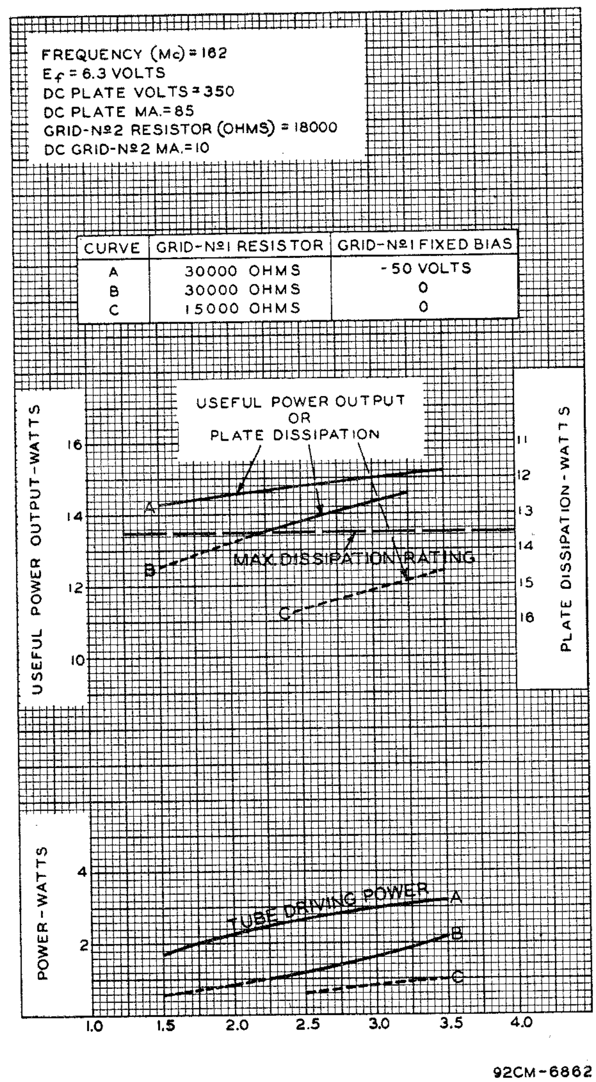


Fig. 7 - ICAS Operation Characteristics for Single-Ended 2E26

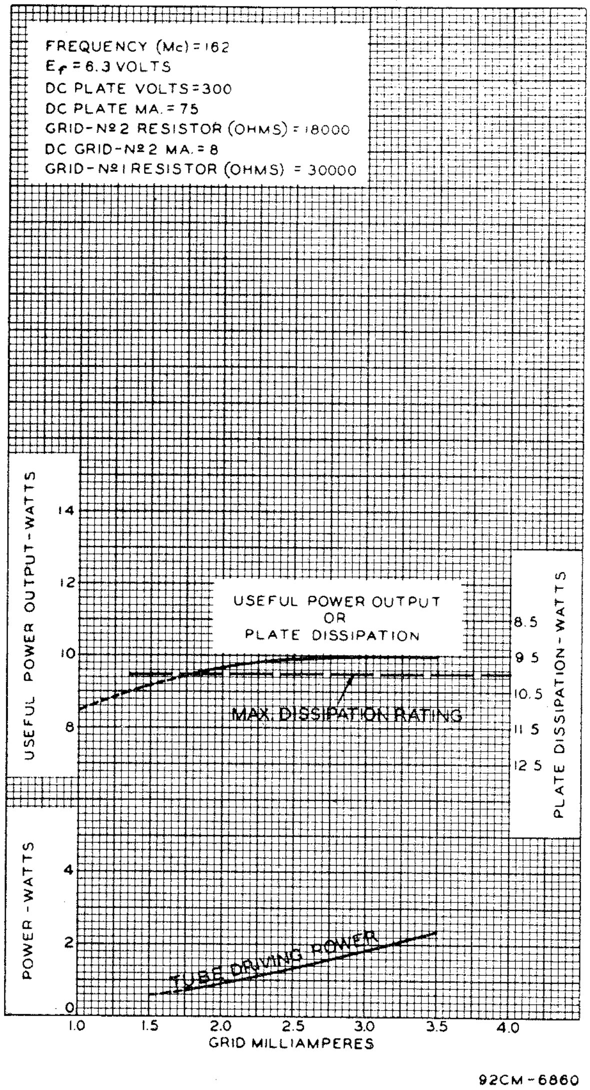


Fig. 8 - CCS Operation Characteristics for Single-Ended 2E26

The maximum ratings of the 2E24 for ICAS class C telegraphy operation of the tube at 162 megacycles are given below.

DC Plate Voltage	450 max.	Volts
DC Grid-No. 2 Voltage .	200 max.	Volts
DC Grid-No. 1 Voltage .	–175 max.	Volts
DC Plate Current	85 max.	Milliamperes
DC Grid-No. 1 Current .	3.5 max.	Milliamperes
Plate Input	30 max.	Watts
Grid-No. 2 Input	2.5 max.	Watts
Plate Dissipation	13.5 max.	Watts

The maximum ratings of the 2E26 for CCS and ICAS class C telegraphy operation of the tube at 162 *megacycles are given as follows:*

	CCS	
DC Plate Voltage	375 max.	Volts
DC Grid-No. 2 Voltage .	200 max.	Volts
DC Grid-No. 1 Voltage .	–175 max.	Volts
DC Plate Current	75 max.	Milliamperes
DC Grid-No. 1 Current .	3.5 max.	Milliamperes
Plate Input	22.5 max.	Watts
Grid-No. 2 Input	2.5 max.	Watts
Plate Dissipation	10 max.	Watts
	ICAS	
DC Plate Voltage	450 max.	Volts
DC Grid-No. 2 Voltage .	200 max.	Volts
DC Grid-No. 1 Voltage .	–175 max.	Volts
DC Plate Current	85 max.	Milliamperes
DC Grid-No. 1 Current .	3.5 max.	Milliamperes
Plate Input	30 max.	Watts
Grid-No. 2 Input	2.5 max.	Watts
Plate Dissipation	13.5 max.	Watts

Vented Baffle Loudspeaker

Reference was made to the vented baffle loudspeaker in the article "The Design of a High Fidelity Amplifier" in Radiotronics 124, and we have received numerous enquiries regarding the detailed construction of a suitable vented baffle. The design given herewith is suitable for most normal 12" loudspeakers having cone resonances between 55 and 65 c/s. No provision has been made for accurate adjustment of the size of the cabinet to suit the cone resonance of the loudspeaker, since the dimensions are fortunately not at all critical. While the impedance characteristic may not be perfectly symmetrical, the general improvements in the performance over the whole bass frequency range will be almost equal to those obtainable with ideal adjustments.

A detailed specification is given below, but may be varied to suit particular conditions.

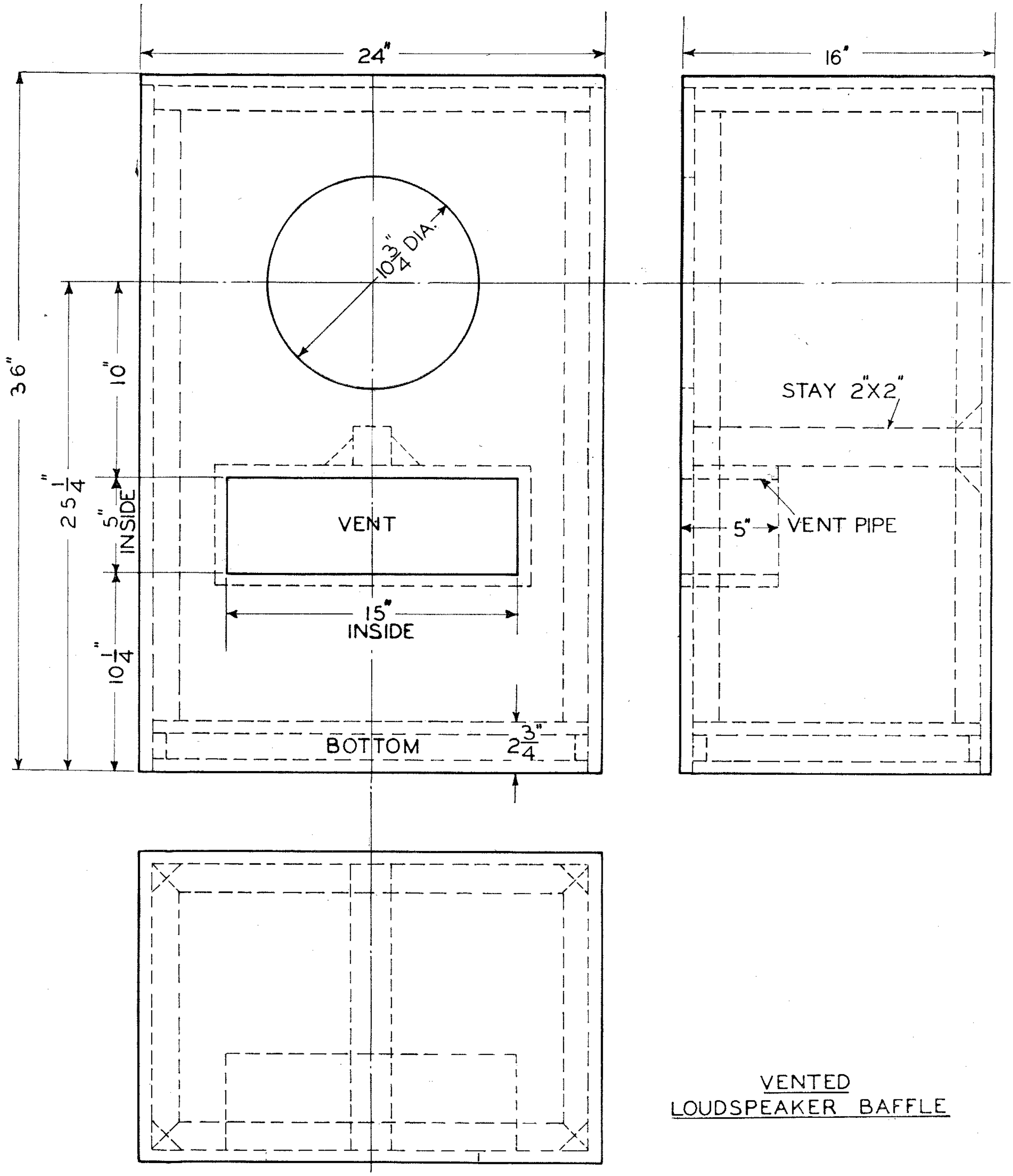
Specification

Timber 5/8" plywood. All joints to be glued, except that the bottom is left free for access to the interior.

Triangular fillets to be screwed and glued to both sides. Fore and aft stay (2" x 2") to be firmly screwed and glued at both ends to withstand internal pressure. Bottom to be glued and screwed to 2" x 1" frame, the latter to be screwed only to main portion of cabinet, using 6 screws on each long side and 4 screws each short side.

Whole of inside cabinet to be covered with cowhair underfelt or thick felt, including inside surface of bottom but excluding the 2" x 2" stay, and leaving a clear ring 2" wide around hole ($10\frac{3}{4}$" dia.) for mounting loudspeaker.

Exterior finish as desired.



VENTED LOUDSPEAKER BAFFLE

(Continued from page 89)

Test Results

(8) HUM OUTPUT.

(A) Tone Control normal position.

(B) Tone Control bass boost position.

	Hum Frequency	Voltage across output transformer primary
A	100 c/s	3.6 millivolts
	200 c/s	2.4 millivolts
B	100 c./s	8 millivolts
	200 c/s	3.6 millivolts

RADIOTRON CIRCUIT No. S115

Below is shown a circuit for a 5″ oscillograph using a type 5BP1. The following notes relate to details not shown on the diagram:—

(1) C_1 approximately 190 $\mu\mu$F: adjust for 2,000 volts output. The frequency of operation for the high voltage power supply is approximately 1 Mc/s. For full constructional details of this supply refer to "Radio Frequency High Voltage Sources" on page 10 of Radiotronics 117.

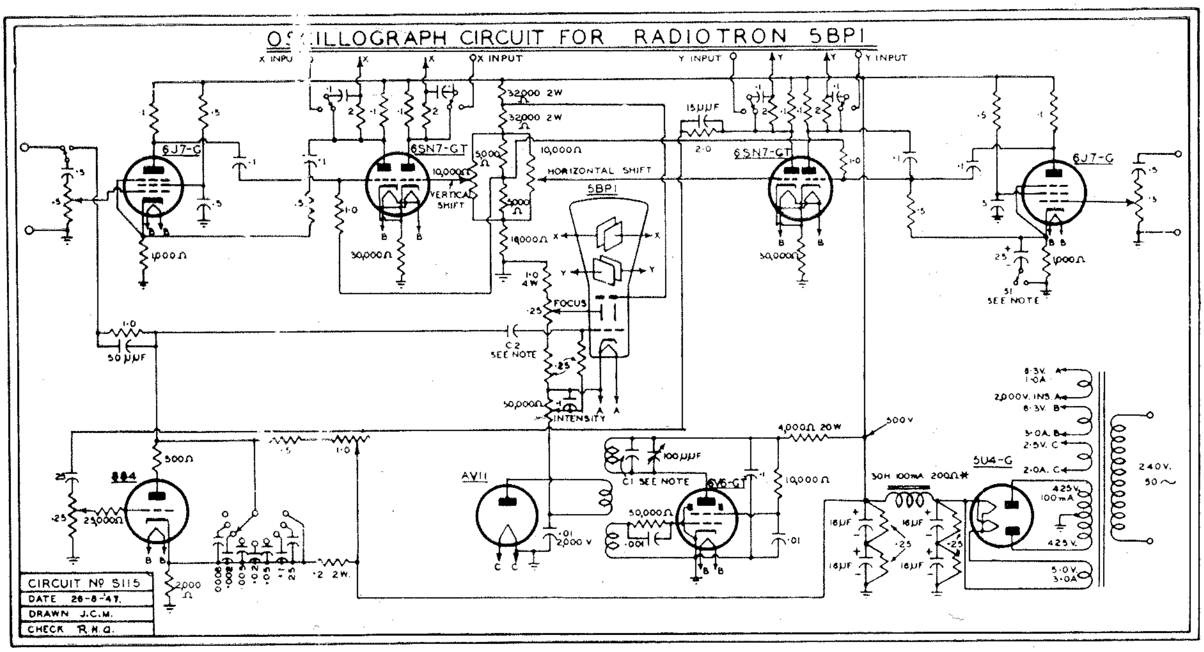
(2) C_2 made up from three 250 $\mu\mu$F 500 V mica capacitors connected in series.

(3) The switch marked S1 may be used to increase the gain of the voltage amplifier by approximately four times.

(4) The range of time base frequencies is from approximately 20 c/s to 20 Kc/s.

(5) The values of resistors are given in megohms (1 watt rating) and capacitors in microfarads, except where otherwise indicated.

(6) *A nominal d.c. resistance of 200 ohms is stated for the power choke. A series resistor may be used to adjust the H.T. to 500 volts.



NEW R.C.A. RELEASES

Radiotron type 1U5 is a multi-unit valve containing an r-f diode for use as detector, and a pentode for use as an a-f voltage amplifier, in portable receivers. It is similar to the type 1S5 but has a different basing arrangement, and utilizes an improved structure which greatly reduces any tendency toward microphonic effects. In addition the diode unit is effectively shielded from the pentode unit to prevent "play-through" (residual volume effect) when the volume control of the receiver is set for minimum gain.

Radiotron type 3E29 is a twin unit beam power amplifier similar to the type 829-B but intended particularly for pulse modulator service. Full details are available on the current HB3 data sheets.

Radiotron type 6BJ6 is a remote cut-off amplifier pentode particularly useful in mobile equipment where heater-current drain is an important consideration and in AC/DC FM and AM receivers. It features a 6.3V 150 mA heater, high transconductance, and low grid-plate capacitance.

Radiotron type 6K5-G was originally in the RCA line, but was withdrawn during the war. It is now planned to again place this valve in the active RCA list for replacement purposes.

Radiotron type 10Y is a transmitting power amplifier triode, having a maximum plate dissipation rating of 10 watts. Full details are available on the current HB3 data sheets.

Radiotron type 12AL5 is a high perveance twin diode like the type 6AL5, but has a 12.6V 0.150A heater, and is intended particularly for use as a ratio detector in ac/dc f-m receivers. In circuits utilizing wide-band amplifiers the low internal resistance of the type 12AL5 makes it possible to obtain increased signal voltage from a low-resistance diode load.

Radiotron type 8014-A is a forced air-cooled transmitting triode intended especially for pulsed operation. Full details are available on the current HB3 data sheets.