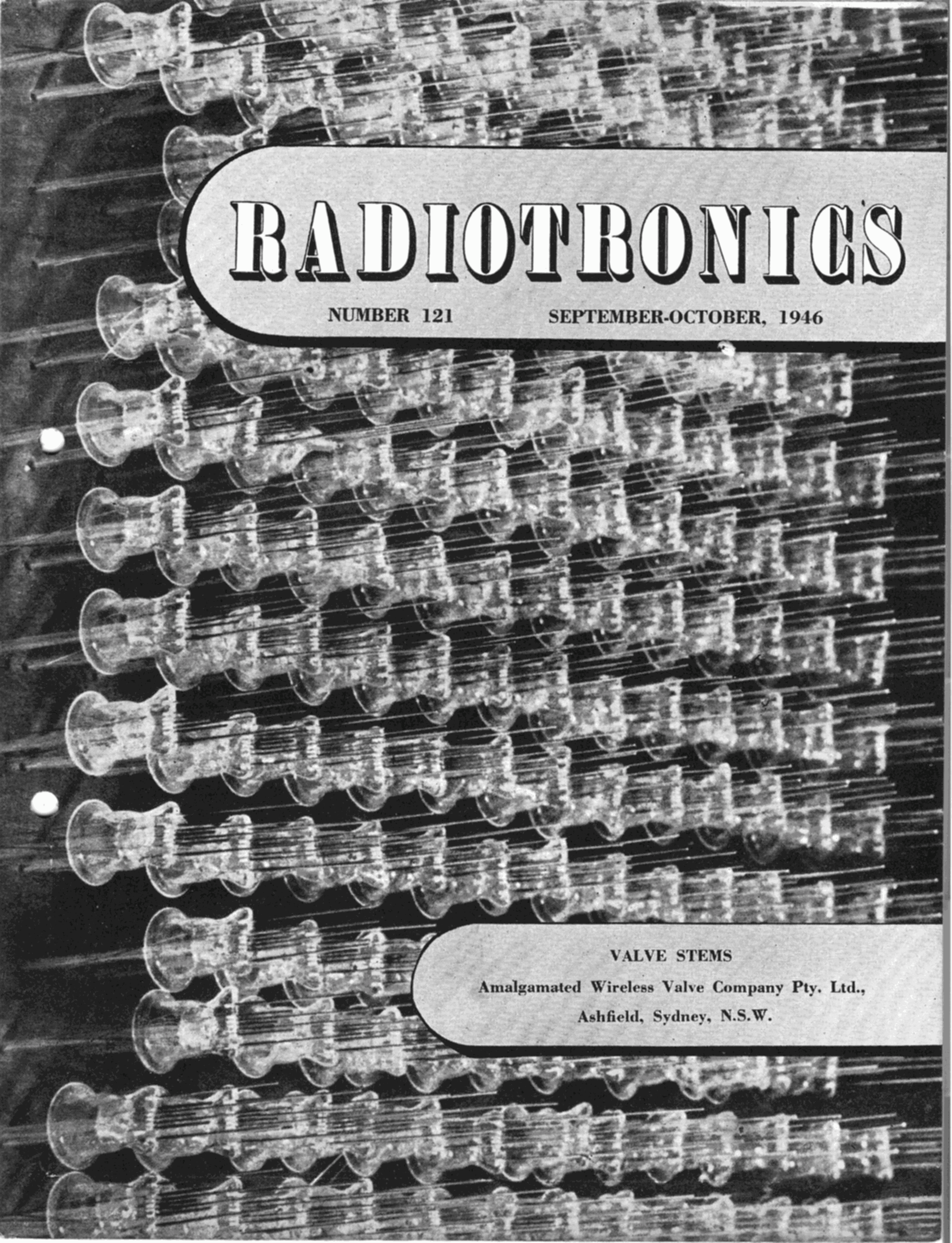
RADIOTRONICS

NUMBER 121 SEPTEMBER-OCTOBER, 1946



VALVE STEMS

Amalgamated Wireless Valve Company Pty. Ltd.,
Ashfield, Sydney, N.S.W.

RADIOTRONICS

Number 121 ★ SEPTEMBER/OCTOBER, 1946

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Technical Editor

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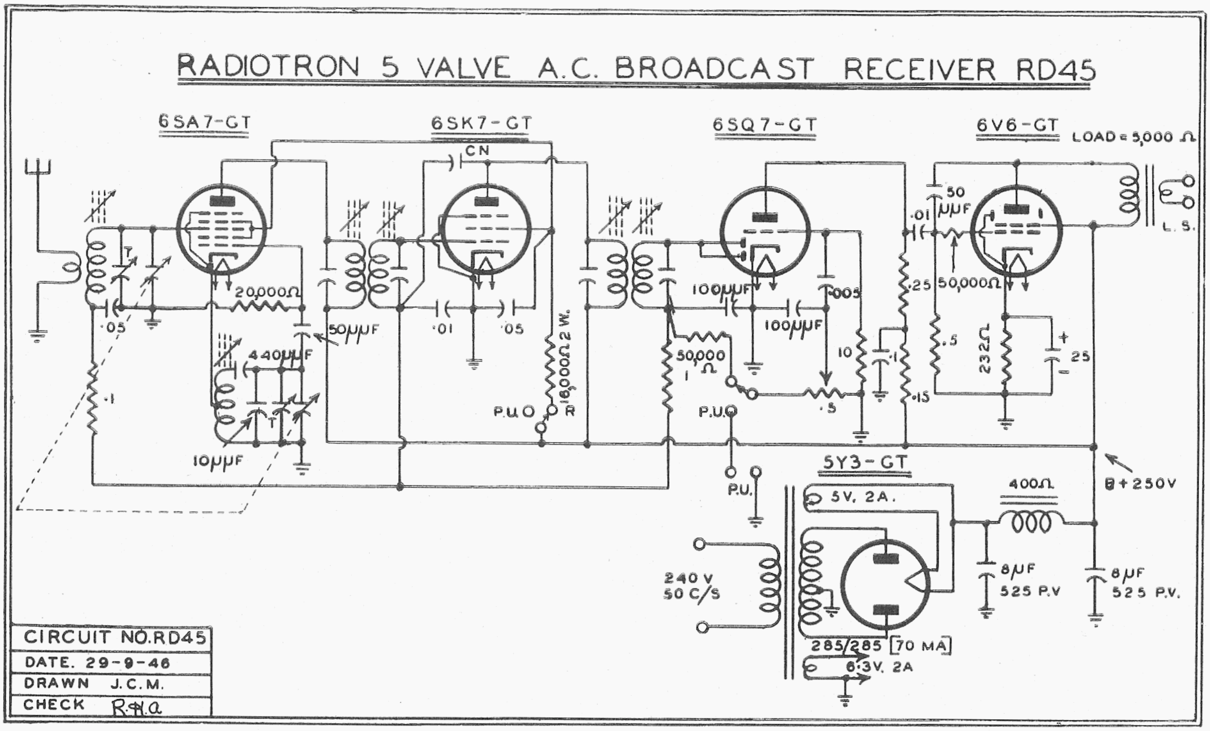
Radiotron Receiver RD45

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



5 Valve A.C. Broadcast Receiver

A 5 valve receiver of simple and economical design is described, which employs a minimum of components consistent with good performance.

While this design is not to be preferred to the previously described RC41 and RC42, for general performance, it may be more suitable in applications where economy is a major factor. Every effort has been made to reach a good compromise between performance and price. It would, of course, be simple to make this receiver either a little better or a little cheaper. The relative merit of a step in either direction is largely a matter of opinion. We feel, however, that this design strikes a good balance and should be quite useful.

GENERAL DESCRIPTION OF THE CIRCUIT.

The 6SA7-GT is again the preferred converter valve—as in all our previous post-war receivers. The circuit arrangement is conventional and needs little comment. There should be no difficulty in obtaining satisfactory oscillator operation, as the high oscillator g_m is a feature of this valve. The oscillator coil used in our model was wound in a single pie of about 80 turns, tapped at the sixth turn. A $\frac{1}{2}'' \times \frac{3}{8}''$ dia. slug was used for adjusting the inductance. This coil is a standard type, and was supplied by Aegis.

The intermediate amplifier valve is a 6SK7-GT which makes possible a high gain with economical transformers. A simple form of neutralizing is used, and although there is only one capacitance to be neutralized, our previous remarks concerning the 6SF7-GT neutralization apply in general to this valve. We believe that the very small increased cost, resulting from the addition of one capacitor and some extra operational time, is more than offset by the elimination of the need for a valve shield, and more uniform performance of production receivers.

Both the 6SA7-GT and the 6SK7-GT operate with diode contact potential and a.v.c. potential as the sole source of bias. To avoid exceeding the recommended maximum screen and plate dissipations for zero signal input, the screen potential is limited to approximately 85 volts.

Some interesting constructional information is included in an R.C.A. application note titled "The 6SK7 as an I.F. Amplifier" which was reprinted in "Radiotronics" No. 100. This information is particularly relevant at this time when single-ended valves are becoming so widely used.

The functions of detector and audio amplifier are performed by the diode and triode sections of the

6SQ7-GT. Simple a.v.c. is employed for economy reasons. Both diodes may be connected together or, alternatively, one only may be used with the other earthed. This latter arrangement has some virtue in that it tends to reduce the residual output, which is sometimes present with the volume control turned right off, due to diode-plate capacitance. The diode connected to pin 4 is used for detection and pin 5 earthed, so that there will be additional shielding provided on the valve socket.

Grid leak bias is used in the audio amplifier. This method of biasing is under suspicion, but we are unable to find any definite fault with it, and undoubtedly it is a convenient arrangement.

The plate load is decoupled from the supply source to reduce hum. This proved to be approximately equal in effectiveness to increasing the second filter capacitor to 16 μF.

An interesting arrangement is used to limit the high frequency rise of output from the 6V6-GT. This consists of a very small capacitor connected between plate and grid to provide negative feedback for the higher frequencies. Besides being slightly more economical than the alternative system of a larger capacitor between plate and B+, there are also the usual advantages of negative voltage feedback for the higher frequencies.

TEST RESULTS

For output of 50 milliwatts absolute.

Input to	Frequency	Input	Ratio	ENSI.
6V6-GT control grid	400 c/s	0.86V	—	—
6SQ7-GT control grid	400 c/s	0.015V	57	—
6SQ7-GT diode	455 Kc/s	0.15V	—	—
6SK7-GT control grid	455 Kc/s	1.08mV	139	—
6SA7 GT control grid	455 Kc/s	19μV	57	2.0
6SA7-GT control grid	600 Kc/s	23.5μV	46	2.5
6SA7-GT control grid	1000 Kc/s	21.0μV	51.5	2.4
6SA7-GT control grid	1400 Kc/s	20.0μV	54	2.42
Aerial	600 Kc/s	3.6μV	6.5	0.41
Aerial	1000 Kc/s	3.7μV	5.7	0.48
Aerial	1400 Kc/s	3.8μV	5.3	0.57

Oscillator Performance

Frequency	i_c	e_k (RMS)	e_0 (RMS)
540 Kc/s	0.32 mA	1.7 V	11.4V
1000 Kc/s	0.42 mA	1.33V	13.6V
1600 Kc/s	0.42 mA	1.1 V	13.2V

Selectivity

Times down	1000 Kc/s input to aerial Bandwith Kc/s
3	6.9
10	10.0
30	13.2
100	17.3
300	21.9
1000	28.5
3000	36.1
10000	47.1

Automatic Volume Control

Input	Output db (0=0.6w)	A.V.C. volts
3 μV	—13.5	0.2
10	+ 1.5	0.6
30	11.0	1.0
100	18.0	2.25
300	21.0	3.5
1 mV	23.5	4.5
3	25.0	5.4
10	26.5	6.5
30	28.0	8.0
100	30.0	11.3
300	32.0	14.5
1 V	35.0	19.0

Audio Frequency Response

(input to pick-up terminals)

Frequency (c/s)	Output (db)
50	— 12
100	— 4
300	0
1000	0
3000	— 3.5
5000	— 7.0
10000	— 12.5

Plate and Screen Dissipation

With no signal input the plate (or screen) dissipation in watts is equal to the plate (or screen) voltage multiplied by the current in amperes.

With signal input, the plate dissipation is equal to the plate input in watts less the power output in watts; the plate input being the plate voltage multiplied by the plate current, both measured with the input signal applied to the grid.

With signal input, the screen dissipation is equal to the screen input—that is, the screen voltage multiplied by the screen current in amperes, both being measured with the input signal applied to the grid.

In all cases, the plate (or screen) voltage should be measured between the plate (or screen) and cathode. If a resistance is in the plate (or screen) circuit, the voltmeter should have a resistance at least 100 times as great as the dropping resistance. Alternatively, correction should be made for the voltmeter loading, or the plate voltage calculated from a knowledge of the supply voltage less the voltage drop in the resistor; in this case it will be necessary to measure the plate (or screen) current and the resistor.

RADIO RECEIVER DESIGN

(Part 3)

TRACKING

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

In the two previous articles on receiver design, consideration was given to various general aspects of receiver layout and construction. It is now proposed to deal with some of the actual details of receiver design.

One of the most important topics for the receiver designer is superheterodyne tracking. This will necessarily cover the computation of the component values for the signal and oscillator circuits. The signal circuits covering the aerial and r-f stages. Before proceeding with an actual design it is worthwhile to consider the methods available and then to see what practical discrepancies are likely to occur in the component values.

General

A number of solutions have been made in the past, for the case where three point tracking is desired between the signal and oscillator circuits in a superheterodyne receiver. All of the methods have had fairly serious defects, the most usual being that no account is taken of the effect of the presence of the oscillator coil feedback winding on the tuning of the oscillator circuit, and the presence of the primary windings in the aerial and r-f coils are usually ignored. In the few cases where attempts have been made to include the effects of the primary windings the solutions are cumbersome, and if a change in any component in the circuits is contemplated, after a solution has been made, a very considerable amount of re-calculation becomes necessary. The most useful practical solution so far made, appears to be that due to R. Payne-Scott and A. L. Green. This method, although it ignores the presence of windings on the coils other than the actual tuned circuits, allows rapid computation for a given set of conditions. Further the use of a slide rule gives sufficient accuracy for most practical cases, which is advantageous since many methods require seven figure log. tables for accurate results. Auxiliary formulae are also derived which, once an initial design has been established, allow a small change in one component of the oscillator circuit to be readily interpreted in terms of the changes required in the other oscillator circuit components, to maintain three point tracking. The discrepancy using this method, between calculated and practical results has usually been found to be so small that inductance alterations can be taken up by means of the usual iron "slug", and the trimmer capacitance takes care of any capacitance errors likely to occur.

So that this method may be readily appreciated it will be illustrated by means of actual solutions. The derivation of the design formulae are given in the reference, quoted below. This reference also gives a series of design charts to allow a more rapid determination of the circuit components. As a final point, the choice of the actual tracking frequencies is important, and in order to minimize errors the upper and lower tracking frequencies are chosen somewhat in from the band limits. The third tracking frequency is most often the arithmetic mean of the outer tracking frequencies, although the geometric mean is sometimes used. No solution has yet been made for the optimum tracking frequencies, and even if this were available it would be subject to practical difficulties but would, of course, be of assistance. Three point tracking is only achieved, when the oscillator circuit uses a series padding condenser, under the condition that the oscillator frequency is higher than the signal frequency by an amount equal to the i-f. If the oscillator frequency is less than the signal frequency, it is necessary to use series padding condensers in the signal circuits to obtain three point tracking, but this application is usually limited to special purpose receivers having small frequency coverages.

Formulae and Examples

In practice it is usually sufficient to design with the tracking frequencies coincident with the band limits, and then to obtain actual circuit tracking at the specified tracking frequencies by adjustment of the signal inductance and capacitance trimmers. The formulae for these conditions are given, but it should be noted that more accurate formulae are available in the reference for the case where the tracking frequencies are not coincident with the band limits.

Figure 1 shows the circuit for which the formulae are derived. The position of the padder (which is fixed for practical convenience) is of some importance, and is preferably placed at the earthly end of the oscillator coil, to keep untracked stray capacitances to a minimum.

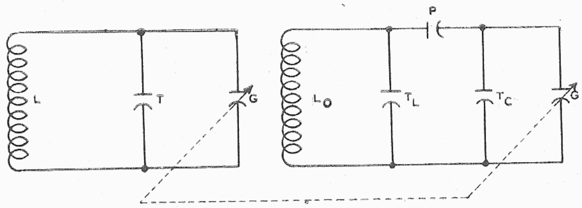


FIG. 1

Signal Circuits

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

where:

T = the total trimmer and stray capacitance across the signal circuit (Aerial or R.F.).

G_{max} = the total incremental capacity of the gang condenser e.g. the capacitance range is 10—400 $\mu\mu$F. G_{max} is 400—10 = 390 $\mu\mu$F.

$$\alpha = \frac{\omega_2}{\omega_1} = \frac{f_2}{f_1}$$

$$= \frac{\text{High frequency tracking point}}{\text{Low frequency tracking point}}$$

As a simple example: a gang is available with a capacity range of 10—400 $\mu\mu$F. and the highest frequency to which it is desired to tune is 1,600 Kc/s and the lowest frequency 550 Kc/s. What is the total value of T which will permit this frequency coverage?

Then

$$G_{max} = 400 - 10 = 390 \ \mu\mu\text{F}.$$

$$\alpha^2 = \left\{\frac{f_2}{f_1}\right\}^2 = \left\{\frac{1600}{550}\right\}^2$$

$$= 8.46$$

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

$$= \frac{390}{8.46 - 1}$$

$$= \frac{390}{7.46}$$

$$= 52.1 \ \mu\mu\text{F}.$$

So if we assume the strays to be about 30 $\mu\mu$F. including the minimum capacity of 10 $\mu\mu$F. for the gang, the trimmer required across the gang is 20 $\mu\mu$F. This could be a variable trimmer to take account of any error in the estimate of 30 $\mu\mu$F.

The calculation of the coil inductance required is the next step. This is found from the expression

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

where:

L = the inductance required.
T has the same meaning as before
$\omega_2 = 2\pi \times f_2$
$= 6.28 \times$ high frequency tracking point

Continuing with the previous example,

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

or for simplification in working

$$L = \frac{25330}{T\ f_2^2}$$

where:

L is in microhenries
T is in micro-micro-farards
f_2 is in megacycles per second

Then:

$$L = \frac{25330}{52.1 \times 1.6^2}$$

$$= 190 \ \mu\text{H (approximately)}.$$

Which would be the approximate value of the inductance required for the aerial and R.F. coil secondaries. This value would, of course, require a slight adjustment in practice. It should be observed that the formulae used in this section are quite general, and could be applied to find the components of any single tuned circuit to give a particular band coverage. More will be said later about the arrangements for aerial and R.F. coil primaries. We will now proceed to determine the constants for the oscillator section, taking the tracking points as the band limits of 550 Kc/s and 1600 Kc/s, and using an i-f of 455 Kc/s.

Oscillator Circuit

The formulae given below will apply where the third tracking frequency is the arithmetic mean of the outer tracking frequencies. In our case this would mean the third tracking frequency (f_3) is

$$\frac{f_1 + f_2}{2} = \frac{550 + 1600}{2}$$

$$= \frac{2150}{2}$$

$$= 1075 \text{ Kc/s.}$$

Before stating the formulae there are several points worth discussing. In figure 1 we see that two capacitances are shown, T_L and T_c. Now these represent trimmers plus stray capacitance, and we could include all of our trimmer capacitance in T_L or in T_c, or else divide the trimmer capacitance between them. Since there is always some value of T_L present due to stray capacitance across the oscillator coil L_0, we can select a fixed value for T_L and make T_c include our variable trimmer. The other case, which sometimes has practical advantages, is to let T_L include a variable trimmer and take a fixed value for T_c. This alternative is discussed in the reference, but for simplicity we will assume a fixed value for T_L and then calculate the values of T_c, P and L_0.

A usual value for T_L would be in the vicinity of 5 to 10 $\mu\mu$F., for our particular case. Take 8 $\mu\mu$F. as a convenient practical value. We are now in a position to proceed with the calculations.

The formulae required are:

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

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

(This value is often critical, and may require the use of logs., or greatest accuracy, when the value is close to 1.)

$$Tc_{max} = \frac{G_{max}}{p\beta^2 - 1}$$

$$P_{min} = P_{max} - Tc_{max}$$

$$P = \frac{P_{min}}{2}\left[1 + \sqrt{\frac{P_{min} + 4\,T_L}{P_{min}}}\right]$$

$$T_c = P_{max} - P$$

$$L_0 = \frac{P_{min} \qquad P_{max}}{Tc_{max}\,P^2\,(\omega_2 + \omega_1)^2}$$

where:

G_{max} has the same meaning as previously.

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

$$= \frac{f_2}{f_1}$$

$$= \frac{\text{highest tracking frequency}}{\text{lowest tracking frequency}}$$

$$\beta = \frac{\omega_2 + \omega_i}{\omega_1 + \omega_i}$$

$$= \frac{f_2 + f_i}{f_1 + f_i}$$

f_i = intermediate frequency.

P = padder capacitance.

T_c = trimmer and stray capacitance across the gang condenser G.

L_0 = Oscillator coil inductance.

Proceeding with our example, we first calculate,

$$\alpha = \frac{f_2}{f_1}$$

$$= \frac{1600}{550}$$

$$= 2.91$$

$$\alpha^2 = 8.46$$

$$\beta = \frac{f_2 + f_i}{f_1 + f_i}$$

$$= \frac{1600 + 455}{550 + 455}$$

$$= \frac{2055}{1005}$$

$$= 2.045$$

$$\beta^2 = 4.18$$

Then

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

$$= \frac{8.46}{4.18} \times \frac{3+2.91}{3+2.045} \times \frac{1+3\ (2.045)}{1+3\ (2.91)}$$

$$= \frac{8.46}{4.18} \times \frac{5.91}{5.045} \times \frac{7.135}{9.73}$$

$$= 1.74$$

$$P_{max} = \frac{G_{max}}{p-1} = \frac{390}{1.74-1} = \frac{390}{.74} = 526\ \mu\mu F.$$

$$Tc_{max} = \frac{G_{max}}{p\beta^2 - 1}$$

$$= \frac{390}{(1.74 \times 4.18) - 1}$$

$$= \frac{390}{6.275}$$

$$= 62.1\ \mu\mu F.$$

$$P_{min} = P_{max} - Tc_{max}$$

$$= 526 - 62.1$$

$$= 463.9\ \mu\mu F.$$

$$P = \frac{P_{min}}{2}\left[1 + \sqrt{\frac{P_{min} + 4T_L}{P_{min}}}\right]$$

$$= \frac{463.9}{2}\left[1 + \sqrt{\frac{463.9 + 4\ (8)}{463.9}}\right]$$

$$= \frac{463.9}{2}\left[1 + \sqrt{\frac{495.9}{463.9}}\right]$$

$$= \frac{463.9}{2}\ (2.035)$$

$= 470\ \mu\mu F.$ (slide rule accuracy)

$$T_c = P_{max} - P$$

$$= 526 - 470$$

$$= 56\ \mu\mu F.$$

$$L_0 = \frac{P_{min}\ \ P_{max}}{Tc_{max}\ \ P^2\ \ (\omega_2 + \omega_i)^2}$$

$$= \frac{463.9 \times 526}{62.1 \times 470^2 \times 4\pi^2\ (f_2 + f_i)^2}$$

$$= \frac{463.9 \times 526 \times 10^6}{62.1 \times 470^2 \times 39.5\ (2.055)^2}\ \mu H.$$

$$= 107\ \mu H.$$

Summarizing our results, the total frequency coverage is 550 to 1600 Kc/s using a gang having a capacity range of 10 — 400 $\mu\mu$F. The signal circuit inductance is 190 μH., and the trimmer capacitance, including gang minimum capacitance, is 52.1 $\mu\mu$F. These apply to the aerial and r-f coils with practical adjustments necessary to allow for the effects of the primary windings. The oscillator circuit inductance is 107 μH., the series padder is 470 $\mu\mu$F., and the total trimming capacity across the gang, including the gang minimum capacity is 56 $\mu\mu$F. The readjustment for the oscillator coil inductance is normally quite small, and could be taken up by the usual iron "slug". The capacitance across the gang would normally be made up of strays (including gang minimum capacity) which would amount to very approximately, 30 $\mu\mu$F. and the remainder could be a variable trimmer with the usual range of about 5 to 30 $\mu\mu$F. The padder normally does not require readjustment. The value chosen for T_L is not critical, provided the error is not excessive, and is automatically taken care of during alignment of the receiver. The complete design appears, at first sight, to be rather long but once the symbols become reasonably familiar the whole solution can be accomplished very quickly.

It is now proposed to show how a small change in one oscillator component can be compensated for in the other components without recalculation. The important point which should be observed is that a range of values for the various components can be chosen, within the maximum and minimum limits calculated, and the most convenient values selected. Three point tracking is still maintained at the selected frequencies, the changes required in the various components, for a selected change in another component, being independent of the tracking and intermediate frequencies, provided these remain unchanged. The relations required are:

$$\delta T_c = -\delta P$$

$$\delta T_L = \left[1 + \frac{2T_L}{P}\right]\delta P$$

$$\delta L_0 = -\left[\frac{2L_0}{P}\right] \delta P$$

δL_0 etc. means the amount of change in the component concerned. Taking as an example our previous results. Suppose the padder available is 475 $\mu\mu$F. and not 470 $\mu\mu$F. as calculated. Then the changes required in the other components are

$$\delta T_c = -\delta P$$
$$= -5\ \mu\mu\text{F.}$$

so that the value of T_c is decreased by 5 $\mu\mu$F. from 56 $\mu\mu$F. to 51 $\mu\mu$F.

$$8\ T_L = \left[1 + \frac{2T_L}{P}\right] \delta P$$

$$= \left[1 + \frac{2 \times 8}{470}\right] 5$$

$$= (1 + .034)\ 5$$
$$= 5.17\ \mu\mu\text{F.}$$

then T_L should be increased by 5.17 $\mu\mu$F.

$$\delta L_0 = -\left[\frac{2L_0}{P}\right] \delta P$$

$$= -\frac{2 \times 107}{470} \times 5$$

$$= -2.28\ \mu\text{H.}$$

So the new value for L_0 is 104.72 μH. These changes can often all be accomplished very readily by slight alteration in the inductance and capacitance trimmer settings.

Method of Using the Results

All the components of the signal and oscillator tuned circuits have been determined with the exception of the aerial and r-f coil primary windings, and the oscillator feedback winding. For the *r-f coil* suppose we require a high impedance primary circuit. Assume the total stray capacitance due to the valve and wiring is 25 $\mu\mu$F. Then following ordinary practice we resonate the primary at about .7 of the lowest signal frequency. This means that the primary inductance of the r-f coil resonates with the 25 $\mu\mu$F. stray capacitance at 550 × .7 = 385 Kc/s and the inductance required is given by

$$L = \frac{25330}{f^2 \times C}$$

where f is in Mc/s.
C is in $\mu\mu$F.
L is in μH.

then

$$L = \frac{25330}{.385^2 \times 25}$$
$$= 6{,}840\ \mu\text{H.}$$

or 6.84 mH.

The coefficient of coupling normally used lies between .15 and .2 for a practical design. This gives a good compromise between tracking and gain. If the coupling is too high tracking is adversely affected. More elaborate solutions are available but the final results are not appreciably different. For the case of a low impedance primary the procedure is the same except that the natural resonant frequency of the primary would be chosen as approximately 1600 × 1.4 = 2,240 Kc/s.

Other methods are sometimes adopted, with high impedance primaries, such as the use of a damping resistor, R of a few thousand ohms. Under these circumstances a smaller winding can be used but the stage gain is materially reduced. The actual values of R and L_p to be chosen can be readily determined from the solution of equivalent transformer circuit, and a compromise between gain and primary inductance arrived at.

The *aerial coil* is a somewhat more difficult problem than the r-f coil, as the nature of the aerial to be used is often not known. A solution can easily be made for the case of the usual standard dummy, or alternatively for the worst type of aerial likely to be encountered in normal service. A procedure often employed with a high impedance primary aerial coil is to make the inductance sufficiently large to resonate below the lowest tuning frequency with the smallest capacity, due to a short aerial, likely to be encountered. The effect of larger aerials is to merely shift the primary resonant frequency to a still lower value because of the increased aerial capacitance. As a typical case suppose the smallest aerial capacitance, which may be an indoor aerial, likely to be used with the receiver is 60 $\mu\mu$F., then the primary inductance for the aerial coil is

$$L = \frac{25330}{f^2 \times C}$$

f is again chosen as .7 of the lowest signal frequency (385 Kc/s as before).

$$L = \frac{25330}{.385^2 \times 60}$$
$$= 2{,}280\ \mu\text{H.}$$
$$= 2.28\ \text{mH.}$$

A usual value for the coefficient of coupling between primary and secondary is .2 to .3, which offers a good compromise between gain, signal to noise ratio, and tracking. Other more elaborate methods can be found in the literature. The low impedance primary aerial coil can be dealt with in a similar manner to that indicated for the r-f coil but using the aerial capacitance value of about 60 $\mu\mu$F., or whatever value is chosen, and k is again about .2 to .3.

It is worth observing that the lower values of k give improved tracking because the effects of the primary windings on the secondaries are reduced as the receiver is tuned across the band. These effects naturally vary with the frequency setting.

The choice between high and low impedance primaries for the aerial and r-f coils is usually a matter of compromise, but it can be taken that for the normal broadcast receiver high impedance primaries are preferable. This subject is fully discussed in reference (3) given below.

The final point to be considered is the method of obtaining the value of inductance required for the oscillator feedback winding. Methods are available which allow the winding to be calculated, but usually the results give inductance values which are far too small. A direct practical approach is usually the best way out of the difficulty, keeping in mind that the feedback winding should be kept small so that its natural resonant frequency is well above the highest oscillator frequency. A small feedback winding, and a fairly high coefficient of coupling, are usually a necessary compromise. The small winding resonating well outside the oscillator tuning range reduces tracking errors to a minimum, provided the value of k is not too high. Usually the coefficient of coupling is about .5 to .6. The practical approach is to use a feedback winding with about a half to one third of the total turns on the tuned section of the oscillator coil, closely coupled, and then to remove turns until the desired value of oscillator grid current is obtained.

Summary

An approximate method has been shown for finding the signal and oscillator circuit components. More accurate methods are available from the references given below, but often the increased accuracy is not required in practice. The method can be applied for any band of frequencies, the steps being the same as outlined.

The procedure for adjusting the oscillator and signal circuits is to set the bandlimits by adjusting the oscillator coil slug at the low frequency limit, and the trimmer condenser at the high frequency limit. The signal trimmer condensers and inductances are then set in the usual manner at the required tracking frequencies, which are usually in from the ends of the band. Tracking is then checked at the third tracking frequency used in the calculations. In the example used the bandlimits are 550 and 1600 Kc/s, (set by the oscillator circuits) and the tracking frequencies could be 600 Kc/s, 1400 Kc/s (set by the signal circuits) and about 1000 Kc/s (actually 1075 Kc/s in our case).

It should be observed that the centre tracking frequency is usually obtained exactly only when the padder is also adjustable. A close approximation is all that is normally required, however, as it is the minimizing of tracking errors across the whole tuning range that is important. The tracking points are only incidental, and if the centre tracking frequency is only a few Kc/s away from the value used in the calculations the increase in tracking error normally will not be appreciable.

Variable inductances and padders could be used in the developmental model, and when the required values are obtained these can then be measured and used as standards, as it may be desirable to use fixed padders and inductances in production. Suitable limits for manufacturing tolerances would have to be set, and these could be determined experimentally.

References

(1) R. Payne-Scott and A. L. Green "Superheterodyne Tracking Charts — II."
A.W.A. Tech. Review, Vol. 5, No. 6, 1941.

(2) A. L. Green "Superheterodyne Tracking Charts — III."
Vol. 6, No. 2, 1943.

(3) K. R. Sturley "Radio Receiver Design," Part 1 — Chapman & Hall.

Revised R.M.A. Type Designations for Other than Receiving and Cathode Ray Tubes

In Radiotronics 112 on page 38 a proposed R.M.A. valve type designation system was reprinted. The system covered transmitting and special purpose valves, but was not intended to cover receiving or cathode ray tube types. Further advice has since been received from the R.M.A. Engineering Department, which rescinds the designation system previously given. A new system of numbering has now been introduced, and a summary of this is set out below.

1. It shall be standard to use the following system of type designations for tubes and devices exclusive of receiving and cathode-ray tubes;

2. The type designation shall consist of a pure numeric starting with 5,500 and shall be assigned consecutively and chronologically in the order of type number request.

3. A new type designation shall be assigned to a new version of a prototype whenever the new version is not completely interchangeable with the prototype.

Typical type designations quoted as examples of these are 5501, 5712, 5923, 6234, 6545.

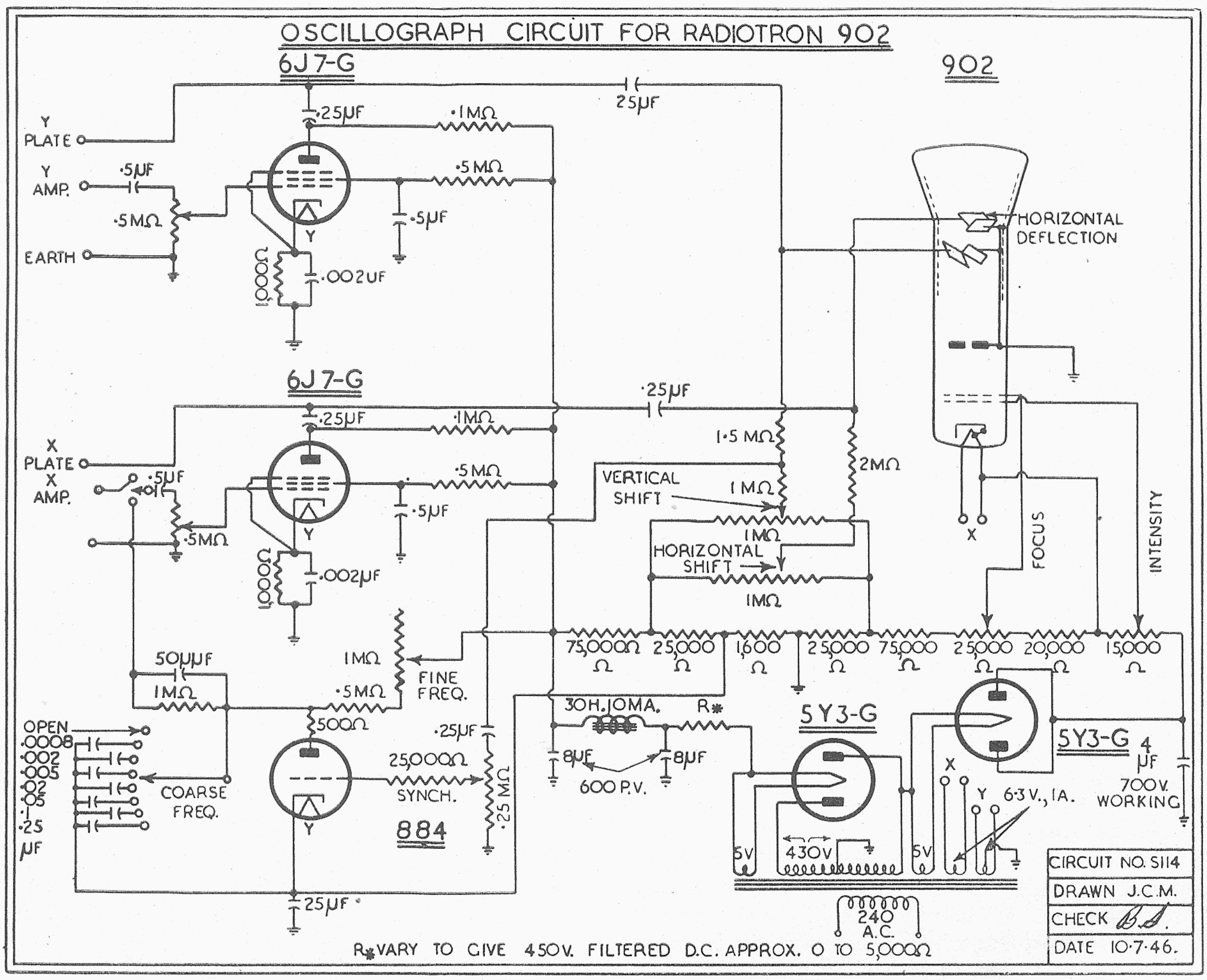
Radiotron 2 Inch Oscillograph Circuit S114

We have received numerous requests for a reprint of the 2 inch Oscillograph Circuit for Radiotron 902 published in Radiotronics 86. The original circuit has been improved by the addition of controls to provide both horizontal and vertical shift, together with a few other minor changes. The new circuit (S114) is shown below, and includes both horizontal and vertical amplifiers, each using type 6J7-G although type 6SJ7-GT could be used as an alternative. The 0.002 μF. cathode bypass condensers are for the purpose of giving some degree of boost of high audio frequencies.

Type 884 gas triode is used in the sweep circuit with quite conventional controls for frequency adjustment and synchronization.

The power is supplied by a 430/430 volt transformer and two rectifiers, each type 5Y3-G or 5Y3-GT. One rectifier is used under full-wave conditions, with a condenser input filter and adjustable series resistor to provide 450 volts between the extreme positive end of the voltage divider and earth. The other rectifier is used under half-wave conditions, with both plates tied, to provide a negative voltage of approximately 600 volts with respect to earth.

The Anode No. 2 of the Cathode Ray Tube is earthed, the cathode being negative with respect to earth, in accordance with normal practice.



The AV10A Ionization Gauge

By J. ANDERSON, A.S.T.C. and D. MILLER, A.S.T.C.

General

The ionization gauge is a thermionic device for measuring very low gas pressures in a vacuum system. We can illustrate its action by considering a triode containing a trace of gas, and connected as shown in Fig. 1.

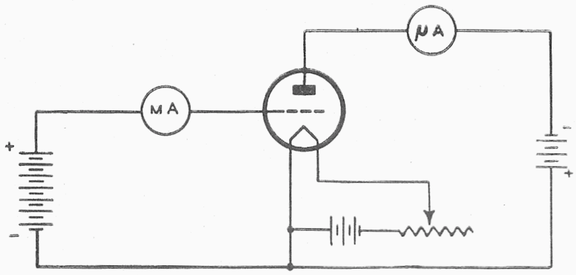


FIG. 1

Electrons are accelerated from the filament towards a positively charged electrode, in this case the grid, while the plate is maintained at a negative potential so that it collects the positive ions formed by ionization of the residual gas by the electrons which pass through the grid spaces. In a particular case we might have a current of 10mA in the grid circuit, and find that we have a plate current of a few micro-amperes. It is easy to understand that the more gas present in the triode, the more ionization and the greater the "gas" current collected by the plate. Hence the anode current is some measure of the gas pressure in the envelope. More correctly, since clearly ionization will increase with increasing electron current to the grid, the ratio of plate to grid current is a measure of the gas pressure.

Basically then, the ionization gauge is an arrangement of three electrodes mounted in an envelope which can be connected to the vacuum system under test; and provided with suitable metering and supply circuits. Of the electrodes we require an electron emitter, an electron collector, and a gas ion collector. Various forms of electrodes are used, but the common triode arrangement of a filament surrounded by a grid surrounded by a plate is quite a good one. It is usual to use the plate as the ion collector and the grid as the electron collector, as this gives the greatest sensitivity; but the reverse arrangement is quite feasible. (Indeed this connection is used in measuring gas currents in an ordinary sealed off valve—the valve is used as its own ionization gauge.)

An ionization gauge must be calibrated before use, and the calibration depends on the type of gas being measured, as is noted later. The normal range of pressure for which the guage can be used is 10^{-2} to 10^{-6} m.m. of Hg, while pressures as low as 10^{-10} have been recorded. At a given pressure, the ratio of plate to grid current is different for different values of grid current. For this reason it is necessary to adjust the grid current to some definite value, usually in the range of 10 to 20 mA.

A good control circuit is necessary to ensure stable operation, as the guage is very sensitive to the slightest fluctuations in supply voltages and gas pressures.

If the normal range of pressures is exceeded, the anode grid current pressure ratio becomes non-linear. Further, in most designs the filament will be damaged if air is admitted during use, as a result of a crack in the vacuum system for example.

The ionization gauge is capable of reading condensible vapour pressures such as water and oil, to which the McLeod gauge will not respond. Instantaneous readings are obtained.

It can be used to record continuously pressure variations in any vacuum system, where the gas or vapour pressures are below 10^{-2} m.m., such as an electron tube undergoing exhaust and activation.

The AV10A Ionization Gauge

DESIGN FEATURES.

This is an inexpensive medium sensitivity type of gauge of robust construction and long life, and is well suited to uses where these features, rather than high sensitivity, are the chief requirements. See Fig. 2.

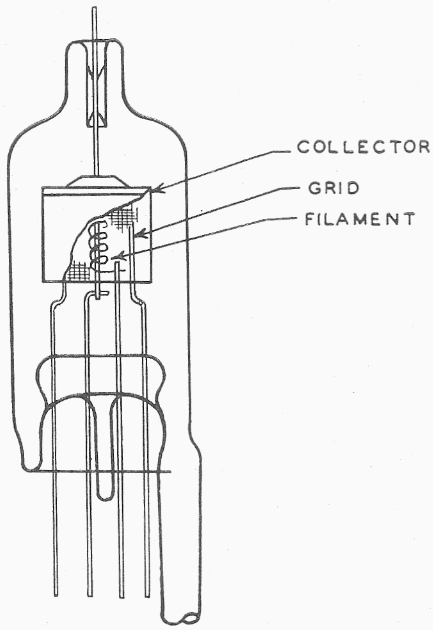


FIG. 2 AV10A IONIZATION GAUGE.

The electrodes and leads are arranged to avoid electrical leakage, which can be troublesome when currents of the order of microamperes are to be measured.

Physical Data

Collector	Nickel cylinder
Grid	Nickel wire mesh
Filament	Tungsten
Envelope	Soft glass
Height	3"
Diameter	$1\frac{1}{8}$" O.D.
Tubulation	Soft Glass $\frac{1}{8}$" O.D.

Standard Operation

Filament	4 to 6 volts; 2.5 to 3.0 amp.
Grid	300 volts max., 20 mA.
Collector	–20 to –40 volts.
Emission	at 5.0 volts filament and 300V on grid approx. 32mA.
Sensitivity	1.5 microamp / mA. / micron for argon at 20 mA. grid current (1 micron = 10^{-3} m.m. Hg.)

Operation

The gauge is preferably sealed in a vertical position directly to the system under exhaust. For reliable use it should be well degassed before operation. A preliminary bake in an oven for at least 10 minutes at 300°C is recommended. With the grid connected to 250V d.c. through a 40 watt lamp and 250V d.c. applied directly to the collector, on increasing the filament temperature (filament current approx. 3.5 amps) the grid becomes red hot and the removal of gas from both the grid and collector is effected in a few minutes. The actual current through the electrodes is about 50 mA. in each case.

In taking the reading the filament current is adjusted to give the desired grid current and the resulting positive ion current is read on a micro-ammeter. The pressure in the gauge is read from the calibration curves as explained in the next paragraph.

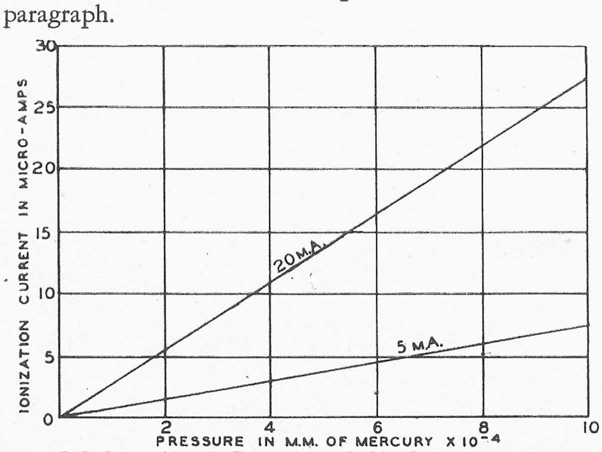


FIG. 3 IONIZATION GAUGE CALIBRATION CURVE

Calibration

The AV10A ionization gauge was calibrated against a McLeod Gauge over the top portion of its range and a curve drawn showing the relation between collector current and pressure at constant grid current. Owing to the linearity between pressure and ionization current below 10^{-3} m.m. this curve may be extrapolated to zero.

To obtain repeatable readings it was essential to degas the gauge completely before calibration*. It was also necessary to provide cold traps between the diffusion pump and the ion gauge and between the McLeod and ion gauge to minimize vapour pressure effects, especially those due to the diffusion pump oil, and the McLeod gauge mercury. Fig. 3 shows curves for type AV10A calibrated with argon.

The relative ionization current varies according to the particular gas. The usual residual gases in a vacuum system, viz.: Air, CO_2, H_2O, N_2 have ionization currents which differ from Argon by a factor of less than 2[1]. Other gases show wide variations in ionization current.[2]

Control Circuit

The use of an ion gauge connected as shown in Fig. 1 is extremely difficult where rapid variation in pressure and supply voltage are encountered. These give rise to wide variations in the grid current and hence the collector current.

*Note: These gauges are thoroughly degassed at manufacture and sealed off evacuated to prevent contamination before use. However, after sealing on to a vacuum system for the first time, or on subsequent use after being exposed to atmospheric pressure, process is usually necessary.

A number of circuits have been developed for maintaining grid current constant with varying pressure and voltage. See (3), (4), (5).

References

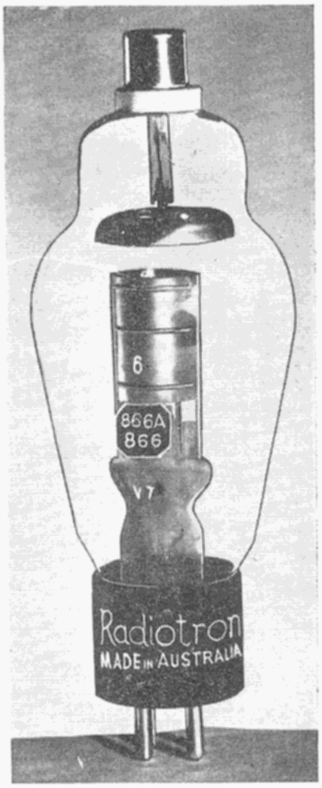
(1) R. H. Healey and J. W. Reed. "Low Pressure Measurements by Means of an Ionization Gauge." A.W.A. Tech. Review, October, 1937.

(2) S. Dushman and A. H. Young. "Calibration of an ionization gauge for different gases" Phys. Rev. December 1 and 15, 1945.

(3) J. B. Hoag and N. M. Smith "A Thyraton controlled Ionization Gauge." Rev. Sc. Inst. 7, 497, 1936.

(4) R. B. Nelson and A. K. Wing. "Emission Regulating Circuit for an Ionization Gauge." Rev. Sc. Inst. 13, 215, 1942.

(5) L. N. Ridenour and C. W. Lampson. "Thermionic Control of an Ionization Gauge". Rev. Sc. Inst. 8, 162, 1937.



Radiotron 866A/866

Radiotron type 866A/866 is a half-wave mercury-vapour rectifier which supersedes the older types 866A and 866. It has a wide application for use in power supplies which require a rectifier capable of a higher voltage and current output than the usual high vacuum types, such as type 5V4-G, etc., but where the output requirements are not sufficiently large to necessitate the use of a rectifier such as type 872-A/872. A common application is in the power supply of fairly small transmitters.

1.4 Volt Miniature Valves

FILAMENT VOLTAGE LIMITS

The maximum filament voltage permitted to be applied to the 1.4 volt miniature valves is 1.6 volts, which is equal to the highest voltage produced by a nominal 1.5 volt A battery. There is therefore no necessity to use a series dropping resistor such as was advised some years ago in connection with the 1.4 volt GT valves. It is possible that the use of this resistor would, to a certain extent, reduce the number of filament breakages during the early life of the valve with new batteries, but this is counter-balanced by the decreased length of life available from A batteries, particularly in portable receivers where the batteries are limited in size.

We therefore do not recommend the use of a series dropping resistor in receivers operated from a nominal 1.5 volt dry cell A battery. This applies equally with the present 1.4 volt GT types as with the 1.4 volt miniature range.

★

New R.C.A. Releases

Radiotron Type 9C25—is a forced-air-cooled grounded-grid type power triode designed for communication and industrial service. It has a maximum rated plate dissipation of 17.5 kilowatts, and can be operated with full plate voltage, plate input, and grid current at frequencies as high as 30 Mc/s. Operation at frequencies up to 100 Mc/s is permissible with reduced ratings.

Radiotron Type 9C27—is a water-and forced-air-cooled power triode designed for communication and industrial service. It has a maximum rated plate dissipation of 25 kilowatts, and can be operated at full ratings on frequencies as high as 30 Mc/s. Operation to 100 Mc/s is permissible with reduced ratings. The type 9C27 is a water-cooled version of the type 9C25.

Radiotron Type 6BG6-G—is a heater-cathode type of beam power amplifier designed for use in applications where high surge voltages occur during short duty cycles. Such applications include hozizontal—deflection circuits in television receiving equipment and in television cameras.

Radiotron Type 12SX7—is a metal twin triode. It is for use wth a 12 cell storage-battery supply, and the filament is rated at 12.6 volts .3 amp. The valve has separate cathodes for each triode section, and the cathode current is limited to 20 mA. maximum in each case. Plate dissipation is 2.5 watts, and the maximum plate voltage is 300.

TYPES DISCONTINUED BY R.C.A.

Type 904—high-vacuum cathode ray tube.

Type 1847—television camera tube.

Radiotron 4-125A/4D21

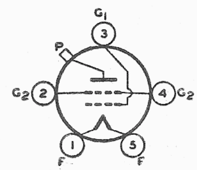
V-H-F Power Tetrode

Ratiotron 4-125A/4D21 is a filament type of four-electrode valve suitable for use as oscillator, amplifier, and class AB_2 modulator. It has a maximum plate dissipation of 125 watts in class C telegraph service, and can be operated with the rated maximum plate voltage of 3,000 volts at frequencies as high as 120 Mc/s. With reduced plate voltage, the 4D21 can be operated at frequencies up to 250 Mc/s.

Having high power sensitivity and high efficiency, the 4D21 is capable of delivering large power output with small driving power. The low value of grid-plate capacitance—0.05 μμF—permits stable operation without neutralization up to about 100 Mc/s. if input and output circuits are adequately isolated.

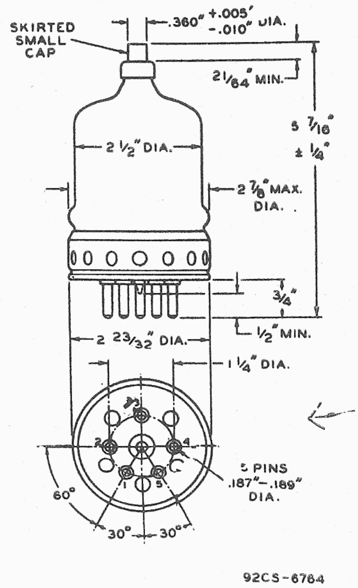
Forced-air cooling of the bulb and plate seal of the 4D21 is required when it is operated near maximum ratings at frequencies above 30 Mc/s. In addition, circulation of air through the perforated shell of the 5-pin base is required to cool the stem.

SOCKET CONNECTIONS
Bottom View



PIN 1: FILAMENT
PIN 2: GRID NO.2
PIN 3: GRID NO.1
PIN 4: GRID NO.2
PIN 5: FILAMENT
CAP: PLATE

DIMENSIONAL OUTLINE



General Data

Electrical:

Filament, Thoriated Tungsten:		
Voltage (AC or DC)	5	Volts
Current	6.2	Amperes
Transconductance, for plate current of 50 ma.	2450	Micromhos
Grid-Screen Mu-Factor	6.2	
Direct Interelectrode Capacitances:		
Grid No. 1 to Plate*	0.05	μμF
Input	10.8	μμF
Output	3.1	μμF

Mechanical:

Mounting Position	Vertical, base up or down
Overall Length	$5\frac{7}{16}'' \pm \frac{1}{4}''$
Seated Length	$4\frac{11}{16}'' \pm \frac{1}{4}''$
Maximum Diameter	$2\frac{7}{8}''$
Cap	Skirted Small
Base	Special Metal-Shell Giant 5-Pin

* With no external shielding and base shell connected to ground.

A.F. Power Amplifier & Modulator —Class $AB_2^{\blacktriangle}$

Maximum Ratings, *Absolute Values:*

DC PLATE VOLTAGE	3000 max.	Volts
DC GRID-No. 2 (SCREEN) VOLTAGE	400 max.	Volts
MAX.-SIGNAL DC PLATE CURRENT**	225 max.	mA.
GRID-No. 2 DISSIPATION	30 max.	Watts
PLATE DISSIPATION**	125 max.	Watts

Typical Operation With Fixed Bias:

Values are for 2 Valves.

DC Plate Voltage	1500	2000	2500	3000	Volts
DC Grid-No. 2 Voltage	350	350	350	350	Volts
DC Grid-No. 1 (Control Grid) Voltage	—41	—45	—43	—51	Volts
Peak AF Grid-No. 1-to-Grid-No. 1 Voltage	282	210	178	198	Volts

Zero-Signal DC Plate Current	87	72	93	55	mA.
Max.-Signal DC Plate Current	400	300	260	260	mA.
Zero-Signal DC Grid-No. 2 Current	0	0	0	0	mA.
Max.-Signal DC Grid-No. 2 Current	34	5	6	3.5	mA.
Effective Load Resistance (Plate to plate)	7200	13600	22200	27700	Ohms
Peak Grid-No. 1 Input Power°	5.2	3.1	2.4	2.5	Watts
Max.-Signal Power Output°°	350	350	400	520	Watts
Total Harmonic Distortion	2.5	1	2.2	1.8	%

Plate-Modulated R.F. Power Amplifier —Class C Telephony

Carrier conditions per Valve for use with a maximum modulation factor of 1.0

Maximum Ratings, *Absolute Values:*

DC PLATE VOLTAGE	2500 max.	Volts
DC GRID-No. 2 (SCREEN) VOLTAGE	400 max.	Volts
DC GRID-No. 1 (CONTROL GRID) VOLTAGE	—500 max.	Volts
DC PLATE CURRENT	200 max.	mA.
PLATE DISSIPATION	85 max.	Watts
GRID-No. 2 DISSIPATION	20 max.	Watts
GRID-No. 1 DISSIPATION	5 max.	Watts

Typical Operation:

DC Plate Voltage	2000	2500	Volts
DC Grid-No. 2 Voltage ♂	350	350	Volts
	50000	70000	Ohms
DC Grid-No. 1 Voltage	—220	—210	Volts
Peak RF Grid-No. 1 Voltage (Approx.)	375	360	Volts
DC Plate Current	150	152	mA.
DC Grid-No 2 Current	33	30	mA.
DC Grid-No. 1 Current	10	9	mA.
Driving Power (Approx.)	3.8	3.3	Watts
Power Output (Approx.)	225	300	Watts

R.F. Power Amplifier & Oscillator —Class C Telegraphy

Key-down conditions per Valve without modulation.

Maximum Ratings, *Absolute Values:*

DC PLATE VOLTAGE	3000 max.	Volts
DC GRID-No. 2 (SCREEN) VOLTAGE	400 max.	Volts
DC GRID-No. 1 (CONTROL GRID) VOLTAGE	—500 max.	Volts
DC PLATE CURRENT	225 max.	mA.
PLATE DISSIPATION	125 max.	Watts
GRID-No. 2 DISSIPATION	30 max.	Watts
GRID-No. 1 DISSIPATION	5 max.	Watts

Typical Operation:

DC Plate Voltage	2000	2500	3000	Volts
DC Grid-No. 2 Voltage	350	350	350	Volts
DC Grid-No. 1 Voltage	—100	—150	—150	Volts
Peak RF Grid-No. 1 Voltage (Approx.)	230	320	280	Volts
DC Plate Current	200	200	167	mA.
DC Grid-No. 2 Current	50	40	30	mA.
Dc Grid-No. 1 Current	12	12	9	mA.
Driving Power (approx.)	2.8	3.8	2.5	Watts
Power Output (approx.)	265	375	375	Watts

▲ Subscript (2) indicates that grid-No. 1 current flows during a part of input cycle.

** Averaged over any audio-frequency cycle of sine-wave form.

° Driver stage should be capable of supplying the No. 1 grids of the class AB_2 stage with the specified driving power at low distortion. The effective resistance per grid-No. 1 circuit of the class AB_2 stage should be kept below 250 ohms.

°° With zero-impedance driver and perfect regulation, plate-circuit distortion will be as indicated. In practical circuit design, the useful power output will be several per cent. less than values shown.

♂ Obtained from fixed supply or through resistor of value indicated in series with modulated plate supply.

Typical Operation:

The following table shows the highest percentage of maximum plate voltage that can be safely used up to 250 Mc/s. Special attention should be given to adequate cooling at the higher frequencies.

FREQUENCY (Mc/s.)	120	150	200	250
MAX. PERMISSIBLE PERCENTAGE OF MAX. RATED PLATE VOLTAGE:				
Class C plate-modulated telephony	100	84	64	—
Class C telegraphy	100	80	64	56

When the 4D21 is operated with the maximum permissible plate voltage at 250 megacycles in class C telegraph service, the power output from the valve is approximately 220 watts, and the required driving power at the valve (including bias loss) is approximately 8 watts.

Grid-No. 1 dissipation of the 4D21 may be calculated approximately from the following equation:—

$$P_g = e_{cmp} I_c$$

where,

P_g = Grid-No. 1 dissipation in watts.

e_{cmp} = Peak positive grid-No. 1 excitation voltage in volts.

I_c = D.C. grid-No. 1 current in amperes.

The value of e_{cmp} may be measured by means of a suitable peak voltmeter connected between filament and grid No. 1. Where no means are available for measuring e_{cmp}, the maximum grid excitation must not be greater than that which causes the power dissipated in the bias source to be 5 watts.