# Radiotronics

Number 123 JANUARY/FEBRUARY 1947

EQUIPMENT AT THE RADIOTRON WORKS, ASHFIELD, N.S.W., FOR TESTING AUSTRALIAN-MADE CONVERTER TYPE VALVES

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Number 123 JANUARY/FEBRUARY 1947

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

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

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

### *PRODUCTION TESTING-CONVERTER VALVES*

*Our cover this issue shows the complete set-up used for the production testing of converter valves in the Radiotron works. Looking from left to right, the equipment comprises a preheater, a push-button relay-operated characteristics test set and a radio frequency (R.F.) noise set in the screened room.*

*The sequence of testing provides that A.C. type valves are first preheated under maximum dissipation and cathode current conditions, then tested for reverse grid current, electrode characteristics, conversion conductance and oscillation in a typical oscillator circuit.*

*Following these tests, valves are finally checked for emission, after which they are operated in the converter stage of a typical radio receiver in the screened room and tested for noise and sensitivity.*

# A Note on Matching By Means of T and H Pads

*By B. Sandel, A.S.T.C.*

A problem sometimes encountered is the connection of a constant impedance source to a number of loads having fixed impedance inputs. A particularly simple solution is possible, which gives correct matching at all junctions, for the special case where the loads all have similar impedances.

Consider fig. 1a which shows a generator E having an internal impedance represented by $Z_g$. This generator is connected by means of an arrangement of T sections (or H sections are equally applicable) to a number of similar loads $Z_L$. The circuit of fig. 1a can be simplified to that of fig. 1b to obtain a solution to the problem. It can be seen that all but one of the terminations have been "folded" into the shunt impedance Z′, and the simple T section shown results.

The T section of fig. 1b can now be solved in the usual manner and $Z_1$, $Z_2$ and Z′ determined. Now it can be readily seen that

$$Z' = \frac{Z_3 \dfrac{Z_2 + Z_L}{n-1}}{Z_3 + \dfrac{Z_2 + Z_L}{n-1}}$$

where n is the number of similar terminations.

From this expression it immediately follows that

$$Z_3 = \frac{Z'}{1 - \dfrac{Z'(n-1)}{Z_2 + Z_L}}$$

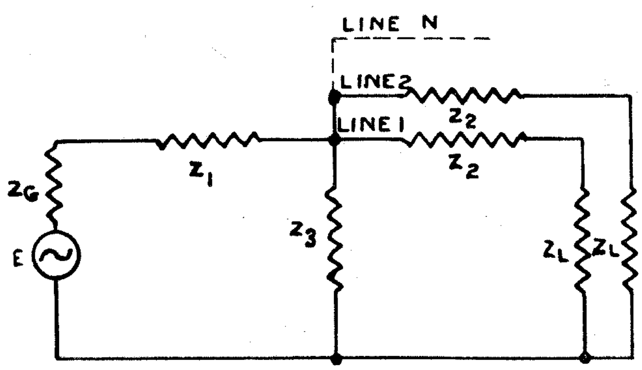



FIG. 1 (a)

and since Z′ and $Z_2$ are determined from the solution of the equivalent T section, and $Z_L$ is known, the solution for the matching sections is complete. The method can be extended to the case where the terminations are not all equal, but this involves a considerable amount of algebraic manipulation.

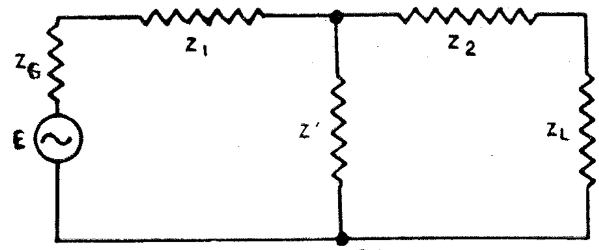



FIG. 1(b)

As an example of the use of the method, suppose an arrangement of T sections is to be constructed of pure resistances to match a generator being an internal resistance of 600 ohms to three lines each having an input resistance of 600 ohms. The total power loss is to be 4 db per section.

From the equivalent T section of fig. 2b which represents the three circuits of fig. 2a,

$$\text{attentuation} = \Theta = 0.115 \times \text{db}$$
$$= 0.115 \times 4 \times 3$$
$$= 1.38 \text{ nepers.}$$

$$Z' = \frac{\sqrt{Z_g\, Z_L}}{\sinh \Theta}$$
$$= \frac{\sqrt{600 \times 600}}{\sinh 1.38}$$
$$= \frac{600}{1.86}$$
$$= 323 \text{ ohms.}$$

$$Z_1 = \frac{Z_g}{\tanh \Theta} - Z'$$

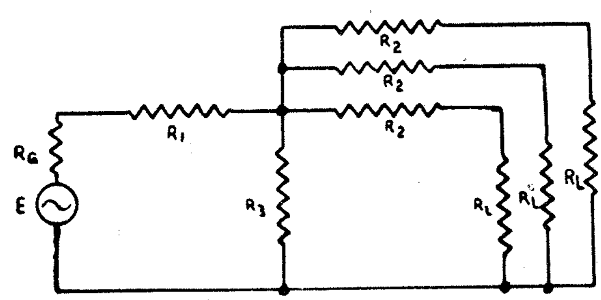



FIG. 2 (a)

$$Z_1 = \frac{600}{\tanh 1.38} - 323$$
$$= \frac{600}{0.881} - 323$$
$$= 680 - 323$$
$$= 357 \text{ ohms.}$$

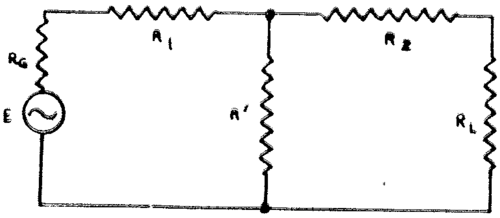



FIG. 2 (b)

$$Z_2 = \frac{Z_L}{\tanh \Theta} - Z'$$

and since the T section is clearly symmetrical, in this particular case,

$$Z_2 = Z_1 = 357 \text{ ohms.}$$

It now remains to determine $Z_3$.

$$Z_3 = \frac{Z'}{1 - \dfrac{Z'\,(n-1)}{Z_2 + Z_L}}$$
$$= \frac{323}{1 - \dfrac{323\,(3-1)}{357 + 600}}$$
$$= \frac{323}{1 - \dfrac{646}{957}}$$
$$= \frac{323}{0.326}$$
$$= 1{,}021 \text{ ohms.}$$

The required network is then made up of the resistances:

$$Z_1 = R_1 = 357 \text{ ohms.}$$
$$Z_2 = R_2 = 357 \text{ ohms.}$$
$$Z_3 = R_3 = 1{,}021 \text{ ohms.}$$

It is necessary to point out that the minimum value of attenuation chosen for the sections must be determined, as otherwise the circuit elements may have negative values. The application to H sections merely means dividing the series arms into two equal half sections. The method could easily be extended, as one typical example, to the design of constant impedance faders, where a typical application would be matching the output of an amplifier to a number of loud speakers. Also, as previously mentioned, the terminations need not be the same in all cases, and where the number of terminations required is not large the amount of algebraic manipulation required is not excessive.

## CHOICE OF SEPARATE HEATER WINDING FOR TYPE 6X5-GT

Type 6X5-GT has a maximum d-c heater-cathode potential rating of 450 volts, so that normally this type may be used quite safely with the heater operated from the same transformer winding as that used for the other valves in the receiver or amplifier. This condition is the same as that which occurs in the case of vibrator operation, when the heater is operated from a common 6 volt accumulator.

It should be noted, however, that in spite of the fact that all valves are carefully tested after manufacture for heater-cathode leakage, and insulation breakdown, it is a matter of experience that a very small percentage of indirectly heated valves do occasionally develop a short between heater and cathode during life. Due, in general, to a chip or fracture in the insulation of the heater. The occurrence of such a short circuit in an indirectly heated rectifier may have serious results.

These remarks should not be taken as being in any way derogatory of the performance of Radiotron type 6X5-GT, which has a much thicker cathode insulation than other types of valves, and is normally quite capable of withstanding a voltage of 450 volts indefinitely.

## GRID BLOCKING

Grid blocking is a phenomenon which occurs under certain conditions and results in a valve becoming completely inoperative but with a heavy constant plate current. The effect is caused through the application of a fairly high positive voltage to the grid which is unable to return to its normal voltage owing to a "kink" in the grid characteristic caused by secondary emission. It may be discovered either by looking for heating on the plate or excessive bulb temperature, or preferably, by the use of a milliammeter to measure the plate current.

This effect may be avoided by preventing any high positive voltages from reaching the grid and it may be cleared temporarily either by switching off the set and switching it on again, or by shortening the grid resistor.

The effect becomes less serious and will be eliminated entirely by decreasing the total resistance in the grid return circuit, but in some cases the value of grid resistance to achieve this result may be too low to give satisfactory operation in the receiver. In such cases it may be necessary to take precautions so as to limit the maximum positive voltage applied to the grid.

# Residual Volume Effect

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

In many radio receivers which use a diode-triode or diode-pentode as a combined detector and audio frequency voltage amplifier, trouble is often experienced with signal output when the volume control is set to its zero position. The effect is quite distinct from the minimum volume usually experienced with receivers employing reflex amplifiers. The causes of the effect can be traced to capacitive and electronic coupling between the diode circuit and the a-f voltage amplifier. The transfer may occur at either audio or intermediate frequencies, or both, and is practically independent of the volume control setting, but usually becomes apparent at the minimum setting of the control.

A complete cure for the trouble can normally only be accomplished by combining the diode with the i-f amplifier stage, and using a separate valve in the audio voltage amplifier circuit. This arrangement is also particularly convenient in so far as it lends itself more readily to various types of feedback circuits, such as those shown in the RC52 and RC41 receiver circuits published in Radiotronics 118 and 119.

To keep the residual volume, or "play-through", as it is sometimes called, to a minimum, it is essential to keep stray capacitive coupling between the diode circuit and the voltage amplifier plate circuit to as low a value as possible. This necessitates care in circuit layout and wiring. Adequate by-passing of the cathode circuit is also necessary when cathode bias is used. The effects of incomplete by-passing can readily be demonstrated by reducing the value of the by-pass condenser and noting the increase in residual volume that occurs. Earthing the grid of the voltage amplifier will usually be found to have negligible effect on the signal output for minimum volume control setting. An improvement is often effected by using diode-pentodes, instead of diode-triodes, as the action of the screen grid, when adequately by-passed, can be utilized to assist in keeping the diode plate coupling to a minimum. In this regard series screen feed enables better by-passing to be more simply achieved than if a voltage divider arrangement is used.

If the receiver uses simple a.v.c., considerable improvement can be obtained by employing only one diode of the duo-diode section for combined detection and a.v.c.; the second diode should be earthed, and it is necessary that this earthed diode be the one whose socket connection lies between the a-f voltage amplifier plate and the detector diode. Even disconnecting the diode, without earthing gives a big improvement.

Various types of sockets have also larger effects than others, as might be expected, and resistive and capacitive effects both play a part, but usually the variation due to this cause is not very great, except where sockets are of particularly poor quality.

Neutralization of the effect has been attempted, but this is usually unsatisfactory, as the cause of the trouble is complex, and the circuits required are undesirable from a cost and manufacturing point of view.

Electronic coupling in the valve itself can occur, but every precaution is normally taken in valve design to minimize coupling between the diode and pentode, or triode, sections. Direct inter-electrode capacitances are also kept as small as is practicable.

To summarize, it appears that the best solution to overcome residual volume effect is to use a separate a-f voltage amplifier, and to combine the diode detector circuit in the valve used in the i-f amplifier stage. This arrangement can, however, lead to other troubles, due to feedback at i.f., between the diode circuit and the signal-grid of the voltage amplifier. This feedback can cause either regeneration or degeneration, and neutralization between the two circuits often becomes necessary. (Suitable methods for neutralizing this effect are described in Radiotronics 118.) As a compromise, careful layout and wiring of a diode-pentode, or diode-triode, combined detector and a-f amplifier will assist in keeping the residual volume to small proportions; the diode-pentode arrangement, particularly with series screen feed, is somewhat preferable to the diode-triode in this regard.

# Peak and Average Currents in Class "B" Amplifiers

In Class "B" Amplifiers it is often required to find the relationship between peak and average currents. If it is assumed that the valve output is distortionless, that the input signal is a pure sine wave, and that the valves are biased exactly to cut-off, the peak current drawn by each of the two valves will be $\pi/2$ times the average current drawn from the power supply as indicated by a d.c. milliameter.

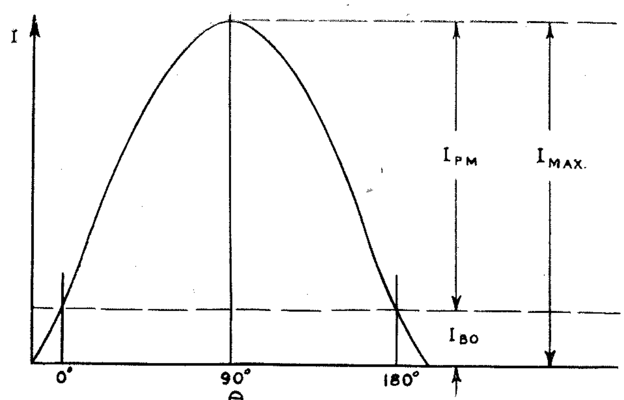



FIG. 1.

If there is a standing quiescent plate current when no signal is applied to the grid, an interesting and simple approximation has been suggested by K. R. Sturley.*

Figure 1 shows the current: time relationship over slightly more than one-half cycle, during which each valve is conducting. The conducting areas below 0° and above 180° are included in this approximation, which is therefore only satisfactory for use when the quiescent plate current is low.

Let $I_{max}$ = peak plate current,
$I_{bo}$ = quiescent plate current (one valve),
$I_{pm}$ = maximum value of varying component of plate current.
and $I_b$ = average value of plate current (both valves) over the complete cycle.

Examination of the figure shows immediately that

$$I_{max} = I_{pm} + I_{bo} \quad \text{. . . . . . . . . . . . . . . . . . .} \quad (1)$$

The average plate current is given by

$$I_b = 0.637 I_{pm} + I_{bo} \text{ approx.} \quad \text{. . . . . . . . . . . .} \quad (2)$$

and it will be seen that if the quiescent plate current is zero this is reduced to

$$I_b = 0.637 I_{pm} \text{ approx.,}$$

which is given on page 291 of the Radiotron Designer's Handbook.

The maximum value of the varying component of plate current is given by

$$I_{pm} = 1.57\ (I_b - I_{bo}) \text{ approx.} \quad \text{. . . . . . . . . . . .} \quad (3)$$

The peak plate current is therefore given by

$$I_{max} = 1.57 I_{bo} - 0.57 I_b \text{ approx.} \quad \text{. . . . . . . . .} \quad (4)$$

It will therefore be seen that the peak value of plate current may be calculated approximately by the use of Equation (4), which only requires knowledge of the average plate current under maximum signal conditions, and of the quiescent plate current.

* *"Peak pulse currents in Class B Amplifiers", Wireless Engineer* 23.277 (*October,* 1946) *page* 286.

# Control of Phase Shift in Cathode Bias Circuits

*By B. Sandel, A.S.T.C.*

It is sometimes necessary, in the design of voltage amplifier circuits, to set the amount of phase shift introduced by the cathode bias circuit to some predetermined value. A typical example occurs in the design of video amplifiers, where it may be desirable to limit the phase shift between the input voltage and the grid-to-cathode voltage to a small value at low audio frequencies, without at the same time introducing appreciable degeneration. The results of the analysis given below, allow any desired value of phase shift to be set by the cathode circuit components. When the value of phase shift is small it is also shown that degeneration due to the cathode circuit is negligible.

From an investigation of various types of cathode bias circuits, the arrangement shown in figure 1 has been found very suitable where the requirements of small phase shift, between the signal voltage $E_s$ and the grid-to-cathode voltage $E_i$, and negligible degeneration are to be fulfilled.

Examination of the circuit of figure 1 shows that the correct value of grid bias can be obtained by the voltage drop across the resistor R. The complete circuit is one that is well known, and the arrangement has often been used to minimize degenerative effects due to the cathode bias circuit.

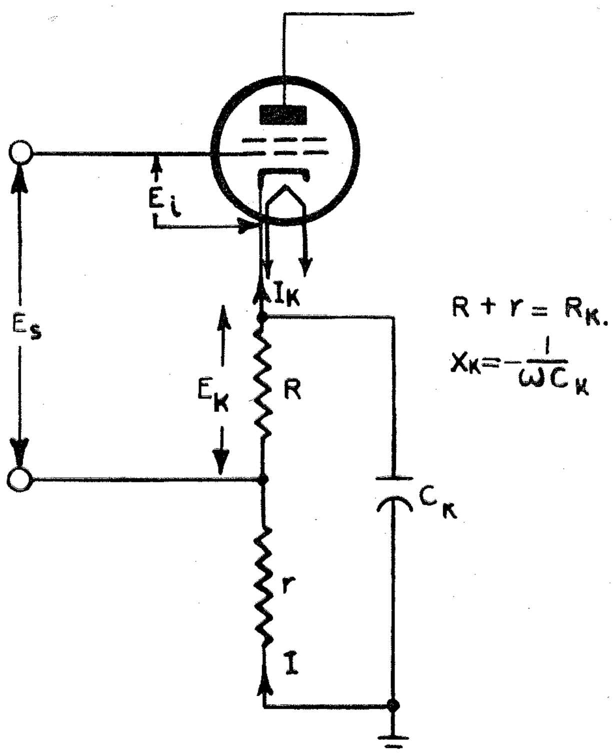



FIG. 1.

From the circuit.

$$I = \frac{jX_k . I_k}{(R + r) + jX_k}$$

$$= I_k \left[ \frac{jX_k \left[(R + r) - jX_k\right]}{(R + r)^2 + X_k^2} \right]$$

also

$$E_i = E_s - E_k$$
$$= E_s - IR$$

$$= E_s - I_k R \left[ \frac{jX_k \left[(R + r) - jX_k\right]}{(R + r)^2 + X_k^2} \right]$$

$$= E_s - I_k R \frac{X^2_k + jX_k (R + r)}{(R + r)^2 + X_k^2}$$

but

$$I_k = g_k . E_i$$

where $g_k = g_m \left(1 + \frac{I_{c2}}{I_p}\right)$ in the usual case where the screen grid is dynamically connected to earth through a by-pass condenser. If the screen is dynamically connected to cathode, instead of to earth, $g_m$ is used in place of $g_k$ in all the equations.

$$\therefore E_i = E_s - g_k E_i R \left[ \frac{X_k^2 + jX_k (R + r)}{(R + r)^2 + X_k^2} \right]$$

and

$$\frac{E_s}{E_i} = \frac{(R+r)^2 + X_k^2 + g_k R \left[X_k^2 + jX_k (R+r)\right]}{(R+r)^2 + X_k^2}$$

$$= \frac{(R+r)^2 + X_k^2 (1 + g_k R) + jg_k X_k R (R+r)}{(R+r)^2 + X_k^2}$$

Write

$A = (R + r)^2 + X_k^2 (1 + g_k R)$
$B = g_k X_k R (r + R).$

Then

$$\frac{E_s}{E_i} = \frac{A + jB}{(R + r)^2 + X_k^2}$$

and

$$\frac{E_i}{E_s} = \frac{(R + r)^2 + X_k^2}{A + jB}$$

converting to polar form

$$\frac{E_i}{E_s} = \frac{(R + r)^2 + X_k^2}{\sqrt{A^2 + B^2}} \, \Theta$$

where

$$\tan \Theta = \frac{-B}{A}$$

so that the phase shift is

$$\Theta = \tan^{-1} \frac{-B}{A}$$

and the degeneration is given by

$$\left| \frac{E_i}{E_s} \right| = \frac{(R + r)^2 + X_k^2}{\sqrt{A^2 + B^2}}$$

In many practical circuits $(R+r)^2 \gg X_k^2 (1+g_k R)$

so that $A = (R + r)^2$ approx.

and $B = g_k X_k R (r + R)$ as before.

The equation for $\Theta$ can now be rewritten as

$$\Theta = \tan^{-1} \frac{-g_k X_k R (r + R)}{(R + r)^2} \text{ approx.}$$

$$= \tan^{-1} \frac{-g_k X_k R}{(R + r)} \text{ approx.}$$

and

$$\left| \frac{E_i}{E_s} \right| = \frac{(R + r)^2}{\sqrt{(R+r)^4 + g_k^2 X_k^2 R^2 (r+R)^2}} \text{ approx.}$$

$$= \frac{R + r}{\sqrt{(R+r)^2 + g_k^2 X_k^2 R^2}} \text{ approx.}$$

It now remains to show how the results may be used. As an example, suppose a type 6AC7/1852 valve is to be used in the voltage amplifier circuit and the phase shift, between $E_i$ and $E_s$, introduced by the cathode circuit is to be limited to 1° at 20 c.p.s. The operating conditions for the type 6AC7/1852 are as listed on the valve data sheet, with 300 volts on the plate, and a plate current of 10mA; screen voltage supply 300 volts through a 60,000 ohm dropping resistor, and a screen current of 2.5ma. The screen is by-passed to ground. A cathode bias resistor of 160 ohms is recommended. Under these conditions, the $g_m$ of the valve is given as 9000 $\mu$mhos.

$$R = 160\Omega.$$

$$g_k = g_m\left(1 + \frac{I_{c2}}{I_p}\right)$$

$$= 9000\left(1 + \frac{2.5}{10}\right)$$

$$= 9000 \times 1.25$$

$$= 11{,}250\ \mu\text{mhos.}$$

In the expression for $\Theta$ it is seen that there are two unknowns, $X_k$ and r. It is necessary to choose the value of $C_k$ or r, and then to calculate the other component. Suppose we select $C_k$ at some commercially available value, and then calculate r for the required phase shift, as this appears to be the most economical procedure. Several values for $C_k$ may be chosen, and the most suitable values for r and $C_k$ selected. Attenuation should then be checked at the lowest frequency of operation, 20 c/s in our example.

Take $C_k = 200\mu F$, then

$$X_k = -\frac{1}{\omega C_k} = \frac{-10^6}{2\pi \times 20 \times 200} = -39.8 \text{ ohms}$$

then from $\tan\Theta = \dfrac{-g_k X_k R}{R + r}$

we have

$$r = \frac{-R[g_k X_k + \tan\Theta]}{\tan\Theta}$$

$$= \frac{-160[11{,}250 \times 10^{-6}\ (-39.8) + 0.0175]}{0.0175}$$

$$= \frac{160 \times 0.43}{0.0175}$$

$$= 3{,}930 \text{ ohms.}$$

A lower value of resistance r may be desirable, as the d.c. voltage drop across the cathode circuit is

$$E = (3930 + 160)\ 12.5 \times 10^{-3}$$

$$= 51 \text{ volts.}$$

A larger value for $C_k$ will allow the value of r to be reduced to a smaller value if required. If the value for r is satisfactory, it now remains to check the effects of degeneration at 20 c/s.

Then

$$\left|\frac{E_i}{E_s}\right| = \frac{R + r}{\sqrt{(R+r)^2 + g_k^2\, X_k^2 R^2}}$$

$$= \frac{160 + 3930}{\sqrt{(160+3930)^2 + (11250 \times 10^{-6})^2 (-39.8)^2 \times 160^2}}$$

$$= 1 \text{ approx.}$$

So that the degeneration is negligible with the component values selected, as would be expected by inspection of the circuit arrangement.

# RCA Application Note

## DESIGN PRECAUTION FOR OSCILLATORS EMPLOYING FILAMENT-TYPE TUBES

Experience with filament-type acorn tubes as oscillators in transmitting equipment has shown that, under some conditions of operation, oscillation may continue after the filament voltage has been removed unless the plate voltage is also removed. When the filament voltage is removed from an oscillator tube having particularly low filament power consumption, continued oscillation frequently takes place because of continued heating of the filament by the plate current.

Continued oscillation has been found most likely to occur (1) with a tube having high emission capability, (2) with an exceptionally well-designed circuit, and (3) with a high value of oscillator plate current; it has been observed with oscillator tubes operated at moderate values of plate voltage and current.

Because of these results in the laboratory and in the field, it is recommended that both the filament voltage and the plate voltage of filament-type miniature, GT, and acorn oscillator tubes used for transmitter purposes be removed when equipment employing these types is "shut down." Usually, a convenient method is to break the minus filament and the minus plate supplies with a single, double-pole switch.

The recommended procedure insures that the oscillator will always stop functioning in the "off" position, saves B power, and avoids interference with reception in combined transmit-receive equipment.

# RCA Application Note

## A DISCUSSION OF NOISE IN PORTABLE RECEIVERS

The trend toward improved performance of portable dry-battery-operated receivers has led to considerations of whether a tube additional to the normal 4-tube complement for such receivers can be used advantageously.

At the outset, it is self-evident that if increased gain is the only requirement for improved performance, the fifth tube should be added as an i-f stage. However, maximum gain may not be the only consideration, since the frequency converter stage when followed by high i-f gain may give rise to intermittent noises not predictable by tube and circuit theory. Using the additional tube as an r-f stage avoids this difficulty and is often preferable design practice for portable receivers utilizing dry-battery tubes. For this reason, this Note considers only the use of the fifth tube in an r-f stage.

On the basis that the fifth tube is a 1T4 used in a resistance-coupled r-f stage having a load resistance of 10000 ohms and a total shunt capacitance of 20 $\mu\mu$f, calculated gain values for the r-f stage are as follows:

| Frequency | Gain per 1000 micromhos |
|---|---|
| 600 kc | 8.0 |
| 1000 | 6.2 |
| 1500 | 4.7 |

While higher values of gain may be obtained by the use of the 1T4 in the more expensive tuned r-f stage, a complete appraisal of the value of an r-f stage must include the matter of noise. However, before considering the noise, let us examine further the r-f gain on the basis of electrode current requirements.

Fig. 1 shows the effect of screen voltage on plate current, screen current, cathode current, and transconductance, for the 1T4. It will be noted that while the cathode current is nearly 5 milliamperes for a screen voltage of 67.5 volts, it is only 1.3 milliamperes for 30 volts, and that the corresponding values of transconductance are 900 and 600 micromhos. From the standpoint of minimizing B-battery current, the 30-volt condition is preferable. However, this condition of operation results in a moderate decrease in transconductance. Thus, for an untuned r-f stage at 600 kc, the gain is 8.0 × 0.6, or 4.8. At higher frequencies, the gains are proportionately less. Since the capacitance in shunt with the load resistor can not be reduced, it is apparent that resort must be had to tuned circuits to obtain improved gain per stage. As an illustration, 1T4 used in a tuned r-f stage gave measured gain values for a transconductance of 600 micromhos as follows:—

| Frequency | Stage Gain |
|---|---|
| 600 kc | 12.5 |
| 1000 | 9.3 |
| 1500 | 6.0 |

In this stage, the plate circuit was tuned to 500 kc in order to increase the gain at the lower frequencies. This procedure tends to compensate for the decreased efficiency of a loop antenna at the lower frequencies.

TYPICAL CHARACTERISTICS

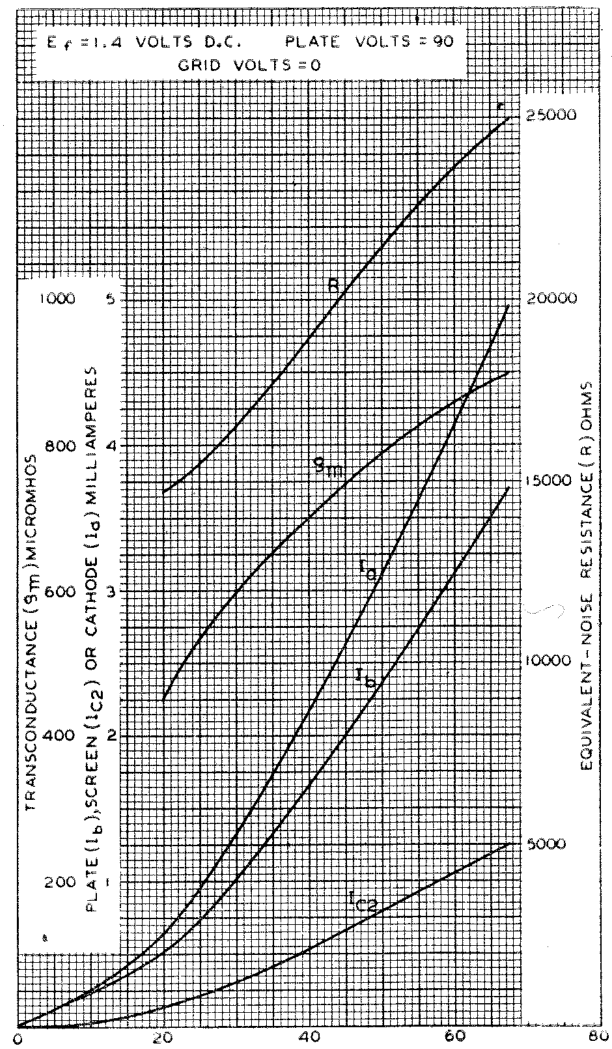



FIG. 1

## Receiver Noise Measurements.

A convenient method of determining the relative noise of receivers is to compare their equivalent noise side-band input (ensi) values. This method is used in the following discussion to compare two portable receivers, both having tuned loops and 1R5 converter stages, but one having a 1T4 tuned r-f stage. Although the following procedure is not based entirely on theoretical considerations, the equations (1), (2) and (3) are directly useful.

The thermal-agitation-noise voltage (rms) developed by a resistor is

$$2\sqrt{KTRf} \qquad (1)$$

where K is Boltzman's constant of $1.372 \times 10^{-23}$ joules per degree Kelvin
T is temperature in degrees Kelvin (usually 300°K)
R is resistance (ohms) when it, as usual, is substantially constant over a narrow bandwidth of frequencies
f is bandwidth in cycles per second.

A tuned loop is assumed to develop thermal agitation noise as though it were a resistor having a pure resistance equal to its resonant impedance. This assumption does not lead to any serious error where the bandwidth of the receiver is not greatly reduced by the selectivity of the tuned loop. A tuned loop known to be typical for portable receivers, gave values of tuned impedance and Q as follows:—

| Frequency | Tuned Impedance of Loop Circuit | Q |
|---|---|---|
| 600 kc | 53000 ohms | 84 |
| 1000 | 88000 | 79 |
| 1500 | 116000 | 61 |

The noise developed in a pentode (such as the 1T4) may be dealt with as though it were produced by a voltage impressed on the grid circuit of the tube. This noise-equivalent voltage, in turn, may be dealt with as though it were the thermal-agitation-noise-voltage of a resistor. The value of this resistor is the noise-equivalent resistance and, for a pentode, may be readily calculated from the following expression.

Pentode noise-equivalent resistance =

$$\frac{2.5}{g_m} + 19.2\,\frac{I_b . I_{c^2}}{I_a . g_m{}^2} \qquad (2)$$

where

$g_m$ is transconductance in mhos
$I_b$ is plate current in amperes
$I_{c^2}$ is screen current in amperes
$I_a$ is cathode current in amperes

An immediate advantage of the equivalent-resistance method is that the equivalent-noise resistance of a tube, the 1T4 for example, may be simply added to the noise-equivalent resistance of a tuned loop (or other input circuit), to determine the total noise-equivalent resistance for the loop and tube together. However, for a receiver circuit, a third component of resistance should be included in order to account for the receiver noise developed beyond the first tube.

At the grid of the following stage — a 1R5 frequency converter in the present case — a measurement (or calculation) of the equivalent-noise voltage (when the first tube, a 1T4, is cold) results in a noise-equivalent resistance which may be directly referred to the grid of the preceding tube by dividing it by the square of the r-f stage gain. This procedure gives the third component of noise-equivalent resistance for the entire receiver.

In this manner, values of noise-equivalent resistance are found for a typical loop used with the 1T4 r-f stage, and for the loop used with a 1R5 frequency converter. If these values of noise-equivalent resistance should be found to be equal, it would be evident that the use of the r-f stage does not reduce noise. A lesser value of noise-equivalent resistance with the r-f stage means, of course, a decreased value of ensi for that receiver. Since equation (1) shows that a resistor develops noise voltage in proportion to the square root of its resistance, the ratio of ensi for the receiver with the r-f stage to the ensi for the receiver without the r-f stage is

Square root of quotient
of equivalent resistance without r-f stage (3)
by equivalent resistance with r-f stage

The equivalent-noise resistance for the 1T4 at a screen voltage of 30 volts is shown in Fig. 1 to be 16500 ohms. Various measurements on the 1R5 at the same screen voltage (30 volts) show its equivalent-noise resistance to be approximately 200,000 ohms throughout the broadcast band. The equivalent-noise resistance of the 1R5 varies with screen voltage in substantially the same manner as shown for the 1T4 in Fig. 1. While the noise of the 1R5 (or of any typical converter) is essentially the direct i-f noise, the result is as though the thermal agitation noise of a resistor (having a resistance equal to the noise-equivalent resistance of the tube) were impressed on the grid at the signal frequency.

## RECEIVER NOISE CALCULATIONS.

In order to compare the performance of a portable receiver having either a tuned or an untuned r-f stage with a receiver without an r-f stage, it follows from the earlier discussion that the comparison must include noise as well as gain. The relative noise values for the different receivers can be calculated in the following manner. The receivers considered are:

1. Receiver with 1R5 converter and no r-f stage
2. Receiver with 1R5 converter and untuned r-f stage using 1T4 with plate load of 10000 ohms shunted by 20 $\mu\mu$f.
3. Receiver with 1R5 converter and tuned r-f stage using 1T4 and having gain values as reported earlier in this Note.

The 1T4 is assumed to have a transconductance of 600 micromhos. The tuned circuit between the 1T4 and the 1R5 is assumed to have an impedance value of 50000 ohms. This latter assumption simplifies the calculation, and is permissible because a large change in this value does not materially affect the final results.

The noise-equivalent resistance ($R_o$) for the receiver without r-f stage is simply the noise-equivalent resistance (200,000 ohms) of the 1R5 plus that of the tuned loop. Hence, $R_o$ for the various frequencies is as follows:—

| Frequency | $R_o$ |
|---|---|
| 600 kc | 253000 ohms |
| 1000 | 288000 |
| 1500 | 316000 |

The noise-equivalent resistance ($R_t$) for the receiver with tuned r-f stage is the noise-equivalent resistance (16500 ohms) of the 1T4 plus that of the tuned loop, plus that (referred to the grid of the 1T4) of the 1R5 and the circuit which couples it to the 1T4. Values of $R_t$ follow:—

| Frequency | $R_t$ |
|---|---|
| 600 kc | 70000 ohms |
| 1000 | 107500 |
| 1500 | 139500 |

Equation (3) may be evaluated from the values of $R_o$ and $R_t$, although the db improvement, 10 log ($R_o/R_t$) may be obtained directly and is given below.

| Frequency | Improvement with tuned r-f stage |
|---|---|
| 600 kc | 5.6 db |
| 1000 | 4.3 |
| 1500 | 3.55 |

These values of db improvement are not subject to any substantial increase by raising the r-f gain, because the referred values of noise-equivalent resistance are already small in comparison with the combined noise-equivalent resistance of the loop and the 1T4.

For the receiver with the tuned r-f stage, the calculation was considerably simplified because the selectivity afforded by the tuned grid circuit of the frequency converter prevents any effective transmission of noise at the intermediate frequency, the image frequency, and at high frequencies which have i-f separation from harmonic frequencies of the oscillator. However, these several components of noise must be considered for the receiver with the untuned r-f stage. In aggregate, they are quite detrimental. Moreover, the values of untuned r-f stage gain given earlier in this Note are considerably less than gain values for the tuned r-f stage. For these reasons, the noise reduction afforded by an untuned r-f stage is considerably less than that which may readily be obtained with a tuned r-f stage.

The case of an untuned r-f stage is considered by means of a numerical example showing the evaluation of the noise-equivalent resistance at 100 kc. Results for 600 and 1500 kc are also given at the end of this Section.

For the assumed case of a 10000-ohm load, a 20 $\mu\mu$f shunt capacitance, and a transconductance of 600 micromhos, the following table shows the calculated r-f stage gain for the signal frequency of 1000 kc, an intermediate frequency of 455 kc, an image frequency of 1910 kc, and the higher frequencies of 2455 and 3365 kc. Inclusion of still higher frequencies would not materially affect the result.

| | I-F | Signal | Image | Higher | Frequencies |
|---|---|---|---|---|---|
| Frequency (kc) | 455 | 1000 | 1910 | 2455 | 3365 |
| Stage Gain | 5.2 | 3.73 | 2.3 | 1.85 | 1.38 |
| Relative Gain | 1.40 | 1.00 | 0.62 | 0.50 | 0.37 |

The 1T4 itself develops noise as though the thermal-agitation-noise voltage of a 16500-ohm resistor were impressed on its grid. However, a bandwidth factor must be used with this value of resistance so as to take into account the amplification of noise at all of the frequencies in the foregoing table. The 1R5 must also be considered in the process. Approximate ratios of 1R5 gain at the above frequencies to the gain at the signal frequency will be used as follows.

| | Frequency | 1R5 Relative Gain |
|---|---|---|
| I-F | 455 kc | 1.25 |
| Image | 1910 | 1.00 |
| 2nd H. Images | 2455, 3365 | 0.60 |

These ratios are typical of most frequency-converter tubes when operated near optimum excitation.

The relative gain factors applicable to noise at the five frequencies considered are the products of the factors in the two previous tables.

Their values are

| | I-F | Signal | Image | Higher | Frequencies |
|---|---|---|---|---|---|
| Frequency (kc) | 455 | 1000 | 1910 | 2455 | 3365 |
| Relative Gain | 1.75 | 1.00 | 0.62 | 0.30 | 0.22 |

The effective bandwidth applicable to the 1T4 noise is that which is applicable to the receiver as a whole, multiplied by the squares of these terms. Another way of arriving at the same result is to state that "the noise-equivalent resistance, considered to be effective at the signal frequency only, is the product of the noise-equivalent resistance of the 1T4 and the sum of the squares of these terms." This statement follows, since the thermal-agitation-noise voltage of a resistor, impressed on the 1T4 grid in a manner which would exclude noise generated outside the band centred at 1000 kc, would produce the same result. Consequently, we can introduce the idea of an equivalent resistance for the 1T4 noise in an untuned stage. When the 1T4 is used in this circuit, the components of noise-equivalent resistance are as follows:—

| Frequency | Noise-Equivalent Resistance |
|---|---|
| 455 kc | 50000 ohms |
| 1000 | 16500 |
| 1910 | 6300 |
| 2455 | 1400 |
| 3365 | 800 |

The total noise-equivalent resistance is 75000 ohms,

More than half of this total is caused by i-f amplification in the r-f stage. This may be prevented by connecting a suitable wavetrap across (or effectively across) the 10000-ohm load resistor in the plate circuit of the 1T4. When this is done, 50000 ohms is removed from the total value (75000 ohms), leaving 25000 ohms. This latter value of noise-equivalent resistance is used for the 1T4 instead of 16500 ohms when the noise-equivalent resistance for the entire receiver with untuned r-f stage is obtained.

The entire noise-equivalent resistance at the 1T4 grid includes 88000 ohms to account for the tuned loop, and 14400 ohms added to account for the 1R5 converter. The latter figure is the value referred to take care of the 200000-ohm noise-equivalent resistance of the converter. The result at the 1T4 grid is approximately 177000 ohms for a receiver without an i-f wavetrap, and 127000 ohms with a wavetrap. The results with an untuned r-f stage are compared with the results when no r-f stage is used, in the table following this paragraph. Although the use of noise-equivalent resistances has avoided any necessity for a knowledge of bandwidth, these resistance values—177000 and 127000 ohms— may be readily interpreted as actual noise voltages by means of equation (1), provided a bandwidth is given or assumed. In this matter, the noise bandwidth is twice the effective a-f bandwidth, because the r-f and i-f circuits and tubes develop noise throughout the entire range of frequencies occupied by the double side-band signal. Also, the effective a-f bandwidth is somewhat less than the apparent frequency coverage, because of the usual drooping or gradual cut-off at both the high- and the low-frequency ends of the a-f response curve. On an assumed basis of an effective a-f bandwidth of 5000 cycles per second, the noise-equivalent resistances of 177000 and 127000 ohms correspond respectively to noise voltages of 5.4 and 4.6 microvolts (rms).

| Frequency | Improvement without i-f wavetrap | Improvement with i-f wavetrap |
|---|---|---|
| 600 kc | 3.2 db | 4.5 db |
| 1000 | 2.0 | 2.8 |
| 1500 | 0.7 | 2.6 |

Although the improvement factors are less than those obtained with the tuned stage, it is important to note that the improvement at the low-frequency end is still considerable, and that this is the range in which improvement is most needed because of the loss of loop sensitivity.

A word of caution is added to assist in dealing with likely discrepancies between calculated and measured data on noise improvement. The above table gives calculated results in db. The magnitudes are small, particularly at the high-frequency end of the broadcast band. When such data are taken, an error of 1 db is probable, except where a very highly refined technique is used. Accordingly, calculated results should be considered only as a guide to results obtainable in practice.

Conclusions:

The decision as to whether an r-f stage can be used advantageously in portable receivers must give consideration to the degree of noise reduction which can be achieved by the r-f stage.

When a tuned r-f stage having only moderate gain is used, converter noise is quite unimportant and the result is a greater noise reduction than obtainable with the use of an untuned r-f stage. However, when an i-f wavetrap is used across the plate load circuit of the untuned r-f stage, the difference in noise is small.

The question of the efficacy of the untuned r-f stage deals with a border-line case, where the final decision may depend upon other considerations.

# Voltmeter Loading

An interesting note by R. E. Lafferty, on a correction formulae for voltmeter loading, was published in the June, 1946, issue of the Proceedings of the I.R.E. (U.S.A.), page 358. The note discusses a simple method for correcting the error due to current drawn by a voltmeter, when used to measure a potential difference across portion of a high impedance circuit.

The correction involves only a simple calculation, and no additional equipment is required. It can be used for either a.c. or d.c. voltages, and is accurate when applied to linear-circuits. However, it should be realized that valves are not true linear elements inasmuch as the current drawn by a particular electrode is not necessarily directly proportional to the voltage applied to it. Experiments indicate that when a valve is working with fixed bias the element resistance changes considerably with variation of applied voltage; thus the use of the method for obtaining a correction factor, given below, cannot be used directly. When the valve operates with self bias, there is a tendency for the valve electrode resistances to remain reasonably constant, and a correction factor can be used which leads to a reduced error in the voltmeter reading. The magnitude of the error in this latter case is of the order of 5 to 14 per cent., but this should be compared to the error of about 60 per cent. usually obtained when uncorrected measurements are made with a 1,000 ohms-per-volt meter.

The correction formula is based on the effect that the additional current drawn by the voltmeter through the resistance $R_M$ (fig. 1), causes the voltage measured across $R_N$ to be lower than it would be in the absence of the voltmeter. By taking voltage readings on two separate ranges, and assuming that

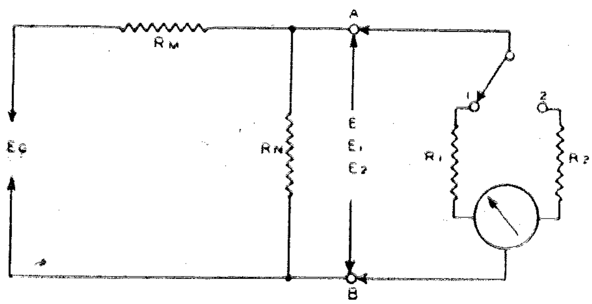


FIG. 1

the voltmeter resistance is proportional to the ranges used, the correction equation allows the unloaded voltage to be calculated. The derivation of the formula is simple and straightforward, but only the results are quoted below.

Referring to fig. 1, the true voltage between A and B is given by

$$E = \frac{(S-1)\ E_1}{S-(E_1/E_2)} \quad \ldots\ldots. \ (1)$$

where E = true voltage with voltmeter disconnected.

$E_1$ = voltage measured on highest voltmeter range that will give a reasonably accurate reading.

$E_2$ = voltage measured on the next voltmeter range lower than that used for $E_1$.

S = ratio of the two voltmeter scales used for reading $E_1$ and $E_2$.
If the voltmeter has two ranges such that

$S = \dfrac{E_1}{E_2} = 2$ then the formula

simplifies to

$$E = \frac{E_1}{2-(E_1/E_2)} \quad \ldots. \ (2)$$

Example: A voltmeter when placed across two terminals reads 110 volts on the 200 volt range. The voltmeter is switched to its 100 volt range and measures 95 volts. Then the voltage across the terminals is from (1)

$$E = \frac{(S-1)\ E_1}{S-(E_1/E_2)}$$

$$= \frac{\left(\frac{200}{100}-1\right)110}{\frac{200}{100}-\frac{110}{95}}$$

$$= \frac{110}{2-1.16}$$

$= 131$ volts

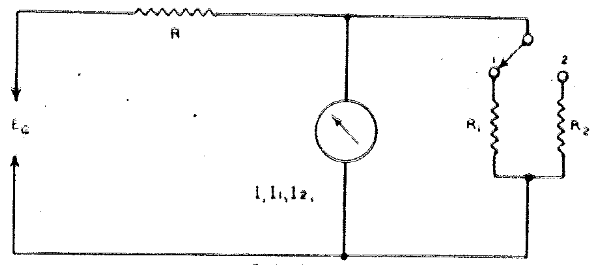


FIG 2

Although in radio work it is often not necessary to correct for meter resistance when measuring current, the following correction formula may be of some interest. The derivation follows somewhat similar lines to that for obtaining the voltmeter formula, and the reference quoted should be consulted by those particularly interested.

Referring to fig. 2

$$I = \frac{(S-1)\ I_1}{(S-I_1/I_2)} \quad \ldots. \ (3)$$

where:

I = actual current in the circuit in the absence of the ammeter.

$I_1$ = current measured on the highest range that will give an accurate reading.

$I_2$ = current measured on the next range lower than that used for $I_1$.

S = ratio of the two current scales used for reading $I_1$ and $I_2$. (Note that for S to be greater than 1 the ratio must be opposite to that used for the voltmeter).

From the above it is seen that all that is required for voltage or current correction is to make two measurements by switching from one range to another. The appropriate formula is then applied to find the true voltage drop across a circuit element. The method requires some care, particularly when used for measuring voltages on valve electrodes, and the results obtained must be interpreted with caution.

# Radiotron 8020

## Half-Wave High-Vacuum Rectifier

*(Interchangeable with Type GL-8020)*

### Electrical:

| | | |
|---|---|---|
| Filament, Thoriated Tungsten: | | |
| Voltage | 5 | Volts |
| Current | 5.5-6.5 | Amp. |
| Direct Interelectrode Capacitance: | | |
| Anode to Filament | 1.4 | $\mu\mu F$ |
| Tube Voltage Drop at 100 ma. | 200 | Volts |

### Mechanical:

| | |
|---|---|
| Mounting Position | Vertical, Base Down |
| Overall Length | $7\frac{1}{2}'' \pm \frac{1}{2}''$ |
| Maximum Diameter | $2\frac{5}{16}''$ |
| Bulb | T-18 |
| Cap | Medium |
| Base | Medium 4-Pin Bayonet |

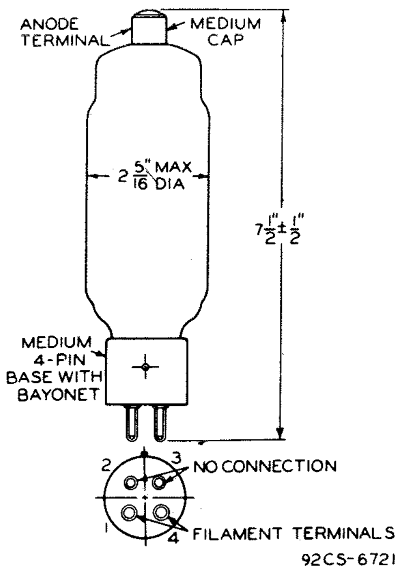
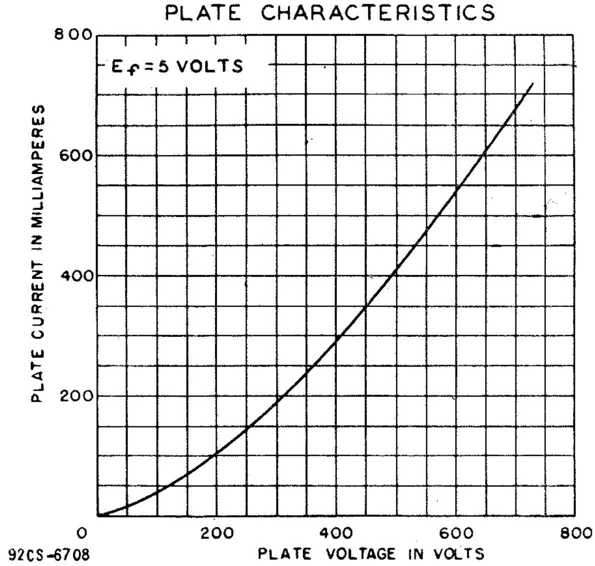



PLATE CHARACTERISTICS




## Rectifier Service

**Maximum Ratings,** *Absolute Values:*

| | | |
|---|---|---|
| PEAK INVERSE ANODE VOLTAGE | 40000 max. | Volts |
| PEAK ANODE CURRENT | 750 max. | mA. |
| AVERAGE ANODE CURRENT | 100 max. | mA. |

## Surge-Limiting Diode Service

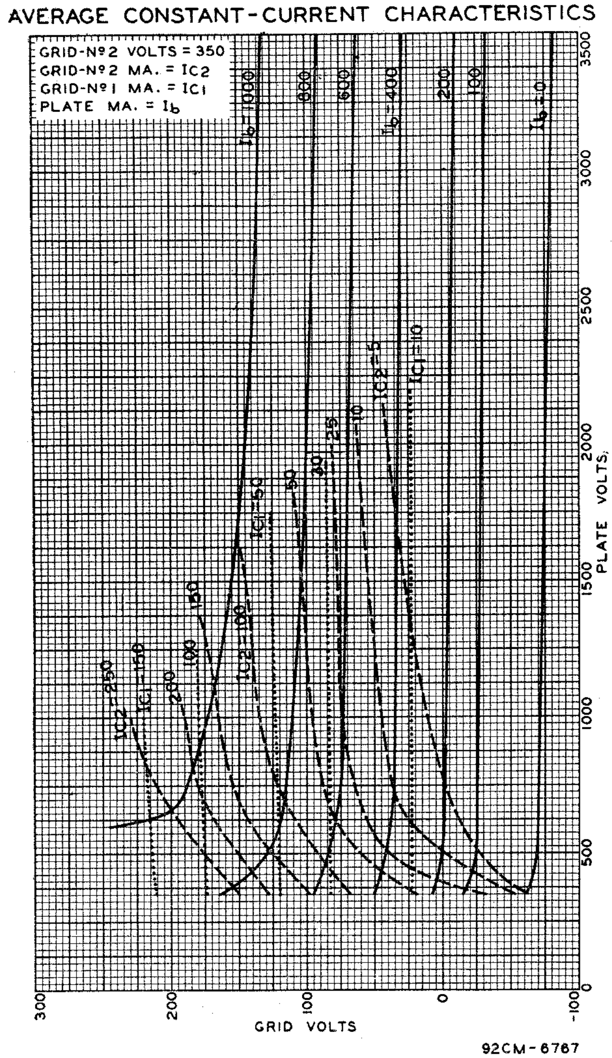
**Maximum Ratings,** *Absolute Values:*

| | | |
|---|---|---|
| FILAMENT VOLTAGE | 5.8 max. | Volts |
| PEAK FORWARD ANODE VOLTAGE | 12500 max. | Volts |
| AVERAGE ANODE DISSIPATION | 75 max. | Watts |

**Typical Operation:**

| | | |
|---|---|---|
| Filament Voltage | 5.5 | Volts |
| Peak Forward Anode Voltage | 10000 | Volts |
| Minimum Peak Anode Current | 2 | Amp |

AVERAGE CONSTANT-CURRENT CHARACTERISTICS



# RADIOTRON 6AS7-G

## Low-Mu Twin Power Triode

Radiotron 6AS7-G is an indirectly heated type of multi-unit valve containing two low-mu triode units. It is particularly suitable for use as a regulator valve in regulated d.c. power-supply units. It is also useful as a booster valve in the scanning circuit of television receivers, or in any application where a low-mu twin triode with high plate current is desired.

High load-current capability together with low valve voltage drop can be obtained by connecting the two units in parallel.

### GENERAL DATA.

**Electrical:**

| | | |
|---|---|---|
| Heater, for unipotential Cathode: | | |
| Voltage (a.c. or d.c.) | 6.3 | Volts |
| Current | 2.5 | Amperes |

**Mechanical:**

| | |
|---|---|
| Mounting Position | Any |
| Maximum Overall Length | $5\frac{5}{16}$" |
| Maximum Seated Length | $4\frac{3}{4}$" |
| Maximum Diameter | $2\frac{1}{16}$" |
| Bulb | ST-16 |
| Base | Medium Shell Octal 8-Pin |

### D.C. AMPLIFIER.

*Values are for each unit.*

**Maximum Ratings, Design-Centre Values:**

| | | |
|---|---|---|
| Plate Voltage | 250 max. | Volts |
| Plate Current | 125 max. | mA. |
| Plate Dissipation | 13 max. | Watts |
| Peak Heater—Cathode Voltage: | | |
| Heater negative with respect to cathode | 300 max. | Volts |
| Heater positive with respect to cathode | 300 max. | Volts |

**Characteristics:**

| | | |
|---|---|---|
| Plate-Supply Voltage | 135 | Volts |
| Cathode-Bias Resistor | 250 | Ohms |
| Amplification Factor | 2.1 | |
| Plate Resistance | 280 | Ohms |
| Transconductance | 7500 | Micromhos |
| Plate Current | 125 | mA. |

**Maximum Circuit Values (for maximum rated conditions):**

| | | |
|---|---|---|
| Grid-Circuit Resistance: | | |
| For Cathode-bias operation† | 1.0 max. | Megohm |

### BOOSTER SCANNING SERVICE.

*Values are for each unit.*

**Maximum Ratings, Design-Centre Values:**

| | | |
|---|---|---|
| Peak Inverse Plate Voltage* | 1700 max. | Volts |
| Plate Current | 125 max. | mA. |
| Plate Dissipation | 13 max. | Watts |
| Peak Heater—Cathode Voltage: | | |
| Heater negative with respect to cathode | 300 max. | Volts |
| Heater positive with respect to cathode | 300 max. | Volts |

**Maximum Circuit Values (for maximum rated conditions):**

| | | |
|---|---|---|
| Grid-Circuit Resistance: | | |
| For Cathode bias operation† | 1.0 max | Megohm |

† Operation with fixed bias is not recommended.

* The duty cycle of the peak inverse voltage pulse must not exceed 15% of one scanning cycle and its duration must be limited to 10 microseconds.

### APPLICATION.

It is essential that precautions be taken in equipment design to prevent subjecting the 6AS7-G to full load-current before its cathode has reached normal operating temperature. The cathode requires approximately 15 seconds to attain normal operating temperature. Unless this precaution is observed, the cathode of the 6AS7-G will be seriously damaged if not completely ruined.

### SOCKET CONNECTIONS.

Pin 1: Grid of Triode No. 2.
Pin 2: Plate of Triode No. 2.
Pin 3: Cathode of Triode No. 2.
Pin 4: Grid of Triode No. 1.
Pin 5: Plate of Triode No. 1.
Pin 6: Cathode of Triode No. 1.
Pin 7: Heater.
Pin 8: Heater.

# TYPES DISCONTINUED BY R.C.A.

**Type 1898** — monoscope using electrostatic deflection.

# NEW R.C.A. RELEASES

**Radiotron Type 1P42** — is a very small, head-on type of high-vacuum phototube designed particularly for control purposes in applications where space limitation is a prime consideration in equipment design. The head-on structure facilitates extremely close spacing of 1P42's in banks for controlling a large number of circuits. The semi-transparent cathode surface on the glass window in the large end of the 1P42 is sensitive to light sources predominating in blue radiation, and has negligible sensitivity to infra-red radiation.

**Radiotron Type 3C33** — is a heater — cathode type of power amplifier containing two high-perveance triode units, and is intended particularly for use in industrial control and voltage-regulator service. The maximum plate dissipation of each unit is 15 watts.

The type 3C33 utilizes single-ended construction and has a 12.6 volt heater. The amplification factor is 11 for each triode unit.

**Radiotron Type 12AU7** — is a twin-triode medium amplifier and is a companion type to the miniature series. The valve features a small glass envelope with integral button 9-pin base, separate terminals for each cathode, and a centre-tapped heater to permit operation from either a 6.3 or 12.6 volt supply.

Type 12AU7 has characteristics which are very similar to the larger types 6SN7-GT and 12SN7-GT.

**Radiotron Type 35B5** — is a miniature beam power amplifier of the heater-cathode type for use in the output stage of a.c./d.c. receivers. Type 35B5 is capable of providing relatively high-power output because of its high-power sensitivity and high efficiency. Within its maximum ratings, performance is equivalent to that of the larger type 35L6-GT.

# Perveance

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

A term often used in connection with vacuum tube design is perveance. A simple explanation of the meaning of the term may be of interest to those who have seen the expression in the literature, but have not understood its significance.

Most vacuum tubes are designed so that they operate with space charge limited current. In other words, the emission from the filament is always more than sufficient to supply the total current required by the rest of the electrodes in the valve. A study of a diode, made up from two parallel planes of infinite extent, operating under these conditions, was made by C. D. Child. He formulated a law which relates the plate current to the plate voltage in such a way that these two quantities are dependent only on the geometry of the valve; this law states

$$i_p = \frac{2.334 \times 10^{-6} e_p^{3/2}}{d^2} \text{ amps/cm}^2 \text{ of plate or cathode surface.}$$

where d is the distance between cathode and plate in cms.

A further study of the problem was made by I. Langmuir for the case where the electrodes in the diode are circular, and he obtained the relationship

$$i_p = 14.66 \times 10^{-6} \frac{e_p^{3/2}}{r_b \beta^2} \text{ amps./cm. length of the concentric cylinders.}$$

where $r_b$ is the radius of the plate, and $\beta$ is a function depending on the ratio of anode radius to cathode radius. This expression can be transformed to one similar to that for the parallel plane diode except that $\beta^2$ is still present in the denominator, and the distance $r_b$ is measured from the axis of the cylindrical system.

The important points to observe are that the $i_p$–$e_p$ relationship is only dependent on tube geometry, and $i_p$ is related to $e_p$ by the three halves power law in each case. We can then say, that in general, for some particular electrode geometry

$$i_p = k e_p^{3/2}$$

This equation can be rewritten as

$$k = \frac{i_p}{e_p^{3/2}}$$

which is in the form of a conductance.

The equation is often written

$$G = \frac{i_p}{e_p^{3/2}}$$

to emphasise the conductance property, and G is called the perveance.

The usefulness of the perveance is that it is constant for a fixed electrode geometry, and is independent of the electrode voltages and currents, as long as the three halves power law holds. The perveance can be used as a figure of merit and serves as a quite useful factor in valve design, not only with diodes but also for multi-electrode valves. Usually this figure of merit is determined experimentally, as the analysis of the $i_p$–$e_p$ relationship, which involves the tube geometry, becomes exceedingly difficult for any but the simplest electrode shapes. A perveance of $1 \times 10^{-6}$ means 1$\mu$A at 1 volt, 1 mA. at 100V., or 1 amp at 10,000V. For those further interested several references are listed below.

REFERENCES.

Radio Design Worksheet No. 49—Perveance—"Radio" June 1946. Y. Kusunose "Design of Triodes" Proc. I.R.E. Oct. 1929. A. L. Samuel "Design of Electron Guns" Proc. I.R.E. April 1945.