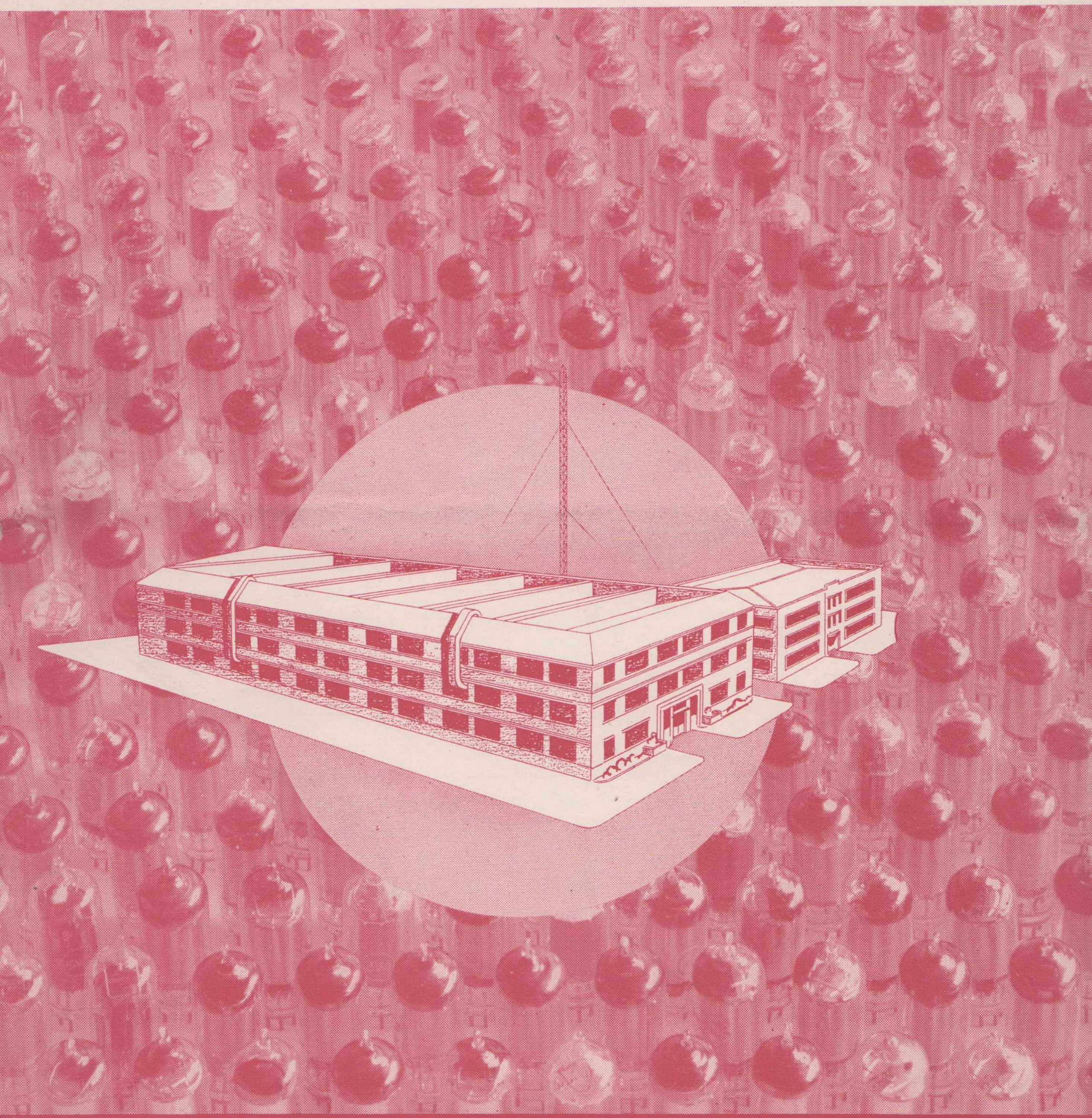
Radiotronics

Number 140 NOVEMBER — DECEMBER, 1949



A background of high-quality, top-performance, seven-pin miniature valves makes an appropriate setting for the Ashfield Works of Amalgamated Wireless Valve Company where the production of Radiotron miniatures has now passed the millionth mark.

NEW RCA RELEASES

Radiotron type 5819—is a new, head-on multiplier phototube, intended for use in scintillation counters for the detection and measurement of nuclear particle radiation, and in other applications involving low-level, large-area light sources. It has high sensitivity to blue-rich light and negligible sensitivity to infrared radiation.

An outstanding feature of the 5819 is its semi-transparent photocathode which has a diameter of 1½ inches and an area of 1.8 square inches. This relatively large cathode area permits very efficient collection of light from large-area light sources, such as are encountered in scintillation counters.

The spectral sensitivity characteristic of the 5819 covers a region in which many organic and inorganic phosphors respond efficiently to radioactive emanations.

Utilizing 10 electrostatically focused dynode stages, the 5819 operated at 90 volts per stage is capable of multiplying treble currents produced at the cathode under weak illumination by an average value of 400,000 times.

Radiotron type 5820—is a television camera tube designed for outdoor-pickup use, but it is also suitable for studio cameras. It features exceptionally high sensitivity, a spectral response approaching that of the eye, stability of performance at all incident light to a deep shadow, and a resolution capability of better than 500 lines at the centre of the picture. The photocathode in the 5820 has a response characterized by high blue sensitivity, high green sensitivity, very good yellow sensitivity, good red sensitivity, and practically no infrared sensitivity. This latter characteristic of the response prevents any color-masking by infrared light, and thus permits portrayal of colors in nearly their true tonal gradation.

DISCONTINUED RCA TYPES

Type 811—Replaced by type 811A.

Type 5731—U-H-F acorn triode, discontinued by RCA.

Type 7193—Considered obsolete by RCA, but small stocks are still held in Australia.

RCA Receiving Tube Manual

Copies of the RCA Receiving Tube Manual (RC15) are now available from the A.W. Valve Company, 47 York Street, Sydney. Price, 5/9 each.

BACK NUMBERS

The following issues of Radiotronics are no longer available: 117, 118, 121, 122, 123, 124, 125, 126, 127, 128, and 129.

Radiotronics

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Radiotron 4 Valve B/C Receiver RD33

By A. WARDALE, Member I.R.E. (U.S.A.).
(Valve Works Laboratory, Ashfield)

The introduction of Radiotron types 6AR7GT and KT61 makes possible the design of a simple "straight" 4 valve receiver having a sensitivity formerly only obtainable with the same number of valves by reflexing. The sensitivity is better than $20\mu V$ at all frequencies, which is ample for most locations.

The receiver described is designed for minimum cost, consistent with satisfactory performance. It is intended only as a mantel model with a small loud-speaker. Hum may be excessive if a large speaker is used because of the simple filtering system employed.

As the a-f amplification is low, resistance-capacitance filtering is adequate and no filter choke is required.

The plate supply voltage for all stages is taken directly from the rectifier cathode and is filtered only by the input capacitor of 24 μF.

The screen voltage for all stages is obtained from a dropping resistor of 15,000 ohms which, in conjunction with another $24\mu F$ capacitor, provides adequate filtering.

The residual hum is about 0.3mW which is quite acceptable with the types of speaker normally used in mantel receivers. At very high signal input levels ($>$ 0.3 volts at the aerial) some modulation hum becomes apparent as shown by the sudden rise in noise level in the accompanying curves. However, as signals of this order are not likely to be received, and as the hum is not less than 40 db below the output from a 30% modulated signal, this effect should not be noticed in use.

The aerial and oscillator coils and intermediate frequency transformers used are of conventional design. The aerial coil is air cored with a secondary Q of 97 at 1000 Kc/s. The oscillator coil is slug tuned with a 410 $\mu\mu F$ fixed padder.

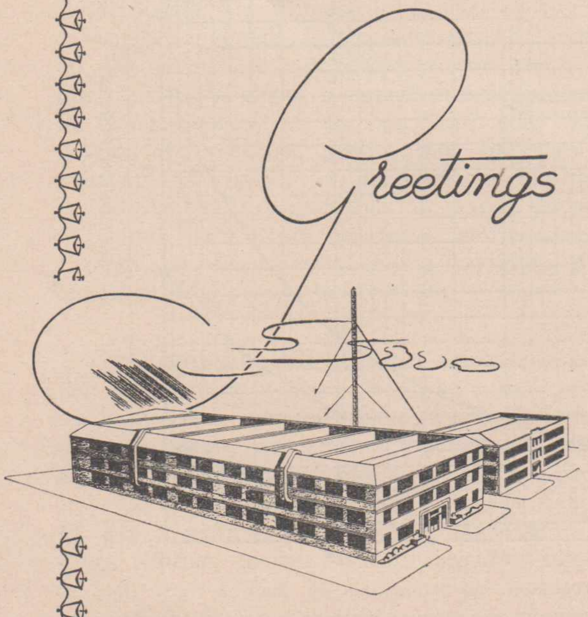
The intermediate frequency transformers are slug tuned. Each winding has an unloaded Q of 105 and is tuned with a 70 $\mu\mu F$ capacitor.

Bias is obtained from a 50 ohm back bias resistor which develops -1.8 volts at zero signal and -2.0 volts at normal signal level. The full bias is applied to the KT61 but because of the diode bias supplied through the a.v.c. network, the bias voltage at the grids of the 6AR7GT and 6A8G is -1.2 volts at zero signal, rising to -2.1 at $20\mu V$ input and progressively to -20 volts at 1 volt input.

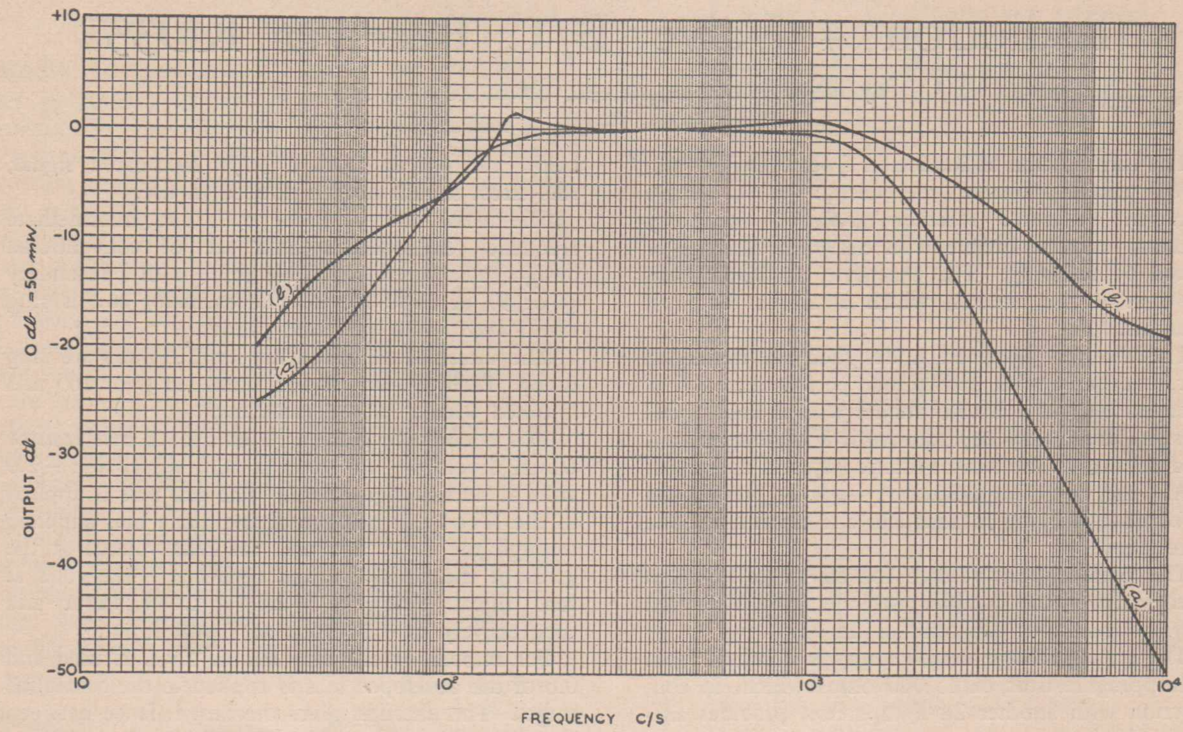
For reasons of economy simple a.v.c. is used and 0.6 of the developed bias is applied to the controlled stages. This fraction gives the best balance between signal handling capacity and small signal sensitivity.

The output stage is designed to develop 1.0W in the plate circuit of the KT61, which it does with 8% total distortion. This output is usually sufficient for normal listening and makes possible a substantial reduction in heat dissipation and current drain, allowing a cheaper power supply to be used.

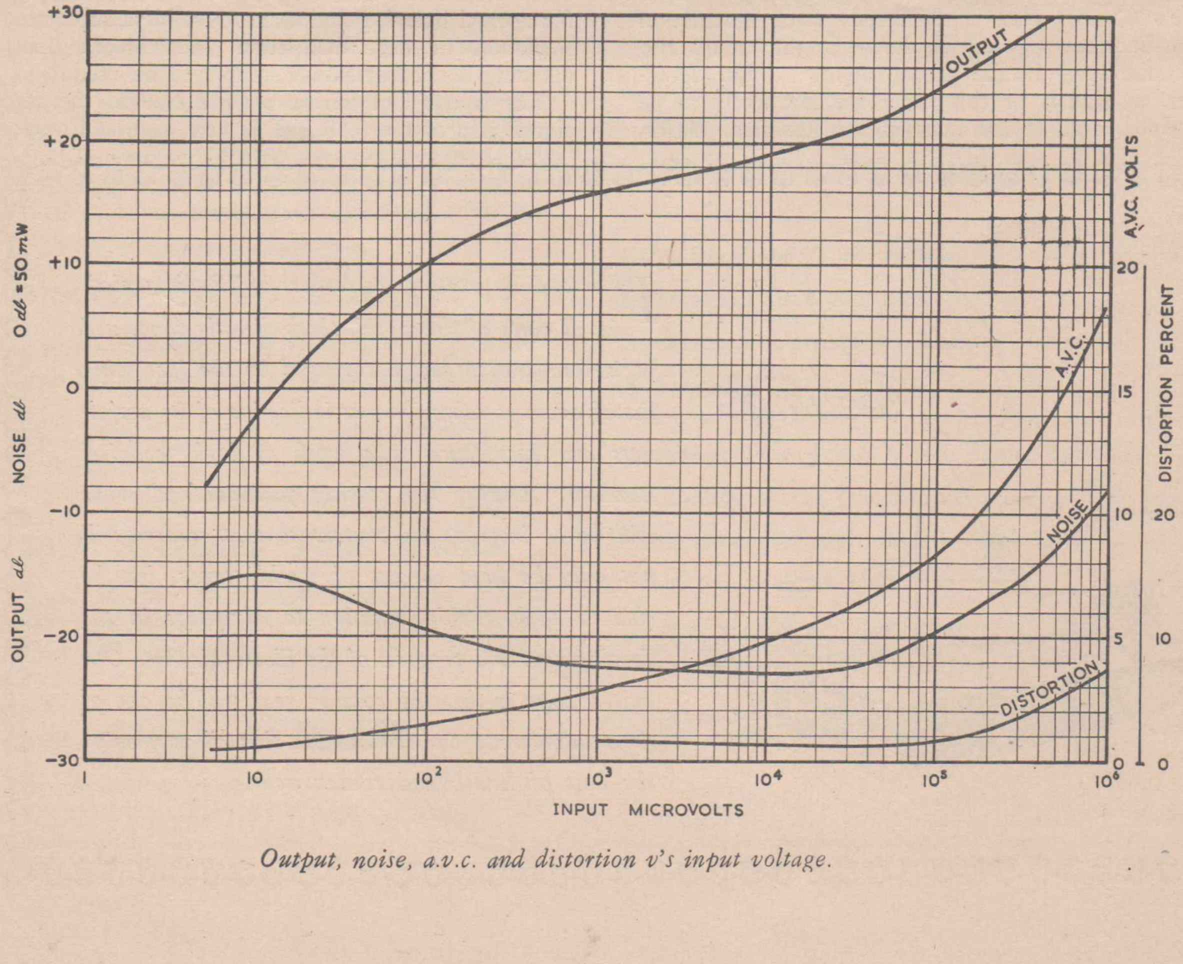
The layout and wiring require care as the zero signal gain of the 6AR7GT is very high and only a



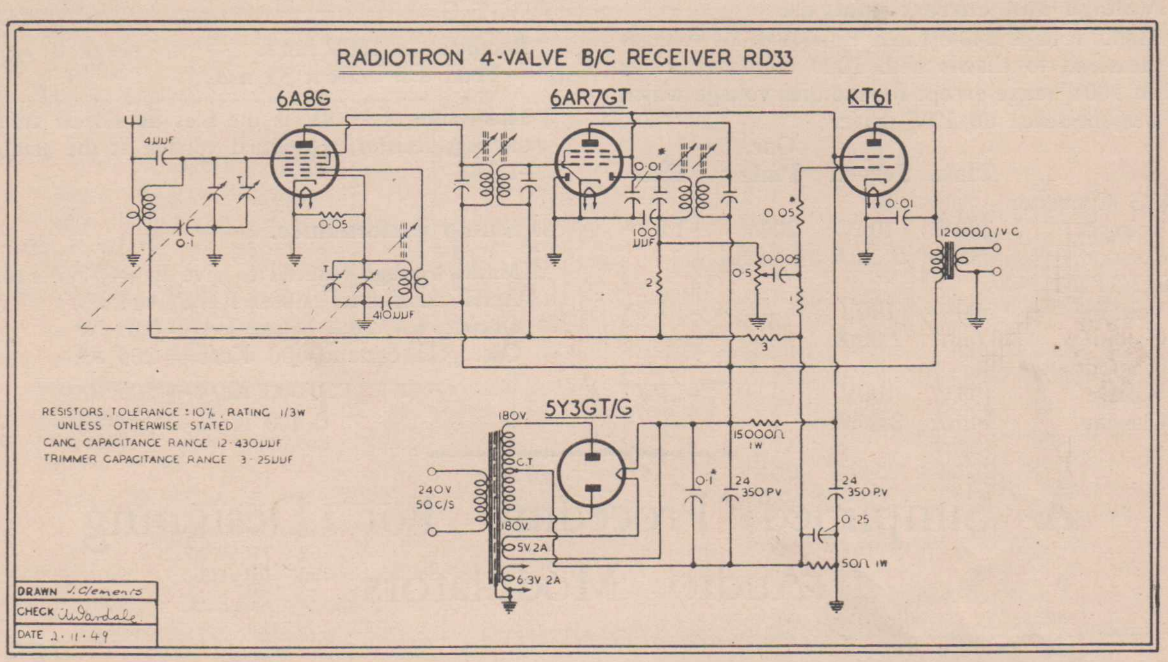
IN extending greetings to readers of "Radiotronics" in all parts of the world, those directly concerned in its preparation feel encouraged by the increasing list of overseas subscribers as indicative of its usefulness and acceptance as a work of high technical standard. During the coming year our endeavour will be to improve its context so that readers will not only be kept abreast of world trends, but be able to apply the information to advantage in the development and progress of the Electronic Art. Our united hope is that the coming year will be one of great moment to our readers and that all may enjoy the best of health and contentment.



Curves of frequency response v's output voltage measured across voice-coil. (a) Overall response; input applied through 1mW signal at 1000 Kc/s, 30% mod. (b) Audio section only input = 0.26V r.m.s.



Output, noise, a.v.c. and distortion v's input voltage.



small amount of external coupling between plate and grid circuits may cause instability.

Those components shown marked in the circuit diagram with an asterisk may, in most cases, be omitted as they are only included as precautions against instability in production and use. The developmental receiver was completely stable without these components and the performance was not noticeably affected in any way.

No oscillator grid capacitor has been used, as the padder, gang and trimmer in parallel provide the capacitance required. In the developmental receiver no difficulties were experienced but the satisfactory operation of the circuit is nevertheless dependent on the characteristics of the oscillator coil, which governs the oscillator grid current and thus the voltages and currents of the converter electrodes. A 100 µµF oscillator grid capacitor may be necessary to ensure satisfactory oscillator performance if changes are made to the operating conditions of the converter or the coil characteristics.

The curve of overall frequency response indicates that the response is down 18 db at 3 Kc/s. With small speakers, especially, the electrical response does not necessarily represent the acoustic output, and in the average receiver the normal rise in response of the speaker around 3 Kc/s is sufficient to give a sensibly linear acoustic response in the upper register. In fact the high frequency response is such that it may be necessary to increase the value of the by-pass capacitor across the speaker to achieve the mellowness of tone preferred by many listeners.

TEST RESULTS RECEIVER RD33

1. Signal voltages for 50mW output

	400 c/s	455 Kc/s	600 Kc/s	1000 Kc/s	1400 Kc/s
KT61 grid	0.26V	—	—	—	—
6AR7GT diode	—	900mV	—	—	—
6AR7GT grid	—	5mV	—	—	—
6A8 grid	—	52µV	61µV	56µV	54µV
Aerial	—	—	18µV	14µV	16µV
Image ratio		—	275	100	50
Noise sensitivity*		—	4µV	3.5µV	3.7µV
Osc. grid current		—	320µA	440µA	490µA

* Input signal (30% modulated) at which noise output is equal to the signal output.

2. Selectivity

Times down	455 Kc/s	1000 Kc/s
2 (6 db)	3.2 Kc/s	2.7 Kc/s
10 (20 db)	12	9.1
10^2 (40 db)	33	32
10^3 (60 db)	55	53

3. Overall distortion for 1.0W output

Measured with 10^3µV input to the aerial. Signal 1000 Kc/s modulated 30% at 400 c/s.

(a) With resistive load:—8% (10% for 1.12W)

(b) Secondary power for same distortion using a 12000/3.5 ohm transformer 0.8W (10% for 0.99W).

4. Modulation distortion at input of 10^3µV at 1,000 Kc/s; modulation frequency 400 c/s.

% mod.	distortion.
20	0.5%
40	2.2%
60	3.2%
80	4.6%
100	6.8%

5. Voltage and current analysis.

Mains voltage 240V r.m.s. no signal, all voltages measured to chassis with 1000 ohm/volt meter on 500V range except for the bias voltage which was measured on 10V range.

	Plate	Screen	Osc. Plate	Bias
6A8G Converter				
Voltage	200V	100V	200V	–1.8V*
Current	3.3mA	2.2mA	3.4mA	—
6AR7GT I-F				
Voltage	200V	100V	—	–1.8V*
Current	8.0mA	2.0mA	—	—
KT61 Output				
Voltage	195V	100V	—	–1.8V
Current	14mA	2.0mA	—	—
5Y3GT/G				
Rectifier	180V r.m.s.	—	—	—

Total d.c. current 35 mA.

* This value of –1.8V is the bias developed across the bias resistor, the actual voltage at the grid is –1.2V.

6. Slump performance.

Mains Voltage reduced to give $E_f = 5.5$V r.m.s.
Aerial Sensitivity 1,000 Kc/s = 17 μV
Output for 10% distortion = 0.85 W
Osc. grid current 600 Kc/s = 205 μA
10,000 Kc/s = 295 μA
1,400 Kc/s = 343 μA

A Simplified Procedure For Designing Audio Modulators

By A. G. NEKUT

Reprinted from Ham Tips by courtesy of the Radio Corporation of America.

In most amateur applications the problem of choosing a suitable audio modulator circuit is affected at the start by certain fixed conditions in the ham shack. Usually, for example, the modulator plate-supply voltage is fixed by the power supplies available. Often the modulation transformer available has an "impedance" rating that may not fit the value of plate-to-plate load resistance published under the typical operating conditions for the modulator valve desired. It is the purpose of this article to present simplified design formulae which will aid in the design of a satisfactory modulator stage.

Because efficiency and economy of operation are usually of the utmost importance, this discussion will be limited to push-pull circuits using (1) beam power valves, (2) power pentodes, or (3) power triodes operating in the positive grid region. Screen-grid type valves may be operated under either high-bias class AB_1 or class AB_2 conditions; triode types operate, of course, under high-bias class AB_2 or class "B" conditions.

Let us start off with values of d.c. plate voltage (E_{bs}) and d.c. plate current (I_{bs}) of the fully loaded class "C" r-f stage which is to be plate modulated. These values have been either computed[1] or obtained from published class C telephony operating conditions for the desired valve type.

The average audio power (W_a) in watts required to fully modulate this input power with sine-wave modulation is obtained as follows[2]:

Required average audio power W_a=

$$\frac{\text{d.c. plate voltage } E_{bs} \times \text{d.c. plate current } I_{bs}}{1.7} \qquad (1)$$

where E_{bs} is in volts and I_{bs} is in amperes.

The a.c. load resistance (R_s) in ohms presented to the modulation transformer secondary by the r-f stage is given by

$$\text{A.C. load resistance } R_s = \frac{0.85\ E_{bs}}{I_{bs}} \qquad (2)$$

Equations (1) and (2) allow for an efficiency factor chargeable to the modulation transformer and arbitrarily set at 85%. No specific allowance has been made for screen-modulation power which is usually negligible if a screen-voltage tap is available on the modulation transformer. If a dropping resistor is used to supply the screen with modulated voltage, the screen current per valve of the modulated stage should be added to I_{bs} before W_a and R_s are computed. It should be noted that satisfactory plate modulation of screen-grid valves often results if the screen is fed from an unmodulated voltage source through an audio choke or a high resistance.

Design procedure

Let us assume that the d.c. plate supply voltage for the modulator stage (E_{bb}) is fixed, and the design problem is to select suitable modulator valves and a modulation transformer to meet the conditions imposed above. The following approximate relations will be used:

For E_{bb} in range from 400 to 750 volts	For E_{bb} in range from 1250 to 3500 volts	
$I_b = \dfrac{0.75\ W_a}{E_{bb}}$	$I_b = \dfrac{0.71\ W_a}{E_{bb}}$	(3)
$W_p = 0.25\ W_a$	$W_p = 0.21\ W_a$	(4)
$W_{in} = 0.75\ W_a$	$W_{in} = 0.71\ W_a$	(5)

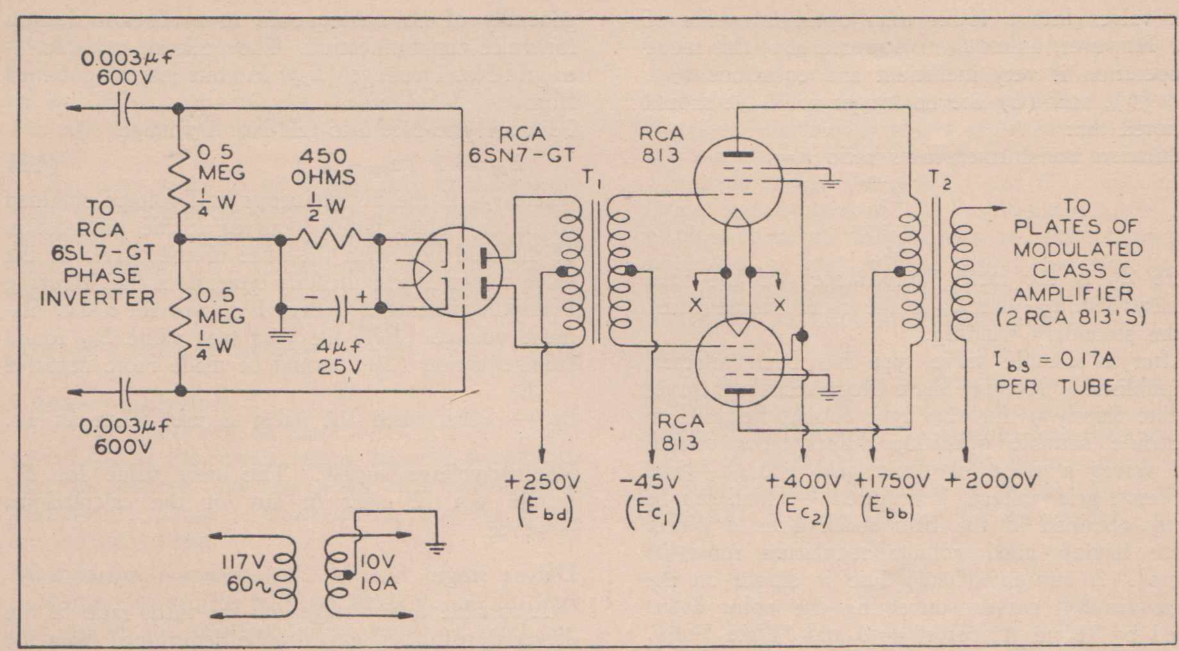


Figure 1. Modulator circuit designed from calculations given in text. It has not been built, and therefore, test data are not available.

Notes:

(1) All power supplies returned to ground.

(2) E_{c1} to be obtained from a source of good regulation (internal impedance equal to or less than 200 ohms).

(3) The 250-volt supply may be obtained from a tap on bleeder for E_{c2} supply. Minimum bleeder current should be approximately 0.05 amperes.

(4) T_1 = Driver Transformer — 5-watt audio level — Total primary to ½ secondary turns ratio = 3.

(5) T_2 = Modulation Transformer — 400-watt audio level — one-half primary to total secondary turns ratio = 0.8.

(6) E_{c2} and E_{bb} supplies should be adequately bypassed to ground for audio frequencies. Radio-frequency bypass capacitors at valve socket may be required under some conditions.

$$R_{pp} = \frac{1.3\ E_{bb}^2}{W_a} \qquad R_{pp} = \frac{1.7\ E_{bb}^2}{W_a} \quad (6)$$

$$r = \sqrt{\frac{R_{pp}}{4R_s}} \quad (7)$$

In the above relations, I_b is the max.-signal d.c. plate current per valve in amperes, W_p is the max.-signal plate dissipation per valve in watts, W_{in} is the max.-signal d.c. power input per valve in watts, and W_a is the audio power output for two valves (push-pull stage) also in watts, all for sine-wave modulation. R_{pp} is the plate-to-plate load resistance presented to the modulator valves, and r is the turns ratio of the modulation transformer defined as

Modulation transformer turns ratio r =

$$\frac{\tfrac{1}{2}\ \text{total number of primary turns}}{\text{number of secondary turns}} \quad (8)$$

It is assumed, of course, that the primary of the modulation transformer is centre tapped and that the secondary feeds the class "C" r-f stage to be plate modulated.

Modulator valve selection

Suitable modulator valves (either screen-grid or triode types) may now be selected on the basis of maximum ratings[3] for either class AB_2 or class B audio service (or class C telegraphy ratings if audio ratings are not given) that are equal to or in excess of the values found from equations (3) to (6). It is evident from inspection of equations (6) and (7) that the selection of E_{bb}, R_s, and W_a automatically fixes the modulation transformer turns ratio, r. If a transformer having a different turns ratio is already available in the ham shack it will be necessary to change either one or all of the three quantities listed in order to make use of this transformer. If the turns ratio of the available modulation transformer is lower than the value given by equation (7), it is possible to operate the modulator valves into a lower than optimum value of R_{pp}. However, unless E_{bb} is lowered also, this mode of operation is very inefficient and equations (3), (4), (5), and (6) are no longer valid. It should be noted that

Modulation transformer turns ratio $r =$

$$\sqrt{\frac{Z_p}{4Z_s}} \qquad (7a)$$

where Z_p is the rated "impedance" of the total primary winding and Z_s is the rated "impedance" of the secondary winding.

After a suitable valve type has been selected, the published "Average Plate Characteristics" curves ("plate family") for this type should be used to determine suitable operating values. For screen-grid valves a value of screen-grid voltage — and suppressor-grid voltage, if required — which can be readily obtained in the ham shack from a power source having good voltage regulation must be selected. A straight (load) line is drawn on the "plate family" curves connecting the point determined by "Plate Amperes" = 0 and "Plate Volts" = E_{bb} to the point determined by "Plate Volts" = 0 and "Plate Amperes" = I'_b where

$$I'_b = \frac{4E_{bb}}{R_{pp}} \qquad (9)$$

The optimum value of grid - No. 1 bias may now be obtained from the relation

$$\text{Optimum grid bias } E_{c1} = -\frac{(e_1 i_2 - e_2 i_1)}{(i_1 - i_2)} \qquad (10)$$

where the values of e_1 and e_2 are convenient intermediate values of grid-No. 1 voltage taken from the intersection of the load line with the bias curves, and i_1 and i_2 are the corresponding plate currents. In this equation it is assumed that the "e" and "i" points chosen lie on a linear portion of the valve's dynamic transfer characteristic and that the plate current of the non-working valve of the push-pull connection is zero. For this reason, the values of "e" and "i" chosen for equation (10) should lie well up on the load line but should not include points near the "knee" of the curve where some non-linearity may usually be expected. The plate dissipation under zero-signal conditions (W_{po}) may now be checked. Proceeding vertically upwards from E_{bb} on the "plate family" curves, read the value of plate current I_{bo} at the value of E_{c1} computed from equation (10). Then,

$$\text{Zero-signal plate dissipation } W_{po} = E_{bb}\, I_{bo} \qquad (11)$$

This value of W_{po} (zero-signal plate dissipation per valve) should not exceed approximately $\frac{1}{3}$ to $\frac{1}{2}$ of the maximum rated plate dissipation of the valve. If the value of E_{c1} found from equation (10) is not sufficiently negative to limit W_{po} to the desired value, it may be made more negative at the expense of only a slight increase in distortion at max.-signal levels; small-signal operation will produce larger amounts of distortion, but this mode of operation is generally of no consequence in modulator designs for voice communication. The peak a-f grid-No. 1-to-grid-No. 1 voltage (E_{gg}) in volts may be obtained from

Peak a-f grid-No. 1-to-grid-No. 1 voltage

$$E_{gg} = 2\ (e_{gm} - E_{c1}) \qquad (12)$$

where e_{gm} is the instantaneous grid voltage obtained from the "plate family" curves at the intersection of the load line with the knee of the curve. If the valve chosen is a filamentary type and if the "Average Plate Characteristics" curve is shown for a d.c. filament voltage (E_f), the grid bias value E_{c1} found from equation (10) should be made more negative by $\frac{E_f}{2}$ volts when the valve is used with an a.c. filament-voltage supply. This new value for E_{c1} should not be used in any of the calculations, however.

Driver stage

A suitable driver stage and the turns ratio of the driver transformer may now be determined from the following considerations. If no current is drawn by grid No. 1 of the modulator valve, any conventional resistance-capacitance-coupled push-pull or phase-inverter voltage amplifier, comprising either triodes or pentodes, capable of supplying the required value of peak a-f grid-No. 1-to-Grid-No. 1 voltage E_{gg} to the modulator circuit may be used.[4] If current is drawn by grid-No. 1 of the modulator valve, the following approximate relations are useful. For conventional low- and medium-mu triodes for the driver stage in push-pull class A or AB_1 connection

The driver transformer turns ratio $r_d =$

$$\frac{2.4\, E_{bd}}{E_{gg}} \qquad (13)$$

and

Driver valve max. allowable plate resistance

$$R_{pm} = \frac{r_d\, E_{bd}}{6.7\, i_{gm}} \qquad (14)$$

where r_d is the driver transformer turns ratio and is defined as

$$r_d = \frac{\text{total number of primary turns}}{\frac{1}{2}\text{ number of secondary turns}} \qquad (15)$$

E_{bd} is the plate supply voltage of the driver stage, i_{gm} is the instantaneous grid current drawn by grid No. 1 of the modulator valve in amperes at the value of e_{gm} used in equation (12), and R_{pm} is the maximum allowable driver-valve plate resistance in ohms. Valves with values of R_p higher than indicated by equation (14) may be used but somewhat higher distortion will result. For single-ended class A driver circuits using conventional low- and medium-mu triodes

$$r_d = \frac{1.2\, E_{bd}}{E_{gg}} \qquad (16)$$

Equations (14) and (15) also apply in this case. The power rating of the driver transformer should be adequate to handle at least the rated power output of the driver valve(s) in conventional class "A" (or AB_1 as the case may be), audio power-amplifier service.

The final value to be determined in computing valve operation is the screen-grid dissipation. Useful relations for approximating the value of average screen current (I_{c2}) in amperes and screen dissipation (W_{c2}) in watts at max.-signal levels are—

$$\text{Average screen current } I_{c2} = \frac{i_{c2m}}{4} \qquad (17)$$

Screen dissipation

$$W_{c2} = I_{c2}\ E_{c2} \qquad (18)$$

where i_{c2m} is the instantaneous value of screen current in amperes flowing when the instantaneous grid-No. 1 voltage is equal to e_{gm}, and E_{c2} is the d.c. screen voltage.

Modulation transformer

Before proceeding with an example to illustrate the use of the relations given above, a brief discussion of modulation transformer "impedance" ratings may prove useful. Modulation transformers are usually rated in terms of primary and secondary "impedance" and audio power (or more properly KVA) capability. The peak a.c. voltage (E_{pm}) that may be applied to ½ of the modulation transformer primary is

Peak a.c. voltage across primary E_{pm} =

$$\sqrt{\frac{W_t\ Z_{pm}}{2}} \qquad (19)$$

where Z_{pm} is the maximum impedance rating of the entire primary in ohms and W_t is the rated audio-power-handling capability of the transformer in watts. Similarly, the peak a.c. voltage (E_{sm}) permissible across the transformer secondary winding (equal to the d.c. plate voltage of the plate-modulated r-f amplifier for 100% modulation) may be found from

Peak a.c. voltage across secondary E_{sm} =

$$\sqrt{2\ W_t\ Z_{sm}} \qquad (20)$$

where Z_{sm} is the maximum secondary-impedance rating of the transformer. Of course, any voltage (and impedance) lower than these maximum rated values may be used. However, in order not to exceed the a.c. current ratings implied in the audio power and impedance ratings of a transformer having a fixed turns ratio, the power-handling capability of a transformer should be reduced approximately in accordance with the relation

$$W'_t = \frac{W_t\ R_s}{Z_{sm}} \qquad (21)$$

where R_s [as defined previously for equation (2)] is less than Z_{sm} and W'_t is the reduced audio-power-handling capability of the transformer. The d.c. current ratings of both primary and secondary windings are assumed to remain constant when the transformer is operated at other than rated impedance levels, although a reduction in primary d.c. current may allow some increase in a.c. current (allowing W'_t as given in equation (21) to be increased somewhat) and a reduction in secondary d.c. current may allow a slight increase in both E_{sm} (as given in equation (20)) and W'_t. For modulation transformers of the "multimatch" type it is assumed (unless information to the contrary is published by the manufacturer) that full power-handling capability has been preserved by proper design for all rated impedance values.

Example

As an example, let us assume that the class "C" r-f amplifier to be modulated is a push-pull circuit using 2 Radiotron-813's with a d.c. plate voltage (E_{bs}) of 2000 volts and a d.c. plate current (I_{bs}) of 0.17 amperes for each valve or 0.34 amperes for both. From equation (1), we obtain

Required average audio power W_a =

$$\frac{E_{bs}\ I_{bs}}{1.7} = \frac{(2000)\ 0.34}{1.7} = 400 \text{ watts}$$

From equation (2), we obtain

A.C. load resistance R_s =

$$\frac{0.85\ E_{bs}}{I_{bs}} = \frac{0.85\ (2000)}{0.340} = 5000 \text{ ohms}$$

If we assume that it is desired to operate the modulator from a 1750-volt supply, equations (3) to (5) yield

Max.-signal d.c. plate current per valve I_b =

$$\frac{0.71\ W_a}{E_{bb}} = \frac{0.71\ (400)}{1750} = 0.162 \text{ amperes}$$

Max.-signal plate dissipation per valve W_p =

$$0.21\ W_a = 0.21\ (400) = 82 \text{ watts}$$

Max.-signal d.c. power input per valve W_{in} =

$$0.71\ W_a = 0.71\ (400) = 284 \text{ watts}$$

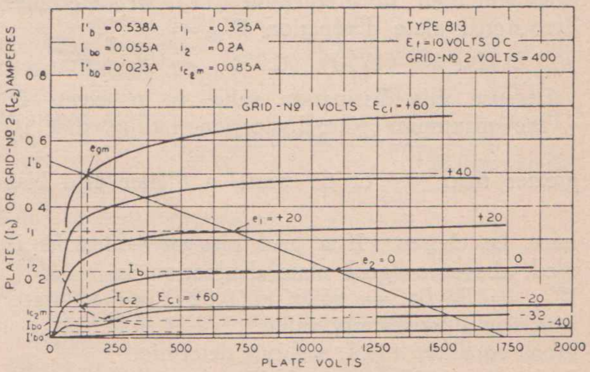


Figure 2. Average plate characteristics of the Radiotron-813.

Inspection of the maximum ratings in the technical data[5] for power valves shows that either the Radiotron-813 or the Radiotron-810 types will easily fulfill all requirements. If a 400-volt screen supply having good regulation is available, the 813 may be chosen to advantage, because this choice will ease the driver stage requirements somewhat in comparison to those required for the Radiotron-810. Equations (6) and (7) give us the required modulation-transformer impedance and turns ratio ratings.

Plate-to-plate load resistance $R_{pp} =$

$$\frac{1.7\ E_{bb}^2}{W_a} = \frac{1.7\ (1750)^2}{400} = 13{,}000 \text{ ohms}$$

Turns ratio of modulation transformer $r =$

$$\sqrt{\frac{R_{pp}}{4R_s}} = \sqrt{\frac{13{,}000}{4\ (5000)}} = 0.806$$

The load line can now be drawn on the curve of "Average Plate Characteristics" shown in Fig. 2 after I'_b is obtained by means of equation (9) as follows

$$I'_b = \frac{4E_{bb}}{R_{pp}} = \frac{4\ (1750)}{13{,}000} = 0.538 \text{ amperes}$$

From equation (10) after points e_1 and e_2 have been selected, we obtained

$$\text{Optimum grid bias } E_{c1} = -\left[\frac{e_1 i_2 - e_2 i_1}{i_1 - i_2}\right]$$

$$= -\left[\frac{20\ (0.2) - 0\ (0.325)}{0.325 - 0.2}\right]$$

$$= -\frac{4}{0.125} = -32 \text{ volts}$$

The value of I_{bo} at a grid bias of –32 volts is obtained from the family of average plate characteristics and then, from equation (11), we determine Zero-signal plate dissipation $W_{po} =$

$$E_{bb} I_{bo} = (1750)\ (0.055) = 96 \text{ watts}$$

Because this dissipation value is in excess of ½ the maximum plate-dissipation rating; that is, greater than $\frac{125}{2}$ or 63 watts, a higher grid bias must be chosen. If a grid bias of –40 volts is used, the zero-signal plate dissipation is

$W_{po} = E_{bb} I_{bo} = (1750)\ (0.023) = 40$ watts

which is a satisfactory value.

From equation (12), we can determine

Peak a-f grid-No. 1-to-grid-No. 1 voltage

$$E_{gg} = 2\ [e_{gm} - E_{c1}] =$$

$$2\ [60 - (-40)] = 200 \text{ volts}$$

For a.c. filament operation, an actual bias of –45 volts is required because the average plate characteristics were taken with a d.c. filament power supply of 10 volts.

If we assume that a push-pull driver stage having a plate supply voltage (E_{bd}) of 250 volts would be most desirable, then from equation (13) we obtain

Driver transformer turns ratio $r_d =$

$$\frac{2.4\ E_{bd}}{E_{gg}} = \frac{2.4\ (250)}{200} = 3$$

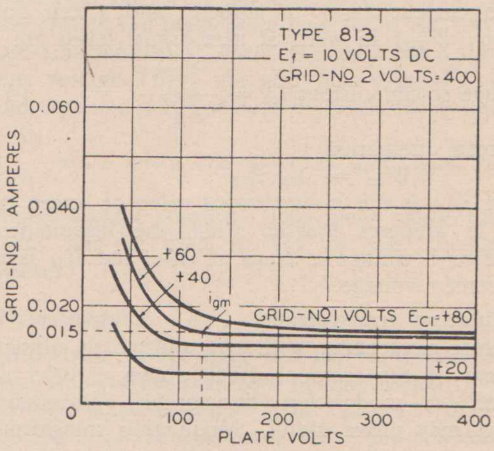


Figure 3. Grid characteristics of the Radiotron 813.

From Fig. 3, at the value of instantaneous grid-No. 1 voltage obtained from the plate family curves at the intersection of the load line with the knee of the curve, $e_{gm} = 60$ volts. At a plate voltage corresponding to the intersection of the load line and the curve of $e_{gm} = 60$, the value of instantaneous grid-No. 1 current (i_{gm}) is 0.015 amperes.

Hence, from equation (14) the maximum allowable plate resistance of the driver valve (R_{pm}) is given by

$$R_{pm} = \frac{r_d\ E_{bd}}{6.7\ i_{gm}} = \frac{3\ (250)}{6.7\ (0.015)} = 7460 \text{ ohms}$$

A Radiotron 6SN7-GT in push-pull class "A" connection will meet the requirements for a driver valve. From Fig. 2 the instantaneous screen current (i_{c2m}) is found to be 0.085 amperes.

From equations (17) and (18), we obtain

Average screen current $I_{c2} =$

$$\frac{i_{c2m}}{4} = \frac{0.085}{4} = 0.021 \text{ amperes}$$

Screen dissipation $W_{c2} =$

$$E_{c2}\ I_{c2} = 400\ (0.021) = 8.5 \text{ watts}$$

This value is well within the ratings for screen power input for the Radiotron 813. All the pertinent design information for the modulator is given in Table I. Fig. 1. is a typical circuit based on these values.

TABLE I.

Audio Modulator using 2 Radiotron-813's in Class AB_2

Values are for 2 valves.

D.C. Plate Voltage	1750 volts
D.C. Grid-No. 3 Voltage	0 volts
D.C. Grid-No. 2 Voltage	400 volts
D.C. Grid-No. 1 Voltage*	–45 volts
Peak A-F Grid-No. 1 to Grid-No. 1 Voltage	200 volts
Zero-Signal D.C. Plate Current	0.046 amperes
Max.-Signal D.C. Plate Current	0.324 amperes
Max.-Signal D.C. Screen Current	0.042 amperes
Effective Load Resistance (Plate-to-plate)	13,000 ohms
Max.-Signal Power Output	400 watts
Output Transformer Turns Ratio, r	0.806
Driver Transformer Turns Ratio, r_d	3
Driver Valve	6SN7-GT (or equivalent)

* For A.C. filament operation

FOOTNOTES.

[1] "Simplifying the Calculation of Transmitting Triode Performance" by E. E. Spitzer, "Ham Tips", Nov.-Dec., 1948.

[2] Although it is true that considerably less average audio power than the value of W_a given above is required for voice modulation, the peak power capability of the modulator must be adequate if severe distortion at the voice peaks is to be avoided. It is necessary, therefore, to compute the modulator circuit constants for sine-wave signal conditions. Somewhat lower values of plate dissipation than those calculated later will result if voice modulation is used exclusively and this fact may therefore be considered in selecting suitable modulator valves on the basis of their maximum plate dissipation rating (and, incidentally, in choosing the d.c. current rating of the modulator plate supply). It is well to remember, however, that if the modulator valves are chosen with a plate dissipation rating that is only "just sufficient" for voice modulation, a sustained whistle into the "mike" or several seconds of r-f, audio circuit, or acoustical feedback, will produce excessive plate dissipation and may result in valve failure.

[3] See footnote 2.

[4] See pages 196ff in RCA Receiving Tube Manual, RC-15.

[5] RCA Tube Handbook HB-3; Headliners for Hams, HAM-103.

Radiotron Type 4-65A V-H-F Power Tetrode

Reprinted by courtesy of the Radio Corporation of America.

Data contained herein is published for general information. No stocks of these valves are at present held in Australia.

The 4-65A is a small power tetrode intended for modulator, power amplifier, and oscillator service. It can be used with full input at frequencies up to 50 megacycles and with reduced input at frequencies up to 250 megacycles. The 4-65A has a maximum plate dissipation of 65 watts.

Features of the 4-65A include a structure with short heavy leads, a thoriated-tungsten filament, and low interelectrode capacitances permitting operation without neutralization up to 110 megacycles.

GENERAL DATA

Electrical:

Filament, Thoriated Tungsten:		
Voltage (a.c. or d.c.)	6.0	volts
Current	3.5	amperes
Mu-Factor, Grid No. 2 to Grid No. 1	5	
Direct Interelectrode Capacitances:		
Grid No. 1 to Plate	0.08	$\mu\mu$F
Input	8	$\mu\mu$F
Output	2.1	$\mu\mu$F

Mechanical:

Mounting Position	Vertical, base down or up
Overall Length	4-3/16" ± 3/16"
Seated Length	3-11/16" ± 3/16"
Maximum Diameter	2-3/8"
Base	Medium-Molded-Flare Septar 5-Pin
Cap	Skirted Small
Bulb	T-16

Bulb and Seals Temperature:

Continuous Service— 200 max. °C

Adequate ventilation around the valve must be provided to prevent the temperature of the bulb and seals from exceeding the specified maximum value.

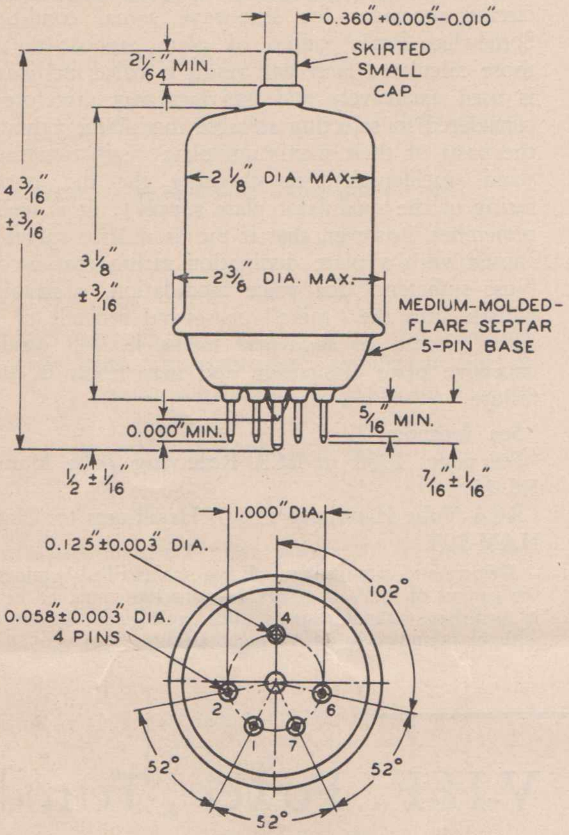
Intermittent Service ("on" period does not exceed 5 minutes and is followed by "off" period of the same or greater duration)— 220 max. °C

When ambient temperature does not exceed 30°C and the operating frequency is below 50 Mc/s, it will not usually be necessary to provide forced-air cooling of the bulb and seals to prevent exceeding the specified maximum temperature value provided a heat-radiating plate connector is used and adequate ventilation is provided.

Components:

Socket .... Johnson No. 122-101, or equivalent
Heat-Radiating Plate Connector
Eimac HR-6, or equivalent

DIMENSIONAL OUTLINE



PUSH-PULL A-F POWER AMPLIFIER & MODULATOR— Class AB$_1$*

Maximum CCS● Ratings, *Absolute Values:*

D.C. Plate Voltage	3000	max.	volts
D.C. Grid–No. 2 (Screen) Voltage	600	max.	volts
Max.–Signal D.C. Plate Current**	150	max.	mA
Max.–Signal Grid–No. 2 Dissipation**	10	max.	watts
Plate Dissipation**	65	max.	watts

Typical Operation:

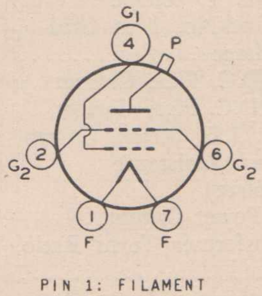
Values are for 2 valves

D.C. Plate Voltage	1000	1500	1750	volts
D.C. Grid–No. 2 Voltage×	500	500	500	volts
D.C. Grid–No. 1 (Control-Grid) Voltage▲	–85	–85	–90	volts
Peak A-F Grid–No. 1–to–Grid–No. 1 Voltage	170	170	180	volts
Zero–Signal D.C. Plate Current	30	30	20	mA
Max–.Signal D.C. Plate Current	170	180	170	mA
Zero–Signal DC Grid–No. 2 Current	0	0	0	mA
Max.–Signal D.C. Grid–No. 2 Current	24	14	17	mA
Effective Load Resistance (Plate to plate)	9000	15000	20000	ohms
Max.–Signal Driving Power (Approx.)	0	0	0	watts
Max.–Signal Power Output (Approx.)	80	145	175	watts

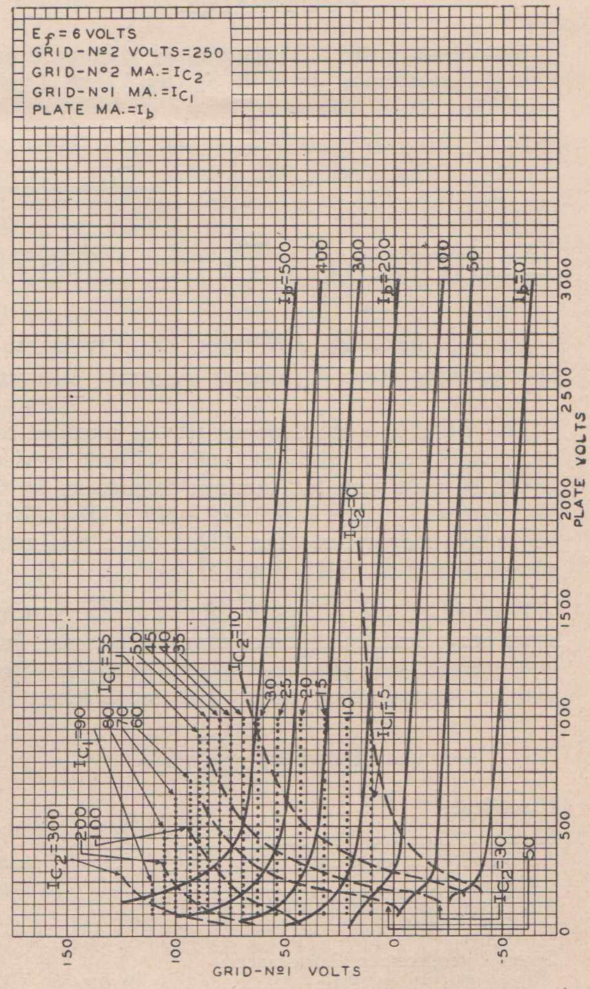
Maximum Circuit Values:

Effective Grid–No. 1–Circuit Resistance .... 250000 max. ohms

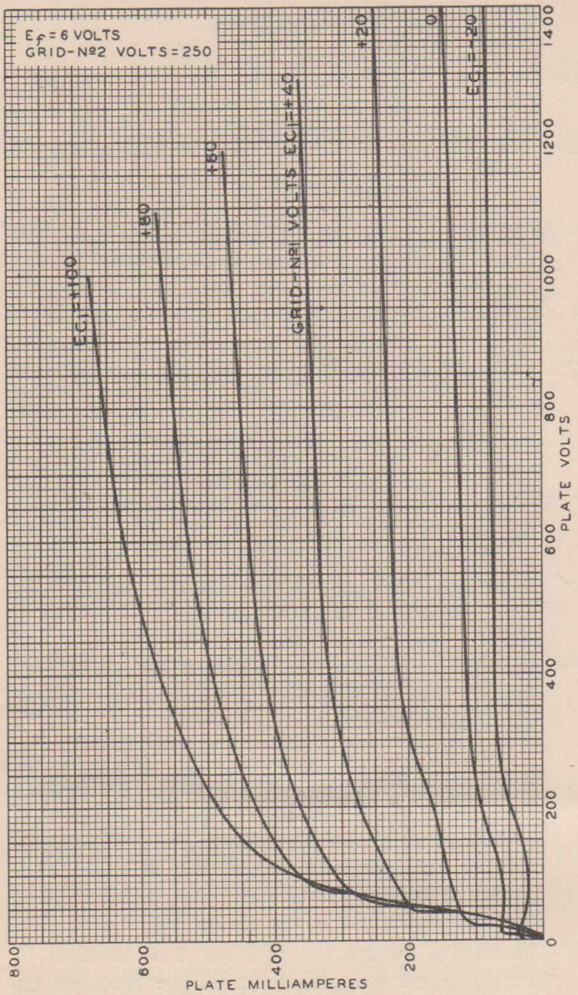
SOCKET CONNECTIONS
Bottom View



PIN 1: FILAMENT
PIN 2: GRID NO. 2
PIN 4: GRID NO. 1
PIN 6: GRID NO. 2
PIN 7: FILAMENT
CAP: PLATE



Average constant current characteristics of type 4-65A.



Average plate characteristics of type 4-65A.

PUSH-PULL A-F POWER AMPLIFIER & MODULATOR— Class AB_2†

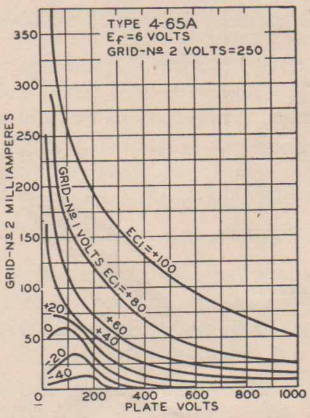
Maximum CCS• Ratings, ***Absolute Values:***

D.C. Plate Voltage	3000 max.	volts
D.C. Grid–No. 2 (Screen) Voltage	600 max.	volts
Max.–Signal D.C. Plate Current**	150 max.	mA
Max.–Signal Grid–No. 2 Dissipation**	10 max.	watts
Plate Dissipation**	65 max.	watts

Typical Operation:

Values are for 2 valves

D.C. Plate Voltage	600	1000	1500	1800	volts
D.C. Grid–No. 2 Voltage	250	250	250	250	volts
D.C. Grid–No. 1 (Control Grid) Voltage:★★					
From fixed supply of	–30	–30	–35	–35	volts
Peak A-F Grid–No. 1–to Grid–No. 1 Voltage	240	210	200	180	volts
Zero–Signal D.C. Plate Current	60	60	60	50	mA
Max.–Signal D.C. Plate Current	300	300	250	220	mA
Zero–Signal D.C. Grid–No. 2 Current	0	0	0	0	mA
Max.–Signal D.C. Grid–No. 2 Current	60	45	30	25	mA
Effective Load Resistance (Plate to plate)	3600	6800	14000	20000	ohms
Max.–Signal Av. Driving Power (Approx.)	3.1	2.5	1.6	1.1	watts
Max.–Signal Peak Driving Power (Approx.)°	6.2	5	3.2	2.2	watts
Max.–Signal Power Output (Approx.)	90	170	250	270	watts



Average characteristics of type 4-65A.

PLATE-MODULATED R-F POWER AMPLIFIER— Class C Telephony

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

Maximum CCS• Ratings, ***Absolute Values:***

D.C. Plate Voltage	2500 max.	volts
D.C. Grid–No. 2 (Screen) Voltage	400 max.	volts
D.C. Grid–No. 1 (Control-Grid) Voltage	–500 max.	volts
D.C. Plate Current	120 max.	mA
Plate Dissipation	45 max.	watts
Grid–No. 2 Dissipation	10 max.	watts
Grid–No. 1 Dissipation	5 max.	watts

Typical Operation:

D.C. Plate Voltage	600	1000	1500	2000	2500	volts
D.C. Grid–No. 2 Voltage°°	250	250	250	250	250	volts
D.C. Grid–No. 1 Voltage■	–100	–110	–125	–125	–150	volts
Peak A-F Grid–No. 2 Voltage°°	175	175	175	175	175	volts
Peak R-F Grid No. 1 Voltage	190	210	225	225	235	volts
D.C. Plate Current	117	120	120	120	108	mA
D.C. Grid–No. 2 Current	40	40	35	33	16	mA
D.C. Grid–No. 1 Current (Approx.)	11	12	12	12	8	mA
Driving Power (Approx.)	2.1	2.5	2.7	2.6	1.9	watts
Power Output (Approx.)	50	95	145	200	225	watts

R-F POWER AMPLIFIER & OSC.—

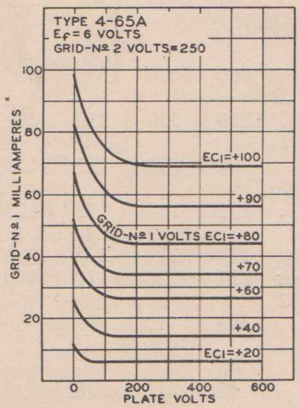
Class C Telegraphy‡
and
R-F POWER AMPLIFIER—
Class C F-M Telephony

Maximum CCS• Ratings, *Absolute Values:*

D.C. Plate Voltage	3000	max.	volts
D.C. Grid–No. 2 (Screen) Voltage	400	max.	volts
D.C. Grid–No. 1 (Control-Grid) Voltage	–500	max.	volts
D.C. Plate Current	150	max.	mA
Plate Dissipation	65	max.	watts
Grid–No. 2 Dissipation	10	max.	watts
Grid–No. 1 Dissipation	5	max.	watts

Typical Operation:

D.C. Plate Voltage	600	1000	1500	2000	3000	volts
D.C. Grid–No. 2 Voltage	250	250	250	250	250	volts
D.C. Grid–No. 1 Voltage	–50	–70	–75	–80	–90	volts
Peak R-F Grid–No. 1 Voltage	145	170	180	175	170	volts
D.C. Plate Current	140	150	150	150	115	mA
D.C. Grid–No. 2 Current	40	40	35	30	20	mA
D.C. Grid–No. 1 Current (Approx.)	13	15	14	12	10	mA
Driving Power (Approx.)	1.9	2.5	2.5	2.1	1.7	watts
Power Output (Approx.)	54	105	170	235	280	watts



Average characteristics of type 4-65A.

FOOTNOTES

ó A flexible lead should be used in making connection to the plate.

* Subscript 1 indicates that grid-No. 1 current does not flow during any part of the input cycle.

• Continuous Commercial Service.

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

× Obtained from a source having good regulation.

▲ Adjusted to give indicated value of zero-signal plate current.

† Subscript 2 indicates that grid-No. 1 current flows during some part of the input cycle.

★★ Adjusted to give indicated value of zero-signal plate current. The d.c. resistance of the bias source should not exceed 250 ohms.

° The driving 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 held at a low value.

°° Modulation voltage for grid No. 2 is obtained by supplying the d.c. grid-No. 2 voltage from the unmodulated plate supply through a series dropping resistor, or by the use of an a-f reactor in the positive grid-No. 2 supply lead, or from a separate winding on the modulation transformer. With either the series-resistor or the reactor method, the a-f variations in grid-No. 2 current resulting from variations in plate voltage as the plate is modulated automatically produce the grid-No. 2 modulation voltage.

■ The use of bias obtained partially from a grid resistor is recommended.

‡ Key-down conditions per valve without amplitude modulation. Amplitude modulation essentially negative may be used if the positive peak of the audio-frequency envelope does not exceed 115% of the carrier conditions.
When the 4-65A is used in the final amplifier or a preceding stage of a transmitter designed for break-in operation and oscillator keying, a small amount of fixed bias must be used to maintain the plate dissipation within rated value. With 2000 volts on the plate, and 250 volts on grid No. 2, a fixed bias of at least –40 volts should be used.

Transmitter Aerial and Site Selection

Reprinted from RCA publication Form 7865 by courtesy of the Radio Corporation of America.

RADIO WAVE PROPAGATION

Types of waves — There are three general forms of wave propagation — surface or ground wave propagation, skywave propagation by reflections from the ionospheres, the optical (or quasi-optical) propagation.

Ground waves are utilized for the lowest radio frequencies, from 10 to approximately 3000 Kc/s. For a given radiated power over typical land surfaces, the attenuation of signal strength vs. distance increases with frequency and with the average resistivity of the soil or earth. Over sea water, the rate of decay is much lower than that of the best earth due to the relatively low resistivity of salt water, so that much greater distances can be covered with given power over sea than over land. Ground waves are so called because their utilization depends upon the nature of the electro-magnetic fields at the boundary between the wave in the atmosphere and the accompanying ground currents propagated in the earth below it. For most effective utilization the wave must be vertically polarized; i.e., the electric vectors lie in vertical planes extending in the direction of propagation.

Sky-waves are so called because they are propagated in space above the earth and reflected earthward from regions of high free-electron density in the upper atmosphere (called ionospheres). The ionospheres thus act as imperfect mirrors for radio waves roughly between 200 and 30,000 Kc/s. Typical propagation over great distances may involve several reflections between earth and ionosphere before arriving at a distant receiving point. The waves travelling in the atmosphere between these boundaries decay at a rate which is inversely proportional to distance. The reflection coefficient of the ionosphere depends upon the particular layer used, the frequency, the angle of incidence, and the immediate state of ionization in the ionosphere layer, this latter being dependent upon solar ultraviolet radiation, sunspot activity and the consequent state of the earth's magnetic field. The layers are seriously modified by sunspots, magnetic storms and aurora borealis.

Ionospheric propagation — The ionosphere layers have been studied very extensively by adaptation of radar principles. Two principal layers have been found. These are designated as the E and F regions, the latter being divided into 2 normal layers during local daylight hours, called F1 and F2. Other layers often exist transiently. The E-layer exists during daylight only, and has a stable height of approximately 170 km. Sporadic layers at E-layer heights may exist at any time. At night there is one stable F-layer with a nominal height of the order of 300 km. After local sunrise, sunlight ionization of the upper atmosphere causes the formation of the E-layer, and the break-up of the night-time F-layer into the F1 at roughly 230 km. average height and the F2 at heights from 325 to 450 km. After local sunset, deionization commences with the disappearance of the E-layer, and the convergence of F1 and F2 layers into the F which alone remains throughout the hours of darkness. Just before local sunrise, the electron-density in the ionosphere is at its lowest and transmission during this period requires lower frequencies and higher power than at other times of the day.

There are critical frequency waves which penetrate these layers and are not reflected. The optimum working frequency for radio transmission is usually 15% below the critical frequency for normal incidence at the point of reflection. This optimum frequency varies throughout the day and night, necessitating changes in operating frequencies over fixed radio circuits for best transmission efficiency. The Central Radio Propagation Laboratory of the U.S. Bureau of Standards, co-operating with agencies throughout the world operating continuous ionosphere sounding stations, compiles and publishes complete ionosphere data monthly.

Medium frequency broadcasting (550-1600 Kc/s) provides primary service with ground waves and sometimes an intermittent skywave service at night. Usually skywaves and ground waves overlap at distances from 50 to 150 kms. from a broadcast station causing severe rapid fading at night. Skywaves between stations are also a serious source of interference. Recent years have seen great strides in the development, design and installation of special directive antennas to permit several broadcasting stations to operate on the same frequency satisfactorily. The engineering principles are included in the North American Regional Broadcasting Agreement.

At frequencies above those which are reflected by the ionosphere (roughly above 30,000 Kc/s) radio propagation becomes optical in character. Within radio-optical line-of-sight, signals are propagated in the space immediately above the surface of the earth, by direct optical transmission. Due to the very narrow angles subtended by practical heights of transmitting and receiving antennas at distances of several miles, the received wave is the composite of the ray transmitted directly in free space, and one or several rays which have been reflected from the ground, the latter of opposite phase and very nearly the same amplitude. High resultant field strengths at distant points are obtained only by providing circuit geometry which will avoid this normal tendency to complete cancellation near the ground so that the *distance travelled by the reflected ray is at least one-sixth wavelength greater than the direct ray*. The lower the frequency, the higher must be the transmitting and receiving antennas above the ground level to achieve this.

Beyond the radio-optical horizon, the wave field present is due to diffraction. At frequencies below about 150 Mc/s, useful service can exist in the diffraction region, thus extending the distance which can be covered from a given antenna height. Above this frequency the diffraction zone vanishes virtually and transmission distance approaches that of the optical horizon. It has been found that frequencies above 5,000 Mc/s are affected by meteorological conditions and the state of the atmosphere which produce transient mirage effects, absorption and dispersion.

BROADCASTING ANTENNAS

Medium frequency broadcasting — 550-1600 Kc/s.—Primary service is rendered by ground-wave coverage and a secondary service is provided at night by skywaves. The object is to get the highest field intensity along the ground. This is done by increasing power, or by increasing antenna efficiency. We will deal only with the latter.

Antenna efficiency is increased in two ways: (A) By decreasing the dead loss in the system, principally the ground, and (B) by controlling the current distribution in the antenna so as to concentrate its energy toward the horizon instead of at high angles where such radiated energy serves no useful purpose.

It has been found by experimental research[1] and proved hundreds of times since, that a radial system of ground wires centred under the antenna, having a length of 0.4 wavelength and 120 radials, is an adequately efficient ground system. More radials which are longer add little to the improvement but increase cost very much. Fewer, shorter wires depreciate performance slowly at first and then rapidly as ground wires are shortened and reduced in number. It is suggested that, as a fixed formula which is safe to use, ground systems be always:

Radials—120 in any case (no ground rods or circular wire connections).

Length—0.3 wavelength min. 0.4 wavelength preferred.

Depth—Immaterial—but 6 to 15 inches is usual.

Wire—No. 10 B & S gauge bare copper (0.1" dia.) minimum size.

The height of a radiator of uniform cross section determines the current distribution. A larger and larger portion of a nearly sinusoidal standing wave of current is obtained as the antenna height is increased[2]. This distribution controls the vertical radiation pattern for the antenna and also its efficiency in producing signals along the ground. This table (see below) contains the main facts, based on 1 kw of power into a vertical radiator (wire or tower of uniform cross-section) and an optimum ground system. The field intensities quoted are unattenuated values at 1 mile (1.61 km).

Tower radiators[3]—Rather than use two towers to support a wire antenna it is more economical to use a single tower of uniform, or very nearly uniform, cross-section as the radiator. If the cross-section deviates much from uniform the current distribution, and therefore the efficiency, is unfavourably modified. Many satisfactory types of tubular masts and cantilever structures, guyed and self-supporting, have become available and are used generally for broadcasting.

Occasionally the need for a directive antenna arises for broadcasting use. There are several hundred in use in North America, using from two to six radiators. These must be specially treated and no general statements can be made beyond the following:—

1. Except in unusual cases, 2 radiators would probably suffice for a directive antenna in most countries at the present time.
2. Every directive antenna, and its associated impedance matching power dividing and phasing

Order of merit	Height Wave-lengths	Height degrees	Unattenuated field intensity at 1 mile	
—	0.125	45	191 MV/M	In this range we specify a 60° antenna as best value for low and medium power stations.
3	0.165	60	193 MV/M	
—	0.25	90	196 MV/M	
—	0.333	120	203 MV/M	
2	0.46	160	220 MV/M	A good intermediate recommendation.
1	0.555	200	256 MV/M	Best antenna for anti-fading reasons.
1*	0.65	235	270 MV/M	Best efficiency but not best anti-fading.
—	0.72	260	200 MV/M	Included to show that beyond 235° no desirable properties remain.
—	0.78	280	110 MV/M	

* Recommended only when sea water or soil of very high conductivity (3×10^{-13} e.m.u. or more) prevails for a distance of 100 miles or more in all directions, and when the operating frequency is less than 1000 Kc/s.

apparatus, must be specially designed and built for the particular job. It is therefore necessary to know the pattern or the performance required. The operating frequency and the power to be used should also be known.

3. The design of the coupling circuits is directly affected by the tower height, operating frequency, spacing, phase relations of the currents, current ratios, power, the place where power is fed to the system, and the kind of transmission lines which are used. There are so many combinations of these factors that every case becomes an individual design problem.

Low frequency broadcast antennas—160-300 kc/s—In countries where l-f broadcasting is used the standard antenna has been a "T" wire radiator supported by two towers, usually 200 to 250 metres high. The antenna is often electrically a quarter-wave length (90°). At these carrier frequencies, plus and minus 10 kc/s, sidebands occupy a relatively wide band, and special means are necessary to reduce antenna selectivity. Tower antennas have special advantages in this regard, due to this relatively large cross-section.

RCA, by means of v-h-f models, has studied the l-f broadcast antenna quite extensively and has developed a special design of vertical tower radiator with intrinsically lower selectivity, thus providing a system having a greater natural bandwidth characteristic. Low-frequency broadcast antennas should be especially designed for maximum practicable bandwidth capabilities.

L-F broadcasting is rendered by direct ground wave transmission. However, radiators of practical heights cannot usually surpass $\frac{1}{4}$ wavelength, so that the special techniques developed for medium frequency antennas of greater electrical height cannot be applied to these frequencies.

High frequency broadcasting—3 to 32 Mc/s—In this band, all transmission is via the ionosphere, so that antenna designs are quite different from those above using vertically polarized waves along the ground. In this band, skywave transmission is used entirely, even for short distances. This subject is so complex that only a few general guiding remarks will be given. These complications arise from the nature of h-f wave propagation.

1. Short distance transmission requires an antenna radiating its energy at very high angles. For tropical broadcasting, a half-wave dipole antenna about one-quarter wavelength above ground, should be used, whatever frequency. The antenna should be horizontal and perpendicular to the principal line of transmission.

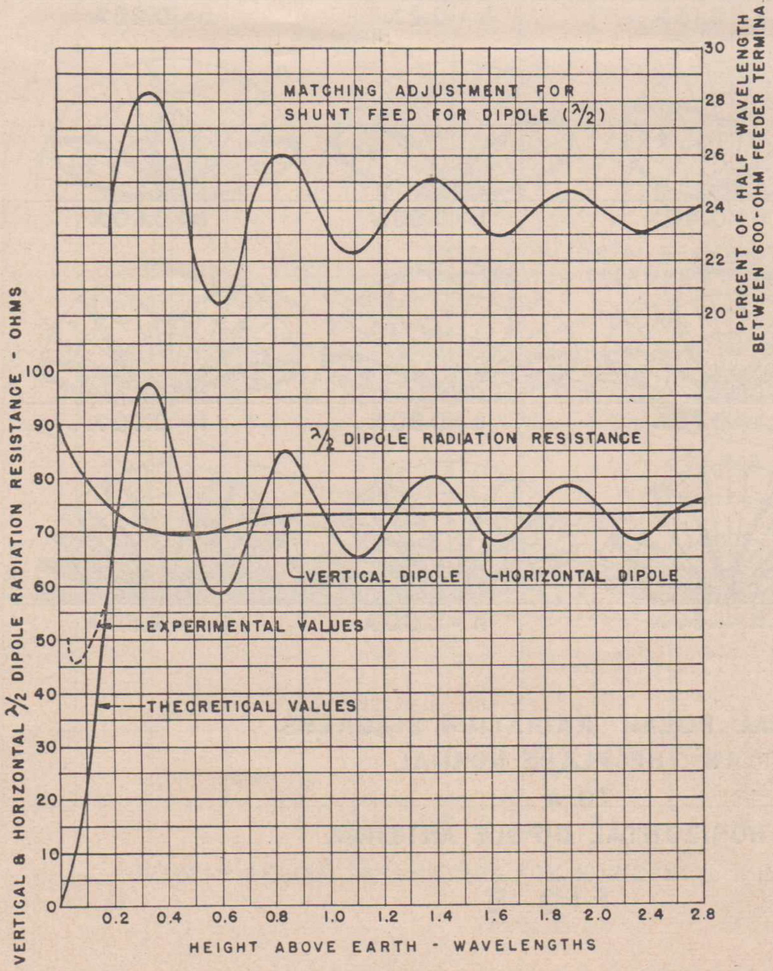


FIG. A

2. Horizontal polarization is most often used for h-f.

3. The longer the circuit, the lower the desirable angle of radiation. For circuits over 2000 miles, the vertical pattern of the antenna should beam the energy at from 5° to 10° above the horizon. For shorter distances essentially vertical transmission becomes necessary.

4. The optimum vertical angle of radiation depends upon which layer of the ionosphere is used, because they vary greatly in height with time of day and season, and with aurora and sunspot activity.

5. Long distance international broadcasting usually requires directive antennas in order to:

   (a) Serve a region to best advantage on the frequency in use.

   (b) Obtain a power gain by directive concentration of power.

   The amount of directivity, and the power gain, depend upon the geography involved.

6. The simplest and least costly directive antenna for long-distance high gain operation, with low vertical radiation, is the 3-wire rhombic antenna. It has the further advantage that several widely different frequencies can be fed to the same antenna, but

limit it to approximately a 2/1 frequency range in most cases.

7. Rhombic antennas giving high angles of radiation also give low gain, and low angle radiation is achieved by high gain antennas. With a high antenna composed of stacked dipoles, one can get low angle radiation with broad pattern in the horizontal plane. The horizontal pattern can be beamed by various numbers of such dipole arrays side by side, but with rhombic and V antennas, both characteristics are interrelated, and low angle radiation is accompanied by a narrow horizontal beam, and high gain, and vice versa.

8. If dipole arrays are used, their selectivity is such that they will accommodate only about 2% band width. Thus a different array would be needed for each international broadcast frequency band, for each direction of transmission.

9. H-F antennas of the horizontal dipole, dipole array, V and rhombic types require no ground systems, and may be built over any type of soil. However, ground reflection losses are lower when the soil conductivity is high.

10. Antennas for high angle radiation can be placed in valleys. Antennas for low angle transmission should have horizons which subtend angles that do not exceed one-half the beam angle.

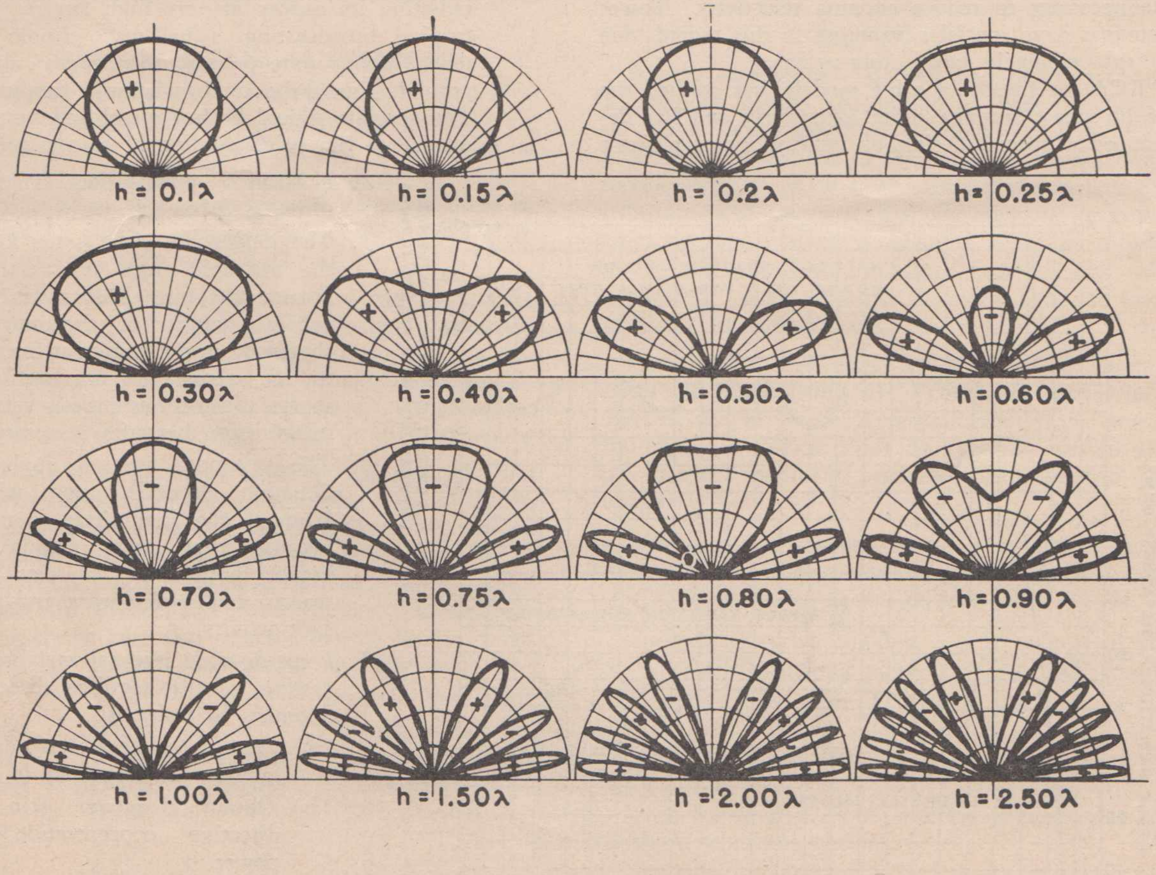
11. The most important types of high gain antennas to consider for concentrated long distance broadcasting are:

    (a) The horizontal-rhombic antenna
    (b) The stacked-dipole antenna.

    (a) Types use low supports but require considerable land.

    (b) Types use less land but require high supports and complex rigging and are frequency selective.

12. The same general remarks apply to h-f communication antennas in this frequency range.



VERTICAL POLAR RADIATION DIAGRAMS
IN THE PLANE NORMAL
TO A
HORIZONTAL DIPOLE ANTENNA

FIG. B

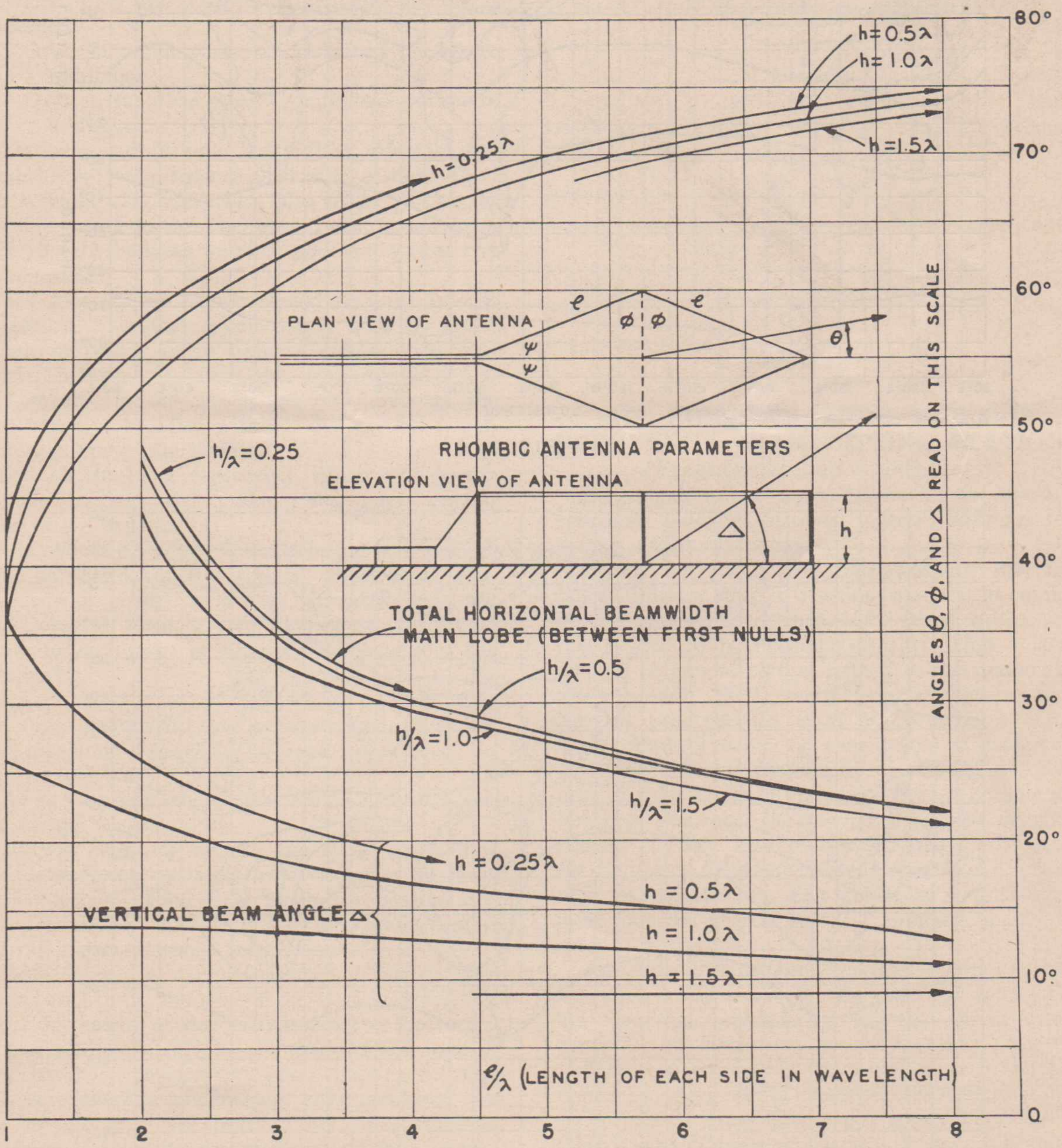
F-M antennas—88-108 Mc/s.—Due to the propagation characteristics of these frequencies antenna designs in this frequency range differ from those discussed above. There is no ionospheric consideration involved in this band, and transmission is based on the principles of optical circuits. The purpose is to concentrate the signal on the horizon, and at the same time have uniform transmission in all directions in the horizontal plane. Any gains realized therefore must come about from the shaping of the vertical radiation pattern. This is done by vertical stacking of radiating elements. Since horizontal polarization is preferred, some special forms of antennas have been evolved. These are principally:

(a) RCA Turnstile Antenna, in from 2 to 10 layers, depending upon the gain desired.

(b) RCA Super Turnstile Antenna, a turnstile arrangement of from 1 to 4 layers of current sheet radiators.

(c) RCA Pylon Antenna, of from 1 to 4 sections.

(d) Loop radiators of from 1 to 10 layers, in several versions.



OPTIMUM DESIGN PARAMETERS FOR HORIZONTAL RHOMBIC ANTENNA

FIG. C

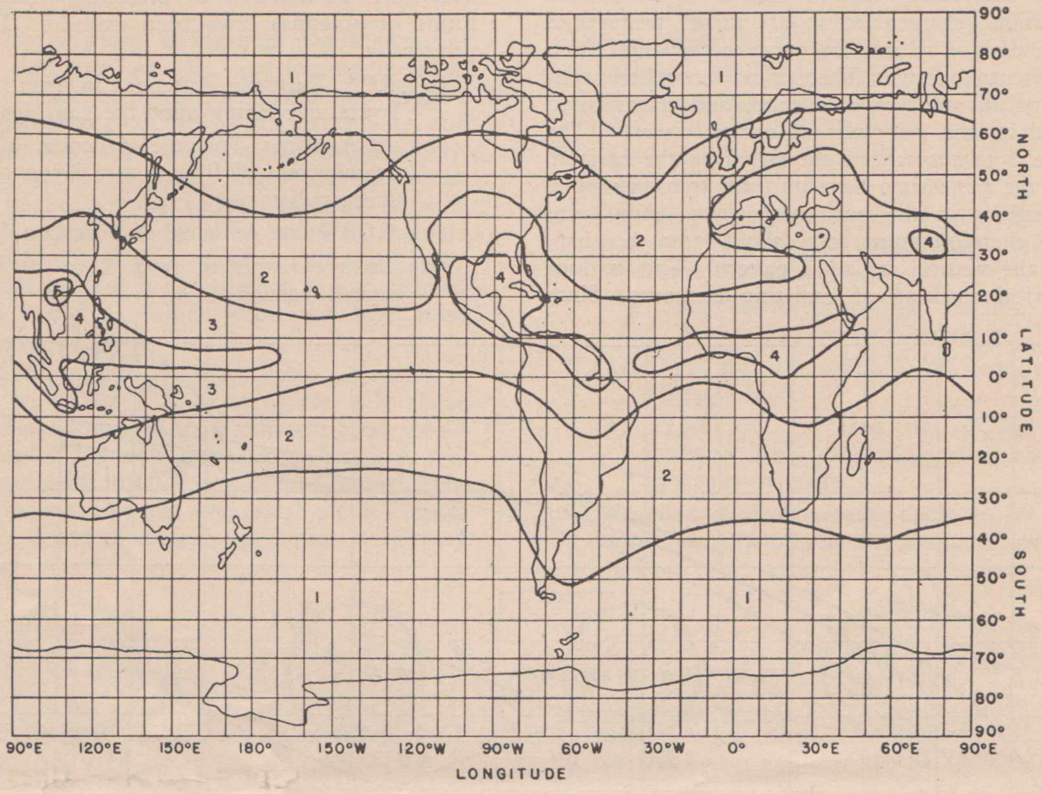

90°E
120°E
150°E
180°
150°W
120°W
90°W
60°W
30°W
0°
30°E
60°E
90°E
LONGITUDE
90°
80°
70°
60°
50°
40°
30°
20°
10°
0°
10°
20°
30°
40°
50°
60°
70°
80°
90°
NORTH
LATITUDE
SOUTH

FIG D

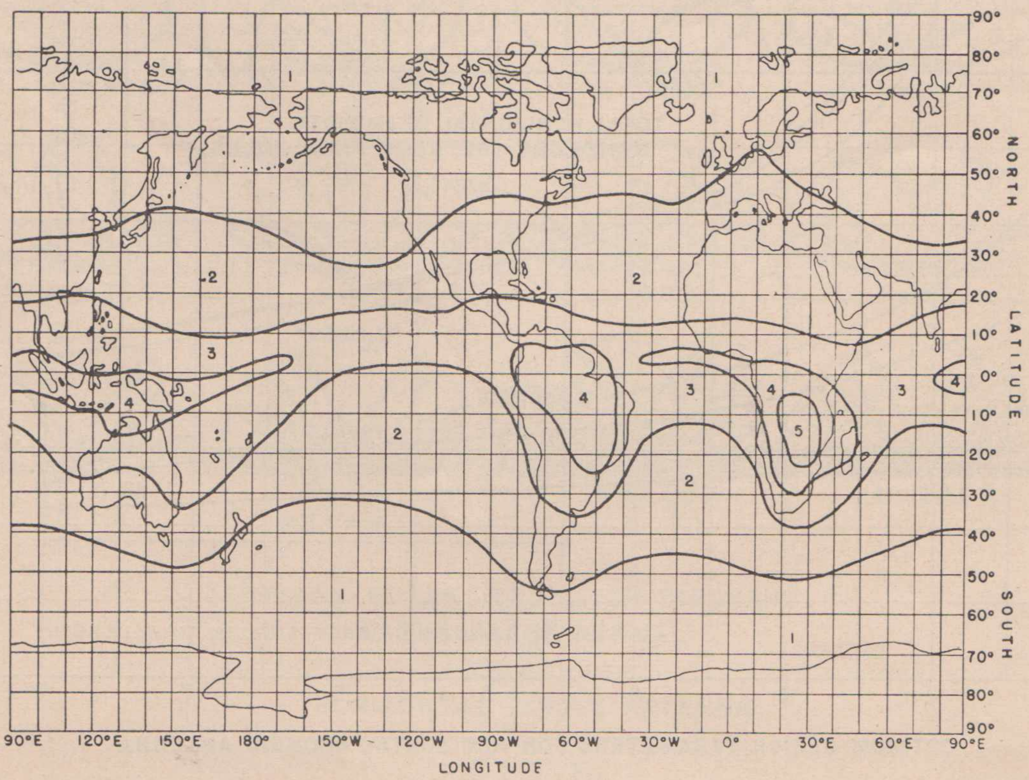

90°E
120°E
150°E
180°
150°W
120°W
90°W
60°W
30°W
0°
30°E
60°E
90°E
LONGITUDE
90°
80°
70°
60°
50°
40°
30°
20°
10°
0°
10°
20°
30°
40°
50°
60°
70°
80°
90°
NORTH
LATITUDE
SOUTH

FIG E

Note: These types all transmit horizontally polarized fields.

The application problem in this field is much simplified by the fact that prefabricated antennas are practical, and they can be handled as catalogue items.

SITE CONSIDERATIONS FOR STANDARD BROADCAST STATIONS

(Medium Freq. — 550 to 1600 Kc/s)

In the absence of local regulations to the contrary, locations should be chosen that, for the power and antenna to be used, the following field intensities are delivered over the city to which the station is associated:

Over all urban parts—25 millivolts per metre minimum.

Over all suburban parts—2 millivolts per metre minimum.

Choose a site having the highest available soil conductivity for a distance of a mile or more from the antenna, and one which also offers as clear as possible a path to the city. Avoid obstructions such as high hills between antenna and the nearest city to be served.

The plot for the station should be large enough to enclose a ground system about one-half wavelength in radius, or as near to this as local conditions permit. An acceptable compromise would be a square plot one wavelength on a diagonal, the whole plot to be filled with a radial ground system centering under the antenna.

Soil should be a consistency which will permit suitable footings for antenna and guys without unnecessary expense.

Site should be where antenna height will not become an obstruction for airways. When local regulations with regard to airways obstructions do not exist, use the formula that the nearest one should locate to an airport is 50 times the height of the antenna*.

Locate the site as conveniently as possible, after the above conditions have been satisfied, to power and telephone lines, and to road or railroad for transportation.

Avoid locations where there are chances of flooding at any season.

Hill-top locations require the same precautions regarding ground systems and antennas as do those on flat terrain. There is no advantage or disadvantage (electrically) to hill-top sites for stations operating on low or medium frequencies.

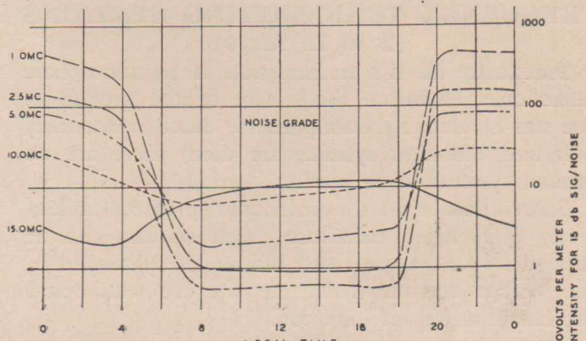
Hills higher than one-half wavelength cast shadows behind such hills, the size of the shadow depending upon the height of the hill. Regions to be served in places where high hills cast shadows will have to be taken into consideration when choosing the site.

Where directive antennas are to be employed, the size of the plot will have to be correspondingly increased to accommodate antennas and their ground systems.

*Consideration to be given to future airport expansion plans.

Try not to locate nearer than 1 wavelength to another broadcast station.

FIG F



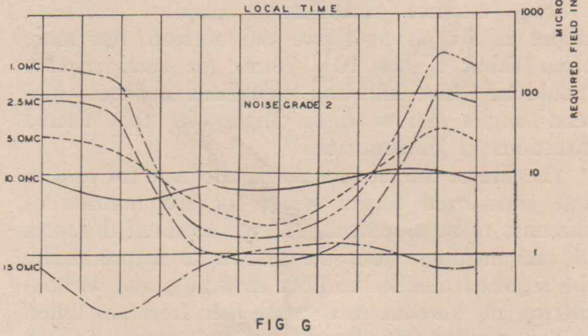


FIG G

SITE CONSIDERATIONS FOR LOW FREQUENCY TRANSMITTING STATIONS

(Frequencies below 400 Kc/s)

In general, follow recommendations for medium frequency broadcast stations, together with the following additional points.

Geology is particularly important for low frequency stations since it is seldom practical to employ ground systems more than 1000 ft. in radius, and this is well below optimum usually. Thus a large amount of ground current returns to the antenna by the soil beyond the limits of the ground wires. Such currents penetrate the earth to considerable depth (see curves of depth of penetration of earth currents for different frequencies and soil conductivities) and the nature of the subsoil is to be considered. For best efficiency, the ground should be of good conductivity to a depth of 10 to 20 metres (the lower the frequency the greater the depth desired), before rock is encountered. This condition is difficult to meet, but the nearest practical approach should be made.

For long distance and marine communication, locations should be at or very near sea water. For low frequency broadcasting, location will be determined by the disposition of desired service areas.

Where directional l-f transmission is to be obtained by using Wave Antenna, site selections are quite different from the foregoing. IN THIS CASE ONLY, soil of poor conductivity is desirable under the antenna, which is a pole line open-wire system running one wavelength or more in the desired direction of transmission. For such an application, the plot consists of a long narrow right-of-way for the pole line.

SITE CONSIDERATIONS FOR HIGH FREQUENCY TRANSMITTING STATIONS (2 to 25 Mc/s)

Practically all h-f transmission is by ionosphere reflection. Therefore the nature of the ground at the site is relatively unimportant. Since horizontally polarized radiation systems are used, there are no ground systems to be used.

Short distance h-f transmission (up to 600 miles) is by high angle radiation. Such stations can be located in valleys, provided the angles subtended by the visible horizons from the site are well below the useful angles of radiation.

Long distance transmission employs radiation angles as low as 5° above the horizon, but more often about 8° or 10°. Sites for such stations should be where hills and mountains and trees subtend angles below these angles in the desired directions of transmission.

The size of the plot should be adequate for present and anticipated future needs for the number of antennas to be employed, and the required distances of the antennas from the station to accommodate the required number without crowding, and without placing any antenna in a strong field from any other.

The site should be conveniently located with respect to roads, railroad, power lines, and telephone line.

The highest antennas may constitute an obstruction to aviation, If no local rules exist with regard to such matters, follow the practice that the site is not nearer than 50 times the height of the highest tower to the nearest airport.

Avoid locations subject to flooding or other troublesome phenomena.

HIGH FREQUENCY RECEIVING STATION SITES

The same general conditions apply as for transmitting stations, but with the additional important consideration of local electrical noise. The site should be far enough away from cities, towns, factories, and especially highways with appreciable motor traffic so as not to be affected by man-made noise. Artificial electrical noise should never exceed natural atmospheric noise under the least conditions of the latter. If power must be generated at the receiving station, the internal combustion engines used should be completely shielded for radio interference. Ignition interference can be very destructive.

STRAIGHT HORIZONTAL HALF-WAVE DIPOLE

A straight conductor parallel to earth, of the order of one-half wavelength long, suitably excited, is one of the simplest and most useful h-f radiating systems. When exactly one-half wavelength long, and in free space, a thin wire antenna has a centre-point impedance of approximately $73 + j\,30$ ohms. When placed parallel to and within about 5 wavelengths of ground, mutual impedance with its image causes the impedance to vary appreciably with its height. Fig. A shows how the radiation resistance varies with height for both horizontal and vertical half-wave dipoles.

The electric field around a half-wave dipole resembles that of the magnetic field of force around a short bar magnet. When in free space, the field intensity at constant distance follows the relation

$F_c = \cos(90^\circ \cos\theta)/\sin\theta$ (θ is angle to the wire)

from which it can be seen that there is zero field intensity in the direction of the dipole and maximum everywhere at right angles to it. It is naturally a directive radiating system. When used in practice over ground, some endwise radiation occurs because of reflections from ground and from the ionosphere and in this direction the field is vertically polarized. Normal to the wire field is horizontally polarized. In intermediate directions there are components of both. As the centre of such an antenna sags more and more both vertically and horizontally polarized fields are radiated normal to the wire.

In most h-f applications, the most important characteristic of such an antenna is its vertical-plane pattern normal to the wire. Over perfectly conducting earth, this has the equation

$$F\phi = \cos\ (H \sin\ \phi + 90^\circ)\ \text{(see Figure B)}.$$

Both transmitting and receiving antennas of this type, used over a given fixed circuit, should be at the same height and perpendicular to the direction of transmission. This insures that transmitting and receiving antennas have complementary characteristics.

Feed systems—The straight half-wave dipole can be centre-fed in series from a balanced feeder, or shunt-fed (also called delta-and-Y-feed). In the former, there is a large mismatch between antenna and feeder impedances for the type feeders commonly used, so that means must be employed to match the line at some point close to the antenna if highest efficiency is desired in long feeders. The shunt-feed system, when correctly performed makes a satisfactory impedance match directly (but never perfectly, unless series capacitors are included to correct for the inductive effect of the enclosed portion of the antenna) for low standing-wave ratio on the feeder. Care must be taken to maintain exact symmetry of connection between line and the antenna. Adjustments for this are shown in upper part of Fig. A.

Feeders should always run normal to the antenna wire for as far as possible. Skew relations cause radiation coupling between feeder and antenna which upset line balance and compromise pattern and efficiency. This is quite general, and should be observed in all antenna designing.

Another feed method, technically inferior but often used for its simplicity, is off-centre feed, with a 1-wire feeder working against ground. A position can be found on the antenna where standing waves on the feeder are quite small for one frequency.

This is a convenience for transmitters having single-end output circuits. The single-wire feeder should be as short as possible.

Radiated field intensities—In free spaces maximum field intensity at 1 mile (d = 1610 metres) with 1000 watts radiated from a half-wave dipole is 138 millivolts per metre. Due to ground reflections, pattern maxima vary from 1 to 2 times this value, depending upon height and ground reflectivity.

Selectivity and bandwidth—The selectivity of a half-wave dipole is maximum for thin wires where it can be considered to have a bandwidth of the order of $\pm$ 2% of optimum frequency. Where larger bandwidths are to be transmitted quite uniformly, the ratio of diameter to length of 1-wire or cage antenna should be the largest practicable. If propagation considerations permit, the heights should be made such as to give maximum radiation resistances as shown in Fig. A. Where bandwidths in excess of $\pm$ 5% in frequency are required, other types of construction should be used, such as folded dipoles of large cross-section.

Antenna potentials—Where high power is to be transmitted, or at high altitudes, antenna insulation and conductor designs require care to details. For h-f, only radial potential gradients need be considered. At high altitudes, pluming may occur, with consequent damage to the system. Fortunately in practice, high power is usually used with directive antennas, and the power is divided between several dipole sections, which tends to minimize this problem. A thin wire dipole gives an end potential of about 3900 volts r.m.s. for 1000 watts antenna input for a height of 0.25λ. It will be higher for smaller heights, and falls to a minimum of about 1700 volts as heights increases to 0.75λ; beyond which it settles down to the free-space value of about 3000 volts. Potentials vary as the square root of the power ratio and as the inverse square root of the capacitance per unit length. For a potential of 3900 volts on a wire 0.101 inch diameter (No. 10 B & S), the radial gradient is of the order of 31 kv/cm. Larger diameter wire or four wires in the form of a cage can be used to reduce gradient and losses due to corona.

Half-wave dipole for receiving—In receiving, one wishes to know the receiver input voltage when the receiving antenna is immersed in a plane wave field of E volts per metre when both the antenna and the receiver are correctly matched to a feeder of known characteristic impedance Z_0. This can be approximated in the following manner:

The area over which a half-wave dipole collects energy is 0.1305 square wave-length. The power within this area from the field is $W = E^2/377$. If the antenna is in free space and has no losses, the power intercepted would be, for optimum antenna orientation.

$W = 0.1305\ \lambda^2 E^2/377$ (watts)

λ = wavelength in metres.

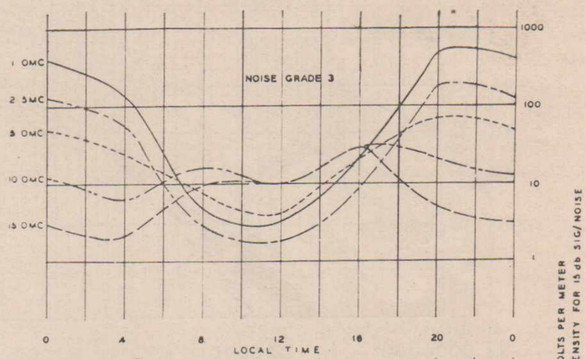
E = field intensity in volts per metre.

so that the receiver input voltage would be

$e = \sqrt{W/Z_0}$ (volts).

Over ordinary ground we may estimate that the effective area with the image might be about 1.8 times that in free space. Mismatches in the system, together with other losses, probably would yield a net value approximately the same as for the free space condition given above.

FIG H



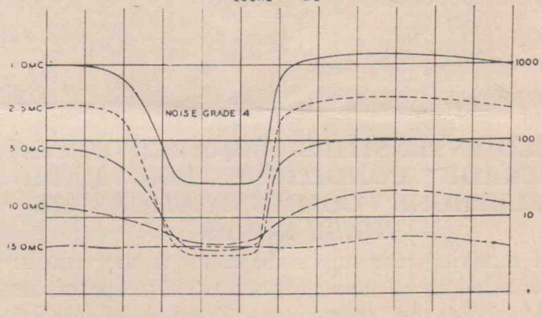


FIG I

DESIGN OF HORIZONTAL RHOMBIC ANTENNAS

The accompanying curve (Fig. C) gives the optimum design for a horizontal rhombic antenna in terms of its physical dimensions in wavelengths for the operating frequency chosen. When the antenna is used for several frequencies, the optimum design is chosen for the highest frequency that it is to be used. At frequencies other than optimum, the performance will be compromised. Tolerable compromises may apply for a 2/1 frequency range, but beyond this the compromise may not be tolerable for a given radio circuit. Therefore good engineering practice will limit the range of frequencies applied to one antenna to something of the order of 2/1 and often less than this.

Rhombic antennas for both transmitting and receiving should preferably be the type using 3 wires per side, spread about 2 metres at the side supports, and converging to a single conductor at the end supports. Single-wire antennas can be used for receiving when utmost economy is essential and where precipitation static is not often encountered.

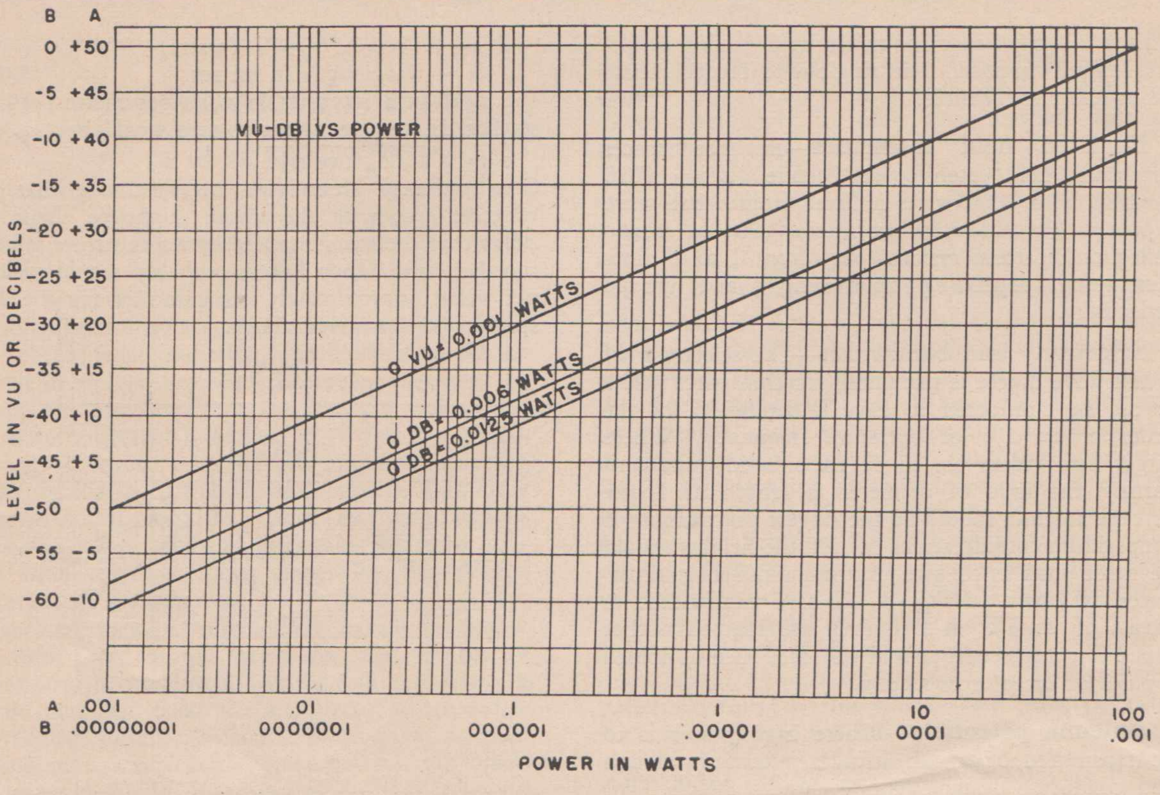


FIG. J

FIELD INTENSITIES REQUIRED UNDER AVERAGE CONDITIONS FOR RADIO TELEPHONE COMMUNICATION, BY NOISE ZONES

In receiving, the important factor is the ratio of signal strength to noise (S/N). The maps of the world show the approximate location of natural atmospheric noise zones for May-September (Fig. D) and November-March (Fig. E). There are five zones indicated as noise grades 1 to 5.

In any zone, there are typical variations of noise throughout a day, and these follow rather constant patterns. However, the information represents only general average conditions and considerable local variations can be expected.

Fifteen decibels S/N is regarded as a minimum for satisfactory commercial telephone service. Accordingly, Figs. F, G, H, and I show average required signal field intensities for 15 db S/N, as it varies with local time and with frequency from 1 to 15 Mc/s for noise zones 1, 2, 3, and 4.

For CW telegraph operations, low S/N can be tolerated. For good quality broadcast service S/N should be much higher, and correspondingly higher field intensities are necessary. These figures are based on information published by the Central Propagation Laboratory of the National Bureau of standards (U.S.A.) and are useful in planning radio systems.

NOTE — RE: RADIO TRANSMITTER POWER

The power of a transmitter to meet a particular radio communication circuit condition can be selected for minimum cost if the frequencies are properly chosen, and the antenna systems for use with those frequencies over the distance involved are satisfactorily engineered. The receiving installation must also be adequately engineered from the points of view of location, antennas, diversity reception, etc.

It will be noted that in high frequency communication, where skywave propagation is employed, the calculation and selection of the working frequency is extremely important; the conditions, as met in practice, usually require 2 frequencies, and sometimes as many as 4 frequencies are necessary to obtain a good grade of service over a 24 hour period. The factors of fading and noise level variation should also be given due consideration; it is common to encounter fading of 10 to 80 decibels and the noise level may vary over a wide range therefore when transmitter power is considered, increments of power by factors of 2 (3 db) or 3 (4.8 db) are insignificant. A signal increase of at least 10 decibels (power increase by a factor of 10) is necessary to achieve a noteworthy improvement in circuit performance.

[1] G. H. Brown—Ground Systems as a Factor in Antenna Efficiency—Proc. I.R.E. June, 1937.

[2] G. H. Brown—A Critical Study of the Characteristics of Broadcast Antennas as Affected by Antenna Current Distribution—Proc. I.R.E. January, 1936.

[3] H. E. Gihring & G. H. Brown—General Considerations of Tower Antennas for Broadcast Use—Proc. I.R.E. April, 1935.

Elimination of Reflections on Video Lines

By C. A. MEYER* and R. G. MIDDLETON*

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Television "ghosts" may arise from several causes. One type of ghost is caused by reception of two or more signals from the same transmitting station. The desired signal travels, in most cases, by direct line-of-sight, while the undesired signal is reflected and delayed along its course. Reflection can take place from buildings, cliffs, suspension bridges, or other large structures, man-made or natural. The reflected signal is delayed with respect to the line-of-sight signal because it travels a longer path, and thereby causes a displaced or ghost image on the television screen.

A second type of ghost is caused by a mismatch of television line to the receiver input circuit. When the impedances of the line and the receiver input circuit differ, not all of the available signal energy is delivered to the receiver. Instead, a portion of the incoming energy is reflected from the receiver back up the line to the antenna. This diverted energy is usually reflected from the antenna terminals back down the line to the receiver, where it causes a displaced secondary image or ghost. When the mismatch between the receiver and the line is very bad, several ghosts can be observed because the signal "bounces" back and forth several times along the transmission line between the antenna and the receiver.

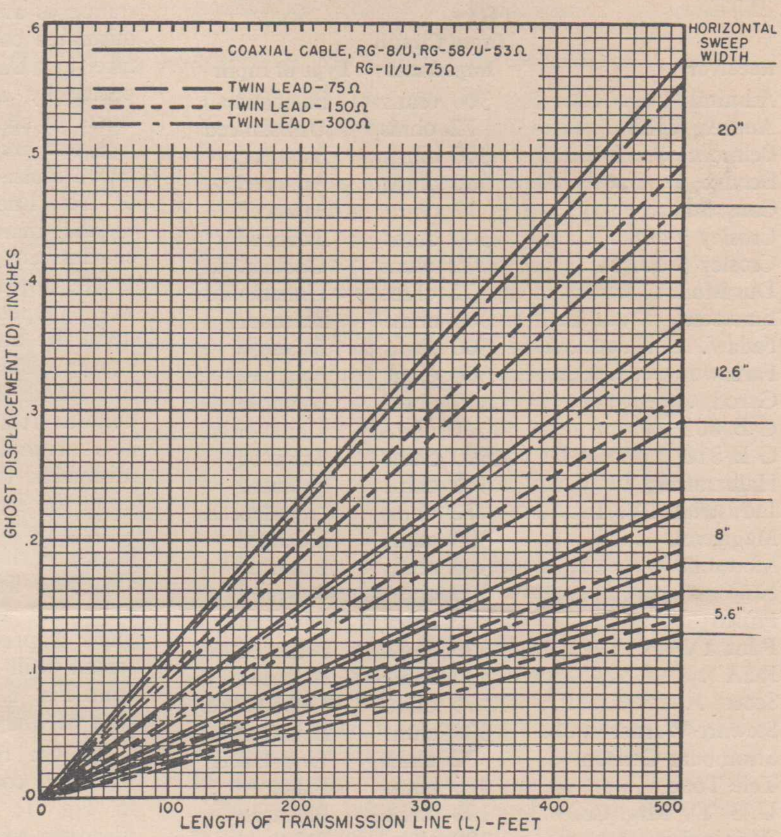

Fig. 1. Chart may be used to determine whether ghost images appearing on television screens are caused by multi-path reception or by line mismatch.

A third type of ghost can result from impaired receiver operation when the video intermediate-frequency amplifier or video amplifier is improperly adjusted. This type of ghost, however, is infrequent.

The first type can be minimized or eliminated, in many cases, by suitable orientation and positioning of a directional antenna. In poor receiving locations, however, it may be impossible to attenuate this type of ghost signal to a level where it is unobjectionable. No other method as yet is commercially available which can eliminate ghosts in such situations.

This article is primarily concerned with the elimination of the second type of ghost: that caused by a mismatch between the transmission line and the receiver input circuit. It should be observed that while a line can be practically matched to a receiver for operation on all television channels, it is not possible to match a line to an antenna over a wide band of frequencies. Hence, a match is sought only at the receiver input terminals. Fortunately, a match at this point suffices for satisfactory, ghost-free reception.

Mismatch arises in practice from several situations. In the case of older installations, consideration may not have been given to matching requirements, and a line may have been used which has a widely different impedance from the impedance of the receiver input circuit. When an old television receiver is traded in on a new receiver, the customer usually fails to consider possible differences in receiver input systems. In some cases, coaxial lines are installed on the assumption that they will help minimize severe noise pickup at street level; but, since coaxial lines have a considerably lower

*Tube Dept., Radio Corporation of America.

(A)

TYPE OF LINE	CHARACTERISTIC IMPEDANCE
Twin Lead	300, 150, 75 ohms
Coaxial RG-8/U, RG-58/U	53 ohms
Coaxial RG-11/U	75 ohms
Twisted Pair	Large Variation

(B)

Receiver	Input Circuit Impedance	Type of Input
Admiral	300 ohms	Balanced
Andrea VJ12	72 ohms	Unbalanced
Belmont	300 ohms	Balanced
Bendix	300 ohms	Balanced
Consolidated	72 ohms	Balanced
Crosley 9-408	300 ohms	Balanced
Crosley 9-407 M	72 ohms	Unbalanced
Du Mont	72 ohms	Unbalanced
Emerson	300 ohms	Balanced
Fada	300 ohms	Balanced
Farnsworth	300 ohms	Balanced
Garod	300 ohms	Balanced
G-E 802	300 ohms	Unbalanced
G-E 810	300 ohms	Balanced
Hallicrafters	300 ohms	Balanced
Industrial TV, Inc.	72 ohms	Unbalanced
Magnavox	300 ohms	Balanced
Motorola	300 ohms	Balanced
National	300 ohms	Balanced
Philco	300 ohms	Balanced
Pilot TV37	300 ohms	Balanced
RCA	300 ohms	Balanced
Scott	72-75 ohms	Unbalanced
Stewart-Warner	300 ohms	Balanced
Stromberg-Carlson	72 ohms	Unbalanced
Tele-Tone	300 ohms	Balanced
U. S. TV Mfg. Co.	100 ohms	Balanced
Westinghouse	300 ohms	Balanced

Table 1. (A) Characteristic impedance of several types of television transmission lines. (B) Characteristics of television receiver input circuits.

characteristic impedance than that of many standard receivers, a mismatch may be encountered. In other cases, the line is selected to "match" the antenna impedance instead of the receiver impedance. This selection, however, is based upon incorrect conceptions and frequently leads to reflections along the line.

Determination of source of ghost

When ghosts are observed on the kinescope screen, it becomes necessary at the outset to distinguish between displaced images caused by multi-path reception, and those caused by line mismatch. These two types of ghosts can frequently be distinguished by switching the television receiver from one station to another. If ghosts are observed on one channel, but not on the other channels, multi-path reception is indicated.

If a reflecting surface is so situated that multi-path reception could be expected from all stations, the switching test will be inconclusive. In such a case, or if only one television station can be received at the time, the chart shown in Fig. 1 usually suffices to distinguish between ghosts caused by multi-path reception and those caused by line mismatch.

The chart is further useful because the switching test may be inconclusive even if more than one station is available. To obtain a mismatch ghost, a mismatch condition must exist at both antenna and receiver. Usually, in one channel, the antenna will practically match the line, although a mismatch exists in all other channels. Accordingly, a switching test might lead to the erroneous conclusion that a multi-path ghost is present, when, as a matter of fact, line reflection is taking place. Reference to the chart, however, will answer the question conclusively.

The displacement of a ghost image on a kinescope screen depends on the horizontal sweep width and upon the time delay between desired and undesired signals. This time delay has a definite relation to line length in the case of mismatch reflection, and a definite relation to path-length difference in the case of multi-path reception. In the case of line mismatch, the duration of the delay also depends upon the type of line used.

Because a multi-path ghost is considerably more displaced than a mismatch ghost (unless an unusually long line is used), differentiation is possible upon this basis. Furthermore, when a mismatch ghost is present, its displacement on the kinescope screen will check closely with the value found from Fig. 1 with respect to horizontal sweep width, type of line, and line length.

To use this chart, adjust the horizontal sweep width to correspond with the nearest value given in Fig. 1; measure the amount of displacement (leading edge of main image to leading edge of first ghost); and determine or estimate the length of the antenna transmission line. Find the displacement (D) on the vertical axis of Fig. 1 and the line length (L) on the horizontal axis. If the intersection of these two co-ordinates is approximately on the diagonal for the sweep width and type of transmission line used, then one may reasonably assume that the ghost is caused by a mismatch.

Let us take the following example: A displacement of 0.15 inch is observed on a 10BP4 10-inch kinescope; the sweep width is 8 inches; approximately 325 feet of RG-58/U coaxial cable connects the receiver to the antenna. Since L and D nearly intersect on the RG-58/U line for a 10-inch tube, the conclusion is drawn that the ghost is caused by mismatch. If the line is short and displacement of the ghost image is very small, it may not be possible to measure this displacement on the screen. The effect in this case is that of "fuzziness" or loss of definition and detail.

Matching considerations

Representative impedances for both receiver and line are shown in Table 1. It is evident that mismatch difficulties may arise from installation of

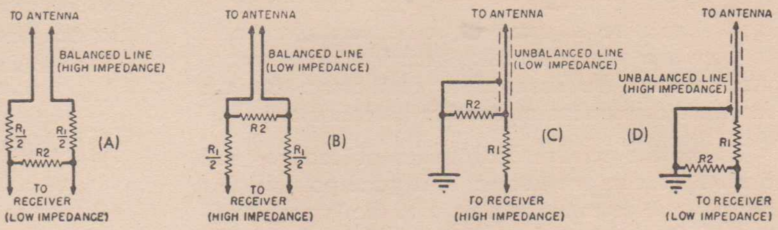

Fig. 2. Pad arrangements for balanced and unbalanced systems.

unsuitable lines, changing receivers, or mixing balanced and unbalanced systems. Television receivers have either balanced or unbalanced input systems, as diagrammed in Fig. 3. The unbalanced input system requires a coaxial line for proper operation; the balanced input requires a twin line (shielded or unshielded). A pair of coaxial lines, or a shielded balanced line provides a shielded balanced input. A twin-lead line provides an unshielded and balanced input.

Receivers with balanced input systems should be operated from balanced lines; receivers with unbalanced inputs should be operated from unbalanced (single coaxial) lines. If line-balance converters to operate over the TV band are not conveniently available, it will be necessary to make use of expedients when a balanced receiver is operated from an unbalanced line, or an unbalanced receiver is operated from a balanced line. These expedients are discussed in more detail at a later point.

RECEIVER IMPEDANCE (OHMS)	LINE IMPEDANCE (OHMS)				
	53	75	106	150	300
72	R_1 = 36 R_2 = 100		R_1 = 62 R_2 = 130	R_1 = 110 R_2 = 100	R_1 = 270 R_2 = 82
100	R_1 = 68 R_2 = 75	R_1 = 47 R_2 = 150		R_1 = 91 R_2 = 180	R_1 = 240 R_2 = 130
150	R_1 = 120 R_2 = 68	R_1 = 100 R_2 = 100	R_1 = 82 R_2 = 200		R_1 = 220 R_2 = 220
300	R_1 = 270 R_2 = 56	R_1 = 270 R_2 = 82	R_1 = 240 R_2 = 130	R_1 = 220 R_2 = 220	

All resistance values are in ohms.
(Pad resistances given in nearest five per-cent RMA preferred values.)
Example: A 300-ohm twin line is to be matched to a 75-ohm balanced receiver.
From table, R_1 = 270 ohms, R_2 = 82 ohms, $\frac{1}{2}R_1$ = 135 ohms.
Pad arrangement is obtained from Fig. 2A

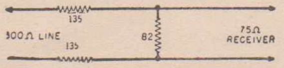

Table 2. Various pad arrangements required when antenna-receiver mismatch occurs.

Padding the line

When a mismatch ghost exists, it can be eliminated by matching the line impedance to the impedance of the receiver input circuit. Matching is accomplished by the insertion of a suitable carbon resistor pad between the line and the receiver input terminals. At television frequencies, both the line impedance and the receiver impedance are resistive for all practical purposes. Accordingly, simple resistive pads serve the purpose and maintain a practical match over all television channels. An unavoidable power loss, however, is the price which must be paid for ghost-free reception, when pads are used. If this insertion loss cannot be tolerated because of low signal level, the only remaining solution is to install a new transmission line which has the same characteristic impedance as the receiver. In many cases, however, the insertion loss of the pad can be tolerated.

To obtain proper impedance relations with minimum insertion loss, L-type pads are recommended. In the case of receivers with balanced input circuits, half the total series resistance (see Fig. 2) is placed in each side of the line. For receivers with unbalanced input circuits the total series resistance is placed in the "hot" side of the line.

The question is sometimes asked why both series and shunt resistors are used to make up a pad. The answer is that two conditions are to be met: the line should "see" its own impedance when looking into the pad-plus-receiver, and the receiver should "see" its own impedance when looking into the pad-plus-line. The reason for the first condition has been explained above. If the receiver does not "see" its own impedance, the input circuit may be disturbed with corresponding impairment of performance.

When a pad is designed, reference should be made to Fig. 2 to determine the required circuit. The series resistance is placed in the high-impedance side of the system. In addition, the total series resistance is placed in the "hot" side of an unbalanced line, but in a balanced line, half the total series resistance is placed in each side. The shunt resistance is placed across the low-impedance side of the system.

Next, the values of R_1 and R_2 are determined. If the impedances of both the line and receiver are known, the resistance values can be found from Table 2. If the impedances are unknown, the values may be found by experiment. To determine the values of R_1 and R_2 experimentally, two (or three) potentiometers having at least 300 ohms of resistance are hooked up into the required pad circuit. Only carbon-type potentiometers should be used because wirewound elements have excessive inductance. The settings of the potentiometers are varied until the ghost disappears. Adjustment of the contrast control may be required to maintain satisfactory picture brightness. The values of potentiometer resistance are then measured with an ohmmeter, and the nearest values of small fixed carbon resistors are made up into a pad and permanently installed.

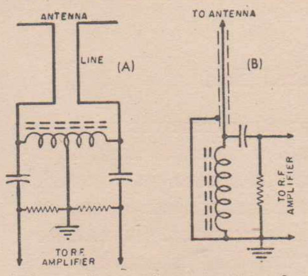

Fig. 3. Conventional balanced input circuit (A) and unbalanced circuit (B).

Inspection of Table 2 shows that each potentiometer must be adjusted to some particular value to obtain a proper match. That is, one potentiometer cannot be changed to compensate for the incorrect adjustment of another potentiometer.

Unless the relative magnitudes of receiver and line impedances are known, it is quite possible that the potentiometer pad will be tried the "wrong way" on the first trial, and no match will be found. In this case, the pad should be reversed and a match sought with the new connections.

If it is found impossible to eliminate the ghost with the variable pad, a defective transmission line is indicated. For example, there may be a break or short in one side of the line. Such defects can usually be detected by a visual or ohmmeter check.

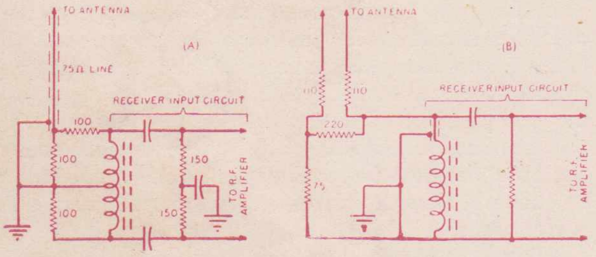

Fig. 4. (A) An unbalanced line matched to a balanced receiver input circuit. (B) A balanced line matched to an unbalanced receiver input circuit.

Balanced input, unbalanced line

When it is desired to operate a receiver with balanced input circuit from an unbalanced line, the input circuit may be arranged as shown in Fig. 4A. As an example, a 300-ohm balanced input circuit is shown in combination with a 75-ohm unbalanced line. One half of the input circuit is shunted with a 150-ohm resistor load. This 150-ohm dummy load provides correct loading for the unenergized half of the input. The other half of the input circuit is energized from the line through a pad designed to match 75 ohms to 150 ohms. The values for both resistances in this pad (as found in Fig. 4A) are 100 ohms.

Unbalanced input, balanced line

When it is desired to operate a receiver with an unbalanced input circuit from a balanced line, it is possible to use the arrangement shown in Fig. 4B. In this diagram, a 300-ohm balanced line is shown properly matched to a 75-ohm receiver. It will be observed that a 75-ohm dummy load is provided to maintain line balance to ground, thus preserving the antenna and line characteristics.

The receiver and dummy load together present a 150-ohm resistance to the 300-ohm line. Consequently, a pad which utilizes two 110-ohm resistors and one 220-ohm resistor is required to match the 300-ohm line to the input system. The pad indicated in Table 2 for matching a 300-ohm impedance to a 150-ohm impedance is comprised of two 220-ohm resistors. Because the line is balanced, however, the 220-ohm series resistance is split into two 110-ohm sections, with one section in each side of the line.

It should be noted that the noise-rejection feature claimed for a balanced line is lost in the circuit of Fig. 4B, because the receiver is energized from only one side of the line. Moreover, the matching losses are doubled when a dummy load is used. Thus, in the circuits given in Fig. 4A and Fig. 4B, the operating losses are about 12 db. This figure, however, is only about 6 db. more than the power loss encountered if a pad were not used. In strong-signal areas, satisfactory ghost-free reception can be obtained with these matching systems. In weak-signal areas, however, it may be necessary to forego these expedients and replace the transmission line with a line having characteristics which match the receiver.

RECEIVER IMPEDANCE (OHMS)	LINE IMPEDANCE (OHMS)				
	53	75	106	150	300
72	L_p = 5 L_m = 0.3 L_o = 4.7		L_p = 6 L_m = 0.6 L_o = 5.4	L_p = 8 L_m = 2.1 L_o = 5.9	L_p = 12 L_m = 6.9 L_o = 5.1
100	L_p = 7 L_m = 1.7 L_o = 5.3	L_p = 5 L_m = 0.4 L_o = 4.6		L_p = 6 L_m = 0.7 L_o = 5.3	L_p = 10 L_m = 4.5 L_o = 5.5
150	L_p = 10 L_m = 4 L_o = 6	L_p = 8 L_m = 1.9 L_o = 6.1	L_p = 5 L_m = 0.5 L_o = 4.5		L_p = 8 L_m = 1.9 L_o = 6.1
300	L_p = 13 L_m = 9 L_o = 4	L_p = 11 L_m = 6.5 L_o = 4.5	L_p = 10 L_m = 4 L_o = 6	L_p = 8 L_m = 1.9 L_o = 6.1	

L_p Power loss due to insertion of pads in db.
L_m Loss due to mismatch in db.
L_o Additional operating loss under matched (ghost-free) conditions in db

Table 3. Power losses in typical unmatched and matched systems.

It is interesting to compare the power loss due to mismatch with the power loss caused by the insertion of a matching pad. Such a comparison is given in Table 3. It can be seen from this tabulation that the power loss due to the addition of the pad for ghost-free reception is between 4 and 6 db. Except in critical areas of low signal strength, this loss is not troublesome. In critical areas, however, it is apparent that replacing the transmission line with one that matches the receiver may not only result in ghost-free reception, but in a signal increase of as much as 13 db.