The
Marconi Review

No. 112 **1st QUARTER 1954** **Vol. XVII**

CONTENTS:

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THE MARCONI REVIEW

No. 112. Vol. XVII. *1st Quarter, 1954.*

Editor: L. E. Q. WALKER, A.R.C.S.

TOLERANCES IN PARAMETERS OF MICROWAVE LENSES

BY D. H. SHINN, M.A. Ph.D., and T. C. CHESTON, B.Sc., A.M.I.E.E.

Simple formulae are derived from which the constructional and electrical tolerances of microwave lenses may be found. It is shown how some errors may be corrected by re-focusing. Tolerances in "phase-corrected reflectors" are treated in an Appendix.

Introduction.

THE designer of microwave lenses must be able to specify the constructional and electrical tolerances. J. R. Risser(1) has derived the most important of these. The following treatment is a generalisation and extension of his work. It also shows how some consistent constructional or electrical "errors" (e.g. change of wavelength) may be corrected by re-focusing.

The maximum permissible errors resulting from the tolerances depend on the application. In most cases an error of $\pm \frac{\lambda}{8}$ is permissible, or even conservative. The effects of various phase errors on the radiation pattern are discussed elsewhere(2, 3, 4).

Whilst the complete formulae for tolerances are very complicated(5), it is possible to obtain simple approximations provided that:—

(*a*) the lens is thin (i.e. fully stepped), or designed so that the rays in the lens are parallel to the axis of the lens, and

(*b*) the focal length of the lens is greater than about 0·6 times the aperture width.

For lenses which do not satisfy both of these conditions the results below must be applied with caution.

Some Fundamental Relations.

The following symbols are used:—

λ = wavelength.

d = electrical path length.

μ = refractive index.

b = plate separation in a metal plate lens, or inner dimension of tube in a square metal tube lens.

f = focal length.

a = aperture width.

F = f/a.

t = thickness.

ρ = radial distance from a point on the inner surface of the lens to the axis of the lens (the inner surface is the surface nearer the focus).

s = axial distance from a point on the inner surface of the lens to the focus.

ε = allowable error in electrical path length measured in wavelengths.

p = number of zone, i.e., number of integral wavelengths by which the electrical path length through the lens exceeds the electrical path length through the pole of the lens.

The electrical path length difference which has to be corrected by the lens can be shown to be approximately $\rho^2/2f$. This expression is exact when the inner surface of the lens is a paraboloid whose focus is at the focus of the lens. If this surface is plane, or a sphere with centre at the focus of the lens, the approximation involved gives adequate results when $F = 1{\cdot}5$, but leads to considerable errors when $F = 1$. These errors will be discussed in Appendix II, but will be neglected elsewhere.

Thus, for a thin lens, or for any lens in which the rays in the lens are parallel to the axis, the electrical path length from the focus through the lens to a reference plane beyond the lens is given by

$$d \doteq \frac{\rho^2}{2f} - (1 - \mu)t - p\lambda + k \tag{1}$$

where k is a constant. In designing the lens the thickness t is so chosen that d is a constant. In order to find the tolerances in the parameters of the lens it is necessary to find the conditions that the error in d shall not exceed $\varepsilon\lambda$. The fundamental formula from which these conditions may be derived is obtained by differentiating equation (1). This yields

$$\delta d = \frac{\rho\delta\rho}{f} - \frac{\rho^2\,\delta f}{2f^2} - (1 - \mu)\,\delta t + t\delta\mu - p\delta\lambda \tag{2}$$

It may be deduced from equation (1) that the maximum value of p is given by

$$p_{max} = \frac{\rho_{max}^2}{2f\lambda} - 1 \qquad = \frac{a}{8F\lambda} - 1 \tag{3}$$

In a metal lens $\mu^2 = 1 - \left(\frac{\lambda}{2b}\right)^2$ Hence, $\frac{\mu\,\delta\,\mu}{1 - \mu^2} = -\frac{\delta\lambda}{\lambda} + \frac{\delta b}{b}$ (4)

Thus, for a metal lens, equation (2) may be written

$$\delta d = \frac{\rho\,\delta\rho}{f} - \frac{\rho^2\,\delta f}{2f^2} - (1-\mu)\,\delta t + \frac{(1-\mu^2)t}{\mu}\left(\frac{\delta b}{b} - \frac{\delta\lambda}{\lambda}\right) - p\delta\lambda \qquad (5)$$

A fully stepped lens need only correct for path length differences up to one wavelength. Hence the greatest variation in thickness is given by

$$t_{\max} - t_{\min} \doteqdot \frac{\lambda}{|1-\mu|} \qquad (6)$$

TABLE I. FORMULAE FOR TOLERANCES.

Each of the errors below produces an error in path length of $\pm\frac{\lambda}{8}$

Lens	$\frac{\delta\lambda}{\lambda}$	Bandwidth	$\delta\mu$	$\frac{\delta b}{b}$
Dielectric, stepped	$\mp\frac{F\lambda}{a - 8F\lambda}$	$\frac{200F\lambda}{a}$ (if $a >> 8F\lambda$)	$\pm\frac{\mu-1}{8}$	
Metal, stepped	$\mp\frac{F\lambda\mu}{a\mu - 8F\lambda}$	$\frac{200F\lambda}{a}$ (if $a\mu >> 8F\lambda$)		$\pm\frac{\mu\lambda}{8(1+\mu)\,(\lambda + (1-\mu)t_{\min})}$
Dielectric unstepped	∞	∞	$\pm\frac{F\lambda}{a}(\mu - 1)$	
Metal, unstepped	$\mp\frac{F\lambda}{a}\frac{\mu}{1+\mu}$	$\frac{200F\lambda}{a}\frac{\mu}{1+\mu}$		$\pm\frac{F\lambda}{a}\frac{\mu}{1+\mu}$

Common to all lenses $\delta f = \mp F^2\lambda$, $\delta t = \pm\frac{\lambda}{8(\mu-1)}$, $\delta s = \mp\frac{F^2 a^2\lambda}{4\rho^2}$, $\delta\rho = \pm\frac{Fa\lambda}{8\rho}$

It may be deduced from equation (1) that the greatest variation in thickness of an unstepped lens is given by

$$t_{\max} - t_{\min} \doteqdot \frac{a}{8F|1-\mu|} \qquad (7)$$

For the purpose of finding errors in path length, the value of the minimum thickness does not matter, provided that the refractive index is constant over the lens. It may therefore be taken to be zero except for the purpose of determining the tolerance in plate spacing (or tube width) in a metal lens.

In the remainder of this article, equations (1) to (7) are used to obtain formulae for tolerances. These formulae are summarised in Table I for a maximum error of $\pm\lambda/8$.

Tolerances and Re-focusing for Unstepped Lenses.

The relevant formulae are (2) and (5), with $p = 0$, and (7). For a lens made without errors in ρ or t,

$$\delta d = -\frac{\rho^2}{2f^2}\delta f + \left(\frac{\rho^2}{2f\,|1-\mu|} + t_{\min}\right)\delta\mu \tag{8}$$

For a metal lens

$$\delta d = -\frac{\rho^2}{2f^2}\delta f + \left\{\frac{\rho^2(1+\mu)}{2f\mu} + \frac{(1-\mu^2)\,t_{\min}}{\mu}\right\}\left\{\frac{\delta b}{b} - \frac{\delta\lambda}{\lambda}\right\} \tag{9}$$

From equations (8) and (9) all the tolerances and conditions for re-focusing can be deduced.

Bandwidth; Re-focusing for Change in Frequency.

A dielectric lens has no bandwidth limitation. The bandwidth of a metal lens is given by

$$\frac{\delta\lambda}{\lambda} = -\frac{8F\lambda}{a}\frac{\mu}{1+\mu}\varepsilon \qquad \therefore \text{bandwidth} = \frac{1600\,F\lambda}{a}\frac{\mu}{1+\mu}\varepsilon\,\% \tag{10}$$

It is possible to compensate for a change in wavelength by re-focusing. The requisite change in focal length is given by

$$\frac{\delta f}{f} = -\frac{1+\mu}{\mu}\frac{\delta\lambda}{\lambda} \tag{11}$$

Refractive Index; Re-focusing for Change in Refractive Index.

For a dielectric lens, where the refractive index is constant throughout the lens, the tolerance in this refractive index is given by

$$\delta\mu = \frac{8\lambda F\,(\mu-1)}{a}\varepsilon \tag{12}$$

An error in dielectric constant can be corrected by changing the focal length by δf, where

$$\frac{\delta f}{f} = \frac{\delta\mu}{\mu-1} \tag{13}$$

Plate Spacing; Re-focusing for Constant Error in Plate Spacing.

For a metal lens the tolerance in plate spacing is given by

$$\frac{\delta b}{b} = \frac{8F\lambda}{a}\frac{\mu}{1+\mu}\varepsilon \tag{14}$$

provided that $a >> 8F\,t_{\min}\,(1-\mu)$

An error in plate spacing throughout the lens (e.g., due to the use of a thickness of metal different from that for which the lens was designed) can be corrected by the change

$$\frac{\delta f}{f} = \frac{1+\mu}{\mu}\frac{\delta b}{b} \tag{15}$$

Tolerances and Re-focusing for Fully Stepped Lenses

The relevant formulae are (2), (3), (5) and (6). For a lens made without errors in ρ or t,

$$\delta d = -\frac{\rho^2\,\delta f}{2f^2} + t\,\delta\mu - p\,\delta\lambda \tag{16}$$

where $t_{max} - t_{min} = \dfrac{\lambda}{|1 - \mu|}$, and $p_{max} = \dfrac{a}{8F\lambda} - 1$

For a metal lens

$$\delta d = -\frac{\rho^2}{2f^2}\,\delta f + \frac{t\,(1 - \mu^2)}{\mu}\left(\frac{\delta b}{b} - \frac{\delta\lambda}{\lambda}\right) - p\,\delta\lambda \tag{17}$$

Bandwidth; Re-focusing for Change in Frequency.

The bandwidth of a dielectric lens is given by

$$\varepsilon\,\lambda = -\left(\frac{a}{8F\lambda} - 1\right)\delta\,\lambda$$

$$\text{If } \frac{a}{8F\lambda} \gg 1,\ \text{bandwidth} = \frac{1600F\lambda}{a}\,\varepsilon\% \tag{18}$$

The bandwidth of a metal lens is given by

$$\varepsilon\,\lambda = -\left(\frac{a}{8F\lambda} - 1\right)\delta\,\lambda - \frac{1+\mu}{\mu}\,\delta\,\lambda$$

$$\text{If } \frac{a\mu}{8F\lambda} \gg 1,\ \text{bandwidth} = \frac{1600F\lambda}{a}\,\varepsilon\,\% \tag{19}$$

If the above inequalities hold, it is possible to compensate for a change in wavelength by altering the focal length by δf, where

$$\frac{\delta f}{f} = -\frac{\delta\lambda}{\lambda} \tag{20}$$

Dielectric Constant and Plate Spacing.

For a dielectric lens the tolerance in dielectric constant is given by

$$\delta\mu = (\mu - 1)\,\varepsilon \tag{21}$$

For a metal lens, the tolerance in plate spacing is given by

$$\frac{\delta b}{b} = \frac{\mu\lambda}{(1+\mu)\,(\lambda + (1-\mu)\,t_{min})}\,\varepsilon \tag{22}$$

Neither an error in μ nor an error in b can be corrected by re-focusing.

Tolerances and Re-focusing Common to Stepped and Unstepped Lenses

Focal Length; Thickness of Lens.

Tolerances in focal length and thickness are given by

$$\delta f = -8F^2\lambda\,\varepsilon \tag{23}$$

and

$$\delta t = \frac{\lambda}{(\mu - 1)}\,\varepsilon \tag{24}$$

Axial and Radial Position of a Tube.

In the construction of a metal tube lens errors may occur in the positioning of the individual tubes. Displacing any one tube axially by an amount δs is equivalent to changing the focal length with respect to that tube. The tolerance in s is therefore given by

$$\delta s = \frac{-2f^2\lambda}{\rho^2}\,\varepsilon \tag{25}$$

The tolerance in the radial displacement of a tube is given by

$$\delta\rho = \frac{f\lambda}{\rho}\,\varepsilon \tag{26}$$

For the extreme tubes, these tolerances are

$$\delta s_{\max} = -\,8F^2\lambda\,\varepsilon \qquad\qquad \delta\rho_{\max} = 2F\lambda\,\varepsilon$$

Re-focusing for Build-up Error.

If, due to errors in construction, there is a continuous linear increase of $\delta\,\rho$ with ρ (" build-up " error) this can be corrected by re-focusing. The requisite alteration is given by

$$\delta f = \frac{2f}{\rho_{\max}}\;\delta\,\rho_{\max} = 4F\;\delta\,\rho_{\max} \tag{27}$$

APPENDIX I

Tolerances for " Phase Corrected Reflectors "

Metal tube lenses have been designed with the outer surface (i.e., the surface further from the focus) short circuited, so that the energy is reflected. The electrical path length for such a lens can be shown to be

$$d \doteqdot 2s + \frac{\rho^2}{2f} + 2\,\mu t + p\lambda \tag{1a}$$

If the outer or inner surface is approximately a sphere with centre at the focus, then

$$s \doteqdot f - \frac{\rho^2}{2f}$$

If this surface is a plane then s is a constant.

For either of these surfaces

$$p\;\max = \pm\left(\frac{a}{8F\lambda} - 1\right) \tag{3a}$$

The value of the bandwidth depends on the number of zones and therefore on the shape of the lens surface. For either of the above surfaces the bandwidth is the same as for a conventional lens. For a paraboloid the bandwidth is infinite.

The other tolerances may be deduced from equations (1a) and (3a), together with

$$t_{\max} - t_{\min} = \frac{\lambda}{2\mu} \tag{6a}$$

They are the same as those for a conventional lens with the exception of

$$\frac{\delta b}{b} = \frac{\mu^2\,\lambda}{(1-\mu^2)\;(\lambda + 2\mu t_{\min})}\;\varepsilon \tag{22a}$$

$$\delta t = \frac{\lambda}{2\mu}\,\varepsilon \tag{24a}$$

$$\delta s = \frac{1}{2}\,\lambda\,\varepsilon \tag{25a}$$

Re-focusing is always possible to compensate for change in wavelength. The amount of re-focusing required depends on the shape of the lens surface. For the above two surfaces this is the same as for a conventional lens, due regard being paid to sign.

APPENDIX II

Errors in the Expressions Deduced.

The electrical path length which has to be corrected by the lens is not exactly $\rho^2/2f$. The exact path length depends on the shape of the lens. As examples of the errors which occur due to this approximation two special types of lens will now be treated more accurately.

In passing, it is worth remarking that the "second order errors", i.e. those obtained from the second partial differential coefficients of the path length d, are negligible except in extreme cases. For example, in re-focusing a lens to compensate for a change in wavelength, the maximum "second order error" is about $\frac{a}{8F}\left(\frac{\delta\lambda}{\lambda}\right)^2$, for a fully stepped lens with a spherical profile. Thus for a 20% change in wavelength, and a lens of aperture and focal length equal to forty wavelengths, the maximum second order error is 0·2 wavelengths. This is just a significant amount. The sum of the first and second order errors can be made to be zero if the amount of re-focusing is changed by about 20%.

Tolerances, etc. For a Fully Stepped Lens with a Spherical Profile

The centre of the sphere is the focus of the lens. The path length to be corrected is $f - \sqrt{f^2 - \rho^2}$, to which $\frac{\rho^2}{2f}$ is only a first approximation. Equations corresponding to those in the main discussion are as follows (the other equations in the main discussions are either unchanged or inapplicable):

$$d = f - \sqrt{f^2 - \rho^2} - (1 - \mu)t - p\lambda + k \tag{1b}$$

$$\delta d = \frac{\rho\,\delta\rho}{\sqrt{f^2 - \rho^2}} - \left(\frac{f}{\sqrt{f^2 - \rho^2}} - 1\right)\delta f - (1 - \mu)\,\delta t + t\delta\mu - p\delta\lambda \tag{2b}$$

$$p_{\max} = \frac{f - \sqrt{f^2 - \rho_{\max}^2}}{\lambda} - 1 \tag{3b}$$

$$\delta d = \frac{\rho\,\delta\rho}{\sqrt{f^2 - \rho^2}} - \left(\frac{f}{\sqrt{f^2 - \rho^2}} - 1\right)\delta f - (1 - \mu)\,\delta t + \frac{(1 - \mu^2)t}{\mu}\left(\frac{\delta b}{b} - \frac{\delta\lambda}{\lambda}\right) - p\delta\lambda \tag{5b}$$

$$\delta d = -\left(\frac{f}{\sqrt{f^2 - \rho^2}} - 1\right)\delta f + t\,\delta\,\mu - p\delta\,\lambda \tag{16b}$$

$$\text{Bandwidth of dielectric lens} = \frac{400\lambda}{a\,(2F - \sqrt{4F^2 - 1}) - 2\lambda}\,\varepsilon\,\%$$

The first approximation to this expression is given in equation (18).
The next approximation can be shown to be given by

$$\text{bandwidth} \doteqdot \frac{1600\lambda F}{a}\left(1 - \frac{1}{16F^2} + \frac{8\lambda F}{a}\right)\varepsilon\,\% \tag{18b}$$

For a metal lens

$$\text{bandwidth} \doteqdot \frac{1600\lambda F}{a}\left(1 - \frac{1}{16F^2} + \frac{8\lambda F}{\mu a}\right)\varepsilon\ \% \qquad (19b)$$

While the bandwidth can be conveniently found from a specification of the maximum allowable parabolic phase error, $\varepsilon\lambda$, at the edge of the lens, the re-focusing required to compensate for change in bandwidth introduces complicated phase changes over the aperture. The best amount of re-focusing depends on the amplitude distribution in the aperture. If the distribution is uniform, the amount of re-focusing required to compensate for a change in wavelength $\delta\lambda$ is given approximately by

$$\frac{\delta f}{f} = -\frac{\delta\lambda}{\lambda}\left(1 - \frac{1}{9F^2} - \frac{5F\lambda}{a}\right) \qquad (20b)$$

For a tapered amplitude distribution the amount of re-focusing is between the values given by (20) and (20b). For a metal lens the last term in (20b) is $5F\lambda/a\mu$. Other tolerances are:

$$\delta f = -\frac{\varepsilon\lambda\sqrt{f^2 - \rho^2}}{f - \sqrt{f^2 - \rho^2}} \qquad = -\frac{\sqrt{4F^2 - 1}\lambda}{2F - \sqrt{4F^2 - 1}}\varepsilon \qquad (23b)$$

$$\delta s = -\frac{\sqrt{f^2 - \rho^2}\lambda}{f - \sqrt{f^2 - \rho^2}}\varepsilon \qquad (25b)$$

$$\delta\rho = -\frac{\sqrt{f^2 - \rho^2}\lambda}{\rho}\varepsilon \qquad (26b)$$

$$\delta f = -\frac{\rho\,\delta\rho}{f - \sqrt{f^2 - \rho^2}} \qquad (27b)$$

Thus, when $F = 1$, the tolerances in focal length and axial position are less by 19·2% than the approximate expressions, the tolerance in radial position is less by 12·5%, and the re-focusing correction for build-up is less by 6·7%. The re-focusing correction for a change in wavelength is less by $\left(11 + \frac{500\lambda}{a\mu}\right)$ % for a metal lens, and by $\left(11 + \frac{500\lambda}{a}\right)$ % for a dielectric lens. The bandwidth of a metal lens is greater than or less than the approximate expression according to whether $\frac{a}{\lambda}$ is less or greater than $\frac{128}{\mu}$. The above figures are, of course, subject to a second order correction of magnitude about $\frac{\delta\lambda}{\lambda}$, $\frac{\delta f}{f}$ or $\frac{\delta\rho}{\rho}$. The expressions (21) and (22) for tolerances in dielectric constant and in plate spacing depend only on the approximation $t_{\max} - t_{\min} = \lambda/|1 - \mu|$. This approximation holds well for all fully stepped lens.

Plane Profile Fully Stepped Lens

It may easily be shown that

$$d = \sqrt{f^2 + \rho^2} - f - (1 - \mu)t - p\lambda + k \qquad (1c)$$

$$\delta d = \frac{\rho\,\delta\rho}{\sqrt{f^2 + \rho^2}} - \left(1 - \frac{f}{\sqrt{f^2 + \rho^2}}\right)\delta f - (1 - \mu)\,\delta t + t\delta\mu - p\,\delta\lambda \qquad (2c)$$

$$\rho_{\max} = \frac{\sqrt{f^2 + \rho^2} - f}{\lambda} - 1 \tag{3c}$$

Bandwidth of dielectric lens

$$= \frac{400\lambda}{a\,(\sqrt{4F^2+1} - 2F) - 2\lambda}\ \varepsilon\ \%$$

$$= \frac{1600\lambda F}{a}\left(1 + \frac{1}{16F^2} + \frac{8\lambda F}{a}\right) \varepsilon\% \tag{18c}$$

Bandwidth of a metal lens

$$\doteqdot \frac{1600\lambda F}{a}\left(1 + \frac{1}{16F^2} + \frac{8\lambda F}{a\mu}\right) \varepsilon\% \tag{19c}$$

The amount of re-focusing for a change in wavelength is given approximately by

$$\frac{\delta f}{f} = -\frac{\delta\lambda}{\lambda}\left(1 - \frac{0{\cdot}4}{F^2} + \frac{5F\lambda}{a}\right) \tag{20c}$$

for a dielectric lens. Other tolerances are:—

$$\delta f = -\frac{\lambda\sqrt{f^2+\rho^2}}{\sqrt{f^2+\rho^2} - f}\,\varepsilon \qquad = -\frac{\sqrt{4F^2+1}\,\lambda}{\sqrt{4F^2+1} - 2F}\,\varepsilon \tag{23c}$$

$$\delta s = -\frac{\sqrt{f^2+\rho^2}\,\lambda}{\sqrt{f^2+\rho^2} - f}\,\varepsilon \tag{25c}$$

$$\delta\rho = -\frac{\sqrt{f^2+\rho^2}\,\lambda}{\rho}\,\varepsilon \tag{26c}$$

$$\delta f = -\frac{\rho\,\delta\,\rho}{\sqrt{f^2+\rho^2} - f} \tag{27c}$$

Thus, when $F = 1$, the tolerances in focal length and axial position are greater by 18·4% than the approximate expressions, the tolerance in radial position is greater by 11·8%, and the re-focusing correction for build-up is greater by 5·9%. The re-focusing correction for a change in wavelength is less by $\left(40 - \frac{500\lambda}{a\mu}\right)$ % for a metal lens and $\left(40 - \frac{500\lambda}{a}\right)$ % for a dielectric lens. The bandwidth is greater by $\left(\frac{6{\cdot}25}{F^2} + \frac{800\lambda F}{a\mu}\right)$ % for a metal lens and $\left(\frac{6{\cdot}25}{F^2} + \frac{800\lambda F}{a}\right)$ % for a dielectric lens.

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SECONDARY BEAMS FROM METAL LENSES

By T. C. Cheston, B.Sc., A.M.I.E.E.

In the following article the conditions for the occurrence of secondary beams are derived from diffraction grating theory and applied to some radio metal lenses.

Introduction

THE radiation patterns obtained from two or more sources contain, in the general case, a number of beams whose maxima are angularly displaced. In particular, this applies to a number of adjacent sources formed by a parallel metal plate or "egg box" medium commonly used for radio lenses. Usually, only the primary beam is required and the others, the secondary beams, are undesirable. The primary beam is that beam which would be obtained if the medium were homogeneous, the secondary beams are those which arise from the periodicity of the structure which may give equi-phase fronts in other directions. The conditions for the occurrence of these beams and their directions are determined by the space and time phase relationships of the sources. The complete theory of secondary beams is not simple. It has been derived by various methods[1-4], which show that the power transmission coefficient through the medium rapidly decreases as secondary beams are formed. It is, therefore, of interest to know under what limiting conditions secondary beams occur, so that the design of radio metal lenses may be such as greatly to reduce them, if not to eliminate them completely. Simple diffraction grating theory gives these conditions and has been used in the following treatment.

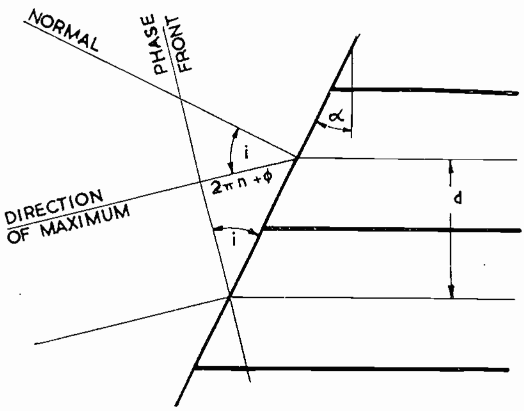

Fig. 1

General Analysis

Fig. 1 shows a metal plate medium cut at an angle α. ϕ is the phase difference between the adjacent sources which are spaced by $d \cdot \sec \alpha$. The direction of the maximum of any beam, defined by its angle i, is seen to be given by

$$\sin i = \frac{\phi + 2\pi n}{\frac{2\pi}{\lambda} d \cdot \sec \alpha}$$

where n is a positive or negative integer or zero. When $n = 0$ then $i = i_0$ where i_0 gives the direction of the maximum of the primary beam. Thus ϕ may be eliminated in the above expression giving

$$\sin i = \sin i_0 + \frac{n}{\frac{d}{\lambda}} \cos \alpha \qquad (1)$$

No secondary beams will occur if equation (1) has no real solution for $|n| \geqslant 1$. This gives the fundamental condition for no secondary beams

$$|\sin i_0| \leqslant \frac{\cos \alpha}{\frac{d}{\lambda}} - 1 \tag{2}$$

which implies $\cos \alpha \geqslant d/\lambda$.

From equation (2) it is apparent that secondary beams will invariably occur if the spacing, $d \cdot \sec \alpha$, exceeds a wavelength.

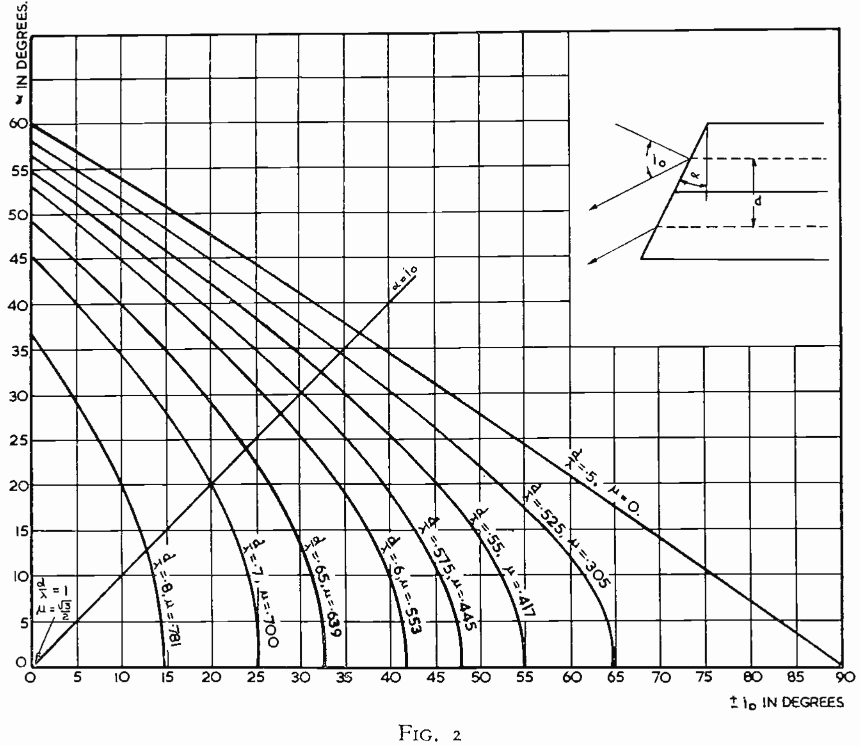

FIG. 2

Maximum possible values of i_0, α, $\frac{d}{\lambda}$ and μ for no secondary beams

No limitation has been set to the polarisation. However, if the polarisation is parallel to the plates, or if the medium is of " egg box " construction, then, ignoring the metal thickness,

$$\frac{d}{\lambda} = \frac{1}{2\sqrt{1 - \mu^2}}$$

where μ is the ratio of wavelength in air to wavelength in the medium (refractive index), and equation (2) may be re-written

$$|\sin i_0| \leqslant 2\sqrt{1 - \mu^2} \cos \alpha - 1 \tag{3}$$

which implies
$$\cos \alpha \geqslant \frac{1}{2\sqrt{1 - \mu^2}}$$

Fig. 2 correlates the maximum values of α, i_0, d/λ and, ignoring the metal thickness, μ, for which no secondary beams occur.

A case of particular interest with radio metal lenses is when the primary beam has a maximum in a direction parallel to the plates, i.e., when $i_0 = \alpha$. The condition for no secondary beams may then be written

$$\cos\alpha \geqslant \frac{2\dfrac{d}{\lambda}}{1 + \left(\dfrac{d}{\lambda}\right)^2} \tag{4}$$

or, ignoring the metal thickness, with the polarisation parallel to the plates or with " egg box " media,

$$\cos\alpha \geqslant \frac{4\sqrt{1-\mu^2}}{1 + 4(1-\mu^2)} \tag{5}$$

With radio metal lenses, the angle α for any part of the aperture is determined by μ and hence the conditions for no secondary beams may be found in general terms for any particular lens.

With metal plate lenses, secondary beams, if present, occur in the plane at right angles to the plates, the H-plane. With fully constraining lenses using the " egg box " construction, secondary beams occur in the two directions parallel to the walls. They may also occur at 45° to these directions, but less readily, since the effective spacing is reduced by a factor of $1/\sqrt{2}$.

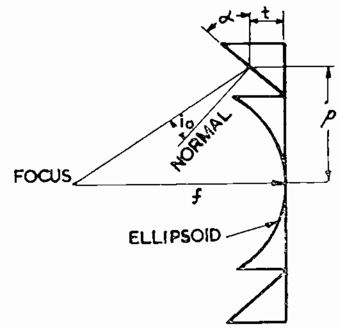

FIG. 3

Stepped Ellipsoidal-Plane Lens. (*Fig.* 3.)

Secondary beams will occur firstly on the stepped (inner) profile, since on the plane (outer) profile $\alpha = i_0 = 0$. With the notation shown, the equation of the lens profile is

$$\rho^2 = 2t\{f(1-\mu) - \mu p\lambda\} - t^2(1-\mu^2) + (p\lambda)^2 + 2fp\lambda \tag{6}$$

where p is the number of the step. After the first zone α remains substantially constant over a zone. To find α therefore, t may be taken as zero after differentiating equation (6), giving

$$\cot\alpha = \frac{d\rho}{dt} \doteqdot \frac{f}{\rho} - \mu\sqrt{\left(\frac{f}{\rho}\right)^2 + 1} \tag{7}$$

The angle i_0 is easily found

$$i_0 \doteqdot \alpha - \tan^{-1}\frac{\rho}{f} \tag{8}$$

Equations (3), (7) and (8) now give the maximum value of $\frac{\rho}{f}$ as a function of μ which conforms with the condition for no secondary beams. The relationship is shown in Fig. 6.

Sphero-Ellipsoidal "Egg Box" Lens. *(Radius equal to focal length.)* *(Fig. 4.)*

Secondary beams can be shown to occur firstly on the ellipsoidal (outer) profile provided $\frac{\rho}{f} < \cdot 85$.

The equation of the stepped ellipsoidal (outer) profile is

$$z = \frac{\mu f - p\lambda}{1 - \mu} - \frac{\mu}{1 - \mu}\sqrt{f^2 - \rho^2} \qquad (9)$$

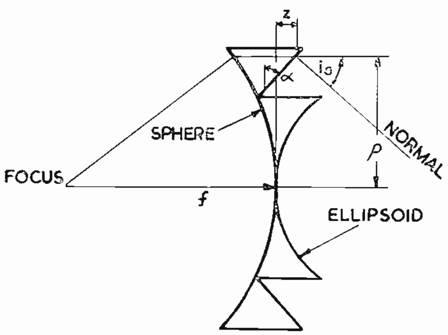

Fig. 4

Differentiation gives

$$\cot \alpha = \frac{1 - \mu}{\mu}\sqrt{\left(\frac{f}{\rho}\right)^2 - 1} \qquad (10)$$

Since $i_0 = \alpha$ equation (5) now yields as condition for no secondary beams

$$\frac{\rho}{f} < \frac{1}{\sqrt{\frac{1 + \mu}{1 - \mu}\left(\frac{4\mu}{3 - 4\mu^2}\right)^2 + 1}} \qquad (11)$$

The limiting values of $\frac{\rho}{f}$ are shown in Fig. 6.

Plane-Hyperboloidal "Egg Box" Lens *(Fig. 5.)*

Secondary beams can be shown to occur firstly on the hyperboloidal (outer) profile. The equation of the lens is

$$t = \frac{\sqrt{f^2 + \rho^2}}{1 - \mu} - \frac{f + p\lambda}{1 - \mu} \qquad (12)$$

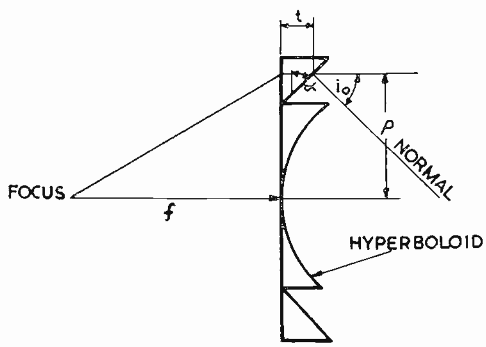

Fig. 5

Differentiation gives

$$\cot \alpha = (1 - \mu)\sqrt{\left(\frac{f}{\rho}\right)^2 + 1} \qquad (13)$$

Since $i_0 = \alpha$, equation (5) now yields as condition for no secondary beams

$$\frac{\rho}{f} < \frac{3 - 4\mu^2}{\sqrt{16 \cdot \frac{1 + \mu}{1 - \mu} - (3 - 4\mu^2)^2}} \qquad (14)$$

The limiting values of $\frac{\rho}{f}$ are shown in Fig. 6.

Correction for Metal Thickness

Fig. 6 gives the limiting values of $\frac{\rho}{f}$ for no secondary beams, for lenses where the metal thickness has been neglected. If it is taken into account, then the limiting values of $\frac{\rho}{f}$ will be a function of $\frac{d}{\lambda}$ as well as of μ since then

$$\frac{d}{\lambda} \not\doteq \frac{1}{2\sqrt{1 - \mu^2}}$$

Reading off values of $\frac{\rho}{f}$ against $\frac{d}{\lambda}$ in Fig. 6, will give a pessimistic answer, i.e., the limiting value of $\frac{\rho}{f}$ which leads to secondary beams is greater than that given.

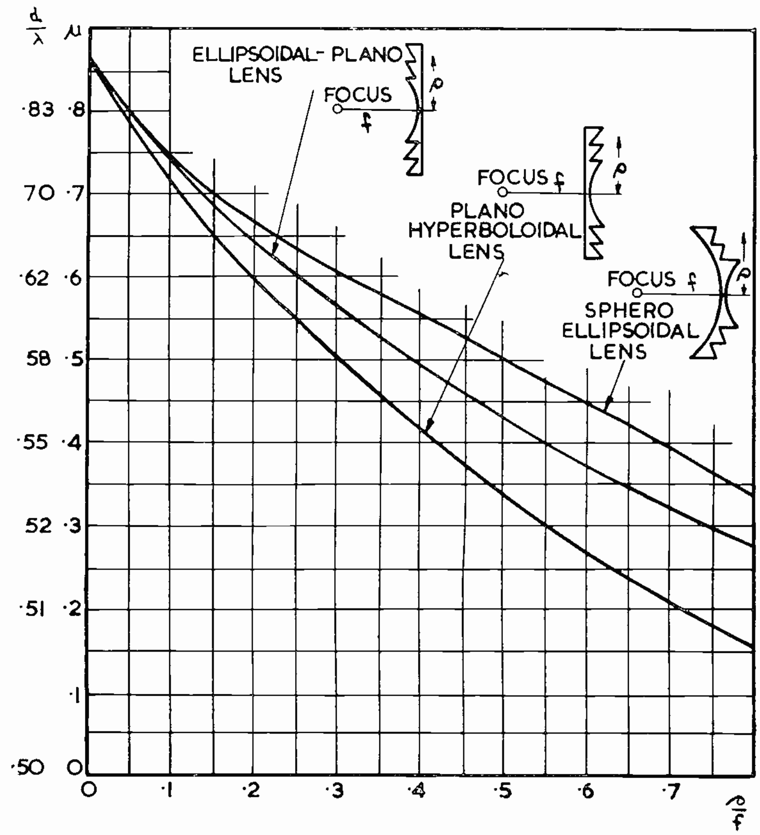

FIG. 6

Maximum values of $\frac{\rho}{f}$ for no secondary beams

This is due to the actual value of μ being smaller than that calculated by ignoring the metal thickness, and hence the medium of the lens is more powerful, requiring a smaller angle α. Similarly, the limiting values of $\frac{\rho}{f}$ read against μ will give an optimistic answer. The correct value will therefore lie somewhere between the two.

Correction for Scanning

With scanning the condition for no secondary beams is more severe for that half of the lens towards which the beam is being scanned and less severe for the other half of the lens, than without scanning. This is readily seen since, on the outer profile, the angle i_0 is increased or decreased by the scan angle. Similarly, with scanning, the angle of incidence on the inner profile is increased for one half of the lens and decreased for the other half.

Conclusions

The conditions for no secondary beams are readily found from equation (2) and Fig. 2. The directions of the maxima of secondary beams, if present, are given by equation (1).

Of the three types of radio metal lenses investigated, the sphero-ellipsoidal lens is least likely to generate secondary beams. When $\mu = \frac{1}{2}$ and $\rho/f \leqslant \frac{1}{2}$ (i.e., when the F number is $F \geqslant 1$) then secondary beams will not occur. The lens, however, is a wide angle lens and secondary beams may occur and certainly will occur if $\rho/f = \frac{1}{2}$ (i.e., if $F = 1$), on scanning. It seems impractical to design radio metal lenses which are free from secondary beams when used for scanning over wide angles, since low values of μ are undesirable from the point of view of match.

Acknowledgment

The author wishes to thank the Board of Admiralty for permission to publish this work which was carried out under an Admiralty Contract.

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HIGH POWER H.F. BROADCAST TRANSMITTERS

By D. F. Bowers and J. F. Ennos, B.Sc. (Eng.), A.C.G.I., A.M.I.E.E.

In the following article the problems involved in the design of a high power broadcast transmitter are considered, and the different requirements of broadcasting in the low, medium, and high frequency bands are stated. A general description of the H.F. Broadcast Transmitter type BD.253 follows, together with the design problems peculiar to a high frequency transmitter of this type.

Introduction

In past years, the position of Marconi's Wireless Telegraph Co., Ltd., in the field of high power short wave broadcasting has been largely maintained by the reputation that the 100 kW. High Frequency water-cooled transmitter type TBS.802 had justly earned for reliability and for ease of operation. It, was, however, a relatively expensive transmitter to purchase and to operate.

Immediately after the war, therefore, the requirements for high power broadcast transmitters were reviewed in the light of current development. A policy was decided upon aimed at reducing the initial and running costs of transmitters without lowering the quality of the equipment. In the execution of this policy a design for a series of 100 kW. transmitters was produced in such a way that three versions could satisfy the requirements for broadcasting in the low (160 kc/s–285 kc/s), medium (525 kc/s–1605 kc/s), and high frequency (5·9 Mc/s–26·1 Mc/s) bands. The three versions have a common design of modulator and power conversion equipment but a different design of the radio frequency circuits.

A number of problems common to each version of the 100 kW. transmitters have a most important bearing on the general design. The problems may be summarized under the headings of valve cooling, overall efficiency, noise level and rectifier design.

The previous design of 100 kW. transmitters employed water-cooled valves which permitted high anode dissipation but involved difficulties with the plumbing of the water cooling system, water insulating columns, heat interchangers, pumps and filters to say nothing of the attendant water flow interlock devices. At the time the design problems were being investigated for the new range of transmitters a range of air cooled valves were becoming available, although these were limited both as to maximum permitted anode dissipation and filament emission. It was decided that the merits of simplicity and ease of servicing which air cooling confers justified the use of air-cooled valves in the new design.

High overall efficiency of high power transmitters is required in order to minimize running costs. Previous designs of high power transmitters employed tungsten filament valves in the radio frequency output and modulator stages. The high filament power required by these valves placed a definite limit on the overall efficiency of the transmitter. It was decided therefore that the new transmitters should use valves then being developed by the English Electric Valve Co. which used thoriated tungsten filaments. The reduction in filament power required by these valves enabled a considerable increase in the overall efficiency of the transmitter to be obtained.

To minimize the level of hum in the transmitter output it was customary to heat the filaments of the final radio frequency stage by direct current. This is a relatively costly and inefficient way to minimize the hum and it was decided in the new design of transmitter to use alternating current for valve filament heating and to reduce the hum level by the use of feedback.

Previous designs of high power transmitters employ single cathode steel tank mercury arc rectifiers for the high tension supply. This type of rectifier has been

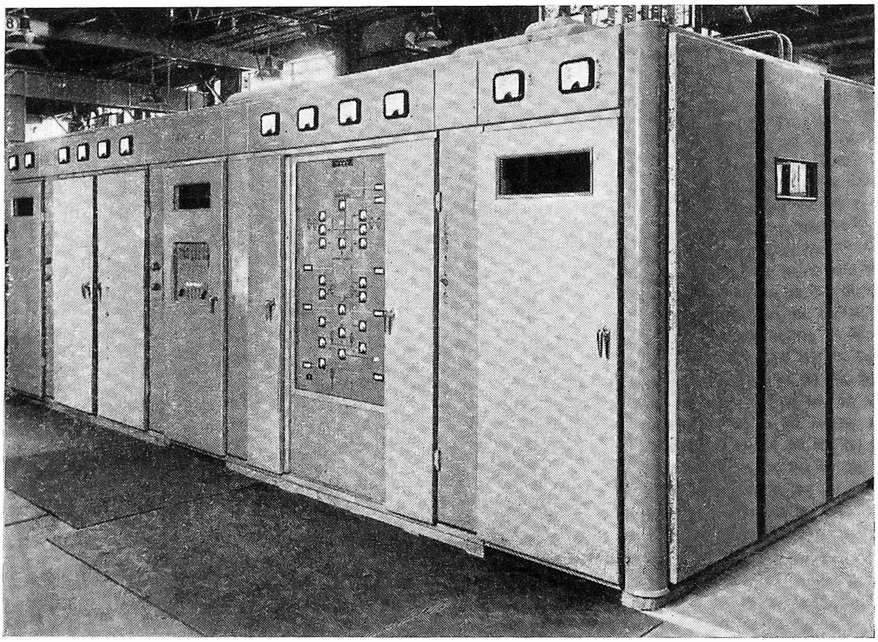

Fig. 1

developed to a high state of efficiency and gives excellent service. The use of a single cathode places limitations on the rectifier design and causes the associated high tension transformer to work with a low utilization factor. The alternative method of deriving the direct current high tension supply from the alternating current mains supply is to use hot cathode mercury vapour rectifiers. Six of these valves are connected in a three-phase full wave rectifier circuit which allows the high tension transformer to operate with a high utilization factor. Valves of this type have filaments and the disadvantage of being expendable. Further, no valves of the required size were available with grid control. The English Electric Co. were aware of these facts and had developed a new version of their proven mercury arc rectifier. This rectifier employs six mercury arc excitron type rectifier tubes which embody the advantages of the single cathode steel tank mercury arc rectifier and the hot cathode mercury vapour rectifier.

These rectifier tubes are of relatively small size and the complete rectifier occupies a smaller space than that required by the steel tank rectifier. In the new design of transmitter this rectifier was mounted within the transmitter enclosure.

The requirements of transmitters operating in the low, medium and high frequency bands differ considerably. These differences can largely be associated with the radio frequency circuits, and have a great influence on their design. Medium and low frequency broadcast transmitters operate on a fixed frequency which is rarely changed. High frequency transmitters, on the other hand, operate on a frequency which may be changed every few hours with variations in propagation conditions. It is essential that frequency changes should be accomplished in the shortest possible time and that they should be capable of being carried out by relatively unskilled personnel.

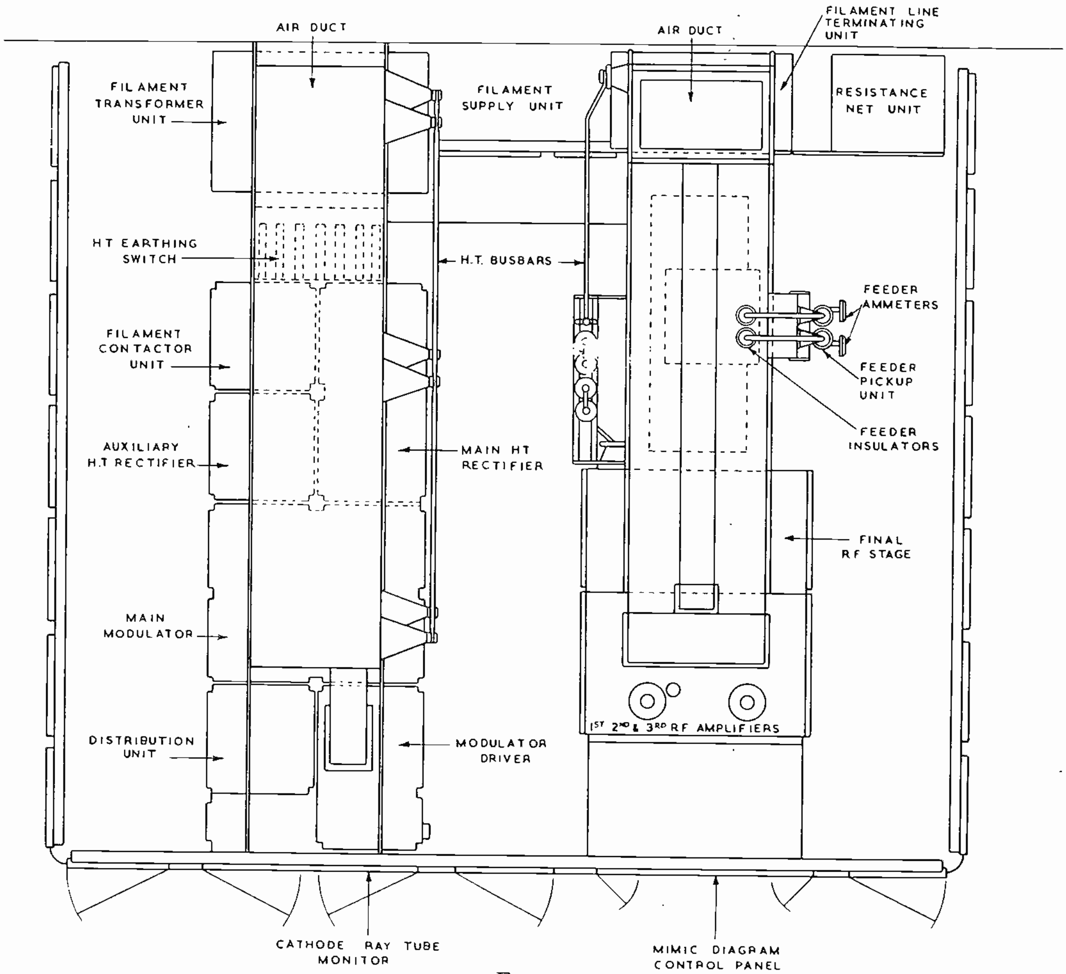

FIG. 2

The following description is confined to considerations of the high frequency version of the 100 kW. transmitter and to the circuits and problems encountered in its design.

General Description

The H.F. Broadcast Transmitter Type BD.253 is illustrated in Fig. 1. It gives a carrier power output of 100 kW. between 5·9 Mc/s and 18 Mc/s thereafter falling to 70 kW. at the highest frequency of 26·1 Mc/s. The carrier can be modulated by audio frequencies in the range 30–10,000 cycles per sec. with a low level of distortion, using the high level Class B modulation system.

Continuously variable tuning elements are used on all radio frequency stages except the final stage anode and feeder coupling circuits where it was found convenient to use plug-in coils. These coils cover the complete frequency range of the transmitter but are normally supplied to operate within the eight "broadcast" bands located between 5·9 and 26·1 Mc/s. Each set of coils is

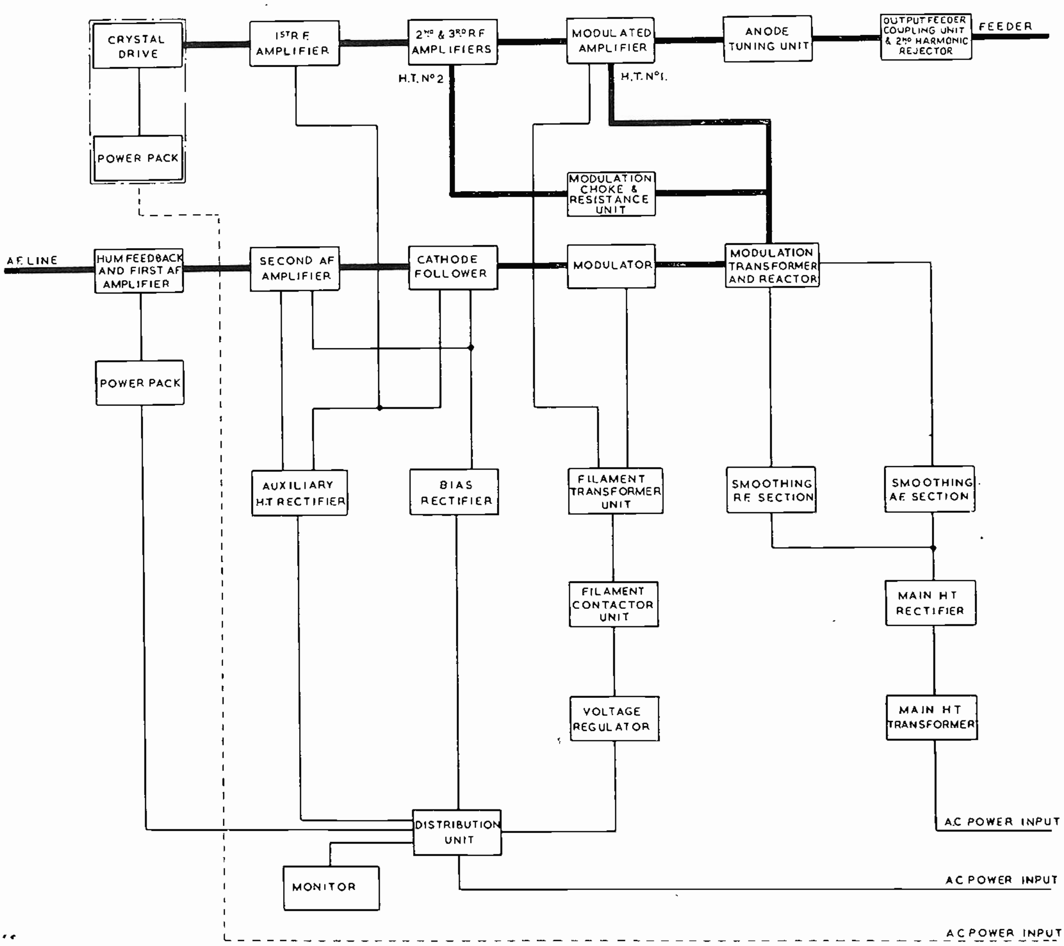

FIG. 3

mounted on two transporter racks which can be accommodated within the transmitter enclosure if desired.

The transmitter enclosure houses all the equipment with the exception of the main exhaust fan and oil filled modulation and high tension components. The position of the various units within the transmitter enclosure is shown on Fig. 2, and the electrical relation between them shown by the block schematic diagram of Fig. 3.

The transmitter is normally designed to operate from a three-phase 4-wire 440-volt 50 cycles per sec. supply, but it may be adapted to operate from high voltages. In cases where the instability of the mains supply voltage is greater than ± 5% special precautions are taken to ensure the correct functioning of the transmitter.

Radio Frequency Chain

General considerations suggest that the radio frequency section of the transmitter should comprise three main units, the radio frequency amplifiers, the central control panel or mimic diagram and a miscellaneous unit, in which would be mounted all components which for one reason or another could be moved away from either of the other two units.

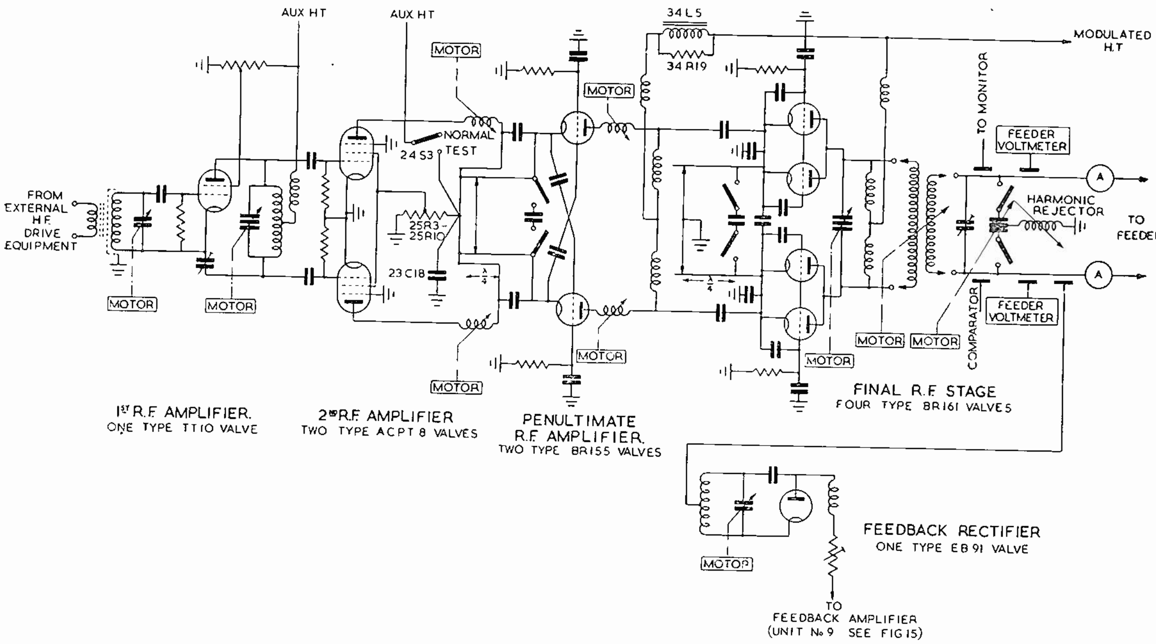

Fig. 4

The radio frequency amplifiers take an input at radiated frequency from an external drive unit at a level of 10 watts in a balanced 100 ohms circuit and amplify it to the rated output power with the following stages:—

First radio frequency amplifier:

One tetrode valve type TT.10 giving a balanced output.

Second radio frequency amplifier:

Two pentode valves type ACPT.8 operating in push/pull.

Penultimate radio frequency amplifier:

Two triode valves type BR.155 operating as a push/pull inverted amplifier.

Final radio frequency amplifier:

Four triode valves type BR.161 operating as a parallel push/pull inverted amplifier.

A simplified circuit diagram of the radio frequency circuits is given in Fig. 4.

To simplify the mechanical arrangements, all the tuning controls are operated electrically by means of standard motor units. These motor units not only enables hand tuning of the circuits to be done conveniently, but also provide for six pre-set spot frequencies to be set up to give rapid frequency changing.

Control of the radio frequency circuits is carried out from the mimic diagram control panel. As can be seen from Fig. 5 the mimic diagram consists of a highly simplified circuit diagram drawn on an insulating panel and mounted behind a protective glass window. Square faced current meters are mounted on the diagram at positions appropriate to the current being measured. Rectangular meters are used to indicate the position of the various variable tuning inductors and capacitors.

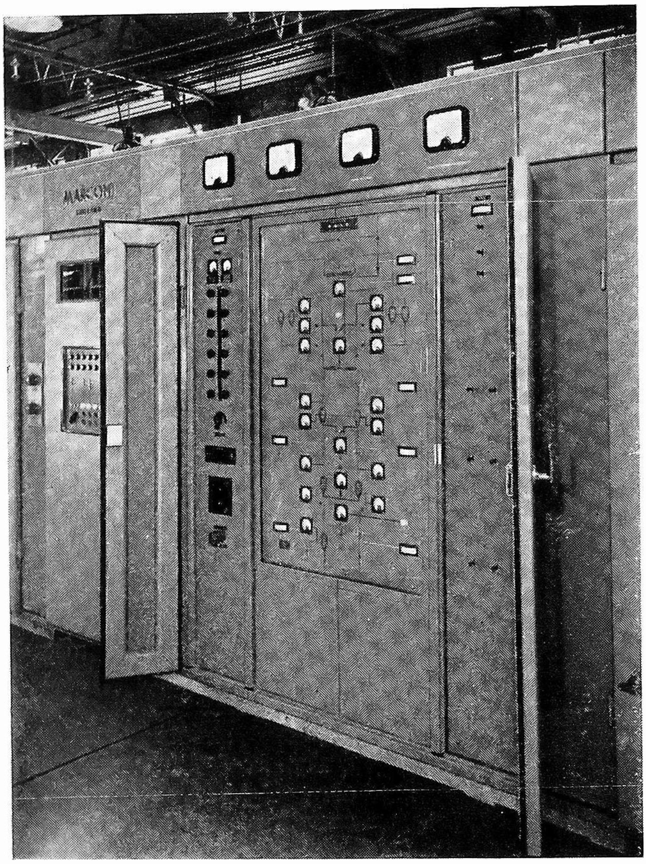

FIG. 5

The operation of the auxiliary and main H.T. rectifiers, the state of the radio frequency circuits and the operation of the overload trip circuits are shown by indicator lamps, mounted on appropriate positions on the diagram.

Behind the small door which is shown opened on the right of the mimic diagram are the keys for hand tuning the radio frequency circuits, while opening the left-hand door discloses all the controls associated with the selection of one of the six spot frequencies which have previously been set up on the transmitter. By means of these latter controls, all circuits, with the exception of the final stage output circuit may be tuned to a pre-selected frequency by the operation of a single switch.

At the rear of the transmitter are housed grid leaks, meter shunts, filament rectifiers and any other components which may be physically separated from the radio frequency circuits. The result is a clean layout in the radio frequency compartments of the transmitter which not only improves the appearance and access but also gives improved radio frequency performance.

The Final Radio Frequency Amplifier and Output Circuits

The choice of an inverted or grounded grid stage for the final radio frequency amplifier introduces several advantages over the more conventional grounded cathode amplifier. The inherent stability of a correctly designed grounded grid amplifier is superior to that of a grounded cathode amplifier and as a result a simpler layout of the high frequency circuits may be achieved. The fact that no neutralizing is required also simplifies the physical arrangements of the circuits and the consequent reduction in capacity from anodes to earth almost halves the anode circulating current. This allows the anode coils to be designed to give adequate coupling to the feeder coils and reduces them to a size more convenient for handling.

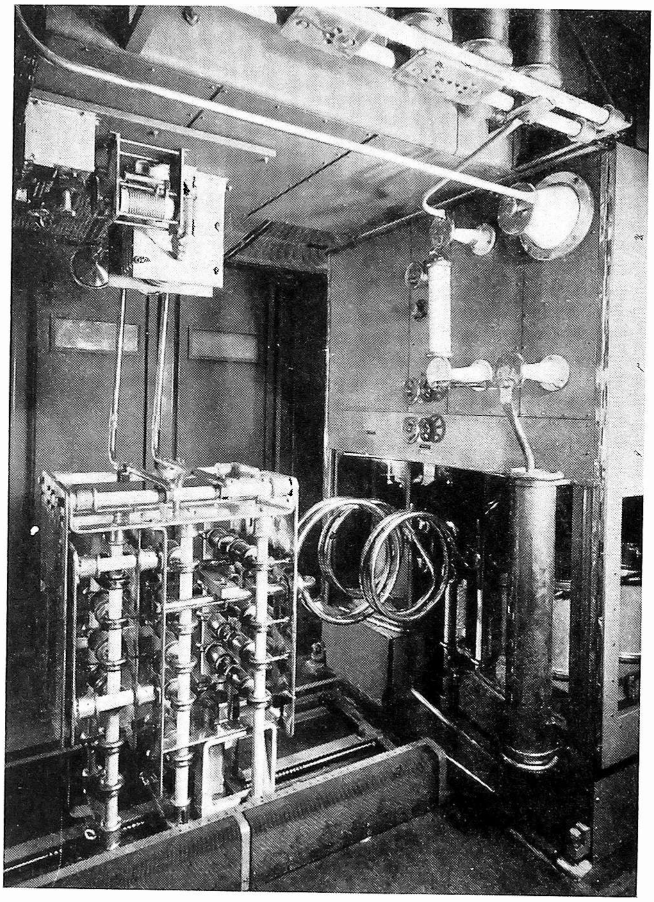

Fig. 6

Fig. 6 shows the final stage anode and output coupling assembly in a working position. It will be seen that the anode coils plug into chuck type holders mounted on the main anode block assembly. This block is designed to give equal loading to the two valves on the same side of the circuit, at the highest radio frequency used, where the valves may effectively be fixed at points of different radio frequency potential along the anode tuning inductance. At the end of the anode assembly opposite to the coil chucks is a variable tuning capacity, used for final adjustment to the tune of the final stage anode circuit.

The feeder tuning circuit is built up in the form of a truck made of bent aluminium sheets supported and insulated from earth by ceramic rod insulators. The coupling circuit inductances are secured to the " coupling truck " by a chuck similar to that used for securing the anode coils to the anode assembly. The self capacity of the " coupling truck " provides the output tuning capacity at the high frequency end of the transmitter frequency range, but at the lower end of the frequency range, additional capacity in the form of vacuum condensers is switched into circuit. The degree of coupling between the feeder and final stage valves is adjusted by traversing the " coupling truck " along a specially prepared rail by means of a pin which engages in a lead screw. This lead screw, which can be seen in Fig. 6 is rotated by a motor unit and controlled from the mimic diagram.

The outgoing feeder is connected to the transmitter coupling truck by means of a sliding, telescopic coupling which is designed to accommodate the movement of the coupling truck without affecting the feeder impedance. To the feeder terminals are connected the harmonic rejector unit, feeder ammeters, and feeder pick-up units. The voltages obtained by use of these pick-up units are used to operate feeder peak voltmeters, monitors, hum feedback and overload protection circuits.

The harmonic rejector is designed to provide a low impedance path to earth for any second harmonic frequency present in the output of the transmitter. It is in the form of a series tuned circuit connected between the feeders and earth and consists of a continuously variable inductance ganged to a variable condenser. The harmonic rejector is tuned by a standard motor unit controlled from the mimic diagram control panel. Using two ganged elements in this manner, a large frequency coverage is obtained from the unit without the use of range switching. The harmonic rejector is linked to the feeder by two hand operated switches. It is of interest to note that if these switches are open, the self capacity of the harmonic filter to the feeder is sufficient coupling for the harmonic rejector to operate satisfactorily at the higher frequencies. By choosing whether or not these switches are open, the frequency coverage of the harmonic rejector is increased and the load which it places on the transmitter, is decreased for a given rejection of harmonic. The harmonic rejector used on this transmitter may be tuned to resonance over a frequency range of 5 Mc/s to 52 Mc/s.

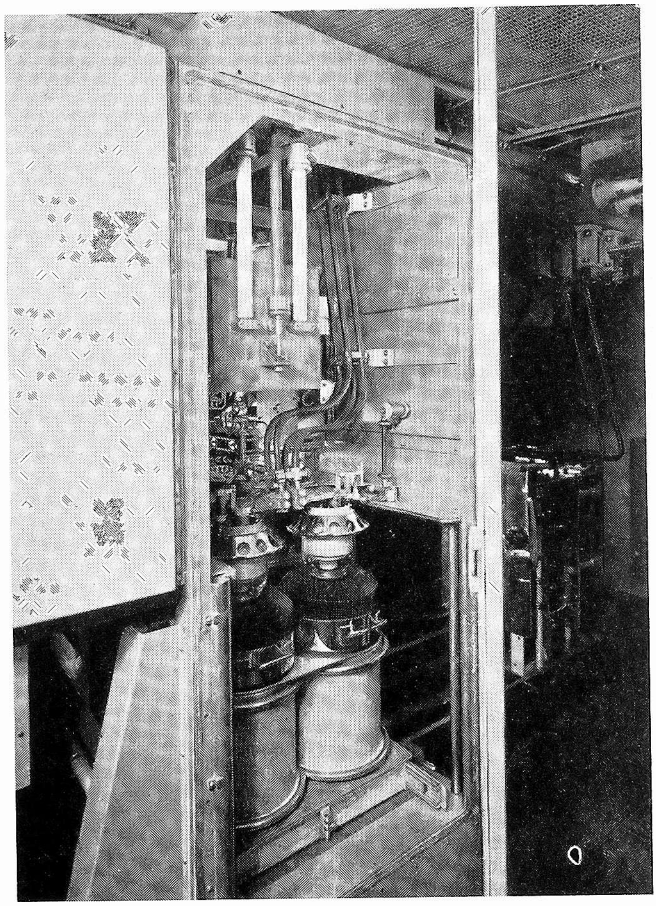

FIG. 7

The final stage cathode and grid circuits are shown on the simplified radio frequency circuit diagram in Fig. 4 and illustrated in Fig. 7. Grid bias for operating the valves as Class C amplifiers is obtained by means of grid leaks, the grid of the valve being earthed to radio frequency through a blocking condenser. Ideally this condenser would consist of a single unit but in practice due to the capacity requirement and the large radio frequency current to be carried it consists of several condensers in parallel spaced round the grid connector of the valve. These condensers are liable to form parallel resonant circuits with their paralleling connectors thus producing, at certain frequencies, a high impedance between the grid and earth instead of the required low impedance. This may cause the stage of self-oscillate, the circuit having the form of an unneutralized grounded cathode stage. In the H.F. Broadcast Transmitter Type BD.253, each final stage valve grid is connected to earth through seven condensers, each comprising two 0·003 micro farad condensers in series. These are arranged round the grid in such a manner that no parallel resonance between the condensers and their connectors exists within the frequency range of the transmitter. It is of interest to note that when investigated in the development stage no less than four different parallel resonances were detected in the final radio frequency grid to earth circuit, the variations being obtained by the different lengths of connections between pairs of condensers.

The impedance between the valve cathodes and earth, across which the radio frequency driving voltage is developed, comprises a balanced radio frequency line. The effective line length is varied by placing a shorting bar across it along its length and by inserting loading condensers across the input end of the line at the lower frequencies. Short circuits are placed across the line by means of electrically operated contactors spaced along its length and controlled by one of the switches shown immediately above the final stage anode assembly in Fig. 6. Insertion of capacity across the line at the lower end of the transmitter frequency range is controlled by another of these switches.

The electrical length of the lines is adjusted to be approximately $\frac{1}{4}$ wave at the frequency of operation and to minimize even harmonics in the final stage cathode driving wave an earth connection is made at the quarter wave point. The length is not critical, any reactance appearing at the valve end of the line due to its incorrect length, being absorbed into the π coupling between the penultimate and final radio frequency stages.

The cathode lines which are accommodated in the overhead air duct are also used as filament bus-bars for the final stage valves, the filament transformers being at radio frequency earth potential. Each of the two cathode lines, one for each side of the push/pull circuit, supplies filament power to two valves and comprise four coaxial cables whose inner conductor is of adequate cross section to carry the filament current of the valves. The lines feeding the valve filaments on one side of the circuit are shown in Fig. 7. De-coupling condensers are connected between the inner and outer conductor at the valve and transformer ends of each of the lines to enable the function of filament heating and cathode impedance to be separated between the inner and outer conductors respectively of the coaxial cables.

The Penultimate and Second Radio Frequency Amplifiers

The Penultimate radio frequency amplifier consists of two triode valves type BR.155 in a push/pull inverted amplifier circuit as shown on Fig. 4. The tuned anode circuit is of the π coupler type, the tuning element being a continuously variable inductance. This variable inductance is built up from a fixed coil of approximately thirteen turns supported from low loss ceramic rods. A contact within the periphery of the turns is rotated and moved axially along the coils by a travelling nut on a lead screw driven from a standard motor unit. This contact is of the single point high pressure type and has adequate thermal capacity to allow the heavy circulating current of this stage to be carried without overheating. Trailing contacts short circuit sections of the coil and minimize any spurious resonances in the unused section.

It has already been stated that the final radio frequency stage is operated as a modulated inverted amplifier, and to achieve full modulation of the carrier wave it would be necessary to modulate the penultimate radio frequency stage. Since the penultimate stage is also operating as an inverted amplifier, on theoretical grounds the transmitter would have to have three modulated stages and could require three sources of modulated H.T. at three different voltages with audio frequency modulation in the same phase.

Considering the valves used in the penultimate and second radio frequency amplifiers we have the BR.155 valve which when operated as a Class C anode modulated stage is limited to an H.T. voltage of 9 kV., and the ACPT.8 valve which is limited under similar conditions to an H.T. of 3 kV. In order to simplify the modulated H.T. supply arrangements the penultimate and second radio frequency amplifiers are operated in series from the modulated H.T. source used for the final radio frequency stage. This is shown on the simplified radio frequency circuit diagram in Fig. 4 and its effects will be discussed after the stages have been described. The anode of the penultimate stage therefore is connected to the H.T. source via the anode tuned circuit, the grid and cathode circuits being at a potential of about 3 kV. D.C. The grid circuit is treated, generally, as that of the final stage, the grid being connected to earth through a suitable arrangement of condensers, bias being provided by a grid leak. The filament supply and cathode impedance of the penultimate radio frequency amplifier follow the practice used on the final stage, the lines being carried in a separate duct and screened from the final stage cathode lines. The controls for varying the effective line length by the application of suitable shorting bars and the addition of loading condensers are also situated above the final stage anode assembly and can be seen on Fig. 6.

The second radio frequency amplifier is a conventional amplifier coupled by a π network to the penultimate stage. The tuning element of this network is a continuously variable inductance but in this case the coil, which is wound on an insulating support, is rotated bodily by a standard motor unit, the contact being moved axially along the coil by the action of the turns themselves bearing on an insulating disc.

The arrangement of the two stages in series can be better explained if we follow the path of the current from the H.T. rectifier through the penultimate amplifier valves and the cathode lines to the filament line terminating unit at the rear of the transmitter (see Fig. 4). After passing through metering and overload protection circuits the current divides, the proportion taken by the second radio frequency amplifier being carried in a co-axial cable running alongside the penultimate stage cathode lines to the second radio frequency amplifier valves, while the balance is returned to earth via a resistance (25R3–25R10). The purpose of the condensers (23C18), choke (34L5) and resistance (34R19) are explained later.

It was found that great difficulty might be experienced when tuning this series chain of radio frequency amplifiers unless one stage was near tune. With the two stages detuned the theoretical tune point could be indicated by maximum anode current as opposed to the more usual minimum. It is well known that as the tune point of a stage is approached, its impedance rises. Since we are dealing with two stages in series this rise in impedance results in a larger proportion of the total H.T. voltage appearing across the stage being tuned. The increase in anode current due to this increase in anode voltage can produce the indication of tune described above. The effect, however, is not consistent and it was realized that to an operator it could be most confusing. Switch (24S3) shown in Fig. 4 was inserted to connect the second radio frequency amplifier (at the earthy end of the series chain) to an auxiliary source of H.T. voltage of 1,800 volts. In this condition, the second radio amplifier can be tuned using normal methods, the interlock system of the transmitter preventing the main H.T. supply from being applied to the penultimate and final radio frequency stages. The second radio frequency amplifier having been tuned, the H.T. circuits are then restored to normal and main H.T. applied to the penultimate radio frequency amplifier which is then tuned without difficulty. With both states initially tuned in this manner final adjustment may be made to either circuit in the normal way.

As has already been stated it had been appreciated that on theoretical grounds the final, penultimate and second radio frequency amplifiers should all be modulated. However, it was found in practice, when modulating three stages, that although the harmonic distortion produced in the transmitter was low when the modulating frequency was low, it rose with increase of modulating frequency, the effect being noticeable at all except very low depths of modulation. It was also noticed that while the driving voltage applied to the cathode of the final stage rose with increase of modulation depth, when the modulating frequency was low, it fell when the modulating frequency was high. Every symptom indicated that the distortion was being caused by insufficient drive to the final radio frequency amplifier. It was suspected, and finally proved, that this was caused by a phase shift between the audio frequency modulating voltages applied to the penultimate and second radio frequency amplifier stages due to the capacity to earth from the grid and cathode circuits of the penultimate stage. Consequently, as the anode voltage of the penultimate stage was increasing to its maximum due to the rise of the modulating voltage, the radio frequency drive applied to the cathode of this stage was not rising in sympathy. Thus the amplitude of the penultimate stage output voltage was not allowed to reach the required value at the peaks of modulation. Attempts to balance this phase shift by adding suitable capacities across the penultimate radio frequency amplifier resulted in self-oscillation, the added capacity tuning with the leakage reactance of the modulation transformer. It was therefore decided to remove the modulation from the second radio frequency amplifier except at low modulation frequencies and this is done by the condenser (23C18) shown in Fig. 4. The audio frequency choke (34L5) and resistance (34R19) reduce the depth of modulation of the H.T. applied to the penultimate and second radio frequency amplifier chain to such a value that over modulation of the driving stages never takes place. With this modification the distortion produced in the transmitter was reduced to the values given in Figs. 16, 17 and 18.

Drive Unit

The drive unit type BD.454 is specially designed for use with this transmitter and will provide the necessary level of drive within the frequency range 3 Mc/s to 26·1 Mc/s with a frequency stability better than the requirement of the Mexico City conference, i.e., $\pm$ 20 cycles per sec. or approximately $\pm$ 7 parts in 10^7 at the highest radiated frequency. A standby variable frequency oscillator of comparatively low frequency stability is an optional feature of this drive unit which is designed for rapid frequency change to suit short wave working conditions.

The First Radio Frequency Amplifier

The first radio frequency amplifier illustrated in Fig. 8 is mounted on a separate pull-out chassis, its supplies being brought in on plugs and sockets. Drive is accepted from an external unit at a level of 10 watts, the input impedance of the stage being 100 ohms balanced to earth. This stage is designed to cover 3 Mc/s to 26·1 Mc/s in six switched ranges, switching being carried out by the transmitter frequency range switch. This switch selects an appropriate balanced to unbalanced tuned transformer which matches the drive input to the grid of the type TT.10 amplifier valve. The anode circuit, taps on which are also selected by the transmitter frequency range switch, gives, by the use of a circuit balancing capacity, a balanced output, to drive the second radio frequency amplifier, from a single valve. In both grid and anode circuits tuning within the switched ranges is by motor driven variable condenser operated from the mimic diagram. The standard motor unit which drives the first R.F. amplifier tuning condenser can be seen in Fig. 8.

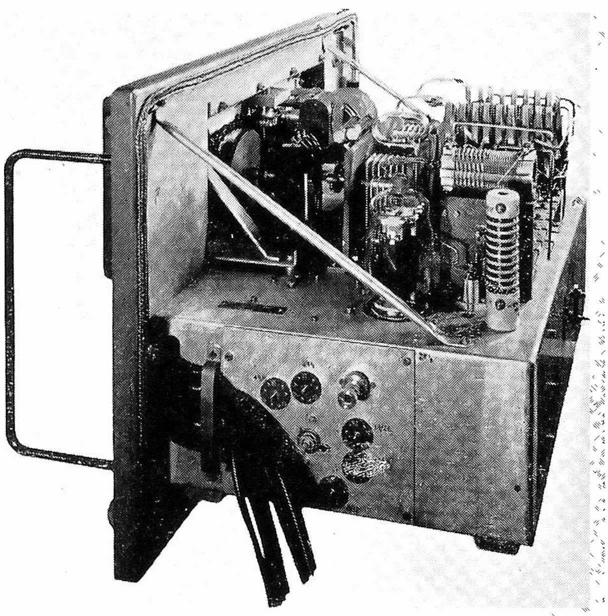

FIG. 8

The Audio Frequency Chain

The modulator for the series of 100 kW. transmitters was designed to satisfy the conditions imposed by its use with the three versions of the radio frequency circuits for low (160 kc/s–285 kc/s), medium (525 kc/s–1,605 kc/s) and high frequency (5·9 Mc/s–26·1 Mc/s) operation.

The 100 kW. M.F. transmitter uses a grounded cathode circuit in the final radio frequency amplifier and the penultimate stage is not modulated. At 100% modulation the modulator must deliver 66 kW. into a load of 920Ω. The shunt capacity presented by the output circuits is approximately 2,600 μμF which has a reactance of 6,150Ω at 10,000 c/s.

The modulator power output requirements at low modulating frequencies when used with the L.F. transmitter are comparable to the M.F. case. The conditions at higher modulation frequencies are more difficult than in the M.F. case as the shunt capacity is increased to 5,300 μμF having a reactance of 3,000Ω at 10,000 c/s. In addition, each installation will have other problems introduced by the aerial system. Economic considerations usually cause L.F. aerials to be of a comparatively high Q value which causes an asymmetrical load to be presented to the modulator at sideband frequencies. Whilst precautions are taken to minimize this effect it has to be considered when designing the modulator.

The high frequency transmitter has somewhat different requirements since both final and penultimate stages are modulated. At 100% modulation of the 100 kW. carrier the modulator must deliver 77 kW. into a load of 840Ω. When working at 26 Mc/s the transmitter output is reduced to 70 kW. and the modulator has to provide 58 kW. into a load of 1,110Ω. The shunt capacity is some 2,200 μμF which gives a reactance of 7,250Ω at 10,000 c/s.

The audio frequency amplifiers require an input level of 1 milliwatt in a balanced 600 ohm line to give 100% modulation of the carrier. A high level of negative feedback is applied between the anode of the main modulator stage and the grid of the first audio frequency amplifier. This not only improves the overall performance of the modulator as regards frequency response and harmonic distortion, but causes the audio input level required for a given depth of modulation to be virtually independent of the power of the carrier being modulated. This is of use when changing frequency from the lower to the upper end of the transmitter frequency

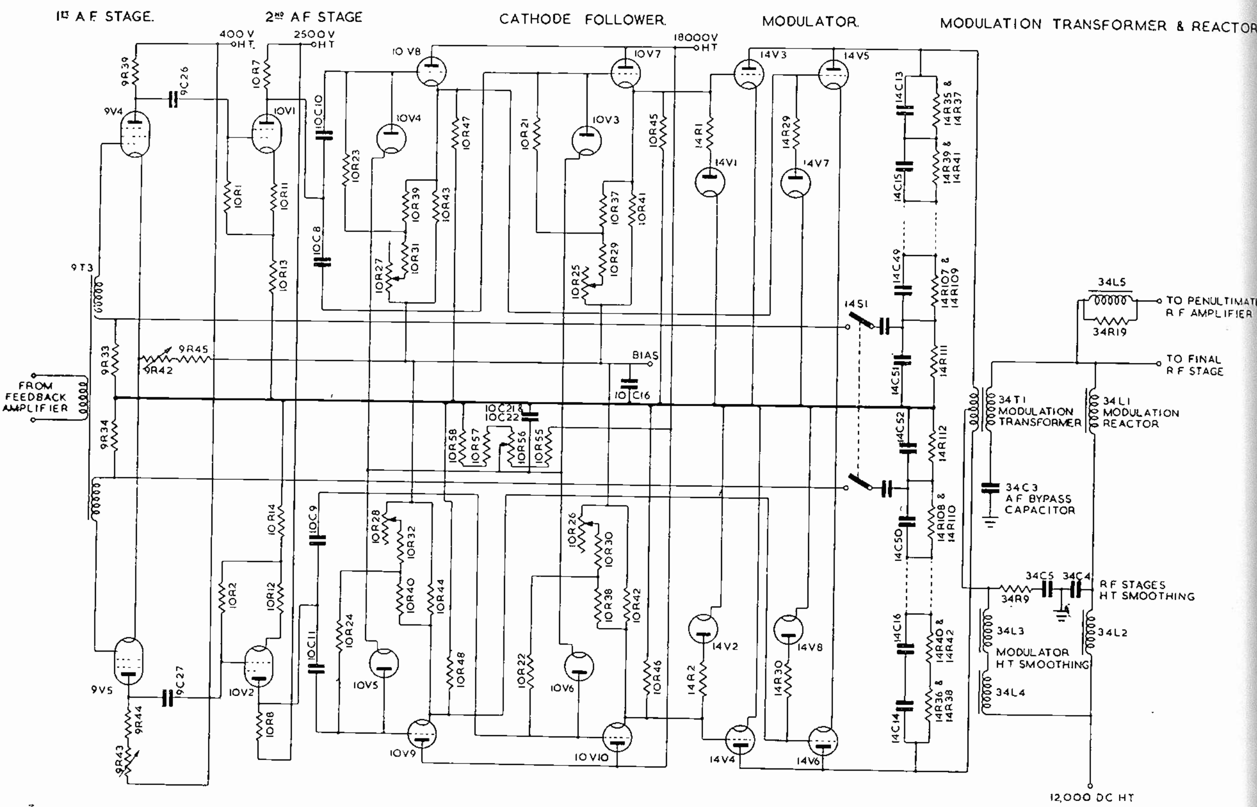

FIG. 9

range where there are limitations in the maximum permitted carrier power. With a reduced carrier, there is, of course, a comparable reduction in the audio frequency power required for 100% modulation, but no change is necessary in the audio input level. Tone line-up of the transmitter, therefore, is not necessary every time the radio frequency is changed.

A simplified circuit diagram of the audio frequency chain is shown in Fig. 9. The output of the feedback amplifier is transformer coupled to the first audio frequency amplifier, provision being made at this point for the injection of negative feedback voltages derived from the main modulator stage. The amplification of the two sides of the audio frequency chain is equalised at this point by means of a common cathode resistance (9R42 and 9R45) and a variable resistor (9R43) in the anode circuit of one of the first audio frequency amplifier valves. The second audio frequency amplifier employs tetrodes operating with internal negative feedback to improve the overall performance of the modulator. A cathode follower driving stage is interposed between the second A.F. amplifier and the main modulator. This circuit gives a low output impedance and allows the main modulator valves to be driven into the positive grid region of their characteristics without undue distortion of the A.F. signal. External controls allow individual adjustment of the bias on each cathode follower valve, and hence the bias on each modulator. By this means correction may be made to allow for differences in characteristics between individual valves which may then be set for their optimum operating points. Diode limiters are connected to the grids of these cathode follower valves and they are biased to prevent over-modulation of the carrier wave and the development of dangerous voltages within the transmitter.

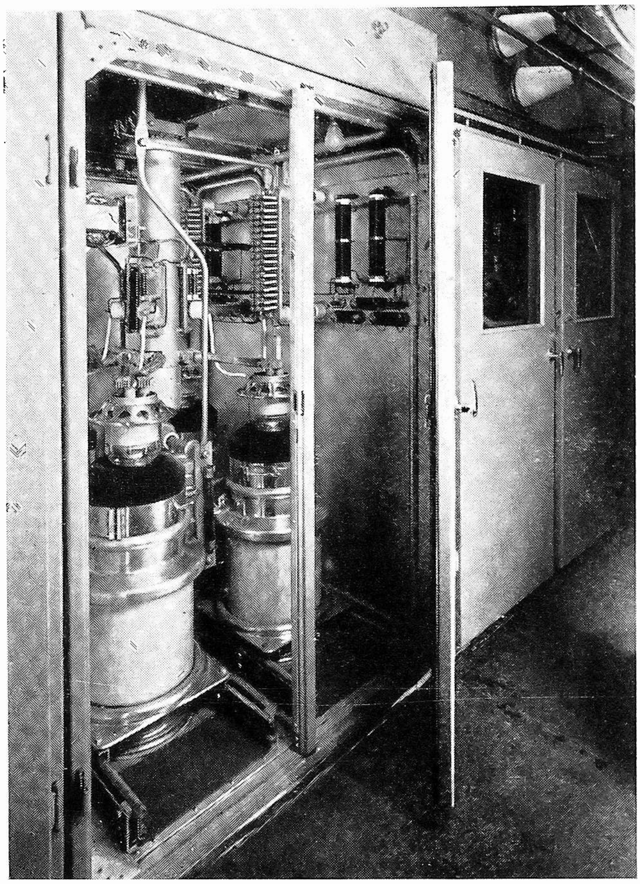

FIG. 10

The main modulator stage which can be seen in Fig. 10 comprises four triode valves type BR.161 in parallel push/pull, damping diodes to prevent dynatron type self-oscillation being connected between each valve grid and cathode. Across the modulator output is a resistance and capacity potentiometer network which provides a small proportion of the modulator output for negative feedback. External to the transmitter and connected to the modulator anodes by low capacity connectors is the push/pull Class B modulation transformer. This matches the modulator output to the load presented to it by the radio frequency stages using the conventional transformer, reactor and audio frequency blocking condenser combination.

Noise Modulation

In spite of the fact that alternating current is used to heat the filaments of the main valves, the level of noise modulation on the carrier wave is extremely good, being considerably better than 60 db below 100% modulation. To achieve this result the following precautions have been taken:—

(1) All directly heated filaments supplied with alternating current are fitted

with adjustable centre-point resistances to ensure the balanced operation of the valve filament.

(2) " Scott " and " Double Scott " connection of filament supplies. The primaries of the two penultimate radio frequency amplifier valve filament transformers are supplied with alternating voltages phased 90° apart from a pair of " Scott " connected auto transformers. The pairs of valves on the two sides of the main modulator stage are similarly treated. The filaments of the four valves in the final radio frequency amplifier are supplied with voltages phased 45° apart from a pair of " Double Scott " connected auto transformers. With the exception of the penultimate radio frequency amplifier filament transformers which are located in the filament line

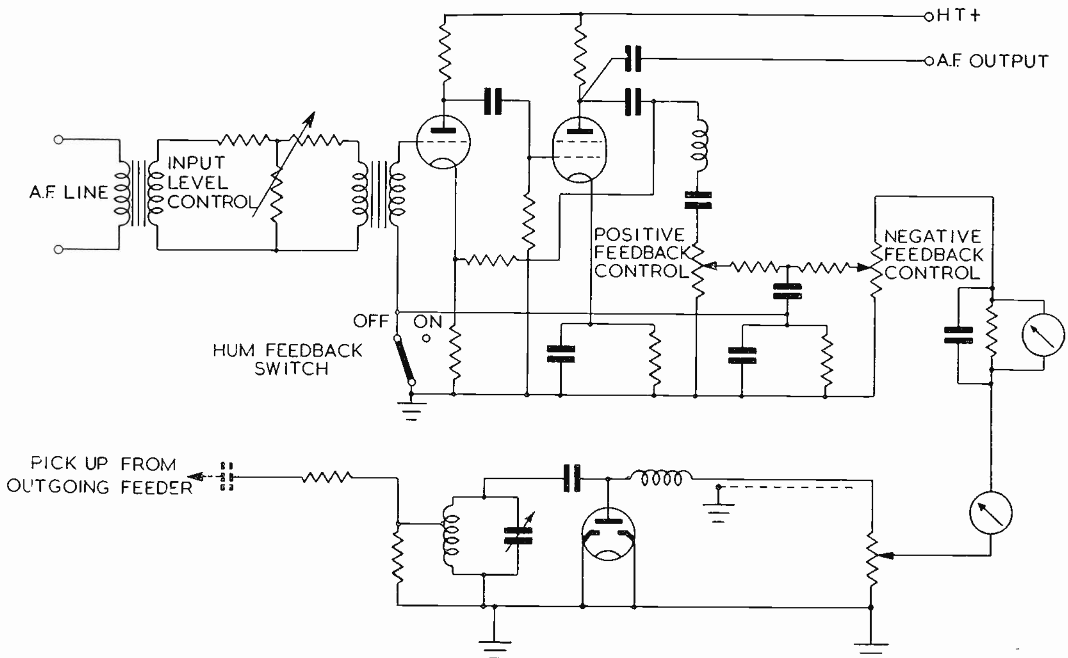

FIG. 11

terminating unit, all filament transformers and auto transformers are located in the filament transformer unit.

(3) Direct current heating of the filaments of the first and second radio frequency amplifiers to minimize both amplitude and phase modulated noise on the carrier wave. The direct current for heating these filaments is obtained from three-phase full wave metal rectifiers located in the filament supply unit.

(4) The use of envelope feedback. A sample of the transmitter output is obtained by means of a capacity pick-up on the feeder and after rectification, it is fed, at the correct level into the feedback amplifier. In the feedback amplifier the rectified output of the feedback rectifier unit is balanced against a sample of the modulator input voltage, as is shown in Fig. 11. The resultant is obviously any distortion product and noise which may have been generated in the transmitter. This resultant is filtered and restricted to noise frequency and fed back into the modulator input.

Transmitter Monitoring

In addition to the large number of meters indicating current and voltage in various parts of the transmitter a cathode ray tube monitoring unit is fitted in order to obtain visual monitoring of audio and radio frequencies. The monitor is mounted on runners within the modulator driver unit and is shown extended in Fig. 12. By means of rotary type co-axial switches, samples of audio frequency from different parts of the audio frequency chain may be displayed on the double beam oscilloscope If required, a sample of the transmitter radio frequency output may be impressed on one set of the "Y" plates of the oscilloscope and examined, either by displaying it against a normal time base, for investigation of the modulation envelope, or against a suitable phased sample of the audio input for examination of modulation trapezium patterns. The monitor unit includes amplifiers of low distortion and of sufficient gain to provide adequate voltage for a time base scan, and sufficient power to drive an external monitoring loudspeaker. The monitor also includes a low distortion rectifier which provides rectified radio frequency to operate an external distortion factor meter and a meter indicating percentage modulation. Using these amplifiers and rectifiers, one may also display a linearity curve of audio frequency input to the transmitter against rectified carrier output.

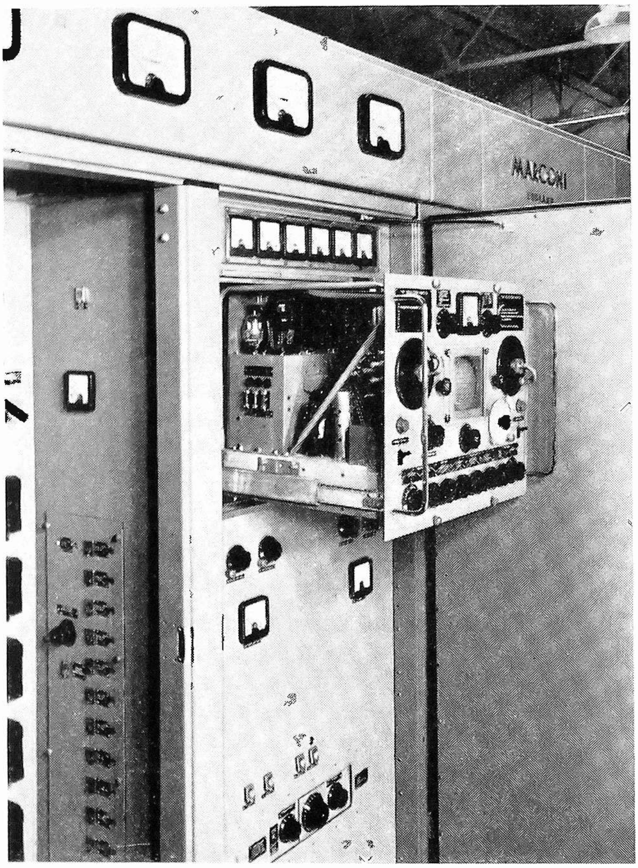

FIG. 12

Power Supplies

All power supplies required for operation of the transmitter are obtained from the three-phase four-wire A.C. input and they may be divided into four main classes.

(1) *Filament Heating.*—As already described, valve heating is mainly by alternating current, single-phase, two-phase "Scott" and four-phase "Double Scott" supplies being obtained from transformers located in the filament transformer unit at the rear of the transmitter. To ensure that the lives of the high power valves are not reduced by excessive current being passed through the filaments when they are cold, the high power filaments are switched on in two stages, current limiting reactors

being inserted in the primaries of the filament transformers for the first 20 seconds of the filament starting sequence. Direct current for heating the first and second radio amplifier filaments is obtained from three-phase full wave rectifiers located in the filament supply unit, the rectifiers also being energized via the current limiting reactors during the first 20 seconds of the filament starting sequence.

(2) *Modulator Bias Supplies.*—This is obtained from a conventional selenium rectifier in a bi-phase bridge circuit.

(3) *Auxiliary H.T. Supply of* 1,800 *Volts and* 2,500 *Volts.*—The 1,800-volt supply is obtained from a bi-phase rectifier circuit using two glass mercury pool arc rectifier tubes type AR.60. This 1,800-volt rectifier supplies the anode voltage for the first radio frequency amplifier and the audio frequency cathode follower stage. It is also used to provide H.T. for the initial tuning of the second radio frequency amplifier. The 2,500-volt supply which is of low current capacity only is required for the second audio frequency amplifier, and this is obtained by putting a 700-volt selenium bridge connected rectifier in series with the 1,800-volt supply.

(4) *Main H.T. at* 11,000 *Volts.*—This rectifier uses six grid controlled mercury pool arc rectifier tubes type AR.61 connected in a three-phase full wave circuit. The rectifier tubes are excited from an auxiliary D.C. source and voltages of one-third and two-thirds of the full value are obtained by removing this excitation from three or two of the tubes respectively. Voltage variations of five-sixths of the full value may be obtained by suitably shifting the phase of the grid control voltage with respect to the anode voltage. The main H.T. starting sequence, which is controlled by a motor driven cam sequence switch applies H.T. to the transmitter in three steps, to minimize surges on the incoming supply system, and to limit the peak current taken from the rectifier in charging the smoothing condensers. The H.T. starting sequence may be stopped at any of the three steps to allow low H.T. voltages to be applied to the transmitter for tuning and checking.

Associated with the rectifier is a high speed relay system which provides rapid suppression of the mercury arc in the event of any fault arising in the rectifier or the transmitter circuits. The high speed suppression system includes circuits to give automatic restoration of the high tension voltage a short time after a fault occurs. In the event of a fault being of a sustained nature, after a pre-selected number of attempts to restore, the main rectifier circuit breaker is opened and an alarm bell is rung.

The transmitter overload protection system, which is coupled into the main H.T. rectifier high speed arc suppression circuits, is of the automatic restoration and indication type. A voltage is obtained from the circuit to be protected by means of a low resistance in series with the protected circuit in the case of D.C. circuits, or from a current transformer in the case of A.C. circuits. If this voltage exceeds a pre-determined value it causes a high speed relay to operate. The operation of this relay causes the main H.T. rectifier arc suppression circuits to function and at the same time, lights a lamp on the transmitter control panel to indicate which part of the circuit is faulty. The fault having cleared by virtue of the H.T. supplies having been removed, the overload relay resets itself, but the indicator lamp remains alight until manually cancelled. Facilities are provided for testing the overload lamps without operating the arc suppression circuits.

In addition to overload protection on rectifier A.C. input circuits and in various valve circuits, the transmitter is further protected against the effect of flash-overs on the feeder or other faults which may result in the feeder presenting an incorrect load to the transmitter, or faults which cause the normal ratio of feeder voltage to radio frequency amplifier cathode current to be disturbed. A sample of the transmitter output is rectified and balanced continuously against a voltage derived from the cathode circuit of the final radio frequency amplifier. Should any fault arise, whether on the feeder or within the final radio frequency amplifier, which disturbs the balance by more than a pre-determined amount, a relay is energized which in turn operates the arc suppression circuits in the main H.T. rectifier.

FIG. 13

Operating conditions at 5·9 Mc/s.

Second R.F. stage:	Anode volts 3·35 kV.	
	Total cathode current 880 mA.	
Penultimate stage:	Total Anode current 1·15 amps.	
Final R.F. stage:	Cathode current left-hand front valve.	3·95 amps.
	Cathode current left-hand rear valve.	3·55 amps.
	Cathode current right-hand front valve.	3·65 amps.
	Cathode current right-hand rear valve.	3·55 amps.
Main H.T. voltage:	11·3 kV.	

Total anode current taken by R.F. stages. 13·5 amps.
Carrier power to aerial 100 kW.
Total power taken from A.C. mains:—
Carrier only. 230 kW. at 0·97 P.F.
Carrier 100% modulated at 1,000 c/s. 367 kW. at 0·96 P.F.
Transmitter overall conversion efficiency (carrier level): 45·2%.
Transmitter noise level: 64 db below 100% modulation.

FIG. 14

Operating conditions at 26·1 Mc/s.

Second R.F. stage:	Anode volts 3·25 kV.	
	Total cathode current 930 mA.	
Penultimate stage:	Total Anode current 1·35 amps.	
Final R.F. stage:	Cathode current left-hand front valve.	2·25 amps.
	Cathode current left-hand rear valve.	3·7 amps.
	Cathode current right-hand front valve.	1·9 amps.
	Cathode current right-hand rear valve.	3·4 amps.
Main H.T. voltage:	11·4 kV.	

Total anode current taken by R.F. stages. 10·3 amps.
Carrier power to aerial. 78 kW.
Total power taken from A.C. mains:—
Carrier only. 185 kW. at 0·98 P.F.
Carrier 100% modulated at 1,000 c/s. 279 kW. at 0·96 P.F.
Transmitter overall conversion efficiency (carrier level): 43·2%.
Transmitter noise level: 62 db below 100% modulation.

Safety Precautions

Safety of operating personnel is assured by a three-stage system of mechanical interlock. With the transmitter off, and the main transmitter isolator open, we have the first stage of interlocking with complete access to all parts of the transmitter. In this condition it is only possible to energize a 50-volt A.C. lighting system, a 50-volt D.C. relay system and a 6-volt indicator lamp system, but this is sufficient to enable the complete transmitter control system to be tested.

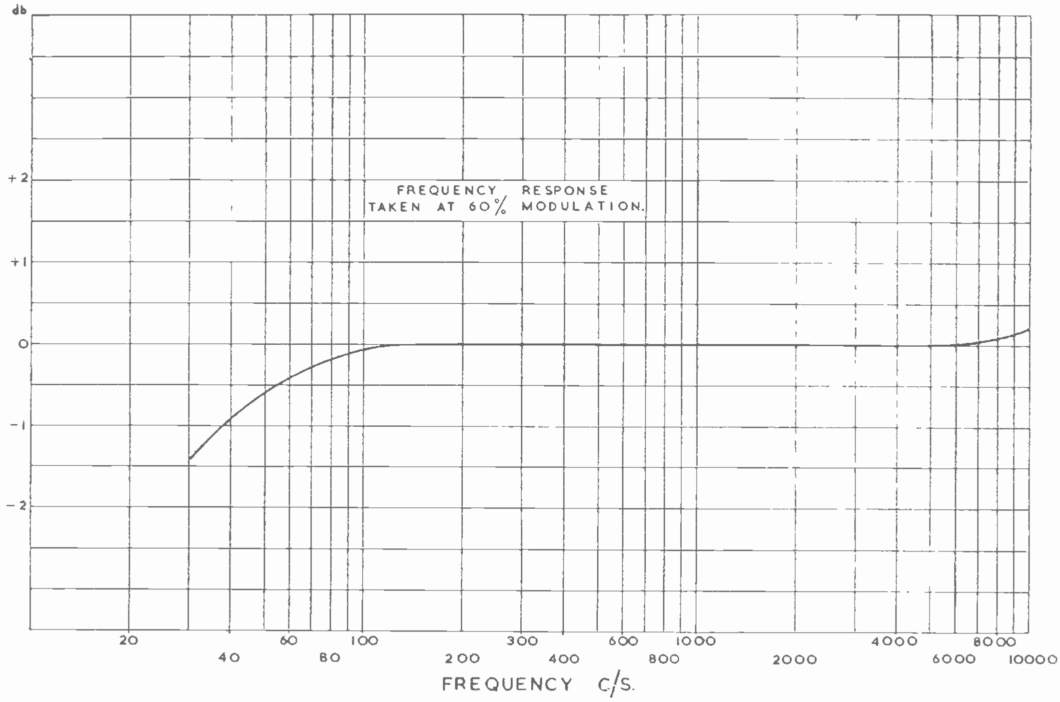
db
+2
+1
0
-1
-2
FREQUENCY RESPONSE
TAKEN AT 60% MODULATION.
20
40
60
80
100
200
300
400
600
800
1000
2000
4000
6000
8000
10000
FREQUENCY C/S.

FIG. 15

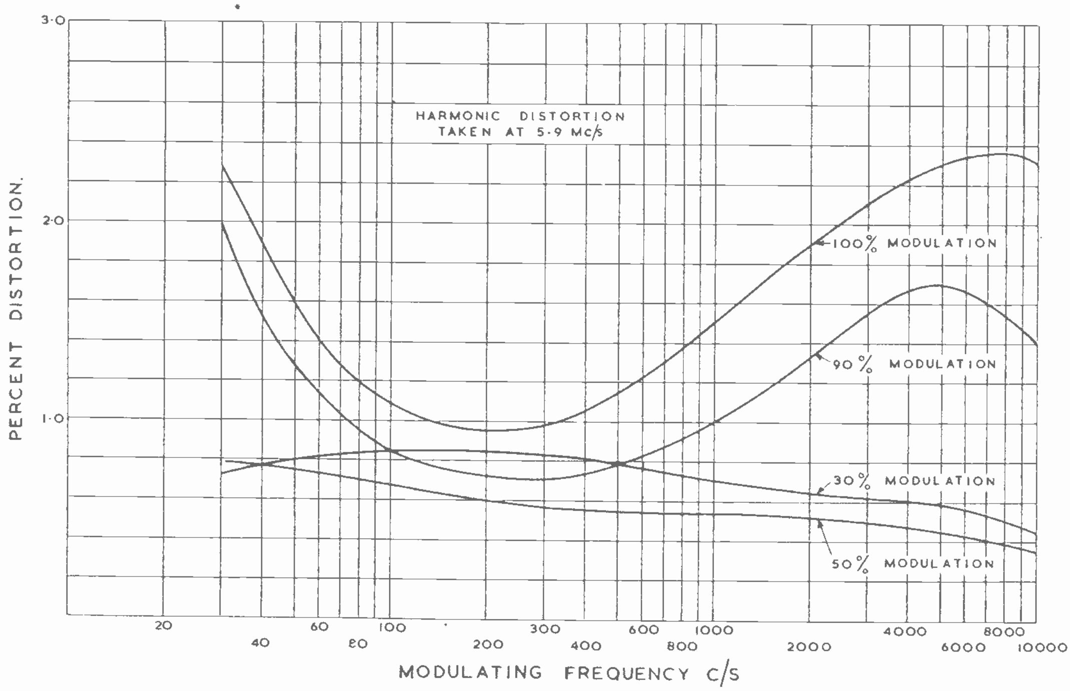
3·0
2·0
1·0
PERCENT DISTORTION.
HARMONIC DISTORTION
TAKEN AT 5·9 Mc/s
100% MODULATION
90% MODULATION
30% MODULATION
50% MODULATION
20
40
60
80
100
200
300
400
600
800
1000
2000
4000
6000
8000
10000
MODULATING FREQUENCY C/S

FIG. 16

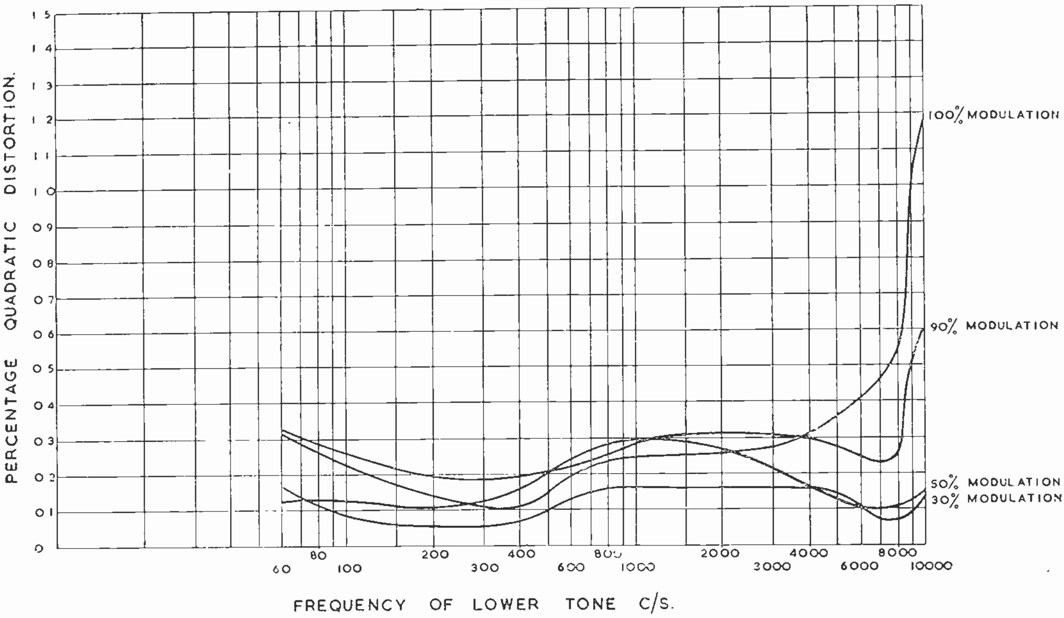

FIG. 17. *Quadratic Intermodulation Distortion*

Difference between tones: 1,000 *c/s*.
For two tones f_1 *and* f_2,

$$\text{Quadratic Distortion} = \frac{\text{R.M.S. value of frequency } (f_1 + f_2)}{\text{sum of R.M.S. values of } f_1 \text{ and } f_2}$$

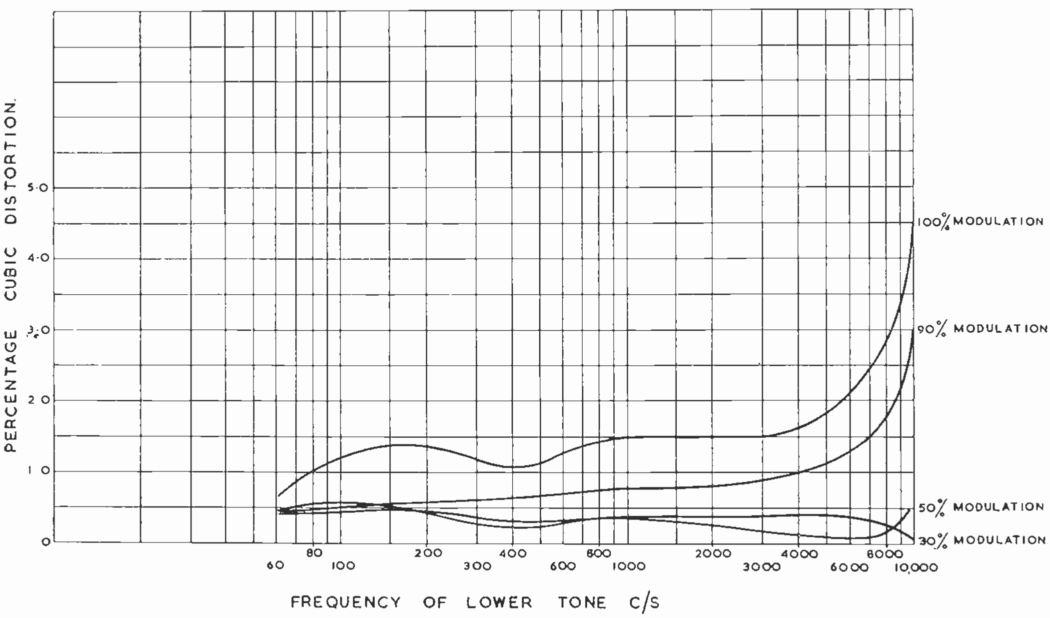

FIG. 18. *Cubic Intermodulation Distortion*

Difference between tones: 1,000 *c/s*.
For two tones f_1 *and* f_2, *Cubic distortion*

$$= \frac{\text{sum of the R.M.S. values of frequencies } (2f_1 - f_2) \text{ and } (2f_2 - f_1)}{\text{sum of the R.M.S. values of } f_1 \text{ and } f_2}$$

All units which are supplied with 230-volt single phase or 440-volt three-phase mains supply are fitted with "Castell" type door locks. When these units are closed and locked, then keys, released from the units, may be inserted and trapped in a suitable switch box. The transmitter isolator switch may then, and only then, be closed. This is the second stage of interlocking and it is now possible to switch on low power supplies, start the main exhaust fan and light the filaments of the high power valves.

Before the outgoing feeder and the high voltage rectifier outputs can be disconnected from earth and dangerous voltages applied to the transmitter, it is necessary to close all doors leading into the transmitter enclosure. This having been done, these doors may be locked from a single handle. The earth connection can now be removed from the outgoing feeder and the high voltage rectifiers. This is the third stage of interlocking and the transmitter may now be operated fully.

Transmitter Performance

Figs. 13 and 14 show typical operating conditions for the transmitter on 5·9 Mc/s and 26·1 Mc/s respectively. These figures show the power input required from the mains supply, various valve feeds and the corresponding radio frequency power supplied to the outgoing feeder. Figs. 15 and 16 show typical frequency response and harmonic distortion characteristics. Intermodulation distortion, as measured by the relative amplitude of a "difference frequency" tone generated when two tones of equal amplitude but different frequencies are used to modulate the transmitter is shown on Figs. 17 and 18.

BOOK REVIEW

It was in the immediate post-war days, when up-to-date books on radio design were almost unknown, that *Radio Designer's Handbook* first appeared in this country as a book of some 350 pages. It became an immediate success and now a fourth edition has been published, again under the editorship of F. Langford-Smith.

The competent indexing of the wide diversity of subjects included under the title of "Radio Design" is a most impressive feature of the book. The thirty-eight chapters are collected together into seven groups in a Chapter Index; each chapter is divided into sections and sub-sections and the headings are collected in a subject index occupying some thirty pages; finally an alphabetical index at the back of the book lists about eight thousand items covered in the text.

In his preface Mr. Langford-Smith states that everything outside the field of radio receivers and audio amplifiers has been excluded to limit the book to a reasonable size; one cannot but agree with the necessity for this restriction, but one wonders whether sixty pages should have been devoted to Valve Testing, and whether the designers for whom the book has been produced will appreciate the inclusion of fifty pages of elementary mathematics. It is clear that an attempt has been made to produce a comprehensive handbook at a reasonable price, and there is no doubt that the editor and his colleagues have succeeded triumphantly. If faults exist they may be found in the rather indistinct typeface employed, and in the size of the volume which tends to become unmanageable. Moreover, one feels that the book is worthy of a better binding.

In a book devoted to a subject of this scope a good bibliography is of the greatest value. Here again nothing but praise can be accorded to the editor; each chapter concludes with a comprehensive list of references compiled from a cross-section of British and American publications.

Finally, a tribute should be paid to the competence with which the editor has handled the work of his collaborators; in the limited period during which this reviewer has "lived with" the book he has been as much impressed by the absence of overlapping from author to author, as by the extremely adequate cross-referencing employed where occasion demands it.

Mr. Langford-Smith's book can be confidently recommended to professional radio engineer and amateur alike as a comprehensive and authoritative addition to the literature.

Radio Designer's Handbook. Edited by F. LANGFORD-SMITH, B.Sc., B.E. Published by Iliffe & Sons, Ltd., at 42s.

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