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# WIRELESS ENGINEER

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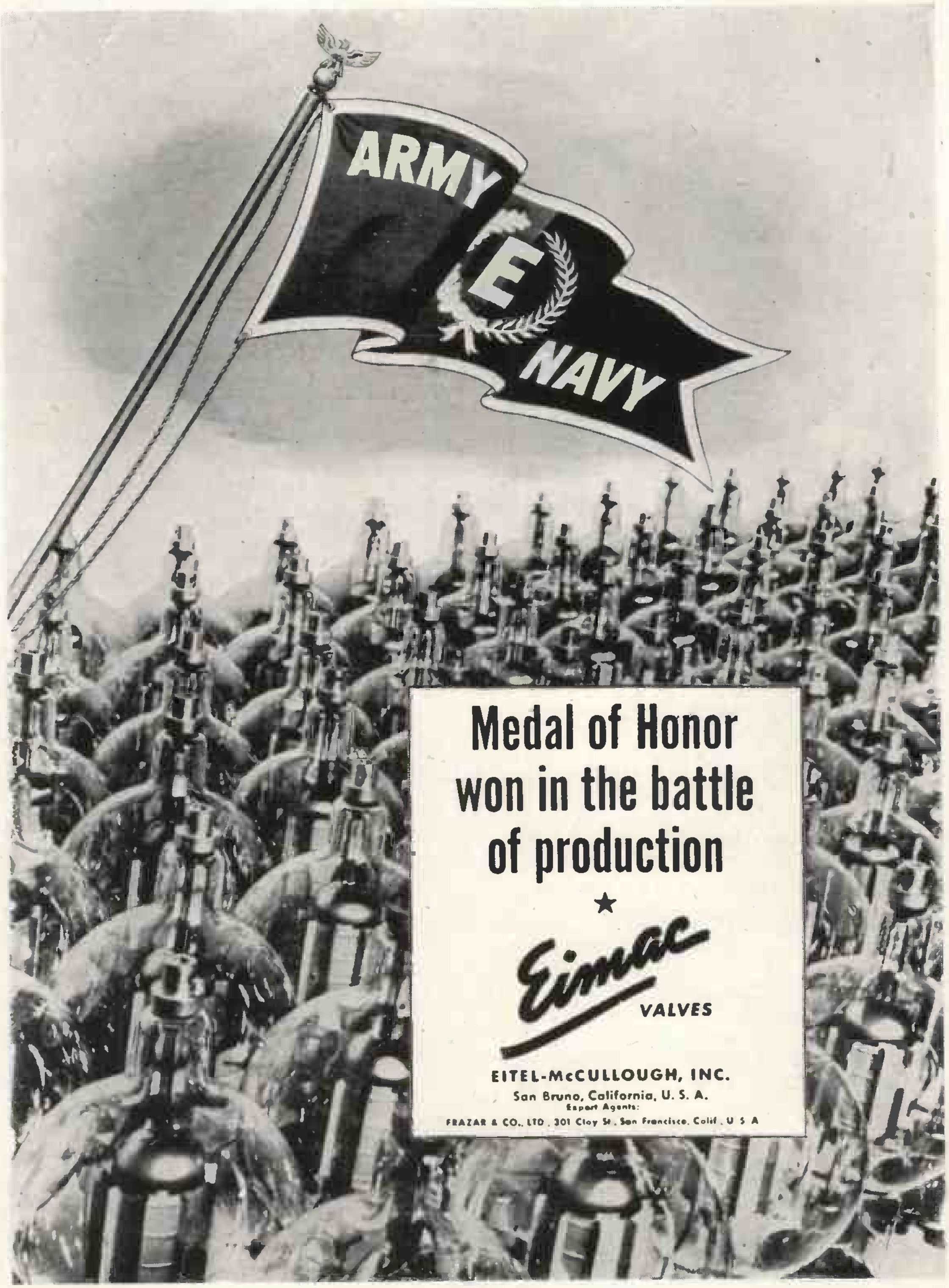
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# WIRELESS ENGINEER

Editor HUGH S. POCOCK, M.I.E.E. Technical Editor Prof. G. W. O. HOWE, D.Sc., M.I.E.E.

VOL. XX FEBRUARY, 1943 No. 233


## Editorial

### Optimum Conditions in Class A Amplifiers

IT is generally known that in a Class A amplifier, that is, an amplifier operating in such a way that the grid voltage never goes beyond the cut-off point, and in which, therefore, a sinusoidal grid-voltage produces an approximately sinusoidal variation of the anode current, the greatest amount of power is supplied to the load—assumed purely resistive—when the resistance $R$ of the load is made equal to the slope resistance $r_a$ of the valve. This assumes that the a.c. grid voltage is fixed, and that the characteristic curves are linear. It is also generally known that if more practical assumptions are made, *viz.*, that the amplitude of the a.c. grid voltage is only limited by the conditions (1) that it should never cause the voltage on the grid to become positive, and (2) that the anode current should never fall below a certain minimum value, maximum power is obtained by making $R$ equal to $2r_a$. The object of the two conditions, is to limit distortion; it can be seen from Fig. 1 that by fixing a minimum anode current, the load-line is limited more or less to the region in which the characteristics are approximately parallel straight lines. It is assumed, moreover, that the quiescent anode voltage $V_{ao}$ is fixed and that the grid-bias and the H.T. supply voltage are adjusted to bring the quiescent point exactly to the middle of the load line $AA'$, for it is only then that the anode current oscillates between the two limiting conditions. In Fig. 1 cotan $\alpha = r_a$ and cotan $\beta = R$; so long as $R$ is fixed at $2r_a$, the load-line must be paralled to $AA'$ and its operative mid-point must lie on the dotted line $pp'$; if $R$ were made equal to $3r_a$ the load line would be parallel to $AA''$ and its operative mid-point would lie on the dotted line $pp''$.

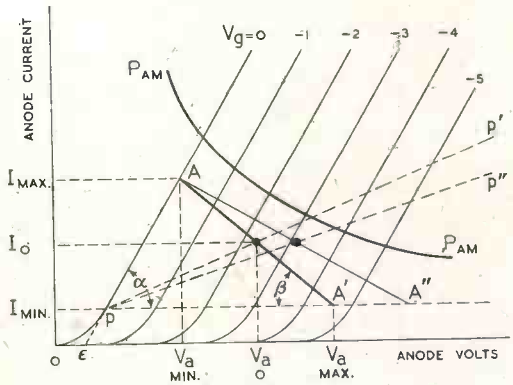


Fig. 1.

In Fig. 2 it is assumed that the quiescent anode voltage $V_{ao}$ is fixed, and the load-lines are drawn for three different values of $R$, the H.T. supply voltage and the grid-bias being suitably adjusted in each case to bring the operative mid-point of the load-line on to the vertical $qr$. The a.c. power supplied to the load is proportional to the product $\hat{v}\hat{\imath}$ *i.e.* to $qs \times qu$ or to $qs \times st$ which is equal to $(pq - ps)st$. Since $ps/st = r_a$, this may be written $\frac{1}{r_a}(pq \times ps - ps^2)$, and to find the condition for this to have its maximum value, we differentiate it with reference to $ps$ and equate to zero. This gives $pq = 2\,ps$ ; hence $ps = sq$ and $R = 2r_a$ as stated above.

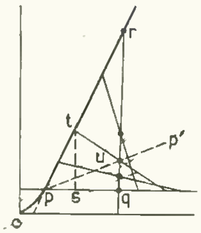


Fig. 2.

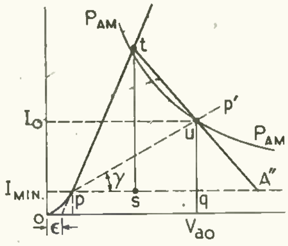


Fig. 3.

The line $pp'$ on which the mid-point of the operative load-line for a given value of $R$ must lie makes an angle $\gamma$ with the horizontal such that

$$\tan\gamma = \frac{qu}{ps + sq} = \frac{st/2}{st\,.\,r_a + st\,.\,R/2} = \frac{1}{2r_a + R}$$

In a recent paper W.B. Nottingham* draws attention to another case in which, instead of specifying the quiescent anode voltage, the quiescent anode dissipation is specified. This must, of course, never exceed a certain value because of the danger of overheating the anode and damaging the valve. The question is, what value $R$ should have in order that the power output may be a maximum without exceeding the specified maximum quiescent anode dissipation $P_{am}$. Since $P_a = V_{ao} I_o$, if the maximum value $P_{am}$ is specified, the maximum value of $I_o$ must be inversely proportional to $V_{ao}$, as shown by the hyperbolic curve $P_{am}$ in Fig. 1. This curve is the upper limit of the mid-point of the operative load-line.

In the case already considered, in which we found the optimum value of $R$ to be $2r_a$, it may be found that this is impracticable, for making $R = 2r_a$ may give an anode dissipation greater than $P_{am}$ if $V_{ao}$ has been chosen at too high a value. If this be the case, $R$ must be increased, as can be seen from Fig. 2, until $I_o$ is reduced to $\dfrac{P_{am}}{V_{ao}}$. The correct value of $R$ can be calculated from the formula

$$R = \frac{V_{ao} - V_{a\,\min}}{I_o - I_{\min}} = \frac{V_{ao} - r_a\,(2I_o - I_{\min}) - \epsilon}{I_o - I_{\min}} \qquad \ldots \quad \ldots \quad (A)$$

The correctness of this can readily be seen from Fig. 1.

To find the largest value that $V_{ao}$ can be given if $R = 2r_a$, without the quiescent dissipation exceeding $P_{am}$, it is only necessary to drop a perpendicular from the point at which $pp'$ intersects the $P_{am}$ hyperbola, since, as we have already seen, if $R = 2r_a$, the quiescent point, *i.e.* the mid-point of any operative load-line parallel to $AA'$ must lie on $pp'$. This is shown in Fig. 3, in which $qu = I_o - I_{\min}$, $st = 2qu = 2(I_o - I_{\min})$ and therefore $ps = 2r_a\,(I_o - I_{\min})$. $V_{ao}$ is made up of the two small portions $\epsilon$ and $r_a I_{\min}$ near the origin, and $pq = 2ps = 4r_a(I_o - I_{\min})$.

Putting $I_o = \dfrac{P_{am}}{V_{ao}}$ we have

$$V_{ao} = \epsilon + r_a I_{\min} + 4r_a\left(\frac{P_{am}}{V_{ao}} - I_{\min}\right)$$

or $\quad V^2_{ao} - V_{ao}\,(\epsilon - 3r_a\,I_{\min}) = 4r_a\,P_{am}$

and therefore

$$V_{ao} = \tfrac{1}{2}\left[(\epsilon - 3r_a\,I_{\min}) + \sqrt{16r_a\,P_{am} + (\epsilon - 3r_a\,I_{\min})^2}\right]$$

If in Fig. 3, as an approximation, $V_{ao}$ is taken as being equal to $pq$ and $I_o$ to $qu$, then

$$st = 2I_o \text{ and } V_{ao} = 2ps = 4r_a\,I_o = 4r_a\frac{P_{am}}{V_{ao}}.$$

Hence $V_{ao} = 2\sqrt{r_a\,P_{am}}$ is approximately the highest permissible quiescent anode voltage if $R = 2r_a$.

If a higher value of $V_{ao}$ is permissible without endangering the valve, then the power output may be still further increased by using a higher value of $R$. This means sliding the mid-point $u$ to the right along the $P_{am}$ curve (see Fig. 3) and thus reducing $I_o$ ; the value of $R$ can then be determined by the formula ($A$) found above. There is, however, a limit to the increase of power even if the valve is not endangered by the increase of $V_{ao}$. The optimum value of $R$ under these conditions can be determined graphically or by trial and error, but as can be seen from Fig. 5 it is of little importance, since the output varies very slightly for large changes in $R$, and the necessary

* *Proc. I.R.E.*, December 1941, p. 620.

value of $V_{ao}$ is probably far above the permissible limit.

In the paper referred to, the calculations are carried out for an actual valve with the following constants :—

$P_{am} = 100$ watts ; $r_a = 1{,}700$ ohms ; $V_{ao}$ should not exceed 1,250 volts ; $\mu = 5.3$ ; $\epsilon = 85$ volts ; $I_{min} = 10\ mA$ (estimated).

The characteristic for $V_g = 0$ and the hyperbola for $P_{am} = 100$ watts are plotted in Fig. 4 ; the horizontal through 10 $mA$ fixes the lower limit of the anode current. The lines $pp_1$, $pp_2$, $pp_3$ and $pp_4$ are drawn at angles $\gamma$ corresponding to four different values of $R$ *viz.* $R = r_a$, $R = 2r_a$, $R = 7.65r_a$ and $R = 24.4r_a$. They cut the hyperbola at the points $u_1$, $u_2$, $u_3$ and $u_4$, which are therefore the quiescent points and the mid-points of the four operative load-lines which are shown in the Figure. The corresponding values of $V_{ao}$ and $I_o$ can be read off the scales and the power outputs calculated by the formula

$$0.5\ (I_o - I_{\min})^2 R.$$

For $R = r_a$, $V_{ao} = 740$ volts, $I_o = 135\ mA$ and the output 13.3 watts. On increasing $R$ to $2r_a$ we have $V_{ao} = 840$ volts,

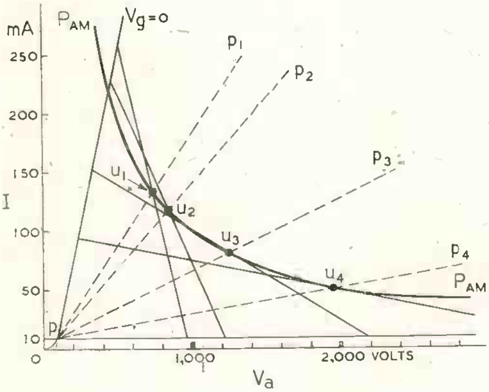


Fig. 4.

$I_o = 119\ mA$, and the output goes up to 20 watts. The approximate formula that we found for the maximum permissible value of $V_{ao}$ with $R = 2r_a$ gives

$$2\sqrt{r_a P_{am}} = 2\sqrt{1{,}700 \times 100} = 825 \text{ volts.}$$

The point $u_3$ corresponds to the limiting value of the quiescent anode voltage recommended by the makers, *viz.*, 1,250 volts ; $I_o$ is therefore 80 $mA$, and $R$ is found by formula $A$ to be $7.65r_a$. The output is thus increased to 31.9 watts. By ignoring the limit set by the makers and increasing $V_{ao}$ to 1,950 volts, we have $u_4$ as the quiescent point ; $I_o = 51\ mA$ and $R$ must be increased to 41,500 ohms, that is, $24.4r_a$. The output under these extreme conditions is 35.3 watts.

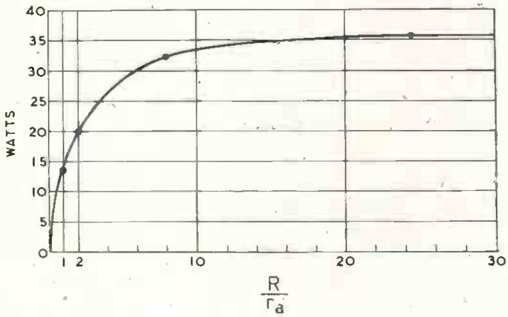


Fig. 5.

In Fig. 5 the output is plotted against the value of $R/r_a$. In all these cases it would be necessary to adjust carefully the grid bias to obtain the value of $I_o$ corresponding to the value of $V_{ao}$.

We thus see that in this actual example it is possible without exceeding the limits set by the makers to increase the output more than 50 per cent. by making $R$ equal to $7.65r_a$ instead of $2r_a$.

G. W. O. H.

## New Year Honours

AMONG the few names of prominent figures in the world of wireless which appeared in the New Year Honours List were Dr. C. Dannatt, head of a section of the Research Department of Metropolitan-Vickers, and C. O. Stanley, managing director of Pye, who were appointed Officers of the Order of the British Empire.

Among those appointed Members of the Order of the British Empire were E. C. S. Megaw, scientific officer in the G.E.C. Research Laboratories, who is a frequent contributor to this journal ; D. C. Birkinshaw, engineer-in-charge of the B.B.C. station at Daventry ; A. L. Chilcot, chief engineer of the Electronics Department of Ferranti, and T. W. Morgan, works manager of Marconi's.

# High-Frequency Resistance of Plated Conductors*

*By R. Faraday Proctor, M.Sc., A.M.I.E.E.*

*(Communication from the Staff of the Research Laboratories of The General Electric Company, Limited, England)*

**SUMMARY.**—Schelkunoff[1] has derived an expression for the H.F. impedance of a laminated, or plated conductor, in terms of the surface and transfer impedances of its constituent layers. (For the definition of these see sections 2 and 3). His expression lends itself to numerical calculations as the various impedances can be readily evaluated with the aid of tables. It also has the advantage of being easily applied to conductors having any number of layers of plating.

In section 2 it is shown how Schelkunoff's expression may be derived, in the case of plane conductors, from the normal uniform line equations, by a simple extension of the method proposed by Howe[2] for calculating the skin resistance of plain conductors.

In section 3 the results are extended to cover conductors of any arbitrary cross-section.

In section 4 expressions are derived for the surface and transfer impedances of hollow cylinders and the surface impedance of solid cylinders or rods. Tables (I and II) have been prepared enabling the surface impedance of a cylindrical, or plane, plated conductor to be readily calculated from equation 15 and the appropriate equations of sections 4(a) and 4(b).

In section 5(a) curves are given of the surface impedance of plane plated conductors composed of tinned copper, copper-plated iron, and nickel-plated copper. Section 5(b) indicates the results which may be expected from the use of copper-plated iron conductors as a means of equalising the attenuation of a co-axial cable over a wide band of frequencies.

## 1. Introduction

A KNOWLEDGE of the H.F. resistance of plated or laminated conductors is of considerable interest to the communication engineer. At very high frequencies, where the penetration depth of the current is small compared with the thickness of the plating, the problem reduces to that of a plain solid conductor composed of the same metal as the plating. At lower frequencies where the current penetration is such that it flows partly in the plating and partly in the underlying metal, the problem becomes considerably more complicated. The solution of this case is given in a paper by Schelkunoff[1] in which he derives an expression for the H.F. impedance of a laminated conductor consisting of $n$ layers. Unfortunately this paper has not been given the attention it deserves by engineers in general owing to its somewhat advanced mathematical nature. The problem of plane plated conductors may, however, be readily undertaken by means of the normal line equations with which telephone and radio engineers are familiar, by an extension of the method described by Howe[2] for the derivation of the H.F. resistance of plain conductors. This approach to the problem has the advantage of bringing out clearly the physical significance of the equations obtained in the more involved case of cylindrical conductors.

## 2. Plane Conductors

Following Howe, consider the case in which the outward and return leads of a transmission line consist of two very wide conducting strips as shown in Fig. 1. In the case of perfect conductors no potential

* MS accepted by the Editor, August, 1942.

differences can exist in the conductors and the electric field intensity in the dielectric is solely in the direction $AB$. The magnetic field is in the direction perpendicular to the plane of the paper, and the energy is propagated through the dielectric in the plane of the paper, in the direction $CD$ as indicated by Poynting's Theorem, the two conductors merely acting as guides. In the case of imperfect conductors, however, potential differences are produced in the metallic conductors in the direction $CD$ due to the currents flowing in them. These cause a flow of electro-magnetic energy into the conductor, in the direction $AB$, with a corresponding reduction of the energy propagated in the direction $CD$.

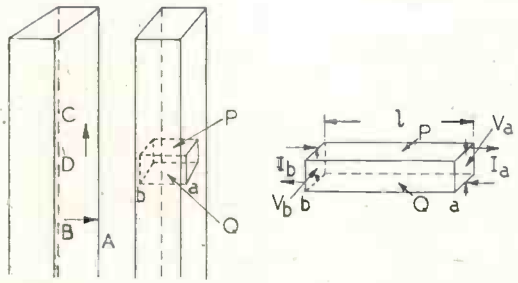


Fig. 1.

We are not now concerned with the transmission of power along the conductors but only with the transmission of power into the conductors, from the outer surface, required to supply the losses in the conductors. It is only necessary to consider the transmission of energy through a square of one cm. side in the surface of the conductor and along the square column of material of which this forms the base. Let this represent a transmission line in which the two opposite faces $P$, $Q$, are assumed to be two resistanceless conductors and the material of the plate is the dielectric.

If the intrinsic permeability $\mu_0$ of free space is taken as $4\pi \times 10^{-9}$ the inductance per cm. length of this fictitious transmission line will be $\mu$ henrys per cm. where $\mu$ is the intrinsic permeability of the conductor. The magnetic permeability of the material in its usually accepted sense will then be $\mu/\mu_0$.

The conductance per unit length, $G$, will be $1/\rho$ where $\rho$ is the specific resistance of the material of the conductor. Since the conductance per unit length is enormous in comparison with the susceptance, the capacitance per unit length may be safely neglected for simplicity.

Let $V_b$, $I_b$ be the voltage and current at the sending end of the transmission line, and let $V_a$, $I_a$ be the voltage and current at the receiving end. Also let $Z_0$ be the characteristic impedance and let $\gamma = \alpha + j\beta$ be the propagation constant. Then

$$V_b = V_a \cosh \gamma l + Z_0 I_a \sinh \gamma l \quad \ldots \quad (1)$$

$$I_b = V_a/Z_0 \sinh \gamma l + I_a \cosh \gamma l \quad \ldots \quad (2)$$

where $l$ is the length of the line

$$Z_0 = \sqrt{\frac{R + j\omega L}{G + j\omega C}} = \sqrt{\frac{j\omega\mu}{G}} = (1+j) \cdot \sqrt{\frac{\omega\mu}{2G}} \quad \ldots \quad (3)$$

$$\gamma = \alpha + j\beta = \sqrt{(R + j\omega L)(G + j\omega C)}$$
$$= \sqrt{j\omega\mu G} = (1 + j)\sqrt{\frac{\omega\mu G}{2}} \quad \ldots \quad (4)$$

If the adjoining column, say the next column above is considered, an exactly similar set of equations is obtained and the fictitious line currents flowing in the adjacent surfaces would be equal and opposite and would hence cancel out. The assumed resistanceless surfaces of the fictitious transmission line may thus be removed without affecting the results. This assumes that there is no phase change in the direction of the main current and power transmission, which while not strictly true is sufficiently accurate for the case in hand. The leakage currents of the fictitious transmission line are the actual currents in the main transmission line.

From the above it will be readily seen that the H.F. impedance, or surface impedance as Schelkunoff calls it, of the conductors of the main transmission line per sq. cm. is given by the input impedance of the fictitious transmission line.

Assume that the fictitious transmission line is terminated by an impedance $Z_r$, such as would be the case where one material is plated with another. The sending end impedance $z_{bb}$ would then be given by

$$z_{bb} = \frac{V_b}{I_b} = Z_0 \cdot \frac{Z_r \cosh \gamma l + Z_0 \sinh \gamma l}{Z_r \sinh \gamma l + Z_0 \cosh \gamma l} \quad \ldots \quad \ldots \quad (5)$$

$$= Z_0 \coth \gamma l \left[ \frac{Z_r \sinh \gamma l + \dfrac{Z_0 \sinh^2 \gamma l}{\cosh \gamma l}}{Z_r \sinh \gamma l + Z_0 \cosh \gamma l} \right]$$

$$= Z_0 \coth \gamma l \left[ 1 - \frac{Z_r \sinh \gamma l + Z_0 \cosh \gamma l - Z_r \sinh \gamma l - \dfrac{Z_0 \sinh^2 \gamma l}{\cosh \gamma l}}{Z_r \sinh \gamma l + Z_0 \cosh \gamma l} \right]$$

$$= Z_0 \coth \gamma l - Z_0 \coth \gamma l \left[ \frac{Z_0 \cosh^2 \gamma l - Z_0 \sinh^2 \gamma l}{\cosh \gamma l (Z_r \sinh \gamma l + Z_0 \cosh \gamma l)} \right]$$

$$= Z_0 \coth \gamma l - Z_0^2 \frac{\cosh \gamma l}{\sinh \gamma l} \left[ \frac{1}{\cosh \gamma l (Z_r \sinh \gamma l + Z_0 \cosh \gamma l)} \right]$$

$$z_{bb} = Z_0 \coth \gamma l - \left[ \frac{Z_0}{\sinh \gamma l} \right]^2 \frac{1}{Z_r + Z_0 \coth \gamma l} \quad \ldots \quad \ldots \quad \ldots \quad (6)$$

$Z_0 \coth \gamma l$ is the open circuit impedance of the fictitious transmission line, that is, the surface impedance $Z_{bb}$ of the outer layer of the composite conductor in the case of energy flowing into the layer in the direction $B$ to $A$, $Z_r$ is the surface impedance of the inner layer.

Putting $I_a = 0$ in equation 2 gives

$$V_a = \frac{Z_0}{\sinh \gamma l} I_b$$

thus $\dfrac{Z_0}{\sinh \gamma l}$ is the mutual, or transfer impedance $Z_{ba}$ in the direction $B$ to $A$.

Equation 6 thus becomes

$$z_{bb} = Z_{bb} - \frac{[Z_{ba}]^2}{Z_{aa} + Z_r} \quad \ldots \quad \ldots \quad (7)$$

$Z_{aa}$ is the open circuit impedance of the fictitious transmission line looking in at the receiving end, that is the surface impedance of the outer layer with reference to energy flowing into the layer in the direction $A$ to $B$. The reason for writing $Z_{aa}$ in the denominator of equation 7 instead of $Z_{bb}$ will be readily seen when the general case is considered. In the case of plane conductors $Z_{bb} = Z_{aa}$.

The surface and transfer impedances may be expressed in terms of the electrical constants of the materials as follows.

$$Z_{bb} = R_{bb} + j\omega L_{bb} = Z_0 \coth (\alpha + j\beta) l$$

Substituting for $Z_0$ from 3 gives

$$R_{bb} + j\omega L_{bb} = (1 + j)\sqrt{\frac{\omega\mu}{2G}} \coth(\alpha + j\beta) l$$

$$= (1 + j)\sqrt{\frac{\omega\mu}{2G}} \frac{\sinh 2\alpha l - j \sin 2\beta l}{\cosh 2\alpha l - \cos 2\beta l}$$

whence

$$R_{bb} = \sqrt{\frac{\omega\mu}{2G}} \frac{\sinh u + \sin u}{\cosh u - \cos u} \quad \ldots \quad (8)$$

$$\omega L_{bb} = \sqrt{\frac{\omega\mu}{2G}} \frac{\sinh u - \sin u}{\cosh u - \cos u} \quad \ldots \quad (9)$$

where $u = 2\alpha l = 2\beta l = l\sqrt{2\omega\mu G}$ (see equation 4).

$$Z_{ba} = \frac{Z_0}{\sinh (\alpha + j\beta) l}$$

$$= \frac{Z_0}{\sinh \alpha l \cos \beta l + j \cosh \alpha l \sin \beta l}$$

$$= (1 + j)\sqrt{\frac{\omega\mu}{2G}} \; \frac{\sinh \alpha l \cos \beta l - j \cosh \alpha l \sin \beta l}{\sinh^2 \alpha l \cos^2 \beta l + \cosh^2 \alpha l \sin^2 \beta l}$$

$$R_{ba} = 2\sqrt{\frac{\omega\mu}{2G}} \frac{\sinh \frac{u}{2} \cos \frac{u}{2} + \cosh \frac{u}{2} \sin \frac{u}{2}}{\cosh u - \cos u} \quad \ldots \quad \ldots \quad (10)$$

$$\omega L_{ba} = 2\sqrt{\frac{\omega\mu}{2G}} \frac{\sinh \frac{u}{2} \cos \frac{u}{2} - \cosh \frac{u}{2} \sin \frac{u}{2}}{\cosh u - \cos u} \quad \ldots \quad \ldots \quad (11)$$

## 3. General Case, Conductors of any Shape

Equation 7 may be extended to cover the case of conductors of any arbitrary cross section as follows :—Consider the four terminal network of Fig. 2 and let

$$V_b = AV_a + BI_a \quad \ldots \quad \ldots \quad (12)$$

$$I_b = CV_a + DI_a \quad \ldots \quad \ldots \quad (13)$$

Further, let the impedance connected across the receiving end be $Z_r$, then

$$V_a = Z_r I_a$$

Proceeding as before, the sending end impedance is given by

$$z_{bb} = \frac{V_b}{I_b} = \frac{AZ_r + B}{CZ_r + D}$$

$$= \frac{A}{C}\left[\frac{Z_r + B/A}{Z_r + D/C}\right]$$

$$= \frac{A}{C}\left[1 - \frac{Z_r + D/C - Z_r - B/A}{Z_r + D/C}\right]$$

$$z_{bb} = \frac{A}{C} - \frac{\left(\frac{AD}{C} - B\right)1/C}{Z_r + D/C} \quad \ldots \quad (14)$$

$A/C$ is the open circuit impedance $Z_{bb}$ of the network looking in from the left-hand end and $D/C$ is the open circuit impedance $Z_{aa}$ of the network looking in from the right-hand end.

The transfer impedance from left to right is $Z_{ba} = 1/C$ and that from right to left $Z_{ab} = 1/C(AD - BC)$

But by the reciprocity theorem $Z_{ab} = Z_{ba}$.

Therefore $AD - BC = 1$.

Hence equation 14 becomes

$$z_{bb} = Z_{bb} - \frac{[Z_{ba}]^2}{Z_{aa} + Z_r} \quad \ldots \quad \ldots \quad (15)$$

As a simple illustration of equation 15, consider a transformer feeding a load $Z_r$.

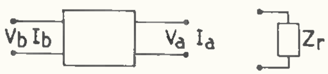


Fig. 2.

Let $R_1$ and $L_1$ be the resistance and inductance of the primary winding and let $R_2$ and $L_2$ be the corresponding values for the secondary. Also let $M$ be the mutual coupling. Then $Z_{bb} = R_1 + j\omega L_1$

$$Z_{aa} = R_2 + j\omega L_2 \,;\; Z_{ba} = j\omega M.$$

The input impedance of the transformer under load conditions is, from 15,

$$z_{bb} = R_1 + j\omega L_1 - \frac{[j\omega M]^2}{R_2 + j\omega L_2 + Z_r}$$

Equations 7 and 15 may be readily extended to the case of a conductor consisting of many layers, since the equation still holds if $Z_r$ is itself the surface impedance of a number of layers.

### 4. Cylindrical Conductors (Tubes and Rods)

A case of particular interest to the communication engineer is that of cylindrical conductors. In this case the normal equations for a uniform line do not apply since the constants of the fictitious transmission line vary with the radius. The appropriate line equations, required for the evaluation of the surface and transfer impedances, may however be readily derived as follows.

Consider the section of an infinite hollow metallic cylinder contained between two planes one cm. apart, see Fig. 3. Let this represent a transmission line in which current flows inward towards the axis over the surface $AB$ and returns along the surface $CD$. At radius $r$ the intrinsic inductance per cm. length of this transmission line will be $\mu/2\pi r$ henrys per cm. The leakage, or conductance, per unit length will be $2\pi rg$ where $1/g$ is the specific resistance of the conductor in ohms per cm. cube. Then neglecting, as before, the capacitance per unit length compared with the leakance

$$\frac{dV}{dr} = \frac{j\omega\mu}{2\pi r} I \quad \ldots \quad \ldots \quad \ldots \quad (16)$$

$$\frac{dI}{dr} = 2\pi rgV \quad \ldots \quad \ldots \quad \ldots \quad (17)$$

Differentiating (16) gives

$$\frac{d^2V}{dr^2} = j\frac{\omega\mu}{2\pi r}\frac{dI}{dr} - j\frac{\omega\mu}{2\pi r^2} I$$

which on substituting for $I$ and $\frac{dI}{dr}$ gives

$$\frac{d^2V}{dr^2} + \frac{1}{r}\frac{dV}{dr} - j\omega\mu gV = 0 \quad \ldots \quad (18)$$

The solution of (18) is

$$V = AI_0(\sigma r) + BK_0(\sigma r) \quad \ldots \quad \ldots \quad (19)$$

where $I_0(\sigma r)$ and $K_0(\sigma r)$ are modified Bessel Functions of zero order and

$$\sigma^2 = j\omega\mu g \qquad \sigma = (1 + j)\sqrt{\frac{\omega\mu g}{2}} \qquad (20)$$

Differentiating (19) gives

$$\frac{dV}{dr} = \sigma[AI_1(\sigma r) - BK_1(\sigma r)] \quad \ldots \quad (21)$$

where $I_1(\sigma r)$ and $K_1(\sigma r)$ are modified Bessel Functions of the first order.

Substituting (21) in (16) gives

$$I = -j\frac{2\pi r\sigma}{\omega\mu}[AI_1(\sigma r) - BK_1(\sigma r)] \quad (22)$$

(a) *Hollow Cylinders*

Let $I_a$ be the line current at the inner radius $a$ and let $I_b$ be its value at the outer radius $b$, then

$$I_a = -j\frac{2\pi a\sigma}{\omega\mu}[AI_1(\sigma a) - BK_1(\sigma a)] \quad (23)$$

$$I_b = -j\frac{2\pi b\sigma}{\omega\mu}[AI_1(\sigma b) - BK_1(\sigma b)] \quad (24)$$

Solving (23) and (24) for $A$ and $B$ gives

$$A = \frac{j\frac{\omega\mu}{2\pi b\sigma}K_1(\sigma a)I_b - j\frac{\omega\mu}{2\pi a\sigma}K_1(\sigma b)I_a}{\Delta} \quad \ldots \quad \ldots \quad (25)$$

$$B = \frac{j\frac{\omega\mu}{2\pi b\sigma}I_1(\sigma a)I_b - j\frac{\omega\mu}{2\pi a\sigma}I_1(\sigma b)I_a}{\Delta} \quad \ldots \quad \ldots \quad (26)$$

where

$$\Delta = I_1(\sigma b)K_1(\sigma a) - I_1(\sigma a)K_1(\sigma b) \quad (27)$$

Substituting for $A$ and $B$ in (19) gives

$$V_a = \frac{-j\frac{\omega\mu}{2\pi a\sigma}K_1(\sigma b)I_0(\sigma a)I_a + j\frac{\omega\mu}{2\pi b\sigma}K_1(\sigma a)I_0(\sigma a)I_b}{\Delta} + \frac{-j\frac{\omega\mu}{2\pi a\sigma}I_1(\sigma b)K_0(\sigma a)I_a + j\frac{\omega\mu}{2\pi b\sigma}I_1(\sigma a)K_0(\sigma a)I_b}{\Delta} \quad \ldots \quad \ldots \quad \ldots \quad (28)$$

$$V_b = \frac{-j\frac{\omega\mu}{2\pi a\sigma}K_1(\sigma b)I_0(\sigma b)I_a + j\frac{\omega\mu}{2\pi b\sigma}K_1(\sigma a)I_0(\sigma b)I_b}{\Delta} + \frac{-j\frac{\omega\mu}{2\pi a\sigma}I_1(\sigma b)K_0(\sigma b)I_a + j\frac{\omega\mu}{2\pi b\sigma}I_1(\sigma a)K_0(\sigma b)I_b}{\Delta} \quad \ldots \quad \ldots \quad \ldots \quad (29)$$

Expressing $V_b$ in terms of $I_a$ and $I_b$ by means of the general network equations (12) and (13) gives

$$V_b = \frac{A}{C}I_b - \frac{AD - BC}{C}I_a$$

or

$$V_b = Z_{bb}I_b - Z_{ab}I_a \quad \ldots \quad \ldots \quad (30)$$

Similarly

$$V_a = Z_{ba}I_b - Z_{aa}I_a \quad \ldots \quad \ldots \quad (31)$$

Comparing (29) and (30) gives

$$Z_{bb} = \frac{j\omega\mu}{2\pi b\sigma\Delta}[I_0(\sigma b)K_1(\sigma a) + I_1(\sigma a)K_0(\sigma b)] \quad \ldots \quad \ldots \quad (32)$$

$$Z_{ba} = \frac{j\omega\mu}{2\pi a\sigma\Delta}[K_1(\sigma b)I_0(\sigma b) + I_1(\sigma b)K_0(\sigma b)] \quad \ldots \quad \ldots \quad (33)$$

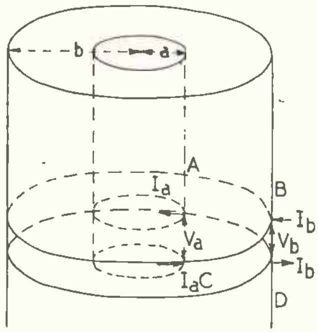


Fig. 3.

And from (28) and (31)

$$Z_{aa} = \frac{j\omega\mu}{2\pi a\sigma\Delta}[K_1(\sigma b)I_0(\sigma a) + I_1(\sigma b)K_0(\sigma a)] \quad \ldots \quad \ldots \quad (34)$$

$$Z_{ab} = \frac{j\omega\mu}{2\pi b\sigma\Delta}[I_1(\sigma a)K_0(\sigma a) + K_1(\sigma a)I_0(\sigma a)] \quad \ldots \quad \ldots \quad (35)$$

Making use of the identity

$$K_1(\sigma r)I_0(\sigma r) + I_1(\sigma r)K_0(\sigma r) = \frac{1}{\sigma r}$$

gives $\quad Z_{ab} = Z_{ba} = j\dfrac{\omega\mu}{2\pi ab\sigma^2\Delta} \quad \ldots \quad (36)$

or substituting for $\sigma^2$ from (20)

$$Z_{ab} = \frac{1}{2\pi abg\Delta} \quad \ldots \quad (37)$$

The exact formulae (32), (34) and (37) do not lend themselves readily to numerical calculations but, as shown by Schelkunoff, simple approximations can be obtained if the modified Bessel Functions $I_0(x)$, $I_1(x)$, $K_0(x)$, $K_1(x)$ are replaced by their asymtotic expansions retaining only the first few terms.

$$I_0(x) = \frac{e^x}{\sqrt{2\pi x}}\left(1 + \frac{1}{8x}\right) \quad \ldots \quad (38)$$

$$K_0(x) = \sqrt{\frac{\pi}{2x}}\, e^{-x}\left(1 - \frac{1}{8x}\right) \quad \ldots \quad (39)$$

$$I_1(x) = \frac{e^x}{\sqrt{2\pi x}}\left(1 - \frac{3}{8x}\right) \quad \ldots \quad (40)$$

$$K_1(x) = \sqrt{\frac{\pi}{2x}}\, e^{-x}\left(1 + \frac{3}{8x}\right) \quad \ldots \quad (41)$$

Substituting the above in (32), (34) and (36) gives after slight approximation

$$Z_{bb} = \frac{\eta}{2\pi b}\left[\coth \sigma(b-a) + \frac{1}{8\sigma}\left(\frac{3}{a} + \frac{1}{b}\right)\right] \quad \ldots \quad (42)$$

$$Z_{aa} = \frac{\eta}{2\pi a}\left[\coth \sigma(b-a) - \frac{1}{8\sigma}\left(\frac{3}{b} + \frac{1}{a}\right)\right] \quad \ldots \quad (43)$$

$$Z_{ab} = \frac{\eta}{2\pi\sqrt{ab}}\, \text{cosech}\, \sigma(b-a) \quad \ldots \quad (44)$$

where

$$\eta = j\frac{\omega\mu}{\sigma} \quad \ldots \quad (45)$$

Substituting (20) and (45) into (42), (43) and (44) and separating the real and imaginary parts gives

$$R_{bb} = \frac{1}{2b}\sqrt{\frac{\mu f}{\pi g}}\left(\frac{\sinh u + \sin u}{\cosh u - \cos u}\right) + \frac{3b+a}{16\pi gab^2} \quad \ldots \quad (46)$$

$$\omega L_{bb} = \frac{1}{2b}\sqrt{\frac{\mu f}{\pi g}}\left(\frac{\sinh u - \sin u}{\cosh u - \cos u}\right) \quad \ldots \quad (47)$$

$$R_{aa} = \frac{1}{2a}\sqrt{\frac{\mu f}{\pi g}}\left(\frac{\sinh u + \sin u}{\cosh u - \cos u}\right) - \frac{3a+b}{16\pi ga^2b} \quad \ldots \quad (48)$$

$$\omega L_{aa} = \frac{1}{2a}\sqrt{\frac{\mu f}{\pi g}}\left(\frac{\sinh u - \sin u}{\cosh u - \cos u}\right) \quad \ldots \quad (49)$$

$$R_{ab} = \frac{1}{\sqrt{ab}}\sqrt{\frac{\mu f}{\pi g}}\, \frac{\sinh\frac{u}{2}\cos\frac{u}{2} + \cosh\frac{u}{2}\sin\frac{u}{2}}{\cosh u - \cos u} \quad \ldots \quad (50)$$

$$\omega L_{ab} = \frac{1}{\sqrt{ab}}\sqrt{\frac{\mu f}{\pi g}}\, \frac{\sinh\frac{u}{2}\cos\frac{u}{2} - \cosh\frac{u}{2}\sin\frac{u}{2}}{\cosh u - \cos u} \quad \ldots \quad (51)$$

where

$$u = t\sqrt{2\omega\mu g} \qquad t = b - a \quad \ldots \quad (52)$$

The second term of equations (46) and (48) represent the first correction for curvature and vanish if the conductors are plane. In this case the equations (46) to (51) become identical with (8) to (11) when expressed per unit length of circumference. Although equations (46) to (51) were derived using asymtotic expansions which were only valid when the argument is large, i.e. at high frequencies, the results are good even at low frequencies provided the tubular conductor is not too thick.

If the frequency is very high equations (46) to (49) become

$$R_{bb} = \frac{1}{2b}\sqrt{\frac{\mu f}{\pi g}} + \frac{3b+a}{16\pi gab^2} \quad \ldots \quad (53)$$

$$\omega L_{bb} = \frac{1}{2b}\sqrt{\frac{\mu f}{\pi g}} \quad \ldots \quad (54)$$

$$R_{aa} = \frac{1}{2a}\sqrt{\frac{\mu f}{\pi g}} - \frac{3a+b}{16\pi ga^2b} \quad \ldots \quad (55)$$

$$\omega L_{aa} = \frac{1}{2a}\sqrt{\frac{\mu f}{\pi g}} \quad \ldots \quad (56)$$

Substituting

$$R_{ob} = \frac{1}{2\pi gbt}; \quad R_{oa} = \frac{1}{2\pi gat}; \quad R_{oab} = \frac{1}{2\pi g\sqrt{ab}t}$$

equations (46) to (51) respectively give

$$R_{bb} = R_{ob}\left[\frac{u}{2}\cdot\frac{\sinh u + \sin u}{\cosh u - \cos u}\right] + \frac{3b+a}{16\pi gab^2} = R_{ob}\,.\,P_1 + \frac{3b+a}{16\pi gab^2} \quad \ldots \quad (57)$$

$$\omega L_{bb} = R_{ob}\left[\frac{u}{2}\cdot\frac{\sinh u - \sin u}{\cosh u - \cos u}\right] = R_{ob}\,.\,P_2 \quad \ldots \quad (58)$$

$$R_{aa} = R_{oa}\left[\frac{u}{2}\cdot\frac{\sinh u + \sin u}{\cosh u - \cos u}\right] - \frac{3a+b}{16\pi g a^2 b} = R_{oa}\cdot P_1 - \frac{3a+b}{16\pi g a^2 b} \quad \ldots \quad \ldots \quad (59)$$

$$\omega L_{aa} = R_{oa}\left[\frac{u}{2}\cdot\frac{\sinh u - \sin u}{\cosh u - \cos u}\right] = R_{oa}\cdot P_2 \quad \ldots \quad \ldots \quad (60)$$

$$R_{ab} = R_{oab}\left[u\cdot\frac{\sinh\frac{u}{2}\cos\frac{u}{2} + \cosh\frac{u}{2}\sin\frac{u}{2}}{\cosh u - \cos u}\right] = R_{oab}\cdot P_3 \quad \ldots \quad \ldots \quad (61)$$

$$\omega L_{ab} = R_{oab}\left[u\cdot\frac{\sinh\frac{u}{2}\cos\frac{u}{2} - \cosh\frac{u}{2}\sin\frac{u}{2}}{\cosh u - \cos u}\right] = R_{oab}\cdot P_4 \quad \ldots \quad \ldots \quad (62)$$

TABLE I

| $u = t\sqrt{2\omega\mu g}$ | $P_1$ | $P_2$ | $P_3$ | $P_4$ |
|---|---|---|---|---|
| 0 | 1.000 | 0.0000 | 1.000 | 0.0000 |
| 0.3 | — | — | — | −0.007550 |
| 1.0 | 1.006 | 0.1664 | 0.9951 | −0.08311 |
| 2.0 | 1.086 | 0.6504 | 0.9254 | −0.3176 |
| 3.0 | 1.378 | 1.340 | 0.6573 | −0.6160 |
| 4.0 | 1.898 | 2.006 | 0.2621 | −0.6941 |
| 5.0 | 2.523 | 2.566 | −0.08046 | −0.5814 |
| 6.0 | 3.003 | 3.021 | −0.2538 | −0.3415 |
| 7.0 | 3.510 | 3.501 | −0.2724 | −0.1237 |
| 8.0 | 4.002 | 3.997 | −0.2066 | +0.01524 |
| 9.0 | 4.500 | 4.499 | −0.1188 | +0.07669 |
| 10.0 | 4.999 | 5.000 | −0.04547 | +0.08374 |

$t$ = thickness of tube in cm.
$\mu$ = intrinsic permeability of the material; that of empty space being taken as $0.01257 \times 10^{-6}$ $(4\pi \times 10^{-9})$.
$g$ = specific conductance of the material in mhos per cm. cube.

To facilitate calculation the terms involving $u$, viz. $P_1$, $P_2$, $P_3$, $P_4$, are tabulated in Table I.

(b) *Solid Cylinders*

For a solid rod the "internal radius" is zero and equation (19) reduces

$$V = AI_0(\sigma r) \quad \ldots \quad \ldots \quad \ldots \quad (63)$$

since in this case $B = 0$ as the E.M.F. along the centre of the rod must be finite whereas $K_0(\sigma r)$ is infinite when $r = 0$.

Similarly

$$I = -j\frac{2\pi r\sigma}{\omega\mu}AI_1(\sigma r) \quad \ldots \quad \ldots \quad (64)$$

Hence

$$Z_{bb} = \frac{V}{I} = j\frac{\omega\mu}{2\pi b\sigma}\frac{I_0(\sigma b)}{I_1(\sigma b)} \quad \ldots \quad \ldots \quad (65)$$

where $b$ is the radius of the rod.

Substituting for $\sigma$ from (20) gives

$$Z_{bb} = \frac{1+j}{2b}\sqrt{\frac{\mu f}{\pi g}}\frac{I_0(\sigma b)}{I_1(\sigma b)} \quad \ldots \quad (66)$$

For large values of $\sigma b$ the modified Bessel Functions may be replaced by their asymtotic expansions, see equations (38) and (40). Substituting these in (66) gives

$$R_{bb} = \frac{1}{2b}\sqrt{\frac{\mu f}{\pi g}} + \frac{1}{4\pi g b^2} \quad \ldots \quad \ldots \quad (67)$$

$$\omega L_{bb} = \frac{1}{2b}\sqrt{\frac{\mu f}{\pi g}} \quad \ldots \quad \ldots \quad \ldots \quad (68)$$

If the rod is so thin or the frequency so low, $|\sigma b| < 6$, that the asymtotic expansions given above are no longer permissible, equation (66) has to be used. Equation (66) can be readily split into its real and imaginary parts by means of the following expressions

$$I_0(\sqrt{j}\,x) = \text{ber}\,x + j\,\text{bei}\,x \quad \ldots \quad (69)$$

$$I_1(\sqrt{j}\,x) = \frac{\text{ber}'\,x + j\,\text{bei}'\,x}{\sqrt{j}} \quad \ldots \quad (70)$$

where ber $x$ and bei $x$ are the real and imaginary parts of $I_0(\sqrt{j}\,x)$ and ber′ $x$ and bei $x$ are the differential coefficients of ber $x$ and

TABLE II

| $v = b\sqrt{\omega\mu g}$ | $P_5$ | $P_6$ |
|---|---|---|
| 0 | 1.000 | 0.0000 |
| 1.0 | 1.005 | 0.1247 |
| 2.0 | 1.078 | 0.4806 |
| 3.0 | 1.318 | 0.9508 |
| 4.0 | 1.678 | 1.373 |
| 5.0 | 2.043 | 1.811 |
| 6.0 | 2.394 | 2.093 |
| 7.0 | 2.743 | 2.451 |
| 8.0 | 3.094 | 2.809 |
| 9.0 | 3.446 | 3.165 |
| 10.0 | 3.799 | 3.520 |

$b$ = radius of rod in cm.
$\mu$ = intrinsic permeability of the material.
$g$ = specific conductance of the material in mhos per cm. cube.

bei $x$ respectively. Tables of ber and bei functions are available to facilitate numerical calculations.

Substituting (69) and (70) in equation (66) and replacing $1/\pi g b^2$ by $R_0$ gives

$$R_{bb} = R_0 \cdot \frac{v}{2} \cdot \frac{\mathrm{ber}\,v\;\mathrm{bei}'\,v - \mathrm{bei}\,v\;\mathrm{ber}'\,v}{(\mathrm{ber}'\,v)^2 + \mathrm{bei}'\,v)^2}$$

$$= R_0 \,.\, P_5 \qquad \ldots \quad \ldots \quad (71)$$

$$\omega L_{bb} = R_0 \cdot \frac{v}{2} \cdot \frac{\mathrm{ber}\,v\;\mathrm{ber}'\,v + \mathrm{bei}\,v\;\mathrm{bei}'\,v}{(\mathrm{ber}'\,v)^2 + (\mathrm{bei}'\,v)^2}$$

$$= R_0 \,.\, P_6 \qquad \ldots \quad \ldots \quad (72)$$

where $v = b\sqrt{\omega \mu g}$ .. .. .. (73)

To facilitate calculation the portion of the equations involving $v$, viz. $P_5$ and $P_6$ have been tabulated in Table II.

## 5. Numerical Examples

(a) *Surface impedance of plane plated conductors*

Figures 4, 5 and 6 give the surface resistance and reactance of plane plated conductors consisting of tinned copper, nickel-plated copper and copper-plated iron respectively. In each case it has been assumed that the base metal is so thick that its surface impedance is practically independent of the actual thickness. This has enabled the curve to be plotted in terms of $R/\sqrt{f}$ and $\omega L/\sqrt{f}$ against $t\sqrt{f}$ (allowing the surface impedance to be readily evaluated at any frequency for various thicknesses of plating) where $R$ is the resistance, and $\omega L$ the reactance in ohms per sq. cm. $f$ the frequency in cycles per sec. and $t$ the thickness of the plating in cms. In calculating the curves the values given in Table III have been taken for the constants of the materials.

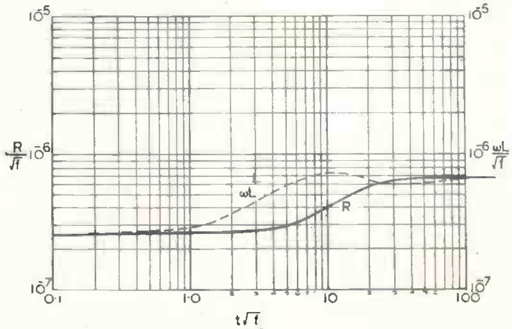


Fig. 4. *Surface resistance and reactance of tinned copper conductor.*

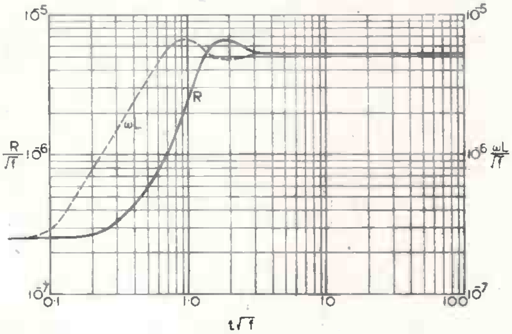


Fig. 5. *Surface resistance and reactance of nickel-plated copper conductor.*

Fig. 5 shows that the thickness of the nickel plating must not be greater than that

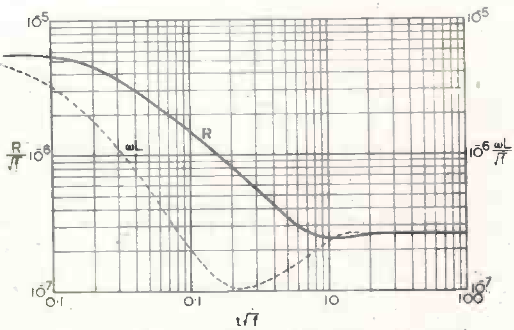


Fig. 6. *Surface resistance and reactance of copper-plated iron conductor.*

TABLE III

| Material | Intrinsic permeability $\mu \times 10^{-6}$ | Permeability $\mu/\mu_0$ | $g$ in mhos per cm.³ | Specific resistance in ohms per cm.³ |
|---|---|---|---|---|
| Copper .. .. | 0.01257 | 1 | $5.800 \times 10^5$ | $1.72 \times 10^{-6}$ |
| Tin .. .. | 0.01257 | 1 | $0.868 \times 10^5$ | $11.5 \times 10^{-6}$ |
| Nickel .. .. | 1.257 | 100 | $1.43 \times 10^5$ | $7.0 \times 10^{-6}$ |
| Iron .. .. | 1.257 | 100 | $1.0 \times 10^5$ | $10 \times 10^{-6}$ |

value which will make $t\sqrt{f}$ equal to 0.2 approximately at the frequency under consideration if it is desired that the surface resistance of the conductor should not be more than 5 per cent. above that of pure copper. For a similar result in the case of tinned copper $t\sqrt{f}$ must not be greater than 2.5 approximately.

Fig. 7 gives a comparison between the surface resistance of a plane copper-plated iron conductor and that of a plain copper conductor having a thickness comparable

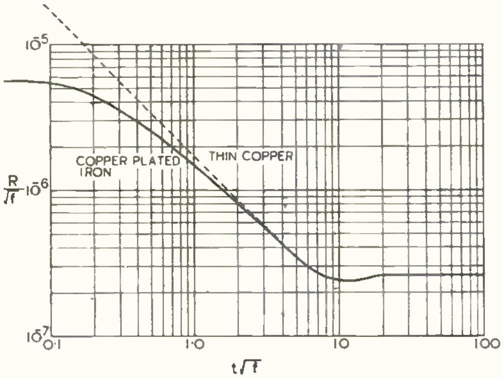


Fig. 7. *Comparison of surface resistance of copper-plated iron conductor and thin copper conductor.*

with the penetration depth of the currents flowing in the conductor. It is seen that for values of $t\sqrt{f}$ greater than 4 the surface resistances of both conductors are practically identical. Below this value of $t\sqrt{f}$ the curves diverge owing to the finite surface impedance of the iron. It is interesting to note in the case of the copper-plated iron, and the plain copper, conductors that the surface resistance is a minimum when $t\sqrt{f} = 12$, its value then being 0.92 of that for an infinitely thick copper conductor.

The "penetration depth" has frequently been defined as that depth at which the magnitude of the H.F. current is (a) $1/e$ of its magnitude at the surface of the conductor; (b) the depth at which the H.F. current is 1 per cent. of its surface value. The values of penetration depth obtained from these definitions may be readily compared with the above thickness for minimum resistance by means of equation (4). If $I_b$ is the magnitude of the current at the surface and $I_t$ its value at a depth $t$

$$\frac{I_b}{I_t} = e^{at} = e^{\sqrt{\frac{\omega\mu g}{2}}\,t}$$

$$2.3026 \log_{10}\frac{I_b}{I_t} = \sqrt{\frac{\omega\mu g}{2}}\,t$$

(a) $\frac{I_b}{I_t} = e$ when $t = \sqrt{\frac{2}{\omega\mu g}}$

(b) $\frac{I_b}{I_t} = 100$ when $\frac{2.3026 \times 2}{\sqrt{\frac{\omega\mu g}{2}}}$

In the case of copper conductors these become

(a) $t\sqrt{f} = 6.6$ $t$ in cm., $f$ in cycles per sec.

(b) $t\sqrt{f} = 30.4$

in comparison with $t\sqrt{f} = 12$ for minimum resistance.

(b) *Co-axial cables*

Jarvis and Fogg[3] have suggested the use of copper-plated iron or thin copper conductors as a means of equalising the attenuation of a co-axial cable over a wide band of frequencies, and as a means of reducing delay distortion.

Fig. 8 illustrates the use of copper-plated iron conductors as a means of equalising the attenuation of a co-axial cable over a wide frequency band. The overall diameter of

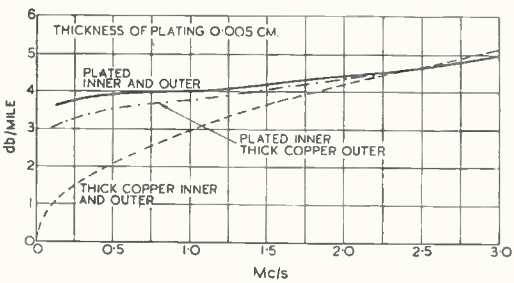


Fig. 8. *Frequency attenuation curves for co-axial cable composed of copper-plated iron conductors. Thickness of plating* 0.005 *cm.*

the inner conductor was taken as 0.32 cm. and the internal diameter of the outer conductor as 1.15 cm., the thickness of the plating being taken as 0.005 cm. The full line gives the frequency attenuation curve for a cable composed of a plated inner and a plated outer conductor, while the dash-dot line gives the corresponding curve for a cable having a plated inner but a plain outer conductor. For comparison the frequency attenuation curve for a similar cable with plain thick copper conductors is shown dotted.

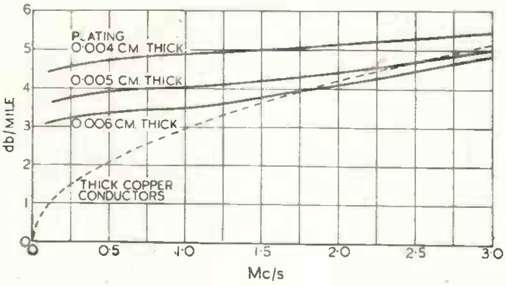


Fig. 9. *Frequency attenuation curves for co-axial cable composed of copper-plated iron conductors for three thicknesses of plating.*

Owing to the greater restriction of the currents to the surface layers of the plated conductors, the surface inductance varies much less with frequency than with a cable composed of thick copper conductors resulting in a more constant velocity of propagation with a corresponding reduction in the delay distortion.

Fig. 9 shows the effect of a small change in the thickness of the plating on the frequency attenuation curve.

Figs. 8 and 9 probably represent about the maximum degree of equalisation that can be attained over this frequency range by the use of copper-plated iron conductors.

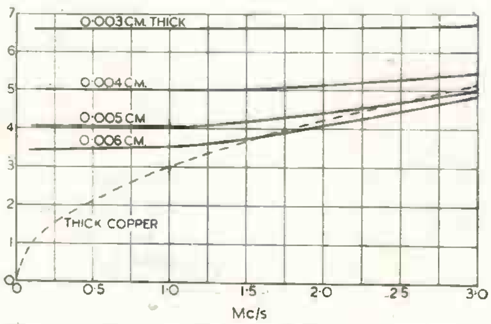


Fig. 10. *Frequency attenuation curves for co-axial cables composed of thin copper conductors of various thicknesses.*

For a higher degree of equalisation it is necessary to use thin copper conductors, having a thickness of, say, 0.003 or 0.004 cm.

Fig. 10 gives the frequency attenuation curves for a similar co-axial cable composed of thin copper conductors for comparison with Figs. 8 and 9. The falling off in the attenuation curves of the plated cable below 0.5 Mc/s due to the finite conductivity of the iron under layer is clearly visible.

### REFERENCES

[1] S. A. Schelkunoff. *Bell S. Tech. Journ.*, Vol. 13, p. 532, 1934.
[2] G. W. O. Howe. *Journ. I.E.E.*, Vol. 54, p. 473, 1916.
[3] R. F. J. Jarvis and G. H. Fogg. *P.O. Elec. Eng. Journ.*, Vol. 30, p. 138, 1937–38.

## Indexes and Binding Cases

The 48-page Subject and Author Index to the Abstracts and References published in *Wireless Engineer* during 1942 will be available on February 15th. As supplies are limited we are asked once again to stress the need for placing orders early. The price will be 2s. 8d., including postage.

Binding cases for Volume XIX, of *Wireless Engineer*, January to December, 1942, are now available and cost 3s. (postage 3d.). Arrangements can be made for binding readers' copies at an inclusive cost of 10s., plus 9d. to cover the postage when returning the bound volume.

## February Meetings

At the next meeting of the I.E.E. Wireless Section, to be held on February 3rd at 5.30 p.m., Professor Willis Jackson, D.Sc., D.Phil., will deliver a paper on "The University Education and Industrial Training of Telecommunication Engineers."

"Industrial Applications of Electronics" is the subject of the paper to be read by J. H. Reyner, B.Sc., A.M.I.E.E., at the next meeting of the Brit. I.R.E., which will be held on February 19th at the Institution of Structural Engineers, 11, Upper Belgrave Street, London, S.W.1.

## Book Received

**D/F Handbook for Wireless Operators.** (2nd Edition). By W. E. Crook. In preparing the second edition of this book, the first edition of which was reviewed in our issue of June 1942, the author has taken the opportunity of adding a chapter on aircraft D/F. It is a strictly practical book in which a knowledge of elementary theory is assumed. Pp. 84. Diagrams 111. Sir Isaac Pitman and Sons, Ltd., 39, Parker Street, London, W.C.2. Price 3s. 6d.

## Back Numbers

Our Publishers advise us that a subscriber who helped to supply the back numbers of *Wireless Engineer* asked for in our November issue is himself in need of the following copies :—August, 1927 ; September, 1928 ; November, 1930 ; December, 1932 ; February and October, 1933.

# Receiver Input Circuits*

## Design Considerations for Optimum Signal/Noise Ratio

*By R. E. Burgess, B.Sc.*

*(Radio Department, National Physical Laboratory)*

**SUMMARY.**—The problem discussed is the design of the inductively coupled input circuit of a receiver for optimum signal/noise ratio.

The general case for any aerial impedance is analysed first assuming that valve noise is small compared with circuit noise, and that the effective temperature of the radiation resistance is the same as that of the circuit. Expressions are derived for the signal/noise ratio at the grid of the first valve and for the optimum coupling, and it is shown that the coefficient of coupling and the $Q$'s of the input transformer should be as large as possible. To extend the analysis to higher frequencies the influence of valve noise on the optimum coupling is discussed.

For a loop aerial the optimum coupling coil inductance is approximately 0.4 that of the loop. A numerical example is calculated for a single-turn loop working over a frequency band of 0.15–6 Mc/s. (2,000–50 m.). In the case of a resistive aerial the optimum coupling is relatively small when valve noise is significant, and, as for the loop aerial, its value is not critical.

For a capacitive aerial the optimum condition occurs in the region of the primary resonance. Since it is sometimes undesirable to have this resonance in the working band, it may be necessary to displace it to one or other end, whichever gives the more uniform sensitivity over the band.

When the aerial is reactive, the signal/noise ratio can sometimes be improved by tuning the primary circuit and this also enables the aerial to be matched to a transmission line. The optimum conditions for the untuned and tuned primary systems are summarised in a Table.

In the Appendix the loss in signal/noise ratio due to a transmission line link is calculated and the conditions for minimising the loss are deduced.

### 1. Introduction

IT is well known that the ultimate sensitivity of a receiving system depends upon the irreducible minimum noise arising in the input circuit and the first valve. Thus a figure of merit for a receiving system should take noise into account, and for this reason the conceptions of effective height and pick-up factor are unsatisfactory. A more useful parameter is the ratio of the signal and noise p.d.'s at the grid of the first valve, in which valve noise can be readily incorporated by using the familiar concept of "equivalent noise resistance."

A generalised analysis of the signal/noise ratio in an input circuit given by the present author in a recent paper [1] in which the problem of assigning an equivalent temperature to the radiation resistance of an aerial was also discussed. However, this analysis, like others which gave a general treatment had the disadvantage of not indicating the *practical* design of input circuits. It is particularly important to use practical variables when determining the optimum conditions and to ensure that these conditions are feasible in a working system.

The object of the present paper is to discuss the more important features in the design of input circuits for maximum signal/noise ratio.

### 2. General Analysis of the Coupled Circuit

The input circuit which is most frequently encountered in a receiver is shown in Fig. 1. The signal source (e.g. aerial) has a signal e.m.f. $e$, and is of impedance $Z$; this is coupled by a coil $l$, $r$ to the first tuned circuit $L_2C_2R_2$ of the receiver by a mutual inductance $M$ corresponding to a coefficient of coupling $k$.

In calculating the noise present in the input circuit, it will be assumed that valve damping and feedback are absent and that all the resistive components of impedance including the radiation resistance are at a common temperature $T$. The assumption

* MS. accepted by the Editor, October, 1942.

that radiation resistance has the same effective temperature $T$ as the rest of the circuit will only lead to an appreciable discrepancy when the radiation resistance is relatively large, as in the case of a resonant open aerial[1]. Since, however, the effective temperature $Tr$ of the radiation resistance will in practice be larger than $T$ the signal/noise ratio will be determined more by

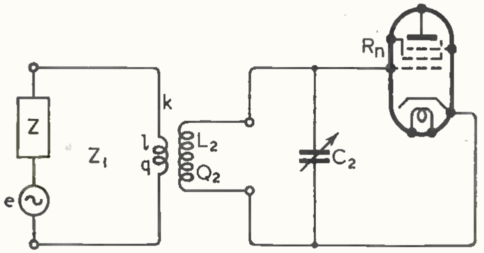


Fig. 1. *The coupled input circuit with untuned primary.*

external noise than by the circuit and thus the coupling condition will become less critical (see for example the curves in Fig. 4 of the paper referred to). It was also shown that the optimum coupling condition is independent of $Tr$.

Now the r.m.s. thermal noise e.m.f. in a resistance $R$ is given by Nyquist's formula :

$$e_n = \sqrt{4kTRB} \quad \ldots \quad (1)$$

where

$k$ = Boltzmann's constant ($1.37 \times 10^{-23}$ joule/deg.)

$B$ = effective bandwidth of receiver (c/s.) which is assumed to be sufficiently small for the impedances of the input circuit to be taken as constant within it.

The shot effect in the first valve is usually represented by the equivalent noise resistance $R_n$, which would produce the same anode current fluctuations if connected in the grid circuit of the valve. By the choice of a suitable valve it will be possible to make valve noise small compared with the circuit noise for frequencies up to about 10 Mc/s., and the signal/noise ratio will now be computed on this basis.

The ratio of the signal and noise voltages at the valve grid will always be less than that of the source itself, owing to the resistances $r$ of the coupling coil and $R_2$ of the secondary. Since the thermal noise is proportional to the square root of the resistance a convenient parameter for expressing the signal/noise ratio at any two terminals is the ratio $\rho$ of the signal p.d. $e'$ between the terminals to the square root of the resistive component $R'$ of the impedance at the terminals. The upper limit to this ratio is the value at the source :—

$$\rho_0 = \frac{e}{\sqrt{R}} \quad \ldots \quad (2)$$

where $R$ is the resistive component of $Z$.

The value of $\rho$ for the secondary circuit is given by

$$\rho = \frac{e'}{\sqrt{R'}} = e \frac{\omega M / |Z_1|}{\sqrt{R_2 + \left|\frac{\omega M}{Z_1}\right|^2 R_1}}$$

$$= \frac{e}{\sqrt{R_1 + \left|\frac{Z_1}{\omega M}\right|^2 R_2}} \quad \ldots \quad (3)$$

where $R_1 = R + r$ is the total primary resistance.

Thus the coupling circuit reduces the signal/noise ratio in the ratio

$$\frac{\rho}{\rho_0} = \frac{1}{\sqrt{1 + \frac{r}{R} + \left|\frac{Z_1}{\omega M}\right|^2 \frac{R_2}{R}}}$$

The signal/noise ratio $\rho$ is proportional to the square root of the available power $e'^2/4R'$ at the two terminals, and this is because the ratio of the square of the noise e.m.f. to $R'$ is $4kTB$ at any two terminals of the circuit, for it is assumed that the temperature and bandwidth are constant throughout the circuit. Thus the loss in signal/noise ratio $\rho/\rho_0$ is the square root of the available signal powers at the source and at the terminals of $C_2$.

In order to determine the optimum coupling condition, it is necessary to specify exactly what variables are involved. In practice the variables are the coefficient of coupling $k$ and the coupling coil inductance $l$ whose resistance $r$ is assumed to be proportional to $l$. Let $\frac{\omega l}{r} = q$ and $\frac{\omega L_2}{R_2} = Q_2$.

Then the reduction in signal/noise due to the coupling circuit is given by the ratio :—

$$\frac{\rho}{\rho_0} = \frac{1}{\sqrt{1 + \frac{\omega l}{qR} + \frac{(X + \omega l)^2 + (R + \omega l/q)^2}{\omega k^2 l Q_2 R}}}$$

$$= \frac{1}{\sqrt{1 + \frac{|Z|^2}{\omega k^2 l Q_2 R} + \frac{\omega l}{R}\left(\frac{1}{q} + \frac{1}{k^2 Q_2 q^2} + \frac{1}{k^2 Q_2}\right) + \frac{2}{k^2 Q_2}\left(\frac{X}{R} + \frac{1}{q}\right)}} \quad \ldots \quad \ldots \quad \ldots \quad (4)$$

The signal/noise ratio in the secondary circuit is independent of $C_2$ since a pure reactance connected in parallel with a network will not affect the ratio of signal to thermal noise at its terminals.

It is seen that in order to make the signal/noise ratio a maximum

(i) $k$, $Q_2$ and $q$ should be as large as possible and

(ii) $\omega l$ should be given its optimum value :

$$\omega l_{\text{opt.}} = \frac{|Z|}{\sqrt{1 + \frac{1}{q^2} + \frac{k^2 Q_2}{q}}} = \frac{|Z|}{\sqrt{1 + \frac{k^2 Q_2}{q}}}$$

very nearly. .. .. (5)

It is of interest to note that as far as the impedance of the source is concerned, only its *modulus* $|Z|$ and not the power factor enters, also that there is a definite optimum coupling condition even when valve noise is negligible as assumed above. This does not arise from the existence of resistance in the coupling coil, since if this is zero ($q = \infty$) a definite optimum condition still exists.

When $l$ is given the optimum value of equation (5) the signal/noise ratio is given by

$$\frac{\rho_{\text{opt.}}}{\rho_0} = \frac{1}{\sqrt{1 + \frac{2}{k^2 Q_2}\left[\frac{X}{R} + \frac{|Z|}{R}\sqrt{1 + \frac{k^2 Q_2}{Q}}\right]}}$$

very nearly .. .. (6)

## 3. The influence of Valve Noise on the Optimum Coupling Condition

In the above it has been assumed that valve noise is small compared with circuit noise, but at frequencies above about 10 Mc/s. this may no longer hold, since the dynamic impedance of the tuned circuit falls off as the frequency increases, and eventually becomes of the same order as the equivalent noise resistance of the valve.

Let $D$ be the effective dynamic impedance of the tuned secondary circuit (when the aerial is coupled to it) and $R_n$ the equivalent noise resistance of the valve. The total effective noise e.m.f. at the grid is then

$$e_n = \sqrt{4kTB(D + R_n)} \quad \ldots \quad \ldots \quad (7)$$

and thus the reduction of the signal/noise ratio below the value at the aerial can be written as

$$\frac{\rho'}{\rho_0} = \frac{\rho}{\rho_0}\sqrt{\frac{D}{D + R_n}} = S\sqrt{\frac{R}{D + R_n}} \quad (8)$$

where $S$ is the transfer ratio from the aerial to the terminals of the tuned circuit and is related to $\frac{\rho}{\rho_0}$ by

$$\frac{\rho}{\rho_0} = S\sqrt{\frac{R}{D}} \quad \ldots \quad \ldots \quad \ldots \quad (9)$$

from the definitions of $\rho_0$ and $\rho$.

It is seen from equation (8) that if the coupling to the aerial is loose or valve noise is large compared with the circuit noise $(D + R_n)$ will be substantially independent of the coupling and the couplings for optimum signal and for optimum signal/noise ratio are the same. Thus the optimum coupling for any value of valve noise will always lie between the coupling for maximum $S$ and for maximum $\frac{\rho}{\rho_0}$ and it is useful to remember this when studying the curves in this paper.

The overall reduction in signal/noise ratio $\frac{\rho'}{\rho_0}$ below the value in the aerial has the factor $\rho/\rho_0$ due to the coupling circuit and $\sqrt{D/D + R_n}$ due to the valve noise. This latter factor is directly related to the representation of valve noise by the "distune ratio" $\Delta$ which is defined as the ratio of the r.m.s. noise voltage with the input circuit normally tuned, to that when it is completely distuned and is thus given by

$$\Delta^2 = \frac{D + R_n}{R_n} \quad \ldots \quad \ldots \quad (10)$$

or $\sqrt{\dfrac{D}{D+R_n}} = \sqrt{1 - \dfrac{1}{\Delta^2}}$ .. .. (11)

Since absolute measurements of signal/noise ratio are rather difficult, it is useful to note that relative values can be obtained by determining the transfer ratio $S$, the effective dynamic impedance $D$ and the distune ratio $\Delta$ and using equation (8).

The signal/noise ratio $\rho/\rho_0$ with valve noise absent, when using the coupling for maximum signal is not likely to be more than 3 db. down on the optimum value at the coupling for maximum $\dfrac{\rho}{\rho_0}$ since the dynamic impedance of the tuned circuit is approximately halved in the condition for maximum signal; with valve noise present the difference is even less.

The quantity $\sigma$ used in the later sections to represent the variation of signal with coupling is defined by

$$\sigma \equiv S\sqrt{\frac{R}{D_0}} = \frac{\rho}{\rho_0}\sqrt{\frac{D}{D_0}}$$

from equation (9) .. .. (12)

where $D_0$ is the original value of the dynamic impedance of the tuned circuit.

It is seen that $\sigma$ is the ratio of the actual transfer ratio $S$ to twice the transfer ratio $0.5\sqrt{D_0/R}$ between resistances $R$ and $D_0$ when they are coupled and matched by an ideal transformer.

In the analysis of the preceding section the optimum $\dfrac{\rho}{\rho_0}$ was determined, and now with valve noise present, it is the maximum of $\dfrac{\rho'}{\rho_0}$ which is required; since in general a decrease of $l$ below the value of $l_{opt.}$ given by equation (5) will increase the effective dynamic impedance $D$ of the secondary circuit, and also the term $\sqrt{\dfrac{D}{D+R_n}}$ the optimum $l$ with valve noise present will be smaller than when it is absent. Furthermore, the coefficient of coupling should still be as large as possible, since for a given mutual inductance $M$ the inductance $l$ can then be made smaller, and with it $r$ which contributes to the thermal noise in the input circuit.

## 4. Inductive Aerial (Loop)

Let the loop inductance be $L$ and resistance be $R$, and $Q = \omega L/R$. Then as $Z = j\omega L + R$ we have from equation (5)

$$l_{opt.} = \frac{L}{\sqrt{1 + \dfrac{k^2 Q_2}{q}}} \quad .. \quad .. \quad (13)$$

which is practically independent of frequency.

Thus in practice the optimum coupling coil inductance will lie between 0.4 $L$ to $L$ depending upon the ratio $k^2Q_2/q$. In the optimum condition we have from equation (6)

$$\frac{\rho_{opt.}}{\rho_0} = \frac{1}{\sqrt{1 + \dfrac{2Q}{k^2Q_2}\left[1 + \sqrt{1 + \dfrac{k^2Q_2}{q}}\right]}} \quad .. \quad .. \quad (14)$$

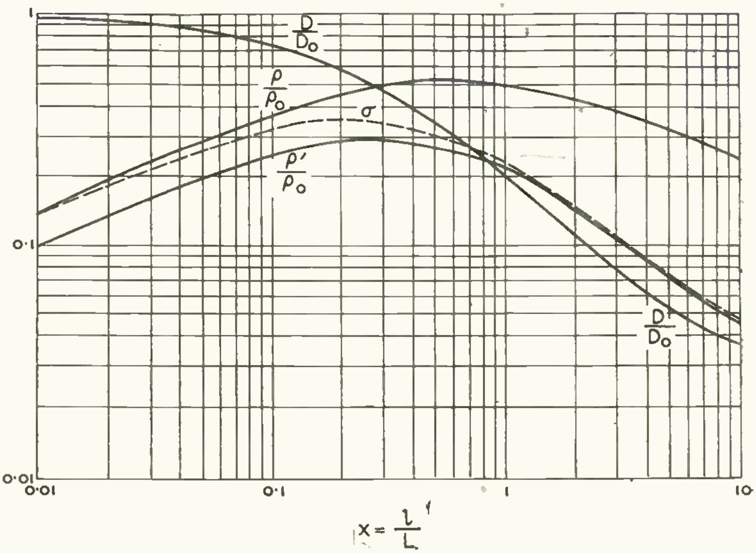


Fig. 2. *Behaviour of the coupled input circuit with an inductive aerial* ($k^2 = 0.75$; $Q = q = 60$; $Q_2 = 160$).

For the general case when $l$ has a value given by

$$l = xL$$

the signal/noise ratio is given by

$$\frac{\rho}{\rho_0} = \frac{1}{\sqrt{1 + \frac{xQ}{q} + \frac{Q}{k^2 Q_2} \frac{(1+x)^2}{x}}} \quad .. \quad (15)$$

Fig. 2 gives curves for the typical case of a closely coupled input transformer:

$$k^2 = 0.75, \quad Q = q = 60, \quad Q_2 = 160.$$

The signal function $\sigma$ has its maximum of 0.35 at $\frac{l}{L} \doteqdot 0.2$ while the signal/noise ratio $\frac{\rho}{\rho_0}$ with valve noise absent reaches a maximum of 0.52 when $\frac{l}{L}$ is 0.58; in practice it will not be possible to exceed appreciably this favourable condition and 0.6 may be taken as the practical upper limit for $\frac{\rho}{\rho_0}$. Curves are also given for $\frac{\rho'}{\rho_0}$ representing (for $R_n = D_0$) the signal/noise ratio with valve noise present: the optimum value of $l/L$ is 0.3 for which $\frac{\rho'}{\rho_0}$ is 0.28. It is seen that the adjustment of $l$ is uncritical and a useful working rule for most cases is:—

$$l_{\text{opt.}} = 0.4\,L \quad .. \quad .. \quad .. \quad (16)$$

As the curves refer to the ratio $\rho/\rho_0$ it is important to make $\rho_0$ as large as possible. Since $\rho_0$ is the ratio of the signal e.m.f. in the loop to the latter's resistance, an important parameter in determining the performance of a loop is the ratio

$$\frac{nA}{\sqrt{R}} = nA\sqrt{\frac{Q}{\omega L}}$$

where $nA$ is the product of area and number of turns of the loop. Thus the loop $Q$ should be as high as possible as should the term $\frac{nA}{\sqrt{L}}$ which is seen to be an important criterion of loop efficiency.

Despite the loss in signal/noise ratio of about 6 db. in the coupled system compared with the directly tuned loop it has the advantage of not requiring separate loop tuning or a symmetrical input circuit, and enables one loop (usually of a single turn) to be used over a wide frequency range while the effect of a transmission line link between loop and receiver is considerably reduced. These advantages may be of decisive importance in some practical installations, especially in mobile apparatus. To illustrate the design of such a coupled loop system consider the use of a single turn loop 50 cm. square and of inductance 2.5 $\mu$H.

Let $Q = q = 60$, and $Q_2 = 160$, which are assumed to be constant over the working band, and $k = 0.5$.

Then from equation (13) the optimum inductance for the coupling coil is found to be

$$l_{\text{opt.}} = 1.9\ \mu\text{H}$$

and in the optimum condition, by equation (14)

$$\frac{\rho}{\rho_0} = 0.36$$

If $E$ is the intensity of signal in $\mu$V/m. the signal/noise ratio in the maximum position of the loop for a receiver with an overall bandwidth $B$ of 10 kc/s. is

$$\frac{S}{N} = \frac{\rho}{\rho_0} \cdot \frac{2\pi nA}{\lambda} \frac{E}{\sqrt{4kTB\frac{\omega L}{Q}}} = 5.1\frac{E}{\sqrt{\lambda}} \quad .. \quad .. \quad (17)$$

If the loop is used in the waveband 50–2,000 m. (6–0.15 Mc/s.) it will be easy to make the valve noise negligible compared with circuit noise and the field intensity required for unity signal/noise ratio at the loop maximum ranges from 1.4 $\mu$V/m. at 50 m. to 8.7 $\mu$V/m. at 2,000 m.

If, as is customary, the loop is linked to the receiver by a transmission line, the effective values of inductance, resistance and e.m.f. looking into the line from the receiver end are the relevant values in this calculation (see Appendix). Such a line should have low loss in order that it should not contribute appreciably to the circuit noise, and the apparent inductance of the loop seen from the receiver should not vary excessively with frequency in the working band.

The high $k$, $Q_2$ and $q$ can be obtained by the use of an iron-cored transformer since in the coupled loop system a screened transformer is usually not needed, and symmetry may be achieved by earthing the centre point of the coupling coil.

## 5. Resistive Aerial

Typical examples of a resistive source are a resonant aerial and a transmission line terminated in its characteristic impedance at its far end. We now have

$$X = 0, \text{ and } |Z| = R$$

and thus equation (5) becomes

$$\omega l_{opt.} = \frac{R}{\sqrt{1 + \dfrac{k^2 Q_2}{q}}} \quad \ldots \quad \ldots \quad (18)$$

and from equation (6)

$$\frac{\rho_{opt.}}{\rho_0} = \frac{1}{\sqrt{1 + \dfrac{2}{k^2 Q_2}\sqrt{1 + \dfrac{k^2 Q_2}{q}}}} \quad (19)$$

However, the degree of coupling given by this "optimum" condition is too great, since a large reduction in the dynamic impedance of the secondary would result.

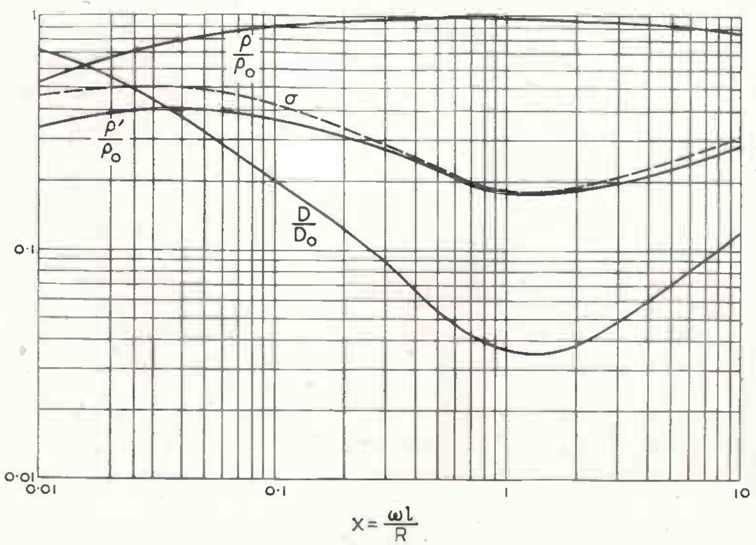


Fig. 3. *Behaviour of the coupled input circuit with a resistive aerial* ($k = 0.5$; $q = 60$; $Q_2 = 160$).

The general expression for $\rho/\rho_0$ is, from equation (4)

$$\frac{\rho}{\rho_0} = \frac{1}{\sqrt{1 + x\left(\dfrac{1}{q} + \dfrac{1}{k^2 Q_2}\right) + \dfrac{1}{x}\,\dfrac{1}{k^2 Q_2} + \dfrac{2}{k^2 Q_2 q}}} \quad \ldots \quad \ldots \quad (20)$$

where $x = \dfrac{\omega l}{R}$

The curves of Fig. 3 have been plotted for the case where

$$k = 0.5, \qquad q = 60, \qquad Q_2 = 160,$$

and as in the case of the loop antenna the variation of signal and signal/noise ratio with the coupling is uncritical. It is seen that only a small value of coupling is required to produce an appreciable reduction in dynamic impedance, while the signal/noise ratio is very little dependent on $x$; the curves can be taken as representing the variation with frequency for a given $l$ or variation with $l$ for a given frequency. It is seen that $\rho/\rho_0$, reaches its maximum in the region of $x = 1$ while $D/D_0$ and $\sigma$ are there at their minima. More exactly $\rho/\rho_0$ has its maximum of 0.97 at $x = 0.77$, but since $D/D_0$ is there 0.04, valve noise would become prohibitive on short waves, and this is illustrated by the curve for $\rho'/\rho_0 (R_n = D_0)$ whose maximum is 0.41 and occurs at the much smaller coupling of $x = 0.035$. If this represents too small an $l$ to be obtained with ease it may be necessary to reduce $k$ as a practical expedient.

The analysis for a resistive source with valve noise present corresponds to the case treated in the paper [1] mentioned earlier. It is there shown that the optimum condition when the primary circuit is resistive is given by

$$\frac{\omega^2 M^2}{R_1 R_2} = \frac{\omega l}{R_1} k^2 Q_2 = \sqrt{1 + \frac{D_0}{R_n}} \quad \ldots \quad (21)$$

and thus

$$x_{opt.} = \frac{\omega l_{opt.}}{R} = \frac{1}{k^2 Q_2}\sqrt{1 + \frac{D_0}{R_n}} \quad \ldots \quad (22)$$

since in the present case $R_1 = R$ very nearly and thus $R_1$ is practically independent of $l$.

The dynamic impedance of the secondary circuit is reduced in the ratio

$$\frac{D}{D_0} = \frac{R_n}{D_0}\left(\sqrt{1 + \frac{D_0}{R_n}} - 1\right) \quad \ldots \quad (23)$$

and the resulting signal/noise ratio is given by:

$$\frac{\rho'_{opt.}}{\rho_0} = \sqrt{\frac{R_n}{D_0} + 1} - \sqrt{\frac{R_n}{D_0}} \quad \ldots \quad (24)$$

The coupling for maximum signal is given by $x_{opt.} = \frac{1}{k^2 Q_2}$ which has the value of 0.025 for the example considered above, at which $\frac{\rho}{\rho_0}$ is 0.7 of its optimum value.

## 6. Capacitive Aerial

The capacitive aerial is one of practical interest, since it embraces open aerials working at frequencies below their resonant frequencies. Let the capacitive source have impedance

$$Z = \frac{1}{j\omega C} + R \qquad \text{and let}$$

$$Q = \frac{1}{\omega C R}$$

then from equation (5)

$$l_{opt.} = \frac{1}{\omega^2 C} \frac{1}{\sqrt{1 + \frac{k^2 Q_2}{q}}} \quad \ldots \quad \ldots \quad (25)$$

which implies an inductance less than that required to resonate at $\omega$ with $C$, such that the resonant frequency of the primary circuit lies above the working frequency. In the optimum condition

$$\frac{\rho_{opt.}}{\rho_0} = \frac{1}{\sqrt{1 + \frac{2Q}{k^2 Q_2}\left[-1 + \sqrt{1 + \frac{k^2 Q_2}{q}}\right]}} \quad \ldots \quad \ldots \quad (26)$$

It is seen from equation (25) that the capacitive antenna is the most difficult to operate over a wide frequency band, since the optimum condition is dependent on frequency and because of the two primary resonance conditions at which the effective dynamic impedance of the secondary falls to a very low value.

The general expression for $\frac{\rho}{\rho_0}$ is

$$\frac{\rho}{\rho_0} = \frac{1}{\sqrt{1 + \frac{xQ}{q} + \frac{Q}{k^2 Q_2}\frac{(x-1)^2}{x}}} \quad (27)$$

where $x = \omega^2 l C$.

This equation can be used to study the variation of $\frac{\rho}{\rho_0}$ with frequency for a given $l$, or with $l$ for a given frequency.

The capacitive aerial has a $Q$ which is usually relatively high by comparison with the $q$ of the primary winding of the transformer, and which tends to be rather more dependent on frequency. In Fig. 4 thecurves are given for

$k = 0.5$, $Q = 240$, $q = 80$, and $Q_2 = 160$.

The dynamic impedance drops to a very low

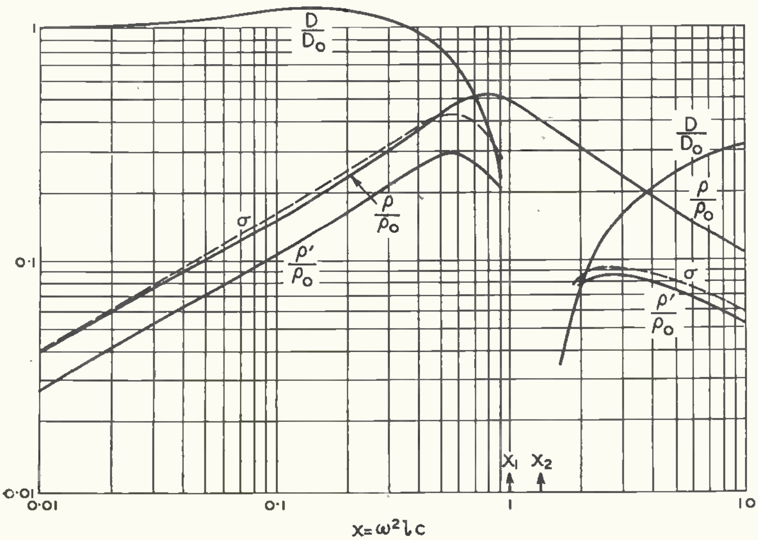


Fig. 4. *Behaviour of the coupled input circuit with a capacitive aerial* ($k = 0.5$; $Q = 240$; $q = 80$; $Q_2 = 160$).

value in the region of the resonances of $C$ with $L_1$ and with its leakage inductance $L_1(1 - k^2)$ :—

$$\left.\begin{aligned} x_1 &= 1 \\ x_2 &= \frac{1}{1 - k^2} \end{aligned}\right\} \quad \ldots \quad (28)$$

On the high frequency side of these resonances ($x \gg x_2$) the dynamic impedance tends to an upper limit given by

$$\frac{D}{D_0} = \frac{(1 - k^2)^2}{1 + \dfrac{k^2 Q_2}{q}}$$

which equals 0.375 for $k = 0.5$ (Fig. 4) when $Q_2/q = 2$. Thus with close coupling $\frac{\rho'}{\rho_0}$ is considerably smaller on the high-frequency side of resonance than on the low frequency side.

It will in general be possible to work on both sides of the resonances ($x < x_1$ and $x > x_2$) but the intermediate region ($x_1 < x < x_2$) is so narrow and the dynamic impedance and secondary tuning so variable that it is unusable. It is seen from equation (26) that $\frac{\rho}{\rho_0}$ rises to a value at the maximum which is practically independent of $k$ for values up to $k = 0.5$, and is given by

$$\frac{\rho_{opt.}}{\rho_0} = \frac{1}{\sqrt{1 + \dfrac{Q}{q}}}$$

very nearly .. .. (29)

but that as $x$ departs from $x_{opt.}$ $\frac{\rho}{\rho_0}$ falls off more rapidly the smaller the factor $k^2Q_2/Q$.

## 7. The Tuned Primary System

In sections 4 and 6 dealing with reactive sources, the primary circuit was untuned; such systems are frequently used as the design is simpler than with a tuned primary, although they have the disadvantages of

(i) the reduction in signal/noise due to the difficulty of reflecting adequate resistance from the primary into the secondary circuit unless a high coefficient of coupling is used, and

(ii) the unsuitability of the circuit for the connection of a transmission line to the secondary owing to the reactive impedance appearing there.

These defects are absent when the primary circuit is tuned and the resulting improvement in signal/noise ratio will now be considered. Let the aerial have an impedance

$$Z = jX + R \quad \text{where} \quad \frac{|X|}{R} = Q \gg 1.$$

The system may be either parallel tuned (Fig. 5a) or series tuned (Fig. 5b). When the tuning condenser $C_1$ is adjusted for primary resonance :

$$1 - \omega C_1 X + \frac{X}{\omega l} = 0 \quad \text{(parallel tuning)}$$

$$1 - \omega C_1 X - \omega^2 l C_1 = 0 \quad \text{(series tuning)}$$

If we now put

$$y \equiv \frac{\omega l}{|X|} = \frac{1}{|1 - \omega C_1 X|} \quad \text{(parallel)}$$

$$y \equiv \frac{|X|}{\omega l} = \left|\frac{\omega C_1 X}{1 - \omega C_1 X}\right| \quad \text{(series)}$$

.. .. (30a, b)

it is found that the equations which follow hold for either method of tuning if the appropriate expression for $y$ is used.

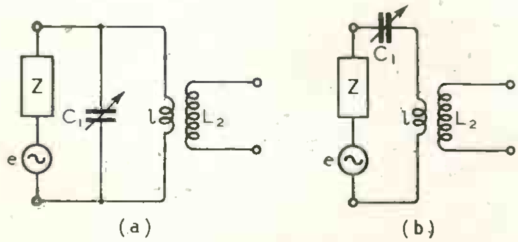


Fig. 5. *The coupled input circuit with tuned primary.* (a) *Parallel tuning;* (b) *Series tuning.*

The reduction of signal/noise ratio in the absence of valve noise is

$$\frac{\rho}{\rho_0} = \frac{1}{\sqrt{\left[1 + \dfrac{Q}{yq}\right]\left[1 + \dfrac{1}{k^2 Q_2}\left(\dfrac{1}{q} + \dfrac{y}{Q}\right)\right]}}$$

.. .. (31)

As in the case of the resistive aerial, we cannot make $k$ large and maximise with respect to $y$ for the resulting reduction in secondary dynamic impedance would lead to a serious reduction in signal/noise ratio due to the valve noise. Hence in calculating the optimum condition it is necessary to take valve noise into account from the start.

Usually $y$ or its range of values is determined by the practical consideration of tuning the primary circuit with reasonable values of capacitance $C_1$. If $k$ is the variable which is adjusted to give maximum signal/noise ratio, it may be shown from the analysis given in the earlier paper[1] (*cf.* equations (21) and (23) of the present paper) that

$$k^2_{opt.} = \frac{1}{Q_2}\left(\frac{1}{q} + \frac{y}{Q}\right)\sqrt{1 + \frac{D_0}{R_n}} \quad .. \quad (32)$$

which gives for the optimum signal/noise ratio with valve noise ($R_n$) present

$$\frac{\rho'_{opt.}}{\rho_0} = \frac{\sqrt{1 + \dfrac{R_n}{D_0}} - \sqrt{\dfrac{R_n}{D_0}}}{\sqrt{1 + \dfrac{Q}{yq}}} \quad .. \quad (33)$$

It is thus best to make $y$ as large as possible, and this means using a large $l$ for parallel tuning, or a small $l$ for series tuning. Since the equations assume different forms depending upon whether the aerial is inductive or capacitive, these two cases will be considered separately :

(*a*) *Inductive Aerial*

If $L$ is the inductance of the aerial we have $y = l/L$ (parallel tuning) or $L/l$ (series tuning) which is independent of frequency ; for comparison with the untuned case treated in section 4, let

$$y = 5$$
$$Q_2 = 160$$
$$Q = q = 60$$

then if

$$R_n = D_0, \; k_{opt.} = 0.030 \text{ giving } \frac{\rho'_{opt.}}{\rho_0} = 0.375.$$

Comparison with the value of 0.24 for $\frac{\rho'_{opt.}}{\rho_0}$ in Fig. 2 (untuned primary) shows that the gain in signal/noise ratio due to tuning is appreciable although there are the disadvantages of an additional tuning control, and the impracticability of using a single loop over a wide frequency band.

(*b*) *Capacitive Aerial.* (*Parallel Tuning*)

If $C$ is the aerial capacitance we have for the parallel tuning from equation (30a)

$$y = \omega^2 lC = \frac{C}{C + C_1}$$

Hence $y$ is always less than unity and varies as the square of the frequency. The need to make $y$ large means that the maximum value of $C_1$ must not be large compared with $C$ and thus the tuning range on the primary becomes restricted ; in practice the extent and number of frequency bands into which the total range is divided depends on the desirable compromise between simplicity and sensitivity.

To compare the tuned and untuned systems, let the frequency band have a ratio of 2 : 1 and thus if $C_1$ is assumed to range from 0 to $3C$, the range of $y$ is from 1 to 0.25. Again let

$$Q = 240 \qquad q = 80 \qquad Q_2 = 160$$

and since the optimum coupling is dependent on frequency let it be chosen to give the optimum at the geometric mean frequency ($y = 0.5$) ; then if

$$R_n = D_0, \; k_{opt.} = 0.011 \text{ giving}$$

$$\frac{\rho'_{opt.}}{\rho_0} = 0.16$$

which indicates a slightly worse performance than the untuned system having $k = 0.5$ (Fig. 4).

It may be concluded that the tuned primary system is superior to the untuned when it is not possible to obtain a coefficient of coupling greater than 0.5. Since the sensitivity with a given $l$ increases as the frequency increases, it is probably best to make the adjustment of $k_{opt.}$ for the lowest frequency in the band.

The above analysis applies to the case where the tuned secondary forms the grid circuit of the first valve. If, however, the secondary winding is connected to a transmission line, the tuned system has the great advantage over the untuned of maintaining an approximately constant resistive termination which may be adjusted to match the characteristic impedance $Z_0$ of the line, thus minimising the line loss and ensuring freedom from resonances when working over a frequency band. The loss in signal/noise ratio due to a transmission line is analysed in the Appendix.

The resistive component of the effective impedance of the secondary circuit $R_2'$ is given by

$$\frac{R'_2}{R_2} = 1 + \frac{k^2 Q_2}{\frac{1}{q} + \frac{y}{Q}} \quad \ldots \quad \ldots \quad (34)$$

and $k$ can be adjusted to equalise $R'_2$ and $Z_0$. When the system is so matched there is an additional loss of 3 db. in signal/noise ratio on the value given by equation (31). Thus if $Z_0$ is considerably greater than $R_2$ as is usually the case, the signal/noise ratio of the complete system with a tuned grid circuit at the far end of the line is given by

$$\frac{\rho}{\rho_0} = \frac{1}{\sqrt{2\left(1 + \frac{Q}{yq}\right)}} \quad \ldots \quad \ldots \quad (35)$$

assuming that the line loss is negligible when it is matched at both ends.

It is seen from equation (22) that in the case of the resistive source the primary impedance is practically wholly that of the source and thus there is no advantage to be gained by tuning the primary either in respect of signal/noise ratio or matching to a transmission line.

A resumé of the optimum coupling conditions for the tuned and untuned primary systems with the various types of aerial, is given in the Table.

## 8. Acknowledgments

The work described above was carried out as part of the programme of the Radio Research Board, to whom this paper was first circulated as a confidential report in September, 1941. It is now published by permission of the Department of Scientific and Industrial Research.

## APPENDIX

### Loss of Signal/Noise Ratio due to a Transmission Line

Let a source of signal e.m.f. $e$ and impedance $Z(= R + jX)$ be connected to a transmission line of length $d$ and constants $L_0$ $R_0$ $C_0$ $G_0$ per unit length. The characteristic impedance of the line

**TABLE**

**Resumé of optimum coupling conditions for the tuned and untuned primary systems.**

$k$ = coefficient of coupling of input transformer.
$l$ = inductance of primary of transformer (coupling coil).

| Circuit \ Aerial | Inductive ($L$) | Resistive ($R$) | Capacitive ($C$) |
|---|---|---|---|
| *Untuned Primary*<br><br>(Not suitable for connection of secondary to transmission line.) | $k$ should be large $l_{opt.} = 0.4\ L$ for all values of valve noise | $k$ need not be larger than about 0.2<br><br>$l$ should be adjusted to its optimum value given by equation (22). | $k$ should be large<br><br>If $k \geq 0.4$, $l$ should be adjusted to make the primary resonant frequency 1.2 times the upper frequency limit of the band.<br>If $k < 0.4$, $l$ should be adjusted to make the primary resonant frequency 0.8 times the lower frequency limit of the band. |
| *Tuned Primary* | Series ($y = L/l$) or parallel ($y = l/L$) tuning may be used. $l$ should be adjusted so that $y$ is large.<br><br>$k$ should be adjusted to the optimum value (generally small) given by equation (32).<br><br>If used for coupling to transmission line, $k$ should be adjusted to give a match from equation (34). | Negligible difference between the tuned and untuned cases. Either may be used for connection to a transmission line, and the matching may be obtained by adjustment of either $k$ or $l$. | Parallel tuning is usually more convenient.<br>Frequency band used with a given $l$ must be adjusted so that the decrease in sensitivity at the low frequency end is not excessive.<br>This fixes the range of $y\left(= \frac{C}{C + C_1}\right)$ and then $k$ should be adjusted to the optimum value given by equation (32) at the low frequency end.<br>If coupling to a transmission line, use equation (34) for adjustment of $k$. |

is given by

$$Z_0 = \sqrt{\frac{R_0 + j\omega L_0}{G_0 + j\omega C_0}}$$

and the propagation constant is given by

$$P \equiv \alpha + j\beta = \sqrt{(R_0 + j\omega L_0)(G_0 + j\omega C_0)}$$

The effective e.m.f. $e'$ appearing at the end of the line is given by

$$\frac{e'}{e} = \frac{1}{\cosh Pd + \frac{Z}{Z_0}\sinh Pd}$$

and the effective impedance $Z'$ there is given by

$$Z' \equiv R' + jX' = \frac{Z\cosh Pd + Z_0 \sinh Pd}{\cosh Pd + \frac{Z}{Z_0}\sinh Pd}$$

and it is the resistive component of $Z'$ which determines the noise at the end of the line.

Now usually $Z_0$ is almost wholly resistive for its imaginary part seldom exceeds about 2 per cent. of the real part, and thus to simplify the analysis $Z_0$ will be taken as a pure resistance which should lead to a close approximation in the final expression for the signal/noise ratio. Using this approximation, it is found that

$$R' = R\left|\frac{e'}{e}\right|^2\left[\cosh 2\alpha d + \frac{\sinh 2\alpha d}{2R}\left(Z_0 + \frac{|Z|^2}{Z_0}\right)\right]$$

The signal/noise ratio $\rho_0$ at the source is given by $e/\sqrt{R}$ while that ($\rho_1$) at the end of the line is related to it by

$$\frac{\rho_1}{\rho_0} = \left|\frac{e'}{e}\right|\sqrt{\frac{R}{R'}}$$

$$= \frac{1}{\sqrt{\cosh 2\alpha d + \frac{|Z|\sinh 2\alpha d}{2R}\left(\frac{Z_0}{|Z|} + \frac{|Z|}{Z_0}\right)}} \quad \ldots \quad (36)$$

If line is loss-free ($\alpha = 0$) the signal/noise ratio at the end of the line will equal $\rho_0$, since it introduces no thermal noise, or in other words $e'^2/R'$ is a constant everywhere.

In the general case $\rho_1$ reaches its maximum when

$$|Z| = Z_0 \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (37)$$

provided that if when $Z_0$ is varied $\alpha$ is constant, or if when $|Z|$ is varied $|Z|/R$ and $\rho_0$ are constant. The adjustment for the optimum condition may be made either by choosing the correct characteristic impedance of line, bearing in mind that $\alpha$ should be as small as possible, or by an adjustment of the impedance $Z$ in such a way that $\rho_0$ is not diminished.

When the impedances are equalised in accordance with equation (37)

$$\frac{\rho_{1\text{opt.}}}{\rho_0} = \frac{1}{\sqrt{\cosh 2\alpha d + \frac{|Z|}{R}\sinh 2\alpha d}} \quad \ldots \quad \ldots \quad (38)$$

and this has an upper limit of $\epsilon^{-\alpha d}$ when $X = 0$ and $Z_0 = |Z| = R$ that is when the impedances of the line and source are matched. It will also be noticed that the loss in signal/noise ratio is greatest when the source is reactive and of small power-factor since $\frac{|Z|}{R}$ can then easily be as large as 100.

To take a numerical example consider a transmission line 10 m. long and of characteristic impedance 100 ohms connected to a loop of 2.5 $\mu$H inductance with a $Q$ of 100. At a frequency of 6 Mc/s. $|Z|$ is also 100 ohms and thus equation (37) is satisfied; if the attenuation constant of the line is 8 neper/km. (2 db./100ft.) it is found from equation (38) that

$$\frac{\rho_{1\text{opt.}}}{\rho_0} = 0.24$$

which corresponds to a loss of 12 db. in signal/noise ratio. If the source were instead a resistance of 100 ohms the loss would only be 0.6 db., and thus the additional loss in the case of a reactive source is clearly illustrated.

## REFERENCE

[1] R. E. Burgess: "Noise in Receiving Aerial Systems." *Proc. Phys. Soc.*, 1941, 53, pp. 293–304.

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# British Standards

IT may interest readers to learn that the latest information regarding the issue of new and revised British Standards, of which there are at present over a thousand, can be obtained from the B.S.I. Library at 28 Victoria Street, London, S.W.1, which is open from 10.0 a.m. to 5.0 p.m., Mondays to Fridays, and at other times by appointment.

The B.S.I. Library also contains a large selection of specifications prepared by the standards bodies in other countries. Extracts from specifications may be made, if desired, and copies of the oversea specifications may be borrowed from the Library.

For readers in the provinces it may be of interest to state that most of the technical libraries, Universities and Central Public Reference Libraries maintain a complete set of British Standard specifications.

### A New Specification

A new British Standard Specification (B.S. 1082) for fixed condensers, based on technical information supplied by the British Electrical and Allied Research Association, has recently been published.

This specification is a revision of the one issued in 1926 which was limited in scope to the small condensers used at that time in wireless receivers. The drafting of a more comprehensive specification was necessitated by the considerable extension of wireless in recent years, and the developments of other activities involving the use of fixed condensers.

The present specification, which is obtainable from the B.S.I., price 2s. (2s. 3d. by post), covers all fixed condensers for general purposes, whatever the nature of the electrodes and insulant. It does not, however, apply to condensers for specialised application.

# Wireless Patents

## A Summary of Recently Accepted Specifications

*The following abstracts are prepared, with the permission of the Controller of H.M. Stationery Office, from Specifications obtainable at the Patent Office, 25, Southampton Buildings, London, W.C.2, price 1/- each*

### ACOUSTICS AND AUDIO-FREQUENCY CIRCUITS AND APPARATUS

547 890.—Cascade arrangement of amplifiers to which a high level of negative feed-back is applied without producing instability due to phase shift.

*Marconi's W.T. Co. and T. H. Price. Application date 13th March, 1941.*

548 013.—Construction of a ribbon type of microphone to give highly-directional discrimination.

*Rediffusion and E. W. Rogers. Application date, 18th February, 1941.*

548 278.—Loudspeaker in which a device, having both compliance and mass, is attached to the diaphragm in order to diminish undesired resonance effects.

*Philco Radio and Television Corporation (assignees of M. E. Swift). Convention date (U.S.A.), 29th January, 1940.*

### AERIALS AND AERIAL SYSTEMS

548 193.—Short-wave directional aerial or " horn " comprising a pair of cone-shaped elements with their apices in close proximity to a small loop or dipole terminating a coaxial feed-line.

*Standard Telephones and Cables (assignees of W. L. Barrow). Convention date (U.S.A.), 9th December, 1939.*

548 202.—Means for automatically adjusting the effective length of a dipole aerial in accordance with the tuning of the circuit to which it is coupled.

*S. Y. White. Convention date (U.S.A.), 29th February, 1940.*

548 218.—Tuning a frame aerial by means of a series coil with a sliding core of powdered iron.

*Marconi's W.T. Co. (communicated by Radio Corporation of America). Application date, 1st April, 1941.*

548 301.—Permeability - tuning arrangement for frame aerials giving constant coupling over a wide range of frequencies.

*Johnson Laboratories Inc. (assignees of W. A. Schaper). Convention date (U.S.A.) 19th February, 1940.*

### RECEIVING CIRCUITS AND APPARATUS

*(See also under Television)*

548 424.—Circuit arrangement for minimising the stray capacitances in a pentode, particularly when used for amplifying and passing pulses to a cathode follower.

*Standard Telephones and Cables and M. M. Levy. Application date, 7th April, 1941.*

548 441.—Means for stabilising the output of a frequency converter, of the kind in which subharmonics are produced in an oscillatory circuit comprising a saturable reactance.

*Telephone Manufacturing Co.; L. H. Paddle; and B. Drake. Application date 16th July, 1941.*

548 475.—Circuit for improving the working characteristic of a diode when used as a mixer in a superhet receiver.

*Philips Lamps (communicated by N. V. Philips' Gloeilampenfabrieken). Application date 29th July, 1941.*

548 501.—Pre-amplifier circuit in which the suppressor grid, say of a pentode, is used for rectification, and the output is applied to reduce the space current in the amplifier stage.

*Standard Telephones and Cables (assignees of F. B. Anderson). Convention date (U.S.A.) 10th August, 1940.*

548 509.—Circuit arrangement in which sharp pulses are generated for the detection of frequency- or phase-modulated signals.

*Standard Telephones and Cables (communicated by Western Electric Co. Inc.). Application date 22nd August, 1941.*

548 539.—Wide-band coupling circuit having a comparatively flat gain characteristic for the I.F. stage of a superhet receiver.

*Philco Radio and Television Corporation (assignees of V. R. Beck). Convention date (U.S.A.) 13th April, 1940.*

548 547.—Broad-band H.F. multi-stage amplifier giving high gain at low signal level and a high signal-to-noise ratio.

*Standard Telephones and Cables and P. K. Chatterjea. Application date 11th April, 1941.*

### TELEVISION CIRCUITS AND APPARATUS

FOR TRANSMISSION AND RECEPTION

548 337.—Construction of a photo-sensitive surface or screen of the transparent or semi-transparent type as used in television transmitters.

*Marconi's W. T. Co. (assignees E. A. Massa). Convention date (U.S.A.), 1st June, 1940.*

548 421.—Television system in which field-synchronising signals are developed independently of the usual line-synchronising pulses.

*Hazeltine Corporation (assignees of H. A. Wheeler). Convention date (U.S.A.), 6th April, 1940.*

548 443.—Means for providing a regulating bias to reproduce the slowly-fluctuating background of illumination in television transmissions.

*Marconi's W. T. Co. (assignees of H. E. Goldstine). Convention date (U.S.A.), 17th July, 1940.*

548 463.—Straightening or sharpening the flyback part of the saw-toothed current applied to the magnetic deflecting coils of a cathode-ray tube, particularly for television scanning.

*The British Thomson-Houston Co. Convention date (U.S.A.), 3rd May, 1940.*

## TRANSMITTING CIRCUITS AND APPARATUS

(*See also under Television*)

548 269.—Means for indicating the peak depth of modulation by measuring the anode current of the reactance valve in a frequency-modulating system.

*A. C. Cossor and D. A. Bell. Application date, 31st March, 1941.*

548 370.—Frequency-modulating system in which pulses are generated at an intermediate stage for the purpose of stabilising the mid-frequency of the carrier wave.

*Marconi's W.T. Co. (assignees of R. D. Kell). Convention date (U.S.A.) 31st May, 1940.*

548 346.—High-frequency generator in which electrons moving orbitally about a centre which itself moves longitudinally are "bunched" or velocity modulated by tangentially-applied fields.

*The British Thomson-Houston Co. Convention date (U.S.A.) 1st March, 1940.*

548 581.—Safeguarding arrangement for a high-powered radio transmitter comprising two oscillation-generators arranged for operation in parallel but energised from different supply sources.

*F. C. McLean. Application date, 22nd July, 1941.*

548 599.—Frequency-modulation system in which a number of spaced side-band groups, each carrying the same intelligence, are formed as an intermediate product.

*Marconi's W.T. Co. (assignees of J. E. Smith). Convention date (U.S.A.) 31st August, 1940.*

## CONSTRUCTION OF ELECTRONIC-DISCHARGE DEVICES

548 235.—Construction and arrangement of the electrodes of a cathode-ray tube designed to produce sharp contrast and half-tone effects in television.

*Standard Telephones and Cables (assignees of R. C. Winans). Convention date (U.S.A.), 25th June, 1940.*

548 570.—Method of assembling and sealing electrodes to the glass envelope of a vacuum discharge tube.

*Standard Telephones and Cables (assignees of L. C. Goodall). Convention date (U.S.A.) 8th June, 1940.*

## SUBSIDIARY APPARATUS AND MATERIALS

548 174.—Apparatus for determining the changes in composition of an electrolyte by its depolarising action.

*Wallace and Tiernan Products Inc. Convention date (U.S.A.), 18th August, 1939.*

548 228.—Mirror-galvanometer method of measuring the value and the ratio of two currents, particularly for testing electric lamps.

*J. J. Dowling and Solus Teoranta. Convention date (Eire), 2nd July, 1940.*

548 296.—Unidirectional inductive device for rapidly tripping a cut-out or safety device, irrespective of the current phase, in an A.C. system.

*The British Thomson-Houston Co. Convention date (U.S.A.), 3rd April, 1940.*

548 299.—Frequency filter for separating closely-grouped ranges of frequency.

*Telefon. Akt L. M. Ericsson. Convention dates (Sweden), 13th July and 24th November, 1938.*

548 345.—Shunt arrangement having a rising impedance-voltage characteristic for minimising the surge when an inductive circuit is broken.

*Igranic Electric Co. and C. E. Randall. Application date 3rd March, 1941.*

548 419.—Method of balancing the load on the supply, particularly in a telegraph system with double-current working.

*Standard Telephones and Cables and T. F. S. Hargreaves. Application date 7th March, 1941.*

548 517.—Method of cutting and mounting a piezo-electric crystal designed to increase the activity and stabilise the frequency of oscillation.

*Standard Telephones and Cables (assignees of S. C. Hight). Convention date (U.S.A.) 18th September, 1940.*

548 518.—Method of generating harmonically-related carrier waves, and for reducing mutual interference, in a multiplex signalling system.

*Standard Telephones and Cables (assignees of W. A. Phelps). Convention date (U.S.A.) 10th September, 1940.*

548 546.—Frequency-selector unit for a complex signal in which each distinctive tone carries some definite form of intelligence or signal.

*Standard Telephones and Cables and P. K. Chatterjea. Application date 11th Aptil, 1941.*

## The Industry

A CHANGE of address is notified by Grampian Reproducers, Ltd., who are now at Hampton Road, Hanworth, Mddx. Telephone: Feltham 2657/8.

The registered trade mark of "Simmonds" has been granted to Simmonds Aerocessories, Ltd., for their self-locking nuts.

## Paper Economies

EXAMPLES of paper economies which have been effected in industry, offices and other spheres of activity are to be seen in the exhibitions, "Paper in Battle Dress" and "Waste Paper Goes to War," which are combined in one show at the Royal Exchange, London. It was opened on Thursday, January 28th, and will remain open from ten to four daily, except Saturdays and Sundays, until February 12th.

### GOODS FOR EXPORT

**The fact that goods made of raw materials in short supply owing to war conditions are advertised in this journal should not be taken as an indication that they are necessarily available for export.**

# Abstracts and References

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*For the information of new readers it is pointed out that the length of an abstract is generally no indication of the importance of the work concerned. An important paper in English, in a journal likely to be readily accessible, may be dealt with by a square-bracketed addition to the title, while a paper of similar importance in German or Russian may be given a long abstract. In addition to these factors of difficulty of language and accessibility, the nature of the work has, of course, a great influence on the useful length of its abstract.*



## PROPAGATION OF WAVES

318. FREQUENCY . . . [Public Lectures on Centimetric Waves, Wave Guides, Cavity Resonators, etc.].—S. Ramo. (*Gen. Elec. Review*, Oct. 1942, Vol. 45, No. 10, pp. 557–567.)

319. "FUNDAMENTALS OF ELECTRIC WAVES" [providing "a Working Knowledge of the Field Theory": Book Review].—H. H. Skilling. (*Science*, 23rd Oct. 1942, Vol. 96, p. 383.)

320. COMMON-WAVE BROADCASTING AND THE INTERACTION OF MODULATION (LUXEMBOURG EFFECT).—F. Vilbig. (*T.F.T.*, May 1942, Vol. 31, No. 5, pp. 121–127.)

After a short review of the researches of Bailey & Martyn on the Luxembourg effect, and also of Grosskopf's investigations (2279 of 1938), the writer points out that although the maximum observed interference-modulation factor is at present only about 1% (so that serious trouble with broadcasting is not involved), the phenomenon would take on a more serious aspect if in the future the power of broadcasting stations were raised. For in order to gain any worthwhile advantages a mere 50 to 100% increase in power would be useless, and it would be a question of something like a ten-fold increase: this would raise the interference factor about ten times. "Thus a natural limit on any permissible increase of station powers is set by the Luxembourg effect." The question therefore arises whether the division of the total output of a station among a number of spaced common-wave transmitters would overcome this difficulty.

The writer begins by calculating the interference factor $M$ at a receiving point E in line with a common-wave station $S_2$, another common-wave station $S_1$, and finally with the "wanted" station S: low aerials (not above $\lambda/4$ in height) are assumed. Later (p. 125) the calculation is repeated for anti-fading (half-wave) aerials at the interference-producing $S_1$ and $S_2$ but a low aerial at S, and then (p. 126) for an anti-fading aerial at S also. Finally the case of three common-wave stations in a line is considered, for low aerials. As a result of these calculations it is found that since the interference factor decreases with decreasing "field-strength loading" of the ionosphere at the points of reflection, the interference caused by a given total radiated output diminishes with increasing number of common-wave transmitters and increasing spacing between them (increasing distance $D_2-D_1$ in Fig. 1): that the use of anti-fading aerials at the interfering common-wave stations, instead of low aerials, diminishes the interference, but that such a change at the "wanted" station leads to little improvement: and finally that with common-wave stations, on the assumption of similar behaviour of the ionosphere at different points, there is, in addition to a constant interference factor, a variable component of interference which fluctuates from place to place and (if the common-wave transmitters are imperfectly synchronised) with time.

321. THE SERVICE AREA OF MEDIUM-POWER BROADCAST STATIONS [500–1500 kc/s].—Patrick. (*See* 603.) From South Africa.

322. CORRECTION TO "THE CALCULATION OF GROUND-WAVE FIELD INTENSITY OVER A FINITELY CONDUCTING SPHERICAL EARTH."—K. A. Norton. (*Proc. I.R.E.*, April 1942, Vol. 30, No. 4, p. 205.) *See* 1596 of 1942: the correction relates only to the graph paper used.

323. Temperature of the Atmosphere [and the Explanation of the Tropopause: Its Rapid Fall at Latitude showing Marked Increase in Cosmic-Ray Intensity, & a Suggested Production of Ozone by Cosmic Rays as the Reason].—T. G. Cowling: L. Jánossy. (*Nature*, 12th Dec. 1942, Vol. 150, pp. 686–687.) Note on a colloquium.

324. The Effect of Temperature on the Production of the Afterglow of Active Nitrogen [Experimental Investigation].—D. E. Debeau. (*Phys. Review*, 1st/15th Sept. 1942, Vol. 62, No. 5/6, p. 302: summary only.)

325. Standards on Radio Wave Propagation: Measuring Methods [for Radio Field Intensity: for Power radiated from an Aerial: for Noise-Field Intensity].—I.R.E. (*Proc. I.R.E.*, July 1942, Vol. 30, No. 7, Part II, 16 pp.)

326. Standards on Radio Wave Propagation: Definitions of Terms.—I.R.E. (*Proc. I.R.E.*, July 1942, Vol. 30, No. 7, Part III, 8 pp.)

327. The Frequency Spectrum [and the Present Confusion of Terminology: C.C.I.R. (1937) Table suggested as Satisfactory, perhaps with an Additional Line for Millimetric Waves].—R. L. Smith-Rose: Fleming-Williams. (*Wireless Engineer*, Aug. 1942, Vol. 19, No. 227, p. 360.) Prompted by Fleming-Williams's proposals, 2293 of 1942.

328. Classification of Frequencies: Need for a Universal System [Editorial on C.C.I.R. Wave-Band System & Fleming-Williams Frequency-Band System (2293 of 1942): Need for System applying both to Frequencies & Wavelengths].—(*Wireless World*, Dec. 1942, Vol. 48, No. 12, p. 277.)

329. Characteristic Impedance of Parallel Wires in Rectangular Troughs [deduced by Joint Use of Conformal Transformation & Method of Images: Cases of the Balanced Two-Wire Transmission Line & Balanced Three-Wire Line with Centre Lines of the Three Wires lying in a Plane, Inner Wire serving as Return for Outer Two.]—S. Frankel. (*Proc. I.R.E.*, April 1942, Vol. 30, No. 4, pp. 182–190.)

330. Skin Effect in Radially Traversed Plane Parallel Circular Plates [regarded as Idealised Case of (*e.g.*) an Earth Connection through a Chassis].—Wolff. (*See* 359.)

331. Formulas for the Skin Effect [in Wires, Transmission Lines, & Coils: for Shielding Effect of Sheet Metal: for Resistance due to Shield near a Coil: etc.].—H. A. Wheeler. (*Proc. I.R.E.*, Sept. 1942, Vol. 30, No. 9, pp. 412–424.) A summary was dealt with in 2298 of 1942.

332. Corrections to "Skin-Effect Formulas". J. R. Whinnery. (*Electronics*, March 1942, Vol. 15, No. 3, p. 94.) *See* 2619 of 1942.

333. The Solution of Maxwell's Equations for Certain Diffraction Problems [of Plane Waves normally incident upon an Aperture in a Plane Opaque Screen in an Isotropic Transparent Medium: the Concept of "Corrugated Waves," & Its Application].—R. A. Woodson. (*Phys. Review*, 1st/15th Sept. 1942, Vol. 62, No. 5/6, p. 297: summary only.)

"Perhaps the most valuable contribution of the corrugated wave theory is its simplicity by which it readily explains phenomena predicted by much more complicated theories."

334. The Influence of an "Uneven" Anisotropy on the Path of Light Rays.—P. Frank. (*Phys. Review*, 1st/15th Sept. 1942, Vol. 62, No. 5/6, pp. 241–243.)

"Examples of 'uneven' anisotropy are provided by the path of light rays through moving bodies or the path of electrons through an electromagnetic field. The formulae we are going to develop will embrace in one mathematical pattern all these problems."

335. The Diffusion from a Point Source: Supplement to the Paper "Some Diffusion Problems" [2292 of 1942].—W. Bothe. (*Zeitschr. f. Phys.*, 8th Sept. 1942, Vol. 119, No. 7/8, pp. 493–497.)

## ATMOSPHERICS AND ATMOSPHERIC ELECTRICITY

336. Cosmic Static [Radiation of Extraterrestrial Origin detectable by Radio Receiving Equipment: Full Description of Highly Directive Apparatus & Auxiliary Equipment: Circuit Analysis: Signal/Ripple Ratio: Thermodynamic Relations: Motor-Car-Ignition Noise: Results: Cosmic Static as the Equivalent of Thermal Agitation in which All Space is the Conductor].—G. Reber. (*Proc. I.R.E.*, Aug. 1942, Vol. 30, No. 8, pp. 367–378.)

This is the work referred to at the end of 3564 of 1942 (Fränz): for the earlier reports *see* 1769 of 1940, and *Astrophys. Journ.*, June 1940, Vol. 91, p. 621. "If this speculation has any foundation, then the intensity/frequency function of cosmic static should be the same as that of thermal-agitation noise; in other words, the intensity per unit frequency band-width should be constant, independent of frequency from any region of the sky": measurements are required at some other frequency several octaves from the present one of 160 Mc/s.

337. The Criterion for Streamer Formation [Experimental Indications that Density of Photoionisation is More Important than Tip Field in Streamer Propagation: but Meek's Criterion for Spark Breakdown still applies in Practice with *K* about 0.1 for Air].—G. L. Weissler & L. B. Loeb. (*Phys.

*Review*, 1st/15th Sept. 1942, Vol. 62, No. 5/6, p. 300: summary only.) *Cf.* Zeleny, 3203 of 1942.

## PROPERTIES OF CIRCUITS

338. The Effect of Fluctuation Voltages on the Linear Detector.—Ragazzini. (*See* 390.)

339. Discussion on "The Distribution of Amplitude with Time in Fluctuation Noise."—Norton: Landon. (*See* 389.)

340. Signal/Noise Ratio of Cathode-Follower [Calculated Comparison between Conventional Pentode Amplifier & Triode Cathode-Follower], *and* Criticism of This Letter.—D. A. Bell; R. E. Burgess. (*Wireless Engineer*, Aug. 1942, Vol. 19, No. 227, pp. 360–361: Oct. 1942, No. 229, p. 450.)

(i) "The signal/noise ratio of Fig. 2 [triode cathode-follower] is 6 or 8 db better than that of Fig. 1 [conventional pentode]": the gain would be less by the factors of 2 and $\sqrt{GZ}$, but the input conductance is reduced by a factor greater than 2, so that where the input conductance of the valve is a serious factor, the difference in gain is not so great as suggested: "the circuit of Fig. 2 may therefore be of interest where it is necessary to obtain the utmost signal/noise ratio." (ii) Various points in Bell's analysis are criticised, and finally: "Although the ratio of shot currents of a pentode and triode of the same slope may be of the order of 2 to 1 [basis of Bell's argument], the over-all improvement due to the use of a triode . . . instead of a pentode will be less than this owing to the thermal noise in the input circuit and in the output impedance. . . ."

341. The Use of Vacuum Tubes as Variable Impedance Elements [in Automatic Tuning, Frequency Modulation, etc: Use of Inverse-Feedback Amplifier to obtain Very Large Effective Capacitance or Very Low Negative Resistance, variable by varying the Amplifier Gain: etc.].—H. J. Reich. (*Proc. I.R.E.*, June 1942, Vol. 30, No. 6, pp. 288–293.)

"Since no general discussion of the entire subject appears to have been presented, this paper may be of some value in suggesting new applications of reactance-tube circuits."

342. Response of Reactive Networks to Frequency-Modulated Signals.—J. D. Weston. (*Wireless Engineer*, June 1942, Vol. 19, No. 225, pp. 251–253.)

"To summarise, we may say that, if the maximum frequency excursion is many times the highest modulation frequency, a reactive network will produce negligible distortion if it has a linear phase characteristic and an amplitude characteristic which is nearly level over the range $\omega \pm \lambda$, where $\lambda$ is the maximum frequency excursion. The use of 'staggered' tuned circuits to produce the required amplitude characteristic is precluded, since the phase shift is then non-linear."

343. Circuit for Neutralising Low-Frequency Regeneration and Power-Supply Hum [in High-Gain Multi-Stage Amplifiers].—Wen-Yuan Pan. (*See* 401.)

344. Differential-Input Circuits for Biological Amplifiers [Survey, including Papers occurring only in Medical Journals], *and* Discussion of the Above Paper.—M. G. Saunders: J. Debski. (*Electronic Eng:g*, May 1942, Vol. 14, No. 171, pp. 760 and 762–763: July 1942, Vol. 15, No. 173, p. 80.)

(i) "In the writer's opinion the circuit of Offner (1713 of 1937) should prove the most satisfactory. . . . In this country it has not, however, been customary to use push-pull throughout. . . . Consequently, one of the other circuits has been preferred, and that of Tönnies (2096 of 1938) is most commonly in use." For diagram corrections *see* June issue, No. 172, Vol. 15, p. 40. (ii) Two points of disagreement.

345. Judging an Amplifier by means of the Transient Characteristic [Derivation of Equations for Transient Characteristics of Various Valve-Coupling Networks, and Comparison between the Corresponding Curves: Question as to the Improvement, at High Frequencies, due to Elaboration of Network: etc.].—J. Haantjes. (*Electronics*, March 1942, Vol. 15, No. 3, pp. 80, 82: summary only.) *See* also 3212 of 1942.

346. Analysis, Synthesis, and Evaluation of the Transient Response of Television Apparatus.—Bedford & Fredendall. (*See* 478.)

347. A Contribution to the Theory of Network Synthesis [for Elimination of Trial-&-Error Design of Corrective Networks for Transient Conditions: Analysis yielding Generalised Infinite-Integral Formula (Eqn. 16) extending Brune's Procedure, based on Steady-State Requirements, to Transient Requirements: Application to finding Linear Passive Network equivalent to a Non-Linear Resistance such as Thyrite].—R. A. Whiteman. (*Proc. I.R.E.*, May 1942, Vol. 30, No. 5, pp. 244–246.)

348. Dividing Networks for Two-Way Horn Systems [Choice of Crossover Frequency: Comparison of M-Derived & Constant-Resistance Types of Network: Air-Core *versus* Iron-Core Coils: etc.].—Campbell. (*See* 456.)

349. L-Type Impedance-Transforming Circuits [for Aerial/Feeder Matching, etc.].—Smith. (*See* 424.)

350. Transformed Networks [for Economy in Material in Band-Pass-Filter Networks without Change in Qualities].—H. J. Griese. (*Wireless Engineer*, Oct. 1942, Vol. 19, No. 229, pp. 463–465.) Full translation of the German paper dealt with in 1936 of 1942.

351. Logarithmic Charts and Circuit Performance [Graphical Method of dealing with Circuit Problems where a Wide Range of Frequencies is involved: Advantages: Use in Teaching].—D. N. Truscott. (*Electronic Eng:g*, May 1942, Vol. 14, No. 171, pp. 744–748, & July 1942, Vol. 15, No. 173, pp. 57–60.)

352. " Electric Circuits " [Book Review].—E. E. Staff of M.I.T. (*Proc. I.R.E.*, May 1942, Vol. 30, No. 5, p. 257.)

353. " Electromechanical Transducers and Wave Filters " [Book Review].—W. P. Mason. (*Journ. Applied Phys.*, Oct. 1942, Vol. 13, No. 10, p. 644.)

354. Radio Bibliography : Filters, Equalisers, Networks : Acoustics, Loudspeakers [and List of Previous Subjects].—F. X. Rettenmeyer. (*Radio* [New York], Oct. 1942, No. 273, pp. 26 and 29..40.)

355. " Chart Atlas of Complex Hyperbolic and Circular Functions (Third Edition, 1924) " *and* " Tables of Complex Hyperbolic and Circular Functions (Second Edition, 1927) " [Book Reviews].—A. E. Kennelly. (*Proc. I.R.E.*, April 1942, Vol. 30, No. 4, p. 211.)

356. The Operative Attenuation of Four-Pole-Network Chains.—W. Klein. (*T.F.T.*, April 1942, Vol. 31, No. 4, pp. 103–108.)

This problem comes to the fore again (after its earlier treatment in connection with audio-frequency technique) owing to the development of wire broadcasting and carrier-current working along overhead lines. A circuit for wire broadcasting often consists of a combination of overhead lines and cable sections which are not matched to each other, and a practical way of calculating its attenuation is very necessary : hence the present theoretical investigation.

Eqn. 9 for the operative attenuation of such a network was already given by Hoecke, but its use is fairly complicated. However, what is often wanted is not the exact value of the attenuation at a given frequency, but rather the limits between which, in a given frequency range, it may under the most unfavourable conditions fluctuate as a result of imperfect matching. The writer therefore obtains from eqn. 9 the simplified eqn. 18 for the operative attenuation, which is here represented as the sum of the quadripole attenuations, the " junction " attenuations (*see* eqn. 9 and text below it), and the " reflection attenuation." The simplification consists largely of the fact that the junction attenuations contain only the characteristic impedances and terminal resistances of the quadripoles, and not their input impedances (so difficult to calculate). They are therefore easily derived, but eqn. 18 also contains the " reflection attenuation," more difficult to obtain : in many practical cases it can be neglected, but its calculation is given in section 2 and illustrated by an example (of a composite wire-broadcasting line) in section 3. With the help of this " reflection attenuation " a simple relation is obtained (eqn. 31) for the input impedance.

357. Characteristic Impedance of Parallel Wires in Rectangular Troughs.—Frankel. (*See* 329.)

358. The Influence of the Connecting Leads in the Measurement of High-Frequency Voltages.—Lang. (*See* 505.)

359. Skin Effect in Radially Traversed Plane Parallel Circular Plates.—H. H. Wolff. (*Arch. f. Elektrot.*, Aug. 1942, Vol. 36, No. 8, pp. 493–513.)

Previous calculations of skin effect have all dealt with cylindrical conductors, in which a practically axial flow of current is assumed, so that the current-density vector has, in the main, the same direction. In practice, however, a number of arrangements occur in which the current flow, starting from a particular point, passes in many directions, and which can be represented to a first approximation, as regards skin effect, by plane parallel circular plates in which the current-density vector is perpendicular to the axis. For example, the earthing of a.c. circuits frequently takes place through metal plates which serve also as the chassis of an apparatus.

The present work deals with the calculation of amplitude and phase conditions, effective resistance, penetration depth, etc., for an idealised version of such a case. Obviously the solution of the problem, thus idealised, will permit exact numerical calculations in practical conditions to be made only in extremely rare cases. On the other hand the application of the formulae now obtained will very often give values of the correct order of magnitude and lead to a satisfactory constructive solution of a given problem. Of special importance for high-frequency apparatus is the fact represented by eqn. 48 and Fig. 8, that for $kd/2 > 2.6$ (where $k = \sqrt{\mu\mu_0\omega \, . \, 10^{-8}/2\rho'}$ and $d =$ plate thickness), that is to say at high frequencies or great thicknesses, the total current $i$ lags by a practically constant amount (approx. $\pi/4$) behind the current density at the surface and thus behind the voltage between points on the surface : a fact which lies at the bottom of frequently encountered and hitherto unexplained phenomena. Also the results concerning the ratio of the effective a.c. resistance to the d.c. resistance, represented generally in eqn. 67 and for high frequencies and great thicknesses in eqn. 68 : the general equation 67 reduces, when $kd$ (*see* above) is not less than 3.7, to the simple statement that the ratio of the resistances is, within at most 5%, equal to $kd/2$. That is, for high $kd$ values the ratio increases in proportion to the thickness of the plate and to the root of the angular frequency (*see* Fig. 13) and in this region it is useless in practice to try to reduce the effective a.c. resistance by increasing the thickness of the plate.

360. Corrections to " Skin-Effect Formulas."—J. R. Whinnery. (*Electronics*, March 1942, Vol. 15, No. 3, p. 94.) *See* 2619 of 1942.

361. Formulas for the Skin Effect.—Wheeler. (*See* 331.)

362. The Calculation of Choke Coils with Several Windings.—J. Böhse & H. Bielert. (*T.F.T.*, May 1942, Vol. 31, No. 5, pp. 128–133.)

Economy in cost and weight is, of course, obtained by providing a single winding with a number of tappings : in this way ten or more inductance values can be provided by one core. Still greater economy, however, is obtained by the use of completely separate windings, which can be connected in all kinds of series and opposed combinations, provided that these windings are properly proportioned. It is shown here that the series of inductance values obtainable with a single core approaches very closely to the theoretical optimum if the various numbers of turns are in the ratio $1 : 3 : 3^2$ . . . But to avoid the serious departure from linearity over a wide range of inductances (Fig. 2) it is desirable to make use of only some of the possible combinations and to cover the whole range by two or more multi-winding chokes: thus for a $p_{max}$ ($= \Delta L/L$) of 10% a range from 1 mH to 2000 mH can be covered by five four-winding or three five-winding chokes: Fig. 5 shows the curve of a five-choke combination covering a range of 0.32–2000 mH in 114 steps.

363. CORRECTIONS TO "ON D.C.-POLARISED A.C. CHOKING COILS AND THEIR BACK COUPLING."—Th. Buchhold. (*Arch. f. Elektrot.*, Aug. 1942, Vol. 36, No. 8, p. 514.) *See* 3547 of 1942.

364. NEGATIVE AND POSITIVE RESISTANCE: SOURCES AND SINKS OF POWER [$I = f(V)$ for Any Two-Terminal Resistive Network: Sinks have Positive Slope, Positive Resistance, and $I$ & $V$ in Phase: Sources have Negative Slope, Negative Resistance, and $I$ & $V$ in Antiphase: Idealised Negative Resistance: Definition of Point Resistance & Incremental Resistance: Appropriate Circuits for Current- & Voltage-Controlled Negative Resistors: etc.].—D. M. Tombs. (*Wireless Engineer*, Aug. 1942, Vol. 19, No. 227, pp. 341–346.)

365. ON THE DEIONISATION IN THYRATRONS OPERATING AS RELAXATION OSCILLATORS.—Upatov. (*See* 551.)

366. DESIGNING A RESISTANCE–CAPACITY OSCILLATOR COVERING FREQUENCIES FROM 40 TO 13 000 C/S IN FOUR RANGES [replacing Beat-Frequency Oscillator: Choice of Number of Sections & Ranges for Phase Shifter: Design of Maintaining Amplifier: etc.].—R. C. Whitehead. (*Wireless World*, Dec. 1942, Vol. 48, No. 12, pp. 278–282.) For another paper on these oscillators *see* Tucker, 512, below.

367. SPARK-SUPPRESSION BY RECTIFIERS [for Preservation of Contacts & Prevention of Radio Interference: Theoretical Investigation, and a Vibrator Circuit giving a Good Square-Wave Voltage].—H. J. Schmidt. (*E.N.T.*, Aug. 1942, Vol. 19, No. 8, pp. 156–160.)

This possible use of dry-plate rectifiers has been dealt with in several patents during the last fifteen years, but the plan has hardly ever come into practice. The usual method of spark-suppression, by a condenser in series with a resistance, quenches the spark at break the better as the condenser is made larger and still more as the resistance is made smaller: at the cost, however, of increasing the spark at make and the tendency of the contacts to stick. The condenser–resistance circuit must therefore be a compromise: but better results can be obtained by making the resistance different in the two directions of the current, by making it, in fact, in the form of a rectifier (Fig. 2): the condenser is not always necessary in that case (Fig. 3), the rectifier alone shunting the working inductance. But this simplified arrangement is not suitable for a supply voltage over about 10 V, and a condenser with rectifier in series is then needed across the contacts themselves (Fig. 4). Each of these circuits has its advantages and disadvantages: even Fig. 4, for instance, has the defect that a buzzer or vibrator on this scheme generates more harmonics than one using the circuit of Fig. 3.

The present paper concentrates on the simple scheme of this Fig. 3, and it is shown that the occurrence of high "extra" voltages in a buzzer can be prevented in this way, but that a subsidiary action of the circuit occurs when, as would happen in such a buzzer (Fig. 6), the contact is made and broken several times within a time-period of the quenching circuit. This unwanted effect is the increasing of the working current taken by the buzzer, and is the more marked the lower the resistance of the rectifier. However, by a suitable choice of this rectifier resistance the surge voltages at break can be suppressed satisfactorily without the current consumption being seriously increased, and a buzzer on these lines will provide a good square-wave voltage form.

## TRANSMISSION

368. THE RETARDING-FIELD OSCILLATOR [Experimental Investigation, using Pair of Matched A.T.40 Valves in Parallel: Comparison of Results with Those of Hollmann and Gerber: Simultaneous Existence of $\alpha$ (Triode-Type) and $\beta$ (Diode-Type) Oscillations, $\alpha$ practically Constant in Wavelength as Circuit Length is increased, $\beta$ decreasing Slightly, but Uniformly: after Critical Position (Transition from B–K to G–M, with Large Increase in Oscillation Amplitude) both $\alpha$ and $\beta$ (still detectable) have Wavelengths governed by External Circuit].—Wm. Alexander. (*Wireless Engineer*, April 1942, Vol. 19, No. 223, pp. 143–147.)

369. VELOCITY-MODULATING GRIDS: AN INVESTIGATION OF THEIR ACTION BY MEANS OF ANALYSIS AHD GRAPHICAL METHODS [Influence of Finite Grid-Distance on Velocity & Density Distributions of Emerging Electrons].—R. Kompfner. (*Wireless Engineer*, April 1942, Vol. 19, No. 223, pp. 158–161.) For previous work *see* 1355 of 1942.

370. ON THE EFFECT OF AN EXTERNAL ELECTROMOTIVE FORCE ON THE OPERATION OF A SPLIT-ANODE MAGNETRON.—S. Ya. Braude. (*Journ. of Tech. Phys.* [in Russian], No. 23/24, Vol. 10, 1940, pp. 1993–2010.)

This is the first part of a paper dealing with the operation of a magnetron with a two-segment anode when in addition to the anode voltage an external e.m.f. is applied to the anode segments (Fig. 2). Equation 1 of the system is written down and its operation is analysed, using the Poincaré-Lyapunov methods. It is shown that if the static characteristic of the magnetron can be represented by an equation of the 3rd degree, and an external e.m.f. is applied to the anode segments, then, according to the average slope of the characteristic and the equivalent impedance of the tuned circuit, three different cases may arise. The first case corresponds to the state of excitation and either pulling-in or beating of frequencies may occur; the amplification of the system is the highest in this case. In the second and third cases there is no excitation of the system and only "forced" oscillations can take place; the amplification of the system is lower than in the first case. For each case formulae are derived for determining the amplitudes of both "forced" and "free" oscillations. The effects of the magnetic field, anode voltage, and emission current on the band-width for pulling-in in the first case are also determined.

371. A New Direct Crystal-Controlled Oscillator for Ultra-Short-Wave Frequencies [using a Mechanical Harmonic (*e.g.* 15th) of an *AT*- or *BT*-Cut Quartz Crystal].—W. P. Mason & I. E. Fair. (*Proc. I.R.E.*, Oct. 1942, Vol. 30, No. 10, pp. 464–472.) In the particular model described, the oscillator was used with two 240H amplifying pentodes to deliver 12 watts to the aerial at 120 Mc/s.

372. The Use of Vacuum Tubes as Variable Impedance Elements [in Frequency Modulation, etc.].—Reich. (*See* 341.)

373. Spectra and Non-Linear-Distortion Factors of Frequency- and Amplitude-Modulated Oscillations: III—Modulation Distortion in Special Cases ["Klirr" Factors for Demodulation at a Loss-Free Self-Inductance: Approximation for Modulating Frequencies Small compared with Carrier Frequency: Modulation Distortion in Multi-Channel Working (F.M: unlike A.M., Combination Tones & Crosstalk may be produced by Too Narrow Filters): Single-Sideband Working].—M. Kulp. (*E.N.T.*, July 1942, Vol. 19, No. 7, pp. 126–135.) For Parts I & II *see* 41 of January.

374. The Zero-Beat Method of Frequency Discrimination [for Frequency-Modulated Transmitters: Defects of Previous Methods: Method characterised by Compactness (suitable for Portable & Semiportable Transmitters), Symmetry of Output with respect to Positive & Negative Frequency Deviation, and Linearity].—C. F. Sheaffer. (*Proc. I.R.E.*, Aug. 1942, Vol. 30, No. 8, pp. 365–367.) U.S. Patent 2 274 434 of 1942.

375. A Modern 10 Kilowatt Frequency-Modulation Transmitter [at W69PH (WCAU), Philadelphia: using the Crosby Stabilising Circuit: capable of less than 1½% Over-All Distortion between 30 & 15 000 c/s at 100 kc/s Deviation].—E. S. Winlund & C. S. Perry. (*Electronics*, March 1942, Vol. 15, No. 3, pp. 40–43.) From R.C.A.

376. A New Frequency-Modulation Broadcasting Transmitter [embodying the "Grounded-Plate" Amplifier, 3232 of 1942].—A. A. Skene & N. C. Olmstead. (*Proc. I.R.E.*, July 1942, Vol. 30, No. 7, pp. 330–335.)

377. A Crystal-Controlled Frequency-Modulation Exciter: a Narrow-Band System for Amateur Use.—W. P. Bollinger. (*QST*, Oct. 1942, Vol. 26, No. 10, pp. 25–27.)

378. Experimental Polyphase Broadcasting [WHO's Experimental Transmitter].—Loyet. (*See* 601.)

379. Class C Telegraphy: the Graphical Determination of Optimum Operating Conditions for Transmitting Valves.—Prinz & Mitchell. (*See* 437.)

380. A Combination Radiotelegraph–Broadcast High-Frequency Transmitter with Many New Features [1 or 30 kW on Telegraphy, 9 kW on Broadcasting: 5.7–22 Mc/s: Phase Modulation for Modulated C.W: Lange Neutralising Circuit for Power Amplifier: etc.].—F. D. Webster & R. E. Downing. (*Communications*, June 1942, Vol. 22, No. 6, pp. 10–13 and 27, 28.) At Lima, Peru.

381. Operating a C.W. Transmitter, with Four Stages all working on Different Voltages, from a Single 1500 Volt Supply [Inefficient Dropping Resistors avoided by Series Connection of Two Stages].—J. Blitch. (*QST*, Oct. 1942, Vol. 26, No. 10, p. 72.)

## RECEPTION

382. Sound Receiver for Television Programmes.—E. Kinne. (*Funktech. Monatshefte*, April 1942, No. 4, pp. 54–56.)

Special attention is given to methods of stabilising the oscillator frequency against temperature changes, such as the use of a Calit and a Condensa F condenser in parallel for the condenser in the tuned circuit $C_1L_3$, or the construction of the iron-cored coil $L_3$ in such a way that warming-up of the iron core produces effects which are opposed to the effects due to the warming-up of the winding. This latter method gives a cheap oscillator whose constancy is such that after a four-minutes' run it can be used as an adaptor for an ordinary broadcast receiver. Similar methods would also be needed for a "straight" receiver.

The sensitivity of the oscillator system to voltage fluctuations is then considered. Quite small over-modulation of the output valve, only too easily produced, will cause serious voltage fluctuations, and consequent oscillator-frequency fluctuations, when the smoothing choke is formed by the loud-speaker-magnet winding of (say) 1500 ohms. One way of avoiding this effect is to make the smoothing choke very low in resistance and to over-dimension the mains transformer and rectifiers. Another way, apparently adopted in the receiver here described, is to decouple the oscillator circuit by a suitable resistance $R_2$ and to connect it in front of the smoothing choke. For hum suppression, each of the heater terminals of the oscillator valve is earthed directly through a non-inductive condenser, $C_2$ and $C_3$. Properly built, the receiver shown is unaffected by mains voltage fluctuations of 10%. If larger fluctuations occur, the anode voltage must be stabilised, at any rate for the oscillator valve.

383. FREQUENCY MODULATION: PART VII, *and* RESPONSE OF REACTIVE NETWORKS TO FREQUENCY-MODULATED SIGNALS.—Sturley: Weston. (*See* 596 & 342.)

384. TRANSIENTS IN FREQUENCY MODULATION [Question of Receiving-Filter Band-Width treated by Consideration of the Transients which will develop in the Receiver upon a Sudden Change in Frequency: of Particular Importance for Certain F.M. Applications such as Television Synchronising Signals: Importance of Amplitude Limiter & Symmetrical Arrangement of Filter Pass-Band: etc.].—H. Salinger. (*Proc. I.R.E.*, Aug. 1942, Vol. 30, No. 8, pp. 378–383.)

385. EFFECT OF AN EXTERNAL ELECTROMOTIVE FORCE ON THE OPERATION OF A MAGNETRON.—Braude. (*See* 370, above.)

386. NOTES ON RECEIVER DESIGN [for Short & Ultra-Short Waves: Suggestions for Improvement of Sensitivity, Stability, Signal/Noise Ratio, & Ease of Control: including the Full Utilisation of the Advantages of the Type 1851 Valve in First R.F. Stage by Use of Push-Pull Tuned R.F. Circuit].—D. Eby, Jr. (*Radio* [New York], Oct. 1942, No. 273, pp. 22 and 40, 41.)

387. COSMIC STATIC [Signal/Ripple Ratio: Thermodynamic Relations: Motor-Car-Ignition Noise: Cosmic Static as the Equivalent of Thermal Agitation in which All Space is the Conductor].—Reber. (*See* 336.)

388. SIGNAL/NOISE RATIO OF CATHODE-FOLLOWER, *and* CRITICISM OF ABOVE LETTER.—Bell: Burgess. (*See* 340.)

389. DISCUSSION ON "THE DISTRIBUTION OF AMPLITUDE WITH TIME IN FLUCTUATION NOISE."—K. A. Norton: V. D. Landon. (*Proc. I.R.E.*, Sept. 1942, Vol. 30, No. 9, pp. 425–429.)

Argument about Landon's results (2129 of 1941). Landon concludes by remarking that "the discussion has brought out important material not previously published. . . . Certain of the details . . . should find important practical applications."

390. THE EFFECT OF FLUCTUATION VOLTAGES ON THE LINEAR DETECTOR [Analysis yielding Expressions for A.F. Noise Spectra & R.M.S. Values of A.F. Noise Outputs in Various Conditions: Demodulation & Modulation-Compression Effects, and Harmonic Distortion, due to Fluctuation Voltages: A.F. Signal/Noise Ratio for Simultaneous Fluctuation Voltages & Modulated Carrier: Experimental Confirmation].—J. R. Ragazzini. (*Proc. I.R.E.*, June 1942, Vol. 30, No. 6, pp. 277–288.)

391. THE NOISE OF DIODES AND DETECTORS [Pyrite, Carborundum] IN THE STATIC AND DYNAMIC RÉGIMES.—H. F. Mataré. (*E.N.T.*, July 1942, Vol. 19, No. 7, pp. 111–126.)

Of the various measurements made to check the theoretical foundations of h.f. noise in saturated and unsaturated valves, all those obtained with triodes proved to be in good agreement with theory, but those obtained with unsaturated diodes differed seriously from the theoretical results: this applied particularly to the values found experimentally for valves with "guarding cones," specially designed for noise measurements (references "6" & "9").

Author's summary:—"The paper deals first with the difficulties encountered in static noise measurements, particularly the separation of the noise sources in the case of small internal resistances of diodes, and consequent small noise-voltages at the input circuit. A noise-factor $p$ is defined which is helpful in obtaining a picture of the noise-behaviour of the valves [eqn. 3 shows how $p$ relates the valve noise to the resistance noise at room temperature: eqn. 7 shows the simple relation between $p$ and the conveniently measured "weakening factor" $F$]. A general behaviour as regards the increase of $p$ with increasing effective voltage was established for diodes [of very different types: Fig. 7, where all the noise-factors $p$, plotted against $U_a$, show more or less the same rising slope: *see* text on p. 115, where also the special behaviour of small (decimetric-wave) triodes is discussed]. The differences in behaviour between diodes and triodes were measured for very different valve types; the explanation is sought in the already known manner by splitting the anode current into a saturated component and an unsaturated component.

"A theory of mixing noise is developed. The noise sources appearing in the heterodyning process are considered separately, starting on the one side from the oscillator and on the other from the mixing stage. The heterodyning noise was measured both in the parallel and push-pull connections [of the two halves of a duo-diode], whereby it was shown that the oscillator generates a considerable noise within the intermediate frequency [Fig. 15: the oscillator wavelength was 7 m, that of the i.f. circuit 150 m]. The oscillator noise can, however, be eliminated by the use of the push-pull connection. A simple calculation [eqns. 29 & 30] shows that the static noise of a valve in the self-excited régime is increased thereby by a factor from 8–10.

"Finally, the noise of [crystal] detectors is investigated. It is found that the detector in the static case and with small biasing voltages and currents reaches the limiting value $p=1$; that is to say it behaves, as regards noise, as its internal resistance at room temperature. Thus for small amplitudes the detector may be superior to the diode as regards noise."

392. THE DIODE AS MIXING VALVE FOR DECIMETRIC WAVES [Theory of Mixing Process in Diodes: Description of Push-Pull Mixer-Diode with Built-In Triode Oscillator System (Min. Fundamental 36 cm) for Reception of Wavelengths down to 9 cm: the Special

Circuit (with $\lambda/4$ Lecher Lines as Input & Oscillator Circuits for Waves below 1 m): Special Plating of Contact Pins: etc.].—M. J. O. Strutt & A. van der Ziel. (*E.T.Z.*, 24th Sept. 1942, Vol. 63, No. 37/38, pp. 450–451: summary, from *Philips tech. Rundschau*, Vol. 6, 1941, p. 289 onwards.) Referred to in 391, above.

393. Departures from Ohm's Law in Solids [and the Existence in Crystalline & Isotropic Solids of a Volume Effect obeying Neumann's Hypothesis, with Resistance varying as a Power Series in Components of Applied Electrical Field: Experiments on Parallel Crystal Plates of Carborundum, Zincite, & Galena].—H. Osterberg. (*Phys. Review*, 1st/15th Sept. 1942, Vol. 62, No. 5/6, p. 297: summary only.)

394. Diode Detection and Selectivity.—O. Tüxen. (*Funktech. Monatshefte*, April 1942, No. 4, pp. 48–54.)

For previous work *see* 2398 of 1941 and 3243 of 1942. The present paper deals with:—equivalent circuits: selectivity of the input circuit, neglecting the non-linear distortions: influence of the non-linear distortions on the selectivity: the threshold of the non-linear distortions: calculation of the largest carrier ratio permissible without reduction of selectivity: conclusions for practical design purposes (including the rule, representing a compromise between the requirements of high amplification, high selectivity, not-too-sharp cutting of the high note-frequencies, and low distortion, that the time constants of the h.f. and l.f. circuits should be about equal: $2RC = r'c' = rc$, where for instance $r = 0.5–1$ megohm, $c = 50–100$ pF).

395. An Evaluation of Radio-Noise-Meter Performance in Terms of Listening Experience [Statistical Investigation on 3 Types of Meter and 3 Types of Noise (Motor, Razor, & D.C. Relay)].—C. M. Burrill. (*Proc. I.R.E.*, Oct. 1942, Vol. 30, No. 10, pp. 473–478.)

A summary was referred to in 1356 of 1941. The need for more definite specification of the transient characteristics of the a.v.c. circuit is further discussed in the paper referred to in 2406 of 1942: otherwise the Joint Coordination Committee's specification is found to be adequate.

396. Spark-Suppression by Dry-Plate Rectifiers [for Preservation of Contacts & Prevention of Interference].—Schmidt. (*See* 367.)

397. Vibrator Wave-Forms [and Their Importance for the Life of the Vibrator].—S. L. Robinson. (*Electronic Eng:g*, June 1942, Vol. 15, No. 172, pp. 32–33.) Following on 1492 of 1942.

398. The Use of Vacuum Tubes as Variable Impedance Elements [in Automatic Tuning, etc.].—Reich. (*See* 341.)

399. Receiver with "Crack-Killing" Circuit and Automatic Tuning Correction.—H. Pitsch. (*Funktech. Monatshefte*, April 1942, No. 4, pp. 59–60.) Telefunken patent, D.R.P. 703 347.

400. A Circuit for Automatic Fading Control [Advantages of Diode over Triode, except that It gives Smaller D.C. Potentials: Method of amplifying These by the Subsequent Triode required for A.F. Amplification].—W. Stoff. (*Funktech. Monatshefte*, April 1942, No. 4, p. 60.) D. S. Loewe Company's patent, D.R.P. 696 188.

401. Circuit for Neutralising Low-Frequency Regeneration and Power-Supply Hum [in High-Gain Multi-Stage Amplifiers: Usual *RC* Decoupling Filters decrease in Effectiveness at the Particularly Troublesome Lowest Frequencies: a Simple Bridge Balancing Circuit effective at All Frequencies].—Wen-Yuan Pan. (*Proc. I.R.E.*, Sept. 1942, Vol. 30, No. 9, pp. 411–412.)

402. Shunt Condenser Aerial Coupling [Valuable Properties: Use in Communication-Type Receivers].—Amos. (*See* 422.)

403. Hearing, the Determining Factor for High-Fidelity Transmission [Volume & Frequency Ranges for Good Fidelity for Music & Speech: Stereophonic System improved by Three Channels: etc.].—Fletcher. (*See* 451.)

404. A Broadcast Receiver which will Not Re-Radiate [Nothing detectable beyond Distance of 25 Feet: primarily for Shipboard Use].—(*Sci. News Letter*, 31st Oct. 1942, Vol. 42, No. 18, p. 288: paragraph only.)

405. Correspondence on "Receivers for the Tropics" [Experiences in Cuba: Corrosion as a Purely Electrochemical Process, and a Suggested Solution of the Problem of Its Prevention].—H. C. Schwalm: W. E. Stewart. (*Electronics*, March 1942, Vol. 15, No. 3, pp. 126–128.) *See* 91 of 1942.

406. Manufacture and Sale of Radio Receivers [Broadcast & Others: New Prohibition of Materials].—(*Génie Civil*, 15th Oct. 1942, Vol. 119, No. 23, p. 288.)

407. Superheterodyne Receiver Tracking [Editorial on Recent Papers].—G.W.O.H. (*Wireless Engineer*, April 1942, Vol. 19, No. 223, pp. 141–142.) The Payne-Scott & Green paper particularly discussed has since been reproduced in *Wireless Engineer*: *see* 408, below.

408. Superheterodyne Tracking Charts: I & II.—A. L. Green: Ruby Payne-Scott & A. L. Green. (*Wireless Engineer*, June 1942, Vol. 19, No. 225, pp. 243–250: July 1942, No. 226, pp. 290–302.) Reprints of papers dealt with in 87 & 1669 of 1942.

409. Short-Wave Spread Bands in Automobile and Home Receivers [Recent Increased Importance: Discussion of Gain & Selectivity Characteristics of Simple Signal-Frequency Circuits giving Good Performance with Short Aerials, and Oscillator Circuits employing Few Switching Elements].—D. E. Foster & G. Mountjoy. (*Proc. I.R.E.*, May 1942, Vol. 30, No. 5, pp. 222–227.) A summary was referred to in 1036 of 1942.

410. Remote Adjustment of Broadcast Receivers [Flexible Cable giving Three Separate Adjustments, by Direct Twist, by Central Bowden Wire, & by Hydraulic Action of a Liquid contained in the Space between Wire & Sheath].—H. Munsch. (*Funktech. Monatshefte*, April 1942, No. 4, p. 60.) A Blaupunkt patent, D.R.P. 705 421.

411. 1941 Radio Audience Statistics [C.A.B. Ratings of Programmes: Highest Ratings of Previous Years about Doubled by President's Speech: 20 Leading Evening Programmes: etc.].—A. W. Lehman. (*Electronics*, March 1942, Vol. 15, No. 3, pp. 84, 86: summary only.)

412. "Radio Reception in Theory and Practice" [Book Review].—V. K. Saksena. (*Proc. I.R.E.*, May 1942, Vol. 30, No. 5, p. 258.) "A very satisfactory introductory book."

## AERIALS AND AERIAL SYSTEMS

413. Ultra-High-Frequency Antennas [Illustrated Account of "Zepp", Coaxial, "J", "Q" (with Cage Transformer), "Delta", & Other Types, and Their Various Advantages].—C. R. Stoll. (*Radio* [New York], Oct. 1942, No. 273, pp. 9–11.) "For mobile use the Marconi type appears to give slightly better results than the usual half-wave", provided it is worked against a well-conducting metal surface.

414. Prevention of Ice-Formation on Aerial of Frequency-Modulation Station W51R [Calrod Heaters built into Cross Arms & automatically switched on in Sleet-Forming Range of 28°–32°F].—(*Gen. Elec. Review*, Sept. 1942, Vol. 45, No. 9, p. 537: paragraph only.)

415. Aircraft Antennas [for Frequencies 2–20 Mc/s: the Icing Problem, & the Influence of Wire Angle relative to Wind Stream: the Use of Shunt-Fed Wing Aerials: Impedance Characteristics of Transverse, Longitudinal, & Shunt-Fed Aerials: Dummy Aerials for Testing: Trailing Aerials & Their Reel Mechanisms: Radiation Test Towers: etc.].—G. L. Haller. (*Proc. I.R.E.*, Aug. 1942, Vol. 30, No. 8, pp. 357–362.)

416. Data Sheets XXXII & XXXIII: the Inductance of Single-Layer Solenoids on Square and Rectangular Formers [*e.g.* for Frame Aerials].—(*Electronic Eng:g*, July 1942, Vol. 15, No. 173, pp. 65–68.)

417. Theoretical Investigations of Radiation Diagrams and Radiation Resistance for Progressive Waves of Various Phase Velocities.—W. Jachnow. (*E.N.T.*, Aug. 1942, Vol. 19, No. 8, pp. 147–155.)

In certain cable and aerial problems account has had to be taken of the difference between the velocities of the waves along the wire and in the surrounding medium, but even then the necessary correction has only concerned the phase or wave-length constant; any influence that a difference in velocities may have on the radiation diagram or the radiation resistance has hitherto been neglected. In the case of the Beverage aerial, where such a difference is brought about artificially, the theoretical problem (as regards a receiving aerial, and using cable theory) was dealt with in 1923 by Beverage, Rice, & Kellog, by Busch, and by Béthenod, with results which did not agree among themselves. The present writer therefore gives a more rigorous theoretical treatment of the effect of different velocities, along the wire and in the medium, on the radiation diagram and the radiation resistance: for the sake of definite conditions of current distribution, only a transmitting aerial is considered.

From the known formula (eqn. 1) for $\mathfrak{H}$ (distant field) based on a light-velocity $c$ both along the wire and in the surrounding medium, the writer arrives at eqn. 2 based on the light-velocity $c$ in the medium and a velocity $v$ along the wire: in this and the following equations, an asterisk represents an "along-the-wire" value, so that, for instance, $\lambda^*/\lambda = v/c = k$, where $k$ may be less than, equal to, or greater than unity. The equation becomes greatly simplified (to eqns. 12 or 13, according to whether $n$ is even or odd, where $l = n \cdot \lambda^*/2$) when the length $l$ of the radiator is a complete multiple of the half-wavelength. Finally, when the wire carries purely progressive waves of the form $i(x, t) = I_0 \sin\omega\,(t - x/v)$, the two alternative equations 17a & b are obtained for $\mathfrak{H}$ (the one involving $\lambda$ and $c/v$, the other $\lambda^*$ and $v/c$). From this relation the radiation diagrams in free space are calculated for a progressive wave of half the velocity of light, for values of $n$ from 1 to 6, and shown in Fig. 2: as is obvious from the formula, radiation disappears in the direction of the wire ($\delta = 0$ or $\pi$), and also in the vertical plane through its mid-point ($\delta = \pi/2$) when $n$ is even, whereas when $n$ is odd this vertical plane shows a maximum field-strength. Comparison of Fig. 2 with Fig. 4 (which represents the ordinary case where $c$ and $v$ are assumed equal) shows that the usual simple relation, "number of half-wavelengths in the wire equals half the number of lobes in the complete diagram," no longer applies: what is clear from Fig. 2 is that the amplitude of the main lobe varies very little for different wave-numbers $l/\lambda^*$ but that its angle to the wire axis fluctuates between 40° and 90° over the wave-number range from ½ to 3.

Fig. 3 shows the similarly calculated diagram for the case when $v = 2c$: here the inclination of the principal lobe remains practically constant at 60° as the wave-number increases from $\frac{1}{2}$ to 3, but its amplitude increases rapidly (to about twice its value at $\frac{1}{2}$), the lobe being increasingly sharply concentrated: the subsidiary lobes are insignificant throughout. For various values of $c/v$, 0, $\frac{1}{2}$, 1, 1.1, 1.5, and 2, envelope curves derived from $\sin\delta/(\cos\delta - c/v)$, one of the two factors into which eqn. 17 can be divided, are given in Fig. 5, and provide a guide to the changing form of the radiation diagram as the wave-number is changed: thus for $c/v=2$ the envelope is extremely flat, indicating that the principal-lobe amplitude hardly varies for different wave-numbers, as has been seen already.

The rest of the paper deals similarly with radiation resistance, for which eqn. 27a is derived for $v<c$, and eqn. 27b for $v>c$: by putting $v=c$ in this, the known formula (eqn. 26) for radiation resistance is obtained. From these formulae the curves of Fig. 9 are calculated, showing for $v/c=\frac{1}{2}$, 1, and 2 the variation of the radiation resistance with increasing wave-number $l/\lambda^*$, while for the particular case of $l/\lambda^*=2$ the variation of radiation resistance as a function of $v/c$ is seen in Fig. 10. This curve shows particularly well the striking sensitivity of the radiation resistance towards small deviations of the wire velocity from the space velocity: a sensitivity which has already been seen to apply also to the radiation diagram.

418. THE INCLINED RHOMBIC ANTENNA [for Reducing the Effect of Fading: Inclination improves Response at both Low & High Angles of Wave-Incidence: Derivation of Equations for determining Inclination to yield Desired Response Patterns].—C. W. Harrison, Jr. (*Proc. I.R.E.*, May 1942, Vol. 30, No. 5, pp. 241–244.)

419. HORIZONTAL-POLAR-PATTERN TRACER FOR DIRECTIONAL BROADCAST ANTENNAS [Derivation of Equations for Relative Field Strength in Horizontal Plane of 3-Element Directional Arrays for Non-Symmetrical Patterns: Mechanical Tracer automatically plotting Horizontal Pattern for 2- or 3-Element Arrays].—F. A. Everest & W. S. Pritchett. (*Proc. I.R.E.*, May 1942, Vol. 30, No. 5, pp. 227–232.) A summary was referred to in 1054 of 1942.

420. A MECHANICAL CALCULATOR FOR DIRECTIONAL ANTENNA PATTERNS [for 2- or 3-Element Arrays: for Horizontal Patterns, and Vertical Patterns when Aerial Heights are Equal: Accuracy well within That of Commercial Field-Strength-Measuring Equipment].—W. G. Hutton & R. M. Pierce. (*Proc. I.R.E.*, May 1942, Vol. 30, No. 5, pp. 233–237.) A summary was referred to in 1054 of 1942.

421. CHARTS FOR THE DETERMINATION OF THE ROOT-MEAN-SQUARE VALUE OF THE HORIZONTAL RADIATION PATTERN OF TWO-ELEMENT BROADCAST ANTENNA ARRAYS [and the Special Properties of an Array with 138° Spacing].—K. Spangenberg. (*Proc. I.R.E.*, May 1942, Vol. 30, No. 5, pp. 237–240.)

422. SHUNT CONDENSER AERIAL COUPLING: AN ANALYSIS OF THE CIRCUIT [Further Notes on Circuit described by Sturley (3334 of 1941): Valuable Properties: Use in Communication-Type Receivers, for Medium & Long Waves].—S. W. Amos. (*Wireless Engineer*, Dec. 1942, Vol. 19, No. 231, pp. 549–554.)

423. THE SELF-IMPEDANCE OF A SYMMETRICAL ANTENNA [Extension of Hallén's Analysis (2763 of 1939) by using his Formula for Input Self-Impedance to determine Input Resistance, Reactance, Impedance-Magnitude, & Phase Angle for Useful Range of Values of $h/\lambda$ & $a/\lambda$: by deriving an Expression for Max. Input Resistance: and by obtaining Expressions for the Resonant & Antiresonant Lengths].—R. King & F. G. Blake, Jr. (*Proc. I.R.E.*, July 1942, Vol. 30, No. 7, pp. 335–349.)

424. L-TYPE IMPEDANCE-TRANSFORMING CIRCUITS [for Aerial/Feeder Matching, etc: the Eight Possible Combinations & Their Properties: Design Curves: Illustrative Problems].—P. H. Smith. (*Electronics*, March 1942, Vol. 15, No. 3, pp. 48–52 and 54, 125.)

425. IMPEDANCE-MEASURING INSTRUMENT [primarily for Matching Measurements on Directional Aerial Systems, where Current Distribution must Not be Disturbed during the Measurement: using Method based on Measurement (by Valve-Voltmeter of Very High Impedance) of Voltages created by Insertion of Physically Small Low-Resistance Series-Resonant Circuit].—C. E. Smith. (*Proc. I.R.E.*, Aug. 1942, Vol. 30, No. 8, pp. 362–364.)

426. HIGH-FREQUENCY ARTICULATED AIR-SPACED CABLES AND VALVE CONNECTOR [" Co-ax " Cables & (*e.g.*) Valve-Grid Connectors].—Telequipment Company. (*Journ. of Scient. Instr.*, Dec. 1942, Vol. 19, No. 12, p. 188.). *See* also 3580 of 1942.

427. GROUND SYSTEM AT KMPC [North Hollywood, Directional Aerial System].—(*Radio* [New York], Oct. 1942, No. 273, pp. 16–17 and 38, 39.)

## VALVES AND THERMIONICS

428. ON THE EFFECT OF AN EXTERNAL ELECTROMOTIVE FORCE ON THE OPERATION OF A SPLIT-ANODE MAGNETRON.—Braude. (*See* 370.)

429. VELOCITY-MODULATING GRIDS: AN INVESTIGATION OF THEIR ACTION BY MEANS OF ANALYSIS AND GRAPHICAL METHODS.—Kompfner. (*See* 369.)

430. THE DIODE AS MIXING VALVE FOR DECIMETRIC WAVES.—Strutt & van der Ziel. (*See* 392.)

431. THE NOISE OF DIODES AND [Crystal] DETECTORS IN THE STATIC AND DYNAMIC RÉGIMES.—Mataré. (*See* 391.)

432. TRIODES WITH SQUARE-MESH GRIDS: CALCULATING THE AMPLIFICATION FACTOR [Grave Analytical Difficulties even with Simplifying Assumptions: Empirical Result (based on Electrostatic-Screening-Effect Measurements) that Square-Mesh Grid is Equivalent to Grating of Parallel Circular Wires with Pitch 0.6 Times the Pitch of the Mesh].—C. C. Eaglesfield. (*Wireless Engineer*, Oct. 1942, Vol. 19, No. 229, pp. 447–450.) From the Mullard laboratories. *See* also 433, below.

433. THE EQUIVALENCE OF PARALLEL-WIRE AND SQUARE-MESH GRIDS.—G. W. O. H: Eaglesfield. (*Wireless Engineer*, Oct. 1942, Vol. 19, No. 229, pp. 443–446.) Editorial on 432, above, giving an analytical mode of attack (already used in 1915 for the calculation of aerial capacitance: *see* footnote "1") leading to the same result as Eaglesfield's experimental method.

434. SPECIALLY STEEP-SLOPE TRIODE WITH CENTRAL FLAT-PLATE ANODE IN SAME PLANE AS, AND BETWEEN, TWO PARTS OF FILAMENT, ALL ENCLOSED IN OVAL CONTROL ELECTRODE.—K. Zwirner. (*Funktech. Monatshefte*, May 1942, No. 5, p. 76.) Tobis Patent, D.R.P. 667 682.

435. THE CHARACTERISTIC CURVES OF THE TRIODE [Simple Scheme by which Complete Static Curves of a Power Triode can be derived by Extrapolation from only Three Curves (Grid-Current, Plate-Current, Total-Current), obtained by Ordinary D.C. Methods at Low Power without Danger of Overheating: and a Simplified Method of Presentation of the Entire Set of Curves, using a New Log-Log Chart].—E. L. Chaffee. (*Proc. I.R.E.*, Aug. 1942, Vol. 30, No. 8, pp. 383–395.)

436. TRACING VALVE CHARACTERISTICS, USING THE CATHODE-RAY OSCILLOGRAPH.—G. Bocking. (*Wireless Engineer*, Dec. 1942, Vol. 19, No. 231, pp. 556–563.)

"The tracing of the characteristics of valves of all classes [and not merely the high-dissipation types operating under positive-grid conditions, for which various c.r.o. techniques have been developed in order to avoid damage to the valves] by means of the c.r.o. now offers such an increase in convenience and so little difficulty of construction and loss of accuracy that suitable apparatus should be a standard piece of equipment in every laboratory." The writer deals both with the "continuous-trace" and "scan" (stroboscopic) methods, and includes the description of an equipment on the former principle, giving $E_a/I_a$ and $E_g/I_a$ curves for all types of valve up to an anode dissipation of 15 watts.

437. CLASS C TELEGRAPHY: THE GRAPHICAL DETERMINATION OF OPTIMUM OPERATING CONDITIONS FOR TRANSMITTING VALVES [Part I—Rules & Graphs: Part II—Theoretical Derivation of the Formulae & Curves].—D. G. Prinz & R. G. Mitchell. (*Wireless Engineer*, Sept. 1942, Vol. 19, No. 228, pp. 401–407.) From the M.O. Valve Company's research staff.

438. THIRTEEN WAYS TO PROLONG [Transmitting] TUBE LIFE.—Heintz & Kaufman, Ltd. (*Radio* [New York], Oct. 1942, No. 273, pp. 15–16.)

439. WATER AND FORCED-AIR COOLING OF VACUUM TUBES: NON-ELECTRONIC PROBLEMS IN ELECTRONIC TUBES [Attempt to give a Common Basis for analysing & comparing Practical Results obtained by Various Experimenters: Better Behaviour of Air-Cooled Valves as regards Arc-Back: Other Advantages & Disadvantages: Possibilities of Aluminium Air-Coolers: etc.].—I. E. Mouromtseff. (*Proc. I.R.E.*, April 1942, Vol. 30, No. 4, pp. 190–205.)

440. "FUNDAMENTALS OF VACUUM TUBES: SECOND EDITION" [Book Review].—A. V. Eastman. (*Proc. I.R.E.*, May 1942, Vol. 30, No. 5, pp. 257–258.)

441. ARMY–NAVY PREFERRED LIST OF VACUUM TUBES [Official Notice].—(*Radio* [New York], Oct. 1942, No. 273, pp. 11 and 38.)

442. THE USE OF SECONDARY ELECTRON EMISSION TO OBTAIN TRIGGER OR RELAY ACTION [similar to That of a Thyratron but capable of Much Higher Operating Speeds].—A. M. Skellett. (*Phys. Review*, 1st/15th Sept. 1942, Vol. 62, No. 5/6, p. 296: summary only.)

443. RANGES OF SECONDARY ELECTRONS IN MAGNESIUM [by Measurements, with High-Energy Primary Electrons, of S.E. Ratio $\delta$ as Function of Layer Thickness: with a Section on Electron Ranges in High-Yield Surfaces and the Effect of Oxygen in producing Good Secondary Emitters].—R. Truell. (*Phys. Review*, 1st/15th Oct. 1942, Vol. 62, No. 7/8, pp. 340–348.)

444. PHYSICAL FOUNDATIONS OF RADIO: IV—COMPOSITE CATHODES, PHOTOSENSITIVE SURFACES, COLD EMISSION [including Thin-Film Field Emission (Malter Effect)].—M. Johnson. (*Wireless World*, Dec. 1942, Vol. 48, No. 12, pp. 291–293.)

445. ON THE DISTRIBUTION OF VELOCITIES IN FIELD-EMISSION ELECTRONS.—G. Richter. (*Zeitschr. f. Phys.*, 8th Sept. 1942, Vol. 119, No. 7/8, pp. 406–414.)

The electrons extracted from cold metals in a vacuum by high fields possess, like thermionic electrons, no single velocity either in magnitude or direction. The velocity distributions normal and tangential to the emitting surface both have their influence on, for instance, the determination of the best possible resolving power of a field-emission electron microscope.

Author's summary:—"The velocity distributions normal and tangential to the emitting surface are derived from field-emission theory. The half-value width of the energy distribution in the former case is found to be about three times as great as in the latter, and for practically attainable field-strengths is of the order of a few tenths of an electron-volt. Measurements so far available, on the other hand, show a distribution width about ten times greater. This great expansion of width compared to the theoretical value is probably to be attributed to errors due to field distortions in the experimental arrangement," the various possible distortions being dealt with in turn.

446. Contacts between Metals and between a Metal and a Semiconductor [Quantitative Development of the Kunming Researches (2501 of 1942): Classical Treatment with help of Results of Wave-Mechanical Theory of Electron Energy States in Solids: Qualitative Treatment of Body in Vacuum and of Two Bodies separated by a Gap].—H. Y. Fan. (*Phys. Review*, 1st/15th Oct. 1942, Vol. 62, No. 7/8, pp. 388–394.)

447. Rate of Crystal Growth in Drawn Tungsten Wires as a Function of Temperature [observed by way of the Thermionic Emission Pattern].—C. S. Robinson, Jr. (*Journ. Applied Phys.*, Oct. 1942, Vol. 13, No. 10, pp. 647–651.)

## DIRECTIONAL WIRELESS

448. Iron-Cored D.F. Loop [from German Aircraft: with Photograph].—(*Wireless Engineer*, Aug. 1942, Vol. 19, No. 227, p. 350.) Giving an increase of 10 db over its air-cored equivalent: *cf.* 2389 of 1942.

449. German Aircraft Radio [Illustrated, from Official Reports: including Emergency Equipment].—(*Electronic Eng:g*, June 1942, Vol. 15, No. 172, pp. 8–12 and 36.)

450. Aircraft Radio Communications [Two-Way Communications, *En Route* Navigation, Instrument Landing: Aerials: Table of Weights: etc.].—H. K. Morgan. (*Journ. Roy. Aeron. Soc.*, Dec. 1942, Vol. 46, No. 384, Abstracts p. 483.)

## ACOUSTICS AND AUDIO-FREQUENCIES

451. Hearing, the Determining Factor for High-Fidelity Transmission [Substantially Complete Fidelity for Orchestral Music given by Volume Range of 65 db & Frequency Range 60–8000 c/s: for Speech, 40 db & 100–7000 c/s: Stereophonic System (2-Channel) up to 5000 c/s preferred to Single-Channel up to 15 000 c/s: Improvement in Stereophonic System by using 3 Channels].—H. Fletcher. (*Proc. I.R.E.*, June 1942, Vol. 30, No. 6, pp. 266–277.)

452. The Technical Basis of Sound Reproduction [and the Work of the Apparatus Approval Committee of the Central Council for School Broadcasting].—L. E. C. Hughes. (*Nature*, 28th Nov. 1942, Vol. 150, pp. 629–630.) Summary of Brit.I.R.E. paper.

453. Problems of Contrast Compression [General Considerations leading to the "Ideal" Compression Characteristic: etc.].—Hildebrand. (*See* 604.)

454. A Frequency-Modulated Reproduction Control for Sound Films [Volume Expansion at Desired Points by F.M. Control Track on Movietone Prints].—J. G. Frayne & F. P. Herrnfeld. (*Electronics*, March 1942, Vol. 15, No. 3, pp. 88 and 92, 94.)

455. Radio Bibliography: Filters, Equalisers, Networks: Acoustics, Loudspeakers [and List of Previous Subjects].—F. X. Rettenmeyer. (*Radio* [New York], Oct. 1942, No. 273, pp. 26 and 29..40.)

456. Dividing Networks for Two-Way Horn Systems [Complementary Loudspeakers covering Modern Wide Range of Frequencies: Choice of Crossover Frequency: Comparison of *M*-Derived & Constant-Resistance Types of Network: Design Data (and Air-Core *versus* Iron-Core Coils: Satisfactory Solution of Iron-Core Difficulties): Intermodulation Test for Distortion: etc.].—C. A. Campbell. (*Communications*, June 1942, Vol. 22, No. 6, pp. 14, 16, 21, 29, and 34.)

457. "Electromechanical Transducers and Wave Filters" [Book Review].—W. P. Mason. (*Journ. Applied Physics.*, Oct. 1942, Vol. 13, No. 10, p. 644.)

458. Recording and Reproducing Standards [N.A.B. Committee's "Technical Standards and Good Engineering Practices" for Electrical Transcriptions & Recordings for Radio Broadcasting].—L. C. Smeby. (*Proc. I.R.E.*, Aug. 1942, Vol. 30, No. 8, pp. 355–356.)

459. Embossed Groove Recording [Simplicity of Operation (No Chip to be removed: No Change of Stylus): Inexpensive Records which may be Filed or Mailed: Poorer Fidelity & Smaller Volume Range than Cutting Process, but Eminently Suitable for Office Dictating Machines, etc.].—L. Thompson. (*Electronics*, March 1942, Vol. 15, No. 3, pp. 30–33 and 76, 78.) From the Sound-Scriber Corporation.

460. "Fun with a Recorder: How to make Good Recordings" [including "30 Most Intriguing Audio-Scripts": Book Review].—Audio Devices. (*Electronics*, March 1942, Vol. 15, No. 3, p. 56.)

461. How Recordings are Made: No. 4—Playback.—C. B. De Soto. (*QST*, Oct. 1942, Vol. 26, No. 10, pp. 54–59 and 114..118.)

462. The Jensen Concert Needle [for Long Life, Fidelity, and Reduced Scratch & Wear: Bent & Flattened End with Wear-Resisting Tip].—P. L. Jensen. (*Radio* [New York], Oct. 1942, No. 273, pp. 24 and 41.)

463. Corrections to "Super-Cardioid Directional Microphone."—B. B. Bauer. (*Electronics*, March 1942, Vol. 15, No. 3, p. 94.) *See* 2696 of 1942.

464. Discussion on "Distortion Tests by the Intermodulation Method" [particularly the Success of the Method in Researches on Electronic Musical Instruments].—B. F. Miessner: Hilliard. (*Proc. I.R.E.*, Sept. 1942, Vol. 30, No. 9, p. 429.) For Hilliard's paper *see* 1709 of 1942.

465. Harmonic Analysis of Waves containing Odd and Even Harmonics [Extension of Author's 1920 Method to include Even Harmonics & D.C. Component: Measurements of Amplitudes at 30° Intervals: Tables].—P. Kemp. (*Electronic Eng:g*, June 1942, Vol. 15, No. 172, pp. 13–18.)

466. An Evaluation of Radio-Noise-Meter Performance in Terms of Listening Experience.—Burrill. (*See* 395.)

467. The Hewlett-Packard Resistance-Tuned Oscillator [Four Models, including Model 200D, 7 c/s to 70 kc/s].—Leland Instruments. (*Electronic Eng:g*, June 1942, Vol. 15, No. 172, p. 38.)

468. Designing a Resistance–Capacity Oscillator covering Frequencies from 40 to 13 000 c/s in Four Ranges.—Whitehead. (*See* 366.)

469. A Note on *R–C* Oscillators [Supplementary Remarks on Bacon's Paper].—Tucker. (*See* 512.)

470. The Vibration Characteristics of Nearly Complete, "Free-Free" Circular Rings [of Piston-Ring Shape].—F. P. Bundy & C. W. Bankes. (*Journ. Applied Phys.*, Oct. 1942, Vol. 13, No. 10, pp. 652–662.)

471. The Reduction of Noise in Rooms, Diminution of Its Interfering Action in Telephony, and Improvement of Audibility in Rooms [General Theoretical Foundations: Measurements on Absorbent Materials ("Akustikplatte," "Ravolitplatte," etc.): Practical Measures taken by German Post Office: etc.].—K. Braun & P. Just. (*T.F.T.*, April 1942, Vol. 31, No. 4, pp. 91–103.)

472. Advances in Acoustical Treatment at New N.B.C. Studios.—(*Electronics*, March 1942, Vol. 15, No. 3, pp. 34–35.)

473. The Optics of Supersonic Waves, and Supersonic Images [Problem of Vision through Opaque Media: Further Development of Pohlman's Technique (196 of 1940) by Use of Quartz-Plate Receiver & Valve Amplification: Some Experimental Results: Possible Applications].—O. Barbier. (*Alta Frequenza*, Aug./Sept. 1942, Vol. 11, No. 8/9, pp. 383–396.)

## PHOTOTELEGRAPHY AND TELEVISION

474. Errors in the Deflection of a Cathode-Ray Beam by Deflecting Systems of Single Symmetry [and the Possibility of Their Employment in correcting Distortion in Television Pick-Up Tubes & Projecting Tubes with Slanting Screens].—G. Wendt. (*Zeitschr. f. Phys.*, 8th Sept. 1942, Vol. 119, No. 7/8, pp. 423–462.)

The commercial cathode-ray tube generally employs for its ray deflection a deflecting field which is symmetrical to two planes passing through the tube axis and perpendicular to each other (Fig. 1): these are fields with "double symmetry." For certain practical purposes, however, fields are used which have only one plane of symmetry: this plane may lie either in the deflecting direction of the field (Fig. 2) or at right angles to it (Fig. 6). It is such deflecting fields of "single symmetry" that are examined here. It is first pointed out that electrostatic deflecting fields may be of single symmetry even though the shape and disposition of the plates are of double symmetry: it is only necessary, for this to occur, for the deflecting potentials applied to them to be "asymmetrical" in the sense of cathode-ray-tube technique.

The properties of an ideal deflection, namely a deflection-increase strictly proportional to the deflecting potentials or currents and a spot which is undistorted throughout the deflection, cannot be maintained over large deflections: "deflection errors" occur. These have already been calculated and discussed for doubly-symmetrical fields (*see* Picht & Himpan, 3091 of 1941, and the references to Wendt and Glaser there given). The calculation for singly-symmetrical fields is now given in Part A, for the case where the plane of symmetry lies in the deflection plane and for the case where it is at right angles to it. Both electrostatic and magnetic fields are considered. The first approximation of this calculation shows the errors as "second-order errors," in contrast to doubly-symmetrical fields, where "third-order errors" appear: the two types of second-order deflection-errors are astigmatism and raster-distortion ("Verzeichnung"), an example of the latter being trapezoidal distortion.

This trapezoidal distortion has useful possibilities. In storage-type television pick-up tubes and in "Aufsichtsröhren" (c.r. tubes for television projection, having the c.r. axis oblique to the fluorescent screen) the desired rectangular scanning raster produced by doubly-symmetrical deflecting organs becomes a trapezoid. Correction of this effect by modification of the saw-tooth deflecting voltages involves a considerable expenditure in apparatus, and it would be very desirable to carry out the correction, instead, by utilising the trapezoidal second-order distortion given by a singly-symmetrical deflecting system. This problem is considered on pp. 443–455, the two possible arrangements being seen in Fig. 10.

In cathode-ray oscillographs, also, the second-order errors play an important rôle, because it is often desirable, for circuit simplification, to connect one plate of each deflecting system to the last anode. This turns the deflecting fields into fields of single symmetry, so that second-order errors appear. On pp. 456–461 the question is examined of how far it is possible to eliminate these distortions by giving the plates a corresponding singly-symmetrical form.

475. THE RELATIVE SENSITIVITIES OF TELEVISION PICK-UP TUBES, PHOTOGRAPHIC FILM, AND THE HUMAN EYE [" True " & " Operating " Sensitivities for All Three : Comparison with an Ideal Picture-Reproducing Device].—A. Rose. (*Proc. I.R.E.*, June 1942, Vol. 30, No. 6, pp. 293–300.) A summary was referred to in 1109 of 1942.

476. THE ELECTROPLANE CAMERA [for Great Depth of Field].—Smith-Dieterich Corporation. (*See* 668.)

477. TRANSIENTS IN FREQUENCY MODULATION [of Particular Importance for F.M. Applications such as Television Synchronising Signals).—Salinger. (*See* 384.)

478. ANALYSIS, SYNTHESIS, AND EVALUATION OF THE TRANSIENT RESPONSE OF TELEVISION APPARATUS.—A. V. Bedford & G. L. Fredendall. (*Proc. I.R.E.*, Oct. 1942, Vol. 30, No. 10, pp. 440–457.)

A summary was referred to in 1104 of 1942. " For the purpose of simplifying the passage between sine-wave response and transient response [response to a Heaviside unit voltage], and interpreting the latter, we present below : (1) a graphical-chart method for analysing the response of a system to a square-wave input signal to obtain the sine-wave phase and amplitude characteristics ; (2) a graphical-chart method for synthesising the response to a square wave from the sine-wave phase and amplitude characteristics ; (3) a method for evaluating the mean steepness of a transient-response wave in terms of the *width of blur* produced in a television image by a wave which is similar in its visual effect and which has a *linear* change from one level to another ; and (4) suggestions for the supplementing of sine-wave measurements by transient measurements " [*see* 479, below].

479. A PORTABLE HIGH-FREQUENCY SQUARE-WAVE OSCILLOGRAPH FOR TELEVISION [by which a Square-Wave (100 kc/s) Response may be Viewed as a Dotted Wave and Recorded as a Series of Readings : Synchronous Sweep & Timing Dots derived from Square-Wave Response of Apparatus under Test].—R. D. Kell, A. V. Bedford, & H. N. Kozanowski. (*Proc. I.R.E.*, Oct. 1942, Vol. 30, No. 10, pp. 458–464.) Primarily for use in connection with 478, above.

480. TELEVISION WAVE-FORMS : AN ANALYSIS OF SAW-TOOTH AND RECTANGULAR WAVE-FORMS ENCOUNTERED IN TELEVISION AND CATHODE-RAY-TUBE PRACTICE, *and* DATA SHEETS XXIX–XXXI.—C. E. Lockhart. (*Electronic Eng:g*, June 1942, Vol. 15, No. 172, pp. 19–22 : pp. 22–26.)

481. COLOUR TELEVISION : PART I [Requirements : Five Alternative Systems, and Reasons for Columbia Broadcasting System's Choice : General Theory of Colour Television (Colour, Flicker, Electrical Characteristics) : Studio Equipment : Receiving Equipment : Patents & Literature References].—P. C. Goldmark, J. N. Dyer, E. R. Piore, & J. M. Hollywood. (*Proc. I.R.E.*, April 1942, Vol. 30, No. 4, pp. 162–182.) *See* also 804 of 1941 and back reference.

482. TRANSMITTING EQUIPMENT FOR TELEVISION REPORTING : II (CONTD.).—H. Weber. (*Funktech. Monatshefte*, April 1942, No. 4, Supp. pp. 13–19.) For other parts *see* 2412 & 3312 of 1942, and August issue, No. 8, Supp. pp. 25–28.

483. TELEVISION : AN AGENCY FOR PREPAREDNESS [New York A.R.P. Training Service, etc.] : WITH A SUMMARY OF SIX MONTHS OF COMMERCIAL OPERATION [and Some Results of Questionnaires to Owners of Receivers : Advertising Rates : etc.].—N. E. Kersta. (*Electronics*, March 1942, Vol. 15, No. 3, pp. 26–29 and 116..124.) *See* also 3635 of 1942.

484. SOUND RECEIVER FOR TELEVISION PROGRAMMES.—Kinne. (*See* 382.)

485. AUTOMATIC RECORDING : THE RAPID DRAWING OF LIGHT-SENSITIVE-CELL CHARACTERISTIC CURVES ACCOMPLISHED AUTOMATICALLY BY THE PHOTOELECTRIC RECORDER.—H. T. Wrobel. (*Gen. Elec. Review*, Oct. 1942, Vol. 45, No. 10, pp. 585–587.) For the high-speed recorder used *see* Clark, 3456 of 1942.

486. STANDARDS ON FACSIMILE : DEFINITIONS OF TERMS.—I.R.E. (*Proc. I.R.E.*, July 1942, Vol. 30, No. 7, Part IV, 6 pp.)

487. SOLUTIONS FOR ELECTROCHEMICAL RECORDING [Twelve Thiosulphate Solutions (based on Five Stock Solutions) with Varying Characteristics as to Tone Gradation, Resistance, etc.].—Ch. P. Fagan. (*Journ. of Scient. Instr.*, Dec. 1942, Vol. 19, No. 12, pp. 184–185.)

## MEASUREMENTS AND STANDARDS

488. REMINISCENCES OF AN INSTRUMENT MAKER.—R. S. Whipple. (*Journ. of Scient. Instr.*, Dec. 1942, Vol. 19, No. 12, pp. 178–183.)

489. WAVELENGTH MEASUREMENTS OF DECIMETRIC, CENTIMETRIC, AND MILLIMETRIC WAVES.—A. G. Clavier. (*Elec. Communication*, No. 4, Vol. 20, 1942, pp. 295–304.) An article prepared before September 1939 but not published.

490. A FREQUENCY-MODULATION STATION MONITOR—Summerhayes. (*See* 595.)

491. THE ZERO-BEAT METHOD OF FREQUENCY DISCRIMINATION [for Frequency-Modulated Transmitters].—Sheaffer. (*See* 374.)

492. A NEW DIRECT CRYSTAL-CONTROLLED OSCILLATOR FOR ULTRA-SHORT-WAVE FREQUENCIES.—Mason & Fair. (*See* 371.)

493. A SECONDARY FREQUENCY STANDARD USING REGENERATIVE FREQUENCY-DIVIDING CIRCUITS [5 Mc/s, 1 Mc/s, 500 kc/s, & 100 kc/s, all from 5 Mc/s Crystal Oscillator (continu-

ously monitored, if Max. Accuracy is desired, against WWV's 5 Mc/s Transmission): Advantages of Frequency Division over Multiplication: etc.].—F. R. Stansel. (*Proc. I.R.E.*, April 1942, Vol. 30, No. 4, pp. 157–162.)

494. A DETERMINATION OF THE ELASTIC CONSTANTS OF BETA-QUARTZ [New Technique for Quick Location & Measurement of Weak Resonance Points of a Piezoelectric Plate: Resonator driven by Frequency-Modulated Carrier, C.R.O. Recording: Approximate Theory of Quartz-Resonator Response to a F.M. Input: etc.].—E. W. Kammer & J. V. Atanasoff. (*Phys. Review*, 1st/15th Oct. 1942, Vol. 62, No. 7/8, pp. 395–400.)

495. BALL ANVIL FOR USE ON THE SCHERR COMPARITOL FOR INSPECTION AND MEASUREMENT OF EXTREMELY THIN WORK [Crystals, Shims, etc.].—George Scherr Company. (*Radio* [New York], Oct. 1942, No. 273, p. 24.)

496. DIRECT FREQUENCY MEASUREMENTS: FEATURING AN AUXILIARY OSCILLATOR OF UNIQUE DESIGN FOR INTERPOLATION [for Carrier-Frequency Measurements within F.C.C. Tolerance: also as Driver for R.F. Bridges, Lining-Up of Tuned Circuits, etc.].—D. Langham. (*Communications*, June 1942, Vol. 22, No. 6, pp. 5–7.)

497. MEASUREMENT OF FREQUENCIES BELOW 15 KILOCYCLES PER SECOND [Survey of Methods, for Frequencies down to 0.2 c/s: including Robinson's Direct-Reading Meter based on Wien Bridge].—J. E. Thwaites & F. J. M. Laver. (*Electronic Eng:g*, May 1942, Vol. 14, No. 171, pp. 767 and 770.) Abstract of part of an I.E.E. paper. For a correction to a diagram *see* June issue, p. 40.

498. THE VIBRATION CHARACTERISTICS OF NEARLY COMPLETE, "FREE-FREE" CIRCULAR RINGS [of Piston-Ring Shape].—F. P. Bundy & C. W. Bankes. (*Journ. Applied Phys.*, Oct. 1942, Vol. 13, No. 10, pp. 652–662.)

499. INSTRUMENTS: TEST AND MEASURING GEAR AND ITS USES: VIII—MULTIVIBRATORS.—W. H. Cazaly. (*Wireless World*, Dec. 1942, Vol. 48, No. 12, pp. 294–297.) Corrections to Part VI (3742 of 1942) are given on p. 297.

500. DATA SHEETS XXXII & XXXIII: THE INDUCTANCE OF SINGLE-LAYER SOLENOIDS ON SQUARE AND RECTANGULAR FORMERS [*e.g.* for Frame Aerials].—(*Electronic Eng:g*, July 1942, Vol. 15, No. 173, pp. 65–68.)

501. IMPEDANCE-MEASURING INSTRUMENT [primarily for Matching Measurements on Directional Aerial Systems].—Smith. (*See* 425.)

502. ON THE DIRECT MEASUREMENT OF THE TOTAL TRANSMISSION EQUIVALENT (ATTENUATION EQUIVALENT) AND PHASE CONSTANT OF LINEAR SYMMETRICAL QUADRIPOLES [*e.g.* Wide-Band Cables] WHEN THERE IS UNDER-MATCHING AT THE OUTPUT TERMINALS.—H. Langer. (*T.F.T.*, April 1942, Vol. 31, No. 4, pp. 115–119.)

The indirect method of determining the attenuation equivalent $b$, from the open-circuit and short-circuit input impedances, becomes inaccurate for the higher values of $b$ (in practice when $b >$ 1 neper). On the other hand the direct method of determination, from measurement of the voltage or current transformation, loses its accuracy at low values of attenuation: eqn. 15 gives errors of nearly 1 to 4% for attenuation equivalents from 1.0 to 0.01 neper, for a 2% matching error. "It is here shown how, starting with a deliberate under-matching, the simultaneous measurement of the input voltage and of the input and output currents will give a good practical determination of the attenuation equivalent and phase constant of a wide-band cable in the region of low attenuation . . . The accuracy will depend on the error limits of the thermoelectric current and voltage meter employed and on the care taken in the carrying out of the measurement. . ."

503. A PORTABLE HIGH-FREQUENCY ANALYSER [*e.g.* for Carrier & Sidebands: on Grützmacher Exploring-Note Principle].—H. Brauner. (*T.F.T.*, April 1942, Vol. 31, No. 4, pp. 109–112.)

Made up of two cases, one containing the exploring-note generator, with its clockwork-driven condenser and low-pass filter, the other the output amplifier: each contains its own mains unit. An additional external recorder is employed. The filter is made of *RC* sections and has a pass band of about 90 c/s which can be varied (and the flank steepness also) by a potentiometer-controlled back-coupling: the result is an excellent filter weighing far less and occupying much less space than an *LC* network. At the input of the output amplifier a switch enables a 1 : 1 or a 3 : 1 sensitivity to be selected: often the carrier amplitude is much larger than that of the sidebands and it is desirable, for greater accuracy, to use the 1 : 1 position for the carrier and then to change over (without affecting the course of the test) to the 3 : 1 position.

504. DISCUSSION ON "DISTORTION TESTS BY THE INTERMODULATION METHOD."—Miessner: Hilliard. (*See* 464.)

505. THE INFLUENCE OF THE CONNECTING LEADS IN THE MEASUREMENT OF HIGH-FREQUENCY VOLTAGES.—J. G. Lang. (*Funktech. Monatshefte*, April 1942, No. 4, pp. 45–47.)

Analysis leading to approximate formulae and general conclusions. To keep down the error due to the leads the main thing to do is to make the length very short compared with the wavelength. A ratio $l/\lambda$ of 0.01–0.02 gives a "negligible" error of 1–2%, so that for 10 Mc/s the leads may be 30–40 cm long. So long as $\lambda$ is greater than $100l$, the leads should be made as low-capacity as possible (thin wire, large spacing between the leads and from earth) so as to keep down the loading of the measuring point, which consists mainly of the sum of the lead capacitance $C$ and that of the input circuit of the voltmeter, $K$. When, especially at frequencies above 30 Mc/s, it is no longer possible to make the lead-length small enough, efforts must be made to keep down the ratio $K/C$. One way of doing this is to make $C$ large, but this raises the loading of the measuring point and gives rise to errors. A better way is to make $K$ small, by using a low-capacity diode, keeping the connection to it as short and free from capacitance as possible, and transferring the charging or blocking condenser to the cathode lead to earth, where it can be made large enough to extend the range of the voltmeter into the l.f. region. The leak resistance also must be as low-capacity as possible, and be mounted close to the valve. A compromise between the contradictory aims of keeping the frequency error due to $K/C$ small by making $C$ greater than $K$, and of keeping down the measuring-point loading, is reached for the cases when $l/\lambda > 0.01$–$0.02$ by choosing the leads so that $C=K$. Formulae are given for the capacitances of a single-pole connector, a concentric lead, and a two-wire lead with spacing much greater than the wire-diameter, so that from this $C=K$ relation the required lead-length can be calculated.

Fig. 3 shows the percentage measuring error $p$ [$100(\mathfrak{U}_2-\mathfrak{U}_1)/\mathfrak{U}_1$ %] as a function of $l/\lambda$ up to 0.12 for seven values of $K/C$ from 0 to 1.5, the scales being so chosen that the error curves take the form of straight lines through zero. Finally a simple equivalent diagram for the leads, as regards the voltage ratio $\mathfrak{U}_1/\mathfrak{U}_2$ only, is discussed (Fig. 4): this represents the leads as a circuit made up of their whole inductance and the half-capacitance effective at the voltmeter terminals: the input capacitance of the voltmeter comes in parallel to the $C/2$. The circuit holds good so long as $l/\lambda$ is not greater than 0.1, but only for the calculation of the voltage relations: it would give wrong results in calculating the loading resistance as offered to the object under test: this point is discussed in the final column.

506. An Electronic Potentiometer [a Degenerative, Slide-Back, Triode Voltmeter combining Advantages of both D.C. Slide-Back & Degenerative Voltmeters: Great Accuracy (to Four Significant Figures) over Whole Range 0–100 Volts: usable also as D.C. Ammeter (with Shunts) and as R.F. Voltmeter & Ammeter (with Diode Voltmeter)].—M. A. Honnell. (*Proc. I.R.E.*, Oct. 1942, Vol. 30, No. 10, pp. 433–436.)

507. Standards on Radio Wave Propagation: Measuring Methods.—I.R.E. (*See* 325.)

508. A Universal Valve Voltmeter with Linear Scale [Cathode-Follower, as an Extreme Form of Negative Feedback, particularly Valuable in production of Highly Stable Valve Voltmeter, less Sensitive but more nearly Linear than Usual Types: Unsuitability of Latter for D.C: a Modified Cathode-Follower Circuit specially suitable for D.C.].—W. F. Lovering. (*Phil. Mag.*, Nov. 1942, Vol. 33, No. 226, pp. 844–846.)

509. A New Large Test Desk for the D.C. Compensation Method [originally applied to Electricity-Meter Checking, but lately used for Precision Measurements on Thermo-elements, Millivoltmeters, etc: a Test Desk based on Experience, and embodying a New Precision D.C. Compensator.]—W. Zschaage. (*E.T.Z.*, 8th Oct. 1942, Vol. 63, No. 39/40, pp. 458–461.)

510. The Neon-Tube Parts Checker: a Simple Means of measuring Resistors, Condensers, and Voltage without a Meter.—W. E. Bradley. (*QST*, Oct. 1942, Vol. 26, No. 10, pp. 18–20 and 90, 92.)

511. Humidity Testing of Components.—P. R. Coursey. (*Wireless Engineer*, June 1942, Vol. 19, No. 225, pp. 255–256.) Reply to Westcombe's criticism, 2446 of 1942.

512. A Note on *R–C* Oscillators [Supplementary Remarks on Bacon's Paper, 2769 of 1942].—D. G. Tucker: Bacon. (*Electronic Eng:g*, May 1942, Vol. 14, No. 171, p. 752.) *See* also Lenihan, p. 772, on Ginzton & Hollingsworth's earlier description. For another paper on these oscillators *see* Whitehead, 366, above.

513. Miniature Instruments [with Scale Lengths from 1.5 to 3.5 Inches: Recent Improvements—Frequency Range to over 150 Mc/s, Construction to stand Shock & Vibration, "Expanded-Scale" Design: etc.].—J. M. Whittenton. (*Gen. Elec. Review*, Sept. 1942, Vol. 45, No. 9, pp. 501–504.) For glass jewels *see* p. 539.

514. Magnetic Materials [and the Improved Design & Performance of A.C. & D.C. Indicating Instruments due to the New Alloys].—A. J. Corson. (*Gen. Elec. Review*, Oct. 1942, Vol. 45, No. 10, pp. 573–575.)

515. The Calculation of Experimental Errors: an Effective Grouping and Analysis of Observational and Experimental Errors—D. Espy. (*Communications*, June 1942, Vol. 22, No. 6, pp. 8–9.)

516. Standards and Standardisation [Editorial, including Proposed "Orders of Precision," whose Numbers become the Logarithm of the Reciprocal of the Tolerance to which the Component is adjusted or can be measured].—(*Wireless Engineer*, Aug. 1942, Vol. 19, No. 227, pp. 339–340.)

517. Standardisation as applied to Industrial Electrical Instruments [Dimensional & Other Limits: Limits of Performance].—K. Edgcumbe. (*Engineering*, 4th Dec. 1942, Vol. 154, p. 452: summary of I.E.E. paper.) For another summary, including the Discussion, see *Electrician*, 11th Dec. 1942, pp. 639–641.

518. Note on Dimensions [Criticism of Guggenheim's Paper].—W. Wilson. (*Phil. Mag.*, Nov. 1942, Vol. 33, No. 226, pp. 842–844.) *See* 3070 of 1942.

519. A Differential Electronic Stabiliser for Alternating Voltages, and Some Applications [Unit designed primarily for Supply for the Testing of A.C. Instruments].—A. Glynne. (*Electrician*, 27th Nov. 1942, Vol. 129, pp. 586–588: summary of I.E.E. paper and Discussion.)

520. Theory and Limitations of the Universal A.C. Bridge.—J. H. Ellison: A. K. McLaren. (*Radio* [New York], Oct. 1942, No. 273, pp. 12–13 and 44.)

521. Studies in Superconductivity: II—Evaporated Lead Films: III—Sn, Cb, Ta, and Pb Wires [Experiments in connection with the Development of a Superconducting Bolometer (3682 of 1942) for Small Amounts of Infra-Red Radiation].—W. F. Brucksch, Jr., W. T. Ziegler, J. W. Hickman. (*Phys. Review*, 1st/15th Oct. 1942, Vol. 62, No. 7/8, pp. 348–353 & 354–356.)

## SUBSIDIARY APPARATUS AND MATERIALS

522. Errors in the Deflection of a Cathode-Ray Beam by Deflecting Systems of Single Symmetry [and Their Utilisation to correct Other Distortions].—Wendt. (*See* 474.)

523. The Influence of an "Uneven" Anisotropy on the Path of Light Rays [or of Electrons].—Frank. (*See* 334.)

524. Correction to Formula in "Errors in Photography of Cathode-Ray-Tube Traces: the Effects of Screen Curvature" [2773 of 1942].—Moss & Cattanes. (*Electronic Eng:g*, May 1942, Vol. 14, No. 171, p. 772.)

525. Image-Field Curvature ["Saucer" Distortion] with Electrostatic Lenses [Experimental Point-by-Point Measurements, with Special C.R. Tube, on Three-Electrode Lenses: Dependence on Focal Length, Lens Dimensions, & Object Distance: Independence of Stop Size (which influences the Accompanying Astigmatism): Distortion 5–7 Times greater than with Optical Lenses].—Gobrecht. (*Arch. f. Elektrot.*, Aug. 1942, Vol. 36, No. 8, pp. 484–492.)

526. Electron Lenses [Part of Lecture on Electron Optics, to Electronics Group of Institute of Physics].—Gabor. (*Nature*, 5th Dec. 1942, Vol. 150, pp. 650–652.)

"In principle, the spherical aberration could be corrected by suitably controlled auxiliary streams of ions or electrons," a line of experiment bristling with difficulties. But regarding the use of the space charge of the ray electrons themselves, "a perfectly stigmatic beam carrying a finite current density is not in principle optically impossible, as the electrostatic energy stored in it is finite," and this method of compensation may unwittingly have been used in c.r. tubes. It would be useless in electron microscopes. Rebsch's hint for the near elimination of spherical aberration in those (*see* reference, and 280 of 1938) seems never to have been adopted. Electron mirrors proved a forlorn hope for the correction of chromatic aberration. The important effects of so-called "polarisation layers": their study has been neglected up to the present, but "until the polarisation layers are eliminated, all electrostatic systems using slow electrons are likely to be disappointing, and this is particularly true of electron mirrors." High voltage ("pinhole apertures") combined with constancy of voltage ("monochromatic illumination") as the panacea in practice. The writer's recent solution of the problem of the fields corresponding to ideally stigmatic lenses: the three-dimensional ideal lens "exists," uniquely, but cannot be realised without space charges. The two-dimensional (cylindrical) ideal lenses may some of them be capable of approximate realisation. These are all derivable from the "stereographic" distribution by any conformal representation in which the action is kept invariant: "it may be hoped that such representations will have other, less ambitious but more immediately fruitful applications, as they allow the transformation of one electron-optical problem into another, and may even hold the key to certain space-charge problems, for which no satisfactory mathematical treatment has yet been found." The complete lecture is appearing in *Electronic Engineering*.

527. Tracing Valve Characteristics, using the Cathode-Ray Oscillograph.—Bocking. (*See* 436.)

528. A Portable High-Frequency Square-Wave Oscillograph for Television.—Kell, Bedford, & Kozanowski. (*See* 479.)

529. Experimental Investigations on the Resolving Power of Fluorescent Screens.—Hinderer. (*Zeitschr. f. Phys.*, 8th Sept. 1942, Vol. 119, No. 7/8, pp. 397–405.)

Author's summary:—"The resolving power of various polycrystalline screens and one single-crystal screen [barium platinocyanide crystal] was determined for ultra-violet excitation by the double-wedge method: the results were the same whether a 'bright' double wedge was used or a 'dark' double wedge [with the former, the resolving power is measured on non-absorbing particles, whilst the latter method makes use of strongly absorbent particles]. The resolving power is greatly influenced by the roughness of the screen surface and the scattering of the exciting and emitted radiation in the fluorescent material. The measured values lie between 5 and 15 $\mu$ for the different screens [14.0 and 14.7 $\mu$ for commercial screens; 12.1–8.3 $\mu$ for screens of the same material as the commercial, but "pressed" to give a smooth surface; 7.3–6 $\mu$ for screens of the same material (the fine-grain variety) with added particles of soot to reduce the scattering; and finally 5.1–4.4 $\mu$ for the barium platinocyanide single crystal]: within the limits of accuracy of the methods, the values agree with those obtained by the 'sharp-edge' method [using a razor-blade edge in place of the double wedge in the optical system of Fig. 3: the substantial agreement between the two methods indicates that the sharp-edge method, generally the simpler, can be used for such measurements instead of the other]. The 'sharp-edge' method is employed to determine the resolving power of a commercial screen excited by a cathode ray"; the actual values ranged from 70 to 109μ, but the writer concludes that for various reasons the true value is the minimum observed figure of 70μ.

530. ON THE MECHANISM OF LUMINESCENCE IN CRYSTALLINE PHOSPHORS.—Schön. (*Zeitschr. f. Phys.*, 8th Sept. 1942, Vol. 119, No. 7/8, pp. 463–471.)

New experimental results and their application to the "model" (primarily applied to zinc and zinc-cadmium phosphors activated with copper, zinc, or surplus base material) developed by the writer and Riehl (1580 of 1940 and back references).

531. MODERN APPLICATIONS OF LUMINESCENT SUBSTANCES [including Radioactive Luminous Compounds].—Levy & West. (*Journ. Roy. Aeron. Soc.*, Dec. 1942, Vol. 46, No. 384, Abstracts p. 480.)

532. TELEVISION WAVE-FORMS [Analysis of Wave-Forms encountered in Cathode-Ray-Tube Practice], *and* DATA SHEETS XXIX–XXXI.—Lockhart. (*See* 480.)

533. CATHODE-RAY-OSCILLOGRAPH DELAYED SINGLE-SWEEP CIRCUIT [to delay Sweep for Any Desired Time after Peak of Repeating Transient, so that Sweep may cover Any Time Interval desired with respect to Peak: primarily for Cardiac Research].—Gilson. (*Electronics*, March 1942, Vol. 15, No. 3, p. 65.)

534. METHODS OF OBTAINING A TIME-LINEAR VOLTAGE RISE IN SAW-TOOTH OSCILLATIONS [for Simplification of the Interpretation of the Resulting Oscillograms].—Johannsen. (*E.N.T.*, Aug. 1942, Vol. 19, No. 8, pp. 137–147.)

For previous work *see* 136 of January and back reference. Dealing first with the exponential curve resulting from the charging of the "kipp" condenser through an ohmic resistance, the writer discusses various methods of obtaining an adequately linear time base from such a circuit: for the first method, eqn. 8 giving the "unlinearity" $l_t$ (defined by eqn. 3) shows that the steps to be taken are to make the working voltage as high as the insulation and loading capacity of the valve allow, and to keep the saw-tooth amplitude small. For most applications this second condition means that the saw-tooth oscillation will have to be amplified, and by rather a high-quality amplifier. Figs. 5 and 6 show what satisfactorily linear results can be obtained in this way. Another method discussed is to obtain a practically linear deflection by selecting a short section of the exponential curve: this plan is subject to limitations which differ according to the type of circuit employed—glow-discharge tube, thyratron, or multivibrator (normal, "retarding-field" and special symmetrical types: references "3," "4," & "5"). Finally, in many cases sufficient linearity can be obtained, even when the amplitude is quite large, by compensating the exponential curve by means of an amplifier, which is needed anyhow for the sake of a symmetrical voltage and which can be given a characteristic opposed to that of the "kipp" oscillation (Figs. 7 & 8).

The above considerations are all concerned with a condenser charged through a constant ohmic resistance, so that the charging current is bound to vary according to the state of charge of the condenser, thus producing non-linearity. Perfect linearity must be obtainable if, on the other hand, the charging current is kept constant, provided that the sum of all the unwanted resistances parallel to the "kipp" condenser is infinitely large. Methods based on this plan are discussed on pp. 140–144. The use of a saturated diode, or of a triode with positive grid bias and a reduced cathode-temperature, as charging resistance for this purpose has certain disadvantages: a better plan is to employ a multigrid valve in the circuit of Fig. 10, the charging current (and hence the "kipp" frequency) being adjusted by varying either the control-grid voltage (left-hand circuit) or the screen-grid voltage (right-hand circuit). Although the latter arrangement gives the simpler circuit, the former is the more generally satisfactory, having a wider range of linearity and offering, also, a partial way of combating the fact that even in a pentode the anode current is not completely independent of the anode voltage: this partial compensation is carried out by the automatic generation of the control-grid bias by a cathode-circuit resistance. Attempts to improve the linearisation by making the cathode resistance carry only the anode current, and not the screen-grid current (which varies in the opposed sense) as well, by the circuit of Fig. 13 are considered to be not worth the extra voltage source involved, because with modern valves the s.g. current is so small in comparison with the anode current. It is mentioned in passing that in some cases an improvement in linearity can be obtained by a slight positive retarding-grid bias: *cf.* Fig. 9.

The drop of anode current with decreasing anode voltage can also be countered by superposing a voltage of suitable characteristic on one grid of the charging valve: assuming the anode current to decrease in proportion to the anode voltage, the grid voltage required is proportional to $-u_a$, and this can be obtained by reversing the anode-voltage phase by a single normal amplifier stage. Such an arrangement is particularly simple when such an amplifier is needed for the symmetrisation of the time-base voltage: Fig. 14 shows the double purpose of the triode stage "V." But complete success in linearisation by this method is only given when the anode current decreases linearly with anode voltage over the whole current-range in question, when the amplifying stage has a constant amplification factor, and when the grid-voltage/anode-current characteristic is linear also. These requirements are not likely to be fulfilled, and a better plan is to modify the circuit of Fig. 14 to that of Fig. 15, where the variations of the charging current produce voltages which, after phase-reversal by the stage "V," are taken back to the control grid of the charging valve.

The efficacy of the various plans described above are illustrated by the oscillograms of Figs. 16–22: the "unlinearity" $l_t$ amounted to 39% for the worst arrangement, a pentode as charging valve, with bias-producing cathode resistance shunted by a condenser: removal of this condenser improved the value of $l_t$ to 29%, while the separation of the third (retarding) grid from the cathode and its provision with a positive 40 volts or thereabouts reduced $l_t$ so much that it could not be observed (Fig. 18). The phase-reversing circuit of Fig. 14, discussed above, yielded the oscillogram of Fig. 19, with $l_t$ = about 2%, while the improved circuit (Fig. 15) reduced the value to below the limit of observation (Fig. 22).

Finally, section D turns from the methods depending on the use of valves as controlled charging resistances and considers what is termed "linearisation by current superposition" in circuits where ohmic charging resistances are used. In Fig. 23, the "kipp" condenser $C$ is charged through two ohmic resistances $R_{L_1}$ and $R_{L_2}$ in series, so that (apart from the action of the right-hand portion of the circuit) its voltage would rise exponentially. If it is to rise linearly, its charging current $i_c$ must remain constant, and since this passes through the neighbouring resistance $R_{L_2}$, the voltage drop in $R_{L_2}$ must also remain constant: but this is only possible if, in the other part of the charging resistance, $R_{L_1}$, a voltage drop occurs which is exactly opposite (in its variation with time) to the voltage at $C$: this is obtained by making $R_{L_1}$ carry, in addition to the actual charging current $i_c$, a superposed current $i_{ü}$ whose required value is given by $i_{ü} = - u_c/R_{L_1} +$ constant. The complete circuit of Fig. 23 involves the provision of a suitable two-stage amplifier, and this would hardly be justified merely for improving linearity: it does, however, also take the place of a special symmetrising amplifier, so that the expenditure is about the same as for an arrangement with charging through a multi-grid valve and with a symmetrising amplifier. The circuit of Fig. 23 presents advantages in linearity over the multi-grid-valve arrangements, both for very small and for large charging currents.

A simplification of Fig. 23, at the sacrifice of complete compensation and of voltage symmetry, is possible if the loading resistance of the control valve is put in the cathode lead instead of the anode lead. This alters the necessary phase conditions so that a second amplifier-stage is no longer needed and the single-triode circuit of Fig. 24 is arrived at. The improved linearity of the saw-tooth wave given by this circuit is seen by comparing Fig. 26 with Fig. 25, and the oscillogram Fig. 27, using the Fig. 25 wave for time-base, with the oscillogram Fig. 28, using the improved wave for time-base: the "unlinearity" drops from about 75% to about 25%, the still incomplete compensation being the price paid for the simplified amplifier. Finally, the possible application is briefly discussed of the "current superposition" method to the compensation of the current fluctuations of a multi-grid valve—a combination, in fact, of the two main principles of linearisation already treated.

535. A Concise Report on a High-Speed Camera for Simultaneous Photographic and Oscillographic Records [Speeds up to & over 1000 Pictures/Second: "Optical-Compensator" Type, with Lens & Rotating Four-Sided Prism, and Lens for Traces from Three-Element C.R.O.].—Baxter. (*Journ. of Scient. Instr.*, Dec. 1942, Vol. 19, No. 12, pp. 183–184.) From E.R.A. Report Ref. G/T141.

536. A Portable High-Frequency Analyser.—Brauner. (*See* 503.)

537. The Automatic Analysis of Ultra-Low-Frequency Transients: Method of A.M. Grass, B.S. [for the 1–50 c/s Potential Waves of Electro-Encephalography: Use of "Variable-Area" Film Track].—Dawson: Grass. (*Electronic Eng:g*, June 1942, Vol. 15, No. 172, pp. 34–35.)

538. A Camera for recording Low-Speed Transients with a Cathode-Ray Oscillograph [Continuous Run of 40 Feet of Unperforated Paper, at Alternative Rates of ½, 1, & 2 Inches/Second].—Dawson. (*Electronic Eng:g*, July 1942, Vol. 15, No. 173, pp. 52–56.)

539. On the Distribution of Velocities in Field-Emission Electrons [of Importance in estimating Resolving Power of Field-Emission Electron Microscopes, etc.].—Richter. (*See* 445.)

540. On the Resolving Power of the Emission Microscope.—Dosse & Müller. (*Zeitschr. f. Phys.*, 8th Sept. 1942, Vol. 119, No. 7/8, pp. 415–422.)

Authors' summary:—"Following on various recent theoretical treatments by different workers, the lower limit for the resolving power of the emission microscope (Johannson immersion objective and the E.W. Müller field-emission microscope), as determined by the distribution of velocities of the electrons, is given [for cold (field) emission use is made of Richter's work, 539, above]. As the field strength $E$ in front of the cathode is raised, the resolving power theoretically sinks at first as $1/E$ (thermal emission) and then changes at high field strengths (field emission) to a constant value, independent of $E$, which is considerably better as a rule for a plane cathode than for a spherical one. The theoretical values for field emission at a spherical cathode agree satisfactorily with the corresponding experimental values measured on the field-emission microscope. The influence of the condition of the cathode surface [its roughness] is briefly discussed."

541. On the Intensity Relations in the Electron Microscope: II—Capability of Enlargement, Graininess, and Resolving Power of Photographic Plates blackened by Electrons.—von Borries. (*Zeitschr. f. angew. Phot.*, Vol. 4, 1942, p. 42 onwards.) For Part III *see* 542, below.

542. On the Intensity Relations in the Electron Microscope: III—Suitability and Limits of Sensitivity of Photographic Plates for Electron-Microscopic Recording.—von Borries. (*Zeitschr. f. Phys.*, 8th Sept. 1942, Vol. 119, No. 7/8, pp. 498–521.)

For previous parts *see* 3079 of 1942, 143 of January, and 541, above. Author's summary:—"The present paper gives, for a number of different emulsions, the energy density of electron radiation, as a function of ray voltage, necessary to produce the blackening $S=0.5$ which is taken as a suitable basis for comparison [*cf.* 3079 of 1942]: this information allows the most sensitive coatings to be chosen. Further, the resolving powers are given for the same plates, so that those types which are specially suitable for subsequent high enlargement can be selected. The most important factor in judging a plate for the combined requirements of sensitivity and resolving power is the 'image-element work' [quantity of energy which must be expended for blackening an element of size $\delta_P$, where $\delta_P$ is the resolving power of the plate]: this quantity is given, as a function of ray voltage, for the same emulsions [Fig. 4]. From the measurements of 'blackening work' and grain size, the amounts of energy are calculated which must be expended for the blackening of a grain in the developed layer.

"In the last part of the paper the limits of object loading, as governed by heat development and ionisation, are determined for a single exposure in which the object is irradiated only during the actual exposure period. It is found that even with this precaution the temperature equilibrium between absorbed and radiated energy comes in, since the thermal capacity of the object plays no part. The unavoidable ionisation is of importance in the problem of electron-microscopic investigation of living materials. The resolving power which seems attainable from this way of regarding the problem is calculated, and is found to be less favourable than the estimates arrived at elsewhere" [von Ardenne, 3595 of 1941].

543. Electron Reflections in MgO Crystals with the Electron Microscope, *and* Electron-Microscope Study of Surface Structure [by Use of Silica Replicas].—Heidenreich: Heidenreich & Peck. (*Phys. Review*, 1st/15th Sept. 1942, Vol. 62, No. 5/6, pp. 291–292: pp. 292–293.)

(i) The alternate light and dark bands are in contrast to those of Hillier & Baker (3076 of 1942) and von Ardenne (1738 of 1941) in that they are uniformly spaced: this and other results indicate that the phenomenon is the result of multiple electron reflections from the crystallographic planes. (ii) A plastic moulding is made in (preferably) polystyrene: a thin-film replica is then obtained by evaporating silica on to the moulding and subsequently removing the silica by ethyl-bromide solvent.

544. A Diffraction Adapter for the Electron Microscope [to convert It into an Electron-Diffraction Camera].—Hillier & others. (*Phys. Review*, 1st/15th Sept. 1942, Vol. 62, No. 5/6, p. 296: summary only.) *See* also 3346 of 1942.

545. A Sensitive Valve-Voltmeter Relay [for Voltage Stabilisation within less than 0.5%].—Orlov & Pirogov. (*Wireless Engineer*, Aug. 1942, Vol. 19, No. 227, pp. 347–350.) The Russian original was dealt with in 1155 of 1942.

546. Electronic Voltage Stabiliser for X-Ray Tube Power Supply at 20 kV: Constancy to ± 1 Part in 50 000 over Long Periods.—Panofsky & others. (*Phys. Review*, 1st/15th Sept. 1942, Vol. 62, No. 5/6, pp. 218–219.) In a paper on a determination of $h/e$. The stabiliser circuit is similar to that dealt with in 1489 of 1942.

547. A Differential Electronic Stabiliser for Alternating Voltages, and Some Applications.—Glynne. (*See* 519.)

548. Inkless Recording: Sensitive and Reliable [Search for Most Practical System leads to Improvement of "Dotting" Principle (Rate increased to 4 per Second) combined with Typewriter Ribbon].—Twiss. (*Gen. Elec. Review*, Sept. 1942, Vol. 45, No. 9, pp. 511–514.)

549. Solutions for Electrochemical Recording.—Fagan. (*See* 487.)

550. Graphite Shielding: Glass, Cardboard, and Similar Materials as Alternate Shields [including Methods of Contact, Splices, etc.].—Porter. (*Radio* [New York], Oct. 1942, No. 273, pp. 14 and 44.)

551. On the Deionisation in Thyratrons operating as Relaxation Oscillators.—Upatov. (*Journ. of Tech. Phys.* [in Russian], No. 23/24, Vol. 10, 1940, pp. 2011–2021.)

The operation of a relaxation oscillator employing a thyratron fed through a diode (Fig. 1) is discussed. Equation 4 determinng the variation of the deionisation current with time is written down as a first approximation, and from this formulae are derived for determining the period of relaxation oscillations under various conditions and taking into account the deionisation effect. Experiments have confirmed that the accuracy of the theoretical results obtained is sufficient for practical purposes. Oscillographic methods have also been developed for measuring certain parameters characterising the deionisation process, namely $\tau$ (deionisation time), $\lambda_o$ (initial conductivity), and $\bar{n}_o$ (initial ion concentration). Results obtained in measuring these parameters in the case of mercury-vapour, neon, and argon thyratrons of Russian manufacture are shown and discussed.

552. The Measurement of Back Currents in Ionic Apparatus.—Granovski & Merzloukhova. (*Journ. of Tech. Phys.* [in Russian], No. 23/24, Vol. 10, 1940, pp. 2022–2032.)

The back current which flows during the half-periods of low conductivity is a factor of great importance in the operation of ionic tubes, and many attempts, briefly reviewed at the beginning of this paper, have therefore been made to devise methods for measuring it. The method proposed by the authors consists in taking off the voltage due to the back current passing through a resistance in the anode circuit of the tube, and after amplification by a d.c. amplifier applying it to a cathode-ray oscillograph (Fig. 2). The design of the resistance and of the amplifier, which have to meet very strict special requirements, is discussed in detail, and a report is given with a number of oscillograms on measurements of the back current in a mercury-vapour thyratron designed for a rectified current of 15 amperes.

553. "THEORY OF GASEOUS CONDUCTION AND ELECTRONICS" [Book Review].—Maxfield & Benedict. (*Proc. I.R.E.*, April 1942, Vol. 30, No. 4, p. 211.)

554. DEPARTURES FROM OHM'S LAW IN SOLIDS [and Experiments on Parallel Crystal Plates of Carborundum, Zincite, & Galena].—Osterberg. (*See* 393.)

555. CONTACTS BETWEEN METALS AND BETWEEN A METAL AND A SEMICONDUCTOR.—Fan. (*See* 446.)

556. HALL EFFECT AND CONDUCTIVITY OF CUPROUS OXIDE [Departure from Exponential Law of Temperature Dependence & Its Two Causes: Deeper Insight into Nature of Ageing given by Experiment on Effect of Electrical Field].—Angello. (*Phys. Review*, 1st/15th Oct. 1942, Vol. 62, No. 7/8, pp. 371–377.)

557. SELENIUM RECTIFIERS AND THEIR DESIGN [Design Procedures & Computations: Nomenclature: etc.].—Yarmack. (*Elec. Communication*, No. 4, Vol. 20, 1942, pp. 275–286.) Published also in the July 1942 section of *A.I.E.E. Transactions*, Vol. 61.

558. SPARK-SUPPRESSION BY DRY-PLATE RECTIFIERS [for Preservation of Contacts & Prevention of Interference].—Schmidt. (*See* 367.)

559. VIBRATOR WAVE-FORMS [and Their Importance for the Life of the Vibrator].—Robinson. (*Electronic Eng:g*, June 1942, Vol. 15, No. 172, pp. 32–33.) Following on 1492 of 1942.

560. CORRECTION TO "A NEW EQUIPMENT FOR TESTS ON THE FUSING BEHAVIOUR OF CONTACTS."—Linckh & Krapf. (*E.T.Z.*, 8th Oct. 1942, Vol. 63, No. 39/40, p. 480.) *See* 183 of January.

561. ELECTRICAL BREAKDOWN IN VACUUM BETWEEN ELECTRODES OF LARGE AREA [500 cm², instead of Usual Small Electrodes: 7 mm Spacing: Voltages Up to 100 kV: Six Factors controlling Breakdown Voltage: Effect of Removal of All Insulation from Regions of Strong Field].—Seifert. (*Phys. Review*, 1st/15th Sept. 1942, Vol. 62, No. 5/6, pp. 300–301: summary only.)

562. THE ELECTRICAL STRENGTH OF NITROGEN AND FREON UNDER PRESSURE.—Skilling & Brenner. (*Elec. Communication*, No. 4, Vol. 20, 1942, pp. 287–294.) Reprinted from the April 1942 section of *A.I.E.E. Transactions*, Vol. 61. A summary was dealt with in 3720 of 1942.

563. HIGH-FREQUENCY ARTICULATED AIR-SPACED CABLES AND VALVE CONNECTOR [" Co-ax " Cables].—Telequipment Company. (*Journ. of Scient. Instr.*, Dec. 1942, Vol. 19, No. 12, p. 188.) *See* also 3580 of 1942.

564. PLASTIC-COVERED CONDUIT [Flexible Metal Conduit protected by Transflex, Irv-O-Lite XTE-30, or Hyflex].—Searle Aero Industries. (*Gen. Elec. Review*, Oct. 1942, Vol. 45, No. 10, p. 595.)

565. INTELIN [Styrene] HIGH-TENSION APPLICATIONS, *and* INTELIN HIGH-FREQUENCY APPLICATIONS [in these Cables the Intelin Insulation has a Dielectric Constant of 2.38 and a Power Factor at 100 Mc/s of 0.0007].—(*Elec. Communication*, No. 4, Vol. 20, 1942, p. 316: p. 316.)

566. ELOXAL-INSULATED WIRES IN ELECTROTECHNICS [in replacing Copper by Aluminium].—Stender. (*E.T.Z.*, 24th Sept. 1942, Vol. 63, No. 37/38, pp. 451–452: summary only.)

567. RUBBER FROM EUROPEAN PLANTS [Eberswalde Investigations]. — Koropp. (*T.F.T.*, April 1942, Vol. 31, No. 4, p. 119: summary only.)

568. THE PROPERTIES OF SOFT RUBBERS FROM THE VIEWPOINT OF CABLE CONSTRUCTION.—Roelig. (*E.T.Z.*, 8th Oct. 1942, Vol. 63, No. 39/40, pp. 465–466.)

569. THE TOLERANCES FOR HARD-PAPER AND HARD-CLOTH PRODUCTS. — Wandeberg. (*E.T.Z.*, 24th Sept. 1942, Vol. 63, No. 37/38, pp. 455–456.)

570. MYCALEX: ITS MANUFACTURE AND PROPERTIES.—(*Electronic Eng:g*, May 1942, Vol. 14, No. 171, pp. 749–752.)

571. THE MECHANICAL PROPERTIES OF GLASS.—Preston. (*Journ. Applied Phys.*, Oct. 1942, Vol. 13, No. 10, pp. 623–634.) " We use glass in this case [200-inch mirror for Mount Palomar], not because it is transparent, but because its general rigidity and permanence of shape are better than steel or concrete."

572. ON THE EFFECT OF THERMAL EXPANSION ON THE ABSORPTION SPECTRUM OF INSULATORS [Lattice Expansion, as well as " Multiple Impacts," contributes to Long-Wave Displacement at High Temperatures].—Möglich & Rompe. (*Zeitschr. f. Phys.*, 8th Sept. 1942, Vol. 119, No. 7/8, pp. 472–481.) Further development of the researches on insulators and phosphors dealt with in 1772, 1773, & 2800 of 1941.

573. FLUORESCENCE OF SOLUTIONS AND DIELECTRICAL PROPERTIES OF SOLVENTS [" (O–O) Separation " & Red Shift explained as Effects of Interaction between Molecules in Liquid State, and discussed in light of Onsager's Theory of Polar Liquids].—Sambursky & Wolfsohn. (*Phys. Review*, 1st/15th Oct. 1942, Vol. 62, No. 7/8, pp. 357–361.)

574. VALIDITY OF THE CLAUSIUS–MOSOTTI FORMULA [Alternative Formula which would affect Many Dipole Moments].—Burniston Brown. (*See* 607.)

575. THE MECHANISM OF POLARISATION OF CERTAIN CRYSTALLINE ORGANIC COMPOUNDS. — Walther, Inge, & Trambitskaya. (*Journ. of Tech. Phys.* [in Russian], No. 23/24, Vol. 10, 1940, pp. 1970–1974.)

Hard substances produced by hydrogenating castor oils in a special manner, and having dielectric constants reaching values of the order of 20–25, have been used lately for impregnating paper condensers. Such high values of $\epsilon$ can only be explained by orientated polarisation, *i.e.* by the rotation either of polarised molecules or of component parts of the molecule. A new substance called "oleovaks" has been developed by the Institute of Oil Industry in Russia, and in this paper a report is given on an experimental investigation of this. Curves are plotted in Figs. 2, 3, and 4 showing the effect of temperatures from $-60$ to $+100$°C on $\epsilon$ and tan $\delta$ of different kinds of "oleovaks" at frequencies of 50, 2000, and $2 \times 10^6$ c/s, and several microphotographs taken with polarised light are shown. A theoretical interpretation of the results obtained is given.

576. ON DISCHARGE-VOLTAGE AND RETURN-VOLTAGE CURVES FOR ABSORPTIVE CAPACITORS [Question of the Validity of Böning's Formulae (*cf.* 2814 of 1942): Application of Principle of Superposition furnishes Rigorous Treatment of Discharge-Voltage & Return-Voltage Curves and a Way of applying These in Measuring Technique].—Gross. (*Phys. Review*, 1st/15th Oct. 1942, Vol. 62, No. 7/8, pp. 383–387.)

577. THE TEMPERATURE COMPENSATION OF CONDENSERS.—Griffiths: Britton. (*Wireless Engineer*, June 1942, Vol. 19, No. 225, pp. 253–255.) Griffiths replies to the criticisms mentioned in 2486 of 1942: Britton criticises his remarks on ceramic condensers, and Griffiths replies in the July issue, No. 226, pp. 308–309.

578. ON THE THEORY OF THE MAGNETIC FIELD [New Formulation of Ampère's "Elementary-Current" Basis of Ferromagnetic Phenomena gives Complete & Consistent Quantitative Picture].—Kneissler-Maixdorf. (*Arch. f. Elektrot.*, Aug. 1942, Vol. 36, No. 8, pp. 471–483.) Further development of 292 of 1942.

579. CORRECTIONS TO "ON D.C.-POLARISED A.C. CHOKING COILS AND THEIR BACK COUPLING." —Buchhold. (*Arch. f. Elektrot.*, Aug. 1942, Vol. 36, No. 8, p. 514.) *See* 3547 of 1942.

580. THE CALCULATION OF CHOKE COILS WITH SEVERAL WINDINGS.—Böhse & Bielert. (*See* 362.)

581. LAMINATION DESIGN: INFLUENCE OF SHAPE ON THE COST AND WEIGHT OF TRANSFORMERS AND CHOKES.—Partridge. (*Wireless World*, Dec. 1942, Vol. 48, No. 12, pp. 286–289.)

582. INTERNAL TRANSFORMATIONS IN THE Fe–Ni–Al ALLOYS.—Lifshits. (*Journ. of Tech. Phys.* [in Russian], No. 23/24, Vol. 10, 1940, pp. 1981–1985.)

The effects of various thermal treatments on the coercive force of the Fe–Ni–Al alloys were investigated experimentally. A number of experimental curves are shown and definite conclusions reached. A theoretical interpretation of the results obtained is made rather difficult by the fact that no structural transformations could be observed in the alloys during thermal treatments, either by microscopic or X-ray methods. Nevertheless, several hypotheses are put forward and discussed in detail.

583. NEW MAGNETIC MATERIALS [Survey, with 16 Literature & Patent References: Data, including for Nicaloi (Hot- & Cold-Rolled, Cold-Rolled *plus* Magnetic Anneal), etc: Permanent-Magnet Materials, including the Latest Alnico IV & V].—Ruder. (*Proc. I.R.E.*, Oct. 1942, Vol. 30, No. 10, pp. 437–440.)

584. NEW KINDS OF STEEL OF HIGH MAGNETIC POWER.—Jonas & van Embden. (*Philips Tech. Review*, Jan. 1941, Vol. 6, No. 1, p. 8 onwards: referred to in 583, above.) *See* 2138 of 1942.

585. PERMANENT MAGNETS [Editorial on Recent Development of Materials, including Ticonal 2A & 40/50, Alnico V, and the "Magnetic Delay" Steel (2518 of 1942)].—G. W. O. H. (*Wireless Engineer*, July 1942, Vol. 19, No. 226, pp. 287–289.)

586. MAGNETIC MATERIALS [and the Improved Design & Performance of A.C. & D.C. Indicating Instruments due to the New Alloys].—Corson. (*Gen. Elec. Review*, Oct. 1942, Vol. 45, No. 10, pp. 573–575.)

587. CATHODIC SPUTTERING: ITS NATURE AND EFFECTS [and Its Industrial Application, *e.g.* to Dry-Plate Photocells].—Haigh. (*Electronic Eng:g*, July 1942, Vol. 15, No. 173, pp. 61–64.)

588. ON THE THEORY OF CATHODE SPUTTERING.—Seeliger. (*Zeitschr. f. Phys.*, 8th Sept. 1942, Vol. 119, No. 7/8, pp. 482–492.)

589. A NEW METHOD OF MAKING ALLOYS OF METALS THAT DO NOT MIX WHEN MELTED. —Samuels & others. (*Science*, 23rd Oct. 1942, Vol. 96, Supp. pp. 8 and 10.)

590. A LITHIUM-BEARING SILVER SOLDER [and Its Advantages for Various Purposes].—Hensel & others. (*Engineering*, 4th Dec. 1942, Vol. 154, p. 460: summary only.)

591. SAVING TUNGSTEN AND COPPER BY USING SILVER.—(*Electronics*, March 1942, Vol. 15, No. 3, p. 104.)

592. NEW SOLDERING FLUXES [leaving Non-Corrosive Residues, and particularly suitable for Fine Work: Lactic Acid with Calsolene-Oil Addition, etc.]—(*Electronic Eng:g*, June 1942, Vol. 15, No. 172, p. 37: summary, from *Tin & Its Uses.*) *Cf.* Homer & Watkins. 2110 of 1942.

## STATIONS, DESIGN AND OPERATION

593. A Transceiver for WERS [War Emergency Radio Service: "a 2½-Metre Outfit from Junk-Box Components"].—Grammer. (*QST*, Oct. 1942, Vol. 26, No. 10, pp. 11–15 and 84, 88, 90.)

594. Frequency Allocations in the WERS: a State-Wide Plan for Clear-Channel Operation without Mutual Interference [Massachusetts Plan].—Ling. (*QST*, Oct. 1942, Vol. 26, No. 10, pp. 16–17.)

595. A Frequency-Modulation Station Monitor [measuring Mean Carrier Frequency, with & without Modulation, and Modulation Percentage: Over-Modulation Alarm & Fidelity-Monitoring Circuit].—Summerhayes. (*Proc. I.R.E.*, Sept. 1942, Vol. 30, No. 9, pp. 399–404.) A summary was dealt with in 1119 of 1942.

596. Frequency Modulation: Part VII—Reception.—Sturley. (*Electronic Eng:g*, May 1942, Vol. 14, No. 171, pp. 755–758.)

597. Aircraft Radio Communications [Two-Way Communications, *En Route* Navigation, Instrument Landing: Aerials: Table of Weights: etc.].—Morgan. (*Journ. Roy. Aeron. Soc.*, Dec. 1942, Vol. 46, No. 384, Abstracts p. 483.)

598. German Aircraft Radio [Illustrated, from Official Reports: including Emergency Equipment].—(*Electronic Eng:g*, June 1942, Vol. 15, No. 172, pp. 8–12 and 36.)

599. Wireless *versus* Wire Broadcasting after the War [Paper dealing "Frankly & Clearly" with Pros & Cons, and concluding that Wire Broadcasting should be pressed on with immediately after a "Victorious Ending" of the War]. — Gladenbeck. (*E.N.T.*, July 1942, Vol. 19, No. 7, p. 136: short summary only.) In a review of the Year-Book dealt with in 2861 of 1942.

600. The Operative Attenuation of Four-Pole-Network Chains [Wire-Broadcasting Lines, etc.]—Klein. (*See* 356.)

601. Experimental Polyphase Broadcasting [on 1 kW, at 1 Mc/s: WHO's Experimental Transmitter: System found Entirely Practical for Modern High-Fidelity A.M. Broadcasting: Peak Power Capability of Output Valves only about 1½ Times the Carrier Power].—Loyet. (*Proc. I.R.E.*, May 1942, Vol. 30, No. 5, pp. 213–222.) For this system *see* Byrne, 4229 of 1939.

602. Paper on the Luxembourg Effect as a Limitation on the Permissible Increase in Broadcasting-Station Power, and Common-Wave Broadcasting as a Possible Cure.—Vilbig. (*See* 320.)

603. The Service Area of Medium-Power Broadcast Stations [500–1500 kc/s: the Average Field-Intensity Increases obtained by Improvement of Each of the Four Factors Power, Area Conductivity, Frequency, & Placing of Station with respect to Distribution of Population].—Patrick. (*Proc. I.R.E.*, Sept. 1942, Vol. 30, No. 9, pp. 404–410.) From the South African Broadcasting Company.

604. Problems of Contrast Compression [General Considerations leading to the "Ideal" Compression Characteristic: the Correct Choice of the "In" & "Out" Control Speeds, governed by the Building-Up & Decay Times of Musical Instruments: a Compression Circuit satisfying Both Requirements].—Hildebrand. (*Funktech. Monatshefte*, April 1942, No. 4, pp. 56–59.)

605. Post-War Broadcasting [Plea for Controlled Volume Compression: Suggestion that Certain Proportion of Broadcasting Time should be Sold to Advertisers for benefit of Hospitals].—King: Anon. (*Wireless World*, Dec. 1942, Vol. 48, No. 12, p. 290: p. 290.)

606. Broadcast Station Maintenance & Operation [Daily Routines & Periodic Checks, etc.].—Wesser. (*Radio* [New York], Oct. 1942, No. 273, pp. 19–21.)

## GENERAL PHYSICAL ARTICLES

607. Validity of the Clausius-Mosotti Formula [Doubts on: Alternative Formula which, if correct, would considerably affect Many Dipole Moments]. — Burniston Brown. (*Nature*, 5th Dec. 1942, Vol. 150, pp. 661–662.)

"The conclusion seems to be that only a macroscopic treatment can be applied and that the formula should be $K = 1 + 4\pi n\sigma$. Thus for the molar polarisation $P$ we should have $P = \{(K-1)/3\} . M/\rho$ instead of the Clausius-Mosotti $P = \{(K-1)/(K+2)\} . M/\rho$".

608. Ionisation of Gases by Collisions of their Own Accelerated Molecules [Measurements on A, $N_2$, $H_2$, & He in the 1000–8000 eV Range].—Berry. (*Phys. Review*, 1st/15th Oct. 1942, Vol. 62, No. 7/8, pp. 378–382.) For a previous paper *see* 1552 of 1942.

609. Mobilities in Nitrogen at High Current Densities.—Bennett. (*Phys. Review*, 1st/15th Oct. 1942, Vol. 62, No. 7/8, pp. 369–371.)

610. Measurement of Electronic Charge [and the Agreement between Oil-Drop & X-Ray Methods (4.8020 & $4.8023 \times 10^{-10}$ e.s.u. respectively)].—Laby. (*Nature*, 5th Dec. 1942, Vol. 150, pp. 648–649.)

611. A New Test of General Relativity using Radioactive Ions, revolved in a Cyclotron, as "Moving Clocks."—Berenda. (*Phys. Review*, 1st/15th Sept. 1942, Vol. 62, No. 5/6, pp. 288–290.) In a paper on the problem of the rotating disc. When brought to rest, the ions should be more radioactive than a sample which had remained at rest.

612. More Magnetic Fallacies [and in particular the Comparison between the Inductances of a Straight Wire & Solenoid].—G. W. O. H : Hatfield. (*Wireless Engineer*, June 1942, Vol. 19, No. 225, pp. 239–242.) Editorial on 2539 of 1942.

613. On the Theory of the Magnetic Field [New Formulation of Ampère's "Elementary Current" Picture].—Kneissler-Maixdorf. (*See* 578.)

614. "Electric and Magnetic Fields" [Book Review].—Attwood. (*Electronic Eng:g*, June 1942, Vol. 15, No. 172, p. 42.)

## MISCELLANEOUS

615. "Chart Atlas of Complex Hyperbolic and Circular Functions (Third Edition, 1924)," *and* "Tables of Complex Hyperbolic and Circular Functions (Second Edition, 1927)" [Book Reviews].—Kennelly. (*Proc. I.R.E.*, April 1942, Vol. 30, No. 4, p. 211.)

616. "Tables of Functions" [1933 & 1938 Editions : Book Review].—Jahnke & Emde. (*Proc. I.R.E.*, July 1942, Vol. 30, No. 7, p. 353.)

617. Note on Dimensions [Criticism of Guggenheim's Paper].—Wilson. (*Phil. Mag.*, Nov. 1942, Vol. 33, No. 226, pp. 842–844.) *See* 3070 of 1942.

618. On fitting Data with Polynomial Functions by the Method of Least Squares.—Weinberg. (*Phys. Review*, 1st/15th Sept. 1942, Vol. 62, No. 5/6, p. 304 : summary only.)

619. The Calculation of Experimental Errors : an Effective Grouping and Analysis of Observational and Experimental Errors.—Espy. (*Communications*, June 1942, Vol. 22, No. 6, pp. 8–9.)

620. Inspection by Sampling : Effect of Number of Samples to Bulk Number.—Parsons. (*Journ. Roy. Aeron. Soc.*, Dec. 1942. Vol. 46, No. 384, Abstracts p. 473.) *See* also *Engineering*, 13th Nov. 1942, p. 395.

621. Standards and Standardisation [Editorial, including Proposed "Orders of Precision"].—(*See* 516.)

622. Harmonic Analysis of Waves containing Odd and Even Harmonics.—Kemp. (*See* 465.)

623. Reminiscences of an Instrument Maker.—Whipple. (*Journ. of Scient. Instr.*, Dec. 1942, Vol. 19, No. 12, pp. 178–183.)

624. Joseph Henry, Pioneer in Space Communication.—Magie. (*Proc. I.R.E.*, June 1942, Vol. 30, No. 5, pp. 261–266.) For an Editorial on this article see *Wireless Engineer*, Dec. 1942, Vol. 19, No. 231, pp. 547–549.

625. Frequency . . . [Public Lectures on Centimetric Waves, Wave Guides, Cavity Resonators, etc.].—Ramo. (*Gen. Elec. Review*, Oct. 1942, Vol. 45, No. 10, pp. 557–567.)

626. The Frequency Spectrum [and the Present Confusion of Terminology].—Smith-Rose. (*See* 327.)

627. Classification of Frequencies : Need for a Universal System [Editorial].—(*See* 328.)

628. Pulsatance, Rotatance, or Velocitance ? [Inappropriateness of "Pulsatance" to represent Vector Angular Velocity : Suggestions wanted : Suggestions].—G. W. O. H : Sturley : Bloch. (*Wireless Engineer*, June 1942, Vol. 19, No. 225, p. 242 : July, No. 226, p. 309 : August, No. 227, p. 361.) Sturley suggests "pulsance" or even "angulocity," and Bloch "phase rate."

629. "Fundamentals of Electric Waves" [providing "a Working Knowledge of the Field Theory" : Book Review].—Skilling. (*Science*, 23rd Oct. 1942, Vol. 96, p. 363.)

630. "Electric Circuits" [Book Review].—E.E. Staff of M.I.T. (*Proc. I.R.E.*, May 1942, Vol. 30, No. 5, p. 257.)

631. "Physik und Technik des elektrischen Fernmeldewesens : I" [Electroacoustics, Sound Films, Telephony & Telegraphy, Voice-Frequency Transmission over Lines : with Appendix of Circular & Hyperbolic Functions : Book Review].—Neckenbürger. (*E.T.Z.*, 8th Oct. 1942, Vol. 63, No. 39/40, p. 478.)

632 Significance and Foundations of Modern Telegraphic Systems [Comprehensive Survey].—Simon. (*Funktech. Monatshefte*, May 1942, No. 5, pp. 61–76.)

633. The Evolution of Wire Transmission [and particularly the "Broad Band" Period beginning 1937].—Gray. (*Elec. Communication*, No. 4, Vol. 20, 1942, pp. 235–245.)

634. Carrier-Frequency Equipment for Short Distances [giving Additional Speech Channel on Two-Wire Circuit over about 12–15 km : with No Valve Apparatus at Far End, only Rectifier & Filters].—Fetzer & Hamm. (*E.T.Z.*, 24th Sept. 1942, Vol. 63, No. 37/38, p. 451 : summary only.)

635. Automatic Monitoring Circuit [primarily to relieve Broadcasting-Station Operators of Extra Strain of Monitoring the "Key" Station's Programme as well as Their Own, to intercept Air Raid Warnings : but with Other Applications also].—Marx. (*Electronics*, March 1942, Vol. 15, No. 3, pp. 39 and 96.)

636. Sweden and Magnetic Mines : "Electric Massaging" of Ships ["by passing Them through a High-Tension Field"].—(*Electrician*, 11th Dec. 1942, Vol. 129, p. 627.)

637. The Problem of the Raising of Output in Manufacturing.—Kienzle. (*Zeitschr. V.D.I.*, 31st Oct. 1942, Vol. 86, No. 43/44, pp. 641–648.)

638. The Complete Utilisation of Scientifically Trained Personnel.—Grundfest. (*Science*, 2nd Oct. 1942, Vol. 96, pp. 318–319.) From the Secretary, American Association of Scientific Workers.

639. The Establishment of a Joint Council of Professional Scientists. — (*Journ. of Scient. Instr.*, Dec. 1942, Vol. 19, No. 12, p. 192.)

640. A Technological High Command : How Fast is the U.S. moving towards It ? Too Slowly . . . — (*Proc. I.R.E.*, July 1942, Vol. 30, No. 7, pp. 309–318 : from *Fortune*.)

641. A Symposium on Radio in the War Effort [by Presidents of R.M.A., N.A.B., & I.R.E.]. —(*Proc. I.R.E.*, Oct. 1942, Vol. 30, No. 10, pp. 479–484.)

(i) " We are constantly concerned with the matter of maintenance and replacement parts . . . We are even investigating the future prospects of utilising Boy Scouts for the necessary servicing of the radio sets in the hands of the public, when the service organisation as we know it has completely gone to war . . . " (ii) What radio broadcasting means in the war effort. (iii) What has been done and what remains to be done, particularly the utilisation of the broadcasting system for information, instruction, and control in air attack.

642. Wartime Engineering [War & the Engineer : Engineering Ethics : the Engineering Virtues : Engineering Methods : etc.].—Goldsmith. (*Proc. I.R.E.*, July 1942, Vol. 30, No. 7, pp. 319–328.)

643. The Engineer in Modern Society.—Van Dyck. (*Proc. I.R.E.*, July 1942, Vol. 30, No. 7, pp. 305–309.) " Why has technology brought about bigger depressions instead of smaller ones, and why more terrible wars instead of none at all ? " " The defences [Pearl Harbour] included devices capable of preventing surprise attacks. Then why was surprise attack possible ? " For a letter to which the Editor calls special attention *see* October issue, No. 10, p. 485.

644. Editorial : Research and Research Workers [Wartime Problems : Cooperative Research : Research & Production]. — G. W. O. H. (*Wireless Engineer*, Sept. 1942, Vol. 19, No. 228, pp. 391–393.)

645. Imagineering [and the Part It should play in the Application of Physics to Industry].—(*Journ. of Scient. Instr.*, Dec. 1942, Vol. 19, No. 12, p. 192 : paragraph on an *Indust. & Eng. Chemistry* editorial.)

646. Engineers train for Victory : the Engineering, Science, and Management Defence Training (ESMDT) Courses.—Dudley. (*Electronics*, March 1942, Vol. 15, No. 3, pp. 36–38 and 103.)

647. *QST* visits Fort Monmouth : Training a Fighting Corps of Radio Operators and Repairmen.—De Soto. (*QST*, Oct. 1942, Vol. 26, No. 10, pp. 28–32 and 96, 98, 106, 108.)

648. Training Signal Instructors [at the School of Signals].—(*Wireless World*, Dec. 1942, Vol. 48, No. 12, pp. 283–285.)

649. Some Notes and Suggestions on the Teaching of Physics [Plea for Greater Exactness in defining Physical Quantities and for Greater Attention to Correct Use of Algebraic Signs : Examples : Improvements of Well-known Experiments : etc.].—Smith. (*Phil. Mag.*, Nov. 1942, Vol. 33, No. 226, pp. 775–815.)

650. Scientific Films and Their Use in the Teaching of Physics [Physical Society Discussion].—Ashhurst & others. (*Nature*, 12th Dec. 1942, Vol. 150, p. 691 : summary only.)

651. " How to Supervise People " [Book Review]. —Cooper. (*Proc. I.R.E.*, July 1942, Vol. 30, No. 7, p. 353.) " There are . . . few supervisors, young or old, whose effectiveness will not be increased by a consideration of the suggestions."

652. More on the Jap Code.—Holden. (*QST*, Oct. 1942, Vol. 26, No. 10, pp. 74–75.) *See* also Millikin, 253 of January.

653. International Systems and Standards [including the Need for an Auxiliary International Language]. — Gregory. (*Nature*, 28th Nov. 1942, Vol. 150, pp. 620–622.) From an address to the ASLIB conference reported on pp. 637–638.

654. " Booklet on Workshop Terms in Polish and English " [Book Review].—(*Electronic Eng:g*, June 1942, Vol. 15, No. 172, p. 37.)

655. " Englisch-Deutsches Wörterbuch für Eisenbahnsicherungswesen und Fernmeldetechnik " [Book Review].—Schmitz. (*E.N.T.*, July 1942, Vol. 19, No. 7, pp. 135–136.)

656. ASLIB [Editorial on the Association of Scientific Libraries and Information Bureaux & Its Services].—(*Electronic Eng:g*, July 1942, Vol. 15, No. 173, p. 51.) " An organisation which deserves to be more widely known among workers in the electronic industry."

657. Universal Decimal Classification [Notice of the Reprinting of a Limited Edition : Orders invited].—British Standards Institution. (*Wireless Engineer*, Dec. 1942, Vol. 19, No. 231, p. 555.)

658. Documentation and Microfilm [Papers & Discussions].—(*Nature*, 21st Nov. 1942, Vol. 150, p. 601 : summary, from *Proc. British Soc. or Internat. Bibliography*.)

659. The Preservation of Engineering Drawings in the Form of Microfilms.—Penther. (*Zeitschr. V.D.I.*, 31st Oct. 1942, Vol. 86, No. 43/44, p. 661.)

660. Electronics on the Air [Fiteen-Minutes' Broadcasts Three Times a Week, sponsored by General Electric: with Stories of Developments made Possible by Electronics].—(*Electronics*, March 1942, Vol. 15, No. 3, p. 104.)

661. Radioelectric Apparatus and Components at the Lyons International Fair, 1942.—(*Génie Civil*, 15th Oct. 1942, Vol. 119, No. 23, pp. 283–284.)

662. Suggested A.F. or Supersonic Communication using Stratum of Bed Rock as Propagating Medium.—Chandler. (*QST*, Oct. 1942, Vol. 26, No. 10, pp. 110 and 112.)

663. The Optics of Supersonic Waves, and Supersonic Images [Problem of Vision through Opaque Media].—Barbier. (*See* 473.)

664. Studies in Superconductivity [Experiments in connection with a Superconducting Bolometer for Infra-Red Radiation].—Bruckschi & others. (*See* 521.)

665. Infra-Red Lamp Heating: Paint Drying and Other Industrial Applications.—Rowland. (*Electrician*, 27th Nov. 1942, Vol. 129, pp. 575–577.)

666. The Relative Sensitivities of Television Pick-Up Tubes, Photographic Film, and the Human Eye.—Rose. (*See* 475.)

667. Physiological-Optical Bases of the Air-Raid Blackout [with Data regarding Threshold of Eye for Various Colours, Nomogram for Relation between Aircraft Height, Surface & Intensity of Light Source, etc.].—Bouma. (*E.T.Z.*, 8th Oct. 1942, Vol. 63, No. 39/40, pp. 473–474: summary, from *Physica*, Vol. 8, 1941, p. 398 onwards.)

668. The Electroplane Camera [for Great Depths of Field: with "Detrar" Objective improved by Replacement of Mechanical Method of Movable-Element Shift by Electronic Method of Oscillation by "Lens Coil"].—Smith-Dieterich Corporation. (*Electronics*, March 1942, Vol. 15, No. 3, pp. 44–47.) *Cf.* 2739 of 1942 and back references.

669. Automatic Recording of Light-Sensitive-Cell Curves by the Photoelectric Recorder.—Wrobel. (*See* 485.)

670. The General Electric Light Meter and Its Adaptation to Various Other Light-Measuring Devices (Ultra-Violet Meter, Exposure Meter, etc.).—Dows. (*Gen. Elec. Review*, Sept. 1942, Vol. 45, No. 9, pp. 505–509.)

671. An Ultra-Violet Recording Photoelectric Spectrophotometer. — Pfister & Rank. (*Phys. Review*, 1st/15th Sept. 1942, Vol. 62, No. 5/6, p. 298: summary only.) For the "extended-range galvanometer-deflection recording system" *see* 3449 of 1942.

672. Electronic Control for Carbon Arcs [in Motion-Picture Projection].—Flaherty. (*Electronics*, March 1942, Vol. 15, No. 3, pp. 65–66 and 70.)

673. Photoelectric Scanner [Light Source, Lens System, & Two Phototubes, in Single Housing, useful wherever Light reflected from a Surface close to Unit can be employed to Control an Industrial Process].—United Cinephone. (*Electronics*, March 1942, Vol. 15, No. 3, pp. 70 and 72.)

674. Phototube-Controlled Revolving Door.—(*Electronics*, March 1942, Vol. 15, No. 3, p. 72.)

675. Photoelectric Intrusion-Detection System [for Protection against Sabotage: Units incorporating Mechanical & Electronic Design Innovations].—Electronic Control Corporation. (*Electronics*, March 1942, Vol. 15, No. 3, pp. 73 and 74.)

676. The Use of Secondary Electron Emission to obtain Trigger or Relay Action [similar to That of a Thyratron but capable of much Higher Operating Speeds].—Skellett. (*Phys. Review*, 1st/15th Sept. 1942, Vol. 62, No. 5/6, p. 296: summary only.)

677. Electrical Extensometers [Carbon-Line Principle: made by Philips Electric Company, Zurich]. — Stettler. (*Journ. Roy. Aeron. Soc.*, Dec. 1942, Vol. 46, No. 384, Abstracts p. 480.)

678. Differential - Input Circuits for Biological Amplifiers [Survey, including Papers occurring only in Medical Journals], *and* Discussion of the Above Paper.—Saunders: Debski. (*See* 344.)

679. The Automatic Analysis of Ultra-Low-Frequency Transients: Method of A. M. Grass, B.S. [for Electro-Encephalography].—Dawson: Grass. (*See* 537.)

680. A Camera for recording Low-Speed Transients with a Cathode-Ray Oscillograph.—Dawson. (*See* 538.)

681. Cathode - Ray - Oscillograph Delayed Single - Sweep Circuits [primarily for Cardiac Research].—Gilson. (*See* 533.)

682. Supplementary Letter on "Electronic Switching in Medical Research: Simplified Apparatus for Multiple Recording" [2888 of 1942].—Dawson. (*Electronic Eng:g*, May 1942, Vol. 14, No. 171, p. 772.)

683. High-Frequency Apparatus: Relaxation of Control Order.—(*Wireless Engineer*, Sept. 1942, Vol. 19, No. 228, p. 400.)

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