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WIRELESS ENGINEER

The Journal of Radio Research & Progress

Vol. XXIII. APRIL 1946 No. 271

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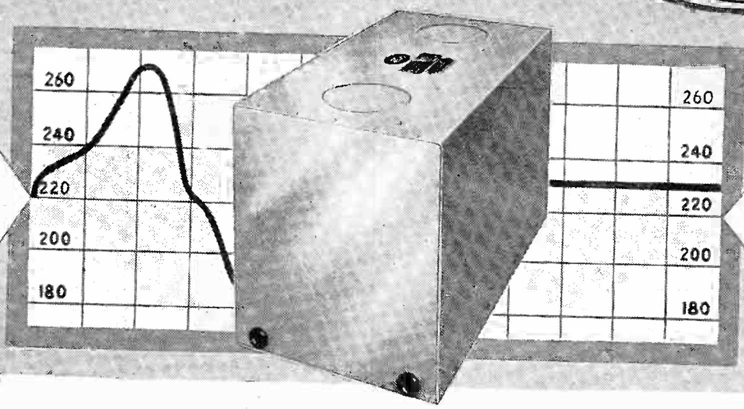

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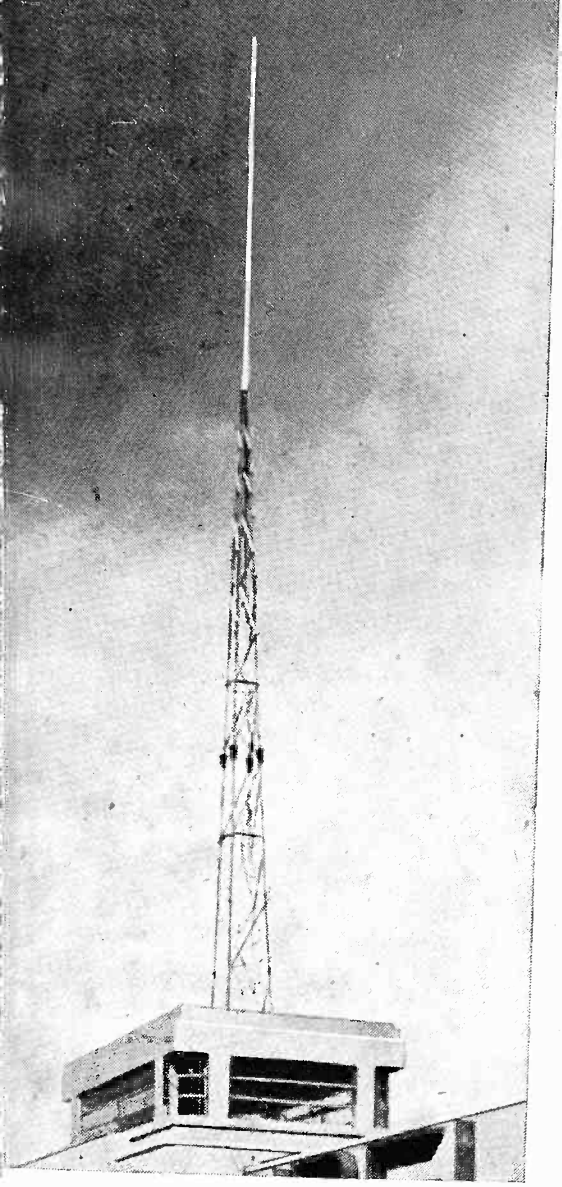

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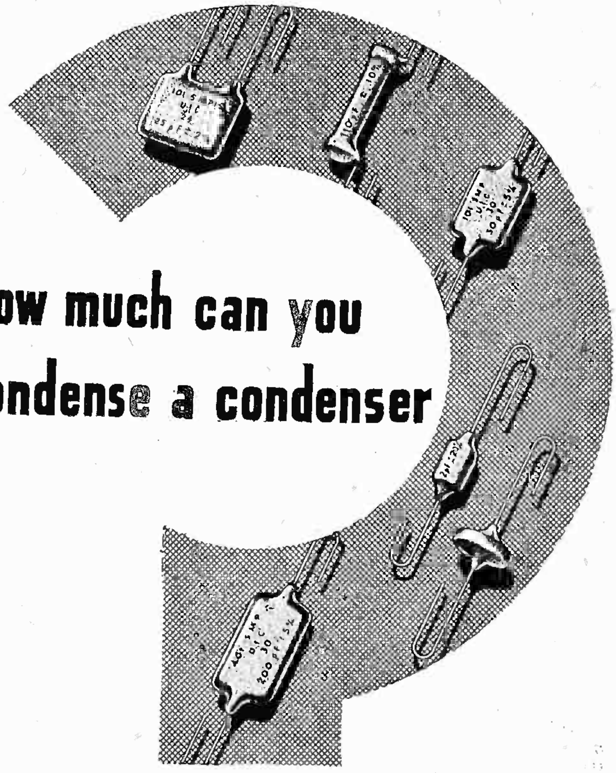

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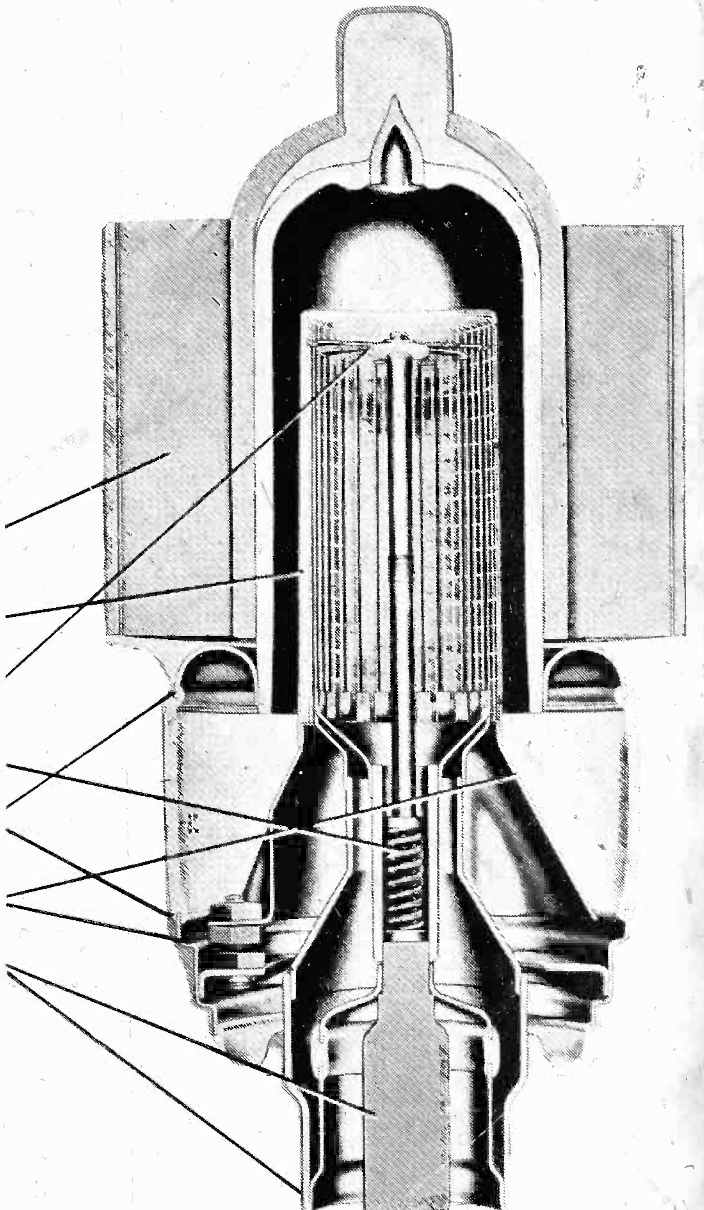

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Printed in England for the Publishers, Iliffe & Sons Ltd., Dorset House, Stamford Street, London, S.E.1, by The Cornwall Press Ltd., Paris Garden, Stamford Street, London, S.E.1.

VOL. 23 No. 271

WIRELESS ENGINEER

APRIL 1946

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EDITORIAL

The Effective Length of a Half-wave Dipole

A HALF-WAVE dipole is usually assumed as an approximation to have a sinusoidal current distribution, and therefore to have an effective length equal to its actual length multiplied by $2/\pi$. This assumes that the field radiated by an aerial of this effective length carrying a current at every point equal to that at the centre of the half-wave dipole would be the same as that radiated by the dipole. This is true of the field strength in the equatorial plane—the direction in which we are often interested—but it is not strictly true in other directions. In the equatorial plane the radiation from all parts of the aerial arrives in phase and can be simply added; we then have

$$\hat{\mathcal{E}} = \hat{H} = \frac{4\pi}{10} \cdot \hat{\imath}\frac{h}{\lambda} \cdot \frac{1}{r}$$

where h is the effective height of the earthed quarter-wave aerial or the effective length of half the dipole and r the distance in centimetres. The current i is in amperes. Putting $h = \frac{\lambda}{4} \times \frac{2}{\pi} = \lambda/2\pi$ we have $\hat{\mathcal{E}} = \hat{H} = \frac{2\hat{\imath}}{10r}$.

It is interesting to note that this is exactly the same formula as that for the strength of the ordinary magnetic field H at a distance r from an infinitely long conductor carrying the same current i amperes.

As one departs from the equatorial plane, however, the equivalence of the actual dipole and the aerial λ/π long, carrying a uniform current is no longer exact. This is due to the effects of phase displacement. Fig. 1 shows the equivalent to the half-wave dipole; its length is λ/π. In the direction shown the radiated field due to the element ds at a distance s from the centre is out of phase with that from the centre by an angle $(2\pi s/\lambda) \cos\theta$. Its effective component, *i.e.*, the component in phase with that from the centre is $\cos[(2\pi s/\lambda)\cos\theta]$, and the mean value of this from $s = -\lambda/2\pi$ to $+\lambda/2\pi$ is easily seen to be $\frac{\sin(\cos\theta)}{\cos\theta}$. Hence the strength of the field at a distance r in this direction is not simply $\frac{2i}{10r}\sin\theta$ but $\frac{2i}{10r}\sin\theta \times \frac{\sin(\cos\theta)}{\cos\theta}$. The values of $\tan\theta \,.\, \sin(\cos\theta)$ are given in Table I.

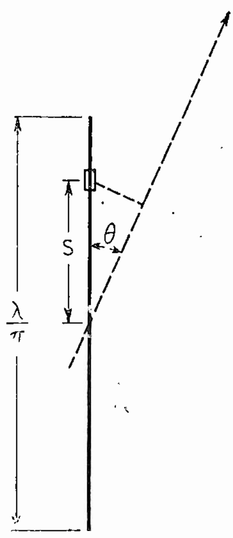

Fig. 1.

In the Editorial of April, 1945, we showed that for the sinusoidal distribution of current along the dipole the field at a distance r was given by the formula $\frac{2i}{10r} \cdot \frac{\cos[(\pi/2)\cos\theta]}{\sin\theta}$. The values for this are given in Table II and it will be seen that as the result of the greater phase displacement, due to the longer length of aerial, the values are lower than those of the equivalent aerial. The difference is not very great, the ratio of the fields increasing from 1 per cent. for $\theta = 70°$ to about 5 or 6 per cent. for $\theta = 20°$.

It can easily be shown that as θ approaches zero, the ratio approaches $(4/\pi)$ sin 1 radian, *i.e.*, 1.071, but the values of the ratio for small values of θ are of little interest because

TABLE I

θ	tan θ	Cos θ		sin (cos θ)	tan θ . sin (cos θ)
		radian	degrees		
10	0.176	0.985	56.4	0.833	0.147
20	0.364	0.94	53.8	0.807	0.294
30	0.577	0.866	49.6	0.76	0.438
40	0.839	0.766	43.8	0.69	0.579
50	1.19	0.643	36.8	0.60	0.714
60	1.73	0.5	28.6	0.48	0.831
70	2.75	0.342	19.6	0.335	0.922
80	5.67	0.174	9.96	0.173	0.981

TABLE II

θ	sin θ	$\pi/2$ cos θ	cos ($\pi/2$ cos θ)	$\dfrac{\cos(\pi/2 \cos\theta)}{\sin\theta}$
10	0.174	88.6	0.0236	0.14
20	0.342	84.6	0.0947	0.28
30	0.5	77.9	0.21	0.42
40	0.643	68.9	0.36	0.56
50	0.766	57.9	0.531	0.693
60	0.866	45.0	0.707	0.816
70	0.94	30.8	0.859	0.913
80	0.985	15.7	0.963	0.978

little power is radiated in these directions. In the directions in which the greater part of the power is radiated the ratio varies from unity to about 1.03. The result of this is that, although the equivalent aerial gives the same field in the equatorial plane, its radiated power exceeds that of the half-wave dipole in the ratio of 80 to 73.2, these being the values of the radiation resistance which we obtained in the Editorial of April, 1945.

G. W. O. H.

CARRIER-FREQUENCY AMPLIFIERS*

Transient Conditions with Frequency Modulation

By C. C. Eaglesfield

(Mullard Radio Valve Company)

SUMMARY.—A treatment is given of the problem of the transient response of a carrier frequency amplifier handling a signal which is frequency modulated. It is found that under certain conditions there is a simple relationship between the responses for amplitude and frequency modulation. Curves are given for staggered circuits, for various numbers of circuits and different amounts of stagger. These curves are also applicable to coupled circuits.

CONTENTS

1. Introduction

THE writer has previously[2,3] dealt with the transient response of carrier frequency amplifiers to a sudden change in the amplitude of the carrier. The problem discussed here is the transient response to a sudden change in the frequency of the carrier. That is to say, the transient response of a frequency-modulation system is now considered, in place of the amplitude-modulation system previously considered.

The " amplifier " will either be a definite network of resistances, capacitances, etc., or else an unspecified algebraic function of the differential operator d/dt. It will not be defined by steady-state amplitude and phase characteristics, since it is not possible to assign to such characteristics shapes which are mutually compatible and yet simple in form.

Very little work seems to have been published on the transient response of carrier systems for either amplitude or frequency modulation, except in terms of arbitrary steady-state characteristics. Bell[1] has dealt with a single tuned circuit with frequency modulation. The present work extends the

* MS. accepted by the Editor, October 1945.

treatment of frequency modulation to any number of tuned circuits, with a reasonable flexibility of tuning adjustment.

With the extension to include frequency modulation, the writer has found it advisable to consider more closely the definitions and terms to be used. In the present work the term " Transient Modulation Ratio", usually shortened to " Modulation Ratio ", is used. This is the ratio of the modulation, however defined, at the output, to the modulation at the input, similarly defined. The definition of modulation depends on the type of system, and must in general be statistical. The necessity for a statistical definition is more apparent for frequency modulation than for amplitude modulation, and for the output than the input. While this is the idea underlying the term " Modulation Ratio", the treatment to be given is somewhat simpler. An input waveform containing an instantaneous change of amplitude or frequency is injected into the amplifier; the modulation of this simple waveform is apparent. The resulting output current is then found, and its effective amplitude or frequency evaluated, using a statistical definition. Then by comparing the variation of amplitude (or frequency) in the output with the variation in the input, the modulation ratio is obtained.

It is found that with frequency modulation, an important simplification results from restricting the modulation of the input to small values. This had previously been found to be so with amplitude modulation[3].

If also the carrier frequency coincides with the central frequency of the amplitude characteristic of the amplifier, this characteristic being symmetrical about the central frequency, there is a simple relation between the modulation ratios for the two systems, amplitude and frequency modulation. (The modulation being small for both systems).

The above statement needs qualifying: it will not be proved as such, though it may well be true. But it is found to be so for an important particular case.

The particular case referred to is that of " staggered " circuits, that is to say circuits tuned to adjacent frequencies in pairs. This arrangement has an amplitude characteristic which is symmetrical about a central frequency, and when the carrier frequency is made equal to the central frequency, then the simple relationship holds between the modulation ratios for amplitude and frequency modulation.

Since double (i.e., coupled) circuits are essentially similar to staggered circuits, the simple relationship also holds for double circuits.

In Section 2 the general equations are found for an arbitrary admittance, and the relationship is shown between the modulation ratios for amplitude and frequency modulation.

In Section 3 these equations are applied to staggered circuits.

In Section 4 numerical results are given for the modulation ratio (frequency modulation) for various numbers of staggered circuits, with different amounts of stagger. It is observed that for a given amount of stagger, the greatest slope of the modulation ratio for frequency modulation is nearly twice the greatest slope for amplitude modulation, and that subsidiary oscillations are much less. Thus if the amount of stagger is adjusted for each system so that the amount of subsidiary oscillation is roughly the same, the greatest slope for frequency modulation is probably rather more than twice that for amplitude modulation.

This comparison is for the same steady-state gain with each system.

However, it should not be forgotten in comparing the two systems, that an amplifier for frequency modulation will usually include an amplitude limiter which will sacrifice some amplification.

2. General Equations for Transient Modulation Ratio

Consider an admittance subjected to a voltage which for time t negative is sinusoidal in form with angular frequency ω_0, and for t positive has an angular frequency ω_0'. That is to say, at $t = 0$ the voltage has a sudden change of angular frequency from ω_0 to ω_0'.

The problem is to find the frequency of the current in the admittance. For t negative the current will clearly have an angular frequency ω_0, and for large positive values of t the current can be expected to have an angular frequency ω_0'.

We can thus expect a transition period during which the angular frequency of the current changes from ω_0 to ω_0'. During this transition period the form of the current will not be a pure sinusoid. It is therefore necessary to define an " effective frequency " for an arbitrary waveform. This definition must be based on statistical quantities.

The voltage waveform can be expressed operationally (in the Heaviside notation) by

$$\cos \omega_0 t + (\cos \omega_0' t - \cos \omega_0 t)\, \mathbf{1} \qquad (1)$$

since equation (1) represents a function which is equal to $\cos \omega_0 t$ for negative time and $\cos \omega_0' t$ for positive time. This follows since the Heaviside step function $\mathbf{1}$ is zero for negative time and unity for positive time. It is convenient to rewrite the expression (1) as

$$e^{j\omega_0 t}(1 - \mathbf{1}) + e^{j\omega_0' t}\, \mathbf{1} \quad \ldots \quad \ldots \quad (2)$$

and use the real part of (2).

The admittance can be written as $A(p)$, that is, an unspecified function of the operator $p(= d/dt)$.

The current in the admittance is then

$$A(p)[e^{j\omega_0 t}(1 - \mathbf{1}) + e^{j\omega_0' t}\, \mathbf{1}]$$

Now apply the Shift Theorem to change this to:—

$$e^{j\omega_0 t}[A(j\omega_0) - A(p + j\omega_0)\, \mathbf{1}] + e^{j\omega_0' t} A(p + j\omega_0')\, \mathbf{1}$$

$$= A(j\omega_0)\left[e^{j\omega_0 t}\left\{1 - \frac{A(p + j\omega_0)}{A(j\omega_0)}\, \mathbf{1}\right\} + e^{j\omega_0' t} \frac{A(p + j\omega_0')}{A(j\omega_0)}\, \mathbf{1}\right] \quad \ldots \quad (3)$$

The factor $A(j\omega_0)$, which is independent of time, is conveniently omitted.

Write
$$\frac{A(p + j\omega_0)}{A(j\omega_0)}\, \mathbf{1} = P(t) + jQ(t) \qquad (4a)$$

$$\frac{A(p + j\omega_0')}{A(j\omega_0)}\, \mathbf{1} = P_1(t) + jQ_1(t) \quad \ldots \quad \ldots \quad (4b)$$

It will be convenient later to restrict the discussion to cases where ω_0' is very nearly equal to ω_0, so that $P_1(t)$ may be assumed equal to $P(t)$ and $Q_1(t)$ to $Q(t)$; also to cases where $Q(t)$ is zero.

Write $\quad P(t) = P \quad$ etc.

Expression (3) then becomes:—

$$e^{j\omega_0 t}(1 - P - jQ) + e^{j\omega_0' t}(P_1 + jQ_1)$$

The real part of this is the current in the admittance:—

$$(1 - P)\cos \omega_0 t + Q \sin \omega_0 t + P_1 \cos \omega_0' t - Q_1 \sin \omega_0' t \quad \ldots \quad \ldots \quad (5)$$

We now require the "effective frequency" of expression (5). The effective angular frequency of a current I will be defined as:—

$$\frac{\text{Average of } dI/dt}{\text{Average of } I}$$

the averages being root-mean-square. In applying this to expression (5) it will be assumed that a duration of time can be chosen for taking the averages, which is long compared to $1/\omega_0$ and $1/\omega_0'$, but during which P etc. are sensibly constant. This amounts to neglecting the derivatives of P etc.

The effective angular frequency ω of expression (5) is given by:—

$$\omega^2 = \frac{[(1 - P)^2 + Q^2]\,\omega_0^2 + (P_1^2 + Q_1^2)\,\omega_0'^2}{(1 - P)^2 + Q^2 + P_1^2 + Q_1^2} \quad \ldots \quad \ldots \quad (6)$$

If in the equation (4a) p is put equal to zero, then for t large:—

$$P + jQ = 1$$

i.e. $\quad P = 1$ and $Q = 0$

Substituting these values of P and Q in equation (6) makes $\omega = \omega_0'$ for t large. For t negative, $P = Q = P_1 = Q_1 = 0$. Thus $\omega = \omega_0$ for t negative.

Equation (6) thus describes a transition in ω from ω_0 to ω_0'.

Now assume that ω_0' is nearly equal to ω_0. Equation (6) becomes:—

$$\omega = \omega_0 + \frac{P_1^2 + Q_1^2}{(1 - P)^2 + Q^2 + P_1^2 + Q_1^2}(\omega_0' - \omega_0)$$

The modulation ratio for frequency modulation is thus

$$\frac{P_1^2 + Q_1^2}{(1 - P)^2 + Q^2 + P_1^2 + Q_1^2}$$

But if ω_0' is nearly equal to ω_0, P_1 will be nearly equal to P, and Q_1 to Q. Assuming $P_1 = P$ and $Q_1 = Q$, the modulation ratio is

$$\frac{P^2 + Q^2}{(1 - P)^2 + P^2 + 2Q^2} \quad \ldots \quad \ldots \quad (7)$$

If further, $Q = 0$, the ratio is

$$\frac{P^2}{(1 - P)^2 + P^2} \quad \ldots \quad \ldots \quad \ldots \quad (8)$$

The expressions (7) and (8) are the small modulation expressions.

The frequency transition has now been expressed in terms of the functions $P(t)$ etc. The problem of amplitude modulation can also be expressed in terms of these functions.

We now consider the same admittance subjected to a voltage which at $t = 0$ undergoes a sudden change of amplitude, being sinusoidal before and after the change with the same angular frequency ω_0.

If the amplitude is 1 before the change and $(1 + m)$ after the change, the voltage can be written as

$$e^{j\omega_0 t}(1 + m\,\mathbf{1}) \quad \ldots \quad \ldots \quad \ldots \quad (9)$$

the real part to be taken.

The current is then :—

$$A(p)e^{j\omega_0 t}(1 + m\,\mathbf{1})$$

$$= e^{j\omega_0 t}\Big[A(j\omega_0) + mA(p + j\omega_0)\,\mathbf{1}\Big]$$

$$= A(j\omega_0)\,e^{j\omega_0 t}\left[1 + m\,\frac{A(p + j\omega_0)}{A(j\omega_0)}\,\mathbf{1}\right]$$

and by equation 4(a), this can be written :—

$$A(j\omega_0)e^{j\omega_0 t}\Big[1 + mP + jmQ\Big] \quad \ldots \quad (10)$$

The factor $A(j\omega_0)$, which is independent of time, is conveniently omitted. Doing this and taking the real part of (10), we get for the current

$$(1 + mP)\cos\omega_0 t - mQ\sin\omega_0 t \quad \ldots \quad (11)$$

To obtain the amplitude of (11), it is necessary first to define an effective amplitude for an arbitrary waveform. Such a definition must be statistical. The effective amplitude of a current I will be defined as

$$\text{(Average of } I)$$

the average being root-mean-square.

In applying this to (11), it will be assumed that a duration of time can be chosen, long compared to $1/\omega_0$ but during which P and Q are sensibly constant.

The effective amplitude of expression (11) is :—

$$\sqrt{(1 + mP)^2 + (mQ)^2} \quad \ldots \quad \ldots \quad (12)$$

The expression (12) can be written

$$1 + m\left[\frac{\sqrt{(1 + mP)^2 + (mQ)^2}}{m} - \frac{1}{m}\right]$$

so that the modulation ratio is :—

$$\frac{\sqrt{(1 + mP)^2 + (mQ)^2}}{m} - \frac{1}{m} \quad \ldots \quad (13)$$

The modulation ratio is thus in general dependent on m; it becomes independent of m if m is either very large or very small :—

$$\sqrt{P^2 + Q^2} \text{ (for } m \text{ very large)} \quad \ldots \quad (14a)$$

$$P \quad \text{(for } m \text{ very small)} \quad \ldots \quad (14b)$$

If $Q = 0$, it is P in both cases.

Comparing the modulation ratio for frequency and amplitude modulation, it is clear that while both are generally expressible in terms of P and Q, the one is given by the other only when $Q = 0$.

We then have, equations (8) and (14),

$$\text{Modulation Ratio} = \frac{P^2}{(1-P)^2 + P^2} \text{ (F.M.)}$$

where P is the modulation ratio (A.M.).

Thus if the modulation ratio (A.M.) is known numerically, it is a simple matter to obtain the modulation ratio (F.M.).

For this simple relation to hold, it is necessary that the function Q be zero, the depth of frequency modulation be very small, and the depth of amplitude modulation be either very small or very large.

In the next section these results will be applied to the particular case of staggered circuits.

3. Application to Staggered Circuits

The modulation ratio for amplitude modulation has already been given [3] for the case of an amplifier chain with "staggered circuits." The reader is referred to the original reference for the full derivation, but for the sake of completeness a brief summary is included here.

Consider an amplifier chain consisting of n stages of which Fig. 1 is a typical stage.

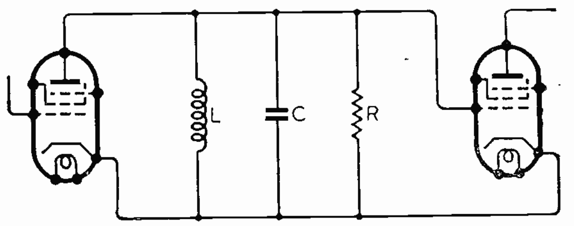

Fig. 1. *A typical amplifier stage.*

For each stage C and R are the same, but in alternate stages L has one of two values, so that one set of circuits is resonant at one frequency and the other set, equal in number, at a second frequency. n is an even integer. Let the two resonant angular frequencies be $(\omega_0 + \omega_1)$ and $(\omega_0 + \omega_2)$ where ω_0 is the injected angular frequency as in Section 2.

Write
$$a = 1/2CR$$
$$k = \frac{\omega_1 - \omega_2}{2a}$$
$$h = \frac{\omega_1 + \omega_2}{2a}$$
$$z = kat$$
$$N = n/2$$
$$\nu = N - \tfrac{1}{2}$$

k is the "stagger" coefficient, and h the "detune" coefficient.

By letting ω_0 tend to infinity, while ω_1 and ω_2 are kept constant, the expressions for the functions P and Q are obtained in the following form (see equation (15) of reference[3]) :—

$$\begin{matrix} P \\ Q \end{matrix} = A \int_0^z e^{N-z/k} \begin{matrix} \cos \\ \sin \end{matrix} \left(\frac{h}{k} z - N\theta \right) \left(\frac{z}{N} \right)^\nu J_\nu(z)\, dz \quad \ldots \quad \ldots \quad (15)$$

In this equation $J_\nu(z)$ is the Bessel function of the first kind of order ν and argument z, and A and θ are given by

$$A = \left[\frac{(1 + k^2 - h^2)^2 + 4h^2}{4k^4} \right]^{N/2} \frac{\sqrt{2\pi N}\, N^N e^{-N}}{N\,!} \quad \ldots \quad \ldots \quad (16)$$

$$\tan \theta = \frac{2h}{1 + k^2 - h^2} \quad \ldots \quad \ldots \quad (17)$$

The problem of finding the modulation ratio with amplitude and frequency modulation for staggered circuits is thus formally solved.

Now consider the case of $h = 0$. This makes $\theta = 0$, and $Q = 0$. Equation (15) becomes

$$P = A \int_0^z e^{N-z/k} \left(\frac{z}{N} \right)^\nu J_\nu(z)\, dz \quad \ldots \quad (18)$$

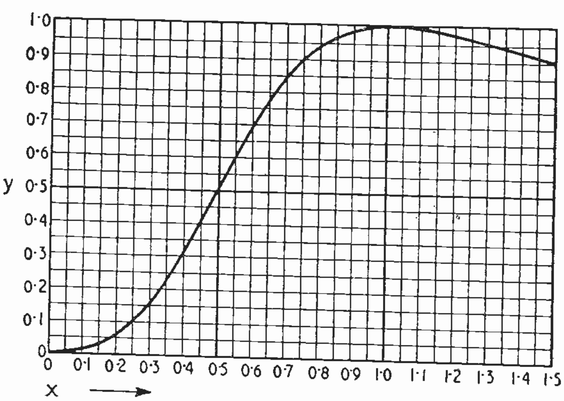

Fig. 2. *The function* $y = \dfrac{x^2}{(1 - x)^2 + x^2}$

Since $Q = 0$, the modulation ratio for amplitude modulation (small modulation) is given by P; and for frequency modulation (small modulation) by

$$\frac{P^2}{(1 - P)^2 + P^2}$$

The condition $h = 0$ implies that the "detune" ratio is zero: that is, that the carrier frequency coincides with the central frequency of the amplitude-frequency characteristic of the amplifier chain.

4. Numerical Results for Staggered Circuits

A number of numerical results have already been given[2,3] for the modulation ratio (A.M.) of staggered circuits. To convert these results into the modulation ratio (F.M.), we require the function.

$$y = \frac{x^2}{(1 - x)^2 + x^2} \quad \ldots \quad \ldots \quad (19)$$

x being the modulation ratio (A.M.), and y the modulation ratio (F.M.). A curve of this function is shown in Fig. 2 (for numerical convenience a table was constructed);

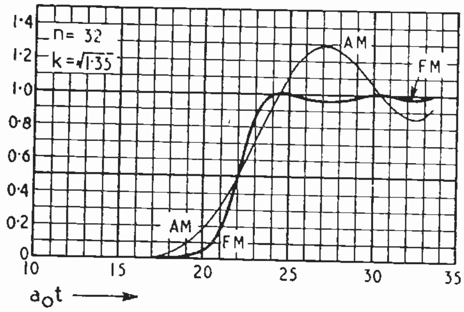

Fig. 3. *The modulation ratio for amplitude and frequency modulation for a particular case of staggered circuits.*

y always lies between 0 and 1; $dy/dx = 2$ for $x = 1/2$; and y droops rather slowly from 1 for x greater than 1.

Fig. 3 shows both the modulation ratio (A.M.) and the modulation ratio (F.M.) for a particular case ($n = 32$, $k = 1.35$). It will be seen that although the modulation ratio (A.M.) is very oscillatory, the restrictive action of (19) makes the modulation ratio (F.M.) much less so, and the maximum slope is greater.

If the modulation ratio (A.M.) had had its greatest slope when its value was 1/2, the greatest slope (F.M.) would have been twice the greatest slope (A.M.). In all the examples given, the greatest slope (F.M.) is somewhat less than twice the greatest slope (A.M.).

Fig. 4 shows the modulation ratio (F.M.) for a number of values of n and k. On the curves the greatest slope and the stationary values are marked. Also with each curve a sketch, which is not to scale, of the corresponding amplitude-frequency characteristic

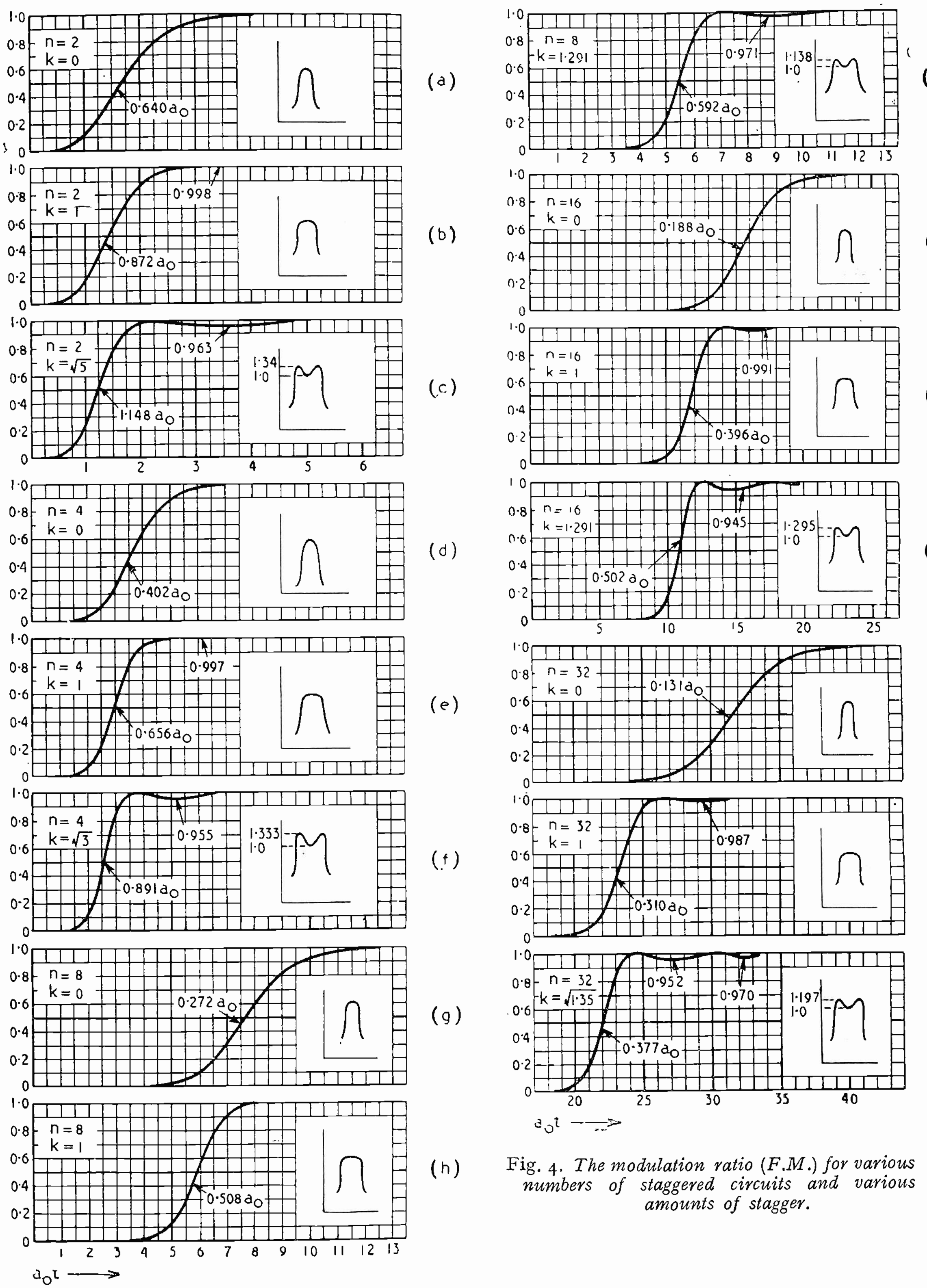

Fig. 4. *The modulation ratio (F.M.) for various numbers of staggered circuits and various amounts of stagger.*

is given. This is useful for visualizing the amount of stagger.

The time scale used in Fig. 4 is a_0t. a_0 is the value of a corresponding to $k = 0$, with the condition that the steady-state gain is independent of k. Thus for each value of n the steady-state gain is the same for the three values of k.

It will be seen that the most noticeable characteristic of the modulation ratio (F.M.) is that it "bounces" against the final value. It cannot go beyond it. It should be remembered, however, that the curves given are for a small modulation depth and for the carrier frequency at the centre of the amplitude - frequency characteristic. For more complicated cases the modulation ratio would probably look rather different.

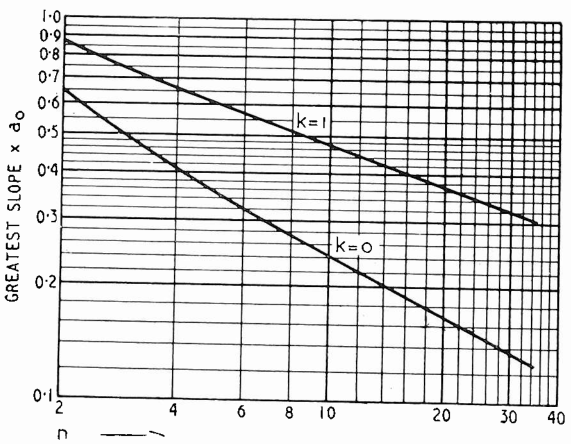

Fig. 5. *The greatest slope of the modulation ratio (F.M.) for various numbers of circuits.*

Fig. 5 shows the greatest slope against n for $k = 0$ and 1.

With slight modification, the curves can be used for double circuits (perhaps more commonly called coupled circuits). These have been dealt with previously[2]. Thus Fig. 4(l), which is drawn for a sixteen-stage amplifier with staggered single circuits, applies to an eight-stage amplifier with double circuits, if the time scale is altered. The rule is to replace a_0 by $2a_0k\sqrt{(1 + k^2)}$.

This makes the eight-stage double-circuit amplifier have the same gain as the eight-stage single-circuit amplifier. The method however breaks down for $k = 0$.

Acknowledgements

The writer is conscious of a debt to Mr. D. P. Dalzell for some general discussion of the matter in this paper as well as the two previous papers on the transient state of carrier frequency amplifiers, in which acknowledgement on specific points was made.

He is grateful to several assistants for aid at different times with the numerical work.

And finally he is grateful to the Directors of The Mullard Radio Valve Company for permission to publish.

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[1] Bell, "Transient Response in Frequency Modulation," *Philosophical Magazine*, Vol. 35, No. 242, p. 143, March 1944.

[2] Eaglesfield, "The Unit Step Response of Carrier Frequency Amplifiers with Single and Double Circuits," *Wireless Engineer*, Vol. XXII, No. 266, p. 523, November 1945.

[3] Eaglesfield, "Carrier Frequency Amplifiers, Transient Response with De-tuned Carrier," *Wireless Engineer*, Vol. XXIII, No. 270, p. 67, March 1946.

BOOK REVIEW

Systems of Electric and Magnetic Units. By Jean Flambouriaris, pp. 120.

This book, published in Athens, is in Greek. The author is the Chief Engineer of the Greek Ministry of Transport and he is obviously well versed in the subject of units. He acknowledges his indebtedness to la Revue Générale and to the Bulletin of the Société Française des Electriciens. The subject of units and dimensions and the choice of fundamental units are dealt with in the first chapter. The second chapter is devoted to mechanical units and the various systems employed. The final chapter deals with electric and magnetic units, the c.g.s. systems, the Giorgi system, rationalization, the Gaussian and Lorentz systems. The book concludes with a comprehensive index and a large table showing the dimensions of all the units in the various systems. G.W.O.H.

Books Received

Television Today and Tomorrow

By Lee de Forest, Ph.D., D.Sc., D.Eng. A popular account of the history of television with an elementary technical description of present day methods. Pp. 176. Hutchinson's Scientific & Technical Publications Ltd., 47, Princes Gate, London, S.W.7. Price 16s.

The Decibel Notation

By Vepa V. Lakshmana Rao, B.E. (Elect.). A detailed explanation of the decibel and the phon and their applications. Conversion tables are included. Pp. 179 + xvi. Addison & Co. Ltd., Madras, India. Price 13s. 6d.

Atomic Spectra

By R. C. Johnson, M.A., D.Sc. This book covers the theory of atomic spectra including the effects of electric and magnetic fields external to the atom. Pp. 120 + viii. Methuen & Co. Ltd., 36, Essex Street, Strand, London, W.C.2. Price 5s.

GRAPHICAL SYMBOLS FOR FILTERS AND CORRECTING NETWORKS*

By G. H. Foot

Introduction

THE present British Standard Symbols† for Filters and Correcting Networks are unsatisfactory because they are limited to a few simple types. Moreover, there is no rational manner in which they can be modified to illustrate new developments or special facilities. As they are also not particularly easy to draw or to understand, for they can be confused with the British Standard Symbols for alternating currents, and as they fail to give much information which is desirable, it is suggested that they should be replaced by the comprehensive system to be described.

The Valve Symbol

It will be useful to consider the reasons for the success of the conventional method of representing a thermionic valve. Four lines are used to show the three electrodes of a triode and its envelope. The nature and function of each electrode is suggested by the type of line, and the circle surrounding them indicates the boundary of the evacuated space. The external connections are shown by taking short lines to this boundary, and these can be numbered to correspond with a standard valve base. The valve characteristics may be specified sufficiently by giving the values of the principal valve parameters. More often the code of the valve is quoted, and this information must be obtained from the maker's catalogue.

Such a symbol has proved to be very satisfactory. In a simple way it gives a great deal of information, and this is presented so that it can be understood by technologists without difficulty. Also, and this is very important, the method of modifying the symbol for more complicated and special types of valves, is obvious. The interpretation of the modified symbols is usually manifest and ambiguity is improbable. The symbolization of screen-grid and pentode valves, and valves with metallization, indirectly-heated cathodes, gas-filling, and variable-mu properties has not been difficult. Many other variants have been treated successfully, including types which are really several valves in one with internal connections and screening. By modifying the shape of the envelope, the symbol for a cathode-ray tube was devised.

All the valve symbols discussed have been found to be satisfactory for the pencil sketches of engineers, yet they are scarcely altered when carefully drawn. The attempts of draughtsmen to make improvements in detail, and the divergences resulting from the independent progress of different countries, have not caused any confusion.

The Rectifier Symbol

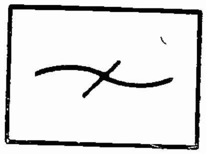
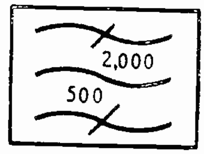
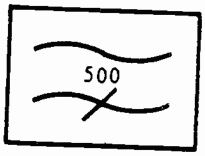
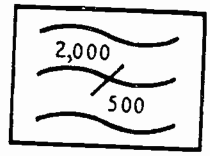
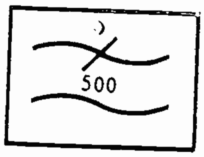
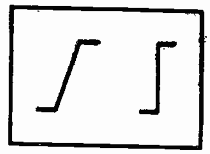
The solution of the problem of the valve symbol is apparently simple and obvious, but in it can be discovered the cardinal principle to be followed in successful electrical symbology. It is necessary that a symbol should enable the apparatus it represents to be identified, and its essential characteristics to be ascertained, without the need for memorizing an arbitrary code. This can be done by basing it on the physical structure of the apparatus. The symbol becomes an elementary picture of the equipment. Sometimes this is not convenient because nothing characteristic of the apparatus is denoted by its construction, or because different arrangements are used to develop the same property.

An example is the rectifier where the symbol diagrammatically represents its performance. When the rectifier is of the multiple type however, the symbol is repeated to indicate the relative positions of the plates. The combination of the electrical and mechanical properties thus portrayed, has been found to present the information which the engineer must have, without including anything unessential.

* MS. accepted by the Editor, November 1945.

† British Standard Graphical Symbols for Telephony, Telegraphy, and Radio Communication. No. 530; 1937.

BRITISH STANDARD SYMBOLS FOR FILTERS AND CORRECTING NETWORKS

Filter—General Symbol.		Filter—Band-Pass. Cut-off frequencies indicated.	
Filter—High-Pass. Cut-off frequency indicated.		Filter—Band Stop. Cut-off frequencies indicated.	
Filter—Low-Pass. Cut-off frequency indicated.		Correcting Network.	

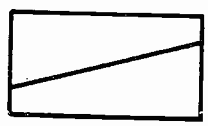
The Basis of the Proposed New Symbols

The shape, size, and mechanical details of a filter (or correcting network) do not determine its electrical behaviour. The value of the filter is assessed after examining its attenuation-frequency characteristic, which is usually given graphically. To an engineer nothing is more strongly associated with a filter than this graph, and nothing more quickly conveys the essential information about performance than a glance at it. It is on this graph that it is suggested the filter symbol should be based. Additional particulars will be necessary, for example, details of the external connections and a description of the construction as far as it affects these. The symbol should therefore include this information.

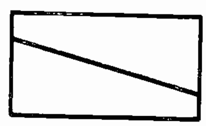
The present British Standard Symbols and the proposed new symbols are shown in the Tables together with some examples to clarify their use. The advantages of the new symbols are most evident in the case of the more complex filters, such as that of Example 4. Filters of this kind are in common use, but the existing symbols are quite inadequate to give the information which is generally required.

Correcting Networks and Amplifiers

The symbol for a correcting network is based on the attenuation-frequency characteristic of the network. This avoids the considerable limitation of having only one symbol for every network, irrespective of the type of correction it provides. Thus networks with more attenuation at high frequencies than at low frequencies are represented by (a) and if the attenuation is less at high frequencies than at low frequencies then the symbol becomes (b)



(a)

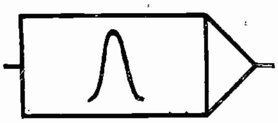


(b)

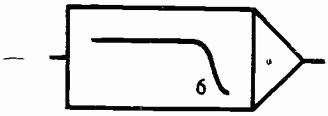
It will be observed that a correcting network can always be distinguished, because the line representing the attenuation goes from side to side of the rectangle. In a filter symbol this line rests on the bottom of the rectangle.

It is usual to obtain the response of radio receiving apparatus as the voltage gain at various frequencies for a constant output. This had led to the custom amongst radio engineers of plotting the gain-frequency characteristic, and this will be a mirror image of the attenuation-frequency characteristic. Such apparatus should, however, be regarded as amplifiers with various types of response curves. They could conveniently be symbolized in the manner shown by sketches (c), (d), and (e), which will be self-explanatory.

(c)



(d)



(e)

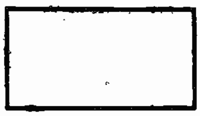
It is not our purpose at the moment to be concerned with this type of symbol, except to show how it could be included in the scheme. It will be sufficiently obvious that the combination of the rectangle with the usual triangular sign removes any possible ambiguity, as well as providing a uniform and valuable method of representing amplifiers comparable with the method for representing filters already discussed.

Many other variations might be described, but not advantageously at the moment. They might easily confuse the reader as to the basis of the proposed plan. Indeed as it is claimed that the new symbols can be developed logically as required, no elaboration is necessary.

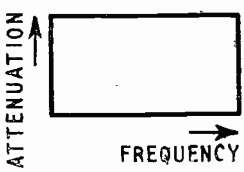
SUGGESTED NEW SYMBOLS FOR FILTERS

Derivation

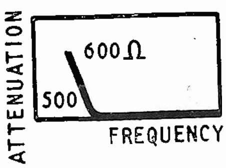
The phyiscal boundary of the filter (or correcting network) is indicated by a rectangle.



The rectangle is considered as a sheet of graph paper with the lower edge as frequency axis, and the left-hand edge as an attenuation axis, the origin being the lower left-hand corner.



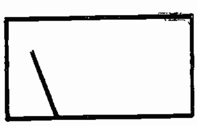
The general form of the attenuation - frequency characteristic is drawn and cut-off frequencies in kc/s are inserted in the acute angle. The characteristic impedance is written over the pass-band.



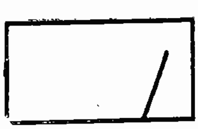
The thickened line in the pass-band shown in the derivation is unnecessary and merges into the boundary of the symbols.

New Symbols

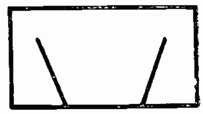
High-Pass.



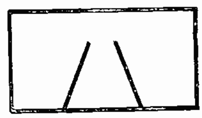
Low-Pass.



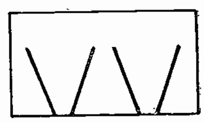
Band-Pass.



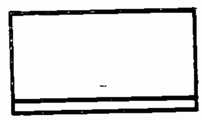
Band-Stop.



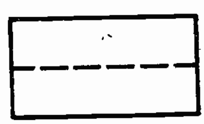
A band-pass filter having two pass-bands which do not overlap.



Unbalance in the filter is shown by a line representing the earth line.

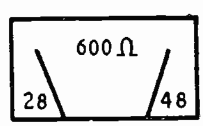


Balance in a filter is shown by a line in the centre—solid for a physical earth connection, otherwise dotted.



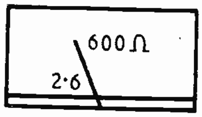
Example 1

A band-pass filter with a pass-band between 28 kc/s and 48 kc/s and with a characteristic impedance of 600 Ω.



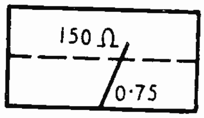
Example 2

An unbalanced high-pass filter with a cut-off frequency of 2.6 kc/s and a characteristic impedance of 600 Ω.



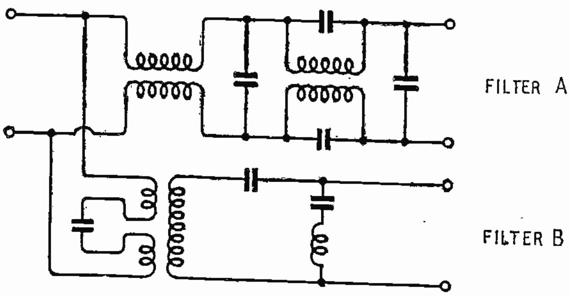
Example 3

A balanced low-pass filter (no physical earth connection) with a cut-off frequency of 750 c/s, and a characteristic impedance of 150 Ω.



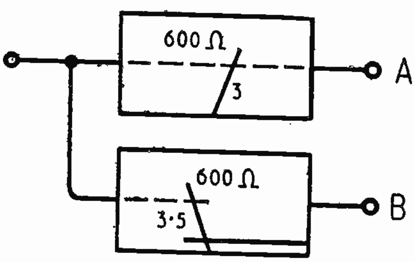
Example 4

The following circuit is of a pair of filters used for separating a voice-frequency telephone circuit from carrier circuits operating on the same cable pair.



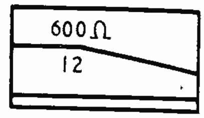
Filter A is a balanced low-pass filter with a cut-off frequency of 3 kc/s, a characteristic impedance of 600 Ω, and no physical earth connection. Filter B is a high-pass filter with a cut-off frequency of 3.5 kc/s and a characteristic impedance of 600 Ω. It has a balanced input where it is connected to the low-pass filter, but one of the coils is used as a transformer and an unbalanced output is provided.

These filters are represented in the new system as follows :—



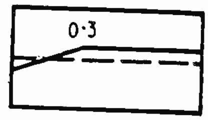
Example 5

A 600 Ω unbalanced correcting network has a constant attenuation to 12 kc/s. The attenuation then decreases as the frequency is raised.



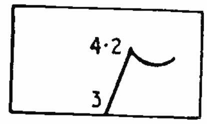
Example 6

A correcting network of the balanced type (no physical earth connection) has a constant attenuation above 300 c/s, but below this frequency the attenuation decreases as the frequency is lowered.



Example 7

A low-pass filter has a cut-off frequency at 3 kc/s, and a point of infinite (i.e. high) attenuation at 4.2 kc/s.



Summary.—As the British Standard Symbols for Filters and Correcting Networks are inadequate, new symbols are proposed. It is shown how the scheme can be extended to include amplifiers.

The advantages of the new symbols are :

(1) They are based on performance graphs which display information which an engineer requires, and which he associates closely with the apparatus itself.

(2) The understanding of the symbols does not depend on memorizing an arbitrary code.

(3) They give information often required, but for which no provision is made in the British Standard Symbols.

(4) The elaboration of a symbol to illustrate a new development or a special feature, is a natural and logical process. An error in the interpretation of a modified symbol is improbable.

(5) The new symbols are very simple, and confusion with other symbols is impossible.

(6) The proposed scheme is flexible, and not likely to be inadequate for future requirements.

SOME EXPERIMENTS WITH LINEAR AERIALS*

By J. S. McPetrie, D.Sc., Ph.D., M.I.E.E., and J. A. Saxton, Ph.D., B.Sc., A.M.I.E.E.

(*Communication from The National Physical Laboratory*)

SUMMARY.—The first part of this paper describes a method by which the polar diagram of any rectilinear aerial can be determined. The method is based on the assumption that elementary current waves are induced in the aerial by an incident electromagnetic field, the final current in the aerial being the summation at each point of the effect of these current waves after reflection at each end of the aerial. If the lengths of the aerial on either side of the point of connection of the aerial to the receiver are equal or differ by an integral number of half wavelengths, the analysis is considerably simplified as the correct polar diagram is obtained if reflection is assumed to occur only once at each end of the aerial. Figures are given in the paper showing close agreement between the theoretical and experimental polar diagrams obtained for aerials of various lengths between one-half and five half-wavelengths. It is also shown experimentally that the polar diagram of any aerial is the same when that aerial is used for transmission or reception.

Part II describes an experimental investigation which has been made of the front-to-back signal ratio of a receiving aerial system consisting of a half-wavelength dipole and single parasitic aerial. When the latter acted as a reflector its optimum length and spacing from the main receiving aerial were approximately 0.5 and 0.1 of the wavelength respectively. For this condition the pick-up in the backward direction was 11 decibels below that in the forward direction. When the parasitic aerial acted as a director, however, the optimum back to front signal ratio was approximately 20 decibels and the length of parasite and spacing from the receiving aerial 0.47 and 0.05 of the wavelength. The overall gain of this aerial system over a simple half-wavelength dipole was also highest when the parasitic aerial was adjusted roughly as a director for optimum back-to-front signal ratio.

PART I

1. Introduction

It has been shown[1] that the current distribution in an aerial may be quite different depending upon whether the aerial is used for transmission or reception. This result suggests the comparison of other properties of transmitting and receiving aerials. A method is described which is suitable for the determination of the polar diagram of a linear receiving aerial. At each point along the aerial two elementary current waves are assumed to be induced by an incident electromagnetic wave. One current wave travels directly towards the point at which the receiver is connected to the aerial and the other wave arrives at the same point after reflection at one end of the aerial. If the difference in the lengths of the aerial on the two sides of the receiver connection is zero or an integral number of half-wavelengths, the correct polar diagram of the aerial is obtained by integrating with proper phase relation the total currents at the receiver connection due to the direct elementary current waves from each point and the indirect waves, the latter after reflection at that end of the aerial at which reflection first takes place. The present paper is limited to a consideration of the cases in which this condition applies.

2. Description of Method

The polar diagram of a receiving aerial represents the variation of input voltage to the receiver with aerial orientation. As the input impedance of the receiver is unaltered by aerial rotation this voltage is proportional to the current in the receiving aerial at the point in the latter from which the aerial output is fed to the receiver. The polar diagram of a receiving aerial, therefore, can be determined completely if the aerial current at this point is known for all orientations of the aerial.

Let AB in Fig. 1 represent a rectilinear aerial and O the point in it at which the aerial output is applied to the receiver either directly or by some coupling device such as a transmission line. If a plane electromagnetic wave of amplitude E is incident on the aerial in the direction making an angle θ with its length, the voltage induced in an

* MS. accepted by the Editor, October 1945.

element of length δx at P is $E \sin \theta \, . \, \delta x$. This voltage developed across the element at P gives rise to two current waves in AB, one travelling in the direction PA and the other in the opposite direction PB. These two elementary current waves undergo successive reflections at each end of the aerial but, in the transient state before these reflections take place, the paths PA and PB appear at P

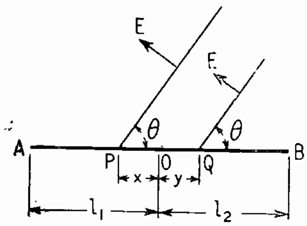


Fig. 1. *Diagrammatic representation of linear aerial.*

as two infinitely long lines each having impedance Z_0 where $2Z_0$ is the input impedance of an infinitely long aerial in free space of the same material and having the same cross-section as the aerial AB.* The equivalent circuit of the aerial measured at P in the transient state can, therefore, be represented by the circuits shown in Figs. 2(*a*) and (*b*). The initial transient current, as seen from Fig. 2(*b*), has amplitude $\dfrac{E \sin \theta \, . \, \delta x}{2Z_0}$. This represents the amplitude of each of the two waves emanating from P.

The final current in the aerial at any point can be obtained by adding the effects of the elementary waves originating at each point along the aerial and their subsequent reflections at each end of the aerial, and any discontinuity at the receiver connection. For the present purpose in which the magitude of the current at O is not required, but only its variation with θ, the orientation of the aerial with respect to the incident wave, it is more convenient to consider the two portions OA and OB of the aerial separately. The current at O, Fig. 1, due to OA is constituted by the vectorial addition at O of all the elementary current waves induced in OA travelling towards O, and all the indirect waves after reflection at the end A. If the magnitude of the current at O due to OA is desired, the subsequent reflections of these current waves at each end of the aerial must be taken into account

For the usual conditions on ultra-short waves, in which the lengths OA and OB are equal, or differ by an integral number of half-wavelengths, the effect of these subsequent reflections at A, B and O can be neglected when only the variation of aerial current with aerial orientation is required. For the determination of the polar diagram of the portion OA of the aerial, therefore, only the current waves induced in OA travelling directly to O and the indirect waves after a single reflection at A will be taken into account in the present paper. The reason that the latter indirect waves must be considered is because the inducing field gives rise at each point to current waves travelling in opposite directions along the aerial, and the phases along OA of the current waves travelling originally in the direction OA are related to incident radiation just as much as the current waves in the opposite direction AO. If the same procedure is followed for the section OB and the currents so obtained for OA and OB added vectorially a formula

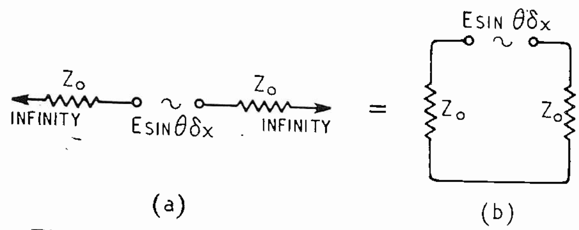


Fig. 2. *Equivalent circuit in transient state of aerial having characteristic impedance Z_0.*

results which, with the limitation as to the position of O in AB mentioned above, gives the correct variation of the current at O with rotation of the aerial and, therefore, the polar diagram of the aerial AB when used for reception. It is shown in an appendix that for an aerial having impedance $2Z_0$, as defined above, the current at a point l_1 and l_2 from each end of the aerial such that

$$l_1 - l_2 = n \, \lambda/2$$

and derived in the above manner from the summation of elementary current waves is given by

$$I = A - jB \qquad \text{(1)}$$

in which A and B are, respectively

$$A = \frac{E\lambda}{2\pi Z_0 \sin \theta} \left\{ \sin \left(\frac{4\pi l_1}{\lambda} \cos^2 \frac{\theta}{2} \right) + \sin \left(\frac{4\pi l_2}{\lambda} \sin^2 \frac{\theta}{2} \right) - \sin \frac{4\pi l_1}{\lambda} \right\} \qquad \text{(2)}$$

* If the aerial AB were equivalent to a transmission line having uniformly distributed constants along its length, $2Z_0$ would represent the characteristic impedance of the aerial. It would also be the input impedance of an infinite length of this transmission line. Schelkunoff[2], however, has shown that for an infinitely long aerial in free space the characteristic and input impedances are not equal.

$$B = \frac{E\lambda}{2\pi Z_0 \sin\theta}\left\{1 - \cos\left(\frac{4\pi l_1}{\lambda}\cos^2\frac{\theta}{2}\right) - \cos\left(\frac{4\pi l_2}{\lambda}\sin^2\frac{\theta}{2}\right) + \cos\frac{4\pi l_1}{\lambda}\right\} \quad \ldots \quad (3)$$

When the receiver is connected to the centre of the aerial the current I (equation 1) is given by

$$I = \frac{E\lambda}{\pi Z_0 \sin\theta}\left\{\cos\left(\frac{2\pi l}{\lambda}\cos\theta\right) - \cos\frac{2\pi l}{\lambda}\right\} \quad \ldots \quad (4)$$

where l is half the length of the aerial.

3. Experimental Procedure

Two methods were used to determine experimentally the polar diagrams of various aerials and thus to test the validity of the formulae given above. One method was followed when only a commercial type of short-wave receiver was available and the second when a field-strength measuring set could be utilized. In each case the transmitting and receiving aerials were arranged horizontally at equal heights above ground of about 1 m and connected respectively to a small oscillator and receiver by means of twisted flexible leads about 2 m in length.

For measurements with the commercial receiver the transmitter was modulated and the audio voltage developed across the telephone terminals of the receiver was observed. This output voltage was maintained constant as the receiving aerial was rotated about a vertical axis by suitably adjusting a piston type of attenuator inserted in the lead between the modulated oscillator and the radiating aerial. In this way the field-strength at the receiver could be varied over a range of 40–50 decibels. As, in this method, the output and, therefore, the input voltage to the receiver is fixed the relation between these two voltages is not required to be known. The necessary adjustment of the attenuator at the transmitter was made by an observer acting under instruction from a second observer at the receiver. This procedure was relatively simple as the distance between transmitting and receiving aerials was only about 50 m.

When a field-strength measuring set was available, however, the oscillator was unmodulated and observation made of the variation of input voltage to the receiver with receiver aerial orientation, no adjustment of transmitter output being required. Because in this latter method, all measurements were made at the receiver it was more convenient than the first method and was followed whenever possible. Various wavelengths between 50 and 100 cm were used in the experiments, but those described in this report were all made on a wavelength 100 cm.

4. Comparison between Theoretical and Experimental Results

Figs. 3 to 8 give at (*a*) the polar diagrams derived from the above analysis for the length of aerial shown in each illustration, while corresponding diagrams obtained experimentally are given at (*b*). The agreement

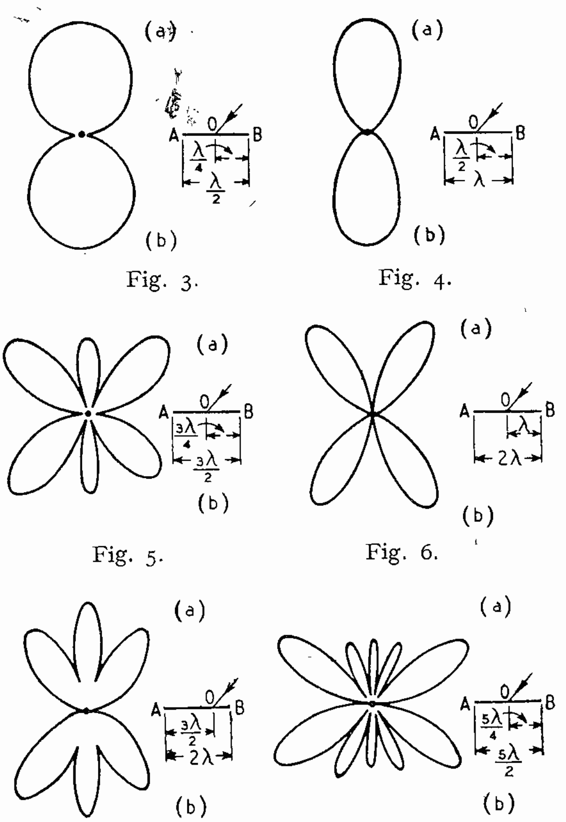
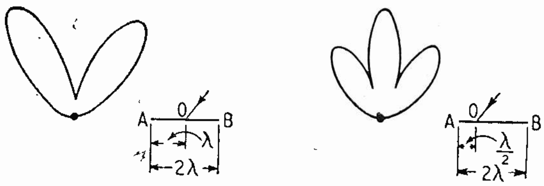


Fig. 3. Fig. 4. Fig. 5. Fig. 6. Fig. 7. Fig. 8.

Figs. 3, 4, 5, 6, 7 and 8. (*a*) *Theoretical and* (*b*) *experimental polar diagrams of receiving aerials of different lengths. AB represents the aerial and O the point of connection to receiver.*

between theory and experiment is seen to be very good. It is particularly interesting to note from Figs. 6 and 7 that in the case of an aerial two wavelengths long the large change in polar diagram with change in position of the connection to the receiver is completely accounted for theoretically.

The analysis outlined in the report, therefore, appears satisfactory for the determination of the polar diagram of any receiving aerial in which the connection to the



Figs. 9 and 10. *Experimental polar diagrams of transmitting aerials. AB represents the aerial, and O the point of connection to transmitter.*

receiver is made at the centre of the aerial or at a point an integral number of half-wavelengths nearer one end than the other. The analysis has been extended to a consideration of rhombic receiving aerials and, again, the agreement between the theoretical and experimental results is good. The polar diagrams of the aerials shown in Figs. 6 and 7 when used as transmitting aerials are given in Figs. 9 and 10. A comparison of Fig. 6 with Fig. 9, and Fig. 7 with Fig. 10 shows that the polar diagram of an aerial is the same whether it is used for transmission or reception. In the determination of the polar diagram of an aerial, therefore, it is largely a matter of personal preference whether the aerial is assumed to be associated with a transmitter or a receiver.

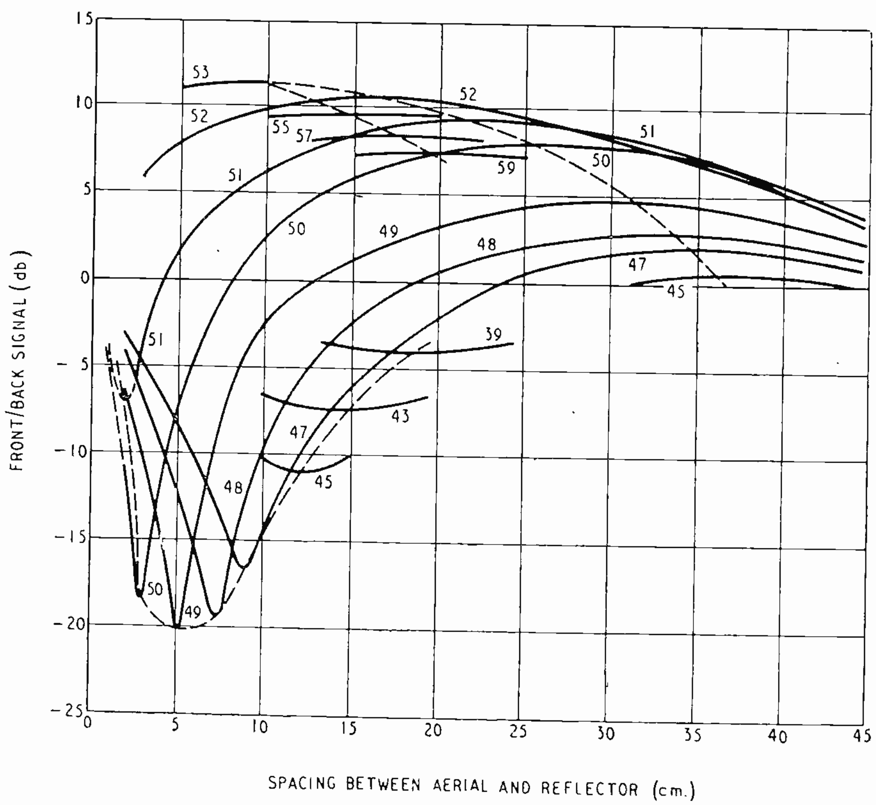


Fig. 11. *Front/back signal with various reflector lengths. λ/2 dipole aerial. λ = 1.05 metres. The numbers on the curves represent reflector lengths.*

Conclusions

The analysis given in the paper is based on the assumption that a receiving aerial of given length is equivalent to an equal length of a uniform transmission line and also that the wavelength of the current waves induced in the wire by external radiation is the same as the wavelength in free space of that radiation. The latter condition is only satisfied if the velocity of propagation of the waves along the aerial is the same as that of light. The close agreement between experiment and theory arrived at on these assumptions shows that such conditions can be considered as exact in most practical cases. Two further conclusions which may be deduced from the work described in this paper are that the polar diagram of any aerial is the same for transmission and reception, and that the correct polar diagram of an aerial can be determined on the assumption that the voltage developed in the aerial is due to discrete current waves induced in the aerial by an incident electromagnetic field.

PART II

1. Introduction

An experimental investigation has been made on certain features of a simple aerial system consisting of a half-wavelength aerial and parasitic reflector. The main purpose of the experiments was to determine the front-to-back signal ratio of such an aerial system and the factors on which this ratio depends.

2. Experimental Procedure and Results

A horizontal electric field with a wavelength of 1.05 m was set up at a receiving site by means of a small screened oscillator connected by a length of about 2 m of twin flexible leads to a horizontal dipole located at a distance of about 50 m from the receiver. The aerial system under investigation was erected at the receiver site, the half-wavelength aerial being placed horizontally and connected to the input terminals of a field-strength measuring set by a vertical length of about 1 m of twin flexible leads. The parasitic aerial was supported at its centre parallel to the main receiving aerial and was constructed of telescopic brass tubing so that its length could be varied. The distance between the main and parasitic aerials could also be varied and the complete receiving aerial system was rotatable about a vertical axis. For various distances between the main and parasitic aerials and for different lengths of the latter, polar diagrams of the receiving aerial system were determined.

Complete diagrams were taken only for a limited number of conditions as the information most required was simply the front-to-back signal ratio. For this purpose the receiver input voltage with the parasitic aerial at a given distance on the side of the main aerial remote from the transmitting aerial was compared with that obtained when the receiving aerial system was rotated about a vertical axis through an angle of 180°.

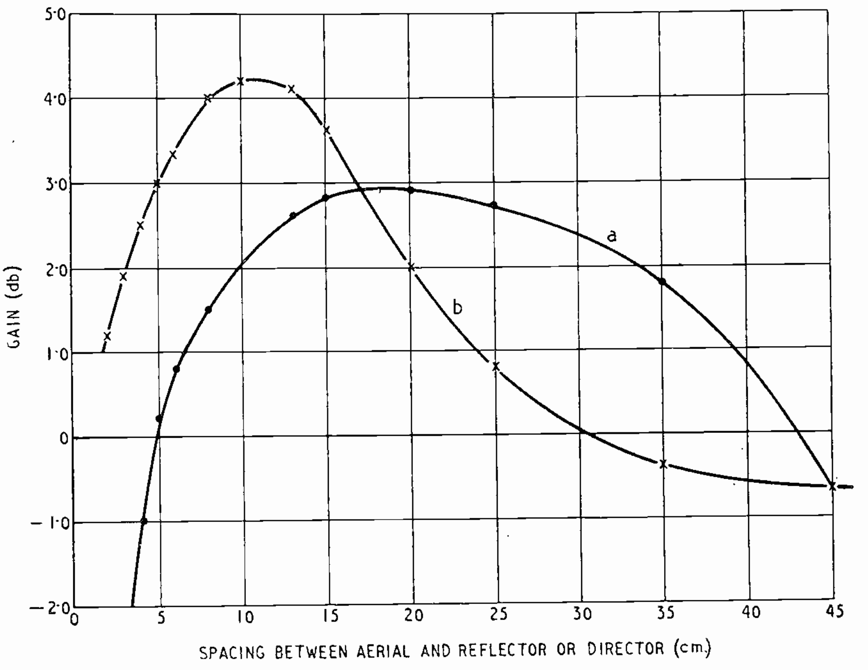


Fig. 12. *Gain of $\lambda/2$ dipole aerial with reflector (a) and director (b) over that of $\lambda/2$ dipole without reflector.*

The results are shown graphically in Fig. 11, in which the ordinates give this ratio of input voltages, termed front-to-back ratio and expressed in decibels, for various spacings between the main and parasitic aerials and for a series of lengths of the latter. The dotted curves in this figure give approximately the optimum front-to-back ratio for any distance.

The front-to-back signal ratio, measured in the direction of the transmitter, can be either positive or negative. When it is positive, the parasitic aerial is on the further side of the receiving aerial, and is known as a reflector. When the ratio is negative the parasitic aerial is on the side of the receiving aerial nearer the transmitter and is known as a director.

It will be seen from Fig. 11 that the maximum positive front-to-back ratio is obtained with a reflector 53 cm in length, which is very nearly equal to half the wavelength (105 cm) used in the experiments;

also the appropriate distance for this maximum condition is 10 cm, or one-tenth of the wavelength. Near the maximum positive front-to-back ratio the magnitude of this ratio does not appear to be critically dependent either on the spacing between the main and parasitic aerials or on the length of the latter. The maximum positive front-to-back ratio of about 11 db is exceeded in absolute value by the negative ratio possible, for example, with the parasitic aerial 49 cm in length, that is, slightly less than a half-wavelength. The curves in Fig. 11 show that the best length of director is 49 cm, or approximately 0.47 of the wavelength, and that the spacing for maximum negative front-to-back ratio is about 5 cm, or one-twentieth of the wavelength. If, therefore, maximum discrimination between two directions 180° apart is required, the parasitic aerial should be used as a director. The magnitude of the front-to-back ratio near the maximum negative value of the latter varies more rapidly with both length of the parasitic aerial and its spacing from the main aerial, than the same ratio near its maximum positive value.

The curves in Fig. 12 show the gain over a single half-wavelength aerial of the aerial system with parasitic aerial adjusted in length and spacing approximately for the maximum positive and negative front-to-back ratios, respectively. The gain with the parasitic aerial acting as a director is seen to be greater than when it is used as a reflector.

The main receiving aerial was connected to the field-strength measuring set by a length of twin flexible leads having a characteristic impedance of about 100 ohms, one end of the leads being inserted directly at the centre of the aerial and no attempt at matching made. The condition for matching was satisfied roughly with the distance between main and parasitic aerials in the region corresponding to maximum positive front-to-back signal ratio. Measurements showed that the matching was far from correct for the small spacing required for optimum negative front-to-back signal ratio. No change in the curves in Fig. 11 would have been obtained if matching had been made between receiving aerial and feeder, but for the small spacings, an in-

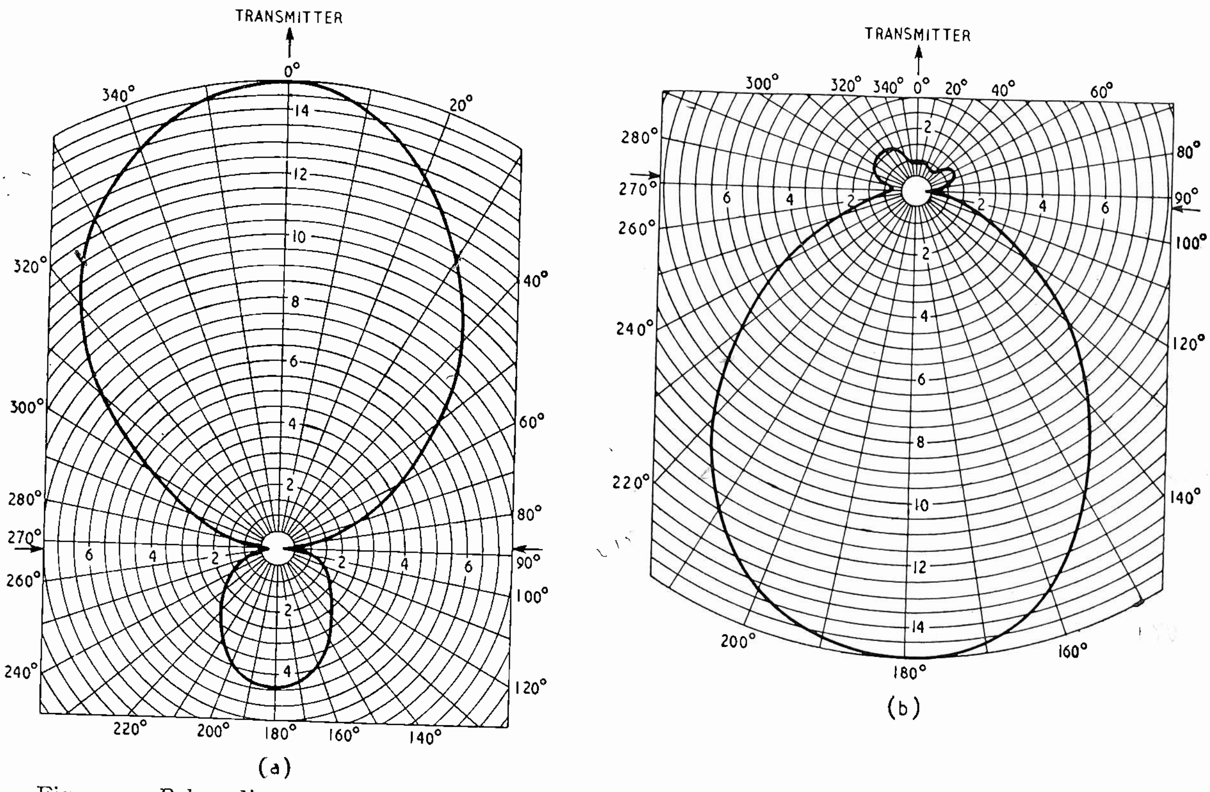


Fig. 13. *Polar diagrams for a dipole with reflector (a) and director (b), at a wavelength of 105 cm. The reflector length was 53 cm spaced 18 cm from the dipole and the gain over the simple dipole was 2.9 db. The director had a length of 49 cm spaced 5 cm from the aerial and had a gain of 2.2 db over a simple dipole.*

crease in gain would have resulted over the curves given in Fig. 12. This increase in gain for such spacing would have made the maximum gain of the aerial system with the parasite acting as a director, greater than that given in Fig. 12 but little, if any, change would have occurred for the greater spacing for optimum reflector conditions. With the parasitic aerial under optimum conditions for a director, therefore, both a better absolute value of front-to-back signal ratio and gain over a half-wavelength dipole would be obtained over those possible with the parasitic aerial acting as a reflector.

In Figs. 13 (*a*) and (*b*) typical polar diagrams are given for the two cases in which the parasitic aerial acts as a reflector and director respectively. The lengths of parasitic aerial correspond to maximum positive and negative front-to-back signal ratios, but the optimum spacing from the main aerial is correct only for the parasitic aerial acting as a director. As mentioned above, however, the change with distance in either polar diagram or front-to-back signal ratio is not critically dependent on spacing near the maximum positive value of that ratio. Incidentally, although the two curves in Figs. 13 (*a*) and (*b*) show an improvement in front-to-back signal ratio for the smaller spacing between main and parasitic aerials, the gains in the two cases over a simple half-wavelength at the receiver are approximately the same.

Acknowledgements

The work described above was conducted as part of the programme of the Radio Research Board to whom confidential reports were circulated during 1941. This paper is published by permission of the Department of Scientific and Industrial Research.

APPENDIX

As shown in Part I of the paper, two elementary current waves each having magnitude $\frac{E \sin \theta \,.\, \delta x}{2 Z_0}$ are induced in the element δx at P (Fig. 1) by the field E incident on the aerial in the direction making an angle θ with its length. These two current waves travel in opposite directions from P. It is convenient to refer all phases to that of the elementary current wave I_0 induced by the field E at the point O in the aerial at which the final aerial current is required.

If λ be the wavelength of the incident radiation the current wave at P lags in phase on that of I_0 by the angle $\frac{2\pi x}{\lambda} \cos \theta$ due to the phase of the field at P lagging on the phase of E at O by the same amount. In travelling directly to O there is an additional retardation in phase of $\frac{2\pi x}{\lambda}$, provided that the wavelength of the current wave in the aerial is the same as that of the incident radiation in free space. The current wave induced at P which travels directly to O lags at O in phase on I_0, therefore, by the angle

$$\frac{2\pi x}{\lambda}(1 + \cos \theta) = \frac{4\pi x}{\lambda} \cos^2 \theta/2.$$

The component of the total current at O due to all the elementary current waves in OA which travel directly towards O is

$$\frac{E \sin \theta}{2Z_0} \int_0^{l_1} exp\left(-j \frac{4\pi x}{\lambda} \cos^2 \theta/2\right) dx$$
$$= j\frac{E\lambda \tan \theta/2}{4\pi Z_0}\left\{exp\left(-j\frac{4\pi l_1}{\lambda}\cos^2 \theta/2\right) - 1\right\} \quad \ldots \quad \ldots \quad (1)$$

The initial current wave at P which travels in the direction PA also has amplitude $\frac{E \sin \theta \, \delta x}{2Z_0}$ and lags in phase on I_0 by the angle $\frac{2\pi x}{\lambda} \cos \theta$. In travelling from P to A and thence to O along the aerial there is an additional phase lag of $\frac{2\pi}{\lambda}(2l_1 - x)$. The reflection coefficient at the end A of the aerial must be such that there is no current at this point. This condition is satisfied if there is π change in phase of the current wave on reflection at A. The current wave from P, therefore, which arrives at O after reflection at A, lags at O in phase on I_0 by the angle

$$\frac{2\pi}{\lambda}(x \cos \theta + 2l_1 - x) + \pi = \frac{4\pi}{\lambda}(l_1 - x \sin^2 \theta/2) + \pi$$

Hence, the component of the total current arriving at O from OA after reflection at A is

$$-\frac{E \sin \theta}{2Z_0}\int_0^{l_1} exp\left\{-j\frac{4\pi}{\lambda}\left(l_1 - x \sin^2 \theta/2\right)\right\} dx$$
$$= j\frac{E\lambda \cot \theta/2}{4\pi Z_0}\left\{exp\left(-j\frac{4\pi l_1}{\lambda}\cos^2 \theta/2\right) - exp\left(-j\frac{4\pi l_1}{\lambda}\right)\right\} \quad \ldots \quad \ldots \quad (2)$$

For a point Q in OB (Fig. 1) distant y from O the elementary current wave leads that at O in phase by the angle $\frac{2\pi y}{\lambda} \cos \theta$. The total phase lag on I_0 of the current wave arriving at O directly from Q is, therefore,

$$\frac{2\pi y}{\lambda}(1 - \cos \theta) = \frac{4\pi y}{\lambda} \sin^2 \theta/2$$

The total current at O from OB due to all such elementary waves from OB is

$$\frac{E \sin \theta}{2Z_0}\int_0^{l_2} exp\left(= j\frac{4\pi y}{\lambda}\sin^2 \theta/2\right) dy$$
$$= j\frac{E\lambda \cot \theta/2}{4\pi Z_0}\left\{exp\left(-j\frac{4\pi l_2}{\lambda}\sin^2 \theta/2\right) - 1\right\} \quad \ldots \quad \ldots \quad (3)$$

Remembering the π phase change on reflection at B the phase at O of the current wave from Q travelling towards B lags on that of I_0 by the angle

$$\frac{2\pi}{\lambda}\left\{2l_2 - y(1+\cos\theta)\right\} + \pi = \frac{4\pi}{\lambda}(l_2 - y\cos^2\theta/2) + \pi$$

Therefore, the component at O of the indirect current wave from OB is

$$-\frac{E\sin\theta}{2Z_0}\int_0^{l_2} exp\left\{-j\frac{4\pi}{\lambda}\left(l_2 - y\cos^2\theta/2\right)\right\}dy$$

$$= j\frac{E\lambda\tan\theta/2}{4\pi Z_0}\left\{exp\left(-j\frac{4\pi l_2}{\lambda}\sin^2\theta/2\right) - exp\left(-j\frac{4\pi l_2}{\lambda}\right)\right\} \quad .. \quad .. \quad (4)$$

The total current at O from AB due to the direct and indirect current wave, the latter after one reflection, is

$$R = A - jB$$

where, remembering that $l_1 - l_2 = \frac{n\lambda}{2}$, we have :—

$$A = \frac{E\lambda}{4\pi Z_0}(\tan\theta/2 + \cot\theta/2)\left\{\sin\left(\frac{4\pi l_1}{\lambda}\cos^2\theta/2\right) + \sin\left(\frac{4\pi l_2}{\lambda}\sin^2\theta/2\right) - \sin\frac{4\pi l_1}{\lambda}\right\}$$

$$= \frac{E\lambda}{2\pi Z_0\sin\theta}\left\{\sin\left(\frac{4\pi l_1}{\lambda}\cos^2\theta/2\right) + \sin\left(\frac{4\pi l_2}{\lambda}\sin^2\theta/2\right) - \sin\frac{4\pi l_1}{\lambda}\right\}$$

Similarly

$$B = \frac{E\lambda}{2\pi Z_0\sin\theta}\left\{1 - \cos\left(\frac{4\pi l_1}{\lambda}\cos^2\theta/2\right) - \cos\left(\frac{4\pi l_2}{\lambda}\sin^2\theta/2\right) + \cos\frac{4\pi l_1}{\lambda}\right\}$$

When $l_1 = l_2 = l$ it is found that

$$|R| = (A^2 + B^2)^{\frac{1}{2}}$$

$$= \frac{E\lambda}{\pi Z_0\sin\theta}\left\{\cos\left(\frac{2\pi l}{\lambda}\cos\theta\right) - \cos\frac{2\pi l}{\lambda}\right\} \quad .. \quad .. \quad (5)$$

Three special cases of equation (5) are interesting

(*a*) $l = (2n+1)\lambda/4$

$$|R| = \frac{E\lambda}{\pi Z_0\sin\theta}\left\{\cos\left[\frac{2n+1}{2}\pi\cos\theta\right]\right\} \quad (6)$$

(*b*) $l = (2n+1)\lambda/2$

$$|R| = \frac{2E\lambda}{\pi Z_0\sin\theta}\left\{\cos^2\left[\frac{2n+1}{2}\pi\cos\theta\right]\right\} \quad (7)$$

(*c*) $l = n\lambda$

$$|R| = \frac{2E\lambda}{\pi Z_0\sin\theta}\sin^2(n\pi\cos\theta) \quad .. \quad .. \quad (8)$$

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CORRESPONDENCE

Letters of technical interest are always welcome. In publishing such communications the Editors do not necessarily endorse any technical or general statements which they may contain.

Pulse Modulation

To the Editor, "Wireless Engineer."

SIR,—I wish to challenge some of the statements made by Roberts and Simmonds in Section 1.4 of their article in the November issue of *Wireless Engineer* (Vol. 22, p. 543).

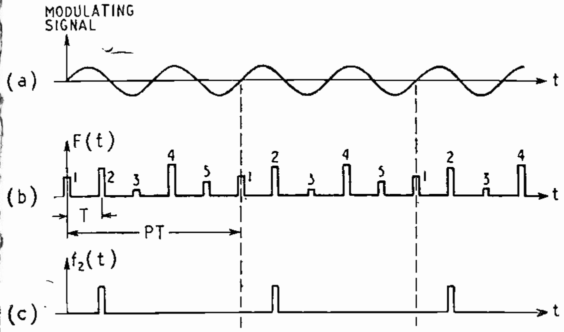
Equation (1) is the Fourier series corresponding to a waveform which repeats itself at intervals equal to $2\pi/\omega$ (I shall, except where noted to the contrary, follow the notation used in Roberts and Simmonds' article). The quantity a_ν is a constant (Sect. 1.4.1, line 9) determined by the shape of the pulses and the integer ν, and for a chain of rectangular pulses of amplitude A it is given by

$$a_\nu = 2AS\frac{\sin\nu\pi S}{\nu\pi S}$$

Since A has been assumed constant, the Fourier series corresponding to the amplitude-modulated pulse chain cannot be obtained by replacing it by $A(1 + m\cos pt)$, which is variable. Furthermore, the modulated pulse chain does not repeat itself at intervals of $2\pi/\omega$, so that at least two of the conditions under which equation (1) was derived no longer hold. A similar criticism can be applied to the procedure followed in Sections 1.4.2 and 1.4.3.

A chain of rectangular pulses is a succession of discrete phenomena to which the idea of continuous variation cannot be applied. For example, in the case of delay-modulation the shape of the pulse does not alter. In the case of amplitude and width-modulation the shape of the pulse is determined by the modulating signal at some instant before the pulse occurs. In all three types of modulation the property of the pulse chain which is being varied is determined by the modulating signal at a number of instants equally spaced along the time axis; what happens to the modulating signal between these instants has no effect on the modulated pulse chain. The mathematical method of obtaining the Fourier series of the modulated chain of pulses must therefore be one which will take this into account. Such a method may be outlined in the following way.

Suppose that the ratio of the pulse to the modulating frequencies, ω/p, is a rational fraction $r = P/Q$, where P and Q are integers prime to one another. The waveform of the modulated pulse chain then repeats itself at intervals equal in length to P pulses or Q cycles of the modulating frequency, since these are equal by the definition of P and Q. In order to fix our ideas let us consider a pulse chain amplitude-modulated by the sinusoidal modulating signal shown in Fig. 1 (*a*). For a ratio $r = 5/2$, the resultant pulse chain will be as shown in Fig. 1 (*b*). This (infinite) pulse chain may be divided into P pulse chains of period PT ($T = 2\pi/\omega$), which I will refer to as *partial pulse chains;* let the Fourier series of the nth one be denoted by $f_n(t)$. Such a partial pulse chain, viz. $f_2(t)$ is shown in Fig. 1 (*c*).



If $F(t)$ denote the Fourier series of the complete modulated pulse chain we have

$$F(t) = \sum_{n=1}^{P} f_n(t).$$

If we assume that the epoch of the leading edge of the pulse is

$$t_n = (n - 1)T - \tfrac{1}{2}ST$$

and that the amplitude of the nth partial pulse chain is given by

$$A_n = A_0(1 + m \sin pt_n)$$

then we obtain for the amplitude of the component in the modulating frequency in $F(t)$ the expression

$$mA_0S \frac{\sin \pi S/r}{\pi S/r}$$

and that the amplitude of the component with pulsatance $\nu\omega \pm p$ is

$$mA_0S \frac{\sin [\pi(\nu \pm 1/r)S]}{\pi(\nu \pm 1/r)S}$$

The differences between these results and those derived from equation (3) are almost trivial, nevertheless in the case of width- and delay-modulated pulse chains the results of a similar chain of reasoning are very different. For a width-modulated pulse chain the amplitude of the component in the modulating frequency is

$$mAS \frac{J_1(2\pi mS/r)}{\pi mS/r} \longrightarrow mAS$$

as $\pi mS/r \rightarrow 0$. $J_1(z)$ is the Bessel function of the first kind of order unity. This departure from linearity is 1.1 per cent for $\pi mS/r = 0.15$, and might not be noticed in practice. For delay-modulated pulse chains a component in the modulating frequency is present and its amplitude is

$$\frac{AS\,\phi_{max}}{r} \frac{\sin \pi S/r}{\pi S/r} \frac{J_1(\phi_{max}/r)}{\phi_{max}/2r} \longrightarrow \frac{AS\phi_{max}}{r}$$

as $r \rightarrow \infty$, that is for a given pulse chain it is proportional to the frequency of the modulating signal.

This latter result is a contradiction of the statement by Roberts and Simmonds in Section 1.4.3 that such a component does not exist. The fact that the amplitude of the signal varies linearly with frequency in no way makes the system private, because it is only necessary to apply it to an RC integrating circuit to obtain a signal whose amplitude is independent of the frequency, provided $1/pCR \ll 1$ for the lowest modulating frequency required. Of course the amplitude of this signal would be small, both because S would be chosen small in a phase-modulated system and also because $1/r < 1$.

I hope to publish a fuller account of these calculations in the near future.

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Rationalization of Publication

To the Editor, "The Wireless Engineer."

SIR,—I would like to make a few comments about Mr. Bell's letter regarding Rationalization of Publication. I am, in general, in complete agreement with his remarks, but I would like to point out that there are many books on limited branches of science or engineering which have been published in this country, many of them before books on the same subjects appeared in America.

I am rather surprised at his suggestion that the types of publication mentioned in his letter are not available as British books by British authors and I certainly consider he is unduly pessimistic in this respect.

I quite agree that the better type of book is international in scope and I consider that technical books should give references to the origin of papers and books from which information has been taken, without regard to the nationality of the authors. In this respect I find that the majority of American books and publications signally fail and I must also admit that many British books fail in the same way but the failing is nothing like so great as in the case of American publications. That I have always held this view may be gauged by the fact that in my book "Time Bases" the following numbers of references are made.

British references	58
American ,, ...	29
German ,, ...	5
French ,, ...	3
Italian ,, ...	1

I am, however, also in agreement with Mr. Bell's suggestion that a definite policy regarding the rationalization of publication is urgently needed and I would like to suggest that the three main Engineering Institutions of this country, together with perhaps the Physical Society, the Chemical Society and others of like nature, might combine for the purpose of originating a common policy in connection with these publications.

It would, of course, be necessary and advisable to obtain the advice of such bodies as the National Physical Laboratory, the Royal Society and various book and periodical publishing firms.

O. S. PUCKLE.
Amersham, Bucks.

NOISE FACTOR OF VALVE AMPLIFIERS

By N. R. Campbell, Sc.D., F.Inst.P.,
V. J. Francis, B.Sc., F.Inst.P., A.M.I.E.E., and E. G. James, Ph.D., B.Sc.
(Communication from the Research Staff of The M.O. Valve Company at the G.E.C. Research Laboratories, Wembley, England)

(Concluded from page 83 of the March issue)

10. Definition of and Formulae for Noise Factor

We are now in a position to discuss noise factor. If the signal generator is independent (in the sense of the last paragraph of Section 2) of the other sources of noise, then the noise of the practical receiver can be written $v_s^2 + v_p^2$ where v_s^2 is the noise at the output due to the signal generator and v_p^2 that due to the other sources. In the ideal receiver v_p^2 is zero by hypothesis; further the signal voltage is the same for the practical and the ideal receiver. We have therefore, from the fundamental definition of N in Section 1,

$$N = \frac{v_s^2 + v_p^2}{v_s^2} \qquad (10.1)$$

$$= 1 + \frac{v_p^2}{v_s^2} \qquad (10.2)$$

We now consider an amplifier of n stages; then

$$v_s^2 = 4k'T \int_0^\infty R_s|A_1|^2.|A_2|^2 \ldots df \qquad (10.3)$$

and v_p^2 will be the sum of the components from all stages which from (9.3) is given by

$$v^2_p = 4k'T \int_0^\infty Z_1|A_2|^2|A_3|^2 \ldots |A_n|^2\, df + 4k'T \int_0^\infty Z_2|A_3|^2|A_4|^2 \ldots |A_n|^2\, df + \ldots + 4k'T \int_0^\infty Z_{n-1}|A_n|^2\, df + 4k'T \int_0^\infty Z_n\, df \qquad (10.4)$$

It should be noted that Z_1 by definition of v_p^2 does not include the noise from the generator.

Accordingly

$$N = 1 + \frac{\left\{ \int_0^\infty Z_1|A_2|^2|A_3|^2 \ldots |A_n|^2\, df + \ldots + \int_0^\infty Z_{n-1}|A_n|^2\, df + \int_0^\infty Z_n\, df \right\}}{\int_0^\infty R_s|A_1|^2|A_2|^2 \ldots |A_n|^2\, df} \qquad (10.5)$$

If the amplifier consists of one stage only, (10.5) reduces to

$$N = 1 + \frac{\int_0^\infty Z_1\, df}{\int_0^\infty R_s|A_1|^2\, df} = 1 + \frac{\int_0^\infty \sum_{pq} (G_{pq} + G_{pqo})|\rho_{kl,\,pq}|^2\, df}{\int_0^\infty R_s|A_1|^2\, df} \qquad (10.6)$$

where $\sum_{pq} G_{pq}$ does *not* include R_s.

In any practicable receiver the range of each integral can be restricted to finite limits, f_1 to f_2, outside which the integrand is zero. So long as the integrand varies appreciably within the limits (which may not be the same for all the integrals), the integrals must be retained, and no general simplification of (10.5) is possible. On the other hand, it is assumed in much noise theory that all the quantities concerned are constant when they are not zero. If this assumption is permissible the integral signs in (10.5) can be omitted, so that (10.5) becomes

$$N = 1 + \frac{1}{R_s|A_1|^2} \left\{ Z_1 + \ldots + \frac{Z_r}{|A_2|^2 \ldots |A_r|^2} + \ldots + \frac{Z_n}{|A_2|^2 \ldots |A_n|^2} \right\} \qquad (10.7)$$

An intermediate condition is when the assumption is true of stage 1, but not of the others. Then (10.5) becomes

$$N = 1 + \frac{1}{R_s|A_1|^2}\left\{\frac{\int_0^\infty Z_1|A_2|^2 \ldots |A_n|^2\,df + \ldots + \int_0^\infty Z_r|A_{r+1}|^2 \ldots |A_n|^2\,df + \ldots}{\int_0^\infty |A_2|^2 \ldots |A_n|^2\,df}\right\} \quad (10.8)$$

11. Application of Noise Factor Formulae

The later stages of the amplifier do not include a signal generator. But the input circuits contain effective generators of E.M.F. derived from the output voltage of the previous stages. In defining the A_r's it was said that the input circuit of the rth stage may be taken to be one having an admittance equal to the admittance looking into the output terminals of the $(r-1)$th stage, which is ${}_{r-1}Y_{kl}$, and containing a generator whose E.M.F. is the voltage across those output terminals when they are disconnected from the input terminals of the rth stage. In calculating the ρ's for the rth stage, i.e. ${}_r\rho_{kl,pq}$, the admittance ${}_{r-1}Y_{kl}$ must be included in ${}_rY_{hj}$; but in calculating Z_r no noise arising from ${}_{r-1}Y_{kl}$ must be included, for any such noise has already been included in Z_{r-1}; that is to say, in calculating the component of the form (8.6), in which (pq) is (hj), G_{pq} must exclude ${}_{r-1}G_{kl}$.

Moreover Section 10 does not specifically mention the possibility that the signal generator may be connected to the input terminals of the first stage, or that the output terminals of stage $(r-1)$ may be connected to the input terminals of stage r, not directly, but through a transformer. In practice a transformer is almost always inserted at least between the signal generator and the first stage, and performs a most important function. The noise factor depends, when all other characteristics are fixed, on the impedance of the signal generator; the effective value of this impedance can be changed by changing the transformation ratio of the transformer. A large part of the practical adjustment of the amplifier to optimum performance consists therefore in varying the transformation ratio of this transformer so as to make the noise factor a minimum for that variation.

Let us suppose, then, that a *PT* (Section 7) has its secondary terminals connected directly to the input terminals (h, j) of the first stage, and its primary terminals to those of the signal generator, the y's of the transformer being adjusted so that, when the Y_1 of Section 7 is Y_s the admittance of the generator Y_2 is Y_u. (Y_u is the quantity so denoted in Sections 4, 5.) Then the (h, j) terminals are connected effectively, by virtue of Thévenin's theorem, by an impedor of admittance Y_u in series with the E.M.F applied by the transformer. The gain from the terminals of this E.M.F. to the (k, l) terminals (the output terminals of the first stage) is therefore, by (3.9), $Y_u \cdot {}_1\rho_{kl,hj}$, where ${}_1\rho_{kl,hj}$ must be calculated with Y_u replacing Y_s. If A_1 is the gain from the terminals of the signal generator to the (k, l) terminals, we have, from (7.7)

$$|A_1|^2 = \frac{|Y_s|^2}{|Y_u|^2}\cdot\frac{G_u}{G_s}\cdot|Y_u|^2|{}_1\rho_{kl,hj}|^2 \quad (11.1)$$

Therefore, from (6.1)

$$R_s|A_1|^2 = G_u|{}_1\rho_{kl,hj}|^2 \quad \ldots \quad (11.2)$$

Accordingly the effect of the *PT* is simply to substitute quantities with suffix u for the corresponding quantities with suffix s in the integrand of (10.3) or (10.5). It should be noted that the quantities in (11.2) are functions of frequency.

From (11.2) it is obvious that, if the signal generator provides a sinusoidal E.M.F. of frequency f_0, and A_s is the gain for this signal from the terminals of the generator to the (k, l) terminals—which is what is usually called the signal gain—then the foregoing argument gives

$$|A_s|^2 = \frac{G_u(f_0)}{R_s(f_0)}\,|{}_1\rho_{kl,hj}(f_0)|^2 \quad \ldots \quad (11.3)$$

where, of course, (f_0) after a quantity indicates that its value is to be taken at the frequency f_0.

Since a perfect transformer cannot be exactly realised, we must, in accordance with Section 7, introduce a conductor of conductance γ in parallel with Y_u. γ will vary when Y_u is varied by adjustment of the transformer, and, when the adjustment is made, will in general vary with frequency, like all the other admittances. In making the Thévenin transformation γ may be excluded from the PT in which case it does not appear in Y_u but its presence will have to be taken into account in estimating the ρ's; further by (8.1), it will provide a component of the noise $4k'T\int_0^\infty \gamma|\rho_{kl,hj}|^2\,df$. It is relevant here to observe that in most valves operating at high frequencies there is appreciable leakage between the (h, j) terminals, which may be represented by a conductor (whose conductance may vary with frequency) between these terminals. So long as there is no inductance in the leads to the electrodes, this conductor is in parallel with that which represents the imperfection of the input transformer; the effect of both may be represented by a single γ.

If there is a transformer between the $(r - 1)$ th and rth stages, it may be regarded as part of either stage, or even divided into two parts, one belonging to one stage and the other to the other, so long as each stage still has definite input and output terminals. Different procedures may be suitable in different circumstances. If it is regarded as part of the $(r - 1)$th stage, it will have three effects. First, it will multiply by the same factor all the gains of the $(r - 1)$th stage; second, it will change the admittance looking into the output terminals of that stage; third (since the transformer cannot be PT), it will introduce a conductance γ in parallel with the output terminals. It will almost always be convenient to regard this conductance as part of the rth stage rather than of the $(r - 1)$th, that is to say as part of ${}_rY_{hj}$ rather than of ${}_{r-1}Y_{kl}$; for, as has been said, ${}_rY_{hj}$ will usually already have a part γ, whose value is merely modified by a part arising from the imperfection of the transformer. We have therefore to consider only the first two effects. Let dashed quantities refer to the stage without the transformer, undashed quantities to the stage with the transformer. Then the second effect will be to substitute ${}_{r-1}Y_{kli}$ for ${}_{r-1}Y'_{kli}$, the relation between these two quantities being determined by the y's of the PT. By (7.7) we shall then have, for the first effect,

$$\frac{|A_{kl,pq}|^2}{|A'_{kl,pq}|^2} = \frac{|\rho_{kl,pq}|^2}{|\rho'_{kl,pq}|^2} = \frac{|{}_{r-1}Y'_{kli}|^2}{|{}_{r-1}Y_{kli}|^2}\cdot\frac{{}_{r-1}G_{kli}}{{}_{r-1}G'_{kli}} = \frac{{}_{r-1}R_{kli}}{{}_{r-1}R'_{kli}} \quad \ldots \quad \ldots \quad (11.4)$$

12. Noise Factor of a Common-Cathode Pentode

It will be well before we proceed to illustrate one of the apparently complicated conclusions presented in Sections 10, 11 by a concrete example. We shall take that of a single stage consisting of a common-cathode pentode without inductance in the electrode leads, fed by a signal generator through a transformer.

In the notation of Section 5, the (k, l) terminals are (1, 3), the (h, j) terminals are (1, 2). The sources of noise are the impedors Y_u and γ between (1, 2), and the electron stream and the anode load Y_v between (1, 3). We have therefore from (10.5) and (11.2)

$$N = 1 + \frac{\int_0^\infty \{\gamma|\rho_{13,12}|^2 + (G_v + G_{13\sigma})|\rho_{13,13}|^2\}\,df}{\int_0^\infty G_u|\rho_{13,12}|^2\,df} \quad \ldots \quad \ldots \quad (12.1)$$

where $G_{13\sigma}$ from (8.11), since $\rho_{kl,pr}$ is here zero, is given by

$$G_{13\sigma} = a\,I_0(\alpha_3{}^2\,\Gamma^2 + \alpha_3\alpha_4) \quad \ldots \quad (12.2)$$

Since (3.11) is true, we can omit the common denominator Δ of the ρ's, and replace them by δ's. Then, from (5.4, 6), remembering that we must now add γ to Y_{02} in (5.1), and that

$$\delta_{13,12} = \Delta_{32}\,; \quad \delta_{13,13} = \Delta_{33} \quad \ldots \quad (12.3)$$

we have

$$N = 1 + \frac{\int_0^\infty \gamma(\alpha_3{}^2 g_2{}^2 + \omega^2 C^2{}_{23})\,df + \int_0^\infty (G_v + G_{13\sigma})\{(G_u + \gamma)^2 + (B_u + \omega C_2)^2\}\,df}{\int_0^\infty G_u(\alpha_3{}^2 g_2{}^2 + \omega^2 C^2{}_{23})\,df} \quad \ldots \quad (12.4)$$

where

$$C_2 = C_{02} + C_{12} + C_{23} + C_{24} + C_{25} \quad \ldots \quad \ldots \quad (12.5)$$

The problem of adjusting the input transformer, and therefore G_u, B_u, so that N is a minimum is in general very complicated. For G_u, B_u are functions of f whose form can be varied by varying the y's of the PT; moreover γ, which is also a function of f, will vary with G_u, B_u. If, however, we are interested only in a frequency band which is so narrow that throughout all quantities may be considered constant, then (12.4) becomes

$$N = 1 + \frac{\gamma}{G_u} + \frac{(G_v + G_{13\sigma})\{(G_u + \gamma)^2 + (B_u + \omega_0 C_2)^2\}}{G_u(\alpha_3^2 g_2^2 + \omega_0^2 C^2_{23})} \quad .. \quad .. \quad (12.6)$$

where G_u, B_u are variable quantities and ω_0 refers to the centre of the band. Further, if it may be assumed that γ varies slowly with G_u, B_u, so that in adjusting them for minimum N, γ may be taken as a known constant, then the conditions for minimum N are

$$\frac{\partial N}{\partial G_u} = 0\,; \quad \frac{\partial N}{\partial B_u} = 0 \quad .. \quad .. \quad (12.7)$$

The second of these gives

$$B_u = -\omega_0 C_2 \quad .. \quad .. \quad .. \quad (12.8)$$

The first gives

$$G_u^2 = \gamma G_c + \gamma^2 \quad .. \quad .. \quad (12.9)$$

where

$$G_c = \frac{\alpha_3^2 g_2^2 + \omega_0^2 C^2_{23}}{G_v + G_{13\sigma}} \quad .. \quad .. \quad (12.10)$$

Consequently

$$N_{min} = 1 + 2\frac{\gamma}{G_c} + 2\sqrt{\frac{\gamma}{G_c}\left(1 + \frac{\gamma}{G_c}\right)} \quad .. \quad .. \quad (12.11)$$

This result shows how important is γ. For, if it were zero, the optimum value of G_u would be zero, and N_{min} would be 1. At the same time, by (11.3), the signal gain would be zero. The problem would become entirely insignificant. That there is a real optimum adjustment of the transformer in this simplified case depends wholly on the presence of a non-zero γ.

13. Effect on the Noise Factor of a Prefixed Stage

We can now approach the problem set forth in Section 1. Suppose again that the given amplifier has n stages. Then its noise factor N_D, measured with a signal generator of internal impedance Z_s, is given by (10.5). Let a pre-amplifier, which we shall call stage 0, be prefixed, and its noise factor N be measured with the same signal generator. Then, if the gains A_1 to A_n are unchanged, we have

$$N = 1 + \frac{\left\{\int_0^\infty Z_0|A_1|^2 \ldots |A_n|^2\,df + \int_0^\infty Z_1|A_2|^2 \ldots |A_n|^2\,df + \ldots + \int_0^\infty Z_n\,df\right\}}{\int_0^\infty R_s|A_0|^2|A_1|^2 \ldots |A_n|^2\,df} \quad .. \quad .. \quad (13.1)$$

If R_s is independent of frequency, so that it can be taken outside the integral in (10.5), and if the variation of A_0 over the band is sufficiently small, then

$$N = 1 + \frac{\int_0^\infty Z_0|A_1|^2 \ldots |A_n|^2\,df}{R_s|A_0|^2\int_0^\infty |A_1|^2 \ldots |A_n|^2\,df} + \frac{\int_0^\infty Z_1|A_2|^2 \ldots |A_n|^2\,df + \ldots + \int_0^\infty Z_n\,df}{R_s|A_0|^2\int_0^\infty |A_1|^2|A_2|^2 \ldots |A_n|^2\,df} \quad .. \quad .. \quad (13.2)$$

$$= 1 + \frac{\int_0^\infty Z_0|A_D|^2 \,.\, df}{R_s|A_0|^2\int_0^\infty |A_D|^2\,df} + \frac{N_D - 1}{|A_0|^2} \quad .. \quad .. \quad (13.3)$$

where $A_D = A_1 \,.\, A_2 \ldots A_n$ .. (13.4)

is the whole gain of the amplifier from the terminals of the signal generator feeding stage 1 to the (k, l) terminals of stage n. Further in virtue of the conditions of the problem, A_D is likely to be appreciably constant through the pass-band, although the components A_r may vary. Accordingly we may omit the integral signs in (13.3), and arrive at the simple formula

$$N = 1 + \frac{Z_0}{R_s|A_0|^2} + \frac{N_D - 1}{|A_0|^2} \quad .. \quad (13.5)$$

It should be noted that (13.5) has been derived on the assumption that the A_r's are unchanged by the prefixing of stage 0. This would imply that the output admittance of stage 0 is equal to the admittance of the generator and is generally not true.

It is most improbable that the gains A_1 to A_n will be unchanged between the measurement of N_D and the measurement of N. The conditions when N_D is measured are likely to be those shown to the right of the dotted line in Fig. 7. The signal generator is applied to the (h, j) terminals of stage 1 through a transformer T_1, and the y's of this transformer are adjusted to make N_D a minimum for variation of those y's. If, when N is measured, the transformer T_1 is removed with the signal generator, A_1 is reduced by the gain in the transformer. If T_1 is left in place, and the (k, l) terminals of stage 0 are connected to its input in place of those of the signal generator, the input admittance of T_1 is changed from Y_s to ${}_0Y_{kl}$, which is not in general even nearly the same as Y_s. Accordingly the gain of T_1 will change, and so will the admittance of ${}_1Y_u$, which is part of the admittance ${}_1Y_{hj}$; accordingly A_1 will change. Further ${}_1Y_{kl}$ will change, and therefore A_2 will change; and so on; and the amplifier will no longer be adjusted for minimum N_D. (It is not certain that adjustment for minimum N will involve adjustment for minimum N_D; but that represents at least a possibility to be explored.)

In order to avoid this difficulty, a transformer T_2 may be inserted between the output of stage 0 and the input of stage 1 when the signal generator is removed, as shown to the left of the dotted line. (Actually, of course, if this is done, T_1, T_2 are likely to be combined.) T_2 transforms the admittance ${}_0Y'_{kli}$, looking into the (k, l) terminals of stage 0 in the absence of the transformer, into ${}_0Y_{kli}$, the admittance looking into the output terminals of the transformer, which are now the (k, l) terminals of stage 0. If it could be arranged that ${}_0Y_{kli} = Y_s$, N_D would again be adjusted to the minimum and A_1 to A_n would retain their original values. But in general this cannot be arranged; ${}_0Y_{kli}$ can be made equal to Y_s at one frequency, but not at all frequencies.

However, if the conditions for (10.8), (13.3, 5) prevail, and the gain of the prefixed valve may be considered constant over the relevant range, it is likely that all the characteristics of stage 0, and in particular, ${}_0Y'_{kli}$, can be considered similarly constant. Then, so long as Z_s is a pure resistance, the desired end can be obtained; and, so long as the condition for minimum N involves the condition that N_D shall be a minimum, minimum N can be obtained by varying only the transformer T_3 between the signal generator and the (h, j) terminals of stage 0.

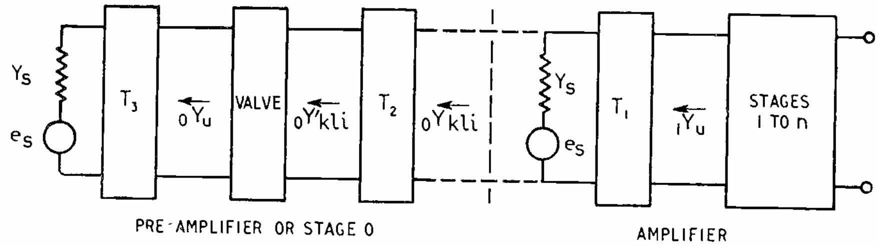

Fig. 7. The use of a pre-amplifier modifies the conditions for the following stage.

By (7.7), if A_0 is the gain of stage 0 without T_2 and A_0 with it,

$$\frac{|A_0|^2}{|A'_0|^2} = \frac{R_s}{{}_0R'_{kli}} \quad \ldots \quad \ldots \quad \ldots \quad (13.6)$$

A'_0 is the A_s of (11.3). Using (6.1b) and (3.16), we have

$$|A_0|^2 = |A'_0|^2 \,.\, R_s \,.\, \frac{|{}_0Y'_{kli}|^2}{{}_0G'_{kli}} = \frac{R_s|A'_0|^2}{{}_0G'_{kli}|{}_0\rho'_{kl,kl}|^2} \quad \ldots \quad \ldots \quad (13.7)$$

When T_2 is introduced, all the gains involved in Z_0 will change in the ratio A_0/A'_0; for they are all gains to the (k, l) terminals of stage 0. Hence, in the second term on the right of (13.3, 5), we may write A'_0 in the denominator so long as we use dashed quantities, when we expand Z_0. Hence (13.5) becomes

$$N = 1 + \frac{Z_0}{R_s|A'_0|^2} + \frac{(N_D - 1)\,{}_0G'_{kli}|{}_0\rho'_{kl,kl}|^2}{R_s|A'_0|^2} \quad \ldots \quad \ldots \quad (13.8)$$

If we now expand Z_0 by (8.7), and, for symmetry, substitute for $R_s|A'_0|^2$ from (11.3), we have

$$N = 1 + \frac{\Sigma_{pq}(G_{pq}+G_{pq\sigma})|{}_0\rho'_{kl,pq}|^2 + (N_D - 1)\,{}_0G'_{kli}|{}_0\rho'_{kl,kl}|^2}{G_u|{}_0\rho'_{kl,hj}|^2} \quad \ldots \quad \ldots \quad (13.9)$$

Now the Σ_{pq} contains a term $G_{kl}|\rho'_{kl,hj}|^2$ where G_{kl} is the conductance of the load which in Sections 4, 5 is called G_v. Hence, in this particular case, the effect on N of the amplifier following the pre-amplifier is the same as if the amplifier produced no noise, but added to the conductance of the load of the pre-amplifier a conductance $(N_D - 1)\,{}_0G'_{kli}$, where ${}_0G'_{kli}$ is the conductance looking into the pre-amplifier from the input terminals of the transformer T_2.

14. Conclusion

In Section 13, some attempt has been made to simplify the fundamental formula (10.5) for particular cases. The validity of the assumptions we make for this purpose are difficult to assess, although it is quite usual for such assumptions to be made.

It is hoped that the theory developed here will allow the validity or otherwise of these assumptions to be verified.

It is hoped also that the theory will make possible the inclusion of all circuit elements (such as stray lead inductances) in the analysis with a view to assessing their effect.

The application of the theory with these objects in mind leads to rather lengthy algebra, and it is proposed to deal with some specific cases in later papers.

15. Acknowledgement

The Authors desire to tender their acknowledgement to the General Electric Company and the Marconiphone Company, on whose behalf the work which led to this publication was done.

CORRECTION

Equation (2.3) of Section 2 should read :—

$$\overline{(v-\bar{v})^2} = \frac{2k'T}{\pi}\int_0^\infty R_{pq}|A_{kl,\,pq}(j\omega)|^2 d\omega$$

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[2] N. R. Campbell, V. J. Francis and E. G. James. "Linear Single-Stage Valve Circuits," *Wireless Engineer*, July 1945, Vol. XXII, p. 333.

[3] D. M. O. North. "Fluctuations in Space-Charge-Limited Currents at Moderately High Frequencies." *R. C. A. Review*, April 1940, Vol. IV. p. 441.

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

573 420.—Capacitance microphone in which one of the plates is relatively massive, for the detection and measurement of mechanical vibrations.

Standard Telephones and Cables Ltd., V. J. Terry, and E. A. Foulkes. Application date 2nd November, 1943.

573 687.—Flexible acoustic fitting for coupling a piezo-electric crystal to a sound-diaphragm.

Brush Crystal Co. Ltd. (assignees of A. L. W. Williams and D. Domizi). Convention date (U.S.A.), 20th March, 1939.

573 688.—Piezo-electric or other drive for a small diaphragm, particularly suitable for a deaf-aid appliance.

Brush Crystal Co. Ltd. (assignees of A. L. W. Williams and D. Domizi). Convention date (U.S.A.) 20th March, 1939.

DIRECTIONAL WIRELESS

573 020.—Variable-frequency type of radio-altimeter arranged to transmit a beam making a conical "sweep" so as also to indicate the speed and course of the aircraft.

R. C. A. Photophone Ltd. and H. W. H. Hodgkins. Application date 3rd July, 1942.

573 436.—Construction and arrangement of the screening-elements of a directional aerial of the kind in which the radiating sections of a conductor are separated by non-radiating sections.

The General Electric Co. Ltd., and R. J. Clayton. Application date 19th August, 1942.

573 744.—Headphone-receiver for navigational signals of the A—N type, with means for modifying the different signal-amplitudes, in order to ease the strain on the pilot's hearing as he approaches the beacon.

Standard Telephones and Cables Ltd. and L. J. Heaton-Armstrong. Application date 24th November, 1943.

TELEVISION CIRCUITS AND APPARATUS

FOR TRANSMISSION AND RECEPTION

572 338.—Louvred device forming a viewing screen for a cathode-ray television receiver.

Philco Radio and Television Corpn. (assignees of G. Zindell, Jr.). Convention date (U.S.A.). 10th September, 1942.

573 008.—Television system in which the image is produced in three dimensions, and is so projected in space (and not upon a flat screen).

J. L. Baird. Application dates 26th August, 27th September, 7th and 11th October, and 26th November, 1943 *and 9th February,* 1944.

573 226.—Multiple-gun cathode ray tube used for scanning a large-sized television screen in successive sections.

A. Soloman. Application date 18th October, 1943.

SIGNALLING SYSTEMS OF DISTINCTIVE TYPE

572 918.—Multi-channel signalling system in which all the frequencies required for inter-communication are controlled by the selected harmonics of a single crystal-oscillator.

The General Electric Co. Ltd., N. R. Bligh, D. M. Heller and L. C. Stenning. Application dates 27th April and 1st July, 1942.

CONSTRUCTION OF ELECTRONIC-DISCHARGE DEVICES

572 540.—Construction and arrangement of the focusing-electrodes for a discharge tube of the velocity-modulation type.

C. S. Bull. Application date 29th April, 1940.

572 813.—Spacing and aligning the focusing and other electrodes of a cathode-ray tube by means of protuberances formed inside and outside the glass bulb.

Ferranti Ltd., S. Jackson, A. L. Chilcot, M. E. Sions and R. W. Sutton. Application date 29th May, 1943.

573 032.—Construction and mounting of the magnetic deflecting coils of a cathode-ray tube.

The British Thomson-Houston Co. Ltd. (communicated by the General Electric Co.). Application date 24th September, 1943.

573 097.—Electron-discharge tube embodying a "Hertzian gap" for detecting and amplifying, or generating, ultra-short waves.

H. A. Malpas. Application date 9th November, 1943.

573 170.—Construction of high-power valve of the kind in which the filament-heating leads consist of two coaxial copper tubes.

Standard Telephones and Cables, Ltd. and W. T. Gibson. Application date 31st March, 1944.

573 193.—Construction of a hollow resonator tube for generating micro-waves by the phase-control of an electron-stream.

Patelhold Patentverwertungs and Elektroholding A.G. Convention date (Switzerland), 1st September, 1942.

573 205.—Arrangement of the control and deflecting electrodes of an electron-discharge device of the beam type, suitable for frequency-doubling or as a relay.

J. H. O. Harries. Application date 5th September, 1940.

573 206.—Construction and mounting of an interwoven spiral cathode, suitable for valves of the beam or deflection type.

J. H. O. Harries. Application date 5th September, 1940.

SUBSIDIARY APPARATUS AND MATERIALS

572 491.—Process for laying an insulating film on sheets of ferrous magnetic material intended for making laminated transformer-cores.

G. R. Shepherd (communicated by Westinghouse Electric International Co.). Application date 25th January, 1944.

572 658.—Process for removing occluded air or gas in the course of the manufacture of dry cells.

Siemens Bros. & Co. Ltd., T. H. Larke and F. G. Kny-Jones. Application date 15th April, 1944.

572 739.—The use of sliding dielectric blocks, having a selected permittivity, as impedance-transformers along a transmission line.

The General Electric Co. Ltd. and D. C. Espley. Application date 11th November, 1941.

572 846.—Selective filter-network having six terminals and comprising both high-pass and low-pass sections.

The General Electric Co. Ltd. and R. O. Rowlands. Application date 19th July, 1943.

572 848.—Assembling and bonding multiple dry-plate rectifier units.

The General Electric Co. Ltd., H. C. Turner and J. Chamberlain. Application date 21st July, 1943.

572 870.—Construction of right-angled branches or stubs for a transmission-line of the coaxial type.

Bruno Patents Inc. Convention dates (U.S.A.) 27th August and 3rd December, 1942.

572 881.—A wave-guide in which part of the guiding surface comprises a quarter-wave "open" grid, which prevents the escape of energy whilst allowing power measurements to be made.

J. Collard. Application date 14th September, 1943.

572 884.—Closed chain of hard-valves operating as a "counter" circuit, and indicating, on the scale of ten, events of the order of 100 000 per second.

Cinema-Television, Ltd. and T. C. Nuttall. Application date 15th February, 1941.

572 999.—Selenium rectifier in which the sensitive material is applied to the electrode from a colloidal suspension in a volatile carrier, the latter then being evaporated.

Standard Telephones and Cables, Ltd. (assignees of O. Saslaw). Convention date (U.S.A.) 24th April, 1942.

573 030.—Resilient metal-ring so shaped as to minimise the tendency of vibration to loosen a thermionic valve in its holder.

A. C. Cossor, Ltd. and A. Levin. Application date 17th September, 1943.

573 099.—Two-core R.F. transformer having a wide band-pass characteristic, and designed to facilitate precision production in quantity.

Philco Radio and Television Corporation (assignees of C. T. McCoy). Convention date (U.S.A.) 29th June, 1943.

573 108.—Electric soldering-iron in which the heat-storing element is under thermostatic control.

Standard Telephones and Cables, Ltd. (trading as Stanelco Products) and J. Handley. Application date 29th November, 1943.

573 124.—Impedance-matching transformer comprising a transmission-line element terminated by a non-reactive load, and a second transmission-line element adjustably shunted across the first.

Standard Telephones and Cables, Ltd. and D. C. Rogers. Application date 1st December, 1943.

ABSTRACTS AND REFERENCES

Compiled by the Radio Research Board and published by arrangement with the Department of Scientific and Industrial Research

The abstracts are classified in accordance with the Universal Decimal Classification. They are arranged within broad subject sections in the order of the U.D.C. numbers, except that notices of book reviews are placed at the ends of the sections. The abbreviations of the titles of journals are taken from the World List of Scientific Periodicals. Titles that do not appear in this List are abbreviated in a style conforming to the World List practice.

ACOUSTICS AND AUDIO FREQUENCIES

534.213.4 : 621.392 815

The Analysis of Plane Discontinuities in [acoustic] Cylindrical Tubes: Part I.—J. W. Miles. (*J. acoust. Soc. Amer.*, Jan. 1946, Vol. 17, No. 3, pp. 259–271.) Using a transmission line analogy the effect of the discontinuity is calculated by considering the higher modes which are excited. The equations of motion of the propagation of a small disturbance in a cylindrical tube yield the two-dimensional wave equation, the solutions of which constitute an infinite set of modes. The analogy with the electrical transmission line is established, and it is shown that each mode requires a separate transmission line.

534.213.4 : 621.392 816

The Analysis of Plane Discontinuities in [acoustic] Cylindrical Tubes: Part II.—J. W. Miles. (*J. acoust. Soc. Amer.*, Jan. 1946, Vol. 17, No. 3, pp. 272–284.) "The theory of plane discontinuities developed in part I [see 815 above] is applied to a 'window' and change of cross section in tubes of both circular and rectangular cross section. In the case of the rectangular tube conformal mapping is introduced as an alternative approach to the problem. Reflection and transmission coefficients are determined from the equivalent circuit of part I, and the theory is also applied to the analysis of certain types of cavity resonators. Finally, the experimental determination of equivalent circuit impedance is discussed."

534.23 817

Acoustic Transmission through a Fluid Lamina.—I. Rudnick. (*J. acoust. Soc. Amer.*, Jan. 1946, Vol. 17, No. 3, pp. 245–253.) Based on a derivation of the acoustic wave equation for moving media, the transmission and reflection coefficients are derived for a fluid lamina in uniform motion. Some measurements are reported in the frequency range 2 kc/s to 14 kc/s for angles of incidence of between 0° and 89°.

534.231 + 621.396.11 818

The Saddle-Point Method in the Vicinity of a Pole with Applications to Wave-Optics and Acoustics.—Ott. (*See* 1024.)

534.232 + 621.396.677 819

Small-Dimensional Radiating and Receiving Systems of High Directivity.—Makov. (*See* 857.)

534.232 820

Acoustic Intensity Distribution from a "Piston" Source: II. The Concave Piston.—A. O. Williams, Jr. (*J. acoust. Soc. Amer.*, Jan. 1946, Vol. 17, No. 3, pp. 219–227.) It is shown how an approximate expansion may be obtained for the r.m.s. excess pressure in the supersonic beam from either a plane or concave piston source when the wavelength is small and the radius of curvature large compared with the aperture of the source. For part I see 2670 of 1945 (Williams and Labaw).

534.321.9 821

New Aspect of Ultrasonics.—B. K. Sahay. (*J. acoust. Soc. Amer.*, Jan. 1946, Vol. 17, No. 3, pp. 285–286.) The study of ultrasonics may reveal a corpuscular nature of sound energy.

534.321.9 : 620.179 822

The Supersonic Reflectoscope, an Instrument for Inspecting the Interior of Solid Parts by Means of Sound Waves.—F. A. Firestone. (*J. acoust. Soc. Amer.*, Jan. 1946, Vol. 17, No. 3, pp. 287–299.) Subsidiary reflections take place from flaws in the metal. The signal received by a piezoelectric crystal is displayed on a c.r. tube.

534.41 + 534.781 823

Visible Patterns of Sound.—R. K. Potter. (*Science*, 9th Nov. 1945, Vol. 102, No. 2654, pp. 463–470.) Phase relationships between fundamental and harmonics may render waveform records of sound dissimilar to the eye when the differences are indistinguishable to the ear. In the sound spectrograph the record obtained on a magnetic tape is played back repeatedly into a scanning filter, the passband of which moves across the frequency spectrum. The legibility of the records thus obtained is discussed. Learning to read visible speech is comparable in difficulty with learning to lip read.

534.522.5 **824**

Lissajous Figures.—I. B. Cohen. (*J. acoust. Soc. Amer.*, Jan. 1946, Vol. 17, No. 3, pp. 228–230.) A description of the harmonograph used by Professor Sabine to draw some Lissajous' figures.

534.833.1 **825**

Demountable Soundproof Rooms.—W. S. Gorton. (*J. acoust. Soc. Amer.*, Jan. 1946, Vol. 17, No. 3, pp. 236–239.) A room having an attenuation of about 43 db is made from panels consisting of two sheets of steel cemented to two sheets of composition board with a rock-wool blanket between. The panels weigh 7 lb/sq. ft. It is easy to construct and dismantle, and its materials can nearly all be re-used.

534.833.4 **826**

The Effect of Position on the Acoustical Absorption by a Patch of Material in a Room.—C. M. Harris. (*J. acoust. Soc. Amer.*, Jan. 1946, Vol. 17, No. 3, pp. 242–244.)

621.395.61 **827**

Gradient Microphones.—H. F. Olson. (*J. acoust. Soc. Amer.*, Jan. 1946, Vol. 17, No. 3, pp. 192–198.) " The response-frequency and directional characteristics of gradient microphones from zero to *n*th order have been obtained. Higher order unidirectional microphones have also been developed. The directivity increases with the order which means a corresponding discrimination against random sound. The accentuation of the low-frequency response, when the distance from a small source is less than a wave-length, increases with the order and the wave-length. This feature may be used to obtain high discrimination against distant sounds in a close talking microphone. The amount of discrimination increases with the order of the microphone."

621.395.623.52 **828**

Generalized Plane Wave Horn Theory. — V. Salmon. (*J. acoust. Soc. Amer.*, Jan. 1946, Vol. 17, No. 3, pp. 199–211.) " By the use of dimensionless variables and simplifying transformations Webster's plane wave horn equation is recast into a form permitting separation of the effects of horn contour and frequency. A generalized expression for the admittance also displays this separation. Further interrelations among the variables are developed which permit the formal synthesis of a horn from a given conductance or susceptance function. The conditions for realizability of the horn thus synthesized are discussed. Several applications of the results are presented, including a comparison with Freehafer's exact theory for the hyperbolic horn."

621.395.623.52 **829**

A New Family of Horns.—V. Salmon. (*J. acoust. Soc. Amer.*, Jan. 1946, Vol. 17, No. 3, pp. 212–218.) " A new family of horns is synthesized in which the exponential forms a central member. This permits the effect of perturbations from the exponential contour to be estimated. From other members of the family unique impedance characteristics are obtained, and are discussed with possible applications in mind."

621.395.623.75 **830**

A High Quality Loudspeaker of Small Dimensions.—P. W. Klipsch. (*J. acoust. Soc. Amer.*, Jan. 1946, Vol. 17, No. 3, pp. 254–258.) Description of a horn designed to be used with a " woofer " (see 727 of 1942 and 1708 of 1943—Klipsch). 330 c/s l.f. cut-off, no peaks or hollows up to 7 kc/s, satisfactory performance to 12 kc/s. Normal radiation angle $\pi/2$ steradians. Negligible harmonic distortion " even at painfully loud intensities".

621.395.625.2/.6 **831**

Film Sound-Recording Methods.—W. Hahm. (*Funktech. Mh.*, Sept./Oct. 1943, Nos. 9/10, pp. 102–104.) A review of the distinctive features of each of six recording systems, variable area and variable density light recording, the Philips–Miller and Tefiphone systems using mechanically engraved film, the Magnetophone system of magnetic recording on a film loaded with ferromagnetic particles, and the Selophone system in which the movement of a vibrating wire of a string galvanometer is recorded photographically.

621.395.625.2 **832**

B.B.C. Disc Recording : Some Technical Details of the New Equipment.—H. Davies. (*Wireless World*, Jan. 1946, Vol. 52, No. 1, pp. 14–18.) The recorder is fed from a 75 W amplifier, with negative feedback obtained from a winding on the cutter head, " radius compensation " to increase the high-frequency response near the centre of the disk, and an " overload protector " to prevent sudden increases of sound from causing harmonic distortion. The traverse head, which carries an oil-damped balanced armature cutter, has automatic positioning and " limit safety " switches. Friction drive to the turntable through an idler wheel prevents vibration and " wow".

621.395.625.2 : 621.396.662.667 **833**

High Fidelity Bass Compensation for Moving Coil Pick-Ups.—F. M. Haines. (*Electronic Engng*, Feb. 1946, Vol. 18, No. 216, pp. 45–46.) High fidelity in the moving-coil pickup has been hampered by lack of adequate bass compensation. A circuit is described which improves bass compensation by the use of frequency-selective negative feedback. Optimum resistance and impedance values for good response over the whole frequency range are discussed.

621.395.625.3 **834**

German Tape-Recording Equipment.—(*Electronic Engng*, Feb. 1946, Vol. 18, No. 216, p. 54.) Another brief account of the " Magnetophon " taken from 538 of March. This system was described in 1028 of 1939 (Lübeck) and back references.

621.395.625.3 **835**

Multiple Wire Recording.—R. J. Tinkham. (*Communications*, Dec. 1945, Vol. 25, No. 12, p. 99.) The master record provides the signal from which a number of recording heads in parallel are operated.

621.395.625.3 836

High Quality Sound Recording on Magnetic Wire.—L. C. Holmes. (*Communications*, Dec. 1945, Vol. 25, No. 12, pp. 44, 46.) A supersonic (30–100 kc/s) current mixed with the signal in the recording head decreases distortion. Wire speed must exceed a minimum value to obtain best signal to noise ratio. Loss after 1 400 repeated playings is about 8 db. Report of an I.R.E. paper.

621.395.645.3 837

A De-Luxe 60-Watt Amplifier.—G. M. King. (*R.S.G.B. Bull.*, Dec. 1945, Vol. 21, No. 6, p. 90.) Circuit and design details of an audio-frequency amplifier, using valves of the 807 or 6L6 type, for use with a low-output high-quality microphone.

621.395.645.36 838

Quality Amplifiers.—(*Wireless World*, Jan. 1946, Vol. 52, No. 1, pp. 2–6.) Details are given of 4, 8 and 12 W circuits with double push-pull resistance-coupled triode stages.

621.395.92 : 534.771 839

The 6BP Audiometer.—A. H. Miller. (*Bell Lab. Rec.*, Dec. 1945, Vol. 23, No. 12, pp. 464–465.) One control dial adjusts frequency over the range 128 to 9747 c/s; a second dial indicates hearing loss over the range − 15 to + 105 db from standard. The patient, using either an air-conduction or a bone-conduction receiver, indicates with a push-button whether or not he hears the tone.

[787.1/.4] : 534.31 840

The Mechanical Action of Instruments of the Violin Family.—F. A. Saunders. (*J. acoust. Soc. Amer.*, Jan. 1946, Vol. 17, No. 3, pp. 169–186.)

789 : 621.38 841

Old and New Sounds produced by Electrical Musical Instruments.—H. Bode. (*Funktech. Mh.*, May 1940, No. 5, pp. 67–74.) A brief survey of the subject of electrical musical instruments, with a description in more detail of the Melodium produced by the author and O. Vierling. The Melodium is a single-note instrument with a keyboard. A resistance-capacitance oscillator has its resistance selected by the key, and gives an output rich in harmonics. The key also sets the frequency of a filter that determines the musical quality of the note. The amplitude is determined by the force with which the key is depressed. Vibrato, pizzicato, and frequency multiplication and division to give coupling and mixture effects can be introduced by the performer.

A short description is given of an organ that depends on neon-lamp oscillators.

534 842

The Theory of Sound [Book Review].—Lord Rayleigh. Dover Publications, New York, 1945, 984 pp., $4.95. (*J. acoust. Soc. Amer.*, Jan. 1946, Vol. 17, No. 3, p. 301.) With an historical introduction by R. B. Lindsay.

534.321.9 843

Grundlagen und Ergebnisse der Ultraschallforschung [Book Review].—E. Heidemann. W. de Gruyter & Co., Berlin, 1939, 287 pp., RM 24. (*Z. Ver. dtsch. Ing.*, 22nd June 1940, Vol. 84, No. 25, p. 444.) A comprehensive survey, with a bibliography of 1346 items.

534.833 844

Schalldämmungs-Messungen im Laboratorium und in fertigen Gebäuden [Book Review].—W. Bausch. R. Oldenbourg, Berlin, 1939, 33 pp., RM 6.40. (*Z. Ver. dtsch. Ing.*, 15th June 1940, Vol. 84, No. 24, p. 424.) Deals with sound absorption measuring technique in relation to sound-proof buildings.

AERIALS AND TRANSMISSION LINES

621.315.2.029.6 845

Problems in the Manufacture of Ultra-High-Frequency Solid-Dielectric Cable.—A. J. Warner. (*Waves and Electrons*, Jan. 1946, Vol. 1, No. 1, pp. 31–37.) A brief description is given of the various types of u.h.f. cables. The manufacture is outlined with an indication of the difficulties of maintaining a constant characteristic impedance and of preventing the braid from becoming buried in the insulating sheath. The relative merits of long- and short-lay braiding and the effect of plating the braid is discussed. The attenuation increases as the lay angle increases but is almost independent of the percentage coverage between 98 and 80%. The attenuation introduced by each component of a typical cable is given. Production methods for the measurement of capacitance, velocity of propagation, attenuation and dielectric strength are outlined. Early international standardization of cables is recommended.

621.315.21.029.6.091 : 621.317.79 846

Attenuation Test Equipment for V-H-F Transmission Lines.—Muller & Zimmerman. (*See* 982.)

621.392 + 621.396.616 + 621.317.763 847

The Transverse Electric Modes in Coaxial Cavities.—Kirkman & Kline. (*See* 888.)

621.392 848

Vectorial Treatment of Transmission Lines.—J. P. Shanklin. (*Electronics*, Dec. 1945, Vol. 18, No. 12, pp. 162..166.) Graphical solution of transmission-line problems by representing current and voltage as vectors, and distance along the line as phase displacement. Typical problems are solved.

621.392 849

Propagation of Electromagnetic Waves in Branched Hollow-Pipe Lines.—E. M. Studenkov. (*J. Phys.*, *U.S.S.R.*, 1943, Vol. 7, No. 6, pp. 308–309.) A brief description of experiments with waveguides using wavelengths of 8–11 cm. Methods are considered whereby the type of standing wave may be transformed by the introduction of sharp bends and branches into the waveguides. Typical transformations are: an E_0 type into an H_1 type, or H_0 into H_1, and *vice versa*.

621.392 : 534.213.4 850

The Analysis of Plane Discontinuities in [acoustic] **Cylindrical Tubes.**—Miles. (*See* 815 & 816.)

621.396.67 851

Concerning Hallén's Integral Equation for Cylindrical Antennas.—S. A. Schelkunoff. (*Proc. Inst. Radio Engrs, N.Y.*, Dec. 1945, Vol. 33, No. 12, pp. 872–878.) An explanation of the quantitative discrepancy between the impedance of a cylindrical aerial calculated by Hallén's formula (2763 of 1939) and that of the author (1049 of 1942). Hallén's solution involves the tacit assumption that the aerial is short compared with the wavelength as well as being thin, and this degrades the successive approximations. Gray's solution (1931 of 1944) does not involve this assumption, and gives satisfactory results. Infinitely long aerials are considered, and the results obtained compared with those of Stratton and Chu (1888 of 1941). An explanation of the discrepancy is given.

621.396.67 **852**

Principal and Complementary Waves in Antennas.—S. A. Schelkunoff. (*Proc. Inst. Radio Engrs, N.Y.*, Jan. 1946, Vol. 34, No. 1, pp. 23–32.) The paper is an extension of the analysis in 1049 of 1942 (Schelkunoff) to cylindrical antennas whose diameters are not small compared with the wavelength. " . . . the principal feature of the method is: waves in infinitely long antennas are considered first; subsequently, the complementary waves are included to express the effect of sudden termination of the wires, particularly with regard to uneven reflection at the wavefront passing through the end of the antenna." The effect of a finite source is discussed, and the simplifications when the antenna is thin are given.

621.396.67 **853**

On the Distribution of Current along Asymmetrical Antennas.—C. W. Harrison, Jr. (*J. appl. Phys.*, July 1945, Vol. 16, No. 7, pp. 402–408.) The first system analysed is a vertical grounded antenna over an infinitely conducting earth, fed at a point remote from the earth. This is treated as an isolated antenna fed symmetrically by two generators at points equidistant from the centre. The second system considered is an isolated linear antenna fed at a point not at the centre, and it is shown that the behaviour of the first system can be derived from that of the second. The basic equations for the current distributions are derived from the field equations, and approximate formulae suitable for computation are given. The results are fundamental to the solution of many advanced antenna problems.

621.396.67 **854**

Symmetrical Antenna Arrays.—C. W. Harrison, Jr. (*Proc. Inst. Radio Engrs, N.Y.*, Dec. 1945, Vol. 33, No. 12, pp. 892–896.) Calculation of the input impedance of a symmetrical aerial system consisting of identical radiators at the vertices of a regular polygon. The modified case of a symmetrical array with an additional central radiator is briefly discussed. Several practical applications of the theory are mentioned, particular reference being made to the case of corner-reflector aerials.

621.396.672 + 621.396.677 **855**

The Half-Rhombic Antenna.—J. H. Mullaney. (*QST*, Jan. 1946, Vol. 30, No. 1, pp. 28–31.) An elementary account of the construction and properties of the half-rhombic antenna, comprising an inverted-V aerial with a single-wire counterpoise.

621.396.674 : 621.314.2 **856**

Loop-Antenna Coupling-Transformer Design.—W. S. Bachman. (*Proc. Inst. Radio Engrs, N.Y.*, Dec. 1945, Vol. 33, No. 12, pp. 865–867.) "The low-impedance loop coupling-transformer circuit is analyzed, expressing the transformer parameters in terms of the circuit and transformer coupling coefficients. Equations are developed which yield optimum design values for the transformer. It is shown that an ideal transformer-coupled loop has 38.4 per cent. of the gain realizable from a direct-connected loop of the same area, assuming the same Q in the transformer secondary as in the direct-connected loop."

621.396.677 **857**

Small-Dimensional Radiating and Receiving Systems of High Directivity.—S. A. Makov. (*C.R. Acad. Sci. U.R.S.S*, 20th Aug. 1940, Vol. 28, No. 5, pp. 418–422.) Acoustic and radio systems, small compared with the wavelength, and with practically constant directivity over a range of frequencies, can be formed from a number of non-directional radiators by suitable adjustment of the phase and amplitude of their outputs. Three equally spaced collinear point sources, separated by $d/2$, the outer pair operated in equal amplitude and phase, the other adjusted in amplitude and phase to cancel the signal in the direction of the line of sources, give a polar diagram of the form $R_a \doteqdot \cos^2 \alpha$ for $0 < d/\lambda < 0.5$. Three-element rectilinear systems directed along the line of sources by making the centre source cancel the effect of the other two in the broadside direction give similar broad frequency bands of operation. The directivity can be improved by using more sources, while maintaining a wide frequency range of operation.

Combinations of such systems in perpendicular directions, or the use of point and annular sources give three-dimensional directivity.

621.396.677.2 + 621.396.71 **858**

Stepping Up from ¼ KW to 5 KW [and the use of a directional aerial].—Griffiths. (*See* 1057.)

621.396.67 + 621.396.11 **859**

Antenne e propagazione delle onde elettromagnetiche (per ingegneri) [Book Review].—B. Peroni. Michele dell'Aira, Rome, 372 pp. (*Wireless Engr*, Feb. 1946, Vol. 23, No. 269, p. 57.) "It can certainly be recommended to any advanced student or radio engineer with a knowledge of Italian or the desire to acquire such knowledge."

CIRCUITS

621.3.012.8 **860**

Determination of the Individual Resistances of a Delta Mesh by Measurements Taken at the Junctions.—L. J. Purnell. (*Elect. Engng, N.Y.*, Sept. 1945, Vol. 64, No. 9, p. 348.) Letter giving formulae.

621.314.2 **861**

Phase Splitter.—L. A. Saunders. (*Electronic Engng*, Feb. 1946, Vol. 18, No. 216, p. 63.) Letter describing an improvement in the phase-splitting circuit given in 295 of February, by the use of a double triode for phase splitting. This circuit will reject in-phase signals as in the Offner biological amplifier (1713 of 1937).

621.314.3 : 578.088.7 **862**

"Biological Amplifiers".—D. Robertson. (*Wireless World*, Jan. 1946, Vol. 52, No. 1, p.27.) Letter on 45 of January and 296 of February (Parnum) discussing reduction of mains-frequency interference.

621.385.032.442 : 621.3.025.4 863

Polyphase from Single Phase for Valve Supply Circuits.—E.M.I. Laboratories. (*Electronic Engng*, Feb. 1946, Vol. 18, No. 216, p. 55.) Circuits for obtaining polyphase supply for filaments of transmitting valves in order to reduce hum modulation.

621.385.2/.3].012.8 864

Triode Equivalent Circuits.—D. A. Bell. (*Electronic Engng*, Feb. 1946, Vol. 18, No. 216, pp. 56–57.) An ordered discussion on equivalent circuits for diodes and triodes, starting from first principles, and paying special attention to phase conventions. Voltage- and current - generator equivalents of amplifying triodes are compared.

621.392.43 865

Graphical Method of Determining the Resistance Transformation of Complex Resistances by means of a Capacitance assumed to be Loss-Free.—Geschwinde. (*Funktech. Mh.*, Sept./Oct. 1943, Nos. 9/10, pp. 89–94.) The purpose of the method is to find the value $-jX_T$ of the loss-free capacitive reactance that must be connected in shunt with a particular impedance $R + jX$ in order that the reactive or resistive part of the equivalent series impedance, $r - jx$, of the combination shall have an assigned value. Formulae are derived and charts are given connecting X/X_T with $R/|X_T|$ and x/X_T with $r/|X_T|$ for different values of R/r between 0.01 and 10. Charts are also given connecting $\pm X_T/X$ with R/r and x/X for different values of $R/|X|$ between 0.01 and 10^4. Examples of the use of these charts for matching *e.g.* an aerial to a transmitter are worked out.

621.392.5 866

General Formulas for "T"- and "Π"-Network Equivalents.—M. B. Reed. (*Proc. Inst. Radio Engrs, N.Y.*, Dec. 1945, Vol. 33, No. 12, pp. 897–899.) ". . . the development of two sets of general formulas which determine a set of "T" or "Π" impedances equivalent to any linear, lumped-constant, four-terminal network."

621.392.5 : 621.316.722.078.3 867

The Theory of the Non-Linear Bridge Circuit as applied to Voltage Stabilizers.—G. N. Patchett. (*J. Instn elect. Engrs*, Part III, Jan. 1946, Vol. 93, No. 21, pp. 16–22.) Analysis of four arrangements of bridge elements obeying the law $V = kI^n$. Formulae are tabulated giving dV_{output}/dV_{input} for various circuit conditions, and giving circuit parameters necessary for maximum power output. A review of the properties of different bridges indicates the use of two non-linear elements in opposite arms to be the best.

621.392.5 : 621.317.725 : 621.385 868

Bridge Circuits with a Non-Linear Element: Valve-Voltmeter with a Stable Zero Adjustment.—Levy. (*See* 976.)

621.392.52 + 621.318.72 869

"Abacs for Filter Design".—T. Roddam. (*Wireless World*, Jan. 1946, Vol. 52, No. 1, p. 31.) Correction of error in 52 of January.

621.392.52 870

Transient Response of Filters [part 1].—D. G. Tucker. (*Wireless Engr*, Feb. 1946, Vol. 23, No. 269, pp. 36–42.) "It is shown that the normal treatments of the transient response of filters are either not sufficiently simple or are not really suitable for all types of filter; the method which considers the filter as an idealized transmission characteristic is satisfactory for multi-section filters and gives results in simple terms, but for single-section filters a special treatment is desirable. A method of determining the build-up and decay characteristics is given for a single section of the "6-element symmetrical" band-pass filter, and the type of equation obtained is of general application to other types of section." The treatment of Carson and Zobel (*Bell Syst. tech. J.*, July 1923, pp. 1–52.) is compared with the idealized transmission characteristic approach. Only band-pass filters are considered, but it is shown that the analysis is, for most purposes, also applicable to low-pass filters. The first of two parts.

621.392.52 871

Filter Design Tables Based on Preferred Numbers: Low-Pass Filters.—H. Jefferson. (*Wireless Engr*, Jan. 1946, Vol. 23, No. 268, pp. 26–28.) Tables are given to facilitate the design of low-pass filters using preferred (standard) values of components. See also 3823 of 1945.

621.392.52 872

Double-Derived Terminations.—R. O. Rowlands. (*Wireless Engr*, Feb. 1946, Vol. 23, No. 269, pp. 52–56.) "A method is described for connecting filters in series or parallel such that the real part of their impedance or admittance function is equivalent to that of a double *m*-derived section."

621.392.52 + 621.318.7 873

Technical Applications of Filters.—G. Guanella. (*Schweiz. Arch. angew. Wiss. Tech.*, Jan. 1945, Vol. 11, No. 1, pp. 16–26.) Describes a wide range of types of filter construction for a wide range of frequencies and applications: use of sliding cylindrical shields as a means of varying the inductance of solenoidal coils: filters embodying quartz crystals, and the great advantages of calcium phosphate in this connexion: resonant cavity filters for use in the decimetre wave range: the use of solenoidal bi-filar windings as a means of discriminating between symmetrical and asymmetrical currents. The article concludes with a brief account of the voltage-distribution method of measuring impedance at very high frequencies.

621.394/.397].645 874

Cathode Bias: Design Data (1).—(*Wireless World*, Jan. 1946, Vol. 52, No. 1, pp. 21–22.) Formulae for calculating the frequency response of valves with automatic bias.

621.394/.397].645 875

Cathode-Follower Circuits.—K. Schlesinger. (*Proc. Inst. Radio Engrs, N.Y.*, Dec. 1945, Vol. 33, No. 12, pp. 843–855.) Analysis of the cathode-follower as an amplifier, an oscillator, or as a discriminator for detection of f.m. signals. The gain/frequency characteristics and bandwidth are calculated for various circuital arrangements and various coupling networks in an amplifier. The input capacitance and conductance are derived as functions of the cathode load. The condition for oscillation is deduced, and the design of a balanced oscillator up to 700 Mc/s is described. Used as a discriminator the cathode follower, though more elaborate than the conventional double-diode, has the advantages of a limiting action, of higher input impedance, and lower output impedance.

621.394/.397].645 **876**

Amplifier Circuit.—J. R. Ford. (*Electronic Industr.*, Oct. 1945, Vol. 4, No. 10, p. 128.) To eliminate distortion without reduction in gain, the grid of a second tube is connected to a point of zero signal potential but appreciable distortion potential, and the amplified output combined in antiphase with the output of the original tube. Summary of U.S. patent 2 379 699.

621.395.645 : 621.395.44 **877**

A New Method of Amplifying with High Efficiency a Carrier Wave Modulated in Amplitude by a Voice Wave.—S. T. Fisher. (*Proc. Inst. Radio Engrs, N. Y.*, Jan. 1946, Vol. 34, No. 1, pp. 3–13.) Description and analysis of a new amplifier circuit. The input wave is divided into sections on an amplitude basis, and after high-efficiency amplification of each section, recombination reproduces the original wave form. The circuit employs valves with suitable separate grid drives, biases and load impedances; in practice 3 or 4 branches give satisfactory compromise between efficiency and complexity. The technique is specially suitable for controlled-carrier systems. Efficiencies about 5 times better than with existing systems are obtainable, and the power dissipation of the output stage is reduced, the advantages being greatest at low modulation levels.

621.395.645.36 **878**

Quality Amplifiers.—(*Wireless World*, Jan. 1946, Vol. 52, No. 1, pp. 2–6.) Details are given of 4, 8 and 12W circuits with double push-pull resistance-triode stages.

621.395.645.36 **879**

Push-Pull Circuit Analysis: Cathode-Coupled Output Stage.—S. W. Amos. (*Wireless Engr*, Feb. 1946, Vol. 23, No. 269, pp. 43–46.) "This circuit is one in which one output valve derives its input by cathode-coupling from the other . . ." "It is the purpose of this analysis to find the effect of varying R_0 [the resistance connecting the junction of the two grid leaks to h.t. negative] on the performance of the circuit and hence to find . . . its optimum value."

621.395.665 **880**

Resistive Attenuators, Pads and Networks: Parts IX and X.—P. B. Wright. (*Communications*, Oct. 1945, Vol. 25, No. 10, pp. 72..106, & Nov. 1945, Vol. 25, No. 11, p. 61.) IX. Consideration of representative examples. X. Particulars of a four-channel series-parallel mixer and fader system. Series concluded. For previous parts see 60 of January and back references.

621.396.615 : 621.394/.397].645.34 **881**

Self-Excitation and Regeneration.—K. Eisenmann. (*Funktech. Mh.*, Sept./Oct. 1943, Nos. 9/10, pp. 95–100.) There can be no doubt that the equating to unity of the product of the amplification (V) and back-coupling factor (K) adequately represents the condition of stationary self-excited oscillations, but it is not immediately apparent that the condition must *necessarily* be fulfilled, especially if an exponentially damped oscillation is introduced as the initiator of the self-maintained oscillation, the frequency of the initiator being not necessarily the same as that of the self-maintained oscillation. The build-up of oscillation is traced in this paper to determine the frequency of the self-maintained oscillation from the frequency of the initiator. Conditions of stability in externally excited regenerative amplifiers are first examined. For similar work by the same author see 3122 of 1944. Self excitation is then considered when a parallel resonant circuit is connected into the regenerative system. It is assumed that switching on the d.c. supply charges the capacitor of the resonant circuit, which starts a damped oscillatory discharge. It is deduced that in addition to this decaying oscillation there is another oscillation that grows exponentially and becomes stationary when KV becomes unity. The amplitude of the stationary oscillation is greater than the initiating oscillation by a factor equal to the reciprocal of the logarithmic decrement of the resonant circuit. The frequency of oscillation differs, but always by very little, from the resonant frequency. A transient fluctuating distortion oscillation is superimposed on the main steady oscillation; it is smaller than the latter by the product of the period of the tuned circuit and its damping factor (R/2L).

621.396.615.072.9 **882**

Forced Oscillations in Oscillator Circuits, and the Synchronization of Oscillators.—D. G. Tucker. (*J. Instn elect. Engrs*, Part I, Jan. 1946, Vol. 93, No. 61, pp. 57–58.) Long abstract of 3825 of 1945.

621.396.615.14.029.63 **883**

A Coaxial Modification of the Butterfly Circuit.—E. E. Gross. (*Communications*, Dec. 1945, Vol. 25, No. 12, pp. 42–43.) The outer conductor of the coaxial cylinder has two 105° sections cut away, and two 75° thick metal sectors rotate between inner and outer conductors. Frequency range, with a lighthouse valve, is 620–1 340 Mc/s, power output 0.15 to 0.3 W. Report of an I.R.E. paper.

621.396.615.17 **884**

Low Frequency Multivibrators.—C. J. Quirk. (*Electronics*, Dec. 1945, Vol. 18, No. 12, pp. 350..360.) Very-low-frequency multivibrators may be obtained by omitting the grid leaks of the valves and relying on the leakage of the capacitors to discharge them. The periodicity is dependent on the quality of the capacitor and possibly on the leakage paths in the valves. Periods up to about 2 hours are discussed.

621.396.615.17 : 621.317.755 **885**

A Single Sweep Time Base.—D. McMullan. (*Electronic Engng*, Feb. 1946, Vol. 18, No. 216, pp. 47–50.) The time base was described in Part I of the paper (see 571 of March). The present paper gives details of the basic circuit and of practical operational points. Performance figures are stated and methods of use described in Part I are discussed in further detail.

621.396.615.17 : 621.317.755 **886**

Time-Base Converter and Frequency Divider.—H. Moss. (*Electronic Industr.*, Nov. 1945, Vol. 4, No. 11, pp. 158..160.) Long abstract with diagrams of 3266 of 1945. See also 308 of February.

621.396.615.17 : 621.317.755 **887**

Synchronous Time Base.—H. Moss. (*Electronics*, Nov. 1945, Vol. 18, No. 11, pp. 458..462.) Summary of 3266 of 1945. See also 308 of February.

621.396.616 + 621.392 + 621.317.763 888

The Transverse Electric Modes in Coaxial Cavities.—R. A. Kirkman & M. Kline. (*Proc. Inst. Radio Engrs, N. Y.*, Jan. 1946, Vol. 34, No. 1, pp. 14–17.) Study of the transverse modes TE $_{l,m,n}$ extending the treatment of Barrow and Mieher (2951 of 1940) and clarifying difficulties in their notation and pictorial representation. In particular the modes with $m = 0$ do not exist but are the limits of the modes for $m = 1$ when the radii of the conductors approach equality. The dependence of the modes on the ratio of the radii is discussed: for a given mode the resonant frequency decreases as the ratio approaches unity. The practical importance of coaxial transverse electric modes is stressed, *e.g.* in a wavemeter.

621.396.619 889

Single-Sideband Generator.—M. A. Honnell. (*Electronics*, Nov. 1945, Vol. 18, No. 11, pp. 166–168.) The system described avoids the use of highly selective filters for eliminating the unwanted sideband. Two separate balanced modulators are used; the carrier and modulation voltages applied to modulator 1 are respectively in quadrature with those applied to modulator 2. By suitably combining the modulator outputs either sideband may be selected. The system is of particular advantage at a fixed modulation frequency: a design of 90° phase shifter is described, however, which has a shift lying within 0.2° of 90° for all frequencies in the band 100 to 5 000 c/s.

621.396.621.52 890

The Super-Regenerative Detector: an Analytical and Experimental Investigation.—Strafford. (*See* 1032.)

621.396.622 : 621.396.619.018.41 891

A New F.M. Detector-Circuit [requiring no limiter].—G. G. (*QST*, Jan. 1946, Vol. 30, No. 1, pp. 26–27.) See 574 of March.

621.396.645.2 892

Pulse Amplifier Coupling.—S. Moskowitz. (*Communications*, Oct. 1945, Vol. 25, No. 10, pp. 58, 88.) An analysis of a method of calculating the response of resistance-capacity amplifier coupling circuits to pulse voltages.

621.396.645.3 : 621.396.621.55 : 621.396.61 893

A Coil-Neutralized Vacuum-Tube Amplifier at Very-High Frequencies.—R. J. Kircher. (*Proc. Inst. Radio Engrs, N.Y.*, Dec. 1945, Vol. 33, No. 12, pp. 838–843.) Neutralization of a triode is achieved (at one frequency) by tuning out the grid-anode capacitance by means of a parallel inductance; at v.h.f. an open λ/2 line can be used without the need for a blocking condenser. The neutralization so obtained is comparable with that existing in small pentodes. A 2-stage amplifier for 140 Mc/s using the Western Electric 299Y triode is described, with details of technique of adjustment and measurement of performance. A 3-db bandwidth of 4.3 Mc/s and a non-oscillating frequency range of 15.4 Mc/s are quoted. Measurements of distortion characteristics using a two-frequency generator and harmonic analyser showed that for a peak power output of 3 W the third-order distortion is – 44 db, and the fifth-order distortion better than – 60 db.

621.396.662 : 621.396.61 894

A Radio-Frequency Auto-Resonator.—Clemens. (*See* 1099.)

621.396.662.34 895

H.F. Band-Pass Filters: Part I.—H. P. Williams. (*Electronic Engng*, Jan. 1946, Vol. 18, No. 215, pp. 24–26.) Theoretical properties in a form suitable for engineers designing radio circuits.

621.396.662.34 896

H.F. Band-Pass Filters: Part II. – Similar Circuits.—H. P. Williams. (*Electronic Engng*, Feb. 1946, Vol. 18, No. 216, pp. 51–54.) A simple method for studying the characteristics of band-pass filters, making use of the fact that they have Q values of 20 and upwards. An approximation is made in the expression for series impedance near resonance, and a parameter for distance off tune is introduced. Various types of coupling are discussed. For part I see 895 above.

621.396.665 897

Radio Design Worksheet: No. 42. – Some Notes on Automatic Volume Control; Parallel Resonance.—(*Radio, N. Y.*, Nov. 1945, Vol. 29, No. 11, pp. 51–52.) Application of simple time-constant theory to the design of orthodox a.v.c. circuits.

621.396.822 898

[Noise] **Fluctuations of Electric Current.**—Bell. (*See* 1038.)

621.396.822 899

A Theory of Valve and Circuit Noise.—Campbell & Francis. (*See* 1037.)

621.397.62 900

The DC Restorer: Its Uses in Television and Radar.—(*Wireless World*, Jan. 1946, Vol. 52, No. 1, pp. 8–11.) A descriptive article.

621.397.813 901

Theoretical Investigation of the Distortion of Television Signals in Valve Circuits.—J. Huber. (*Schweiz. Arch. angew. Wiss. Tech.*, Mar. 1945, Vol. 11, No. 3, pp. 77–86.) Essentially a theoretical analysis of the response of a typical circuit consisting of a valve with a 4-pole network in the anode circuit, to the kind of excitation which occurs in the reception of the modulation corresponding to the usual types of test-picture. The method of analysis is the use of the Laplace transform in the form $f(p) = p\int_0^\infty \epsilon^{-pt} F(t)dt$, following Wagner. The results of the analysis are applied in detail to particular types of anode-circuit network, *e.g.* a resistance-capacitance amplifier, and a band-pass filter network, the equations of which are expressed in matrix form, following Feldtkeller. To be continued.

621.3.01/.02 902

Elementary Electric-Circuit Theory [Book Review].—R. H. Frazier. McGraw-Hill, New York & London, 1945, 434 pp., $4.00. (*Nature, Lond.*, 5th Jan. 1946, Vol. 157, No. 3975, p. 4.) "... the author has gone so thoroughly into basic circuit theory that the term 'elementary' cannot be justified." "A very real and useful approach to the relationship between the fundamental essentials ... and their applications. ..." See also 69 of January.

GENERAL PHYSICS

530.145.6 : 537.212 **903**

On Different Ways of Representing a Homogeneous Electric Field in Wave Mechanics.—F. Sauter. (*Ann. Phys., Lpz.*, 1943, Vol. 43, Nos. 6/7, pp. 404–416.) A mathematical discussion of the wave-mechanical analogy of the non-uniqueness of the scalar and vector potentials corresponding to the electric and magnetic field vectors of the classical theory.

534.321.9 **904**

New Aspect of Ultrasonics.—B. K. Sahay. (*J. acoust. Soc. Amer.*, Jan. 1946, Vol. 17, No. 3, pp. 285–286.) The study of ultrasonics may reveal a corpuscular nature of sound energy.

535.14 + 537.122 **905**

A New Classical Theory of Photon and Electron.—B. M. Sen. (*Sci. Culture*, Jan. 1946, Vol. 11, No. 7, pp. 387–388.) Short preliminary report of a theory designed to overcome deficiency in the Einstein equation $W = h\nu$ associating a quantum $h\nu$ with every e.m. wave, which is regarded as an over-generalization.

537.122 **906**

The Size of an Electron and the Nature of its Mass.—G.W.O.H. (*Wireless Engr*, Feb. 1946, Vol. 23, No. 269, pp. 33–35.) An editorial in which the radius of the electron is derived on the assumption that its apparent mass is inertial and due to its magnetic field. The charge may be considered to be distributed either uniformly over its surface or uniformly throughout its volume, leading respectively to values of 1.88 and 2.256×10^{-13} cm for the electronic radius.

537.212 **907**

Erratum: A Generalization of the Dielectric Ellipsoid Problem.—R. Clark Jones. (*Phys. Rev.*, 1st/15th Nov. 1945, Vol. 68, Nos. 9/10, p. 213.) Correction to 4068 of 1945.

537.22 **908**

The Basic Mechanisms of Static Electrification.—L. B. Loeb. (*Science*, 7th Dec. 1945, Vol. 102, No. 2658, pp. 573–576.) A short review of processes which produce static electricity. The five mechanisms discussed are (a) electrolytic effects, (b) contact electrification between surfaces, (c) spray electrification, (d) frictional electrification, (e) electrical effects in gases and flames. The need is stressed for further work to clarify the exact mechanism involved in each of these processes.

537.52 **909**

On the Mechanism of Spark Formation.—S. Teszner. (*C. R. Acad. Sci., Paris*, 5th & 19th March 1945, Vol. 220, Nos. 10 & 12, pp. 307–309 & 390–392.) An attempt to explain the discrepancy between the time observed to be required for a spark to build up in a gas (order of 10^{-8} sec) and the time implied by Townsend's theory (10^{-4} – 10^{-5} sec), in terms of the effect of negatively charged particles that are in front of and are carried along by the avalanche.

537.523.2 **910**

On the Velocity-Distribution of Electrons in [cold-cathode] **Field Emission.**—Müller. (*See* 1104.)

537.529 + 621.315.611.015.5 **911**

The Electrical Breakdown of Solid Dielectrics.—Krassin. (*See* 948.)

537.533.8 **912**

On Secondary Emission from Films of Pure Metals in the Ordered and Unordered States, and their Penetrability by Secondary Electrons.—Suhrmann & Kundt. (*See* 939.)

538.32 : 621.385.832 **913**

Another Problem of Two Electrons.—G.W.O.H. (*Wireless Engr*, Jan. 1946, Vol. 23, No. 268, pp. 1–2.) A discussion of an alleged fallacy relating to the forces between two electrons following parallel paths in a c.r. tube beam. It is shown that the apparent fallacy arises from a misinterpretation of the idea of relative velocity as between the electrons and the remainder of the system and an observer, and detailed analysis of the system as a unit shows the behaviour to be quite self-consistent. See also 2765 of 1944.

621.385.833 **914**

Electron Beams in Strong Magnetic Fields.—Pierce. (*See* 1112.)

535.22 **915**

The Velocity of Light [Book Review].—N. E. Dorsey. The American Philosophical Society, Philadelphia, 1944, 110 pp., $2.50. (*Rev. sci. Instrum.*, Oct. 1945, Vol. 16, No. 10, pp. 288–289.) " . . . will undoubtedly serve for many years as a convenient reference book on the subject." See also 590 of March.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.5 : 621.396.82 **916**

Listening in on the Stars.—O. G. Villard, Jr. (*QST*, Jan. 1946, Vol. 30, No. 1, pp. 59..122.) The short-duration whistles heard in certain circumstances when listening to a short-wave transmitter are explained in terms of interference introduced by radiation reflected from meteor trails. Methods of recognizing and observing the whistles are described. See also 3408 of 1944.

523.72 : 621.396.822 **917**

Signals from the Sun.—E. V. Appleton. (*Wireless World*, Jan. 1946, Vol. 52, No. 1, p. 29.) Article based on 323 of February (Appleton).

551.51.053.5 : 550.38 : 621.396.11 **918**

On the Influence of the Terrestrial Magnetic Field on the Reflection of Radio Waves from the Ionosphere.—Ginsburg. (*See* 1026.)

551.55 **919**

Wind Measurements at 30 Km.—N. K. Johnson. (*Nature, Lond.*, 5th Jan. 1946, Vol. 157, No. 3975, p. 24.) Letter reporting measurements on bursting smoke shells from a high-velocity gun in S.E. England. The wind is mainly easterly in summer with a mean velocity of 12 metres per second, and it is westerly in winter with a mean velocity of 37 metres per second.

551.5 **920**

Descriptive Meteorology [Book Review].—H. C. Willett. Academic Press, Inc., U.S.A., 1944, 310 pp, $4.00. (*Science*, 14th Dec. 1945, Vol. 102, No. 2659, pp. 624–625.) "... one of the few meteorological texts designed for use in undergraduate courses where students have had preparation in college mathematics and physics . . . indeed a welcome contribution . . ."

LOCATION AND AIDS TO NAVIGATION

621.3(41) "1939/1945" **921**

"British Electrical Engineers and the Second World War".—P. Dunsheath. (*J. Instn elect. Engrs*, Part I, Jan. 1946, Vol. 93, No. 61, pp. 17–27.) The full paper, of which summaries were noticed in 415 of February and 729 of March.

621.385 + 621.396/.397 + 621.365.5 **922**

Results Achieved by a Large [French] **Industrial Research Centre** [Le Groupe Industriel des Compagnies Françaises de T.S.F. Associées] **During and Despite the Occupation.**—Brenot. (*See* 1105.)

621.396.9 **923**

Hyperbolic Navigation.—(*Electrician*, 1st Feb. 1946, Vol. 136, No. 3531, pp. 294–296.) Summary of a joint meeting of the Institution of Electrical Engineers and the Institute of Radio Engineers. . A. Pierce read a paper, over short-wave radio rom America, on "An Introduction to Hyperbolic Navigation". A discussion followed, a part of which was also heard in both countries.

21.396.9 **924**

"Loran" Guides Pilots.—W. Davis. (*Sci. News Lett., Wash.*, 3rd Nov. 1945, Vol. 48, No. 18, p. 275–276.) A non-technical account of the istory of Loran. Peacetime applications and costs re discussed.

21.396.9 **925**

Specialized Loran.—(*Sci. News Lett., Wash.*, 4th Nov. 1945, Vol. 48, No. 21, p. 324.) Sky-wave ynchronized (S.S.) Loran has an over-land range of bout 1 400 miles after sunset, compared with 00 miles obtained by using the ground wave. rief note only.

1 396.9 **926**

Loran — the Latest in Navigational Aids : Part I, undamental Principles : Part II, Ground Station quipment.—A. A. McKenzie. (*QST*, Dec. 1945, ol. 29, No. 12, pp. 12–16, & Jan. 1946, Vol. 30, o. 1, pp. 54–57.)

1.396.9 **927**

Fundamentals of Radar : 4. — Pulse Methods pplied to Navigation.—(*Wireless World*, Jan. 46, Vol. 52, No. 1, pp. 23–26.) An outline of the stem using three fixed transmitting stations and a obile receiver (Gee), and of the system using two ed transmitting-receiving stations and a mobile sponder (Oboe).

1.396.9 **928**

The Radar.—S. K. Mitra & J. N. Bhar. (*Sci. lture*, Jan. 1946, Vol. 11, No. 7, pp. 343–353.) semi-technical account of the principles and plication of radar and various new aids to vigation. For Part 3 see 328 of February.

1.396.9 **929**

Manufacturer's Radar Display.—(*Electrician*, 14th c. 1945, Vol. 135, No. 3524, pp. 659–660.) scription of an exhibition by Ferranti at Moston.

621.396.9 **930**

Elements of Radar : Part I, Radar Pulses.—J. McQuay. (*Radio Craft*, Dec. 1945, Vol. 17, No. 3, pp. 169..204.)

621.396.9 **931**

Army and Navy Radar.—(*Electronic Industr.*, Oct. 1945, Vol. 4, No. 10, pp. 98–99.) Photographs of operational radar installations in the field and on board ship.

621.396.933.1 **932**

An Ultra-High-Frequency Radio Range with Sector Identification and Simultaneous Voice.—A. Alford, A. G. Kandoian, F. J. Lundburg & C. B. Watts, Jr. (*Waves and Electrons*, Jan. 1946, Vol. 1, No. 1, pp. 9–17.) A combination of two perpendicular aerial arrays, each consisting of three horizontal loops in line, the centre one common, is used to give aircraft course indication with unambiguous sector identification. One array gives an aural course by means of interlocked U and D morse patterns, using 1 020 c/s modulation frequency. The other provides a perpendicular visual course by the simultaneous transmission of 90 c/s and 150 c/s modulation, indication being by a centre-zero meter. The carrier (125 Mc/s) with voice modulation is radiated from the centre loop only. Earlier schemes and developments leading up to the final system are also described.

MATERIALS AND SUBSIDIARY TECHNIQUES

53.081.7(0 : 025.345) **933**

Card File of Physical Constants of Substances ["Annual Tables of Physical Constants"].—(*Nature, Lond.*, 29th Dec. 1945, Vol. 156, No. 3974, p. 775.) Numerical data of physical properties of substances are printed on 5 × 3 inch library cards, with one or more cards for each substance. It is planned to issue about 2 000 substance cards per year.

533.5 **934**

A Useful Seal for Dynamic Vacuum Systems.—D. D'Eustachio. (*Rev. sci. Instrum.*, Dec. 1945, Vol. 16, No. 12, pp. 377–378.) A "fairly soft" rubber stopper under compression seals electrodes or glass tubing, of up to several inches diameter, down to pressures of 10^{-6} mm of mercury.

533.5 **935**

Steel-to-Glass Vacuum Seal.—(*Rev. sci. Instrum.*, Oct. 1945, Vol. 16, No. 10, p. 298.) Note on a new R.C.A. technique due to G. R. Shaw.

533.5 **936**

The Hydrogen Gauge — An Ultra-Sensitive Device for Location of Air Leaks in Vacuum - Device Envelopes.—H. Nelson. (*Rev. sci. Instrum.*, Oct. 1945, Vol. 16, No. 10, pp. 273–275.) "... a sealed-off, highly evacuated ionization gauge having a section of its envelope made of thin palladium sheet which, when heated is highly permeable to hydrogen. In use the gauge is connected to the manifold of a vacuum system with the palladium section isolating the vacuum of the gauge from that of the manifold. When the manifold or any vacuum device connected to it is probed with hydrogen, leaks in the system are indicated by an increase in the ion current of the gauge." Leaks greater than 10^{-4} litre-micron per second are detected.

533.5 : 621.315.626 937
Ferranti Vacuum-Seal Bushing.—(*See* 1074.)

533.56 938
Small Oil Diffusion Type Vacuum Pump.—C. H. Bachman. (*Rev. sci. Instrum.*, June 1945, Vol. 16, No. 6, p. 153.) Description and drawing of a two-stage glass pump occupying a space about 5 × 6 × 2 inches. Speed 2 L/sec.

537.533.8 939
On Secondary Emission from Films of Pure Metals in the Ordered and Unordered States, and their Penetrability by Secondary Electrons.—R. Suhrmann & W. Kundt. (*Z. Phys.*, 23rd Feb. 1943, Vol. 120, Nos. 5/6, pp. 363–382.) Thin films of copper, gold, silver, and beryllium were condensed on to a glass plate in a vacuum at 83°A, and their secondary emission characteristics were studied, before and after raising to room temperature to convert the films from the unordered to the ordered state. Their penetrability was investigated by condensing the films on to a cooled highly-emitting beryllium cathode. It is concluded that the decrease in the secondary emission of compact metal films on passing from the ordered to the unordered states is only partially due to increased penetration by primary electrons. It must further be assumed that there are fewer absorbing and emitting centres in the ordered state.

539.234 940
Techniques for Evaporation of Metals.—L. O. Olsen, C. S. Smith & E. C. Crittenden, Jr. (*J. appl. Phys.*, July 1945, Vol. 16, No. 7, pp. 425–434.) Heaters are used in the form of open helical wire coils, close wound conical wire baskets, or ceramic crucibles formed on wire baskets. Evaporation has been carried out at temperatures up to 2 000°C and at pressures below 5×10^{-5} mm of mercury. Thirty-four elements have been studied for evaporation behaviour, and the preferred techniques for small and large quantities of each metal are tabulated.

539.234 : 621.396.611.21 941
Quartz Crystal Plating.—H. G. Wehe. (*Bell Lab. Rec.*, Dec. 1945, Vol. 23, No. 12, pp. 475–479.) Silver, aluminium or gold is deposited by evaporation onto the quartz in a high vacuum. The best distance of the quartz from the plating metal is dependent on a number of factors, which are discussed.

620.197 942
Surface-Conversion Coatings.—G. W. Jernstedt. (*ASTM Bull.*, Dec. 1945, No. 137, pp. 29–35.) Methods of producing protective surface coatings on metals by chemical reaction are described with details of the thickness of the films produced and their resistance to corrosive treatments. Most conversion coatings form a good base for protective lacquer or paint.

620.199.1/.9 943
Interference-Free Weatherometer.—W. B. R. Agnew. (*Electronics*, Dec. 1945, Vol. 18, No. 12, pp. 160–161.) " Accelerated weathering is applied to plastic and rubber materials and measured with photoelectric equipment. Automatic cycling of wetting and irradiation is provided and no radio interference is caused. The action is about ten times faster than outdoor exposure."

621.314.632 : 546.289 944
Germanium Rectifiers.—(*Electronic Engng*, Feb. 1946, Vol. 18, No. 216, p. 50.) Describes the use of germanium crystals as contact rectifiers for u.h.f. They may also be useful for voltage regulation.

A brief account of an I.R.E. paper by E. Cornelius, derived from a report in *Electronic Industr.*, Dec. 1945, Vol. 4, No. 12, p. 80.

621.315.53/.54] : 621.315.68 945
Contact Connections in Aluminium and Magnesium Conductors.—R. Schulze & B. Zeisz. (*Z. Ver. dtsch. Ing.*, 8th June 1940, Vol. 84, No. 23, pp. 399–402.) At 300° the oxide layer on aluminium is very loosely attached and easily removed, and aluminium wires can easily be coated with a metal that melts at about this temperature ; or a short length of " Cupal " (copper plated aluminium wire) can be welded on to the end of the aluminium conductor. Wires so prepared can be soldered. Screw terminals or clamps for aluminium conductors must be of spring construction to allow for the flow of the aluminium. The junction of stranded aluminium conductors presents special difficulties and usually requires welding into special end-pieces. Strips can be joined by screwing or clamping, with a paste containing bronze or similar turnings which are driven into the contact surface by the pressure of the connexion. A similar technique can be used with magnesium strip.

621.315.61 946
Study of the Electrical Conductivity of Rare Oxides : Lanthanum, Praeseodymium, Neodymium and Samarium.—M. Foëx. (*C.R. Acad. Sci., Paris*, 12th March 1945, Vol. 220, No. 11, pp. 359–361.) Graphs are given showing the resistivities versus temperature of La_2O_3, Pr_2O_3, Pr_8O_{11}, Nd_2O_3 and Sm_2O_3 in atmospheres of air, oxygen, nitrogen and hydrogen, over the range of about 300–1200°C. The resistivities lie between 10^8 and 10^4 Ω.cm except for Pr_8O_{11} in oxygen, for which it is ess than about 10^2 Ω.cm.

621.315.611 : 621.317.3 947
Resonance Methods of Dielectric Measurement at Centimetre Wavelengths.—Horner, Taylor, Dunsmuir, Lamb & Jackson. (*See* 966.)

621.315.611.015.5 + 537.529 948
The Electrical Breakdown of Solid Dielectrics.—A. K. Krassin. (*C.R. Acad. Sci., U.R.S.S.*, 20th Sept. 1940, Vol. 28, No. 8, pp. 694–696. In German.) Consideration of the reason for the destruction of the material by the discharge. The appearance of a large number of electrons in motion in the insulator causes a loosening of the bond between the nodal particles of the lattice; thus the electric field producing the electron current can finally destroy the connexion between the heavy charged particles that has already been modified by the electrons. The breakdown field alone would not destroy the lattice.

621.315.612.6.017.143.029.58/.62 949
Dielectric Losses of Different Glasses in the Short Wave Region and their Dependence on Temperature.—M. J. O. Strutt & A. van der Ziel. (*Physica, Eindhoven*, July 1943, Vol. 10, No. 7, pp. 445–450.) Results are given of measurements on heavy lead glass, soda glass (16% Na_2O) and boron glass (23% B_2O_3) at frequencies of about 12 and 50 Mc/s. Values of tan δ are plotted against

temperature over the range 50–250°C. Earlier results at frequencies between 500 and 3.10^5 c/s are plotted for comparison. At the higher frequencies log tan δ is no longer proportional to temperature; the rate of change of log tan δ with temperature increases with temperature. The dielectric losses and permittivity vary less with temperature as the frequency is increased. The dielectric losses and change in permittivity are greater for the soda glass than for the other two.

The existence of this paper was noted in 1685 of 1944.

621.315.616.98 **950**

"Frozen" Electricity: The Electret.—T. A. Dickenson. (*Radio Craft*, Nov. 1945, Vol. 17, No. 2, pp. 88–89.) Manufacture, properties, and uses of wax blocks with permanent electric polarization similar to the permanent magnetism of magnets.

621.318.22 **951**

Physical Basis and Properties of Modern Magnet Steels and the Construction and Testing of Permanent Magnets.—E. Spahn. (*Schweiz. Arch. angew. Wiss. Tech.*, Oct. 1944, Vol. 10, No. 10, pp. 313–322.) A comprehensive survey covering the relation between magnetic characteristics and micro-structure, for pure metals and alloys, the properties of permanent magnets, the characteristics of modern magnet steels, method of measurement of magnetic characteristics, and methods of magnetization.

621.318.322 **952**

Hipersil - A Greatly Improved Transformer Iron.—T. D. Yensen. (*J. appl. Phys.*, July 1945, Vol. 16, No. 7, pp. 379–385.) Hipersil is a silicon-iron alloy with 3–4% silicon, in which hysteresis loss has been reduced by the removal of impurities, increase in grain size, and arrangement of the crystals to their optimum orientation. Losses are roughly 50% lower and flux densities 20–30% higher than in standard silicon-iron sheet for the same magnetizing force. Properties are shown graphically.

621.319.7 : 621.315.626 **953**

Graphical Field Mapping Applied to Bushing Insulator Design.—E. G. Wright & S. Deutscher. (*Beama J.*, Jan. 1946, Vol. 53, No. 103, pp. 29–37.) Following Kuhlmann (*Arch. Elektrotech.*, 1915, Vol. 3, pp. 203–225.) the case of three-dimensional fields with rotational symmetry for several dielectrics in series is developed for the practical uses of solid and oil-filled bushings.

621.791.76 **954**

A Simple Method for Spotwelding Thin Wires.—L. Copeland & K. Rothschild. (*Rev. sci. Instrum.*, Oct. 1945, Vol. 16, No. 10, pp. 293–294.) A capacitor charged from a potentiometer connected across a 110 V d.c. supply provides the energy to weld the wires. The energy required can be adjusted to optimum by the potentiometer tapping. The capacitance used is of the order of 100 μF depending on the size of the wires to be welded and the strength of the weld required.

669–181.32 **955**

On the Structural Changes in Metals by Cold Working.—W. Kranert & H. Raether. (*Ann. Phys., Lpz.*, 1943, Vol. 43, Nos. 6/7, pp. 520–537.) A very thorough investigation, by electron diffraction, of the effects of various kinds of cold working (polishing, pressing, rolling, etc.) on single-crystal surfaces of copper, gold, iron and nickel. (The subject has some bearing on the behaviour of sliding contacts and metal surfaces). The authors' conclusions are:—1. "... under the specified cold-working conditions, the surface of the copper is transformed into an extremely finely crystallized layer, a few tens of Ångstrom units in thickness, which, with increasing depth, gradually changes to a normal crystalline powder structure, and finally, at a depth of a few microns, reverts to the undisturbed mono-crystal form. 2. ... the diagrams indicate that the disruption of the structure can be carried down to less than one or two 'elementary cells' [*Elementarzellen*], so that no distinction can be drawn between the fine-crystal and the quasi-fluid conditions. ... In certain cases (for example, copper, iron, nickel) a thin fine-crystal oxide sheet covers the worked surface."

679.5 **956**

Plastics for the Amateur: Part I.—A. G. Chambers. (*R.S.G.B. Bull.*, Dec. 1945, Vol. 21, No. 6, pp. 85–87.) Details of material characteristics, uses, cementing methods, etc., of a number of plastics, including phenol-formaldehyde; urea-formaldehyde; methyl methacrylate ("perspex" etc.); polyethylene; polyvinyl chloride; polystyrene, etc.

679.5 **957**

A Review of Plastic Materials.—H. L. Brouse. (*Proc. Inst. Radio Engrs, N.Y.*, Dec. 1945, Vol. 33, No. 12, pp. 825–834.) "Common plastic materials are classified according to their thermal characteristics and the methods of fabrication. Typical physical and electrical properties are presented for each of the general class of materials. It is pointed out that such values are average, and do not indicate the range that can be obtained by selection of a specific molding powder. Common trade names, compositions, and their manufacturers are presented in the form of a cross-index for reference purposes. Factors governing the performance and the choice of a particular material are discussed. Expansion coefficients, as applied to the use of metal inserts molded into the material, light transmission properties, and the fabrication of laminates are briefly treated in the Appendix."

679.5 : 539.4 **958**

Strength and Permissible Loading of Polyvinyl Chloride Plastics.—W. Buchmann. (*Z. Ver. dtsch. Ing.*, 22nd June 1940, Vol. 84, No. 25, pp. 425–431.) A detailed discussion of the use of such materials as structural members, with particular reference to Igelit PCU (Vinidur). The chief point emphasized is that the duration of the stress has a marked effect, as in metals at high temperature, and that reliable data must therefore be based on very long period tests, lasting in some cases more than 1 000 hours. The paper gives data for a wide range of stresses and temperatures.

679.5 : 621.396.6 **959**

Plastics in Radio.—L. Laden. (*Radio News*, Dec. 1945, Vol. 34, No. 6, pp. 25–28, 159.) The properties and application of thermo-setting plastics and of the newer thermoplastic materials are described. Tables give dielectric constant, power factor, breakdown voltage, and resistivity of the important plastics, and indicate suitable materials and manufacturing methods for a large range of radio parts.

679.5 **960**
Modern Plastics [Book Review].—H. Barron. Chapman & Hall, London, 1945, 680 pp., 42s. (*Electronic Engng*, Jan. 1946, Vol. 18, No. 215, p. 31.) See also 359 of February.

679.5 **961**
The New Plastics [Book Review].—H. R. Simonds, M. H. Bigelow & J. V. Sherman. D. Van Nostrand Co., New York, 1945, 320 pp., $4.50. (*Science*, 7th Dec. 1945, Vol. 102, No. 2658, p. 600.) " . . . a semi-technical volume devoted almost exclusively to plastic materials, processes and uses which have been developed since 1940." " . . . an excellent attempt to condense the more important information into one volume."

MATHEMATICS

517.432.1 **962**
Symbolic Calculus and Discontinuities.—M. Bayard. (*Rev. gén. Élect.*, Dec. 1945, Vol. 54, No. 12, pp. 373–379.) It is shown that any uniform function, even if it has discontinuities of the first or second kind, can be rendered continuous by representation in terms of an integration of unit or unit-impulse functions. The procedure also leads to a criterion for the validity of Heaviside's asymptotic development of the image-function illustrated by reference to a classical case in which Heaviside's expansion is invalid.

517.93 **963**
Nonlinear Springs.—M. A. Sadowsky. (*J. Franklin Inst.*, Dec. 1945, Vol. 240, No. 6, pp. 469–476.) A new method for solving the nonlinear oscillatory equation $\ddot{y} + S(y) = 0$, giving an implicit presentation of the solution. Certain forms of $S(y)$, containing an arbitary parameter, are suitable for curve fitting and give solutions in terms of elementary functions. The simpler solutions are tabulated.

621.396.11 + 534.231 **964**
The Saddle-Point Method in the Vicinity of a Pole with Applications to Wave-Optics and Acoustics.—Ott. (*See* 1024.)

MEASUREMENTS AND TEST GEAR

621.3.018.41 : 621.3.081.3 **965**
New Frequency Unit.—P. M. Honnell. (*Elect. Engng, N.Y.*, Nov. 1945, Vol. 64, No. 11, p. 422.) A suggestion for a logarithmic scale of frequency, the unit defined as 1 marconi = $\log_{10} f$, with f in c/s.

621.317.3 : 621.315.611 **966**
Resonance Methods of Dielectric Measurement at Centimetre Wavelengths.—F. Horner, T. A. Taylor, R. Dunsmuir, J. Lamb & W. Jackson. (*J. Instn elect. Engrs*, Part III, Jan. 1946, Vol. 93, No. 21, pp. 53–68.) A detailed account of the theory, design and operation of apparatus for measuring the permittivity and power factor of solid dielectrics at frequencies > 600 Mc/s. The relative merits of a short-circuited coaxial resonator operating in the principal mode, and of hollow cylindrical resonators in the E_{010} and H_{01n} modes are studied. The theory of resonance when wholly or partially filled with "lossy" dielectric is developed, connecting the dielectric properties with resonant frequency and Q.

621.317.32 : 621.3.015.33 **967**
Measurement of Impulse Voltages.—J. M. Meek. (*Electrician*, 30th Nov. 1945, Vol. 135, No. 3522, pp. 608–610.) Summary of I.E.E. paper and report of discussion. See also 365 of February.

621.317.33 **968**
Measurement Applications of the Dynatron.—W. M. Ross. (*Electronics*, Dec. 1945, Vol. 18, No. 12, pp. 320..328.) The dynatron circuit is useful for measurement of impedance and as an oscillator in test work. Reactance can be measured by a substitution method, a calibrated capacitor being used to restore the frequency to its original value after the unknown reactance component has been introduced into the circuit. Resistive components of impedance can be measured by finding the difference in grid bias settings required for oscillation to take place with and without the impedance connected. The use of the dynatron circuit in an a.f., r.f., or modulated r.f. oscillator gives good frequency stability and ease of amplitude control.

621.317.33 : 621.396.611.21 **969**
The Measurement of the Activity of Quartz Oscillator Crystals.—A. J. Biggs & G. M. Wells. (*J. Instn elect. Engrs*, Part III, Jan. 1946, Vol. 93, No. 21, pp. 29–36.) A useful criterion of "activity" is the value of the anti-resonant impedance of the crystal with a standard value of total shunt capacitance, *e.g.*, 30 $\mu\mu$F. Three general methods of measuring this impedance are given; (*a*) an attenuation method of measurement in terms of a standard resistance, (*b*) a substitution method using the crystal as the anode load in an r.f. amplifier, and (*c*) a substitution method using the crystal as the oscillatory circuit in a valve oscillator. Activity meters for the frequency ranges 50 kc/s–20 Mc/s, 80 kc/s–3 Mc/s, and >3 Mc/s are described.

621.317.33 : 621.396.645 **970**
Measurement of Amplifier Input Impedance.—D. L. Waidelich. (*Communications*, Dec. 1945, Vol. 25, No. 12, pp. 48..97.) The voltage required at the generator is measured, for constant amplifier output, when the generator is connected to the amplifier input (1) directly, (2) through a known resistance, (3) through a known capacitance. From the three equations thus obtained the input reactance and resistance can be computed.

621.317.34 : 621.317.72 : 621.385 : 621.396.322 **971**
The Rectification of Signal and Noise.—V. J. Francis & E. G. James. (*Wireless Engr*, Jan. 1946, Vol. 23, No. 268, pp. 16–25.) "It is shown that the d.c. component of rectified noise, obtained with a square law triode, is not proportional to the mean square noise unless the valve is biased to cut-off. Unless this condition is fulfilled, errors as large as 3 db may be obtained if the assumption of proportionality is made. It is shown that a linear diode with a resistive load shunted by a large capacitor provides an accurate method of measuring noise. The mean rectified current with no applied signal is directly proportional to the r.m.s. noise input, the factor of proportionality being a function of the load resistance multiplied by the diode conductance.

When both noise and signal are present, the ratio of the rectified current to r.m.s. noise input is a function of the signal/noise input ratio and the product of load resistance and diode conductance. For the case of a diode operating on the exponential part of the characteristic, the rectified current is obtained for an input consisting of a mixture of signal and noise. Results are given for various values of signal input voltage and r.m.s. noise input. An indication is given of the type of transition to be expected between the exponential characteristic and the linear characteristic as the mean diode current increases."

621.317.35 **972**

A Simplified Method of Wave Analysis.—W. L. Cassell. (*Communications*, Dec. 1945, Vol. 25, No. 12, pp. 38..103.) The presence and value of any required harmonic can be determined by the addition of ordinates of superimposed equi-angular divisions (equal in number to the order of the harmonic sought) of one cycle of the complex wave. The superposition and addition of the divisions can be performed on a cathode-ray tube by adjustment of the time base.

621.317.361 **973**

Checking UHF Oscillator Stability.—L. E. Pinney. (*Electronics*, Dec. 1945, Vol. 18, No. 12, p. 139.) A simple untuned mixer circuit that may be connected to the input of a communications receiver to obtain a heterodyne beat between a standard oscillator and that under test.

621.317.42 **974**

The Fluxball: a Test Coil for Point Measurements of Inhomogeneous Magnetic Fields.—W. F. Brown, Jr. & J. H. Sweer. (*Rev. sci. Instrum.*, Oct. 1945, Vol. 16, No. 10, pp. 276–279.) "The value of a magnetic field at a point is equal to the volume average of the field over the interior of any sphere centred at the point and not containing any sources of the field. It is therefore possible to wind a coil on the surface of a sphere in such a way that the flux through the coil is proportional to the field at a point . . . Design formulas are derived and construction procedures described."

621.317.714 **975**

Measurement of Electric Currents by Means of a Mercury Manometer.—A. Kolin. (*Rev. sci. Instrum.*, Dec. 1945, Vol. 16, No. 12, pp. 378–379.) The mechanical forces produced when current flows through a mercury film in a magnetic field alter the level in a capillary tube. The sensitivity is 1 mm per ampere in a field of 2 300 gauss with a resistance of $10^{-3}\ \Omega$. A magnetic rather than an ohmic shunt should be used to increase the range of the instrument.

621.317.725 : 621.385 : 621.392.5 **976**

Bridge Circuits with a Non-Linear Element: Valve-Voltmeter with a Stable Zero Adjustment.—M. Levy. (*Wireless Engr*, Jan. 1946, Vol. 23, No. 268, p. 3–7.) The instability of the "zero" of valve voltmeters due to variations of supply voltage is overcome (*a*) by biasing the valve so that over the expected range of variation of supply voltage the valve is effectively a linear device or (*b*) by inserting in one of the other arms of the bridge circuit a compensating network, composed of a neon tube and resistors, which simulates the non-linear characteristic of the valve. Stability of zero of about 1% of full-scale deflexion for 30% variation of anode voltage, or 10% variation of heater voltage is obtained.

621.317.73.029.62 **977**

VHF Impedance Measurements.—D. S. Henry. (*Electronics*, Dec. 1945, Vol. 18, No. 12, pp. 156–159.) Impedance measurements in the range 50–150 Mc/s are made by connecting the unknown across a high-*Q* circuit coupled to a resonant circuit. The resistive and reactive components are measured in terms of the ratio of the voltages across the resonant circuit, with and without the unknown impedance connected, and the change in capacitance required to tune it to resonance. A balanced valve voltmeter is used both to measure the voltages and as a resonance indicator.

621.317.738 **978**

Microfarad Meters: Their Advantages and Limitations.—R. P. Turner. (*Radio News*, Dec. 1945, Vol. 34, No. 6, pp. 46..147.) A.c. milliammeter, a.c. ohmmeter, and ratio dynamometer instruments give direct readings if the power factor is very small. Heterodyne meters are bulky, but readings are independent of power factor, voltage, and waveform.

621.317.75 **979**

Radio-Frequency Spectrum Analyzers.—E. M. Williams. (*Proc. Inst. Radio Engrs, N.Y.*, Jan. 1946, Vol. 34, No. 1, pp. 18–22.) The resolving power S of a radio-frequency spectrum analyser of the continuously tuned type is defined as the width in frequency, at points 3 db down, of the trace of a constant-frequency c.w. signal. Experimentally this has a maximum value given by $S = 1.3\ (F/T)^{\frac{1}{2}}$ (F = frequency band scanned, T = time of scan), and requires an i.f. bandwidth $\Delta f \approx (F/2T)^{\frac{1}{2}}$. Arguments based on the properties of a single tuned circuit are used to support these formulae. By similar methods the resolution required to separate the sidebands and the sideband envelope of pulsed signals is deduced. Spectra of pulse- and frequency-modulated signals are shown.

621.317.763 + 621.392 + 621.396.616 **980**

The Transverse Electric Modes in Coaxial Cavities.—Kirkman & Kline. (*See* 888.)

621.317.79 **981**

Probe Error in [waveguide] **Standing-Wave Detectors.**—W. Altar, F. B. Marshall & L. P. Hunter. (*Proc. Inst. Radio Engrs, N.Y.*, Jan. 1946, Vol. 34, No. 1, pp. 33–44.) The errors due to reflection from a probe are analysed assuming that it behaves as a shunt admittance. Measurements confirm the validity of the assumption over a wide range of penetrations up to 65% of the guide height. Even where bad distortion of the pattern is produced, three readings at specified probe positions are sufficient to give the true standing-wave ratio. Application of the results enables deeper penetrations to be used with increase of sensitivity. Geometrical representations of the vector relationships and curves of experimental patterns with various probe penetrations and standing-wave ratios are given.

621.317.79 : 621.315.21.029.6.091 **982**

Attenuation Test Equipment for V-H-F Transmission Lines.—F. A. Muller & K. Zimmerman. (*Communications*, Dec. 1945, Vol. 25, No. 12, pp. 54..60.) Impedance matching is obtained by adjusting the inductive coupling between test oscillator and line; input and output voltages on the line can then be measured with a valve voltmeter. A frequency range of 100 to 400 Mc/s is covered by two oscillators. The apparatus is described.

621.317.791 **983**

The Audio Chanalyst.—A. Liebscher. (*Radio News*, Dec. 1945, Vol. 34, No. 6, pp. 32..100.) Another account of the instrument described in 3940 of 1945.

621.317.794 **984**

Alternating-Current Bolometer for Infra-Red Spectroscopy.—C. H. Schlesman & F. G. Brockman. (*J. opt. Soc. Amer.*, Dec. 1945, Vol. 35, No. 12, pp. 755–760.) The 1 000 c/s a.c. bridge is screened and contains resistance elements to improve temperature compensation. The detector amplifier has a maximum sensitivity of 0.1 μV for full-scale deflexion of the high-speed recorder used. The bridge power supply, derived from a tuning fork and amplifier, is voltage stabilized to better than 7 mV in 240 V over a half-hour. Sensitivity, limited by power supply instability, is 10^{-3} degree C for full scale, giving a minimum detectable change of 10^{-5} degree C.

621.396.62 **985**

Dynamic Handful Signal Tracer.—R. Bloom. (*Radio Craft*, Nov. 1945, Vol. 17, No. 2, pp. 98, 113.) A small indicator constructed from a midget radio receiver for r.f. and a.f. signal tracing when servicing receivers.

621.317.35 **986**

Waveform Analysis: A Guide to the Interpretation of Periodic Waves, including Vibration Records [Book Review].—R. G. Manley. Chapman & Hall, Ltd., London, 1945, 275 pp., 21s. (*Nature, Lond.*, 29th Dec. 1945, Vol. 156, No. 3974, p. 765.) See 3712 of 1945. "The author is to be congratulated on the completeness of his practical treatment, and the advice he has given in interpretation . . ."

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

535.6.08 : 612.1 : 621.317.39 **987**

A Simplified Photoelectric Colorimeter for Blood Analysis.—H. L. Andrews & B. L. Horecker. (*Rev. sci. Instrum.*, June 1945, Vol. 16, No. 6, pp. 148–152.) A comparison spectrophotometer suitable for clinical purposes, based on infra-red absorption phenomena. A meter indication derived from a photocell coupled to a linear amplifier gives percentage transmission as an indication of haemoglobin content.

545.721 : 545.83 : 621.38 **988**

Microanalytic Measurement of Oxygen Production.—P. Pringsheim. (*Electronics*, Dec. 1945, Vol. 18, No. 12, pp. 340..350.) Silica gel dyed with *e.g.* acriflavine, and highly evacuated, is phosphorescent to light, and the decay time after switching off the irradiating light source is dependent on the oxygen concentration in the surrounding medium, but is not affected by the other uncontaminated atmospheric gases. This property is used in measuring partial pressures of very small quantities of oxygen. The light-sensitive phosphor is irradiated periodically, and the decay time measured by an electronic device.

578.088.7 : 621.314.3 **989**

"Biological Amplifiers".—Robertson. (*See* 862.)

620.179 : 534.321.9 **990**

The Supersonic Reflectoscope, an Instrument for Inspecting the Interior of Solid Parts by Means of Sound Waves.—Firestone. (*See* 822.)

621.31 : 621.38 **991**

Application of Electronics in the Electric Power Industry.—C. F. Wagner. (*Elect. Engng, N.Y.*, Sept. 1945, Vol. 64, No. 9, pp. 323–327.) A general account of the use of mercury-arc rectifiers, a.m., f.m., and single-sideband carrier communications, dust precipitation, lighting, and cable testing.

621.317.39.082.7 **992**

An Electrical Moisture Meter.—C. F. Brockelsby. (*J. sci. Instrum.*, Dec. 1945, Vol. 22, No. 12, pp. 243–244.) "A moisture meter depending on the variation of dielectric constant is described. Measurements are made at a low radio frequency. The range of direct reading is 8–28% moisture content for wheat (for which the instrument was primarily designed); over the greater part of this range the probable error of a single reading is approximately ± 0.4% moisture content, taking one recognized oven method as the standard. Changes in moisture content of 0.2% can be detected." See also 388 of February.

621.317.39.082.7 **993**

An Electrical Moisture Meter.—L. Hartshorn & W. Wilson. (*J. Instn elect. Engrs*, Part I, Jan. 1946, Vol. 93, No. 61, pp. 56–57.) Long abstract of 388 of February.

621.365.2.076.7 **994**

Electronic Regulator for Arc Furnaces.—(*Electronics*, Nov. 1945, Vol. 18, No. 11, pp. 314..316.) A device to control the lowering of the electrode in an arc furnace, utilizing the furnace current and furnace voltage as initiating control-parameters.

621.365.5 + 621.365.9 + 621.369 **995**

The Place of Radiant, Dielectric and Eddy-Current Heating in the Process Heating Field.—L. J. C. Connell, O. W. Humphreys, & J. L. Rycroft. (*J. Instn elect. Engrs*, Part I, Dec. 1945, Vol. 92, No. 60, pp. 464–465.) Long summary of 145 of January.

621.365.5 + 621.385 + 621.396/.397 **996**

Results Achieved by a Large [French] **Industrial Research Centre** [Le Groupe Industriel des Compagnies Françaises de T.S.F. Associées] **During and Despite the Occupation.**—Brenot. (*See* 1105.)

621.365.5 **997**

An Introduction to the High-Frequency Induction Furnace.—J. H. H. Teece. (*J. Instn elect. Engrs*, Part I, Jan. 1946, Vol. 93, No. 61, pp. 50–51.) Abstract of an I.E.E. Students' Section Paper.

621.365.52 **998**

High Frequency Induction Furnace.—N. C. Saha. (*Electrotechnics*, Sept. 1945, Nos. 17/18, pp. 49–58.) A short general account of the main types of h.f. alternators, the design of the furnace, the theory of induction heat treating and its main uses.

621.365.92 **999**

Dielectric Heating Design Chart : A Correction.—(*Electronic Engng*, Feb. 1946, Vol. 18, No. 216, p. xiii following p. 60.) A correction to charts appearing in 3724 of 1945 (Maddock).

621.365.92 **1000**

Dielectric Heating Fundamentals.—D. Venable. (*Electronics*, Nov. 1945, Vol. 18, No. 11, pp. 120–124.) The basic molecular theory underlying the process of dielectric heating is briefly described. Power requirements, thermal losses, characteristics of the load, and types of networks for matching the load to the oscillator are discussed.

621.365.92 : 679.5.023 **1001**

The Hardening of Synthetic Resins in a High-Frequency Field.—H. Stäger & F. Held. (*Schweiz. Arch. angew. Wiss. Tech.*, Sept. 1944, Vol. 10, No. 9, pp. 259–268.) A detailed experimental study of the hardening process: includes measurements of the temperature changes in the specimen, and the variations of the electrical constants in the course of the hardening process.

621.383 : 621.385.15 : 621.314.2 : 535.24 **1002**

The Possibility of Comparing Unmodulated Luminous Fluxes by means of an Electron-Multiplier Photocell followed by an A.C. Amplifier.—Le Blan. (*See* 1088.)

621.385.833 + 537.533.72 **1003**

Rigorous Treatment of the Electrostatic Immersion Lens Whose Axial Potential Distribution is Given by $\phi(z) = \phi_0 e^{K \arctan z}$.—R. G. E. Hutter. (*J. appl. Phys.*, Nov. 1945, Vol. 16, No. 11, pp. 678–699.) This is the simplest electrostatic lens for which Newton's image equations are satisfied throughout the field. The general solution of the paraxial-ray differential equation is derived, from which all the important optical quantities can be obtained graphically.

621.385.833 + 537.533.72 **1004**

The Class of Electron Lenses which Satisfy Newton's Image Relation.—R. G. E. Hutter. (*J. appl. Phys.*, Nov. 1945, Vol. 16, No. 11, pp. 670–678.) A simple construction can be used to find the image of any object if the cardinal points of the lens are known, provided the object and image lie outside the region of the electromagnetic fields. It is shown that a class of electron lenses exists for which these image relations are satisfied for any object and image position. The results are compared with those of Glaser and Lammel (202 of 1942).

621.385.833 + 537.533.73 **1005**

High Dispersion Electron Diffraction by Primary Magnification.—G. L. Simard, C. J. Burton, & R. B. Barnes. (*J. appl. Phys.*, Dec. 1945, Vol. 16, No. 12, pp. 832–836.)

621.385.833 **1006**

Electron Microscopical Replica Techniques for the Study of Organic Surfaces.—R. B. Barnes, C. J. Burton & R. G. Scott. (*J. appl. Phys.*, Nov. 1945, Vol. 16, No. 11, pp. 730–739.)

621.385.833 **1007**

A Shadow Casting Unit for the RCA Electron Microscope.—H. R. Crane, H. Levinstein, & R. C. Williams. (*Rev. sci. Instrum.*, Oct. 1945, Vol. 16, No. 10, p. 296.)

621.385.833 **1008**

A Bibliography of Electron Microscopy. III.—C. Marton & S. Sass. (*J. appl. Phys.*, July 1945, Vol. 16, No. 7, pp. 373–378.) For parts I and II see 545 of 1944 and 165 of 1945.

621.385.833 **1009**

Cinematography with the Universal Electron Microscope.—M. von Ardenne. (*Z. Phys.*, 23rd Feb. 1943, Vol. 120, Nos. 5/6, pp. 397–412.) A further stage in the developments described in 1774 and 1775 of 1942. The equipment enables a cinema film to be taken of a moving or changing subject in an electron microscope, up to 25 pictures a second, or a long series of pictures up to 2 per second using a "Leica-format". Some photographs are reproduced.

621.385.833 **1010**

The Resolving Power of the Field-Emission [cold cathode] **Electron Microscope.**—E. W. Müller. (*Z. Phys.*, 23rd Feb. 1943, Vol. 120, Nos. 5/6, pp. 270–282.) One limitation on resolving power is the emission-velocity distribution. For a given velocity distribution and a given radius of curvature of the point-cathode, the resolving power can be calculated, and the theory is found to be confirmed by experimental data; but the highest resolving power so far obtained, about 10 Å, is not limited by these factors, but by non-uniformity of the cathode surface.

621.385.833 **1011**

The Resolving-Power of the Self-Emission Electron Microscope.—A. Recknagel. (*Z. Phys.*, 23rd Feb. 1943, Vol. 120, Nos. 5/6, pp. 331–362.) A homogeneous accelerating field and a perfect electron-lens are assumed, and the resolving power calculated, on the basis of the Schrödinger wave equation, in so far as it is limited by spherical and chromatic aberration and by wave phenomena. The least resolvable distance is of the order of the wavelength at the cathode, decreasing with increase of field strength, but less rapidly than predicted by geometric-optical theory. With further increase of field strength, the electron stream extends more and more beyond the limits of the permissible geometric-optic cone, and there is consequently a theoretical upper limit to the resolving power. For previous work see 514 of 1942.

621.385.833 **1012**

Beam Current Stability in RCA Electron Microscopes.—F. W. Cuckow. (*Rev. sci. Instrum.*, Oct. 1945, Vol. 16, No. 10, p. 293.) Comment on 2306 of 1945 (Crane).

621.398 : 629.13 **1013**

Army Target Plane Controlled by Radio Transmitter.—(*Electronics*, Dec. 1945, Vol. 18, No. 12, p. 300.) A u.h.f. ground transmitter is modulated by five separate audio frequencies selected by a lever simulating the control column of an aircraft. A brief description.

621.9 : 621.38 **1014**

Machine Tool Control.—R. A. Stremel. (*Electronic Industr.*, Nov. 1945, Vol. 4, No. 11, pp. 102–103.) Form and thread milling machine automatically controlled by means of an electronic timer. No further setting required after the first cut. The circuit diagram is given.

623.422.31 : 621.318.572 **1015**

Machine Gun Rate-of-Fire Indicator.—A. D. Peterson. (*Electronics*, Dec. 1945, Vol. 18, No. 12, pp. 134–138.) The reciprocating action of the slide mechanism actuates a micro-switch to initiate pulses which are applied to a series of five Eccles-Jordan trigger circuits with a counting ratio of 16 to 1. The last trigger circuit controls a gate circuit which allows a capacitance to charge for a period of seventeen shots. The final potential across the capacitance is measured by a valve voltmeter calibrated in rates of fire.

623.455 : 621.383 **1016**

Small Arms Ammunition.—R. S. J. Spilsbury & A. Felton. (*Electrician*, 28th Dec. 1945, Vol. 135, No. 3526, pp. 739–743.) Light falling on the base of a metal cartridge case passes through the fire holes to a photocell. A relay mechanism is actuated, causing the case to be accepted or rejected according to the adequacy of the area of the fire holes.

629.13 : 621.317.39 **1017**

Circuit of Electronic Capacitance-Type Fuel Gage.—(*Electronics*, Nov. 1945, Vol. 18, No. 11, pp. 324..332.) The fuel rises between two concentric aluminium tubes thus varying their mutual capacitance. The change in capacitance is measured on a self-adjusting bridge. Mass of fuel in the tank is measured, rather than its volume, as thermal expansion of the liquid is offset by a fall in permittivity.

629.13 : 621.317.39 **1018**

Electronic Fuel Gaging.—(*Electronics*, Dec. 1945, Vol. 18, No. 12, pp. 234..244.) A 10 kc/s oscillator feeds two similar L-C circuits. One capacitor is pre-set for reference, the other, in the fuel tank, varies in capacitance with the quantity of fuel dielectric present. A meter shows the ratio of amplitudes of oscillation of the two circuits.

629.13 : 621.43.044.1 : 621.3.029.56 **1019**

High-Frequency Aircraft Ignition System.—(*Electronics*, Dec. 1945, Vol. 18, No. 12, pp. 264..272.) The magneto output at about 1 200 V is converted into 2–3 Mc/s pulses by an impulse-type frequency converter in each spark plug circuit, and transformed to 10 000 V by a transformer built into the end of the plug. The advantages are enumerated. A short note.

677.024 : 621.383 **1020**

Phototube Weft Straightening in Textile Industry.—(*Electronics*, Nov. 1945, Vol. 18, No. 11, pp. 316..324.) Two spots of light, with associated phototube circuits, explore the fabric in directions symmetrically disposed to the direction of travel of the fabric. When the weft is not at right angles to the warp the impulses from the two channels do not occur at the same frequency, and the frequency difference is used to control the operation of the straightening motor.

615.84 **1021**

Technic of Electrotherapy and Its Physical and Physiological Basis [Book Review].—S. L. Osborne & H. J. Holmquest. Chas. C. Thomas, Springfield, Ill., 1944, 780 pp., $7.50. (*Rev. sci. Instrum.*, Oct. 1945, Vol. 16, No. 10, p. 283.) "From the physicist's point of view it is one of the few works of this kind which furnishes a thoroughly adequate and sound background for the . . .results produced with physical therapy methods . . ."

PROPAGATION OF WAVES

621.3.011.2 + 621.011.5].029.62 : 631.437 **1022**

The Electrical Properties of Soil at Wavelengths of 5 Metres and 2 Metres.—J. S. McPetrie & J. A. Saxton. (*J. Instn elect. Engrs*, Part I, Jan. 1946, Vol. 93, No. 61, p. 58.) Long abstract of 711 of March.

621.396.1 + 621.396.8 **1023**

The Evolution of the Technique of Long-Distance Telecommunications.—Rabuteau. (*See* 1048.)

621.396.11 + 534.231 **1024**

The Saddle-Point Method in the Vicinity of a Pole with Applications to Wave-Optics and Acoustics.—H. Ott. (*Ann. Phys., Lpz.*, 1943, Vol. 43, Nos. 6/7, pp. 393–403.) A study of the integration in the complex plane of expressions of the form $e^{ikR\cos(\theta-\alpha)}A(\theta)\,d\theta$ leading to solutions for the vicinity of a pole in terms of Fresnel integrals. The method is applied to a vertical dipole assumed to be located above the surface of an earth of finite conductivity, and leads to a general formula for the distant field in any direction which includes Sommerfeld's classical formula as a special case. The treatment is considered to throw some new light on the question of the "surface wave". "The splitting off of a Zenneck-wave term from the total radiation is thus seen to be, in certain conditions, not so arbitrary a matter as has been widely suggested ; this surface wave does not extend far into the atmosphere, but appears, so to speak, as a very narrow fringe [*Saum*] to the wave field."

621.396.11 : 551.51.053.5 : 523.746 **1025**

The Bright New World—of Sunspots.—E. H. Conklin. (*QST*, Jan. 1946, Vol. 30, No. 1, p. 43.) Mainly an account of how world-wide ionospheric research during the war has extended knowledge of the characteristics of long-distance propagation *via* the F layer.

621.396.11 : 551.51.053.5 : 550.38 **1026**

On the Influence of the Terrestrial Magnetic Field on the Reflection of Radio Waves from the Ionosphere.—V. L. Ginsburg. (*J. Phys., U.S.S.R.*, 1943, Vol. 7, No. 6, pp. 289–304.) "The question of the influence of the terrestrial magnetic field on the reflection of radio waves and signals from an inhomogeneous ionized layer (Heaviside layer) is considered [theoretically]. In particular the propagation of waves at a small angle to the direction of the magnetic field is investigated, and it is shown that in this case a very peculiar splitting of the reflected signal into three pulses, and not into two, as observed in other cases, must take place."

621.396.11 : 741.021 1027
Drafting Aids to Relay Profiling.—(*See* 1087.)

621.396.11.029.62 1028
Field Intensities Beyond Line of Sight at 45.5 and 91 Mc.—C. W. Carnahan. (*Communications*, Dec. 1945, Vol. 25, No. 12, pp. 46..92.) Tests on a 70-mile link. For equal transmitter power 45.5 Mc/s signals exceeded to 10 μV/m all the time (2 months), while 91 Mc/s signals exceeded this value only 65% of the time. Report of an I.R.E. paper.

621.396.67 + 621.396.11 1029
Antenne e propagazione delle onde elettromagnetiche (per ingegner) [Book Review].—Peroni. (*See* 859.)

RECEPTION

621.396.61.029.64 1030
Duplex Phone on 5300 Megacycles.—Merchant & Harrison. (*See* 1100.)

621.396.62 1031
Pye Model 15A [receiver: test report].—(*Wireless World*, Jan. 1946, Vol. 52, No. 1, pp. 12–13.)

621.396.621.52 1032
The Super-Regenerative Detector: an Analytical and Experimental Investigation.—F. R. W. Strafford. (*J. Instn elect. Engrs*, Part III, Jan. 1946, Vol. 93, No. 21, pp. 23–28.) An approximate theoretical analysis on the basis of an exponential build-up of oscillations in the grid circuit. The circuit gives high amplification with inherent a.v.c. action, but with severe distortion for deeply modulated signals. The selectivity of the input circuit is dependent on the frequency and waveform of the quenching voltage. A low quenching frequency with a waveform giving a slow rate of rise (*e.g.* sinusoidal) gives optimum selectivity and signal to noise ratio. On the other hand the output at modulation frequency rises with increase in quenching frequency. Impulsive interference may be substantially suppressed. F.m. reception is discussed briefly.

621.396.621.53 : 621.385.2 1033
Diode Input Resistance.—R. E. Burgess. (*Wireless Engr*, Jan. 1946, Vol. 23, No. 268, p. 29.) Letter pointing out computational errors in 3343 of 1941 (Bell).

621.396.622 : 621.396.619.018.41 1034
A New F.M. Detector Circuit [requiring no limiter].—G.G. (*QST*, Jan. 1946, Vol. 30, No. 1, pp. 26–27.) See 574 of March.

621.396.645 : 621.317.33 1035
Measurement of Amplifier Input Impedance.—Waidelich. (*See* 970.)

621.396.82 : 523.5 1036
Listening in on the Stars.—Villard. (*See* 916.)

621.396.822 1037
A Theory of Valve and Circuit Noise.—N. R. Campbell & V. J. Francis. (*J. Instn elect. Engrs*, Part III, Jan. 1946, Vol. 93, No. 21, pp. 45–52.) Detailed discussion of fluctuation noise on the basis of Campbell's theorem, emphasis being laid on the rigour of the fundamental postulates in the statistical theory of random processes. Campbell's theorem is applied to the shot noise in a tuned circuit connected to a temperature-limited diode and to thermal noise after reference to the equipartition principle leading to Johnson's formula. A general theory of valve and circuit noise in complicated systems is presented using a determinant method. The modification of the theory for the effects of space charge is based on the treatment of Thompson and North. The effects of finite transit time in a valve and in a resistor are discussed, and doubt expressed regarding Spenke's assertion that Johnson's formula holds at very high frequencies.

621.396.822 1038
[Noise] **Fluctuations of Electric Current.**—D. A. Bell. (*J. Instn elect. Engrs*, Part III, Jan. 1946, Vol. 93, No. 21, pp. 37–44.) A theoretical review of shot noise and thermal noise. The thermodynamical basis of Nyquist's theorem and Schottky's formula for the temperature-limited diode are briefly discussed. The controversial case of the space-charge-limited diode is considered in detail, the mutually consistent theories of Rack, Spenke and North being contrasted with that of Bell which predicts appreciably higher values for the space-charge reduction factor F^2 at large values of $eV/k\theta$. A suggestion by J. R. Pierce reconciling the two theories is not accepted by the author. Negative-grid triodes, screen-grid valves (with partition noise), secondary-emission amplifiers, receiver noise figures, and the noise in an aerial are considered briefly.

621.396.822 1039
Fluctuation Voltages in Receiver Input Circuits.—J. R. Ragazzini. (*Proc. Radio Cl. Amer.*, Dec. 1945, Vol. 22, No. 2.) Noise due to thermal agitation of electrons. The analogy with Brownian movement is used to develop equations for noise voltages. If valve noise is negligible compared with circuit noise in a receiver, signal to noise ratio can be improved by overcoupling.

621.396.822 : 621.317.34 : 621.317.72 : 621.385 1040
The Rectification of Signal and Noise.—Francis & James. (*See* 971.)

621.396.826 : 621.396.619.018.41 1041
Frequency-Modulation Distortion Caused by Multipath Transmission.—M. S. Corrington. (*Proc. Inst. Radio Engrs*, Dec. 1945, Vol. 33, No. 12, pp. 878–891.) A theoretical and experimental study. When the amplitude of the direct and reflected waves are about equal at the receiver, interference at certain frequencies may cause one or more "holes" in the carrier too deep for the limiter to deal with, giving rise to unwanted amplitude variations at the discriminator. At these points there is also a very rapid phase shift causing large peaks or dips in the discriminator output. The distortion increases with the deviation and path difference, and is worse on the higher audio frequencies. Expressions for the distorted waveform are derived, and observed waveforms are shown.

621.397.62 1042
Extended-Range Television Reception: Part II.—Wilder. (*See* 1093.)

1043
Correction.—In 410 of February (Frank), *for* Vol. 35 *read* Vol. 34.

STATIONS AND COMMUNICATION SYSTEMS

621.3(41) "1939/1945" **1044**

"British Electrical Engineers and the Second World War".—P. Dunsheath. (*J. Instn elect. Engrs.*, Part I, Jan. 1946, Vol. 93, No. 61, pp. 17–27.) The full paper, of which summaries were noticed in 415 of February and 729 of March.

621.391.32 : 621.395.44 : 622.33 **1045**

The Employment in Mines of Telephony through the Ground.—Burgholz. (*Génie civ.*, 1st/15th Nov. 1944, Vol. 121, Nos. 21/22, p. 180.) Short abstract of a paper in *Glückauf*, 29th Apr. 1944. The use of an l.f. carrier current is recommended.

621.395.44 **1046**

Design of Carrier-on-Cable Equipment.—T. S. Skillman. (*Communication Rev.*, 1945, Vol. 1, No. 1, pp. 1–24.) Design aspects of multi-channel equipment for use on balanced multi-pair telephone paper-insulated cables are discussed. A description is given of a seventeen-channel system operating with such cables using the standardized frequency spacings.

621.396/.397 + 621.365.5 + 621.385 **1047**

Results Achieved by a Large [French] **Industrial Research Centre** [Le Groupe Industriel des Compagnies Françaises de T.S.F. Associées] **During and Despite the Occupation.**—Brenot. (*See* 1105.)

621.396.1 + 621.396.8 **1048**

The Evolution of the Technique of Long-Distance Telecommunications.—M. G. Rabuteau. (*Onde élect.*, Dec. 1945, Vols. 20/25, No. 225, pp. 140–154.) A survey in anticipation of forthcoming international discussion. The potentialities of the various frequency bands are classified thus :—(1) 10 kc/s to 150 kc/s—stable long-distance communication by high-power installations : (2) 150 kc/s to 3 Mc/s—short and medium distances : (3) 3 Mc/s to 30 Mc/s—ionospheric propagation, subject to diurnal and seasonal effects : 3 Mc/s to 6 Mc/s—short distances, by day : 6 Mc/s to 9 Mc/s—long distances, by night : 9 Mc/s to 15 Mc/s—day-night "transition" frequencies for long distances : 15 Mc/s to 30 Mc/s—long distances by day. "In general, to assure optimum conditions for medium and long distances by day and by night, three frequencies are necessary, one for day, a higher frequency for night, and an intermediate frequency." (4) 30 Mc/s to 300 Mc/s—optical ranges : (5) above 300 Mc/s—not yet used for long-distance telecommunication, but likely to be used for optical-range relay links. The propagation characteristics of these bands are illustrated by numerous graphs and diagrams.

Possible means of increasing the traffic-carrying capacity of available bands include :—(1) Stabilization of carrier frequencies : (2) minimization of transmitted bandwidth (*e.g.*, single sideband working), use of multiplex systems, based on pulse or frequency modulation : (3) elimination of parasitic radiation (carrier harmonics) : (4) use of highly directive systems, enabling many different routes to operate on the same frequency.

621.396.619.018.41 **1049**

FM for Ham Use.—R. Frank. (*Radio News*, Dec. 1945, Vol. 34, No. 6, pp. 44–45.) Transmitter amplifiers can be run with larger inputs than for a.m., and modulation is carried out at low level. Reactance-tube modulation is recommended with a maximum modulation frequency of 2 500 c/s and deviation ratio of unity.

621.396.619.018.41 **1050**

Frequency Modulation.—R. Schulz. (*Funktech. Mh.*, July/Aug. & Sept./Oct. 1943, Nos. 7/8 & 9/10, pp. 84–88 & 100–102.) A review of the properties and advantages of f.m. compared with other systems, particularly a.m., and an account of basic f.m. circuits.

621.396.619.018.41 **1051**

Frequency Modulation.—K. R. Sturley. (*J. Instn elect. Engrs*, Part III, Jan. 1946, Vol. 93, No. 21, pp. 11–15.) Discussion of 4047 of 1945.

621.396.619.018.41 : 621.396.826 **1052**

Frequency-Modulation Distortion Caused by Multipath Transmission.—Corrington. (*See* 1041.)

621.396.619.16 **1053**

Pulse Modulation.—D. Cooke. (*Wireless Engr*, Jan. 1946, Vol. 23, No. 268, p. 29.) A plea for early rationalization and standardization of nomenclature applied to various simplex and multiplex pulse-modulation communication systems. Some of the terms used in 183 of January (Roberts & Simmonds) are considered suitable but "pulse delay modulation" is considered inapt. "Pulse phase modulation" is suggested as an alternative.

621.396.619.16 : 621.397.5 **1054**

Pulse Modulation.—Bedford. (*See* 1092.)

621.396.65.029.64 **1055**

Microwave Radio Relay to Replace Telegraph Lines.—(*Electronics*, Dec. 1945, Vol. 18, No. 12, pp. 300..308.) The relay systems include towers about 30 miles apart with four paraboloid aerials, two transmitting in opposite directions and two receiving from opposite directions. The operating frequency is about 4 000 Mc/s with a bandwidth of 150 kc/s, sufficient for 270 multiplex or 1 080 single telegraph circuits, or 25 ordinary telephone circuits. The stations are automatic and unattended. A brief description.

621.396.65.029.64 : 621.396.619.16 **1056**

AN/TRC-6 – A Microwave Relay System.—H. S. Black. (*Bell Lab. Rec.*, Dec. 1945, Vol. 23, No. 12, pp. 457–463.) Eight speech channels are provided with pulse-position modulation on a frequency of 5 000 Mc/s, which is relatively free from interference. A 57-inch-diameter paraboloid reflector gives a narrow beam of radiation, so that a transmitter power of two watts gives adequate signal strength at the receiver over an optical path of 100 miles. The antenna system is carried on a 50 ft tower. Relay stations have provided communication over a distance of 1800 miles. The equipment was used for Army communications to rear areas on the Western Front with conspicuous success. See also 740 and 741 of March.

621.396.71 + 621.396.677.2 **1057**

Stepping up from ¼ KW to 5 KW [and the use of a directional aerial].—A. E. Griffiths. (*Communications*, Dec. 1945, Vol. 25, No. 12, pp. 33..85.) The conversion of broadcast station KOTA (1 380 kc/s) to the use of higher power and a directional antenna system is described, with particular reference to the phasing and feeding of the three transmitting towers to give the required field-strength polar diagrams.

621.396.721 **1058**
Station Design and Planning: Part I.—W. H. Allen. (*R.S.G.B. Bull.*, Dec. 1945, Vol. 21, No. 6, pp. 88–89.) Seeks to answer the question " What apparatus shall I require ? " for the amateur transmitter. The items detailed are:—aerial; transmitter; receiver; test and maintenance equipment.

621.396.82 **1059**
Radio Counter Measures.—(*Wireless World*, Jan. 1946, Vol. 52, No. 1, p. 18.) A very short account of methods used to confuse enemy communications and radar.

621.396.82 **1060**
QRM – The Electronic Life Saver: Part I.—P. Robbiano. (*QST*, Jan. 1946, Vol. 30, No. 1, pp. 12–18.) An outline of some of the radio-counter-measures work of the U.S.A. Descriptions are given of jamming transmitters developed in the frequency range up to 3 000 Mc/s: one of these is the " Resnatron " which develops about 50 kW (average power) in the frequency range 480–600 Mc/s. The application of noise modulation to jamming transmitters is discussed, and brief mention is made of the use of " window " as a passive device for interfering with enemy radar equipments.

621.396.82 : 621.396.9 **1061**
Radar Countermeasures.—R. G. Peters. (*Communications*, Dec. 1945, Vol. 25, No. 12, pp. 50..93.) Description of land-based and airborne methods of locating and jamming enemy radar stations, with tabulated data on jamming transmitters and searching receivers.

621.396.931 : 625.1 **1062**
Milestones in Railroad [radio] **Communications.**—H. H. Hasselbacher. (*Telegr. Teleph. Age*, Nov. 1945, Vol. 63, No. 11, pp. 22..38.) An historical account of the special requirements. The F.C.C. have reserved 60 channels of 60 kc/s bandwidth in the frequency range 152 to 162 Mc/s. An experimental system using top-loaded mobile aerials has realized full operational expectations.

621.396.931 : 625.1 **1063**
Railroad Radio Communications.—R. A. Clark, Jr. (*Communications*, Dec. 1945, Vol. 25, No. 12, pp. 62..66.) Terminal operations require ranges of a few miles, main line operations require ranges of up to hundreds of miles. Inductive carrier systems using from about 50 kc/s to 200 kc/s can be used for both types of operation. Frequencies of 30 to 2 000 Mc/s have been used for terminal operation.

621.396.97 : 534.76 **1064**
Binaural Stereophonic System of the [French] **State Broadcasting Department.**—J. Cordonnier. (*Génie civ.*, 1st July 1944, Vol. 121, No. 13, p. 105.) A short account of the demonstration of a system, duplicate throughout, with double transmitters and receivers. Transmission was by cable for the demonstration.

SUBSIDIARY APPARATUS

533.5 **1065**
The Hydrogen Gauge—An Ultra Sensitive Device for Location of Air Leaks in Vacuum-Device Envelopes.—Nelson. (*See* 936.)

535.81 : 679.5 : 621.397.62 **1066**
Molded " Lucite " Lens.—(*See* 1095.)

620.199.1/.9 **1067**
Interference-Free Weatherometer [accelerated weathering tester].—Agnew. (*See* 943.)

621.3.083.7 **1068**
Selsyn Indicating Systems for Remote Readings.—F. Wisk. (*Electronics*, Dec. 1945, Vol. 18, No. 12, pp. 276..296.) Explanation of the basic principles.

621.314.2 **1069**
Small Mains Transformers: Suggestions for Preferred Types of Standard Size.—L. A. Sherwood. (*Wireless World*, Jan. 1946, Vol. 52, No. 1, pp. 30–31.) The design of transformers for 10 to 1 000 W could be limited to 7 types of laminations with one bobbin size for each type of laminations.

621.314.2 : 621.396.674 **1070**
Loop-Antenna Coupling-Transformer Design.—Bachmann. (*See* 856.)

621.314.632 : 546.289 **1071**
Germanium Rectifiers.—(*See* 944.)

621.314.65 : 621.396.71 **1072**
The Application of High-Voltage Steel-Tank Mercury-Arc Rectifiers to Broadcast Transmitters.—P. A. T. Bevan. (*J. Instn elect. Engrs*, Part I, Jan. 1946, Vol. 93, No. 61, p. 53.) Long abstract of 206 of January.

621.314.65 : 621.396.71 **1073**
High-Voltage Steel-Tank Mercury-Arc Rectifier Equipments for Radio Transmitters.—J. C. Read. (*J. Instn elect. Engrs*, Part I, Jan. 1946, Vol. 93, No. 61, pp. 54–55.) Long abstract of 205 of January.

621.315.626 + 533.5 **1074**
Ferranti Vacuum-Seal Bushing.—(*Beama J.*, Jan. 1946, Vol. 53, No. 103, p. 28.) A brief description of a glass insulator, bonded to a copper flange and end cap that can be soft soldered. The insulator is rated for 1 000 V r.m.s. a.c. or 1 000 V d.c. working.

621.316.722.078.3 : 621.392.5 **1075**
The Theory of the Non-Linear Bridge Circuit as Applied to Voltage Stabilizers.—Patchett. (*See* 867.)

621.316.722.078.3 : 621.396.682 **1076**
Voltage-Regulated Power Supplies: Part II.—G. E. Hamilton & T. Maiman. (*Communications*, Dec. 1945, Vol. 25, No. 12, pp. 70–71.) Typical design data are given. A diagram of a power unit is given for: 0–500 V output, ranges 0–10, 0–50, 0–100 and 0–500 V, maximum output current 300 mA, regulation better than 1% over the whole voltage range, ripple voltage 10 mV. For part I see 756 of March.

621.316.974 : 621.318.4.017.31 **1077**
Power Loss in Electromagnetic Screens.—C. F. Davidson, R. C. Looser & J. C. Simmonds. (*Wireless Engr*, Jan. 1946, Vol. 23, No. 268, pp. 8–15.) An

empirical investigation of plane screens formed of concentric circular loops of wire shows that for efficiency of screening the screen must be made large enough for the eddy currents in the outer rings to be not greater than about one-tenth of the current in the loops carrying maximum eddy current. When the screen is between two small inductance coils, maximum eddy current occurs in those loops having a diameter of the same order as that of the inducing coil. The power loss occurs chiefly in these loops, and can be minimized by employing heavier conductors for their construction. Fair agreement is obtained between measured values of the increase of the resistance of screened coils and values deduced on the assumption that the eddy current distribution will approximate to that in a continuous sheet.

621.317.33 : 621.396.611.21 **1078**

The Measurement of the Activity of Quartz Oscillator Crystals.—Biggs & Wells. (*See* 969.)

621.317.755 **1079**

A Small Oscilloscope Using the 913.—E. M. McCormick. (*QST*, Jan. 1946, Vol. 30, No. 1, pp. 32–34.) A simple oscilloscope in which the use of a high-voltage transformer is avoided by using 6H6 valves directly connected to the 115 V a.c. supply in voltage doubling circuits. The instrument is not suitable for use at video frequency. Diameter of tube 1 inch, maximum sensitivity (with amplifier) about 1 inch/V.

621.317.755 **1080**

An Introduction to the High-Voltage Cathode-Ray Oscillograph.—J. B. Higham. (*J. Instn elect. Engrs*, Part I, Jan. 1946, Vol. 93, No. 61, pp. 51–52.) Summarizes the features of the modern high-speed oscilloscope. A cold cathode is usual, but electron guns employing thermionic emission are foreseen. Summary of an I.E.E. Students' Section Chairman's address.

621.318.323.2.029.5 **1081**

Iron Powder Cores and Coils.—H. W. Jaderholm. (*Proc. Inst. Radio Engrs*, Dec. 1945, Vol. 33, No. 12, p. 904.) The apparent permeability of a short cylindrical core is expressed in terms of the true (or "ring") permeability and the ratio of core length to diameter. The formulae of Bozorth and Chapin are used (*J. appl. Phys.*, May 1942, Vol. 13, No. 5, pp. 320–326.).

621.318.4.013.22 **1082**

Coil Systems for Producing Transverse and Longitudinal Magnetic Gradients.—G. H. Shortley & A. May. (*J. appl. Phys.*, Dec. 1945, Vol. 16, No. 12, pp. 841–843.) "A coil system is described which produces very constant magnetic gradients throughout a considerable part of the volume occupied by the coil assembly. One modification produces constant transverse and longitudinal gradients of the field transverse to a long cylindrical region; another modification produces constant transverse and longitudinal gradients of the magnetic field throughout a spherical volume. Gradient data are given for both modifications."

621.384.6 (091) **1083**

Historical Development of the Betatron.—D. W. Kerst. (*Nature, Lond.*, 26th Jan. 1946, Vol. 157, No. 3978, pp. 90–95.)

621.396[.65 + .71 + .812.3 **1084**

"Recent Developments in Communication Engineering".—A. H. Mumford. (*J. Instn elect. Engrs*, Part I, Jan. 1946, Vol. 93, No. 61, pp. 41–50.) The full paper, of which a summary was noticed in 4046 of 1945. Reprinted in *J. Instn elect. Engrs*, Part III, Jan. 1946, Vol. 93, No. 21, pp. 2–11.

621.396.662.2 **1085**

On the Theory of the Progressive Universal Winding.—A. W. Simon. (*Proc. Inst. Radio Engrs*, Dec. 1945, Vol. 33, No. 12, pp. 868–871.) An extension of previously developed theory (1348 of 1937 and 1464 of 1945 – Simon) to progressive universal winding, with an exact formula for calculating the gear ratio to be used. "The theory of the spiral ridges and close-packed layers of progressive universal coils is developed, and formulas giving the slope of the spiral ridge and the dimensions of the close-packed layers are deduced. The principles underlying the selection of the optimum number of cross-overs per turn are discussed, and an example of the application of the formulas to the design of a progressive universal coil is given."

621.396.69 + 621.38 **1086**

The Physical Society's Exhibition.—(*See* 1131/1133.)

741.021 : 621.396.11 **1087**

Drafting Aids to Relay Profiling.—F. J. Bingley. (*Electronics*, Dec. 1945, Vol. 18, No. 12, pp. 316..320.) Plotting of profiles of terrain along a line-of-sight television path is facilitated by the use of a special T-square with cross-members calibrated in height units. Modifications may be introduced to allow for the curvature of the earth and refraction of waves in the atmosphere.

TELEVISION AND PHOTOTELEGRAPHY

621.383 : 621.385.15 : 621.314.2 : 535.24 **1088**

The Possibility of Comparing Unmodulated Luminous Fluxes by means of an Electron-Multiplier Photocell followed by an A.C. Amplifier.—L. Le Blan. (*C. R. Acad. Sci., Paris*, 19th March 1945, Vol. 220, No. 12, pp. 394–396.) Constant sine-wave modulation is applied to one of the electrodes of the electron multiplier. The operating point is selected so that the modulation of the electron current is mainly at the second harmonic of the modulating voltage, so that by tuning the amplifier to this second harmonic the effect of capacitance coupling of the modulating voltage to the amplifier at fundamental frequency is suppressed.

621.397.2 **1089**

Comments on Existing Television Systems from a Measurement Viewpoint.—J. Minter. (*Communications*, Dec. 1945, Vol. 25, No. 12, pp. 47..80.) Proposals are made for changing the standard receiver selectivity characteristics to simplify alignment. The type of receiver alignment used prior to 1941 gives less trouble due to phase shift than that in use now. Report of an I.R.E. paper.

621.397.3 1090
A Simple Optical Method for the Synthesis and Evaluation of Television Images.—R. E. Graham & F. W. Reynolds. (*Waves and Electrons*, Jan. 1946, Vol. 1, No. 1, pp. 18–30.) An optical method of simulating television picture reproduction is described, using an out-of-focus 35 mm motion-picture projector and a lined graticule placed in front of the receiving screen. A theoretical analysis is given of the appearance of the resulting image and its dependence on various factors such as degree of defocusing of the image, shape of projector aperture, and characteristics of the line screen. The results are compared with those of television image analysis, and shown to be in close agreement. Photographs of simulated television pictures are given, which illustrate the spurious effects introduced by the scanning process.

621.397.3 1091
Television Resolution as a Function of Line Structures.—M. Cawein. (*Proc. Inst. Radio Engrs*, Dec. 1945, Vol. 33, No. 12, pp. 855–864.) Theoretical discussion of the distortion of recurrent rectangular impulses by transmission through finite apertures in an optical system. The Fourier integral representation is used to evaluate the effect of a finite cut-off frequency, and the response obtained by the synthesis of harmonics is expressed to a close approximation in terms of the sine integral. The widening of a pulse undergoing a number of aperture processes is evaluated. The results are applied to explaining the relative merits of 16 and 35 mm film for television image transmission.

621.397.5 : 621.396.619.16 1092
Pulse Modulation.—L. H. Bedford. (*Wireless Engr*, Feb. 1946, Vol. 23, No. 269, pp. 56–57.) In the "videosonic" television system (see 230 of January) the sound is conveyed by width-modulated pulses. It is clear: "(*a*) That there is an absolute limit for the audio band at half the strobe frequency, (*b*) That restriction to this band must be carried out by filtering both at the transmitter and at the receiver, and (*c*) That the linear superposition principle applies so that no complications arise on considering complex wave forms." For the present 405-line standard "one may anticipate an audio-frequency band of perhaps 4 000 c/s, which seems rather a severe restriction . . ."

621.397.62 1093
Extended-Range Television Reception : Part II.—M. P. Wilder. (*QST*, Jan. 1946, Vol. 30, No. 1, pp. 35–40, 102.) Design and construction of the signal-, intermediate-, and video-frequency portions of a sensitive television receiver. The signal-frequency amplifier is operated at a distance from the rest of the receiver to minimize interference. A filter is incorporated in the transmission line to the mixer with the object of eliminating further interference from powerful transmissions in the 5–13 Mc/s band. For part I see 463 of February.

621.397.62 1094
The DC Restorer : Its Uses in Television and Radar.—(*Wireless World*, Jan. 1946, Vol. 52, No. 1, pp. 8–11.) A descriptive article.

621.397.62 : 535.81 : 679.5 1095
Molded "Lucite" Lens.—(*Rev. sci. Instrum.*, Oct. 1945, Vol. 16, No. 10, p. 298.) Note on moulded aspherical lenses for correcting sphérical aberration in television projection. A product of E. I. DuPont de Nemours & Co.

621.397.813 1096
Theoretical Investigation of the Distortion of Television Signals in Valve Circuits.—Huber. (*See* 901.)

621.397 + 654.17 1097
Television To-day and To-morrow [Book Review]. —L. de Forest. Hutchinson's Scientific and Technical Publications, London, 1945, 176 pp., 16s. (*Electronic Engng*, Feb. 1946, Vol. 18, No. 216, p. 62.) "The whole book is infused with a tremendous rising tempo of enthusiasm . . ."

TRANSMISSION

621.385.032.442 : 621.3.025.4 1098
Polyphase from Single Phase for Valve Supply Circuits.—E. M. I. Laboratories. (*See* 863.)

621.396.61 : 621.396.662 1099
A Radio-Frequency Auto-Resonator.—J. F. Clemens. (*QST*, Jan. 1946, Vol. 30, No. 1, pp. 65–66.) An arrangement for automatically tuning the amplifier stage of a transmitter from a remote point. It comprises a motor-driven tank capacitor, in parallel with which is a small capacitor varied cyclically at 114 c/s. When the stage is correctly in tune there is a 228 c/s component in the anode current of the amplifier ; when not in tune there is a 114 c/s component. The appearance of the 228 c/s voltage is used to stop the motor driving the tank capacitor.

621.396.61.029.64 1100
Duplex Phone on 5300 Megacycles.—R. Merchant & A. E. Harrison. (*QST*, Jan. 1946, Vol. 30, No. 1, pp. 19–24.) A description of a simple transmitter-receiver, using reflector frequency-modulation of the 2K43 reflection klystron. Separate parabolic mirrors are used for transmitting and receiving. Two stations of this type have achieved reliable communication over an optical path about 5 miles in length.

621.396.612.1.029.64 1101
Micro-Electromagnetic Waves.—M. G. Kelliher & E. T. S. Walton. (*Wireless Engr*, Feb. 1946, Vol. 23, No. 269, pp. 46–51.) A description of a low-power spark oscillator operating on centimetre wavelengths, and a tunable detector with associated amplifier suitable for examining the properties of the radiation. Experiments are described showing that some twenty to thirty wavetrains are emitted during each half of the spark repetition period. Each wavetrain is shown to comprise at least thirty oscillations : the reason for this low value of damping coefficient is discussed. The energy spectrum from the oscillator is analysed experimentally ; the peak wavelength is found to be 2.7 times the total length of the dipole across which sparking occurs. This is rather higher than the theoretical factor, the discrepancy being attributed to self-capacitance associated with the oscillatory dipole.

621.396[.65 + .71 + .812.3 **1102**
"Recent Developments in Communication Engineering".—A. H. Mumford. (*J. Instn elect. Engrs*, Part I, Jan. 1946, Vol. 93, No. 61, pp. 41–50.) The full paper, of which a summary was noticed in 4046 of 1945. Reprinted in *J. Instn elect. Engrs*, Part III, Jan. 1946, Vol. 93, No. 21, pp. 2–11.

VALVES AND THERMIONICS

537.52 **1103**
On the Mechanism of Spark Formation.—Teszner. (*See* 909.)

537.523.2 **1104**
On the Velocity-Distribution of Electrons in [cold-cathode] **Field Emission.**—E. W. Müller. (*Z. Phys.*, 23rd Feb. 1943, Vol. 120, Nos. 5/6, pp. 261–269.) Recent work (see Haefer, 1680 of 1941) has shown good confirmation of the wave-mechanics theory of field emission as given by Sommerfeld, Bethe and Nordheim, in respect of current density, but is less satisfactory in respect of velocity distribution. This is attributed to defects in the methods of measurement used hitherto. The paper describes an investigation based on the use of retarding potentials, in which the previous sources of error have been eliminated. It is shown that the velocity distribution curve has the expected half-value width of 0.4 V, and that the distribution is independent of direction relative to crystal orientation.

621.385 + 621.396/.397 + 621.365.5 **1105**
Results Achieved by a Large [French] **Industrial Research Centre** [Le Groupe Industriel des Compagnies Françaises de T.S.F. Associées] **During and Despite the Occupation.**—P. Brenot. (*Onde élect.*, Sept. 1945, Vol. 25, No. 222, pp. 29–46.) Work on valves included preparation of the following types (*a*) E3056, 300 kW class C amplifier for medium waves, (*b*) television tetrodes, two working as class B amplifiers give 40 kW at 4½ m wavelength, (*c*) a water-cooled 25 kW pentode, (*d*) an air-cooled 7 kW pentode, (*e*) a series of 10 glass-based pentodes, 5 single, 5 double, for wavelengths between 60 cm and 6 m, and power between 2 W and 1 300 W, (*f*) v.m. tubes giving 250–300 W c.w. at 15 and 20 cm wavelengths, (*g*) v.m. tubes nearly ready are for 20–30 cm wavelengths giving 10 kW c.w. and a few hundred kW pulsed peak power, (*h*) v.m. amplifiers giving 250–300 W c.w. at 15–25 cm, (*i*) magnetrons for wavelengths between 10 and 20 cm giving about 100 kW pulsed peak power at 10 cm with 40% efficiency.

Work on multiplex radio communications was carried out at various wavelengths of which the shortest was in the range 21–24 cm. For the latter a 350 W klystron was developed in 1941. The advantages of f.m. and pulse modulation for multiplex systems were investigated with metre and decimetre waves. For decimetre systems relay stations were made to receive and transmit at slightly different wavelengths.

Other work with metre and decimetre waves included investigation of the properties of wave-guides and horns, the development of a radar obstacle-locating system, and of a landing beacon system.

Work was also done on television (including the use of 20 cm klystron amplifiers for single-sideband modulation), electron microscope, a 13 and 50 m broadcasting station, telegraph receivers, undulators, goniometers, and an industrial plant for the surface hardening of steel by 500 kc/s induction heating.

621.385.032.22 **1106**
Effect of Surface Finish and Wall Thickness on the Operating Temperature of Graphite Radio-Tube Anodes.—L. L. Winter & H. G. MacPherson. (*Proc. Inst. Radio Engrs*, Dec. 1945, Vol. 33, No. 12, pp. 834–837.) Experimental investigation of the effect on internal and external temperatures of cylindrical graphite anodes by brightness measurements at 6 530 Å using an optical pyrometer. The emissivity is calculated from the temperature difference: it is 0.89 for a sanded surface and 0.82 for a polished surface. It is concluded that temperature differences not greater than 30° C will occur in radio valve anodes operating at about 1 000° C as a result of differences in surface finish and wall thickness of 0.07–0.1 inch.

621.385.1 : 621.396.611.4 **1107**
Frequency Stabilisation of Resonators Influenced by Electron Discharge.—E.M.I. Laboratories. (*Electronic Engng*, Feb. 1946, Vol. 18, No. 216, p. 61.) Variation of the electron beam may cause frequency variation or detuning of the resonator of a v.m. valve. This difficulty can be surmounted by using a second electron discharge which can be varied so as to modify the unwanted changes in the main beam. Application to a reflection type klystron is described.

621.385.2/.3].012.8 **1108**
Triode Equivalent Circuits.—Bell. (*See* 864.)

621.385.3/.8 **1109**
On the Theory of the Passage of an Electron Current Through Triodes and Multielectrode Tubes and on the Notion of Control Voltage in the Presence of Space Charge.—G. A. Grünberg. (*J. Phys., U.S.S.R.*, 1943, Vol. 7, No. 6, pp. 279–285.) ". . . a method for the approximate calculation of the current passing through a triode is given, under the condition that the current in the tube is limited by a space charge. It is shown that this method leads automatically to the introduction of the notion of a control voltage, giving at the same time the limits of validity of this notion, starting from a consideration of the processes in the tube in the presence of a space charge. The value ϕ_{st} of the control voltage turns out to be somewhat different from the expression $\phi_{st} = (\phi_g + D\phi_a)/(1 + D)$, which corresponds to the usual formula of van der Bijl-Barkhausen-Schottky, and in the case of a cylindrical triode is approximately equal to $\phi_{st} = (\phi_g + D\phi_a)/[1 + D(r_a/r_g)^{\frac{2}{3}}]$ (r_a and r_g are the radii of the anode and of the grid, D—the usual "static" permeability of the tube). It is shown that this difference is of an essential character, and that in the case of grids with sufficiently thin wires a formula of the type $\phi_{st} = (\phi_g + D\phi_a)/[1 + D(r_a/r_g)^{\frac{2}{3}}]$ is the only possible, inasmuch as the control voltage, linearly depending on grid and plate potentials, actually exists. Similar results are obtained for plane triodes. At the end of the paper the method is applied to the determination of the current passing through a cylindrical diode; the comparison of the result obtained with Langmuir's exact solution enables one to estimate the error introduced by the approximate character of the method."

621.385.3 1110
Transmitter Valves – II.—M. Matricon, J. Chantereau, R. Montagne & A. Laurent. (*Onde élect.*, Dec. 1945, Vols. 20/25, No. 225, pp. 155–167.) In two sections: (*a*) The control of materials in valve manufacture. The maintenance of a high standard of uniformity and reliability in valve manufacture requires a routine of precise control of the materials employed. The methods used by the French Thomson-Houston Company include ordinary quantitative chemical analysis of metals and alloys, spectrographic tests of pure metals, mechanical tests on metals, micro-metallography, and X-ray diffraction for the study of the crystalline condition. (*b*) Demountable continuously-evacuated valves. A detailed illustrated description of demountable valves in metal and fused silica, particularly for short wavelengths. Output powers range from 120 to 145 kW for wavelengths 16 to 49 metres. For part I of this paper see 490 of February.

621.385.832.032.29 1111
High-Frequency Deflection System.—G. Rudenberg. (*Electronics*, Dec. 1945, Vol. 18, No. 12, pp. 332..340.) Summary of 3026 of 1945.

621.385.833 1112
Electron Beams in Strong Magnetic Fields.—J. R. Pierce. (*Phys. Rev.*, 1st/15th Nov. 1945, Vol. 68, Nos. 9/10, pp. 229–230.) A mathematical note dealing with the form of the electron beam emitted from a circular disk cathode in the presence of a strong uniform magnetic field applied in a direction normal to the cathode.

621.396.822 1113
A Theory of Valve and Circuit Noise.—Campbell & Francis. (*See* 1037.)

621.396.822 1114
[Noise] **Fluctuations of Electric Current.**—Bell. (*See* 1038.)

621.385 1115
Radio Valve Vade Mecum, 1945 [Book Review].—P. H. Brans. Algemeene en Technische Boekhandel, Antwerp. (*Wireless Engr*, Jan. 1946, Vol. 23, No. 268, p. 15.) Gives in Flemish, French, English and German, characteristics and base connexions of many British, U.S., and Continental (including Russian) valves.

MISCELLANEOUS

001.83 1116
International Scientific Co-Operation.—F. J. M. Stratton. (*Nature, Lond.*, 26th Jan. 1946, Vol. 157, No. 3978, pp. 96–98.) An account of the organization and activities of the International Council of Scientific Unions, and of the recent meeting of its Executive Committee.

001.89 : 359 1117
The Engineer's Place in Naval Research.—W. G. Schindler. (*Proc. Inst. Radio Engrs*, Dec. 1945, Vol. 33, No. 12, pp. 823–824.) Review of organization of technical research for the U.S. Navy in both government and industrial laboratories.

001.891 1118
Scientific Research and Industrial Planning: British Association Conference.—(*Nature, Lond.*, 5th Jan. 1946, Vol. 157, No. 3975, pp. 8–11.) The conference lasted two days and many speakers eminent in the scientific world took part. "The general theme . . . was essentially the human element in research: the research worker himself and the conditions which are most likely to stimulate creative work; the place of the human factor in production and the contribution which scientific research can render to efficiency." At the end a resolution was passed approving the statement of President Truman, Mr. Attlee and Mr. MacKenzie King on atomic energy.

001.92 : [5 + 6 1119
Science and the Public.—(*Nature, Lond.*, 26th Jan. 1946, Vol. 157, No. 3978, pp. 110–111.) Report of a Royal Institute of Chemistry discussion on "The Publicity of Science, with particular reference to Chemistry", covering the press, exhibitions, films and broadcasting.

06.054 1120
The Presentation of Technical Developments Before Professional Societies.—W. L. Everitt. (*Proc. Instn Radio Engrs, Aust.*, Sept 1945, Vol. 6, No. 3, pp. 15–17.) Reprint of 497 of February.

06.055 1121
Preparation and Publication of I.R.E. Papers.—H. M. Stote. (*Waves and Electrons*, Jan. 1946, Vol. 1, No. 1, pp. 5–9.) Description of the Editorial Department's method of dealing with papers for publication in *Proc. Inst. Radio Engrs, N.Y.*, and in *Waves and Electrons*, with advice to authors on their preparation.

061.22 : 621.396 (054) 1122
Waves and Electrons.—A new monthly publication of the Institute of Radio Engineers, bound together with the monthly parts of *Proc. Inst. Radio Engrs, N.Y.*, intended to contain papers on the engineering development of radio and electronic equipment, tutorial and historical papers, etc.

061.3 : 621.396] (7/8) 1123
The Inter-American Radio Conference of Rio.—A. L. Budlong. (*QST*, Dec. 1945, Vol. 29, No. 12, pp. 33–35.) The principal objective of the third Inter-American Radio Conference, held in Rio de Janeiro, September, 1945, was to rewrite the Habana Convention and to revise the organization of the Inter-American Radio Office.

6(07) 1124
A Reply From Industry to the Colleges.—M. M. Boring, A. R. Stevenson, Jr., & K. B. McEachron, Jr. (*Elect. Engng, N.Y.*, Nov. 1945, Vol. 64, No. 11, pp. 398–400.) A discussion on the future curricula of U.S. engineering colleges, suggesting that war-time accelerated courses should be abandoned and specialization avoided in favour of broad engineering fundamentals. An attempt to stimulate ingenuity and encourage leadership is advocated, and emphasis given to preparation for citizenship by the inclusion of non-technical subjects.

6(07) 1125
Education and Training for Engineers: Second Report – Part-Time Further Education at Technical Colleges.—I.E.E. Subcommittee. (*J. Instn élect. Engrs*, Part I, Nov. 1945, Vol. 92, No. 59, pp. 416–422.) Report of a discussion of 3160 of 1945.

6(07) 1126
The Education of an Engineer as an Administrator.—W. E. Clegg. (*J. Instn Engrs, Aust.*, Sept. 1945, Vol. 17, No. 9, pp. 173–180.)

6(07) 1127
"A Senior Engineer's Views on Technical Education and Training".—S. H. Richards. (*J. Instn elect. Engrs*, Part I, Jan. 1946, Vol. 93, No. 61, pp. 33–36.) Discusses implementation of the two I.E.E. Reports on Education and Training of Engineers (2545 of 1943, and 3160 of 1945). Inaugural address of Chairman, Measurements Section, I.E.E.

6(07) : [621.396/.397 1128
"Apprenticeship and Training Systems in the Radio Industry".—(*J. Instn elect. Engrs*, Part III, Dec. 1945, Vol. 92, No. 20, pp. 259–260.) Report of I.E.E. discussion led by J. Greig considering the electrical and mechanical craftsmanship required in industry, and the relative importance of physics and engineering in the training of technicians for research and development.

616-001 : 537.533 1129
Dangers Inherent in Scattered Cathode Rays.—L. L. Robbins. (*Science*, 14th Dec. 1945, Vol. 102, No. 2659, p. 623.) A note on burns experienced after brief exposure to scattered electrons. In some respects these are similar to ordinary X-ray or thermal burns. Such scattered electrons appear to have a very limited depth of penetration.

621.316.9 1130
Electrical Accidents.—"Supervisor". (*Electrician*, 28th Sept. 1945, Vol. 135, No. 3513, pp. 319–320.) Discussion of "Electrical Accidents", annual report of the Senior Electrical Inspector of Factories. For summaries of the report, see *Electrician*, 31st Aug. 1945, Vol. 135, No. 3509, pp. 216–218, also 4104 of 1945.

621.38 + 621.396.69 1131
Electronics at the Physical Society's Exhibition.—(*Electronic Engng*, Jan. 1946, Vol. 18, No. 215, pp. 13–20.) Incomplete descriptive alphabetical list of equipment of general electronic interest shown at the exhibition.

621.38 + 621.396.69 1132
British Electronic Apparatus : Further Examples of Electronic Equipment shown at the Physical Society's Exhibition.—(*Electronic Engng*, Feb. 1946, Vol. 18, No. 216, pp. 34–44.) Continuation of 1131 above, including classified list of products and exhibitors.

621.38 + 621.396.69 1133
Physical Society's Exhibition.—(*Wireless Engr*, Feb. 1946, Vol. 23, No. 269, pp. 58–62.) The exhibition ". . . was characterized by two things —the overwhelming preponderance of wireless and allied equipment and the enormous variety of such apparatus." The review is divided into the following sections :—measuring instruments : a.f. oscillators : v.h.f. equipment : industrial electronics : components, accessories and materials.

621.396.721 1134
Ham Radio.—R. Washburne & A. Williams. (*Radio News*, Dec. 1945, Vol. 34, No. 6, pp. 38..92). Future of the U.S. amateur field. Code of conduct and postwar frequency allocations. Application of wartime developments in components, tubes, etc., and amateur use of new techniques of f.m., facsimile and radar. A 9-page article.

621.396.721 1135
The New Licence.—(*R.S.G.B. Bull.*, Dec. 1945, Vol. 21, No. 6, pp. 83–84.) The conditions for the new transmitting licence for amateurs in Great Britain.

621.396.82 : 34 1136
Wireless Legislation.—(*Wireless World*, Jan. 1946, Vol. 52, No. 1, pp. 6–7.) A short account of the attitude of English law to radio transmissions and interference.

658.5 : 621.396.6 1137
Case Studies of Production.—M. Lechner. (*Electronics*, Dec. 1945, Vol. 18, No. 12, pp. 140–146.) The output of electronic apparatus was greatly increased by the use of suitable jigs and fixtures along the production line, and by instilling a sense of responsibility and interest into the persons concerned.

001.89 : 6 1138
The Future of Industrial Research [Book Review].—Standard Oil Development Company. Journal of Chemical Education, Easton, Pennsylvania, 1945, 173 pp. (*Rev. sci. Instrum.*, Oct. 1945, Vol. 16, No. 10, pp. 289–290.) Combination of twelve papers presented by prominent Americans, with comments and discussion. ". . . it undoubtedly is the most faithful and compact record of qualified opinion on the subject yet offered."

621.3 1139
Electromagnetic Engineering : Vol. I. Fundamentals [Book Review].—R. W. P. King. McGraw-Hill Co., New York, 1945, 580 pp., $6.00. (*Rev. sci. Instrum.*, Oct. 1945, Vol. 16, No. 10, pp. 287–288.)

621.3 (031) 1140
"Electrical Engineer" Reference Book [Book Review].—E. Molloy, M. G. Say, R. C. Walker & G. Windred (Eds.). Geo. Newnes Ltd., London, 42s. (*Electronic Engng*, Feb. 1946, Vol. 18, No. 216, p. 62.) "The professional engineer will find this book an indispensable source of reference and information . . .".

621.392.081 1141
The Decibel Notation and its Applications to Radio Engineering and Acoustics [Book Review].—V. V. L. Rao. Addison & Co., Madras, India, 1944, 176 pp., $3.00 (in U.S.A.). (*Proc. Inst. Radio Engrs*, Dec. 1945, Vol. 33, No. 12, p. 903.)

621.396(05) 1142
Annales de Radioélectricité [Book Review].—(*Wireless Engr*, Feb. 1946, Vol. 23, No. 269, p. 57.) The first number of a new journal published by the "Centre d'information et de documentation" of the Compagnie Générale de T.S.F. "promises to make a very useful contribution to the current literature of radio research and development".

621.396.029.6 1143
U.H.F. Radio Simplified [Book Review].—M. S. Kiver. D. Van Nostrand Co., New York, 1945, 236 pp., $3.25. (*Proc. Inst. Radio Engrs*, Dec. 1945, Vol. 33, No. 12, p. 903.) ". . . intended primarily for the use of radio amateurs ; laboratory and technical assistants . . . and others who have some technical knowledge and experience, would profit by a simple descriptive treatment of the subject."