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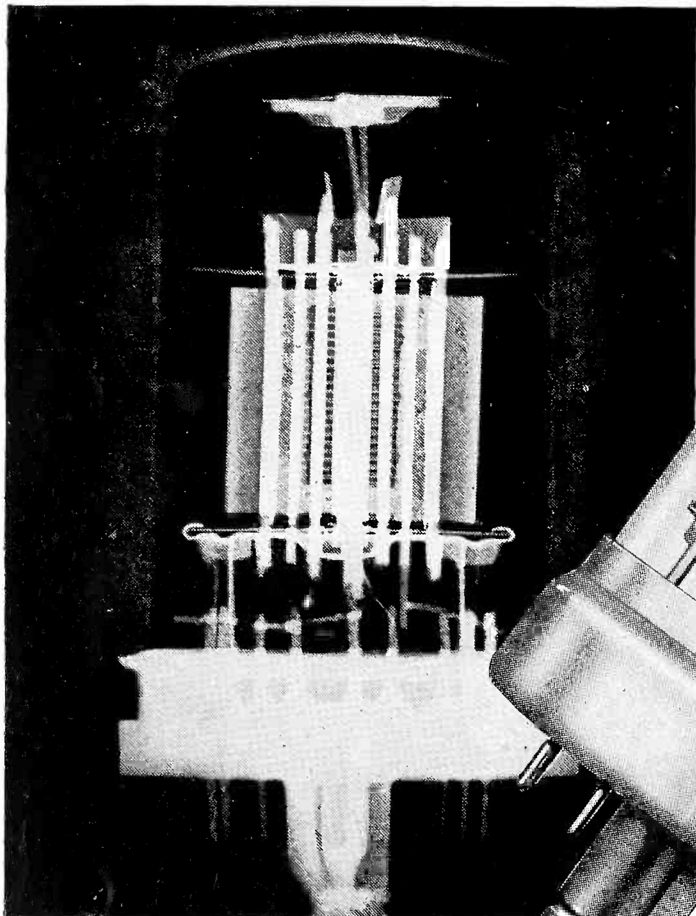

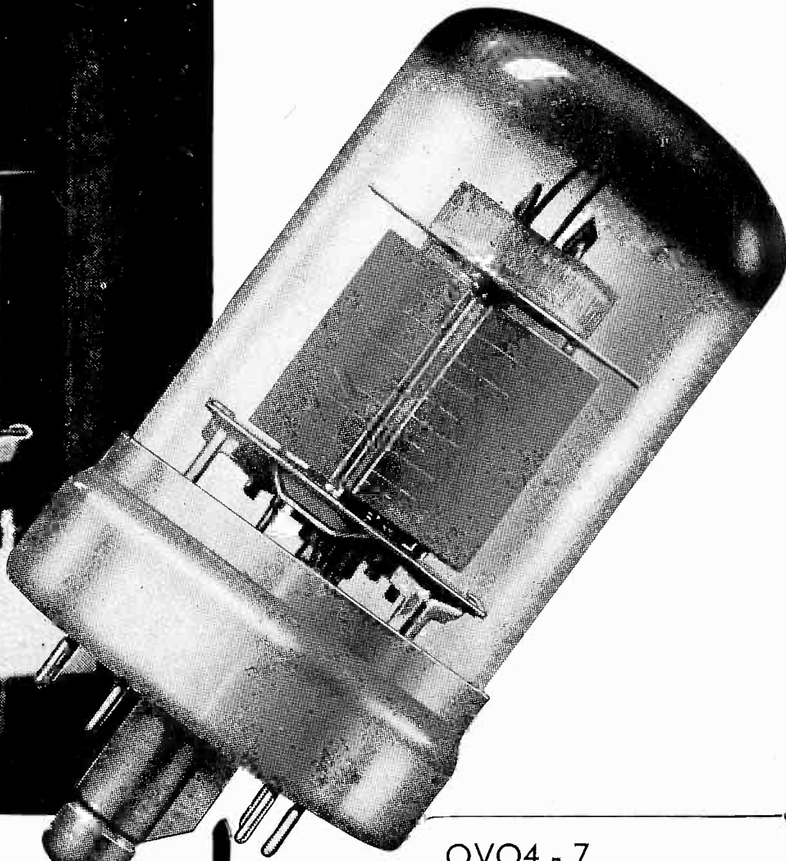

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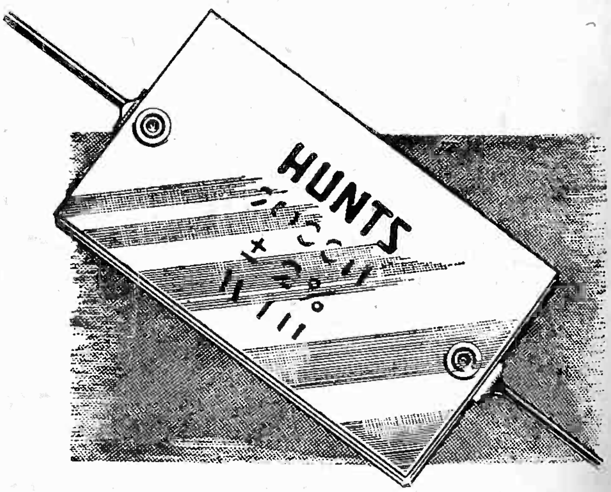

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EDITORIAL

Simple Transmission Formula

UNDER this title H. T. Friis of the Bell Telephone Laboratories, published a note in the *Proc. I.R.E.* of May 1946. He makes some novel and interesting suggestions concerning the way in which the relation between the radiated and received power should be stated. The basic idea is that an aerial has an effective area which is defined, whether the aerial is actually used for transmitting or receiving, by assuming it to be used as a receiver of a linearly polarized plane wave. If P_0 is the power flow per unit area in the neighbourhood of the aerial, and P_r is the power available at the output terminals of the aerial then the effective area A_r is defined as P_r/P_0. An aerial can have different effective areas in different directions, and the effective area will be modified by variations of the load resistance into which the aerial feeds. In any transmission formula the areas must, of course, be those in the direction of transmission. If P_t be the power fed into the transmitting aerial at its input terminals and P_r the power available at the output terminals of the receiving aerial then

$$P_r/P_t = A_r A_t/d^2\lambda^2$$

where A_r and A_t are the effective areas of the two aerials and d their distance apart. Since the right-hand side of the formula has dimensions L^4/L^4, it is immaterial what units are employed, so long as they are the same throughout.

Considering a small dipole of negligible resistance and length l with a load resistance equal to its radiation resistance, i.e. a small uniform current element, situated in an electric field of strength $\mathcal{E}$ r.m.s. volts per unit length, then

$$I = \frac{\mathcal{E}l}{R_l + R_{rad}} = \frac{\mathcal{E}l}{2R_{rad}} \text{ and}$$

$$P_r = I^2 R_l = \frac{\mathcal{E}^2 l^2}{4R_{rad}} \text{ watts.}$$

Now for such an element

$$R_{rad} = 80\pi^2 l^2/\lambda^2 \text{ ohms}$$

hence $$P_r = \frac{\mathcal{E}^2\lambda^2}{320\pi^2} \text{ watts}$$

and since $P_0 = \mathcal{E}^2/120\pi$ watts

$$A = P_r/P_0 = 3\lambda^2/8\pi = 0.1193\lambda^2.$$

This would apply to a Hertzian oscillator with such large terminal capacitances that the current is approximately uniform over its length.

In the case of a half-wave dipole the current is distributed sinusoidally over its length and since its effective length is therefore $2l/\pi$

$$I = \frac{2}{\pi} \cdot \frac{\mathcal{E}l}{2R_{rad}} \text{ and } P_r = \frac{\mathcal{E}^2 l^2}{\pi^2 R_{rad}} \text{ watts.}$$

In this case $R_{rad} = 73.2$ ohms (see Editorial April 1945) and $l = \lambda/2$,

therefore $$P_r = \frac{\mathcal{E}^2\lambda^2}{4\pi^2 \times 73.2}$$

and $$A = P_r/P_0 = \frac{120\lambda^2}{4\pi \times 73.2} = 0.1305\lambda^2.$$

It is interesting to note that this effective area of a half-wave dipole is a little more than that of a rectangle of the same length as the aerial with a width equal to half the length. It is the power transmitted through this rectangle that the aerial abstracts and delivers to the load resistance. It will be noticed that the effective area of the Hertzian oscillator for the same wavelength is only slightly less, although its length may be very much less, but to obtain the assumed uniformity of current, large terminal capacitance is essential and we may picture the aerial as a short wire joining the centres of two large parallel plates. Hence the effective area gains in width what it loses in length.

Isotropic Aerial

This is the name given by the author to a hypothetical aerial which has the same radiation intensity in all directions. It will have an effective area equal to two-thirds of that of the small dipole, for if we consider a spherical shell around the transmitter, the energy per cm^3 will be the same everywhere if the transmitter is isotropic, and the energy in any narrow zone will be proportional to the volume of the zone, that is, to $\sin\theta$ where θ is the angle to the polar axis, whereas, if the transmitter is an ordinary small dipole the energy density is proportional to $\sin^2\theta$, and therefore the energy in any narrow zone will be proportional to $\sin^3\theta$. On integrating from 0 to π to find the total energy, $\sin\theta$ gives 2 and $\sin^3\theta$ 4/3; hence for the same current and the same radiation in the equatorial plane the isotropic aerial radiates 1.5 times as much power as the dipole. Its radiation resistance is therefore 1.5 times as great and its received power for a given P_0 only two-thirds of the received power of the dipole; its effective area is therefore two-thirds of that of the dipole.

We have seen that for a short dipole $A = 3\lambda^2/8\pi$; hence for the isotropic aerial $A = \lambda^2/4\pi$. If such an isotropic aerial were radiating a power P_t, the value of P_0 at a distance d would be $P_t/4\pi d^2$ which we see is equal to $P_tA/d^2\lambda^2$, and if A_t and A_r are the respective effective areas of the sending and receiving aerials, we have

$$P_r/P_t = A_rA_t/d^2\lambda^2$$

Any change in the character of either aerial causes an appropriate change in its effective area which must be known before the above formula can be used.

Broadside Arrays

A broadside array of n aerials each $\lambda/2$ long and $\lambda/2$ apart with a reflector that doubles the gain is found to have an effective area of $n \times \lambda/2 \times \lambda/2$ that is, about equal to the actual area. This is stated to agree with calculations published by Pistolkors in *Proc. I.R.E.* in 1929.

Experiment shows that aerials with parabolic reflectors have an effective area equal to about two-thirds of the projected area of the reflector; for a long horn it was found to be 81 per cent, and for an optimum horn, designed to give maximum gain, 50 per cent of the aperture area.

In conclusion, the author says "It is suggested that radio engineers hereafter give the radiation from a transmitting aerial in terms of the power flow per unit area, which is equal to P_tA_t/λ^2d^2, instead of giving the field strength in volts per metre. It is also suggested that an antenna be characterized by its effective area instead of by its power gain or radiation resistance. The ratio of the effective area to the actual area of the aperture of an antenna is also of importance in antenna design, since it gives an indication of how efficiently the antenna is utilizing the physical space it occupies."

With regard to the first suggestion, since $P_0 = \mathcal{E}^2/120\pi$ it seems immaterial whether one uses the power flow P_0 per unit area or the field strength $\mathcal{E}$, but if one uses P_0 then the effective area is obviously more convenient than the radiation resistance; if, however, one is given the field strength $\mathcal{E}$ then the radiation resistance is the more convenient characteristic. "Power gain" seems a vague and unsatisfactory term. If an aerial has an extensive reflector such as a horn or parabola the area of the aperture is quite definite, but in many cases "the actual area of the aperture of an antenna" can bear little relation to the effective area.

G. W. O. H.

OSCILLATOR POWER RELATIONS*

By R. E. Burgess, B.Sc.

(Communication from the National Physical Laboratory)

SUMMARY.—The amplitude and power relations are derived for a class of valve-maintained oscillators in which the source of power can be represented as a negative conductance element which has a characteristic limited by a term proportional to a higher odd-power of the voltage. The analysis is based on the classical work of E. V. Appleton and B. van der Pol. The coupling conditions for obtaining the maximum output power from such a source are deduced and shown to differ fundamentally from the impedance-match conditions appropriate to linear systems. It is shown that the intrinsic oscillatory-circuit losses are purely parasitic in the transfer of power to an external load-circuit, and there is no question of a resistance match.

The response of such an oscillator circuit to a small external e.m.f. having a frequency different from the oscillation frequency is considered. It is found that the circuit effectively has a positive conductance which is proportional to the excess negative conductance producing oscillation.

1. Introduction

IT is frequently required to couple a load to a valve-maintained oscillator and to know how the power obtained will depend upon the coupling, and in particular how to obtain the maximum power in the load. In such problems it is important to remember that the valve-maintained oscillator is a non-linear source of power, and that the usual linear-network theorems (e.g., Superposition or Thévenin's) are inapplicable. Derived quantities such as impedance and admittance will, in such cases, depend upon arbitrary definition, and it is therefore preferable to describe the behaviour of such systems in terms of directly observable quantities such as current, voltage and power.

The present note illustrates this by reference to a class of oscillators in which the non-linear element which both maintains and limits the oscillation is representable as a negative conductance which can be expressed as a simple binomial function of voltage. The analysis is based on the classical work of Appleton and van der Pol[1, 2, 3, 4, 5] and its development is limited to power relationships, and their significance in practical problems.

2. Amplitude Relations

The circuit considered is shown symbolically in Fig. 1, in which L, C and G represent the oscillatory circuit and N the negative conductance element which maintains oscillation. It is assumed that the current-voltage characteristic of N has the form

$$i = -av + bv^{2n+1} \quad \ldots \quad (1)$$

where the first term represents the main component of the falling characteristic and the second term the essential non-linearity which limits the amplitude of oscillation. This type of characteristic is a sufficiently close representation of a wide range of actual cases to enable the essential features of their behaviour to be deduced. The omission of even-power terms is without effect on the

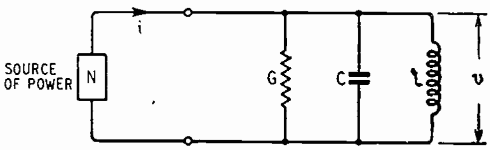

Fig. 1. *Schematic circuit of valve oscillator.*

amplitude while the inclusion of only one odd-power term avoids the possibility of hysteresis and the existence of more than one stable amplitude[3]. Fig. 2 illustrates the characteristics for $n = 1$, 4 and ∞. The latter case may be termed "ideal" and is a limiting form not practically realizable.

The turning values v_1 of v are given by

$$v_1 = \pm \left(\frac{1}{2n+1} \cdot \frac{a}{b}\right)^{1/2n} \quad \ldots \quad (2)$$

and are of particular importance. In terms of v_1, the characteristic becomes :—

$$i = -av\left\{1 - \frac{1}{2n+1} \cdot \left(\frac{v}{v_1}\right)^{2n}\right\} \ldots \quad (3)$$

The range $-v_1$ to $+v_1$ is that for which N has a negative conductance. The conductance is conveniently expressible as

$$g = \frac{di}{dv} = -a\left[1 - \left(\frac{v}{v_1}\right)^{2n}\right] \quad \ldots \quad (4)$$

* MS accepted by the Editor, February 1946.

For the calculation of the amplitude and power relationships it will be assumed that the potential difference v across the oscillatory circuit is a pure sine wave

$$v = A \sin \omega_0 t \qquad \ldots \quad \ldots \quad \ldots \quad (5)$$

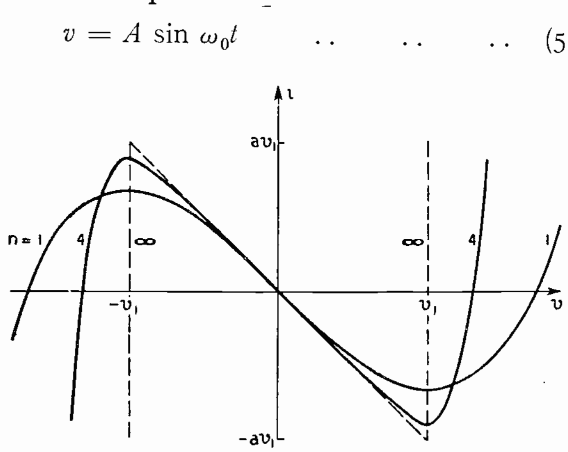

Fig. 2. *Typical current-voltage characteristic of* N $(i = -av + bv^{2n+1})$

Small departures from this wave form will not materially affect the amplitude and power relationship with which we are now concerned.

The amplitude of oscillation, A, can be deduced from the balance over a period of oscillation, T of the power Gv^2 expended in the circuit conductance G and the power $-iv$ supplied by the source.

$$O = \int_0^T (Gv^2 + iv)dt$$
$$= \frac{A^2T}{2}(G - a + A^{2n}b\,F_n)$$

where

$$F_n = \frac{1}{\pi}\int_0^{2\pi} \sin^{(2n+2)}\theta\, d\theta$$
$$= \frac{3.5\ \ldots\ (2n+1)}{4.6\ \ldots\ (2n+2)} \qquad \ldots \quad \ldots \quad (6)$$

is a numerical factor, less than unity, which depends upon the characteristic of the negative conductance element.

Thus,

$$A = \left(\frac{a - G}{bF_n}\right)^{1/2n} \qquad \ldots \quad \ldots \quad \ldots \quad (7)$$

or, in terms of v_1

$$A = v_1\left(\frac{2n+1}{F_n} \cdot \frac{a-G}{a}\right)^{1/2n} \qquad \ldots \quad (8)$$

Now a is the magnitude of the negative conductance N for small oscillations, and for oscillation to occur, a must exceed G, the amplitude depending on the excess negative conductance $(a - G)$. In the limiting case of $G = 0$ the amplitude A has its maximum possible value of

$$A_0 = v_1\left(\frac{2n+1}{F_n}\right)^{1/2n} \qquad \ldots \quad \ldots \quad (9)$$

which decreases from $2v_1$ at $n = 1$ to $1.44v_1$ at $n = 4$ and tends to v_1 as n tends to infinity. The maximum possible amplitude of oscillation is therefore equal to $2v_1$ which is only attainable with a cubic characteristic and zero circuit losses.

3. Optimum Power Relations

It will now be assumed that the effective conductance of the oscillatory circuit is continuously variable, either by adjustment of the coupling to a load or direct variation of its effective resistance or in any other way. The power dissipated in the circuit is

$$P = \tfrac{1}{2}A^2G = \tfrac{1}{2}v_1^2G\left[\frac{2n+1}{F_n}\ \frac{a-G}{a}\right]^{1/n} \qquad \ldots \quad \ldots \quad (10)$$

This reaches a maximum value of

$$P_{\max} = \tfrac{1}{2}v_1^2a\,.\,H_n \qquad \ldots \quad \ldots \quad (11)$$

where

$$H_n = \frac{n}{n+1}\left(\frac{2n+1}{n+1}\cdot\frac{1}{F_n}\right)^{1/n}$$
$$= \frac{4n}{n+1}\left(\frac{n\,!^2}{2n\,!}\right)^{1/n} \qquad \ldots \quad \ldots \quad (12)$$

when the circuit conductance G has its optimum value

$$G_{\text{opt}} = \frac{n}{n+1}\,.\,a \qquad \ldots \quad \ldots \quad \ldots \quad (13)$$

As a function of n, H_n varies as shown in the following table :—

n	1	2	3	4	5	6	---	∞
H_n	1	1.089	1.105	1.106	1.103	1.099	---	1

that is to say, it is a slowly varying quantity which differs very little from unity. *Therefore the maximum power available from the negative conductance is very approximately equal to that which would be expended in a resistance* $1/a$ *by a sinusoidal oscillation of amplitude* v_1.

Hence the slope of the characteristic of N at the origin and the range of voltage over which the slope is negative are the significant parameters in determining the available power of the element.

4. Power Transfer to a Load

In many cases the oscillatory circuit will be used to supply power to an external load. Such cases, however, do not need any separate analysis, for the above discussion of the optimum power relationships is quite independent of the constitution of the conductance G, which may be wholly due to the intrinsic resistance of the oscillatory circuit, or may be chiefly due to the coupling of the oscillatory circuit to its load. Since, in practice, the useful power is that delivered to the external load, the conductance of the oscillatory circuit itself can be taken into account by subtracting it from the linear term " a " of the negative conductance of the source of power. If " a " is reduced in this way, the relations of the preceding section will be applicable to the transfer of power to an external load G.

The important thing to note is that there is no question of matching the tuned circuit to the load as is sometimes wrongly assumed. The tuned circuit is necessary in order to provide the desired sine-wave oscillation and to determine the frequency but it is an agent as far as the power transfer is concerned, and the loss involved is purely parasitic.

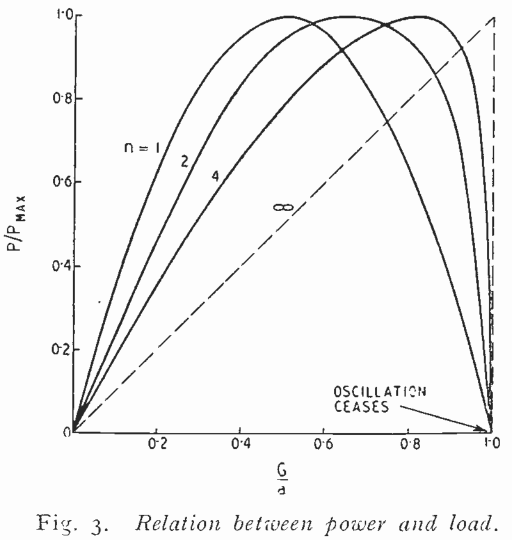

Fig. 3. *Relation between power and load.*

The system is comparable with the transfer of power from a linear source of power to a load by way of a tuned circuit, the analysis of which has been given by Colebrook.[6]

5. Variation of Power and Amplitude with Load

From equations (10) and (11) we can express $P/P_{\max}$ as a function of G/a, i.e.,

$$\frac{P}{P_{\max}} = \frac{n+1}{n} \cdot \frac{G}{a}\left[(n+1)\left(1-\frac{G}{a}\right)\right]^{1/n} \quad \ldots \quad \ldots \quad (14)$$

This is shown in Fig. 3 for $n = 1, 2, 4$ and ∞. It will be seen that for the ideal negative conductance ($n \to \infty$), the power increases linearly with G, since the amplitude remains constant up to $G = a$, at which oscillation will cease abruptly.

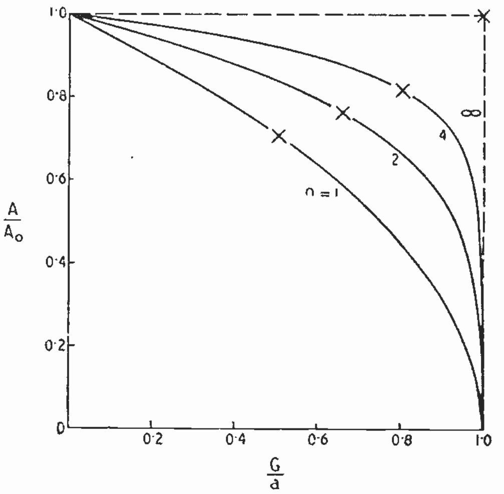

Fig. 4. *Dependence of amplitude on load. Crosses denote load for maximum load power.*

It is noted that the dependence of power on load is more critical than in the case of a linear source of power obeying Thévenin's Theorem and this is due to the reduction of amplitude (eventually to zero) as the loading is increased.

From equation (8), the variation of the amplitude of oscillation with the loading G is given by

$$\frac{A}{A_0} = \left(1 - \frac{G}{a}\right)^{1/2n} \quad \ldots \quad \ldots \quad (15)$$

where A_0 is the amplitude corresponding to $G = 0$. This relation is shown in Fig. 4 and it will be seen that as n increases, the amplitude tends to remain more constant up to the point at which oscillation ceases.

The dependence of the amplitude and power on the load emphasizes the essential differences between a linear and a non-linear source of power. Clearly one cannot ascribe an effective resistance to an oscillator without introducing an arbitrary definition or method of measurement upon which the apparent value will depend. For this reason it is desirable to consider such non-linear systems in terms of observable quantities, viz., current, voltage or power.

6. Response of the Oscillator to an External E.M.F.

The response of an oscillator to a small external e.m.f. of frequency different from the oscillation frequency is of interest in a number of connections. For example:

(i) the influence of the inherent thermal and shot noise in the oscillator on its output voltage;

(ii) the reception of signals by an oscillating detector;

(iii) the magnitude of the sidebands when the oscillator is amplitude-modulated.

Let the applied e.m.f. have the form

$$e = E \sin \omega t$$

and be impressed in series with the inductive arm L of the circuit. The differential equation of the system is then

$$L\frac{d}{dt}\left[i + Gv + C\frac{dv}{dt}\right] + v = E \sin \omega t$$

or $$v'' + \frac{Gv' + i'}{C} + \omega_0^2 v = \omega_0^2 E \sin \omega t \qquad (16)$$

where $\omega_0^2 = 1/LC$ and the primes denote time differentials.

The resultant voltage across the oscillatory circuit has the form appropriate to the sum of a self-maintained oscillation and a forced vibration, the impedance of the circuit being taken as negligible at harmonic and heterodyne frequencies:

$$v = A \sin \omega_0 t + U \sin (\omega t + \phi) \qquad (17)$$

It will be assumed that the applied e.m.f. is small so that U is small compared with A, and it is found on substituting for v in equation (16) and expanding as far as U^2/A^2, that the amplitude of the self-oscillation of angular frequency is given by

$$A \approx \left(\frac{a - G}{bF_n}\right)^{1/2n} \left[1 + n(n+1)\frac{U^2}{A^2}\right]^{-1/2n}$$
$$\approx \left[A_1^2 - (n+1)U^2\right]^{1/2} \qquad (18)$$

where $A_1 = \left(\dfrac{a - G}{bF_n}\right)^{1/2n}$ is the amplitude in the absence of an external e.m.f. (Equation 7).

The amplitude of the forced vibration of angular frequency ω is given by

$$U = \frac{E}{\sqrt{(1 - \omega^2 LC)^2 + \omega^2 L^2 \{n(a - G)\}^2}} \qquad (19)$$

which is the form appropriate to a linear circuit having the constants L, C and G' where the effective conductance G' is given by

$$G' = n(a - G) \qquad (20)$$

which may be greater or less than the true circuit conductance G.

The effective bandwidth of the system between half-power points for an external e.m.f. is thus

$$B = \frac{G'}{2\pi C} = \frac{n(a - G)}{2\pi C} = \frac{f_0}{Q} n\left(\frac{a}{G} - 1\right) \qquad (21)$$

where $Q = \dfrac{\omega_0 C}{G}$ is the magnification of the oscillatory circuit by itself.

As might be expected G' is equal to the sum of the circuit conductance G and the average value of the conductance of N over a cycle of oscillation:

$$G + \bar{g}$$
$$= G + \frac{1}{2\pi}\int_0^{2\pi}\left[-a + (2n+1)b(A \sin \theta)^{2n}\right] . d\theta$$
$$= G + \left[-a + (n+1)(a - G)\right]$$
$$= n(a - G) \qquad (22)$$

The effective magnification U/E and hence the selectivity of the circuit for a small applied voltage of frequency different from the oscillation frequency is seen to be the greater the nearer the system is to the edge of oscillation (i.e. the smaller $a - G$ is). If the excitation (a/G) is great, however, the circuit is apparently heavily damped and will have low selectivity.

7. Acknowledgements

The author is indebted to Mr. F. M. Colebrook for suggestions regarding the problem and its presentation. The work was carried out as part of the programme of the Radio Research Board and this paper is published by permission of the Department of Scientific and Industrial Research.

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SIGNAL-NOISE RATIO AT V.H.F.*

By M. J. O. Strutt and A. van der Ziel

(Natuurkundig Laboratorium der N. V. Philips' Gloeilampenfabrieken.)

SUMMARY.—In Section 1 the two basic concepts of correlation between fluctuation currents or voltages and of the decomposition of fluctuating quantities into singly periodic components are discussed briefly. Section 2 considers the different sources of noise at the input of a triode used in an amplifier circuit. The fluctuation voltage as well as the signal voltage between grid and cathode are evaluated. Thereupon the corresponding signal current and fluctuation current in the short-circuit connection between anode and cathode are given. By division of the latter by the former current the noise ratio is found in Section 3. This noise ratio consists of two components. The one due to the presence of the valve may be minimized by a proper choice of aerial coupling and of detuning (if necessary) of the input circuit. The most favourable values of these two quantities are calculated. The minimum value of the valve's contribution to the noise ratio thus obtained is shown in a set of curves as dependent on two parameters. These results are discussed in Section 4. An experimental example is given, showing the reduction of the noise ratio by detuning the input circuit. This experimental evidence is shown to confirm the theoretical results.

Section 5 deals with a grounded-grid stage. Upon proper handling the equations of the grounded-cathode case are shown to be valid with irrelevant changes of symbols. Hence the chief results are found to be also dependent on the same set of curves. A considerable possible decrease of noise ratio is conjectured. A velocity-modulation valve is considered in Section 6 instead of the density-modulation triode in the preceding sections. Here, again, the discussion and equations of Section 3 may be directly applied with slight changes. Due to the special conditions of the present case the minimum noise ratio turns out to be rather high. A special device by which it might be brought down considerably is indicated. Section 7 deals with the requirements as to valve construction resulting from the preceding discussion. The main points are: uniform electron paths and speeds in the valve and low dielectric and other losses at the input electrodes. It is shown that the proposed reduction of noise ratio applies also to wide-band reception (e.g. 10 Mc/s band-width). In order to compensate the loss of gain incurred by the reduction of noise ratio suitable feedback may be applied. The proposed reduction is not confined to amplifier stages but may also be applied to mixer stages. In the Appendix the interrelation of the noise figures introduced by K. Fraenz, D. O. North, H. T. Friis and the present noise ratio are discussed.

CONTENTS

Introduction

N ultra-high frequency reception the signal-to-noise ratio is often much lower than with similar signals at lower equencies. On the other hand, man-made oise and noise from extra-terrestrial sources ay be less. Hence an increase of the said tio by proper stage design and valve conruction seems especially appropriate in the wer meter-wave and in the decimeter-wave ands.

Several definitions of noise figures have been iblished since 1941 (references [3, 4, 5, 6, 8, 10, 13, 18]). seems, however, that some discussion is ill needed before the universal adoption of one particular figure is secured. Therefore, no explicit use will be made of any of these definitions in this article. Instead, we shall make use of a noise ratio, defined in Section 2.

Two concepts are fundamental to a full understanding of the present discussion. The first is the concept of the correlation of spontaneous fluctuations. Two spontaneous fluctuation currents flowing through the same conductor are completely correlated, if the resulting r.m.s. fluctuation current is obtained by adding the individual r.m.s. current values. They are incorrelated if the mean-square value of the resulting current is obtained by adding the mean-square values of the individual currents. We do not need the definition of partly correlated currents.

The second is the concept of the decomposition of spontaneous fluctuation currents into an infinite sum of single harmonic components. The latter represent simple alternating currents and we may use them in the calculations as such. If the frequency-band within which the spontaneous fluctuation currents

* MS. accepted by the Editor, January, 1946.

are considered is sufficiently narrow compared to its average frequency value, the current in question behaves as a singly periodic alternating current of fluctuating amplitude.

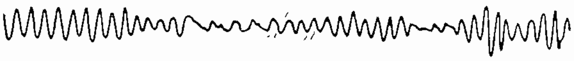

Fig. 1. *Fluctuating current corresponding to a frequency band Δf centred round a frequency f_o such that $f_o \gg \Delta f$. In this case $f_o \approx 14\ \Delta f$, and the amplitude shows random fluctuations of average period approximately equal to $1/\Delta f$.*

The average period of fluctuation is approximately equal to the reciprocal value of the said frequency-band. Perhaps Fig. 1, showing the experimental representation of such a fluctuating current may be helpful to visualize its action in a complex network, since it is similar to that of an ordinary alternating current.*

2. Noise Ratio of a Grounded-cathode Amplifier

Signal currents and voltages will be denoted by I and V respectively with suffixes to indicate different individual values. Fluctuation currents and voltages in the above sense will be denoted by i and v respectively, again using suffixes to indicate different values. The signal at the amplifier input may be supposed to be due to a receiving aerial of effective resistance R_a at its terminals. A triode will be used in the stage under consideration in order to avoid unessential complications attending the use of tetrodes or pentodes (the latter are dealt with in Section 4). The cathode is supposed to be the electrode which is common to input and output, thus forming a so-called "grounded-cathode" amplifier stage. The anode will be assumed to be connected to the cathode by a lead of very small impedance compared to the internal valve impedance between anode and cathode. We shall evaluate the signal current I and the fluctuation current i flowing in this short-circuit connection. The ratio $\overline{i^2}/I^2$ is called the *noise-ratio* of the amplifier stage. The horizontal dash above i^2 denotes its average (arithmetic mean) value, taken either by considering a lapse of time, very long compared with the main periods involved in i, or by considering a great number of identical amplifiers and averaging the values of i^2 obtained. This noise ratio is related by simple formulæ to the noise-figures proposed in previous publications. We need not, however, go into these relations here (see Appendix).

Coupling between the signal source, e.g. the aerial output, and the input tuned-circuit is represented symbolically by a transformation ratio n, stepping up the source-resistance R_a to n^2R_a in parallel with the impedance of the said tuned-circuit. The signal source is also the seat of spontaneous fluctuations. In series with the above resistance R_a two constant-voltage generators of zero internal impedance are assumed, one supplying the signal voltage V_a and the second the fluctuation voltage v_a. The latter corresponds to the small frequency band Δf centred round the signal frequency and may be assumed to be due to thermal electronic fluctuations in the resistance R_a at a temperature T_a, thus[1,14,17]: $\overline{v_a^2} = 4kT_aR_a\Delta f$ where k is Boltzmann's constant ($1.38.10^{-23}$ joule per degree Kelvin). Instead of the said constant-voltage generators it is simpler to make use of constant-current generators of infinite internal impedance supplying signal and signal-source fluctuation currents to the valve's input circuit.

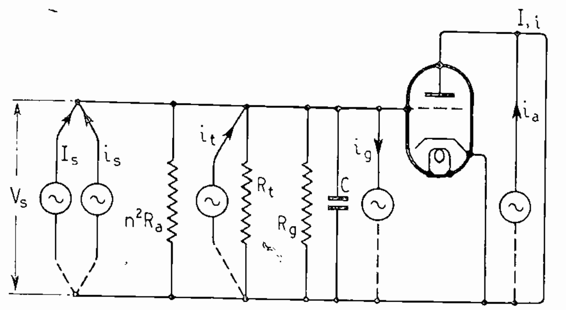

Fig. 2. *Input circuit connected to cathode and grid of a triode showing the transformed signal-source resistance n^2R_a, the tuned circuit's resonant resistance R_t and the valve's input resistance R_g due to electron and to inductive effects. The constant-current generators supply the signal current I_s, the source's fluctuation current i_s, the tuned circuit's fluctuation current i_t, the induced grid fluctuation current i_g and the shot-effect fluctuation current i_a. The signal current I and resulting fluctuation current i in the short-circuit anode lead are calculated.*

Taking into account the transformation ratio mentioned above, we obtain a current supplied by the constant-current signal generator:

$$I_s = \frac{nV_a}{n^2R_a} = \frac{V_a}{nR_a} \quad \ldots \quad \ldots \quad \ldots \quad (1)$$

* Due to Kappler, see *Annalen der Physik*, Vol. 2, 1931, p. 233.

The constant-current noise generator supplies the current

$$\overline{i_s^2} = \frac{\overline{v_a^2}}{n^2 R_a^2} = \frac{4kT_a R_a \Delta f}{n^2 R_a^2} = \frac{4kT_a \Delta f}{n^2 R_a} \qquad (2)$$

The complete input-circuit is shown in Fig. 2. Parallel to the transformed source-resistance n^2R_a is the resonant impedance R_t of the tuned circuit, the input resistance R_g between grid and cathode of the valve due exclusively to electronic causes or to lead inductances and a capacitance C indicating a detuning (if desired) of the total input circuit.

Two sources of noise, besides the one already mentioned, have still to be taken into account. The first one is the thermal noise of the resistance R_t at room temperature T. It may be represented by a constant-current generator supplying the fluctuation current

$$\overline{i_t^2} = \frac{4kT\Delta f}{R_t} \qquad \ldots \quad \ldots \quad \ldots \quad (3)$$

The second one is due to the induced fluctuation current at the grid[6,12,16], and will be indicated by i_g. It may be described as being due to thermal electronic fluctuations in a resistance R_g at a temperature T_g, thus[12]

$$\overline{i_g^2} = \frac{4kT_g\Delta f}{R_g} \qquad \ldots \quad \ldots \quad \ldots \quad (4)$$

Due to the said current generators a signal voltage V_s and a noise voltage v_s arise between grid and cathode at an angular frequency ω

$$\left.\begin{aligned} V_s &= \frac{I_s}{\dfrac{1}{n^2R_a} + \dfrac{1}{R_t} + \dfrac{1}{R_g} + j\omega C} \\ \overline{v_s^2} &= \overline{\left| \frac{i_s + i_t - i_g}{\dfrac{1}{n^2R_a} + \dfrac{1}{R_t} + \dfrac{1}{R_g} + j\omega C} \right|^2} \end{aligned}\right\} \qquad (5)$$

The vertical dashes in the latter expression indicate its modulus whereas the horizontal dash denotes averaging, taking due account of the non-correlation between i_s, i_t and i_g. What we are interested in are the signal and the noise currents in the short-circuit anode lead due to the above voltages. Denoting the transadmittance of the valve from grid to anode by S we obtain an anode signal current $I = SV_s$. As to the noise current i in the anode lead, this consists firstly of Sv_s and secondly, of the noise current i_a circulating in this lead if the grid is connected to the cathode across a low impedance while maintaining the previous d.c. operating conditions. Thus

$$\overline{i^2} = \overline{\left| \frac{S(i_s + i_t - i_g)}{\dfrac{1}{n^2R_a} + \dfrac{1}{R_t} + \dfrac{1}{R_g} + j\omega C} + i_a \right|^2} \qquad \ldots \quad \ldots \quad (6)$$

The noise ratio is obtained by division of Equ. (6) by $|SV_s|^2$, wherein V_s is given by Equs. (5) and (1).

3. Conditions for Minimum Noise Ratio

In studying the conditions for a minimum noise-to-signal ratio two outstanding facts have to be borne in mind. Firstly, the complete correlation of i_g and i_a in an ideal triode, both fluctuations being due to the same electron stream through the valve. Secondly, the phase-angles of the trans-admittance S and of i_g relative to i_a due to electron transit times. We may express these angles thus[18]

$$\left\{\begin{aligned} S &= |S|\, e^{-j\phi_o} \\ i_a &= |i_a|\, e^{-j\phi_a} \\ i_g &= j\,|i_g|\, e^{-j\phi_g} \end{aligned}\right. \qquad j = +\sqrt{-1} \qquad \ldots \quad \ldots \quad (7)$$

Only the relative phase-angle $\phi_g - \phi_a$ of the correlated currents i_g and i_a is relevant.

The modulus $|S|$ of the transadmittance does not vary appreciably from low to extreme-high frequencies[19]. At a frequency of 600 Mc/s, ϕ_0 may be 120° in a triode of the usual electrode-dimensions. The modulus of i_a also does not vary appreciably with frequency up to the said value in many cases. Finally, if the ratios R_t/R_g and R_t/n^2R_a are increased by increasing the quality-figure of the tuned circuit until they become large compared to unity, the expressions $1/R_t$ and i_t may be dropped out of Equ. (6). We shall assume that this condition is satisfied. Inserting Equ. (7) we thus obtain a noise ratio:

$$\left.\begin{aligned} \frac{\overline{i^2}}{I^2} &= \frac{\dfrac{\overline{v_a^2}}{R_a^2} + \overline{i_g^2}\dfrac{R_g}{R_a} w}{V_a^2/R_a^2} \\ w &= \overline{\left| 1 - \left|\frac{i_a}{i_g S R_g}\right| \frac{1}{j} e^{j\phi} \left(\frac{R_g}{n^2R_a} + 1 + j\omega C R_g\right) \right|^2} \times \frac{n^2R_a}{R_g} \\ \phi &= \phi_o + \phi_g - \phi_a \end{aligned}\right\} \qquad \ldots \quad \ldots \quad (8)$$

A minimum noise ratio obviously coincides with a minimum value* of w. Introducing the symbols

$$\left|\frac{i_a}{i_g S R_g}\right|^2 = a^2, \quad \frac{R_g}{n^2 R_a} = x, \quad \omega C R_g = y, \quad \ldots \quad \ldots \quad (9)$$

the expression for w becomes

$$w = 1/x \times [1 - a\{(x+1)\sin\phi + y\cos\phi\}]^2 + a^2/x \times \{(x+1)\cos\phi - y\sin\phi\}^2 \quad \ldots \quad \ldots \quad (10)$$

The values of x (i.e., the coupling ratio n) and of y (i.e., the detuning) may now be chosen so as to minimise w. Varying y a minimum is reached if

$$ay = \cos\phi, \quad \ldots \quad \ldots \quad \ldots \quad (11)$$

this minimum value being

$$\mathrm{Min}_y\,(w) = \frac{\{a(x+1) - \sin\phi\}^2}{x}$$

Varying x, this expression attains a minimum value, if :

$$a^2x^2 = (a - \sin\phi)^2 \quad \ldots \quad \ldots \quad (12)$$

The condition (11) fixes the detuning capacitance. This may be positive if $\cos\phi$ is positive and negative (i.e., inductive detuning) if $\cos\phi$ is negative. As to the latter condition (12) we shall assume a unequal to $\sin\phi$. For if these values were equal we would obtain $x = 0$ and hence no signal would result at the valve input ($I_s = 0$). Two cases corresponding to Equ. (12) have still to be considered : firstly $(a - \sin\phi) > 0$ and, secondly, $(a - \sin\phi) < 0$. In the first case, the minimum value of w is

$$\mathrm{Min}_{xy}\,(w) = 4a\,(a - \sin\phi) \quad \ldots \quad (13)$$

and hence is positive, while in the second case it is zero. Theoretically this second case is extremely interesting, as the contributions of the valve would then completely disappear from the noise ratio. This would hence be equal to the noise ratio at the terminals of the signal source (e.g., receiving antenna). From a practical point of view it is not so important if w disappears completely. The only important condition is that $\overline{i_g{}^2}w$ should be small compared to $\overline{i_s{}^2}$ in Equ. (8). In words :—the contribution of the valve to the noise ratio should be negligible.

We shall now deal with the case when no detuning is applied. Hence $y = 0$ and the minimum condition for w becomes

$$a^2x^2 = 1 - 2a\sin\phi + a^2 \quad \ldots \quad (14)$$

* See Appendix for further definition of w.

the minimum value of w being

$$w_{\min} = 2a(\sqrt{1 - 2a\sin\phi + a^2} + a - \sin\phi) \quad \ldots \quad \ldots \quad (15)$$

When $\sin\phi = 1$ and $a \leqslant 1$ this minimum value becomes zero. Comparing Equs. (14) and (12) it is seen that x^2 is larger and hence n^2 smaller in the former case. Thus, optimal coupling is weaker if detuning is applied, and especially so, if $\sin\phi$ is small.

4. Discussion and Experimental Evidence

The reduction and eventually more or less complete compensation of the valve's contribution to the noise ratio is due to a fluctuation voltage component, completely correlated to i_g, arising across the input circuit. By proper detuning this voltage causes an anode fluctuation current component, due to the valve's transadmittance, which is counterphase (phase-difference of π radians) with respect to i_a. Hence the correlation

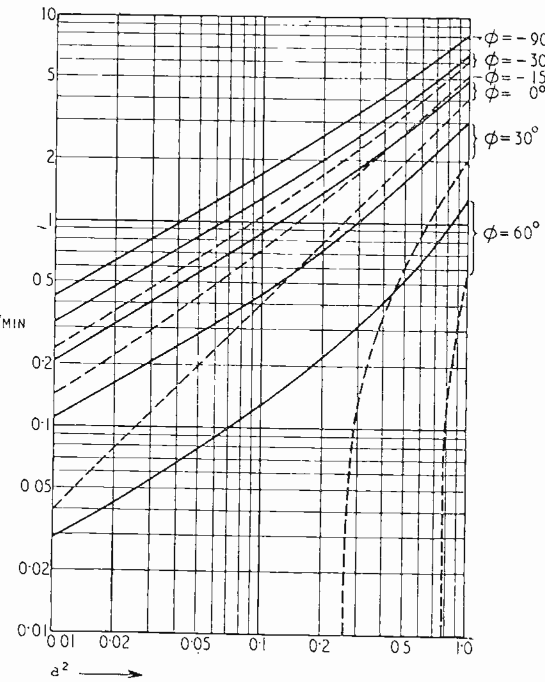

Fig. 3. *The minimum value of the expression w (vertical scale) of Equs. (8) and (10) varying x and y of Equ. (9) as dependent on a^2 of Equs. (9) and on ϕ of Equ. (8). Full curves without, and dotted curves with, optimal detuning.*

of i_g to i_a is essential. In an ideal plane triode this correlation exists to a fairly complete degree.

With the aid of the so-called equivalent noise-resistance R_n of the valve [6], defined by :

$$\overline{i_a{}^2} = 4kT\,|S^2|\,R_n\Delta f, \quad \ldots \quad \ldots \quad (16)$$

T being again the room temperature (e.g., 293 degrees Kelvin), the expression (9) for a^2, using Equ. (4) may be written as :

$$a^2 = \frac{R_n}{R_g} \quad \frac{T}{T_g} \qquad \ldots \quad \ldots \quad \ldots \quad (17)$$

By applying Equs. (13) and (15) the minimum value of w may be calculated as dependent on a^2 for different values of ϕ with or without detuning. This is shown in Fig. 3. It appears from these curves that detuning has relatively little effect at negative values of ϕ. As the noise ratio of Equ. (8) may be written as[18]

$$\frac{\overline{i^2}}{I^2} = \frac{kT\Delta f\,(T_a/T + wT_g/T)}{V_a^2/4R_a} \qquad \ldots \quad (18)$$

by using Equs. (2) and (4), this Fig. 3 is directly suited to the discussion of the relative importance of the two contributions to this ratio, corresponding to the two terms between brackets in the numerator of (18). The ratio of T_a/T is often larger than 1, e.g. of the order of 2 or more, whilst T_g/T is usually, with oxide-coated cathodes, of the order of 5. Assuming 2 and 5 for these ratios, we would obtain the result that w should be below, say, 0.2 in order to make a further decrease of w relatively uninteresting. From Fig. 3 we may see, if ϕ and a^2 are given, whether this may be realized. As an example, we assume a^2 may be 0.5 at 1 metre wavelength. Obviously, by Fig. 3, w cannot attain values below 0.2 at this value of a^2 if ϕ is zero or negative with or without detuning. If, however, ϕ could be, say, 60° positive, such values could easily be obtained.

If a tetrode or a pentode is used instead of a triode, the fluctuation current i_a consists of two parts, one of which is completely correlated to i_g, the other being, however completely uncorrelated to i_g[6, 16]. Hence, the reduction of the noise ratio is mainly confined to the first part of i_a in this case. Denoting the said decomposition of i_a by

$$\overline{i_a^2} = \overline{i_{a_1}^2} + \overline{i_{a_2}^2} \qquad \ldots \quad \ldots \quad \ldots \quad (19)$$

the reduction of w obtainable under ideal conditions, could roughly not be better than to $\overline{i^2_{a2}}/\overline{i_a^2}$ multiplied by the original value of w.

The noise ratio of a particular valve, a pentode, used in the first stage of an amplifier[8], was measured at 300 Mc/s. The measured figures represent the expression in brackets of Equ. (18) with $T_a = T$ (see Fig. 4, taken from a paper by W. Kleen[8]). It is seen that a reduction of this figure from about 21 to 17 is obtained by detuning, leaving, however, the coupling unaltered. In the valve under consideration $\overline{i^2_{a_1}}$ accounts for about 40 per cent and $\overline{i^2_{a_2}}$ for about 60 per cent of $\overline{i^2_a}$. Hence the reduction should be less than from the original 21 to about 13. Due to non-optimal coupling at the detuned position and to the figures of Fig. 4 including some additional components besides wT_g/T, this optimal reduction is not obtained. This experiment tends to show, however, that worth-while reductions may actually be effected using suitable valves and circuits. We may also conclude from Figs. 3 and 4 that ϕ is positive in the present case.

5. The Grounded-grid Amplifier Stages

In the above discussion the cathode was assumed to constitute the electrode common to input and output. In recent years, however, amplifier circuits have been discussed, in which the control grid is the common electrode[2, 7, 9]. The possibilities of reducing the noise ratio in amplifiers using a grounded-grid valve will now be discussed.

The input circuit is exactly the same as that of Fig. 2. Instead of connecting the anode to the cathode by a short-circuit lead it should now be thus connected to the grid (of course, always taking due care of direct current connections). Assuming again a triode, the fluctuation current in the said anode-grid lead, if the input is short-circuited, is again i_a. But the fluctuation current i_g to the grid must now be replaced[16] by $i_a + i_g = i_c$. This current i_c is completely correlated to i_g in an ideal triode, as is i_a. With these changes we obtain instead of Equ. (8) the noise ratio

$$\left.\begin{aligned} \frac{\overline{i^2}}{I^2} &= \frac{\dfrac{\overline{v_a^2}}{R_a^2} + \overline{i_c^2}\,\dfrac{R_g}{R_a}\,u}{V_a^2/R_a^2} \\ u &= \overline{\left|1 - \left|\frac{i_a}{i_c S R_g}\right| \frac{e^{j\psi}}{j}\left(\frac{R_g}{n^2 R_a} + 1 + j\omega C R_g\right)\right|^2} \times \frac{n^2 R_a}{R_g} \\ \psi &= \phi_0 - \phi_a - \pi/2 \qquad i_c = |i_c| \end{aligned}\right\} \qquad \ldots \quad \ldots \quad (20)$$

The value of R_g in this case is different from the grounded-cathode case. At r.f.,

R_g is approximately equal to $1/S$ in the present case.

This ratio attains a minimum value if u is as small as possible. Using the notations :

$$\left.\begin{aligned} \overline{i_c^2} &= 4kT\,|S^2|\,R_c\Delta f, \\ b^2 &= \left|\frac{i_a}{i_c S R_g}\right|^2 = \frac{R_n}{|S^2| R_g^2 R_c} \\ x &= \frac{R_g}{n^2 R_a} \quad y = \omega C R_g \end{aligned}\right\} \quad .. \quad (21)$$

the expression for u becomes :

$$u = 1/x \times [1 - b\{(x+1)\sin\psi + y\cos\psi\}]^2 + \\ + b^2/x \times \{(x+1)\cos\psi - y\sin\psi\}^2 \quad .. \quad .. \quad (22)$$

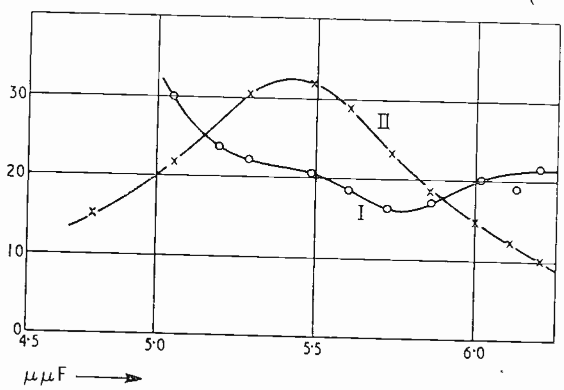

Fig. 4. *Experimental curve representing the expression between brackets in the numerator of Equ. (18) with $T_a = T$ (curve I, vertical scale) as dependent on the input circuit's tuning capacitance (horizontal scale) and at constant coupling, optimal for noise ratio at the tuned position. Curve II represents the input signal voltage squared (vertical linear scale) as dependent on tuning capacitance, and hence affords a check on the tuning position.*

Comparing this Equ. (22) with Equ. (10) we see that the procedure used in dealing with the latter may be repeated in the present case. Hence the resulting minimum conditions and values of u may be taken from Equs. (11) to (15), replacing w by u, a by b and ϕ by ψ. The curves of Fig. 3 are also valid in the present case, replacing a by b and ϕ by ψ. Going by the experimental evidence of ϕ being positive and small in the grounded-cathode case corresponding to Fig. 4, we might conjecture ψ to be near $-\pi/2$ in the present case at not too high frequencies. This would mean nearly zero optimal detuning by the equivalent of Equ. (11). A considerable reduction of the noise ratio may be expected also in the present case by the proposed coupling and (if necessary) detuning.

6. Velocity Modulation Valves

We consider a velocity-modulation valve[15] with two adjacent input electrodes and two similar output electrodes, both pairs being passed by the electron stream. The spontaneous fluctuations in density of this stream at a particular point of its path may be due to initial density fluctuations at a previous point or to spontaneous velocity fluctuations at such point, transferring to density fluctuations along the path. With the usual operational data it may be shown that the latter effect is in most cases unimportant.

The initial density fluctuations will decrease along the path due to velocity-distribution of electrons in the stream. This effect, too, may be shown to be usually negligible. Hence in dealing with spontaneous fluctuations we may confine our attention to the effects caused by initial density fluctuations of the stream. The ensuing fluctuation voltages and currents are completely correlated. The output-electrodes are assumed to be connected by a short-circuit lead. We shall evaluate the fluctuation current as well as the signal current in this lead and thus obtain the noise ratio.

The input-circuit coincides with Fig. 2, replacing i_g by i_f, denoting the induced fluctuation-current due to the spontaneous fluctuations of the electron stream passing the input electrodes. The transadmittance from input to output will again be indicated by the complex symbol S. The distance between the two adjacent input electrodes of the velocity-modulation valve as well as between the two adjacent output electrodes will be assumed so short that the product of angular frequency ω and electron transit time between the electrodes of each pair is small compared to unity. With this assumption a simple relation exists between i_f and the spontaneous fluctuations of the electron stream. If I_d denotes the average direct current corresponding to the stream and e the electron charge, we have :

$$\overline{i_f^2} = 2eI_d F^2 \Delta f \quad .. \quad .. \quad .. \quad (23)$$

The multiplier F^2 is unity if the stream issues from a cathode operating under saturated conditions and may be smaller than unity if I_d is less than the saturation current of the cathode. Equs. (5) and (6) apply exactly to the present case if i_g is replaced by i_f. The signal current in the short-circuit output lead is $I = SV_s$ and the fluctuation current is :

$$\overline{i^2} = \left| \frac{S(i_s + i_t - i_f)}{\frac{1}{n^2R_a} + \frac{1}{R_t} + \frac{1}{R_g} + j\omega C} + i_f^{-j\omega t_0} \right|^2 \quad \ldots \quad \ldots \quad (24)$$

t_0 indicating the electron transit time along the beam from the input to the output pair of electrodes. Thus Equs. (7) have to be replaced by :

$$\left.\begin{aligned} S &= |S|\, e^{-j\phi_0} \\ i_a &= |i_f|\, e^{-j\phi_a} \qquad \phi_a = \omega t_0 \\ i_f &= |i_f| = j\,|i_f|\, e^{-j\phi_f} \quad \phi_f = \pi/2 \end{aligned}\right\} \quad (25)$$

Inserting this notation into the noise ratio obtained by division of (24) by $|SV_s|^2$ we obtain instead of Equs. (8)

$$\left.\begin{aligned} \frac{\overline{i^2}}{I^2} &= \frac{\frac{\overline{v_a^2}}{R_a^2} + \overline{i_f^2}\frac{R_g}{R_a}p}{V_a^2/R_a^2} \\ p &= \left|1 - \left|\frac{1}{SR_g}\right| \frac{e^{j\phi_1}}{j}\left(\frac{R_g}{n^2R_a} + 1 + j\omega CR_g\right)\right|^2 \times \frac{n^2R_a}{R_g} \\ \phi_1 &= \phi_0 + \phi_f - \phi_a = \phi_0 + \pi/2 - \omega t_0 \end{aligned}\right\} \quad \ldots \quad \ldots \quad (26)$$

Thus a similar expression to w of Equ. (8) arises in the present case, as denoted by p above. Introducing the abbreviations

$$c = \left|\frac{1}{SR_g}\right| \qquad x = \frac{R_g}{n^2R_a} \qquad y = \omega CR_g \quad \ldots \quad \ldots \quad (27)$$

an expression for p, very similar to Equ. (10), is obtained :

$$p = 1/x \times [1 - c\{(x+1)\sin\phi_1 + y\cos\phi_1\}]^2 + \\ + c^2/x \times \{(x+1)\cos\phi_1 - y\sin\phi_1\}^2 \quad \ldots \quad \ldots \quad (28)$$

The discussion of this expression in order to calculate its minimum value with x and y as variables is obviously similar to that of Equ. (10) and the results of the latter discussion may be directly applied to the present case. Thus by Equ. (11), the optimal detuning is

$$cy = \cos\phi_1$$

or

$$\frac{\omega C}{|S|} = -\sin(\phi_0 - \omega t_0) \quad \ldots \quad \ldots \quad (29)$$

whilst optimal coupling becomes by Equ. (12)

$$c^2x^2 = (c - \sin\phi_1)^2$$

or

$$\left|\frac{1}{n^4S^2R_a^2}\right| = \left\{\frac{1}{|SR_g|} - \cos(\phi_0 - \omega t_0)\right\}^2 \quad \ldots \quad \ldots \quad (30)$$

Now, by the theory of velocity-modulation valves [11, 15]

$$S = \tfrac{1}{2}\, j\, \frac{I_d}{V_d}\,\omega t_0\, e^{-j\omega t_0} = \frac{I_d}{2V_d}\,\omega t_0\, e^{-j(\omega t_0 - \pi/2)} \quad \ldots \quad \ldots \quad (31)$$

wherein V_d denotes the steady voltage corresponding to the average electron speed along the path between input and output. Hence $\phi_1 = \phi_0 + \pi/2 - \omega t_0$
$= \omega t_0 - \pi/2 + \pi/2 - \omega t_0 = 0$,
and $c - \sin\phi_1$ is thus positive. This makes a complete annihilation of p impossible, its minimum value corresponding to the conditions (29) and (30) being

$$\text{Min}\,(p) = 4c\,(c - \sin\phi_1) = 4c^2 = \frac{4}{|SR_g|^2} \quad \ldots \quad \ldots \quad (32)$$

We have not yet discussed the value of R_g to be inserted into the above equation. By the condition imposed above on the electron transit time t_1, between the two input electrodes, we obtain :

$$\frac{1}{R_g} = \frac{I_d}{V_d}\frac{(\omega t_1)^2}{24} \quad \ldots \quad \ldots \quad \ldots \quad (33)$$

Hence $4c^2 = \dfrac{4}{|SR_g|^2} = \dfrac{(\omega t_1)^4}{36\,(\omega t_0)^2}$

and this value, by the assumption that $\omega t_1 \ll 1$, is surely much smaller than unity. The further discussion may be based on Fig. 3, replacing a^2 by c^2, ϕ by $\phi_1 = 0$ and w by p.

The equivalent of Equ. (18) is in the present case [11]

$$\frac{\overline{i^2}}{I^2} = \frac{kT\Delta f\left(\frac{T_a}{T} + \frac{2eI_dF^2}{kTR_g|S|^2}\right)}{V^2/4R_a} \quad \ldots \quad (34)$$

The expression between brackets may be regarded as a suitable measure for the relative noise at the output (see Appendix). Whereas this expression is of the order of 20 at 300 Mc/s for a pentode valve according to Fig. 4, it is much higher in the present case. Assuming $|S| = 0.12$ mA/V, $R_a = R_g = 20$ kilo-ohms, $F^2 = 1$ and $T_a = T$ (room temperature) we obtain [11] about 2 500. Thus we may conclude that velocity-modulation valves are not well suited for reception purposes under these conditions. The large noise ratio is obviously due to i_f.

It is possible, however, to devise special measures, based on the same principles, which could bring the above figure down to, say, below 100. One of these devices consists of an extra pair of electrodes preceding the input electrodes, as viewed in the direction of the electron stream. By connecting a suitable impedance between these electrodes a fluctuation voltage may be generated across the said impedance, causing a fluctuation current in the output lead, and entailing a reduction of the original fluctuation current and hence of the noise ratio. To achieve this the distances between the three pairs of electrodes thus needed, as well as the impedances at the first two pairs, have to be chosen carefully.

7. Valve-construction and Stage Design

The reduction of the noise ratio as set forth in Sections 3 to 6 is hampered if the individual electron paths and motions are unequal. This may be easily seen by going through the argument of Section 3, assuming the valve to consist of two parallel sections with different values of R_g, S, ϕ_a and ϕ_g. Obviously, one minimum value of w will be obtained corresponding to each section and the corresponding minimum conditions as to detuning and coupling will in general not coincide. Seeking a most favourable compromise, we shall still be left with a higher resulting value of w than would correspond to the minimum value for each of the sections. Using Fig. 3, a decision may be reached as to what degree of inequality is still tolerable in this respect without undue rise of the noise ratio. Valve construction should comply with these requirements.

It is assumed in Sections 3 to 6 that the resonant resistance R_t of the tuned circuit may be raised to such a level that the corresponding spontaneous fluctuations of input voltage are relatively unimportant. If a resistance exists between the input electrodes which is not due to electronic causes or to lead inductances, but to losses such as dielectric losses or series resistance of electrode leads, this desirable increase of R_t will be limited. In fact, this resonant resistance, including the said losses, must always remain smaller than the said shunt loss-resistance. Additional non-correlated fluctuations will arise from this cause and entail a rise of noise ratio. This may be accounted for by the addition of a term n^2R_a/R_t to the expression between brackets in the numerator of Equ. (18) and of a term R_g/R_t in the round brackets of w [Equ. (8)]. It is thus possible to judge the value of R_t required to make the contribution of these terms to the over-all noise figure negligible. R_t should be large compared with R_g and this is hard to achieve with velocity-modulation valves.

Strictly, the reduction of noise ratio as proposed above applies only to a frequency band Δf which is very narrow compared to its corresponding average frequency. In what way do our results apply to wide-band reception ? An answer to this question under practical conditions may be obtained from Fig. 4. The overall resonant resistance of the input-circuit was less than a thousand ohms at 300 Mc/s. The total tuning capacitance being of the order of 12 $\mu\mu$F, the frequency-interval corresponding to 0.1 $\mu\mu$F detuning capacitance is about 1.25 Mc/s. Considering the flatness of the curve II of Fig. 4 corresponding to 0.1 $\mu\mu$F detuning capacitance, the proposed reduction of noise ratio obviously also applies to wide-band reception.

By optimal coupling and (if necessary) detuning of the input circuit the gain of the amplifier stage will become less than the optimal gain obtainable with this stage under favourable conditions with respect to gain. Though a high gain is not so important in itself, a low gain figure may cause an increase of overall noise ratio of the receiver as the noise introduced by the second stage could make an appreciable contribution to this ratio. Now, by a suitable feed-back from the output to the input of a stage its noise ratio is only altered to a minor and often negligible degree[6]. But stage gain may be considerably increased by such feed-back. Hence it may afford a suitable means of compensation for the loss of gain incurred by the coupling and detuning corresponding to a minimum noise ratio. Care should, of course, be taken that no appreciable additional contributions to the overall noise ratio are introduced by the feed-back circuit itself. Preferably, this should not incorporate unsuitable resistance elements.

In the above discussion only amplifier stages were considered. Similar means of noise-reduction may, however, be also applied to *mixer* stages[6], taking due account of noise correlation. If the first stage of a receiver contains a triode with the input signal as well as the local oscillator voltage acting between cathode and grid, reasoning similar to that of Sections 2 to 5 may be applied. The transadmittance S has to be replaced by the *conversion* transadmittance S_c, this being in general complex at the present frequencies, as is S. The value of R_g, due to the action of the local oscillator voltage, will generally be *higher* than with amplifier operation. By the application of the coupling and detuning similar to those of Sections 3, 5 and 6 mixer stages of low noise ratio may be obtained, particularly suitable for the input of receivers at the present frequencies. A complete discussion of such mixer stages will be published elsewhere.

It should be pointed out that the application of A.V.C. by bias regulation might upset the results of the detuning and input-coupling under discussion. By the application of proper delay or of different methods of A.V.C., specially devised for cases like the present one, an increase of noise ratio as well as of selectivity due to A.V.C. may be avoided. It is hoped to deal with these points more fully in a future article.

APPENDIX

Noise Ratio and Noise Figures

It may be well to compare briefly the noise ratio, used in the present article, with some of the different definitions of noise figures, proposed in previous publications. The first of such figures[3,4,8] refers to the ratio of the mean-square noise voltage to the signal-voltage squared at the input electrodes of the first valve. The noise introduced by the valve itself, including its shot noise in the anode-circuit, was transferred to this input circuit by the well-known artifice of introducing a corresponding noise voltage generator of zero internal impedance in series with one of the valve's input terminals. The valve itself could then be assumed to be free of noise. The signal voltage is supplied by a signal generator of voltage V_a and of zero internal impedance in series with an ohmic resistance R_a. The available power of this signal source, i.e., the maximum power to be drawn from it for dissipation in a further circuit connected to its terminals, is $V_a^2/4R_a$. Now by the definition mentioned the available signal power, necessary to make the said ratio of mean-square noise voltage to signal-voltage squared at the valve's input terminals equal to unity, is used as a measure for the relative noise of the stage. Dividing this power by $kT\Delta f$ (T being the room temperature) the *noise-figure* N is obtained.

We shall show that our noise ratio (8) or (18) corresponds very closely to the said noise figure. In fact, the ratio of mean-square noise voltage to the signal voltage squared at the input terminals is *exactly* equal to the noise ratio (8) or (18). Hence, if this ratio is assumed to be unity, the expression between brackets of Equ. (18) coincides with the above noise figure N.

In a more recent definition[5,13] the available noise power at the output terminals of a stage is divided by the available signal power at these terminals to obtain the noise ratio. The noise figure F of a stage is then said to be the ratio of its output noise ratio to the output noise ratio of the preceding stage. By the method of short-circuiting the output terminals our noise ratio coincides exactly with the ratio of available noise power to available signal power at the output. Hence, using the notation of Equ. (18) the noise figure F is given by:

$$F = \frac{kT\Delta f\left(\frac{T_a}{T} + \frac{T_g}{T}w\right)}{V_a^2/4R_a} \frac{V_a^2/4R_a}{kT_a\Delta f}$$

$$= 1 + \frac{T_g}{T_a}w = \frac{T}{T_a}N = N,$$

if $T_a = T$, as was assumed in the original definition of F.

Thus the quantity w which we have extensively used in the preceding sections is simply:

$$w = (F - 1)\frac{T_a}{T_g}$$

While the noise figure N corresponding to two subsequent stages of suffixes a and b, noise figures N_a and N_b and gain figures g_a and g_b is:

$$N = N_a + \frac{N_b - T_a/T}{g_a}$$

the corresponding relation in the F figures is:

$$F = F_a + \frac{F_b - 1}{g_a}$$

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TRANSIENT RESPONSE OF TUNED-CIRCUIT CASCADES*

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SUMMARY.—An analysis is given of the transient response of two, of four and then of any number N tuned circuits, all with the same resonance frequency, connected in cascade with a forward transmission coupling but no mutual coupling. It is shown that the envelope of the build-up or decay of a pulse applied at the resonance frequency is determined by the usual negative exponential term ($\epsilon^{-\beta t}$) applying to a single circuit, with a factor consisting of the first N terms of the series expansion of the corresponding positive exponential. The envelope of the transient component of the output signal when the applied frequency is not very close to resonance is given by $\epsilon^{-\beta t}$ multiplied by the Nth term of the same series expansion of $\epsilon^{\beta t}$. A comparison of the responses of tuned-circuit cascades with those of underived band-pass filters shows that for equal component qualities in the two cases, two tuned circuits in cascade are approximately equivalent, for pulse transmission applications, to a single-section band-pass filter using 50 per cent more components.

LIST OF CONTENTS

1. Introduction

THE use of a series of tuned circuits coupled together by mutual inductance, mutual capacitance, or resistance is well known, and the response of such an arrangement is analysed in various text-books and papers. The use of a cascade of tuned circuits which have transmission coupling but no mutual coupling is not so often referred to, however, although such an arrangement occurs frequently in practice. It is the purpose of the first parts of this paper to deduce the transient response of such a cascade, and it will be seen that very simple and elegant expressions are obtained. In spite of their simplicity, however, two independent searches of the literature have failed to reveal any previous publication of them.

It is also fairly well known that coupled tuned circuits are equivalent to certain types of conventional "3-element" band-pass filter[1]; although the latter are generally designed in accordance with Zobel's theory[2], it is nevertheless a fact that improved filtration can be obtained from single sections by a more empirical or *ad hoc* design such as is generally adopted for coupled tuned circuits.

The relationship between a cascade of tuned circuits without mutual coupling and a conventional band-pass filter is, however, of a different nature, and has probably not been investigated before. There is no direct circuit equivalence, but it is shown in the last part of this paper that there is a relationship in transient response, and that for the same Q-value of the resonant elements the overall performance of two tuned circuits in cascade is almost equivalent to that of an underived band-pass filter whole section, on a balance between transient response and steady-state discrimination. As the filter has three inductors and three capacitors, it is evident that the two tuned-circuit cascade effects an economy of components.

The same methods as are applied in this paper for determining the transient response of tuned-circuit cascades, and comparing them with band-pass filters, may be applied for determining the transient response of cascades of RC circuits and comparing them with low-pass filters. It is necessary only to replace the envelope functions by actual current or voltage functions when a d.c. signal is applied, and to regard a low-pass filter as equivalent to a band-pass with bandwidth equal to twice the low-pass cut-off

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frequency, and with midband frequency equal to zero.

Although the author's point of view is that of multi-channel pulse transmission systems in which the pulse is used to operate a relay* (e.g., a voice-frequency telegraph system or a multi-channel telephone trunk signalling system), an obvious application of the work is to multi-stage amplifiers with LC or RC interstage couplings. Such amplifiers are used in television with a.c. or d.c. pulses, but it must be pointed out that in the case of the LC circuits considered here, the Q-value is assumed high, so that the circuits are not suitable for d.c. pulses, being intended for a.c. pulses. Published treatments[10, 11] of cascaded stages in television video-frequency amplifiers (generally best treated by empirical or numerical methods) are listed in Section 7.

LIST OF SYMBOLS

R, L and C = the parallel components of a shunt-tuned circuit.

$\beta = \frac{1}{2RC} = \frac{\omega_0}{2Q}$ where ω_0 = resonance angular frequency, and $Q = \omega_0 RC = \frac{R}{\omega_0 L}$ for the parallel circuit used. β_1 and β_2 are also used where two tuned circuits of unequal Q are concerned.

ω = any applied frequency.

$x = \frac{\omega}{\omega_0}$

n = fractional bandwidth of filter $= \frac{\text{bandwidth}}{\text{mid-band frequency}}$.

$F_N(t)$ = output amplitude-time function of N tuned circuits in cascade.

N = number of tuned circuits in cascade.

2. Properties of a Single Tuned Circuit

A comprehensive account of the response, both steady-state and transient, of a single tuned circuit has been given by the author in another paper[4], although the subject is also dealt with in various text-books[5, 6]. Since the analysis of tuned-circuit cascades depends on the response of a single circuit, a summary is given here of the more important results. The tuned circuit considered is a parallel combination of R, L and C (as shown in Fig. 1) fed from a constant-current source, such as a pentode valve.

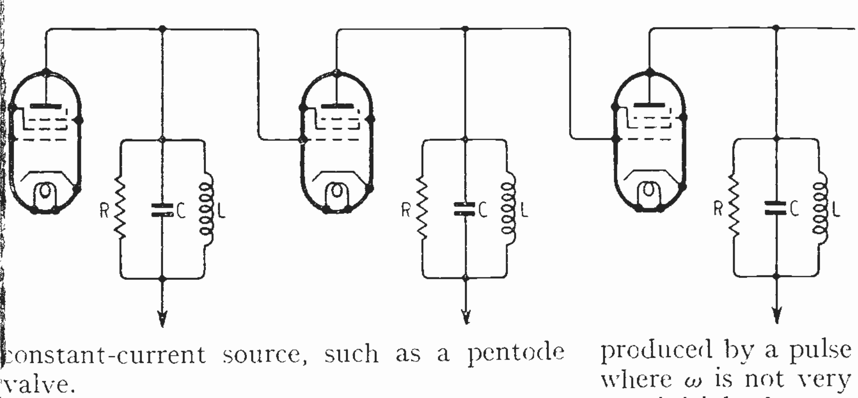


Fig. 1. *Cascade of tuned circuits without mutual coupling, and with blocking capacitors or coupling windings omitted for simplicity.*

We use $\beta = \frac{1}{2RC} = \frac{\omega_0}{2Q}$, and $Q = \frac{R}{\omega_0 L}$;

we assume $Q \gg 1$; $\omega_0 = \frac{1}{\sqrt{LC}}$

ω = any applied angular frequency;

$x = \omega/\omega_0$

$\sin(\omega t + \phi)$ = suddenly applied signal,

ϕ = initial phase angle.

(a) Steady-state voltage attenuation ratio (relative to resonance frequency)

$$= Q(1/x - x) \qquad \ldots \quad \ldots \quad \ldots \quad (1)$$

provided x is sufficiently different from unity to give

$$Q^2 (1 - x^2)^2 \gg 1.$$

(b) Build-up of a pulse* applied at the resonance frequency

$$= (1 - \epsilon^{-\beta t}) \sin(\omega_0 t + \phi) \qquad \ldots \quad (2)$$

The corresponding decay curve is, of course, $\epsilon^{-\beta t} \sin(\omega_0 t + \phi)$.

The steady-state amplitude is assumed to be unity.

(c) The transient produced by a pulse applied at frequency ω, where ω is not very close to ω_0, depends on the initial phase angle, ϕ, of the applied signal at the moment of switching.

For an initial phase angle of zero, the transient has a peak value equal to that of the steady-state output, so that the combined output signal is (relative to unity peak at resonance)

$$\frac{\cos \omega t - \epsilon^{-\beta t} \cos \omega_0 t}{Q(1/x - x)} \qquad \ldots \quad \ldots \quad (3)$$

* Hence the emphasis placed on the "half-amplitude" values in later parts of the paper; for a fuller discussion see Reference 3.

* i.e., the response to a Heaviside Unit Step envelope.

Thus the output $F_2(t) = \int_0^t \epsilon^{-\beta_1 u} \beta_2 \epsilon^{-\beta_2(t-u)} \,.\, du$

$$= \frac{\beta_2}{\beta_2 - \beta_1} (\epsilon^{-\beta_1 t} - \epsilon^{-\beta_2 t}) \quad \ldots \quad \ldots \quad (10)$$

For the case of equal Q-values, i.e. when $\beta_1 = \beta_2$, this is again indeterminate, and for this special case we obtain the solution

$$F_2(t) = \beta t \, \epsilon^{-\beta t} \quad \ldots \quad \ldots \quad \ldots \quad (11)$$

For the special case of one Q-value being twice the other, we obtain

$$F_2(t) = \epsilon^{-\beta_2 t} (1 - \epsilon^{-\beta_2 t}) \text{ when } \beta_1 = 2\beta_2$$

$$\text{and } F_2(t) = 2\epsilon^{-\beta_1 t} (1 - \epsilon^{-\beta_1 t}) \text{ when } \beta_1 = \beta_2/2 \quad \ldots \quad \ldots \quad (12)$$

These time functions are illustrated in Fig. 3, where curve 1 shows the output transient from the first circuit, curve 2 shows the output transient from two circuits when the first circuit has the higher Q and $\beta_1 = \beta_2/2$, and curve 3 shows the output from two circuits of equal Q. The peak values of the transients are 0.5 and 0.368 for the curves 2 and 3 respectively. The transient peak occurs at a time $2/3\beta_1$ seconds from switching in the case of unequal Q-values, and at a time $1/\beta$ seconds in the case of equal Q-values.

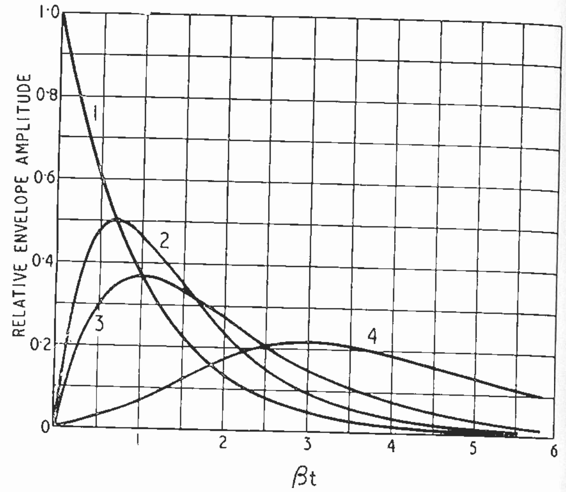
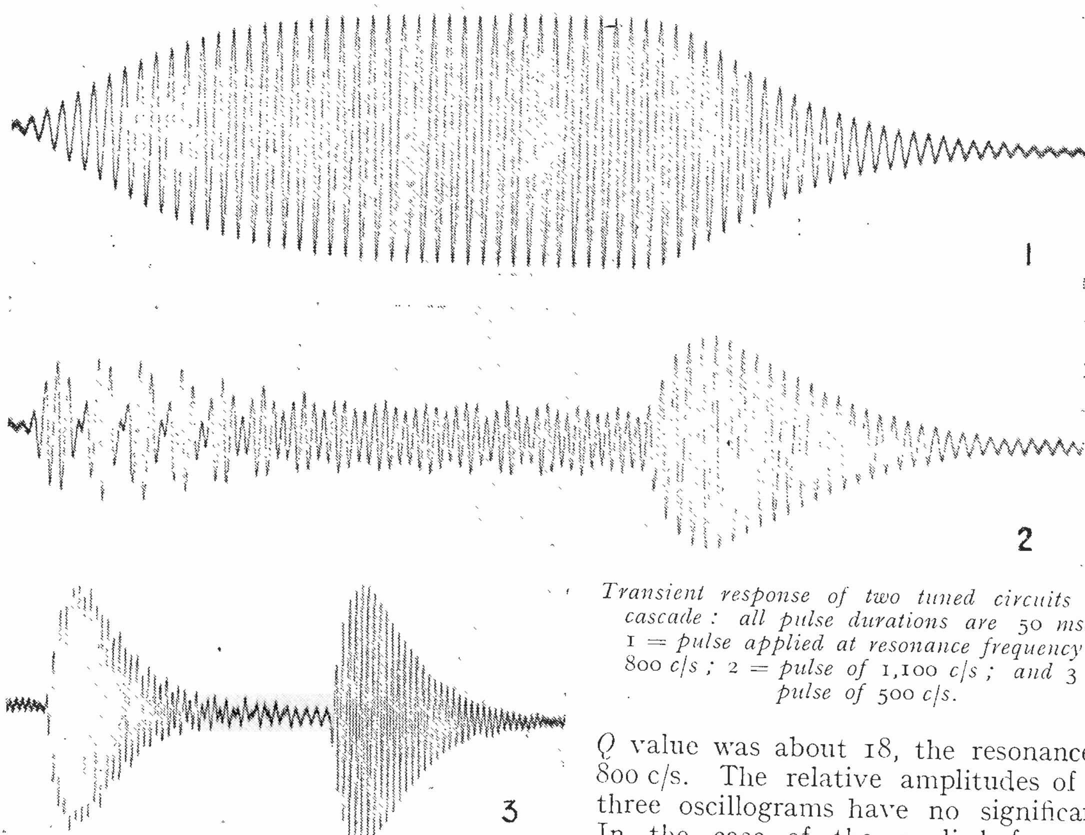


Fig. 3. *Transient envelopes for applied pulse frequency not close to resonance frequency. Curve 1 = one tuned circuit; curve 2 = two tuned circuits, $Q_1 = 2Q_2$; curve 3 = two tuned circuits, equal Qs; curve 4 = four tuned circuits, equal Qs.*

Oscillograms 2 and 3 show observed waveforms corresponding to the case of equal Q-values. As for Oscillogram 1 the Q value was about 18, the resonance at 800 c/s. The relative amplitudes of the three oscillograms have no significance. In the case of the applied frequency



Transient response of two tuned circuits in cascade: all pulse durations are 50 msec. 1 = pulse applied at resonance frequency of 800 c/s; 2 = pulse of 1,100 c/s; and 3 = pulse of 500 c/s.

= 1,100 c/s, the steady-state output is an appreciable proportion of the peak transient output. Apart from the beat produced on the make (due to the steady and transient components being of different frequency) there is the addition of a new transient to be considered, produced by the unit-step-modulated 1,100 c/s being applied to the second circuit. This transient is not in phase with the larger one, owing to the phase-shift at 1,100 c/s through the tuned circuit. If, as in the oscillograms, the initial phase is not controlled, there is a variation of peak transient amplitude due to the varying initial phase, and this is evident in Oscillogram 2. In Oscillogram 3 the steady-state and additional transients are practically negligible.

4. Any Number of Tuned Circuits in Cascade

Although, in practice, large numbers of identical tuned circuits are not likely to be connected in cascade, nevertheless the case is well worth examining because of the simplicity of the relationships. All Q-values are assumed equal.

Consider, to begin with, four circuits in cascade. We can regard this as two sets of two circuits, so that for Duhamel's Integral (Equ. 6) we have :—

$E(t) = 1 - \epsilon^{-\beta t}(1 + \beta t)$ for pulse at resonance frequency

and $E(t) = \beta t\,\epsilon^{-\beta t}$ for pulse not at resonance frequency

$$\phi(t) = 1 - \epsilon^{-\beta t}(1 + \beta t)$$

so that

$$\frac{d}{d(t-u)}\phi(t-u) = \beta^2 t\,\epsilon^{-\beta t+\beta u} - \beta^2 u\,\epsilon^{-\beta t+\beta u}$$

Calling the output time function $F_4(t)$, we have, for the first case,

$$F_4(t) = \int_0^t [1 - \epsilon^{-\beta u}(1 + \beta u)] [\beta^2 t\,\epsilon^{-\beta t}\epsilon^{\beta u} - \beta^2 u\,\epsilon^{-\beta t}\epsilon^{\beta u}]\,du$$

which, on integrating term by term, gives

$$F_4(t) = 1 - \epsilon^{-\beta t}\left(1 + \beta t + \frac{\beta^2 t^2}{2} + \frac{\beta^3 t^3}{6}\right) \quad \ldots \quad \ldots \quad (13)$$

as the output from four circuits when the applied pulse is at the resonance frequency.

For the second case,

$$F_4(t) = \int_0^t \beta u\,\epsilon^{-\beta u}(\beta^2 t\,\epsilon^{-\beta t}\epsilon^{\beta u} - \beta^2 u\,\epsilon^{-\beta t}\epsilon^{\beta u})\,du$$

$$= \frac{\beta^3 t^3}{6}\epsilon^{-\beta t} \quad \ldots \quad \ldots \quad (14)$$

giving the output from four circuits when the applied pulse is not at the resonance frequency. For a case such as this, of four circuits, the effect of the steady-state component can fairly safely be ignored.

Graphs of the functions (13) and (14) are shown as curve 4 on Figs. 2 and 3 respectively, where they can be compared with the results for two circuits in cascade. The slope of the curve at half-steady-state amplitude for the pulse at resonance frequency is compared in Table 1 with the earlier results, and it will be seen that the addition of two circuits to two circuits decreases the slope by 33 per cent. The peak amplitude of the transient when the applied frequency is not the resonance frequency is 0.225 of the peak transient amplitude out of the first circuit.

Consider now a number N of circuits in cascade. It can be shown by performing a few integrations for numbers of circuits other than 2 or 4 that the general results are of the form :—

For applied pulse at resonance frequency*

$$F_N(t) = 1 - \epsilon^{-\beta t}\left(1 + \beta t + \frac{\beta^2 t^2}{2!} + \frac{\beta^3 t^3}{3!} + \ldots + \frac{\beta^{N-1} t^{N-1}}{(N-1)!}\right) \quad \ldots \quad (15)$$

and for applied pulse not at resonance frequency

$$F_N(t) = \frac{\beta^{N-1} t^{N-1}}{(N-1)!}\epsilon^{-\beta t} \quad \ldots \quad \ldots \quad (16)$$

These results are rather striking, and yield the following conclusions :—

(a) When there is an infinite number of circuits, the last term of equation (15) is the expansion of $\epsilon^{\beta t}$, so that $F_\infty(t) = 0$, and build-up never takes place. This is obviously true from physical considertations.

(b) For any given value of N, the maxi-

* This expression for $F_N(t)$ is actually the expansion of the Incomplete Gamma Function

$$\frac{1}{(N-1)!}\int_0^{\beta t} \epsilon^{-\beta t}(\beta t)^{N-1}\,d(\beta t)$$

so that numerical cases can be worked out by reference to tables of this function.[13] This result has been published (since the present paper was written) by Eaglesfield[12, 14], whose method of derivation is completely different from that given here.

The expression (15) is also the same as Poisson's Exponential Probability Summation, and numerical cases can be determined very quickly by reference to tables published by Molina[15], or to charts, published originally by Campbell[16] and Thorndike,[17] and republished by Doust and Josephs.[18]

mum value of equation (16) is found to occur when

$$t = \frac{N - 1}{\beta} \qquad \ldots \qquad (17)$$

which gives an apparent delay time to the peaks of $1/\beta$ seconds per tuned circuit.

(c) The maximum amplitude of equation (16) [occurring at the time given by (17)] is

$$\frac{(N-1)^{N-1}}{(N-1)!}\epsilon^{-(N-1)} \qquad \ldots \qquad (18)$$

Fig. 4 shows a graph of the relationship between this maximum value of the transient and N. It will be seen that the transient

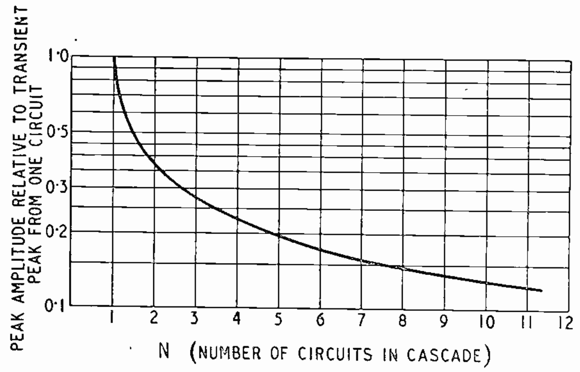


Fig. 4. *Amplitude of transient peak in relation to number of tuned circuits in cascade.*

attenuation of the cascade is increased very little by adding circuits after the fourth or fifth. An increase from 2 to 4 increases the transient attenuation by only 4.8 db.

5. Comparison of Tuned-Circuit Cascades with Band-Pass Filters

As stated in Section 1, there is no direct circuit equivalence between tuned-circuit cascades and band-pass filters, but a comparison of their performances can be made, since to a large extent the frequency function

$$(1/x - x)$$

determined the performance of both.

In practical systems it is likely that the determining design feature will be the slope of the build-up curve at half-steady-state amplitude for applied pulses of the resonance or midband frequency. Thus the comparison is made strictly on this basis; i.e., that the slope must be the same for any circuit combination considered.

There is no need to set out in detail how all the comparison values for tuned-circuit arrangements are derived; all the necessary data has already been given in preceding sections. In the case of band-pass filters, only underived sections are considered, and the derivations of the expressions used below are largely given in other papers by the present author[8, 9]. A more detailed account of the derivations is given in the Appendix.

5.1 *Non-dissipative Filters*

The comparison is made in Table 2, which is self-explanatory. The bandwidth of the filter is $n\omega_0$ radians/sec, where ω_0 is the midband angular frequency. It is assumed that the slope of the midband build-up curve of the filters is $\frac{n\omega_0}{2\pi}$ for one or two sections, which is true to an accuracy of about 15 per cent. [See reference (9), Section 2.3]. If then this is equated to the tuned-circuit slopes given in Table 1, a

TABLE 2

Comparison of Underived Band-pass Filter Performance with Cascaded Tuned Circuit Response

(Comparison based on all arrangements having the same slope of build-up curve at half-steady-state amplitude when applied frequency is equal to midband or resonance frequency)

	1-section filter	2-section filter	1 tuned circuit	2 tuned circuits (equal Q values)	4 tuned circuits (equal Q values)
Transient peak Discrimination (Voltage Ratio)	$(\pi/2n)(x - 1/x)$	$(\pi/n)(x - 1/x)$	$(\pi/2n)(x - 1/x)$	$\frac{0.85\pi}{n}(x - 1/x)$	$\frac{0.92\pi}{n}(x - 1/x)$
Steady-state Discrimination (Voltage Ratio)	$(1/n^3)(x - 1/x)^3$	$(1/n^6)(x - 1/x)^6$ very approximately	$(\pi/2n)(x - 1/x)$	$\frac{0.95}{n^2}(x - 1/x)^2$	$\frac{0.18}{n^4}(x - 1/x)^4$
Time required to attain 90 per cent steady-state Amplitude	approx. $\frac{2.2\pi}{n\omega_0}$	$\frac{2.9\pi}{n\omega_0}$	$\frac{2.3\pi}{n\omega_0}$	$\frac{2.45\pi}{n\omega_0}$	$\frac{2.8\pi}{n\omega_0}$

relationship between Q (for the tuned circuits) and n is established, thus

$$Q = \frac{2\pi A}{n} \qquad \ldots \qquad \ldots \qquad \ldots \qquad \ldots \qquad (19)$$

where A = 0.25 for one tuned circuit,
= 0.155 for two tuned circuits,
= 0.103 for four tuned circuits.

This value of Q is then substituted in the various tuned-circuit equations.

(Note $\beta = \omega_0/2Q$).

The features compared are those which affect the design of multi-channel pulse transmission systems. The actual shape of the various envelopes of the pulses or transients are not compared, although there is a certain amount of correspondence here too.

From Table 2 it will be seen that there is no real general identity between any tuned circuit cascade and a filter. If transient attenuation (e.g., interference suppression or separation from adjacent channels) alone is important, then a single tuned circuit is as good as a 1-section filter using three times as many components. But if steady-state attenuation has to be considered (as it generally must be), then roughly three tuned circuits are required to be equivalent to a 1-section filter.

5.2 *Dissipative Filters*

The effect of dissipation in the filter components has been ignored in Table 2. If the filters have a narrow bandwidth, dissipation will have a large effect in practical cases where ordinary inductors are used. If we take the case in which the filter inductors have the same Q-value as the tuned-circuit inductors, then comparing two tuned circuits with a one-section filter, we have, from equation (19), $Q = \frac{0.31\pi}{n} \approx 1/n$. It was shown in an earlier paper[8] that the steady-state midband loss ratio of a one-section underived filter is approximately

$$1 + 2/Qn$$

which in the present case is approximately $1 + 2 = 3$, i.e. the steady-state midband loss is about 10 db, irrespective of the actual bandwidth or Q value. This quantity must therefore be subtracted from the steady-state discrimination shown in Table 2. For the transient condition this does not apply, because the loss is added to the signal whatever the applied frequency may be, since the transient is of midband frequency in any case.

It will be seen, then, that if the quality of the components is the same for both tuned-circuit cascade and filter, then a two-tuned-circuit cascade will approximate to the same steady-state discrimination as a one-section filter over at least that part of the band not too remote from the pass-band (which is the most critical part in most practical instances), and will have a superior transient discrimination over the whole band. It can therefore be concluded that two tuned circuits in cascade are, on the balance, roughly equivalent in overall performance to a one-section filter using 3 inductors and 3 capacitors; there is a consequent economy of components in using the cascade connection.

APPENDIX I

Derivation of the Expressions in Table 2.

The expressions for transient and steady-state attenuation of the tuned circuit cascades of 1, 2 and 4 circuits are readily derived by the insertion of equation (19) into the various equations given earlier in the paper, e.g. equations (1) and (5) for the steady-state, and equations (3), (11) and (14), or (18), (or alternatively Fig. 3) for the transient state. The time required to attain 90 per cent steady-state amplitude is taken from Fig. 2.

For the band-pass filters, the steady-state attenuation is obtained from Reference 8, Part 1, Section 2, equation (4). This gives the insertion loss of one whole section between terminating resistances equal to the design resistance. Therefore, in taking the square of this as the loss ratio for two sections, an approximation is made because the effect of the impedance match at the junction between sections is not allowed for.

The transient peak attenuation (or discrimination) is obtained from Reference 9; Section 2.2, equation (5), gives the result shown to apply to a two-section filter; and from Section 2.3 and Fig. 1 of the same reference we get the fact that the attenuation is roughly half as great for a single-section filter.

It is to be noted that in all these comparisons the effect of initial phase angle is neglected; it is assumed the angle is zero, corresponding to an applied signal sin ωt. It was shown in Section 2 of the present paper that the transient peak for tuned circuits is $1/x$ times as large when the initial phase angle is $\pi/2$. It is shown in Section 2.2, equation (8) of Reference 9 that the same relationship applies in the case of band-pass filters. Thus the comparison is unchanged whatever initial phase angle is taken.

The time required for filter build-up curves to reach 90 per cent steady-state amplitude is taken from the results given in Reference 9.

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[6] Massachusetts Institute of Technology, "Electric Circuits," p. 356 onwards (John Wiley).

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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.

Carrier-Frequency Amplifiers

To the Editor, "Wireless Engineer."

SIR,—In a recent paper ("*Wireless Engineer*," Vol. 23, No. 271, pp. 96–102, April, 1946) I discussed the transient response of a carrier-frequency amplifier to a sudden change in the frequency of the carrier. Professor van der Pol has drawn my attention in private correspondence to a difficulty in the analysis. The point is as follows:

If an admittance $A(p)$ is 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', then the current in the admittance is given by

$$i = (1-P)\ \cos\omega_0 t + Q\sin\omega_0 t + P_1\cos\omega_0' t - Q_1\sin\omega_0' t$$

This is equation (5) of the original paper; the coefficients P, Q, P_1, Q_1 are functions of t and are given by

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

$$P_1(t) + j\,Q_1(t) = \frac{A(p + j\omega_0')}{A(j\omega_0)}\,\mathbf{1}$$

In these expressions, p is the differential operator d/dt, $\mathbf{1}$ is the Heaviside step function, and $j = \sqrt{-1}$.

To obtain the effective frequency of i, the following definition is used:—

$$\text{Effective Frequency} = \frac{\text{Average of } di/dt}{\text{Average of } i}$$

the averages being root-mean-square.

In the original paper the assumption was made that for taking the averages a duration of time could be chosen, which was long compared to $1/\omega_0$ and $1/\omega_0'$, but during which P, etc., were sensibly constant. But in calculating the averages it was also implicity assumed that the duration of time was long compared to $1/(\omega_0 - \omega_0')$. A little consideration will show that, in any practical application of the equations to a network, this last assumption is likely to conflict with the assumption that P, etc., remain constant during the time interval. For if the network has a fairly well-defined pass-band, and ω_0 and ω_0' are placed as far apart as possible (on the edges of the band), this will make $1/(\omega_0 - \omega_0')$ as small as possible.

But it is known, in a general way, that the transition time is of the order of the reciprocal of the pass-band, so that P, etc., will not be constant during $1/(\omega_0 - \omega_0')$; even less will they be constant during a time interval which is long compared to $1/(\omega_0 - \omega_0')$.

In order to simplify the expressions in the averages, it is now assumed that the time interval for taking the averages is short compared to $1/(\omega_0 - \omega_0')$. In squaring the expression for i it is now necessary to retain the cross-product terms which were previously neglected. They lead to terms in $\cos(\omega_0 - \omega_0')\,t$ and $\sin(\omega_0 - \omega_0')t$, and with the new assumption they are approximately unity and zero respectively.

Recalculating the averages on this basis, it will be found that

$$\begin{aligned}2\,\bar{i}^2 &= (1-P)^2 + Q^2 + P_1^2 + Q_1^2 + 2\,(1-P)\,P_1 - 2\,QQ_1\\ &= 1 + 2\,(P_1 - P) + (P_1 - P)^2 + (Q_1 - Q)^2.\end{aligned}$$

Now since $(\omega_0 - \omega_0')$ is being taken very small, it is clear that $(P_1 - P)$ and $(Q_1 - Q)$ are also very small.

Thus $2\,\bar{i}^2 = 1$

That is to say, the amplitude remains constant during the transition.

Handling $\overline{\dfrac{di}{dt}}^2$ in the same way, we get for the effective frequency ω:—

$$\omega^2 = \left[(1-P)^2 + Q^2\right]\omega_0^2 + \left[P_1^2 + Q_1^2\right]\omega_0'^2 + 2\,(1-P)\,P_1\,\omega_0\,\omega_0' - 2\,QQ_1\,\omega_0\,\omega_0'$$

By assuming $(\omega_0 - \omega_0')$, $(P_1 - P)$, $(Q_1 - Q)$ all small, this is reduced to the simple expression:

$$\omega = \omega_0 + (\omega_0' - \omega_0)\,P.$$

That is, the transient modulation ratio is P, a quantity which can be seen from its definition to vary from zero at $t = 0$ to unity at $t = \infty$.

It should be noted that the new assumption, that ω_0' is nearly equal to ω_0, is equivalent to postulating a small depth of frequency modulation.

Now it was shown in the original paper (Equation 14*b*) that the transient modulation ratio for a small amplitude modulation was also P.

We thus have the interesting result that the transient modulation ratio is the same for either frequency or amplitude modulation, the modulation being small in each case, and in the form of a unit step. But it is possible to go further: since the transient modulation ratio is an impedance it follows that the output modulation for any waveform of input modulation can be obtained from the step response (Borel's Theorem). It thus follows that the modulation ratio is the same for both amplitude and frequency modulation, whatever the waveform of the input modulation, provided it is small.

It may be well to repeat this statement in a slightly different way.

If a carrier frequency, with a small amplitude modulation, is applied to the input terminals of a linear four-terminal network, then the amplitude modulation at the output terminals will differ from the input modulation. Also, if the carrier has a small frequency modulation, the output frequency modulation will differ from the input modulation. The distortion of the modulation is independent of the type of modulation, amplitude or frequency, but depends entirely on the network.

The actual waveform of the output modulation can be calculated, if the network and the waveform of the input modulation be known. But unless at least one of these is simple in form, the calculation would be laborious.

In the original paper, equations (6), (7), (8), and Figs. 4, 5, should be disregarded. They have no practical application.

However, the curves given previously for amplitude modulation apply directly to frequency modulation. They will be found in *Wireless Engineer*, Vol. 22, No. 266, p. 523, November 1945, and also Vol. 23, No. 270, p. 67, March 1946.

I am extremely grateful to Professor van der Pol for drawing my attention to this matter, and to Dr. F. L. Stumpers for communicating his notes on my paper, in which by a slightly different treatment he reached substantially the result given here.

Bournemouth. C. C. EAGLESFIELD.

BOOK REVIEWS

Introduzione alla Radiotelemetria (Radar)

By Prof. Ugo Tiberio. Pp. 277 and 137 Figs. Editore Rivista Marittima Roma. 300 Lira.

The author is Professor of Electrotechnics at the Royal Naval Academy and has evidently made a special study of radar. The book presupposes a general knowledge of radio principles and practice and begins straight away with references to the experiments of Breit and Tuve and of Appleton and Barnett. In the third line there is a reference to Heavisyde (*sic*) and the *imosfera* (presumably a misprint for the ionosfera), which is a bad start. The book is divided into three parts and at the end a footnote states that there is a fourth part which will be published separately. Part I is a general introduction to the methods of radiolocation, the use of cathode-ray tubes, the reflection of waves, etc. Part II is concerned with the circuit arrangements for generating the impulses or waves, special short-wave aerials and reflectors, and the corresponding receiving apparatus. Part III is devoted to the theory of the transmission and reflection, polar curves of radiation, pulse distortion, and allied problems.

G. W. O. H.

Télétransmissions par Ondes Porteuses dans les Réseaux de Transport d'Energie a Haute Tension

By ANDRÉ CHEVALLIER. Pp. 111 + x and 124 Figs. Dunod, 92, rue Bonaparte, Paris (VI).

The development of an underground 60 000-volt network in 1920 necessitated new methods of protection which were developed by M. Fallou. This work, interrupted by his death, has since been taken up and developed successfully. The book is based on courses of lectures given by the author since 1942 at the Ecole Supérieure d'Electricité, It deals with the superposition of high-frequency currents on the high-voltage grid, their modulation. transmission, etc. These high-frequency carrier waves are used to transmit the necessary information, measurements, controls, etc. After explaining modulation and detection, the author considers the various ways in which a 3-phase line can be used to transmit the high-frequency carrier wave, and develops the transmission formulae for the various arrangements. The capacitors employed to couple the h.f. apparatus to the high voltage line, which may be at 220 kV, present some problems; high-pass and band-pass filters are considered as are also rejector circuits, spark-gap protective devices and quartz filters. A chapter is devoted to the problems involved in using such a system for telephony and another chapter to its application to various types of protection. The book contains a large number of diagrams of connections, graphs, etc.; it should certainly be studied by those interested in the protection of large high-voltage networks by means of superposed high-frequency carriers.

G. W. O. H.

University of Glasgow

THE retirement of Professor G. W. O. Howe from the James Watt Chair of Electrical Engineering is taking place at the end of the present session. We are asked to say that a Presentation to mark the occasion is being arranged and contributions are invited from old students and other friends.

The arrangements are in the hands of Dr. A. J. Small, Electrical Engineering Dept., The University, Glasgow, W.2, to whom contributions should be sent.

H.S.P.

Parsons Memorial Lecture

THE Parsons Memorial Lecture for 1946, arranged by the Institute of Civil Engineers, will be given on September 26th, 1946, at 5.30 p.m. at the Institution of Civil Engineers, Gt. George St., London, S.W.1. The lecture is by Sir Hugh Chance and is on "Recent Developments in Optical Glass Manufacture".

Members of the Institution of Electrical Engineers are invited to attend.

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

575 306.—Visual warning signal to prevent any improper manipulation of the controls of an automatic or record-changing gramophone, during the playing period.

Industriakt Luxor. Convention date (Sweden) 20th February, 1943.

575 327.—Valve potentiometer for stabilizing the voltage across a variable resistance-load, and particularly for controlling the volume of an electronic musical organ.

A. H. Midgley. Application date 15th December, 1943.

575 443.—Balanced bridge circuit, including temperature-compensating coils, for detecting land-mines, booby-traps, or the presence of submarines at sea.

Cinema-Televison Ltd. and S. S. West. Application date 29th March, 1943.

575 574.—Coupling arrangement in which fixed primary and secondary coils are associated with a fixed intermediate coil and with a movable non-magnetic cylinder, in order to simulate a resistance coupling, say for detecting buried metals.

Hazeltine Corporation (assignees of L. F. Curtiss). Convention date (U.S.A.) 31st December, 1942.

575 671.—Apparatus for observing by oscilloscope the nature of acoustic and other transients and for making magnetic records for reproduction.

The Brush Development Co. (assignees of S. J. Begun). Convention date (U.S.A.) 1st January, 1942.

AERIALS AND AERIAL SYSTEMS

575 097.—Moulded terminal block of polythene for coupling and supporting say a coaxial feed-line to dipole aerials.

A. D. Ferguson and Metropolitan-Vickers Electrical Co. Ltd. Application date 26th November, 1941.

575 534.—Aerial for radiating micro-waves consisting of an unsheathed rod of dielectric, part of its length being of uniform and the rest of tapering cross-section.

Western Electric Co. Inc. Convention date (U.S.A.) 17th December, 1942.

DIRECTIONAL WIRELESS

575 154.—Blind-landing system in which the transmitter radiates a frequency-modulated wave and the aeroplane develops a control current based on the difference between the direct wave and that received after reflection from the ground.

Hazeltine Corp. (assignees of H. M. Lewis). Convention date (U.S.A.) 26th February, 1942.

575 430.—Highly-directive aerial array, including phase-inverting or suppressor elements, particularly suitable for relaying frequency-modulated programmes from the studio to the main transmitter.

The British Thomson-Houston Co. Ltd. Convention date (U.S.A.) 31st December, 1942.

575 432.—Radiolocation equipment in which the common transmitting and receiving aerial is formed by the flared end of a wave-guide, the body of which comprises automatic blocking and acceptor circuits.

Western Electric Co. Inc. Convention date (U.S.A.) 30th January, 1943.

RECEIVING CIRCUITS AND APPARATUS

(See *also under Television*)

575 188.—Receiver in which the signal and " noise " voltages are developed across impedances having long and short time-constants, respectively, in order to balance out the disturbances.

The British Thomson-Houston Co. Ltd. Convention date (U.S.A.) 31st October, 1942.

575 250.—Time-base voltage-generator, say for a c.r. tube, in which a valve is negatively back-coupled through a time-constant network which may be of the integrating or differentiating type.

A. C. Cossor Ltd. and J. W. Whiteley. Application date 17th February, 1942.

575 329.—Push-button tuning system for a superhet set in which each selected frequency depends upon two tuned circuits, one pre-set and the other continuously variable.

A. C. Cossor Ltd. and D. A. Bell. Application date 21st December, 1943.

575 493.—Portable receiver fitted with tuning controls which are so arranged that they can be conveniently operated when the set is being carried by one hand.

Philco Radio and Television Corp. (assignees of R. J. Whipple). Convention date (U.S.A.) 11th February, 1943.

575 583.—Visual tuning indicator, requiring no d.c. amplifier, for frequency or phase-modulated signals.

Marconi's W.T. Co. Ltd. (assignees of M. G. Crosby). Convention date (U.S.A.) 27th March, 1943.

575 627.—Construction of flat, ribbed valve socket or holder, particularly for the midget type of tube.

Cinch Manufacturing Corp. (assignees of S. M. Del Camp). Convention date (U.S.A.) 16th December, 1942.

575 639.—Amplifier consisting of single valve-stages coupled in cascade through anode and cathode impedances which are selected to produce zero phase-displacement over a range of frequencies.

B. M. Hadfield. Application date 14th December, 1943.

575 796.—Making a permanent record of short time-intervals, as in radiolocation, by means of a device in which a stylus is traversed bodily over a time-base resistance or potential-divider.
N. P. Hinton and C. S. Wright. Application date (for a secret patent) 14th May, 1942. Published 30th November, 1945.

TELEVISION CIRCUITS AND APPARATUS

For Transmission and Reception

575 060.—Television C.R. tube in which the transparency of a screen, subjected to external light, is modified by the heating-effect of the scanning-stream.
G. Liebmann and Cathodeon, Ltd. Application date 20th November, 1941.

TRANSMITTING CIRCUITS AND APPARATUS

(See also under Television)

575 023.—Construction of wave-guide designed to admit of bending or twisting without loss of transmission-efficiency.
Callenders Cable and Construction Co. Ltd., D. T. Hollingsworth and C. W. J. Morley. Application date 6th July, 1944.

575 156.—Frequency-discrimination circuit for stabilizing a short-wave oscillator say between modulation intervals.
The General Electric Co. Ltd. and J. B. Lovell-Foot. Application date 18th June, 1943.

575 228.—Construction of wave-guide the transmission-efficiency of which is unimpaired by flexing or by mechanical vibration.
W. D. Allen. Application date 8th March, 1944.

575 347.—Screwdriver device for adjusting the tuning of radio components which are normally enclosed or sealed against the high humidity of tropical climates.
Standard Telephones and Cables Ltd. and G. Newton. Application date 1st March, 1944.

575 395.—The use of a thermister—a device having a temperature coefficient of resistance—for stabilizing the frequency of a two-stage resistance-coupled oscillator.
Standard Telephones and Cables Ltd., G. C. Hartley and J. A. B. King. Application date 15th March, 1944.

575 511.—Purely-resistive terminating-device for a coaxial transmission-line, including a crystal for measuring the shunt voltage.
Marconi Instruments Ltd. and E. W. Hunt. Application date 1st February, 1944.

575 577.—Wave-guide comprising an adjustable phase-shift section for transmitting a wide band of frequencies, and rotary joints for converting the waves from linear to circular polarization.
Western Electric Co. Inc. Convention date (U.S.A.) 3rd November, 1942.

575 739.—Variable-impedance device for a wave-guide in which a diaphragm acting as a reactance is associated with one or more adjustable probes acting as capacitances.
The British Thomson-Houston Co. Ltd., L. W. Brown and J. H. Nicoll. Application date 30th March, 1944.

SIGNALLING SYSTEMS OF DISTINCTIVE TYPE

575 213.—Speech-free signalling and supervisory circuits arranged between the frequency zones in a carrier-wave multi-channel system.
Automatic Telephone and Electric Co. Ltd. and L. J. Murray. Application date 7th March, 1944.

575 384.—Compact network, simulating a double transmission-line, for the generation of periodic square-shaped signalling impulses.
The British Thomson-Houston Co. Ltd. and K. J. R. Wilkinson. Application date 3rd February, 1942.

575 462.—Two-way signalling system in which the same aerial serves to transmit morse and to receive incoming speech or other signal in the intervals of transmission.
Standard Telephones and Cables Ltd. and E. G. Seath. Application date 17th March, 1944.

CONSTRUCTION OF ELECTRONIC DISCHARGE DEVICES

574 758.—Arrangement of the cooling-fins and transverse baffle-plates for a high-powered electron-discharge tube.
Standard Telephones and Cables, Ltd. and W. T. Gibson. Application date 15th February, 1944.

574 934.—Toroidal resonator, of the rhumbatron type, wherein bimetallic tuning control is operated (a) by the rise in working temperature, or (b) by the deliberate application of heat.
The M-O Valve Co., Ltd., N. L. Harris and J. W. Ryde. Application date 22nd April, 1940.

574 878.—Electron-discharge tube of the "light-transformer" type, designed to facilitate the application of a photo-sensitive coating to the cathode in the course of manufacture.
J. D. McGee. Application date 17th November, 1942.

574 935.—Electron-discharge tube designed to facilitate the connection of an external concentric feed-line to one of the electrodes, particularly for velocity-modulation.
The M-O Valve Co., Ltd., N. L. Harris and R. W. Sloane. Application date 3rd June, 1940.

574 943.—Coupling circuit for taking power from magnetron oscillators of the kind in which the anode-block comprises a number of cavity resonators.
Western Electric Co., Inc. Convention date (U.S.A.) 1st May, 1942.

574 967.—Electron-discharge tube in which the electron stream is first deflected to and fro through a perforated screen to produce a frequency-multiplying effect before being fed to a hollow resonator or rhumbatron.
The M-O Valve Co., Ltd., J. W. Ryde and R. W. Sloane. Application date 8th October, 1940.

574 968.—Tuning a velocity-modulating device, or rhumbatron, lying partly within an electron-discharge tube, by means external to the evacuated tube.
The M-O Valve Co., Ltd., N. L. Harris and R. F. Proctor. Application date 8th October, 1940.

574 969.—Construction and arrangement of the hollow-resonator in an electron-discharge tube designed for velocity-modulation.
The M-O Valve Co., Ltd. and N. L. Harris. Application date 8th October, 1940.

574 972.—Short-wave oscillator wherein the electron stream is reflected back through the original bunching resonator across a path of specified field-gradient.

S. Hill. Application date 17th January, 1941.

574 997.—Construction of the target or reflecting electrode associated with a rhumbatron resonator in order to facilitate degassing of the electron-discharge tube.

J. E. I. Cairns. Application date 8th August, 1941.

575 123.—Discharge-tube in which the grids of two coaxial resonators are relatively adjusted by magnetic or other means external to the tube.

Standard Telephones and Cables, Ltd. (assignees of C. V. Litton). Convention date (U.S.A.) 20th April, 1940.

575 249.—Electron-multiplier with means for focusing the primary and secondary electrons in order to increase the ratio of transconductance to plate current.

Marconi's W.T. Co. Ltd. (communicated by Radio Corporation of America). Application date 10th September, 1941.

575 320.—Process for making "honeycomb" electrodes or grids intended to respond to the action of a high-speed electron-beam.

The British Thomson-Houston Co. Ltd. and W. J. Scott. Application date 3rd November, 1943.

575 747.—Process for making the flattened viewing-end of the glass bulb of a cathode-ray tube.

Chance Brothers Ltd., W. M. Hampton and N. J. B. Raymond. Application dates 4th April and 10th August, 1944.

575 824.—Moulding process for making metal-to-glass seals, say in the manufacture of electron-discharge devices.

Standard Telephones and Cables Ltd. and T. W. Wingent. Application date 13th April, 1944.

575 962.—Cathode-ray tube in which the end of the glass bulb is lenticular with one of its surfaces cylindrical, in order to minimize distortion.

Chance Bros. Ltd., J. G. Holmes and J. English. Application date 17th March, 1944.

SUBSIDIARY APPARATUS AND MATERIALS

574 840.—Equipment, including mechanical gear, for selecting and stabilizing the generation of a number of "slave" frequencies derived from a "master" oscillator, such as a piezo-electric crystal.

The General Electric Co., Ltd. and A. S. Gladwin. Application date 1st March, 1943.

574 916.—High-frequency induction-heater designed to offset the magnetomotive forces set up within the work.

Standard Telephones and Cables, Ltd. (assignees of V. W. Sherman). Convention date (U.S.A.) 25th February, 1943.

574 996.—The use of calcium-fluoride powder as a stabilizing agent for the spark-gap of a high-frequency ignition system.

The Plessy Co. Ltd. and H. V. G. Stubbs. Application date 1st July, 1941.

575 068.—Process for attaching electrodes to piezo-electric elements of the fusible or multi-plate type.

The Brush Development Co. (assignees of C. K. Gravley). Convention date (U.S.A.) 16th December, 1942.

575 223.—Combination of a standard-frequency oscillator with a continuously-variable oscillator in order to allow the "spot measurement" of a drifting frequency.

H. J. Finden and The Plessy Co. Ltd. Application date 8th March, 1944.

575 252.—Arrangement of an auxiliary priming capacitor across the glow-discharge tube of a detonating or like relay.

D. Weighton and Pye Ltd. Application date 25th February, 1943.

575 256.—Impedance bridge which is unresponsive to voltages that are up to 90° out-of-phase with those to be measured.

Honeywell-Brown Ltd. Convention date (U.S.A.) 24th October, 1942.

575 411.—Process for forming selenium "wafers" as used for electric rectifiers and light-sensitive cells.

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

575 447.—High-frequency testing equipment of the type in which a signal-generator is coupled through an attenuating circuit to the apparatus under test.

W. L. Watton. Application date 17th December, 1943.

575 463.—Magnetometer bridge with accessories for measuring the total intensity as well as the horizontal and vertical components, say of the earth's magnetic field.

H. Hughes and Son Ltd., E. Smith and A. J. Hughes. Application date 18th March, 1944.

575 719.—Construction designed to reduce the weight of a power-supply transformer, particularly for use on aircraft.

Standard Telephones and Cables Ltd. (assignees of F. W. Edmonds). Convention date (U.S.A.) 10th April, 1943.

575 805.—Method of coupling and calibrating a series arrangement of piezo-electric oscillators in order to increase the power that the resulting unit can handle.

B. Tenenbaum. Application date 6th April, 1944.

576 097.—Process and apparatus for bonding together the layers of plywood by high-frequency current.

The General Electric Co. Ltd. and E. McP. Leyton. Application date 5th October, 1942.

576 192.—Optical method of preparing long calibration-strips for use with an oscillator of wide frequency-range.

Standard Telephones and Cables Ltd. and I. V. Feldhusen. Application date 14th April, 1944.

576 473.—Cathode-heating switch having a bimetallic strip under a "click" control for the sharp operation of a glow discharge tube.

The General Electric Co. Ltd., W. G. Branson and R. W. Strong. Application date 14th May, 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 : 621.392 2447

Electro-Mechanical Analogy in Acoustic Design.—A. M. Wiggins. (*Radio, N.Y.*, April 1946, Vol. 30, No. 4, pp. 28–29.) An explanation and justification of the analogy whereby mechanical problems can be solved by the solution of equivalent electrical circuits. The method is applied to a unidirectional microphone.

534.2 2448

The Absorption of Sound of High Frequency in Metals.—L. Gurevich. (*Zh. eksp. teor. Fiz.*, 1944, Vol. 14, No. 6, pp. 202–204.) From an equation (1) determining the change in the number of sound quanta (phonons) resulting from their interaction with electrons, a formula is derived for calculating the absorption coefficient τ_g. It appears that τ_g is proportional to the sound frequency. It is also shown that for frequencies exceeding the inverse value of the time of the free travel of electrons, sound is absorbed during an interval of the order of the sound period, *i.e.* propagation cannot take place.

This paper is related to 2232 of 1937 (Landau & Rumer).

534.321.9 2449

Ultrasonic Velocity in Water.—P. L. F. Jones & A. J. Gale. (*Nature, Lond.*, 16th March 1946, Vol. 157, No. 3985, p. 341.) A graph is given showing the velocity as a function of temperature. There is a maximum of 1552.7 m/s at 72.7°C. See also B. K. Singh, *Nature, Lond.*, 1945, Vol. 156, p. 569.

534.321.9 2450

On the Measurement of Ultra-Sound Absorption in Gases by Spherical Waves Methods.—P. Krasnushkin. (*Zh. eksp. teor. Fiz.*, 1944, Vol. 14, No. 5, pp. 152–155.) The advantages of using spherical instead of plane waves for the measurements are pointed out, and the following two new methods proposed: (*a*) A point receiver is moved along the axis of the central diffraction lobe of the radiation field of a point radiator, and amplitudes V of the field are measured with respect to distance R between the radiator and the receiver. (*b*) The receiver is replaced by a metallic plane that reflects the waves back to the radiator. With a continuous movement of the plane the acoustic reactance of the radiator and therefore the anode current I_a of the oscillator are varied. A formula determining the relationship between I_a and d (distance between the plane and the radiator) is given.

The results obtained by both methods in room air are shown in a table. It appears that for the frequencies used (400–710 kc/s) $d\lambda^2$ remains constant within 7%, and its average value is 23.7×10^{-5}, *i.e.* it exceeds by 44% the value given by the classical theory.

534.321.9 : 621.396.9 2451

Ultrasonic [radar] **Trainer Circuits.**—Larsen. (*See* 2582.)

534.417 : 534.88 2452

Navy Releases Sonar Story.—(*Electronics*, May 1946, Vol. 19, No. 5, pp. 284..294.) A general account of the system and its history. See also 1750 of July (Lanier & Sawyer).

534.43 : 621.395.61 2453

A New Moving-Coil [gramophone] **Pickup.**—(*Electronic Engng*, July 1946, Vol. 18, No. 221, pp. 224–226.) Detailed description of the "Lexington" pickup, which has a flat response from 30 c/s to 12 kc/s, with a weight of ½ oz on the record. Sapphire or steel needles of special shape are used.

534.43 : 621.395.61 : 538.652 2454
Torsional Magnetostriction [gramophone] **Pickup.**—S. R. Rich. (*Electronics*, June 1946, Vol. 19, No. 6, pp. 107–109.) The device makes use of the variation of magnetic reluctance in a wire subjected to torsion in a magnetic field. It has a small moving mass, low distortion, and wide frequency response. The torsional magnetostriction element also operates successfully as a recording mechanism.

534.43 : 621.395.645.3 2455
Unique Phono Amplifier.—Pett. (*See* 2515.)

534.61 : 621.317.35 2456
Range Extender for General Radio 760A Sound Analyzer.—J. D. Cobine & J. R. Curry. (*Rev. sci. Instrum.*, May 1946, Vol. 17, No. 5, pp. 190–194.) Details of a circuit to extend the frequency range to 1 Mc/s by a heterodyne principle.

621.395.2 + 621.395.625 2457
Nuernberg Trials Recording System.—P. C. Erhorn. (*Electronic Industr.*, June 1946, Vol. 5, No. 6, pp. 70..114.) A block diagram of the equipment is given, with a general description of the circuits. Eight microphones and five hundred pairs of headphones are catered for, and provision is made for various recorders and for broadcasting.

621.395.613.32 2458
Microphones: Part 3 (cont.).—S. W. Amos & F. C. Brooker. (*Electronic Engng*, July 1946, Vol. 18, No. 221, pp. 221–223.) A description of various makes of microphone of the moving coil, condenser, and piezoelectric types, and of their equivalent circuits, including the acoustic networks incorporated to maintain an even response curve. A polar diagram for a typical pressure-operated microphone shows variation of directional properties with frequency. For parts 1 and 2 see 1755 of July; for part 3(a) see 2120 of August.

621.395.614 2459
Sound-Pressure Measurement Standard.—F. Massa. (*Electronics*, May 1946, Vol. 19, No. 5, pp. 218..228.) A microphone comprising a pile of piezoelectric crystal plates in a rigid housing has wider frequency and dynamic ranges than other microphones generally available for making absolute sound measurements.

621.395.623.54 : 621.395.92 2460
A New Earpiece [for deaf-aid equipment].—C. M. R. Balbi. (*Wireless World*, June 1946, Vol. 52, No. 6, p. 179.)

621.395.623.7 2461
Corner [loudspeaker] **Deflector Baffles.**—(*Wireless World*, June 1946, Vol. 52, No. 6, p. 181.) The walls of the room housing the loudspeaker are used as elements in the combined horn and baffle system. Two outward radiating paths of logarithmically increasing section are produced, and a diffuser is incorporated to give even distribution of high frequencies. There is ample bass response, and the full rated power is delivered without any signs of overloading. A short illustrated description of the system.

621.395.625.2 2462
Embossing Sound on Film.—S. Kempner. (*Radio News*, June 1946, Vol. 35, No. 6, pp. 36..108). A general description of the Recordgraph and its performance. See also 537 of March.

621.395.625.3 2463
The German Magnetophon.—R. A. Power. (*Wireless World*, June 1946, Vol. 52, No. 6, pp. 195–197.) A description of the magnetic recording equipment in which the medium is a polyvinylchloride strip impregnated with an equal weight of finely powdered magnetic Fe_2O_3. Compared with other tape or wire recorders the equipment offers (*a*) better quality (25–10 000 c/s, dynamic range about 70 db with 2% distortion), (*b*) a lighter, tougher, and cheaper medium, (*c*) easy cutting and splicing of the tape, (*d*) facility for writing notes, titles, etc., along the roll. See also 834 of April and back references.

621.395.645.3 2464
Additional Notes on the Parallel Tube Amplifier.—F. C. Jones: J. Velasco. (*Radio, N.Y.*, June 1946, Vol. 30, No. 6, pp. 26..38.) Notes on an experimental high-fidelity amplifier with a parallel-tube output and separate high- and low-frequency tone controls. Follows a previous paper on the subject, 263 of February (Jones). A comment on the latter by Velasco is included.

621.395.82 : 621.395.645 : 621.317.79 2465
Measuring Audio Intermodulation.—Pickering. (*See* 2641.)

621.396.667 2466
Low-Frequency Correction Circuit.—(*See* 2532.)

621.396.667 2467
Tone Correction.—Gregory. (*See* 2533.)

AERIALS AND TRANSMISSION LINES

621.392 2468
On the Calculation of the Radiation Field of a Waveguide.—N. Maloff. (*Zh. eksp. teor. Fiz.*, 1944, Vol. 14, No. 6, pp. 224–225.) In calculating the field at the open end of a waveguide from Kirchhoff's formula it is usual to assume that the configuration of this field is the same as that of the field inside an infinitely long waveguide. In the present paper the validity of this assumption for the H_{01} mode in a cylindrical guide is examined by checking whether the ratio A_2/A_1 remains equal to unity for all values of the ratio λ/λ_0 $(= \Lambda)$, where A_1 is the energy flux through the cross-section of the waveguide, A_2 the energy flux through a sphere at the centre of which the opening of the waveguide is located, λ the free-space operating wavelength, and λ_0 the critical wavelength. The calculated results, which are collected in a table, throw considerable doubt on the validity of the assumption, particularly in the region of most practical interest, *i.e.* when $\lambda/\lambda_0 > 0.8$.

621.392 2469
On the Propagation of Electromagnetic Waves in Curved Pipes.—M. Jouguet. (*C. R. Acad. Sci. Paris*, 18th Feb. 1946, Vol. 222, No. 8, pp. 440–442.) A theoretical analysis of a curved guide of circular cross-section excited in the E_0 and H_0 modes. It is concluded that for E_0 the curve causes no change in phase velocity and that the E and H fields are not orthogonal. Analysis for H_0 gives incompatible equations, from which it is concluded that waves cannot be propagated in this mode.

621.392 2470
Wave Guide Transmission Systems.—T. Moreno. (*Electronics*, June 1946, Vol. 19, No. 6, pp. 136–141.) A sequel to 2136 of August. A discussion of the attenuation and standing waves produced by various joints and bends used in waveguides. If the inner radius of a bend is greater than a guide wavelength λ_g, the voltage standing wave ratio produced will be under 1.05. A rectangular guide of length $2\lambda_g$ or more, twisted by 90° about its axis, will introduce a s.w.r. generally less than 1.1. Graphs showing the required dimensions of corner connectors for minimum reflection are given, and couplings, tee-joints, matching diaphragms, and coaxial-line transformers discussed.

621.392.2 2471
Propagation along a Line having only Distributed Resistance and Capacitance which are Functions of Position but have a Particular Relationship to Each Other.—M. Parodi. (*C. R. Acad. Sci., Paris*, 3rd Sept. 1945, Vol. 222, No. 10, pp. 257–259.) A mathematical paper. For a particular relationship between C and R, given in the paper, the differential equation can, by change of variable, be transformed to one with constant coefficients and solved explicitly.

621.392.2 2472
Remarks on the Equations of Propagation on any Line.—F. Raymond. *C.R. Acad. Sci., Paris*, 2th April 1945, Vol. 220, No. 14, pp. 497–500.) A mathematical paper. It gives formally the solution for the propagation of any disturbance along a transmission line of which the characteristics L, C, R, and G are functions of the distance along the line.

621.392.2 2473
Propagation along any Polyphase Symmetrical Line.—M. Parodi & F. Raymond. (*C.R. Acad. Sci., Paris*, 9th April 1945, Vol. 220, No. 15, pp. 522–523.) A very general formal matrix analysis of a symmetrical system of n lines with non-uniform characteristics.

621.392.2 2474
Transmission Problems.—R. H. Paul. (*Electrician*, 26th April & 3rd May 1946, Vol. 136, Nos. 3543 & 3544, pp. 1097–1099 & 1165–1167.) A discussion of " some rigorous methods of solving problems connected with long transmission lines without having recourse to hyperbolic functions of complex angles or convergent series."

621.392.2 2475
Some Novel Expressions for the Propagation Constant of a Uniform Line.—J. L. Clarke. (*Bell Syst. tech. J.*, Jan. 1946, Vol. 25, No. 1, pp. 156–157.) By a simple extension of well-known equations for the characteristics of a line, the attenuation constant is expressed in terms of the electrostatic and electromagnetic energies per unit length of line, and the characteristic impedance is expressed in terms of the phase velocity.

621.392.2 2476
Propagation Characteristics of a Uniform Line.—F. Macdiarmid & H. J. Orchard. (*Wireless Engr*, June 1946, Vol. 23, No. 273, pp. 168–171.) A geometrical method of deriving the real and imaginary components of the propagation constant of a uniform transmission line is give The form of the expressions thus derived is such as to indicate the physical significance of the different line parameters.

621.392.2 2477
Simplified Input Impedance Chart for Lossless Transmission Lines.—L. Mautner. (*Communications*, May 1946, Vol. 26, No. 5, pp. 44–45, 63.) A chart giving the range of input impedance for lines of length $n\lambda/8$ terminated resistively and reactively.

621.392.2 : 621.396.61 2478
Transmission Lines as Resonant Circuits.—L. R. Quarles. (*Communications*, May 1946, Vol. 26, No. 5, pp. 22..52.) Formulae for calculating the dimensions for any reactance using either coaxial or twin lines, with some examples of their application. Part 1 of a 3-part series, for part 2 see 2480 below.

621.392.43 2479
Graphical Calculation of Double Stubs.—R. C. Paine. (*Radio, N.Y.*, June 1946, Vol. 30, No. 6, pp. 23–25..37.) Circle diagrams, together with a parabolic locus defining the admittance of stubs, can be used to solve double-stub transmission-line matching problems. The diagrams are given, and examples are worked out.

621.392.52 2480
Transmission Lines as Filters.—L. R. Quarles. (*Communications*, June 1946, Vol. 26, No. 6, pp. 34..48.) The design of filters for the following types of application are considered: (*a*) suppression of unwanted r.f. harmonics on transmission lines, by means of a shunt stub; (*b*) band-pass intercircuit coupling with filters of T configuration; (*c*) wide-band matching filters. Part 2 of a series beginning with 2478 above.

621.396.67 2481
Coaxial Feed F.M. Loop Antennas.—A. G. Kandoian. (*Electronic Industr.*, May 1946, Vol. 5, No. 5, pp. 74..126.) Paper based on 1180 of May by the same author.

621.396.671 : 621.396.822 2482
Fluctuation Noise in a Receiving Aerial.—Burgess. (*See* 2699.)

621.396.674 2483
Inductive Tuned Loop Circuits: Parts 1 & 2.—W. J. Polydoroff. (*Radio, N.Y.*, April & May 1946, Vol. 30, Nos. 4 & 5, pp. 21–22 & 20–22.) The advantages of permeability-tuned loop aerials in respect of signal/noise ratio and directional discrimination against interference are described. The need for balancing and shielding the loop to obtain these advantages is explained.

621.396.674 : 621.317.79 2484
An Improved Method of Testing Loop Receivers.—W. J. Polydoroff. (*Radio, N.Y.*, June 1946, Vol. 30, No. 6, pp. 15–17, 36.) A single-wire transmission line is strung across a screened room. One end of the line is connected to a signal generator, the other is terminated with its characteristic impedance. The radiation simulates the field of a horizontally propagated vertically polarized wave. The receiver loop under test is supported underneath the transmission line.

621.396.674 : 621.318.323.2.029.5 2485

Iron-Cored Loop Receiving Aerial.—R. E. Burgess. (*Wireless Engr*, June 1946, Vol. 23, No. 273, pp. 172–178.) " The complex effective permeability of a mass core is expressed in terms of the relevant factors, and the imaginary part is related to the eddy current loss in the particles, which should predominate over other components of loss.

" The increase of pick-up due to a spheroidal core is calculated and it is shown that the core should be elongated in a direction parallel to the axis of the loop. The effect of a hollow spheroidal core is discussed and it is found that in a typical case 80 per cent of the iron can be removed before the increase of pick-up is halved ; the effect of spacing the winding from the core is treated approximately.

" Recommendations are made regarding the design for maximum sensitivity."

An editorial comment (G. W. O. H.) appears in the same journal, pp. 156–157.

621.396.677 2486

Radar Technique.—W. T. C. (*Wireless World*, May 1946, Vol. 52, No. 5, pp. 151–154.) A review of papers on waveguide and aerial techniques presented at the I.E.E. Radiolocation Convention.

621.396.677 2487

A Current Distribution for Broadside Arrays which Optimizes the Relationship between Beam Width and Side-Lobe Level.—C. L. Dolph. (*Proc. Inst. Radio Engrs, W. & E.*, June 1946, Vol. 34, No. 6, pp. 335–348.) " A one-parameter family of current distributions is derived for symmetric broadside arrays of equally spaced point sources energized in phase. For each value of the parameter, the corresponding current distribution gives rise to a pattern in which (1) all the side lobes are at the same level ; and (2) the beam width to the first null is a minimum for all patterns arising from symmetric distributions of in-phase currents none of whose side lobes exceeds that level."

Design curves expressing both the value of the parameter and the relative current values as functions of side-lobe level are given for the cases of 8-, 12-, 16-, 20-, and 24-element linear arrays.

621.396.677 2488

Long-Wire Antennas.—W. V. B. Roberts. (*QST*, June 1946, Vol. 30, No. 6, pp. 36–39.) A simplified qualitative treatment of the operation of rhombic and V aerials. The power gains of the rhombic and the half-wave dipole are compared.

621.396.677 + 621.396.61].029.63 2489

CQ 2 400 Mc/s : Transceivers and Antennas for the 13-Centimeter Band.—Koch & Floyd. (*See* 2758.)

621.396.677.029.64 2490

Radio Lenses.—W. E. Kock. (*Bell Lab. Rec.*, May 1946, Vol. 24, No. 5, pp. 193–196.) The phase velocity of a radio wave propagated between parallel metal plates is greater, on waveguide principles, than the velocity of propagation in free space. A pile of equally spaced parallel plates therefore acts like a block of material with refractive index less than that of free space. Converging lenses have been made by shaping the edges of the plates in such an array to the profile of a concave lens. A general description of th principle and illustrations of lenses are given Beams 0.1° wide have been obtained. Othe possible applications of the principle are mentioned

621.392 249

Problèmes de Propagations Guidées des Onde Électromagnétiques. [Book Review]—L. de Broglie Gauthier-Villars, Paris, 1941, 160 fr. (*Wireles Engr*, June 1946, Vol. 23, No. 273, p. 171.) " . . . carefully prepared review of the subject. . ."

CIRCUITS

621.3.017 249

Loss due to Shunt or Series Resistance Inserte between Matched Source and Sink.—(*Radio, N.Y.* April 1946, Vol. 30, No. 4, p. 38.) A chart giving the loss as a function of the ratio of the shunt o series resistance to the load resistance.

621.314.2 249

Equivalent Capacitances of Transformer Windings —W. T. Duerdoth. (*Wireless Engr*, June 194 Vol. 23, No. 273, pp. 161–167.) " The paper show that the distributed capacitances between windings or windings and screens, of transformers may b represented by lumped capacitances provided tha the magnetic coupling between the turns of winding is perfect. Expressions have been obtaine for the equivalent capacitances of a number different arrangements including windings i layers, sections, and with screens."

621.316.722.078.3 : 621.392.5 249

The Theory of the Non-Linear Bridge Circuit a Applied to Voltage Stabilizers.—G. N. Patchet (*J. Instn elect. Engrs*, Part I, April 1946, Vol. 9 No. 64, pp. 189–190.) Long summary of 867 April.

621.316.974 : 621.318.4.017.31 249

Power Loss in Electromagnetic Screens.—Sioco (*See* 2716.)

621.318.572 249

Design and Use of Directly Coupled Pentod Trigger Pairs.—V. H. Regener. (*Rev. sci. Instrum* May 1946, Vol. 17, No. 5, pp. 180–184.) Discussio of a trigger circuit using two pentodes with dire plate to screen intercoupling. Comprehensive cha acteristic curves are given for a typical trigg using 6AK6 pentodes, showing the effect of biasin either the control or suppressor grids of one both valves. Circuits for a pulse generator an an electronic switch are given, and the limits input for successful operation are deduced fro the curves. Another circuit, in which each su pressor is capacitively coupled to the screen the same valve, may be used to obtain triggeri with pulses of one sign only.

Scaling circuits up to scale of eight are brief mentioned.

621.318.572 249

Decade Counting Circuits.—V. H. Regene (*Rev. sci. Instrum.*, May 1946, Vol. 17, No pp. 185–189.) A simple ring-of-ten counter d signed round the directly coupled pentode trigg discussed in 2496 above. The essential characte istic of the circuit is that it has ten possible equi brium conditions. Two detailed circuits are give one of which will count sharp pulses up to requency of 10^5 c/s. The other will do the same or impulses of arbitrary shape and frequency. The number of pulses counted by each ring of ten may be indicated by the position of the spot on a athode-ray tube. Multiplicity of circuits and ubes enables decimal counting to be obtained to ny required number.

21.392.43 : 621.365.92 **2498**

Coupling Method for Dielectric Heating.—R. C. Kleinberger. (*Electronic Industr.*, June 1946, Vol. , No. 6, pp. 78–79.) The necessary impedance atching to obtain maximum power transfer from ansmission line to load can be most conveniently btained by the use of adjustable stubs. Procedure nd equations are given whereby stubs may be esigned to effect approximate tuning of the load npedance and matching of the transmission line. inal adjustments are determined by actual trial.

21.392.5 + 621.395.665 **2499**

Radio Design Worksheet: No. 47—Bridged and H Attenuators; Diode Conduction.—(*Radio, N.Y.*, April 1946, Vol. 30, No. 4, pp. 36–37.)

1.392.5 **2500**

Solving 4-Terminal Network Problems Graphically: Part 2.—R. Baum. (*Communications,* ay 1946, Vol. 26, No. 5, pp. 40..53.) Further scussion of the Smith diagram and inversion arts, and an illustration of the technique by e solution of a problem containing tuned circuits, sistances, and lines. For part 1 see 1786 of ly.

1.392.5 **2501**

Determination of a Class of Coupled Circuits th n Degrees of Freedom, having the same Natural equencies as a Given Assemblage of Coupled cuits.—M. Parodi. (*C. R. Acad. Sci., Paris,* th Jan. 1946, Vol. 222, No. 5, pp. 281–283.) mathematical paper. It derives formally, by trix methods, the values of the circuit elements C and R) for all members of a class of coupled cuits having the same natural frequencies as a en assembly of such circuits. The demonation depends on the fact that the determinant a product of matrices is equal to the product the determinants of the matrices.

.392.52 **2502**

Preferred Numbers and Filter Design.—P. nchet: H. Jefferson. (*Wireless Engr,* June 46, Vol 23, No. 273, p. 179.) A comment, in nch, on 3823 of 1945 (Jefferson) with Jefferson's ly. See also 871 of April (Jefferson).

.392.52 **2503**

Filter Design Tables Based on Preferred Numbers: gh-Pass Filters.—H. Jefferson. (*Wireless Engr,* y 1946, Vol. 23, No. 274, pp. 197–199.) Tables given for the design of constant-k high-pass ers of T- or π-sections. See also 871 of April 3823 of 1945 (Jefferson).

.392.52 **2504**

Tunable Rejection Filter.—R. C. Taylor. *ans. Amer. Inst. elect. Engrs,* May 1946, Vol. 65, 5, pp. 263–267.) The theory and design of a dge-type narrow-band filter. Range of adjustnt and effect of component variations are dissed with the aid of impedance circle diagrams. proximate formulae for attenuation and bandth are deduced.

621.392.52 **2505**

Transmission Lines as Filters.—Quarles. (*See* 2480.)

621.394/.396].645 **2506**

Radio Design Worksheet: No. 48—Reactive Feedback Factors.—(*Radio, N.Y.*, May 1946, Vol. 30, No. 5, p. 24.) Analysis of reactive feedback in an amplifying stage having an elliptical load line.

621.394/.397].645.2 **2507**

The Cathode-Coupled Amplifier.—K. A. Pullen, Jr. (*Proc. Inst. Radio Engrs, W. & E.*, June 1946, Vol. 34, No. 6, pp. 402–405.) Further applications of the double-triode cathode-coupled circuit previously described by Sziklai & Schroeder (3811 of 1945), including its use as h.f. amplifier, multivibrator, a.f. and r.f. oscillators, resonant-resistance meter, and mixer circuit. Design data, including gain characteristics for a typical valve (6SN7), are given. See also 2157 of August (Crosby).

621.394/.397].645.2 **2508**

Wide-Band Amplifiers: Part 3.—(*Wireless World,* May 1946, Vol. 52, No. 5, pp. 161–162.) An analysis of band-pass coupling by critically coupled equally damped circuits, showing that the arrangement gives less gain for the same bandwidth than that of stagger-tuned circuits described in 1789 of July.

621.394/.397].645.29 **2509**

An Analysis of Cascode Coupling.—R. G. Middleton. (*Radio, N.Y.*, June 1946, Vol. 30, No. 6, pp. 19, 32.) A graphical analysis of a "cascode" amplifier based on the family of plate current curves. In this circuit the signal is applied equally to the grids of two amplifying valves, of which the cathode of one is connected directly to and in series with the anode of the other. The basic d.c. amplifier may be adapted for a.c.

621.394.645.35 : 621.317.715 **2510**

A Contact Modulated Amplifier to Replace Sensitive Suspension Galvanometers.—Liston, Quinn, Sargeant & Scott. (*See* 2629.)

621.394.645.35 : 621.383 **2511**

Direct-Current Amplifier for a Photocell with Low Insulation Resistance or Large Dark-Current.—J. Dubois. (*C. R. Acad. Sci., Paris,* 28th May 1945, Vol. 220, No. 22, pp. 768–770.) Note on a modification of the input circuit to a B405 amplifier valve that gives a substantial improvement of sensitivity by maintaining the operating point on the linear part of the characteristic.

621.395.645 + 621.396.621 **2512**

Superamp with Tuner.—C. G. Brennan. (*Radio Craft,* May 1946, Vol. 17, No. 8, pp. 539, 563.) Details of a high fidelity a.f. amplifier and superheterodyne receiver.

621.395.645 **2513**

Additional Notes on the Parallel Tube Amplifier.—Jones: Velasco. (*See* 2464.)

621.395.645 **2514**

Negative Feedback and Hum.—"Cathode Ray". (*Wireless World,* May 1946, Vol. 52, No. 5, pp. 142–145.) A series of experimental results is given for typical triode and pentode amplifier stages. It is concluded that (*a*) feedback from the anode, when the load is transformer coupled, is generally bad practice unless the h.t. supply is very smooth; (*b*) with triodes, when feedback is generally not used, parallel feed should be used; (*c*) with tetrodes and pentodes, freedom from hum is obtained by smoothing the screen supply, or by use of feedback; (*d*) a transformer-coupled pentode is remarkably humfree without feedback. See also 1477 of June (Builder).

621.395.645 : 534.43 **2515**
Unique Phono Amplifier.—C. E. Pett, Jr. (*Radio News*, May 1946, Vol. 35, No. 5, pp. 50..90.) Constructional details of an amplifier with circuits for bass and treble boost and for contrast expansion.

621.395/.397].645 : 621.314.25 **2516**
Phase Inverters.—H. A. Bustard. (*Radio News*, Feb. 1946, Vol. 35, No. 2, pp. 57..104.) Circuit diagrams of nine different types, with a detailed discussion of their design and operation.

621.395.645 : 621.317.733 **2517**
A Convenient Amplifier and Null Detector.—Scott & Byers. (*See* 2634.)

621.395.645 : 621.395.665.1 **2518**
A Volume Expander Compressor Preamplifier.—R. C. Moses. (*Radio News*, June 1946, Vol. 35, No. 6, pp. 32..149.) Constructional details of a preamplifier with a maximum overall gain of 110 db. The time delay of the automatic gain control can be adjusted to give a minimum rise time of 3 milliseconds, and maximum decay time of 500 milliseconds.

621.396.6.018.1 **2519**
Phase Relationships.—M. G. Scroggie: J. H. Barrett. (*Wireless World*, May 1946, Vol. 52, No. 5, pp. 170–171.) Critical discussion of 1794 of July (Cooper).

621.396.61 **2520**
Tuned Circuits for the U.H.F. and S.H.F. Bands.—F. C. Everett. (*Communications*, June 1946, Vol. 26, No. 6, pp. 19–21, 51.) A review of fixed and variable tuned circuits including a more detailed description of a symmetrical wide-band cylindrical circuit similar to those described in 1797 of July (Gross).

621.396.611 : 536.7 **2521**
Boltzmann's Law of Slow Transformation and the Theory of Electromagnetic Cavities.—T. Kahan. (*C. R. Acad. Sci., Paris*, 2nd Jan. 1946, Vol. 222, No. 1, pp. 70–71.) Derivation of a general formula for the Q of a cavity which "leads to new methods of determining dielectric constants, multiplication factors, and magnetic permeabilities at hyperfrequencies."

621.396.615 **2522**
A Study of Locking Phenomena in Oscillators.—R. Adler. (*Proc. Inst. Radio Engrs, W. & E.*, June 1946, Vol. 34, No. 6, pp. 351–357.) "Impression of an external signal upon an oscillator of similar fundamental frequency affects both the instantaneous amplitude and the instantaneous frequency. Using the assumption that time constants in the oscillator circuit are small compared to the length of one beat cycle, a differential equation is derived which gives the oscillator phase as a function of time. With the aid of this equation, the transient process of 'pull-in' as well as the production of a distorted beat note are described in detail.

"It is shown that the same equation serves to describe the motion of a pendulum suspended in a viscous fluid inside a rotating container. The whole range of locking phenomena is illustrated with the aid of this simple mechanical model."

621.396.615 **2523**
Notes on the Stability of LC Oscillators.—N. Lea. (*J. Instn elect. Engrs*, Part I, May 1946, Vol. 93, No. 65, pp. 235–236.) Summary of 569 of March.

621.396.615 : 621.396.611.21 + 621.317.361 **2524**
Series-Resonant Crystal Oscillators.—F. Butler. (*Wireless Engr*, June 1946, Vol. 23, No. 273, pp. 157–160.) Most crystal oscillators use the crystal in the parallel-resonant mode. Quartz crystals however possess a series resonant mode that has the advantage of somewhat higher constancy of frequency. The frequency, in the series mode, is unaffected by changes in parallel reactance (*e.g.* holder capacitance) but is affected by changes in series reactance. Circuits of the Hartley type are described in which the crystal is connected between the cathode of the valve and the centre of the oscillatory coil.

621.396.615.17 **2525**
Kinematic Definition of Relaxation Oscillations.—J. Abelé. (*C. R. Acad. Sci., Paris*, 9th April 1945, Vol. 220, No. 15, pp. 511–513.) Van der Pol (1930 Abstracts, p. 503) defined relaxation oscillations in terms of a nonlinear second-order differential equation. The present author calls this a "dynamic" definition, and proposes a "kinematic" definition analogous to the definition of sinusoidal oscillations as the projection of a circular motion. Consideration is given to the projection of the end of a uniformly rotating vector on an axis that oscillates according to a fixed law relative to the rotation. "The curves so obtained are analogous to those of van der Pol who used a more laborious and less accurate method of graphical integration." A more detailed account is to appear elsewhere.

621.396.615.17 **2526**
Linear Saw-Tooth Oscillator.—W. T. Cocking. (*Wireless World*, June 1946, Vol. 52, No. 6, pp. 176–178.) A modification of the transitron time base, operating with a single pentode, and essentially a combination of the pre-war transitron and the Miller integrator developed during the war. The control grid, cathode, and anode are used for the linearizing action, and the screen and suppressor grids are resistance-capacitance coupled to give a transitron type of circuit, thus providing a self-oscillating linear timebase.

621.396.619.16 : 621.396.9 **2527**
Radar Technique.—W. T. C. (*Wireless World*, May 1946, Vol. 52, No. 5, pp. 156–158.) Review of papers on pulse circuits presented at the I.E.E. Radiolocation Convention.

621.396.645 **2528**
Intermediate Frequency Amplifier Stability Factors.—D. L. Jaffe. (*Radio, N.Y.*, April 1946, Vol. 30, No. 4, pp. 26–27, 55.) The stability is determined by the plate-grid capacitance of the individual tubes, wiring, overall gain, and coupling between the input and output. The last is important for amplifiers with gains in excess of 80 db. It is

shown, by considering the phase of the feedback current, that the maximum anode impedance for stability is $\sqrt{(2/g_m\omega_0C_{gp})}$ for a single tuned circuit, and $\sqrt{(2g_m/\omega_0C_{gp})}$ is the maximum stable gain. The corresponding figures for double tuned circuits critically coupled are $\sqrt{(0.79/g_m\omega_0C_{gp})}$ and $\sqrt{(0.79g_m/\omega_0C_{gp})}$. Values for the latter are shown in graphs against frequency for a number of commonly used valves.

621.396.645.3.029.58 **2529**

Long Leads aren't Necessary.—Shuart. (*See* 2772.)

621.396.66 **2530**

Clamping Circuits.—J. McQuay. (*Radio Craft*, May 1946, Vol. 17, No. 8, pp. 541..561.) "A clamping circuit maintains either the positive extreme or the negative extreme of a waveform within the limits of a desired reference level of voltage."

621.396.662.2.029.6 **2531**

V.H.F. Coil Design.—Meyerson. (*See* 2732.)

621.396.667 **2532**

Low-Frequency Correction Circuit.—(*Wireless World*, June 1946, Vol. 52, No. 6, pp. 199–200.) Design of a circuit giving a rising response characteristic at the lower frequencies, such as is required for gramophone reproduction: the circuit also gives a small amplification.

621.396.667 **2533**

Tone Correction.—L. Gregory. (*Wireless World*, June 1946, Vol. 52, No. 6, p. 204.) Brief description of a circuit combining bass and treble boost with negative feedback.

621.397.645.2 : 621.396.621.54 **2534**

I.F. Amplifiers in Television Receivers.—M. H. Kronenberg. (*QST*, June 1946, Vol. 30, No. 6, pp. 62–65.) Circuits are given for two 12.75–Mc/s amplifiers with 2.5–Mc/s and 4–Mc/s bandwidth respectively, having attenuation at the 8.25–Mc/s sound channel. Design formulae are discussed.

621.392 **2535**

Heaviside's Electric Circuit Theory. [Book Review]—H. J. Josephs. Methuen & Co., London, 115 pp., 5s. 6d. (*Wireless Engr*, July 1946, Vol. 23, No. 274, p. 200.) One of the Monographs on Physical Subjects. "...a valuable addition to the Heaviside literature."

GENERAL PHYSICS

535.343.4 + 621.317.1.011.5 + 621.396.11.029.64] : 546.171.1 **2536**

Ammonia Spectrum in the 1 cm Wavelength Region.—Bleaney & Penrose. (*See* 2662.)

535.43 : 537.122 **2537**

On the Theory of the Scattering of Light on Free Electrons.—M. Al'perin. (*Zh. eksp. teor. Fiz.*, 1944, Vol. 14, Nos. 1/2, pp. 3–13.) The existing methods for studying the problem are valid for small intensities of the incident wave only. It is possible, however, by choosing suitable variables to find an exact solution of the Dirac equations for an electron in the field of a plane wave. This is done in the present paper, and the solution found (17) is used to derive a formula (47) similar to the one obtained by Klein & Nishina (1929 Abstracts p. 588) but applicable to large intensities of the incident radiation and taking into account the possibility of a simultaneous absorption of several quanta. The results obtained are discussed in the light of quantum electrodynamics.

535.5 **2538**

A Graphical Method for determining the Refractive Index and Thickness of Thin Films.—I. Obreimoff. (*Zh. eksp. teor. Fiz.*, 1944, Vol. 14, Nos. 10/11, pp. 431–438.) It is assumed that the film has a uniform thickness h and a constant refractive index n_2, and that a plane wave falls on the film at an angle i_1. If, under these conditions, E_p and E_s denote respectively the component of the electric vector in the plane of incidence and the component perpendicular to it, then $E_p{}^r$ and $E_s{}^r$ in the reflected wave will be reflected differently, *viz.*, $E_p{}^r = \rho_p E_p$ and $E_s{}^r = \rho_s E_s$. Moreover there will be a phase difference δ between these components. A system of equations (2) determining the relationship between ρ_s, ρ_p and δ was derived by Vlasoff who also pointed out that if γ and δ are determined experimentally ($\rho_p \rho_s =$ tan γ), then h and n_2 can be calculated from equations (2). With the many measurements required, however, the calculations would be too laborious, and a number of nomograms are given in order to simplify these, as well as those required in the measurements of γ and δ. Numerical examples are worked out, and the accuracy of the method is estimated.

536.7 : 621.396.611 **2539**

Boltzmann's Law of Slow Transformation and the Theory of Electromagnetic Cavities.—Kahan. (*See* 2521.)

537.226 **2540**

The Theory of the Polarization of Dipole Liquids in Strong Electric Fields.—A. Anselm. (*Zh. eksp. teor. Fiz.*, 1944, Vol. 14, No. 9, pp. 364–369.) It was shown by the author in previous papers (*Zh. eksp. teor. Fiz.*, 1942, Vol. 12, p. 274, & 1943, Vol. 13, p. 432) that the theory of the inner field proposed by Debye for interpreting the polarization of dipole liquids in weak fields is incorrect. The same considerations also apply to Debye's theory of polarization in strong fields. In the present paper the author develops a new theory from the method used by Kirkwood (*J. Chem. Phys.*, 1939, Vol. 7, p. 911) investigating the polarization in weak fields. It is pointed out, however, that Kirkwood, having derived formula (1) for determining the permittivity ϵ in a weak field has attempted to calculate M_∞, the electric moment appearing in an infinite dielectric with a fixed orientation of one of its molecules. Such attempts are bound to fail with the present state of knowledge of intramolecular forces in a liquid, so the author proposes to treat M_∞ as a parameter which characterizes the molecular interaction, and which can be determined experimentally. Accordingly, a formula (14) is derived for calculating the dielectric constant ϵ' in strong fields. It is possible to check the new theory experimentally by considering other phenomena determined quantitatively by M_∞. Thus it is shown that the value of $(\epsilon' - \epsilon)/\epsilon$ depends on M_∞. A comparison between the theoretical and experimental values of the ratio for water and nitrobenzene indicates that the theoretical results are of the correct order of magnitude.

537.312.62 2541

Notes on the Theory of Superconductivity.—V. Ginsburg. (*Zh. eksp. teor. Fiz.*, 1944, Vol. 14, No. 5, pp. 134–151.) The theory is discussed in the light of the latest experimental and theoretical investigations under the following headings: (*a*) main properties of superconductivity; (*b*) phenomenological electrodynamics; (*c*) microscopic aspect of superconductivity; (*d*) energy spectrum and properties of the electron liquid; (*e*) statistical and some other properties of superconductors.

The main conclusion reached is that the theory of superconductivity is closely associated with the electron theory of metals in a normal state. Efforts therefore should be directed towards further development of the latter theory, but on the basis of the electron liquid concept, *i.e.* without using the electron gas hypothesis.

A list of 16 references is given.

537.525 : 535.34 2542

On the Absorption of Light by a Plasma.—A. Kompaneyets. (*Zh. eksp. teor. Fiz.*, 1944, Vol. 14, No. 6, pp. 171–176.) It is known that free electrons do not absorb light. It would therefore appear that a completely ionized gas at a sufficiently high temperature would have a very low absorption coefficient. To verify this a mathematical investigation of the propagation of electromagnetic oscillations in a plasma is presented. It is shown that owing to the forces acting between the electrons and the positive ions of the plasma the latter possesses a considerable absorption coefficient. This absorption, as distinct from the photoelectric absorption, does not decrease with the frequency of the light wave and the temperature of the plasma. A formula (37) is derived determining the absorption coefficient, and methods are indicated for carrying out the necessary calculations.

538.222 : 538.56 2543

A New Method for Investigating Paramagnetic Absorption.—S. Altschüler, E. Zavoiski & V. Kozirev. (*Zh. eksp. teor. Fiz.*, 1944, Vol. 14, Nos. 10/11, pp. 407–409.)

538.3 2544

Electromagnetic Field Equations for a Conducting Medium with Hysteresis.—M. Rozovski. (*Zh. eksp. teor. Fiz.*, 1944, Vol. 14, Nos. 10/11, pp. 402–406.) In Maxwell's equations, it is usually assumed that $B(z, t) = \mu H(z, t)$. If, however, the magnetic lag is taken into account, the latter formula must be so modified as to reflect the dependence of $B(z, t)$, not only on the value of $H(z, t)$ at the given moment, but also on the states of $H(z, t)$ preceding this moment. Using the relationship (3) between $B(z, t)$ and $H(z, t)$, introduced by Volterra, an integral-differential equation (11) of a more general character is derived from Maxwell's equations. It is shown that this equation can be solved by the Fourier method.

538.31 2545

Two Electromagnetic Problems.—G. W. O. H. (*Wireless Engr*, July 1946, Vol. 23, No. 274, pp. 181–182.) If a current-carrying solenoid, placed in a magnetic field from a source remote from the solenoid, is reversed in direction, the reduction of magnetic energy within the solenoid is balanced by the increase outside it, and all the work done against the field appears as energy in the electric circuit.

538.31 2546

On the Parametric Oscillations of an Iron Body in an Alternating Magnetic Field.—S. Rytoff. (*Zh. eksp. teor. Fiz.*, 1944, Vol. 14, No. 9, pp. 370–378.) An electromechanical system is considered consisting of a circular loop and an iron ball capable of moving along the axis of the loop. If a current is passed through the loop the ball will be attracted into the loop, and the coefficient of elasticity of the system for a current of frequency ω will vary between zero and maximum values with a frequency of 2ω. This is therefore an oscillating system with a periodically variable parameter determining its natural frequency, and, as is known, the equilibrium of such a system may become unstable under certain conditions.

A mathematical analysis of the system is offered, and equation (4) determining the appearance of oscillations is derived. It is shown that there are discrete regions of instability which can be reached by varying the strength of the loop current. A detailed description of experiments is given in which the following two types of oscillations were observed: (*a*) oscillations at the current frequency with small amplitudes and only slight nonlinearity; and (*b*) oscillations at fractional current frequencies with large amplitudes and a strongly pronounced nonlinearity. The parametric interaction between the loop current and eddy currents in the ball is also briefly discussed.

538.32 : 621.385.832 2547

Problem of Two Electrons.—R. E. Burgess: G. W. O. H. (*Wireless Engr*, June 1946, Vol. 23, No. 273, p. 178.) Discussion of 913 of April (G. W. O. H.). Burgess points out that the apparent paradox in the problem of two electrons is easily resolved by application of the principle of special relativity. G. W. O. H. replies editorially in the same journal, pp. 155–156. See also 587 of March (Tripp).

538.56 : 517.948.3 2548

The Boundary Problem of Electrodynamics and Integral Equations of Certain Diffraction Problems.—Ya. Feld. (*Zh. eksp. teor. Fiz.*, 1944, Vol. 14, No. 9, pp. 330–341.) In a number of problems of electrodynamics it is required to determine the electromagnetic field set up by given exciters in a space bounded by metallic surfaces. Problems of this type can be reduced to the following: it is required to find, in a space v bounded by a surface s, a field with the tangential component of the electric vector vanishing at the surface s. In the present paper the case of harmonic oscillations only is considered, and a solution (1) of the problem is derived. The results obtained can be used to reduce some of the problems of electromagnetic diffraction to Fredholm's integral equations of the first kind. This is shown in a number of examples dealing with the diffraction of electromagnetic waves at an aperture in an infinite plane. Methods for solving the equations so derived are also indicated.

541.133 : [621.3.029.5/.6 2549

The Variation of the Electrical Conductivity of Electrolytes with Frequency.—N. Maloff. (*Zh. eksp. teor. Fiz.*, 1944, Vol. 14, No. 6, pp. 221–223.) In a previous investigation (2682 of 1940) into the electrical conductivity of highly concentrated solutions of sodium chloride in water (up to 0.3 mol/litre) the author found a considerable decrease in the conductivity at high frequencies. In the present paper a mathematical analysis of the phenomenon is proposed based on a study of Belikoff (*Zh. eksp. teor. Fiz.*, 1939, Vol. 9, p. 969) of the movement of ions in electrolytes and on the conductivity of electrolytes at low frequencies. To simplify the discussion it is limited to the case of a symmetrical binary monovalent electrolyte, and it is shown that Belikoff's equations when extended to higher frequencies indicate a fall in the conductivity. In the case of low concentrations this becomes apparent at frequencies within the range of centimetre waves, while with high concentrations the effect begins to take place at frequencies of the order of 10^6 c/s. A physical interpretation of the results obtained is also given.

621.314.6 : 621.383.2 **2550**

Experimental Behaviour of a Photoelectric Cell under the Influence of an Alternating Potential of Very High Frequency.—Charles. (*See* 2725.)

621.314.6 : 621.383.2 **2551**

Theory of the Behaviour of a Photoelectric Cell under the Influence of an Alternating Potential of Very High Frequency.—Charles. (*See* 2726.)

621.314.63 + 621.315.34 **2552**

An Anomalous Phenomenon in Thermal Rectification in Lead Sulphide (Preliminary Communication).—Kh. Amirkhanoff. (*Zh. eksp. teor. Fiz.*, 1944, Vol. 14, No. 6, pp. 193–194.) Experiments were conducted with lead sulphide obtained chemically in the form of a black powder. Samples at room temperature and a pressure of 10 000 kg/cm² possessed a hole-type conductivity, and a resistivity of 5.10^5 Ω cm. After heat treatment at 200–300°C, depending on the duration of the treatment, the resistivity decreased, and the hole-type conductivity was replaced by that of the electron type. In one sample, however, conductivity of the electron type changed again to the hole type after a current had passed for 30–60 sec. The phenomenon also occurred when the direction of the temperature gradient was changed. No residual polarization or other effects were observed. A table of the experimental data is given.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.7 + 525.24] : 551.51.053.5 **2553**

The Application of Solar and Geomagnetic Data to Short-Term Forecasts of Ionospheric Conditions.—A. H. Shapley. (*Terr. Magn. atmos. Elect.*, June 1946, Vol. 51, No. 2, pp. 247–266.) The ways in which recurrence tendencies of geomagnetic activity, reports of solar activity, and various solar-terrestrial relationships are used in preparing forecasts at the Dept. of Terrestrial Magnetism, Carnegie Institution of Washington. Forecasts are compared with magnetic activity over a 15-month period, and are 70% satisfactory. Analysis of coronagraphic and spectro-heliographic data with magnetic activity show that for two years there was a decided tendency for disturbances to occur when solar regions identified by these observations were east of the central meridian of the Sun. A minimum in solar activity occurred early in 1944 as indicated by reduction of solar and geomagnetic data.

Recorders have been constructed which give an instantly visible record of variation of the earth's magnetic field.

"Solar-geomagnetic relationships are still too general to be the sole factor in detailed forecasts. The manifestation, if any, of the solar cause of geomagnetic disturbance has not yet been found."

523.7 + 525.24] " 1946.01/.03 " **2554**

Solar and Magnetic Data, January to March 1946, Mount Wilson Observatory.—S. B. Nicholson & E. S. Mulders. (*Terr. Magn. atmos. Elect.*, June 1946, Vol. 51, No. 2, pp. 284–286.)

523.72 **2555**

The Solid Angle of the Corpuscular Solar Radiation.—M. N. Gnevishev & A.I.Ol. (*Terr. Magn. atmos. Elect.*, June 1946, Vol. 51, No. 2, pp. 163–170.) The solid angle may be evaluated by using data on (*a*) duration of magnetic storms; (*b*) equinoctial increase of geomagnetic activity; (*c*) correlation coefficients between magnetic activity and sunspot area in the central zone with different radii; (*d*) the lag of the 11-year variation of geomagnetic activity.

The researches indicate an angle of from 8° to 9°.

523.72 **2556**

A Theoretical Discussion of the Continuous Spectrum of the Sun.—G. Münch. (*Astrophys. J.*, Nov. 1945, Vol. 102, No. 3, pp. 385–394.) Experimental data are examined in relation to the theory of radiative equilibrium. The intensity distribution and the law of darkening in the different wavelengths can be explained in terms of the absorption coefficient. The required variation of this coefficient in the visual and near infra-red regions of the spectrum is also explained.

523.72 : 525.24 **2557**

Geomagnetic Data on Variations of Solar Radiation: Part 1 — Wave-Radiation.—J. Bartels. (*Terr. Magn. atmos. Elect.*, June 1946, Vol. 51, No. 2, pp. 181–242.) Homogeneous time-series for W (a wave type of solar radiation) and P (a particle type) are derived from magnetic observations. Daily values for the deviations of W from a normal value are inferred from the variation of the horizontal intensity. These are compared with tables derived for the solar activity R and radiation P, the correlation for " slow " variations between R and W being well marked. " Fast " variations in R due to solar rotations are followed after a time lag by similar variations in W. The physical meaning of W and its extraction from geomagnetic records are discussed.

523.746 " 1945 " **2558**

Final Relative Sunspot-Numbers for 1945.—M. Waldmeier. (*Terr. Magn. atmos. Elect.*, June 1946, Vol. 51, No. 2, pp. 267–269.)

523.746 "1946.01/.03" **2559**

Provisional Sunspot-Numbers for January to March, 1946.—M. Waldmeier. (*Terr. Magn. atmos. Elect.*, June 1946, Vol. 51, No. 2, p. 274.)

523.746.5 **2560**

A Prediction of the Next Maximum of Solar Activity.—M. Waldmeier. (*Terr. Magn. atmos. Elect.*, June 1946, Vol. 51, No. 2, p. 270.) The maximum should be expected to take place as early as 1947.6. A table gives the smoothed monthly relative numbers for the epoch two years before to five years after the maximum.

523.78 : 525.23 2561
Atmospheric-Electric Potential-Gradient in Kokkola, Finland, during the Solar Eclipse of July 9, 1945.—E. Sucksdorff. (*Terr. Magn. atmos. Elect.*, June 1946, Vol. 51, No. 2, pp. 171–176.) Measurements were made using a bifilar electrometer and an ionium-collector. The eclipse caused a marked and smooth diminution of the potential gradient which began two hours prior to the beginning of the eclipse and continued even after the end of the visual eclipse.

523.78 : 551.51.053.5 : 621.396.11 2562
The Influence of an Eclipse of the Sun on the Ionosphere. — R. L. Smith-Rose. (*J. Brit. Instn Radio Engrs*, June 1946, Vol. 6, No. 3, pp. 82–97.) A survey of the structure of the ionosphere and its effects on propagation. The information gained from radio observations during solar eclipses is outlined. Experimental evidence shows that the main source of ionization is ultra-violet radiation from the sun; the possible contribution in the F_2 region from incident neutral particles is an open question. A more complete understanding is required of long-distance transmission, particularly at very low frequencies. A bibliography of 20 items is given.

533.6.013.22 2563
On Atmospheric Turbulence.—A. M. Obukhov. (*J. Phys., U.S.S.R.*, 1942, Vol. 6, No. 5, pp. 228–229.) Abstract of a paper of the Acad. Sci., U.S.S.R.

550.38 2564
Geomagnetic Secular Variations and Surveys.— J. A. Fleming. (*Proc. phys. Soc.*, 1st May 1946, Vol. 58, No. 327, pp. 213–246.)

550.38 "1945.07/.09" 2565
Five International Quiet and Disturbed Days for July to September, 1945.—W. E. Scott. (*Terr. Magn. atmos. Elect.*, June 1946, Vol. 51, No. 2, p. 284.)

550.38 : 523.75 2566
Relations between Magnetic Disturbances and Solar Eruptions.—M. Burgaud. (*C. R. Acad. Sci., Paris*, 18th Feb. 1946, Vol. 222, No. 8, pp. 449–450.) Observations lead to the following conclusions: (*a*) magnetic storms do not depend on the size or growth of sunspots and faculae, but on violent eruptions; (*b*) a storm can be caused by an eruption on any part of the solar disk, but maximum effects are associated with eruptions near the central meridian; (*c*) magnetic disturbances have taken place in the absence of visible sunspots; they follow eruptions.

550.385 "1945.10/1946.03" 2567
Principal Magnetic Storms [reported from various observatories].—(*Terr. Magn. atmos. Elect.*, June 1946, Vol. 51, No. 2, pp. 287–301.)

550.385 "1946.01/.03" 2568
Geomagnetic Storms.—(*Curr. Sci.*, May 1946, Vol. 15, No. 5, p. 146.) A brief review of the more intense storms recorded at the Alibag Magnetic Observatory in the period January to March 1946.

550.385 "1946.03.23/.29" 2569
Two Notable Geomagnetic Storms.—(*Nature, Lond.*, 6th April 1946, Vol. 157, No. 3988, p. 435. Reprinted *Terr. Magn. atmos. Elect.*, June 1946, Vol. 51, No. 2, pp. 283–284.)

550.385 "1946.03.28" 2570
Geomagnetic Storm at Elisabethville, March 28, 1946.—W. E. Scott. (*Terr. Magn. atmos. Elect.*, June 1946, Vol. 51, No. 2, pp. 281–283.)

551.51.053.5 "1940/1944" 2571
Annual Variation of the Values at Noon of the Critical Frequencies of the Ionized Layers at Tromsö during 1940, 1941, 1942, 1943, and 1944.—L. Harang. (*Terr. Magn. atmos. Elect.*, June 1946, Vol. 51, No. 2, pp. 275–277.) The monthly mean values show a continuous decrease for all three layers compared with those taken before 1940, the decrease being particularly pronounced for the F_2 layer.

LOCATION AND AIDS TO NAVIGATION

534.417 : 534.88 2572
Navy Releases Sonar Story.—(*See* 2452.)

551.576 : 621.383 2573
On Atmospheric [cloud-height] **Sounding.** — Barthélemy. (*See* 2657.)

621.396.674 : 621.318.323.2.029.5 2574
Iron-Cored Loop Receiving Aerial.—Burgess. (*See* 2485.)

621.396.677.1 2575
Aperiodic Combination of an Antenna and a Frame. Application in Direction-Finding to an Aperiodic Arrangement for Indicating Sense.— F. Carbenay. (*C. R. Acad. Sci., Paris*, 2nd Jan. 1946, Vol. 222, No. 1, pp. 63–64.) The potential difference at the terminals of a small resistance at the base of an aerial can be written $u = RCh\,dF/dt$ (R = resistance, C = aerial capacitance, h = effective height, F = vertical electrical intensity, t = time). The e.m.f. in the frame is $e = NS\,dH/dt$ (N = turns, S = area, H = horizontal magnetic intensity). The resultant of the two in series is zero or a maximum if $RChc = NS$ (c = velocity) which is a relationship independent of signal frequency. The arrangement is used in connexion with atmospheric locators.

621.396.9 2576
[I.E.E.] **Radiolocation Convention.**—(*Engineering, Lond.*, 29th March, 5th, 19th & 26th April 1946, Vol. 161, Nos. 4185, 4186, 4188 & 4189, pp. 306–308, 319–320, 367 & 389–390.) Summary of the proceedings. For other accounts see *Engineer, Lond.*, 29th March 1946, Vol. 181, No. 4707, pp. 296–297; *Radio, N.Y.*, May 1946, Vol. 30, No. 5, pp. 26..48; see also 1850 of July.

621.396.9 2577
Radar in War and Peace.—L. N. Ridenour. (*Elect. Engng, N.Y.*, May 1946, Vol. 65, No. 5, pp. 202–207.) The fundamental principles of radar, with brief notes on outstanding war-time achievements. The post-war possibilities of pulse navigation systems, radar beacons, and relay radar are discussed.

621.396.9 2578
Marine Radar for Peacetime Use.—L. H. Lynn & O. H. Winn. (*Trans. Amer. Inst. elect. Engrs*, May 1946, Vol. 65, No. 5, pp. 271–273.) Brief description of a commercial radar system for use on cargo vessels. Simplicity of operation for non-technical personnel is claimed.

621.396.9 2579
Radar for Blind Bombing : Part 2.—J. V. Holdam, S. McGrath & A. D. Cole. (*Electronics*, June 1946, Vol. 19, No. 6, pp. 142–149.) The conclusion of 2196 of August, giving details, with circuit diagrams, of the modulators, scanners, r.f. systems, receiver-indicator systems, and synchronizers in the APQ-13 and APS-15 versions of the H2X equipment.

621.396.9 2580
Airborne Radar for Navigation and Obstacle Detection.—R. C. Jensen & R. A. Arnett. (*Trans. Amer. Inst. elect. Engrs*, May 1946, Vol. 65, No. 5, pp. 307–313.) An account of wartime achievements as a basis for peacetime applications.

621.396.9 2581
Early Fire-Control Radars for [U.S.] **Naval Vessels.**—W. C. Tinus & W. H. C. Higgins. (*Bell Syst. tech. J.*, Jan. 1946, Vol. 25, No. 1, pp. 1–47.) An account of the development in the Bell Telephone Laboratories of equipments "Mark I" to "Mark IV". The first operated at 680–720 Mc/s using 2-kW pulses of adjustable duration 1–5 μs, repetition rate 1 640 p/s, obtained from a pair of "doorknob" valves (see 126 of 1938—Samuel). The aerial was an array of 8 half-wave dipoles in line with a parabolic cylinder reflector 6 ft square, giving a beam width of 12° and gain of 22 db. The receiver operated with 30-Mc/s i.f., 1-Mc/s band-width, and had a noise factor of 24 db. The equipment gave range to an accuracy of about ± 200 yd up to 10 miles or more, and azimuth to 1–2°.

The Mk II radar was superseded by Mk III before production began. Aerial lobe-switching was used in Mk III to give azimuth determination and tracking to ± 15 ft. A new type of range presentation gave an accuracy better than the requirement of ± 50 yd. A magnetron (W. E. 701-A) for 700 Mc/s, based on the 3-kMc/s British cavity type, gave 40-kW, 2-μs pulses, and the receiver, using "light-house" tubes for r.f. amplification, had a noise factor of 9 db, so that there was a substantial improvement of range. A gas-discharge (t.r.) switch was used for aerial duplexing (see 2784).

The Mk IV set had two aerial arrays one above the other, and used lobe-switching in elevation as well as azimuth, but was otherwise similar to Mk III.

All types were first installed in 1941.

621.396.9 : 534.321.9 2582
Ultrasonic [radar] **Trainer Circuits.**—F. J. Larsen. (*Electronics*, June 1946, Vol. 19, No. 6, pp. 126–129.) A 15-Mc/s pulsed ultrasonic beam is projected in a water trough which has a special map made on its bottom surface with graduated surface roughening to represent terrain of different kinds. Sound waves reflected from the rough parts of the map are received by the crystal transducer and used to operate radar equipment. The beam is rotated, and the device is used to give p.p.i. presentation of the map. Auxiliary equipment is used to simulate aerial bombing runs. A general description is given, with details of some of the circuits.

621.396.9 : 621.317.79 2583
Techniques and Facilities for Microwave Radar Testing.—Green, Fisher & Ferguson. (*See* 2645.)

621.396.9 : 621.385.18 2584
The Gas-Discharge Transmit-Receive Switch.—Samuel, Clark & Mumford. (*See* 2783.)

621.396.9 : [621.396.11.029.64 + 538.569.4 2585
Radio Echoes from the Planets.—W. D. Hershberger. (*Science*, 22nd March 1946, Vol. 103, No. 2673, p. 371.) A note on the effect of absorption of microwaves by various gases. See also 1336 of May.

621.396.9 : 621.396.82 2586
Radar Countermeasures.—O. G. Villard, Jr. (*Proc. Radio Cl. Amer.*, March 1946, Vol. 23, No. 3, pp. 7–15.)

621.396.933 : 629.1.052 2587
Pulse-Type Radio Altimeter.—A. Goldman. (*Electronics*, June 1946, Vol. 19, No. 6, pp. 116–119.) General description and circuit details of a high-altitude altimeter, designated SCR-718-C, that operates on the radar principle. The sine-wave output from a crystal-controlled oscillator is clipped, differentiated, and amplified in pulse-generating circuits to form the modulating signal for the 440-Mc/s transmitter, which provides 0.25-μs signals of 5–10 W. The cathode-ray indicator has a circular sweep, and the signals received directly from the transmitter and after ground reflection produce radial deflexions. The height is given by the angular separation of the deflexions. 5 000-ft and 50 000-ft scales are provided. Accuracy 50 ft.

621.396.933.23 2588
Use of Microwaves for Instrument Landing : Parts 2 & 3.—D. F. Folland. (*Radio, N.Y.*, April & May 1946, Vol. 30, Nos. 4 & 5, pp. 23–25, 55 & 16–19.) Further details of the Sperry blind-landing system. For part 1 see 1866 of July.

MATERIALS AND SUBSIDIARY TECHNIQUES

531.788.7 2589
Ionization Gauge Control Unit.—A. H. King. (*J. sci. Instrum.*, April 1946, Vol. 23, No. 4, p. 85.) A device for maintaining constant filament emission.

533.5 2590
An Apparatus for Stirring under Vacuum.—B. R. Atkins. (*J. sci. Instrum.*, April 1946, Vol. 23, No. 4, p. 84.)

533.5 : 621.3.032.53 2591
Coppered-Tungsten Seals through Hard Glass.—A. L. Reimann. (*J. sci. Instrum.*, June 1946, Vol. 23, No. 6, pp. 121–124.) The fine longitudinal cracks in tungsten wire which cause air leaks may be filled by plating the wire with Cu, with or without an added layer of Ni, and fusing the coating to the wire in a hydrogen furnace. The wire is then plated further with Cu and sealed into a glass with suitable properties depending on the diameter of the wire and the thickness of the plating.

535.37 2592
On the Inhibiting Effect of Oxygen on the Fluorescence of Solutions.—C. Chéchan. (*C. R. Acad. Sci., Paris*, 2nd Jan. 1946, Vol. 222, No. 1, pp. 80–82.)

535.37 2593

Note on the Behaviour of Zinc Sulphide Phosphors under Conditions of Periodic Excitation.—M. P. Lord & A. L. G. Rees. (*Proc. phys. Soc.*, 1st May 1946, Vol. 58, No. 327, pp. 280–289.) The effect of periodic excitation on luminescent solids and the electronic processes involved in the emission of luminescent radiation are examined theoretically, showing that the phase shift with respect to the exciting radiation and the ratio of the maximum to the minimum emitted intensities are the significant factors in the time function of the luminescent intensity. The variations of these parameters with intensity and period of excitation can be used to distinguish between the various mechanisms of the luminescent process, as is illustrated by an experimental examination of zinc sulphide and zinc cadmium sulphide phosphors. These phosphors show a semi-quantitative agreement with the characteristics of a simple ionization-recombination process, deviations from the theory being attributed to the activation of these phosphors by more than one type of activator atom.

535.37 2594

Note on the Rapid Determination of Decay Characteristics of Luminescent Solids.—M. P. Lord & A. L. G. Rees. (*Proc. phys. Soc.*, 1st May 1946, Vol. 58, No. 327, pp. 289–291.) Square-wave light pulses from a gaseous discharge tube, obtained by using a small vibrator to interrupt the d.c. supply to the tube, and used in conjunction with a cathode-ray display, afford a convenient and rapid method of investigation.

536.48 + 539.893 2595

Measurements at Low Temperatures and High Pressures: Part I.—Development of the Method for obtaining High Pressures at Low Temperatures.—B. Lazareff & L. Kahn. (*Zh. eksp. teor. Fiz.*, 1944, Vol. 14, Nos. 10/11, pp. 439–447.) The method proposed is based on the fact that certain substances such as water, bismuth, antimony and gallium increase in volume during the transition from a liquid into a solid state. In a bomb containing such a substance high pressures can thus be generated and transmitted to a body immersed in it. Accordingly a bomb was developed (Fig. 2) utilizing water and intended for investigating superconductivity by the induction method. A description is also given of a simple device for accurate measurements of the pressure by observing the expansion of the bomb. Pressures of the order of 2 000 kg/cm² were obtained with this bomb at the liquid helium temperature. It is indicated that, with bismuth and gallium, pressures up to 10 000 kg/cm² would be possible.

536.55 2596

Temperature Indicating Compounds.—G. A. Williams. (*Electronic Engng*, July 1946, Vol. 18, No. 221, pp. 208–212.) An account, with coloured illustrations, of methods of estimating temperature by the use of (*a*) paints formulated to change colour at given temperatures in the range 80–800°C, and (*b*) compounds available in the form of crayons, emulsions, or pellets, which melt at sharply defined temperatures in the range 52–982°C.

537.226 2597

The Theory of the Polarization of Dipole Liquids in Strong Electric Fields.—Anselm. (*See* 2540.)

539.234 : 535.87 2598

Anti-Reflexion Films Evaporated on Glass.—J. Bannon. (*Nature, Lond.*, 6th April 1946, Vol. 157, No. 3988, p. 446.) Reflection from a glass surface may be reduced substantially by evaporating a metallic fluoride on to the surface. Magnesium fluoride is particularly suitable, and the results of various laboratory evaporation processes with this material are given.

620.197 : 621.357.7 : 669.55.6 2599

Corrosion-Resisting Properties of Electrodeposited Tin-Zinc Alloys.—R. M. Angles & R. Kerr. (*Engineering, Lond.*, 29th March 1946, Vol. 161, No. 4185, pp. 289–292.) Alloys with tin content varying from 0 to 100% were tested on iron and steel. The 78% alloy gives the greatest protection and is able to withstand a reasonable amount of deformation by bending or cupping.

621.314.63 + 621.315.34 2600

An Anomalous Phenomenon in Thermal Rectification in Lead Sulphide (Preliminary Communication).—Amirkhanoff. (*See* 2552.)

621.314.63 + 621.315.34 2601

The Asymmetry of Conductivity in Electronic Semiconductors. — Kh. Amirkhanoff. (*Zh. eksp. teor. Fiz.*, 1944, Vol. 14, No. 6, pp. 187–192.) The following factors are enumerated, with which the appearance of the asymmetry is associated: (1) difference in the shape of the two contacts (plane, needle); (2) difference in the specific conductivity of the two electrodes; (3) temperature gradient.

Experimental data supplemented by the author's own experiments are surveyed, and the theoretical implications of each of the above factors discussed. The effects of combining several of these factors are then considered under the following headings: (*a*) Contact between a metallic needle and a crystal.—Heat is generated at the point of the contact, and thus both factors 1 and 3 are effective in this case. The two effects are cumulative or oppose each other according to the type of conductivity of the semiconductor (electrons, holes). (*b*) Solid-plate rectifiers.—No definite indication is available as to whether factors 1 and 3 are co-existent in this case with factor 2. (*c*) Thermal rectification.—This is effective simultaneously with factors 1 and 2. It is pointed out that in the case of copper oxide thermal rectification can only take place when there is a high-resistance layer on the electrode. This is proved by a number of experiments in which copper-oxide plates were etched by nitric acid. The results of these experiments given in table 2 show that thermal rectification becomes negligible in plates with polished surface.

621.314.632 2602

The Effect of Temperature Gradient on the Rectifying Action of Copper Oxide Rectifiers.—Kh. Amirkhanoff. (*Zh. eksp. teor. Fiz.*, 1944, Vol. 14, No. 6, pp. 195–201.) Experiments were conducted with plates of various types with facilities for varying the temperature at both sides of the plate within a range −15 to 150°C. The apparatus used is described, and the results obtained are shown in three tables. The main conclusion reached is that heating the upper electrode (oxide) and cooling the lower (copper) considerably improves the rectifying action. At the same time the sensitivity of the plate is also raised. It appears that under these conditions the thermal rectification is superimposed on the normal rectifying process. It is therefore suggested that in practice provision should be made for cooling only the copper surface of the rectifier elements. This would not only decrease the forward resistance, but also raise the permissible current density, as shown by Sharavski (307 of 1938).

Some of the results obtained in these experiments are also discussed from the point of view of an investigation of the barrier layer in copper-oxide rectifiers by means of a thermal sonde, reported by the author elsewhere.

621.315.59 : 621.396.822 **2603**

Voltage Fluctuations in Electronic Semi-Conductors.—B. I. Davidov. (*J. Phys., U.S.S.R.*, 1942, Vol. 6, No. 5, p. 230.) A general formula for the square of the voltage fluctuation is given, which reduces to the expressions for shot effect and Johnson effect under appropriate conditions. Abstract of a paper of the Acad. Sci., U.S.S.R.

621.315.61 : [621.315.2/.3 **2604**

Insulated Wire and Cable in Communications Today.—A. P. Lunt. (*Communications*, June 1946, Vol. 26, No. 6, pp. 30, 53.) The paper is concerned with available insulating materials, their properties and use in radio applications. Recommended types of insulation for a wide range of applications are given in tables.

621.315.611 **2605**

The Electrical Strength of Solid Solutions and their Melting Temperature.—N. Bogdanova. (*Zh. eksp. teor. Fiz.*, 1944, Vol. 14, Nos. 1/2, pp. 30–31.) A report of an experimental investigation. Data obtained by von Hippel (1186 of 1938) were used to plot the curves in Fig. 1 (system KCl – RbCl: system KCl–KBr also behaves in a similar manner) and Fig. 2 (system NaCl – Ag Cl). In the first case the maximum electrical strength is obtained with a melting temperature of the order of 750°C. In the second case there is a linear decrease in electrical strength with increase in melting temperature. Systems KI–NaI and KI–KBr were also investigated, and the results are plotted in Figs. 3 and 4 respectively. It appears that the electrical strength of these solutions increases with the melting temperature. The effect of the composition of solid dielectric solutions was also investigated, but no sharp strengthening of the dielectric with the introduction of admixtures was observed.

621.316.86 : 546.281.26 **2606**

Silicon Carbide [non-ohmic] **Resistors.**—F. Ashworth, W. Needham & R. W. Sillars. (*Engineering, Lond.*, 29th March 1946, Vol. 161, No. 4185, p. 295.) Summary of an I.E.E. paper. See also 1885 of July.

621.318.22/.23 **2607**

Modern Hard Magnetic Materials.—K. Hoselitz. (*J. sci. Instrum.*, April 1946, Vol. 23, No. 4, pp. 65–71.) A survey of the preparation and properties of the alloys, and data for the design of permanent magnets made with them.

621.318.322.029.54/.64 **2608**

The Permeability of Ferromagnetic Materials at Frequencies between 10^5 and 10^{10} c/s.—J. T. Allanson. (*J. Instn elect. Engrs*, Part I, May 1946, Vol. 93, No. 65, pp. 234–235.) Summary of 645 of March.

621.318.323.2.029.3 **2609**

Study of Ferromagnetic Cores exposed to Sinusoidal Induction [frequencies up to 10 kc/s].—I. Épelboim. (*C. R. Acad. Sci., Paris*, 30th April 1945, Vol. 220, No. 18, pp. 651–653.)

621.38/.39](058.7) **2610**

1946 Electronic Engineering Directory.—(*Electronic Industr.*, May 1946, Vol. 5, No. 5, 48 pp., following p. 90.) A directory of American sources of supply of electronic and radio equipment and materials.

621.386.1 : 548.73 **2611**

X-Ray Studies of Surface Layers of Crystals.—E. J. Armstrong. (*Bell Syst. tech. J.*, Jan. 1946, Vol. 25, No. 1, pp. 136–155.) When a crystalline substance is sawn, ground, lapped, or polished, the crystal structure adjacent to the worked surface is distorted, and the disturbance is detectable by X-ray diffraction. A single crystal spectrometer, in which the intensity of reflection of X rays by the surface is measured, can be used to detect highly distorted layers. Less distorted surfaces can be detected by means of a double crystal spectrometer or by photography of X rays transmitted through the crystal. In quartz crystals, the amount of misorientation is mainly of the order of a minute of arc, but some material may be misoriented by three or four degrees. In addition there is usually some randomly oriented powder which can be detected only by electron diffraction. The distorted layers may be removed by etching. The paper reviews the appropriate X-ray techniques, and gives examples of their application. 38 references are given.

669.231.635.8 : 621.326.21 **2612**

Welding Small Platinum Heaters and Electrodes.—A. R. Morris. (*J. sci. Instrum.*, April 1946, Vol. 23, No. 4, p. 84.) Simple device to facilitate hammer welding of thin platinum wire or foil.

621.793 **2613**

Metallizing Non-Conductors. [Book Review]—S. Wein. Metal Industry Pub. Co., New York, 62 pp., $2.00. (*Electronics*, May 1946, Vol. 19, No. 5, p. 324.) A survey of commercial methods of forming metallic films on almost any type of surface, including chemical formulae and detailed procedures.

MATHEMATICS

517.92 : 518.42 : 531.721 **2614**

Simple Differential Equations arising in Physics; Rapid Solution by using Hatchet Planimeters.—A. Callender. (*J. sci. Instrum.*, April 1946, Vol. 23, No. 4, pp. 77–81.)

517.947.44 **2615**

On the Near-Periodicity of Solutions of the Wave Equation: Part 3.—S. L. Soboleff. (*C. R. Acad. Sci. U.R.S.S.*, 10th Oct. 1945, Vol. 49, No. 1, pp. 12–15. In French.) For parts 1 and 2 see 1901 & 1902 of July.

518.5 **2616**

A Slide Rule for the Addition of Squares.—W. E. Morrell. (*Science*, 25th Jan. 1946, Vol. 103, No. 2665, pp. 113–114.) Instructions for making a slide rule for solving problems of the form $d = (x^2 + y^2 + \ldots)^{\frac{1}{2}}$.

518.6 **2617**

On Relaxation Methods: a Mathematics for Engineering Science.—R. V. Southwell. (*Proc. roy. Soc. A*, 21st Aug. 1945, Vol. 184, No. 998, pp. 253–288.) Review of a powerful method for obtaining approximate solution of boundary value problems (*e.g.* Laplace's or Poisson's equation), illustrated by application to various mechanical and electrical systems.

533.56 : 517.948.3 **2618**

The Boundary Problem of Electrodynamics and Integral Equations of Certain Diffraction Problems.—Feld. (*See* 2548.)

538.3 **2619**

The Transformation of the Integral of Retarded Potentials to the Liénard-Wiechert Formulae.—E. Durand. (*C. R. Acad. Sci., Paris*, 28th Jan. 1946, Vol. 222, No. 5, pp. 284–286.) The derivation is based on Jacobi's theorem on the changing of variables in definite integrals.

518.2 **2620**

Circular and Hyperbolic Functions, Exponential and Sine and Cosine Integrals, Factorial Function and Allied Functions, Hermitian Probability Functions (British Association Mathematical Tables, Volume I). [Book Review] — Cambridge Univ. Press, London, 2nd edn. 1946, 72 pp., 10s. (*Proc. phys. Soc.*, 1st May 1946, Vol. 58, No. 327, pp. 339–340.)

MEASUREMENTS AND TEST GEAR

621.317.1 : 621.396.621.029.6 **2621**

V.H.F. Receiver [selectivity] **Measurements.**—H. Gordon & L. George. (*Electronics*, June 1946, Vol. 19, No. 6, pp. 214..218.) Most v.h.f. signal generators have frequency scales that are too coarsely calibrated to show small frequency increments to the required accuracy. Measurements of the signal-frequency increments are made by observing the change in the i.f. of the receiver under test.

621.317.1.011.5 + 621.396.11.029.64 + 535.343.4] : 546.171.1 **2622**

Ammonia Spectrum in the 1 cm Wavelength Region.—B. Bleaney & R. P. Penrose. (*Nature, Lond.*, 16th March 1946, Vol. 157, No. 3985, pp. 339–340.) Measurements were made over the pressure range 0.2–600 mm Hg by observing, at the higher pressures, the decrease of power transmitted in a waveguide filled with the gas, and at the lower pressures, the damping of a cavity resonator. The results are briefly described and discussed. The absorption spectrum at 1.2 mm Hg is shown graphically for the wavelength range 1.15—1.48 cm. See also 1934 Abstracts p. 260 (Cleeton & Williams).

621.317.1.011.5 : 621.392 **2623**

The Theory of the Second Method of Drude.—B. K. Maibaum. (*Zh. eksp. teor. Fiz.*, 1944, Vol. 14, Nos. 10/11 & 12, pp. 448–458 & 501–513.) One of the more serious difficulties arising in the use of this method of measuring the dielectric constants and specific conductivities of dielectrics with the aid of a Lecher system is the possibility of negative displacements (see 471 of 1941—Maibaum), *i.e.* of conditions under which an increase, instead of a decrease, in the spacing between the two bridges becomes necessary. Six different existing theories of the method are examined critically, and it is shown that none of them is applicable to the above conditions. The effects of the following factors on the sign of the displacement are discussed: (1) ϵ' and ϵ'' (real and imaginary parts respectively of the complex dielectric constant); (2) suspension straps of the capacitor; (3) increase in diameter of the line conductors at the points at which the straps are connected; (4) the presence of an inductively coupled indicator. The effects of various constants of the capacitor on the results obtained by the Drude method are also discussed, and formulae for calculating these constants are given together with a description of the necessary experiments. In part 2 equations (1) for calculating ϵ' and ϵ'' are quoted, and a general solution of these is found. Simplified formulae (46) are also derived in which the attenuation of the measuring system and the effect of the suspension straps are taken into account.

621.317.32 : 621.3.015.33 **2624**

The Influence of Irradiation on the Measurement of Impulse Voltages with Sphere-Gaps.—J. M. Meek. (*J. Instn elect. Engrs*, Part II, April 1946, Vol. 93, No. 32, pp. 97–115.) Full paper of which summaries were noted in 365 of February and 967 of April.

621.317.33 : 621.611.21 **2625**

The Measurement of the Activity of Quartz Oscillator Crystals.—A. J. Biggs & G. M. Wells. (*J. Instn elect. Engrs*, Part I, April 1946, Vol. 93, No. 64, pp. 191–192.) Long summary of 969 of April.

621.317.335 **2626**

Measurement of the Ratio of Two Small Capacitances using a Tetrode Electrometer.—J. Lacaze. (*C. R. Acad. Sci., Paris*, 18th June 1945, Vol. 220, No. 25, pp. 876–877.)

621.317.35 : 534.61 **2627**

Range Extender for General Radio 760A Sound Analyzer.—Cobine & Curry. (*See* 2456.)

621.317.361 + 621.396.615 : 621.396.611.21 **2628**

Series-Resonant Crystal Oscillators.—Butler. (*See* 2524.)

621.317.715 : 621.394.645.35 **2629**

A Contact Modulated Amplifier to replace Sensitive Suspension Galvanometers.—M. D. Liston, C. E. Quinn, W. E. Sargeant & G. G. Scott. (*Rev. sci. Instrum.*, May 1946, Vol. 17, No. 5, pp. 194–198.) A high gain amplifier with low input impedance for measurement of small d.c. and very low frequency a.c. voltages and currents. The input is chopped by mechanical contacts, amplified, rectified by other synchronous contacts, and recorded by a milliammeter. Contact troubles are avoided in the input circuit by extreme rigidity, and by the use of gold for contacts and a special cadmium-tin mixture for soldering, as both have a low thermal e.m.f. against copper. The input circuit is also carefully shielded with mu-metal, and the size of all loops minimized. The equipment will operate from mains or batteries, and has a noise level of the order of 10^{-9} V, the actual value depending on the input circuit. On test, a zero drift of 4.10^{-9} V was recorded over a period of 4 hours.

621.317.715.5 **2630**

The Use of a Moving Coil Galvanometer for Recording at Frequencies Higher than its Own.—D. C. Johnson. (*J. sci. Instrum.*, June 1946, Vol. 23, No. 6, pp. 113–114.) The speed of response of the galvanometer can be increased at the expense of sensitivity by introducing "electromagnetic stiffness" by means of external resistive and reactive circuits.

621.317.72.029.62 **2631**

A Field-Intensity Meter for V.H.F.—D. C. Summerford. (*QST*, June 1946, Vol. 30, No. 6, pp. 40–42.) Constructional details of a small instrument incorporating an acorn triode rectifier for rough measurements at frequencies up to 225 Mc/s.

621.317.725 **2632**

A Correction Formula for Voltmeter Loading.—R. E. Lafferty. (*Proc. Inst. Radio Engrs, W. & E.*, June 1946, Vol. 34, No. 6, p. 358.) The finite resistance of a current-operated voltmeter causes errors due to the load it imposes on the source being measured. A simple formula is derived which enables the true voltage to be obtained from measurements on two voltage ranges which have a known internal resistance ratio. A similar equation may be used to correct the reading of an ammeter having appreciable resistance.

621.317.725 **2633**

Suppressed-Range Recording Peak Voltmeter.—F. G. Brockman. (*Rev. sci. Instrum.*, May 1946, Vol. 17, No. 5, pp. 177–179.) Description of an instrument designed to detect small fluctuations (0.000 95 V) in relatively large audio-frequency voltages (5–30 V). A single-valve circuit is used, but stabilized filament and h.t. supplies are required. Calibration methods are given in detail.

621.317.733 : 621.395.645 **2634**

A Convenient Amplifier and Null Detector.—H. H. Scott & W. F. Byers. (*Gen. Radio Exp.*, March 1946, Vol. 20, No. 10, pp. 1–3.) The high-gain amplifier is suitable for the range 20 c/s to 100 kc/s. A semi-logarithmic valve voltmeter is included to provide null deflexion indications. Miniature valves are used throughout. The circuit diagram and gain/frequency curves are given.

621.317.733.085.3 **2635**

Phase Sensitive [a.c.-] **Bridge Detector.**—P. H. Hunter. (*Electronic Industr.*, June 1946, Vol. 5, No. 6, pp. 60–61.) The detector consists of a twin-triode network essentially similar to a full-wave grid-controlled rectifier system. It needs no d.c. power supply, and indicates the direction as well as the magnitude of the bridge unbalance. The inclusion of an amplifier increases the sensitivity. The device may be useful in industrial process recording and automatic control as well as in ordinary bridge work.

621.317.738 **2636**

Production Bridge for Incremental [inductance] **Tests.**—W. Muller. (*Electronic Industr.*, May 1946, Vol. 5, No. 5, pp. 72..122.) The apparatus contains d.c. and a.c. power sources with an oscillograph as detector. A switch permits operation as either a Hay or an Owen bridge. Inductances from 1 mH to 50 H may be measured with d.c. up to 1 A, 50–500 H with 0.15 A and above 500 H with 15 mA.

621.317.78 : 621.317.382 **2637**

Oscillographic Arrangement for Measuring Small Powers.—J. Benoit. (*C. R. Acad. Sci., Paris*, 2nd Jan. 1946, Vol. 222, No. 1, pp. 59–60.) The voltage across the load is applied to the X-plates, and the voltage across a very small resistance in series with the load is applied, through a valve with a capacitive load, to the Y-plates, so that the area enclosed by the c.r. trace is proportional to the power in the load.

621.317.78.029.5/.6] : 621.326 **2638**

Load Lamp for Microwave Power Measurements.—J. E. Beggs. (*Electronics*, June 1946, Vol. 19, No. 6, pp. 204..210.) To reduce lead inductance and glass losses in small load lamps, a concentric-line construction has been used with multiple filaments of fine wire strung from inner to outer line. Low-loss sealing glass and a gettered vacuum prevent breakdown when measuring pulsed h.f. power.

621.317.79 : 621.314.24 : 621.396.823 **2639**

Maintenance Testing of Dynamotors.—H. M. Tremaine. (*Electronics*, June 1946, Vol. 19, No. 6, pp. 158..168.) Description and circuit diagram of an equipment for the rapid testing of load characteristics and radio noise interference.

621.317.79 : 621.318.4 **2640**

Coil Short Tests.—N. L. Chalfin. (*Electronic Industr.*, May 1946, Vol. 5, No. 5, p. 77.) A solenoidal iron-cored transformer, with two secondaries connected in opposition, is unbalanced when a coil with a shorted turn is placed on the core near one of the secondaries. The resulting voltage operates a relay or indicator.

621.317.79 : 621.395.82 : 621.395.645 **2641**

Measuring Audio Intermodulation.—N. C. Pickering. (*Electronic Industr.*, June 1946, Vol. 5, No. 6, pp. 56..125.) Sine waves at 100 c/s and 7 kc/s are fed to an amplifier under test. The 100 c/s component is filtered from the amplifier output, leaving the 7 kc/s as a carrier with intermodulation sidebands 7 kc/s $\pm$ 100 c/s. This combined signal is amplified and applied to a linear detector which feeds a carrier-level meter and a modulation meter calibrated to read percentage distortion. The theory of the method is summarized, and results of typical measurements are shown graphically.

621.317.79 : 621.396.62 **2642**

Multipurpose Tester.—B. White. (*Radio Craft*, May 1946, Vol. 17, No. 8, pp. 534..567.) Description of an r.f. and a.f. signal tracer that incorporates a volt-, ohm-, and milliampere-meter.

621.317.79 : 621.396.645 **2643**

Visual Radio Alignment.—E. J. Thompson. (*Radio Craft*, May 1946, Vol. 17, No. 8, pp. 540..574.) Description and circuit diagram of a frequency-modulated signal generator ("wobbulator"), mean frequency 415–450 kc/s, bandwidth adjustable 0–40 kc/s.

621.317.79 : 621.396.712 **2644**

Portable Precision Amplifier-Detector.—F. A. Peachey, S. D. Berry & C. Gunn-Russell. (*Wireless Engr*, July 1946, Vol. 23, No. 274, pp. 183–192.)

Description of an instrument for tone-level measurements at 50–10 000 c/s to the nearest 0.1 db over the range + 20 db to − 50 db with respect to 1 mW in 600 Ω. Facilities are provided for measuring peak programme level and noise level down to − 110 db. The two main attenuator controls are geared together so that the algebraic sum of the attenuators is seen directly on the calibrated dial through a window.

621.317.79 : 621.396.9 **2645**

Techniques and Facilities for Microwave Radar Testing.—E. I. Green, H. J. Fisher & F. J. Ferguson. (*Trans. Amer. Inst. elect. Engrs*, May 1946, Vol. 65, No. 5, pp. 274–290.) Equipments and procedures developed at the Bell Telephone Laboratories for testing radar apparatus in the range 500–25 000 Mc/s. The requirements are outlined, and the following items are described: signal generators, their design and application to receiver testing; frequency measurement by the use of tuned cavities; power measurement by the use of thermistors, and, for high power, by the use of directional couplers of known loss in conjunction with thermistors; echo boxes, their design, properties, and uses for testing overall performance; spectrum analysis with echo boxes; standing-wave measurements with various devices; directional couplers; attenuators and pads; oscilloscopes; range calibration: computer test sets.

621.317.79 : 621.397.62 **2646**

A Television Signal Generator: Part 2 — Monoscope and Video Circuits.—R. G. Hibberd. (*Electronic Engng*, July 1946, Vol. 18, No. 221, pp. 204–207.) A detailed description, with circuit diagrams. The frame and line scanning generators each consist of a thyratron feeding a cathode follower with a suitable output stage matched to the low-impedance scanning-coil system, synchronizing impulses being fed to the grid of each thyratron. The monoscope output passes to a video amplifier with a cathode-follower input and output, and then to the impulse-mixing unit where the blanking and synchronizing impulses are mixed into the video signal. All units are supplied from a stabilized power pack. For part 1 see 2255 of August.

621.317.79 : 621.397.645 **2647**

Transient Video Analyzer.—C. Moritz. (*Electronics*, June 1946, Vol. 19, No. 6, pp. 130–135.) A description, with circuit details, of a test set combining a five-signal transient generator and wide-band oscilloscope for checking the performance of wide-band amplifiers connected between them. The five signals are (*a*) a 30-c/s square-wave, (*b*) a 5-kc/s saw tooth, (*c*) a 10-μs pulse repeated at 5 000 per sec, (*d*) a step function, (*e*) a spike function. The sweep of the oscilloscope is synchronized with the signal. The oscilloscope amplifier is designed to have higher fidelity than any ordinary equipment likely to be tested.

621.317.79.029.5 **2648**

LCR Meter for Amateur Use.—W. B. Bernard. (*Radio News*, June 1946, Vol. 35, No. 6, pp. 40..125.) Constructional details of an instrument which measures Q from 10 to 600, capacitance up to 350 μμF accurately and up to 1 μF less accurately, and which can be used as a modulated or unmodulated local oscillator with an output of 1 W in the frequency range 90 kc/s–36 Mc/s.

621.318.4.013.22 **2649**

On a System of Coils producing a Uniform Magnetic Field for a Narrow Wilson Chamber.—Nageotte. (*See* 2720.)

621.396 : 621.317 **2650**

Application of Radio Technique to General Measurements.—G. R. Polgreen. (*J. Instn elect. Engrs*, Part I, April 1946, Vol. 93, No. 64, p. 160.) Abstract only.

621.396.615.029.5/.63 **2651**

Test Oscillator TS-47/APR.—D. W. Moore, Jr. (*Radio News*, May 1946, Vol. 35, No. 5, pp. 32–34, 08.) Detailed description of a robust (army) test oscillator covering 40–500 Mc/s in two bands, with 1% frequency accuracy. Provision is made for 1 000-c/s sine-wave modulation or pulse modulation.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

621.3.078 **2652**

On Increasing the Stability of Self-Regulation by Means of Back Coupling.—Peshkoff. (*See* 2784.)

621.317.39.082.7 **2653**

Rapid Moisture Testing of Granular Material.—J. H. Jupe. (*Electronics*, May 1946, Vol. 19, No. 5, pp. 180..186.) An account of the instrument described in 388 of February (Hartshorn & Wilson).

621.365.92 : 664.84 **2654**

Electronics in Processing Foods.—S. R. Winters. (*Radio News*, June 1946, Vol. 35, No. 6, pp. 31..123.) Report on use of dielectric heating in food preservation. See also 393 of February (Moyer & Stotz).

621.38 : 6(048) **2655**

Electronic Uses in Industry.—W. C. White. (*Electronic Industr.*, June 1946, Vol. 5, No. 6, pp. 66..111.) The fourth of a series of selected references, published annually. About four hundred titles are given, with a subject index. For previous lists see 2844 of 1945.

621.383 : 522.2 **2656**

Photoelectric Sight for Solar Telescope.—W. O. Roberts. (*Electronics*, June 1946, Vol. 19, No. 6, pp. 100–103.) A separate guiding telescope is attached to the main telescope, and a disk which masks most of the solar light is automatically centred within 1 second of arc by means of four photocells and associated amplifiers and relays.

621.383 : 551.576 **2657**

On Atmospheric [cloud-height] **Sounding.**—R. Barthélemy. (*C. R. Acad. Sci., Paris*, 18th Feb. 1946, Vol. 222, No. 8, pp. 450–451.) Brief account of an optical echo-sounding device for cloud-height determination apparently identical in principle with the equipment described in 1943 of July (Moles). Extreme range 7 000 metres.

621.383.078 : 778.6 **2658**

Photoelectric Controls for [photographic] **Color Printing.**—J. Robins & L. E. Varden. (*Electronics*, June 1946, Vol. 19, No. 6, pp. 110–115.) A discussion of the problems involved in the production of high quality photographic colour prints on a large scale. Details of photoelectric circuits used in various processes are given.

621.385.833 : 537.133 **2659**
On a Project for a Proton Microscope.— C. Magnan, P. Chanson & A. Ertaud. (*C. R. Acad. Sci., Paris*, 28th May 1945, Vol. 220, No. 22, pp. 770–772.) By using protons instead of electrons the resolving power due to diffraction will be improved by a factor of 40 for the same aperture and equal energy. A resolving power of 3 Å is expected with a magnification of 20 000. Enlargement of the photograph increases magnification to 600 000.

621.389 : 535–1 **2660**
Night Vision with Electronic Infrared Equipment.—(*Electronics*, June 1946, Vol. 19, No. 6, pp. 192..204.) A description of the application of infra-red image converters and searchlights in German military equipments.

621.389 : 623.555.2 : 535–1 **2661**
Electronic Night Sight.—W. MacD. (*Electronics*, June 1946, Vol. 19, No. 6, p. 95.) See also 2346 of August.

621.392.43 : 621.365.92 **2662**
Coupling Method for Dielectric Heating.—Kleinberger. (*See* 2498.)

621.396.610.16 : 621.385.38 **2663**
Thyratron Pulser Tube for Industrial Microwaves.—(*See* 2770.)

621.398 : 534.43 **2664**
[Gramophone] **Remote Record-Selection System.**—F. M. Berry. (*Electronics*, June 1946, Vol. 19, No. 6, pp. 104–106.) An oscillator transmits r.f. pulses by a carrier-current system to discriminator circuits in the gramophone cabinet. Twelve carrier frequencies and the use of both alternations of the power line permit the selection of 24 records.

621.398 : 621.397 **2665**
Tele-Guided Missiles.—(*See* 2738.)

623.26 : 621.396.9 **2666**
Mine Detectors.—(*Wireless World*, May 1946, Vol. 52, No. 5, pp. 166–168.) An account of the British Army detector No. 4, with circuit diagram. See also 1626 of June (West).

623.454.25 : 621.396.9 **2667**
The Radio Proximity Fuze.—L. G. Hector. (*Proc. Radio Cl. Amer.*, March 1946, Vol. 23, No. 3, pp. 3–6.)

786.6 : 621.383 **2668**
Photo-Electronic Organ : Part 1.—R. E. Campbell & L. E. Greenlee. (*Radio News*, June 1946, Vol. 35, No. 6, pp. 25..110.) A beam of light passes through a tone pattern printed on a rotating wheel to a photocell. The impulses are amplified and reproduced on a loudspeaker. A detailed description of the construction of the instrument.

621.385.833 **2669**
The Electron Microscope. [Book Review]—D. Gabor. Hulton Press, 1945, 104 pp., 4s. 6d. (*Wireless World*, May 1946, Vol. 52, No. 5, p. 168.) Concerned with the electron-optical aspects.

PROPAGATION OF WAVES

523.78 : 551.51.053.5 : 621.396.11 **2670**
The Influence of an Eclipse of the Sun on the Ionosphere.—Smith-Rose. (*See* 2562.)

541.133 : 621.3.029.5/.6 **2671**
The Variation of the Electrical Conductivity of Electrolytes with Frequency.—Maloff. (*See* 2549.)

621.396.11 **2672**
Theory of the Coastal Refraction of Electromagnetic Waves.—G. Grünberg. (*Zh. eksp. teor. Fiz.*, 1944, Vol. 14, Nos. 3/4, pp. 84–111. English version in *J. Phys., U.S.S.R.*, 1942, Vol. 6, No. 5, pp. 185–200.) The difficulties of investigating the propagation of electromagnetic waves in the presence of three different media (air, sea and land) are pointed out, and a simplified treatment of the problem is presented. The propagation of the waves along a boundary plane between the media (air and ground) is considered, and approximate boundary conditions (10) for the electric field component normal to the surface of the ground are derived using the fact that for sufficiently high conductivity of the medium the fall of the intensity of the field penetrating it does not depend on the character of the field outside the medium. From these conditions, assuming that the sea is an ideal conductor and also that the surface of the earth is a horizontal plane, an integral equation (21) of the component is obtained. This is the main equation of the problem, an exact solution of which would give a complete answer. In the present paper an approximate solution (41) of the equation is found and its implications discussed in detail for the case of plane waves impinging on an unlimited rectilinear coast. General considerations are also given regarding an approximate solution of the equation for the case of small islands of an arbitary shape. Formulae are given for evaluating definite and indefinite integrals containing Bessel functions.

This is the paper referred to in 1830 of 1943 and 3380 of 1944.

621.396.11.029.64 **2673**
An Experimental Investigation of the Reflection and Absorption of Radiation of 9-cm Wavelength.—L. H. Ford & R. Oliver. (*Proc. phys. Soc.*, 1st May 1946, Vol. 58, No. 327, pp. 265–280.) Using angles of incidence ranging from 45° to 80°, measurements were made on the reflecting power of the surfaces of level and uneven bare ground, vegetation-covered ground, tap water, and a 4% salt solution. "Specular reflection was found to occur only from very level surfaces; the absorptions of these surface media were measured, and from the combined measurements of reflection and absorption their electrical constants were derived.

"Rough surfaces, either of bare ground or vegetation-covered ground, gave values of reflection coefficient in general agreement with the optical rule that regular reflection is only observed from an uneven surface if the product of the depth of the surface irregularities and the cosine of the angle of incidence is a small fraction of the wavelength. If this fraction exceeded ⅛ λ, the values of reflection coefficient measured were about 0.1.

"The calculated values of reflection coefficient corresponding to dry soil, wet soil, and sea water, for angles of incidence varying from 0° to 85° are given in an appendix to the paper."

621.396.11 : 551.51.053.5 "1946.05" **2674**
Short-Wave Conditions : Expectations for May.—T. W. Bennington. (*Wireless World*, May 1946, Vol. 52, No. 5, p. 165.)

621.396.11 : 551.51.053.5] " 1946.06 " 2675
Short-Wave Conditions: Expectations for June.—T. W. Bennington. (*Wireless World*, June 1946, Vol. 52, No. 6, p.207.)

RECEPTION

621.396.61 + 621.396.621].029.63 2676
Getting Started on 420 Mc/s.—Hoisington. (*See* 2757.)

621.396.61 + 621.396.677].029.63 2677
CQ 2 400 Mc/s : Transceivers and Antennas for the 13-Centimeter Band.—Koch & Floyd. (*See* 2758.)

621.396.61.029.63 2678
A U.H.F. Ham Transceiver.—I. Queen. (*Radio Craft*, May 1946, Vol. 17, No. 8, pp. 545..588.) Description of a small equipment for 420–450 Mc/s.

621.396.621 + 621.395.645 2679
Superamp with Tuner.—Brennan. (*See* 2512.)

621.396.621 2680
The Radio News Circuit File.—(*Radio News*, May & June 1946, Vol. 35, Nos. 5 & 6, pp. 60..82 & 62..72.) Circuit diagrams and parts lists of 21 American post-war commercial broadcast receivers, arranged so that they can be cut out and attached to 5 in. × 3 in. filing cards.

621.396.621 2681
Looking Over the Postwar Receivers.—B. G. (*QST*, June 1946, Vol. 30, No. 6, pp. 24..108.) Review of the mechanical and electrical features of the Hammarlund HQ–129–X. The noise limiter is explained with the aid of a circuit diagram.

621.396.621 2682
High-Level Detector.—J. C. Rankin. (*Electronics*, May 1946, Vol. 19, No. 5, pp. 212..218.) Audio-frequency amplification is eliminated from a radio receiver by using a low-impedance copper-oxide rectifier immediately before the loudspeaker and following an i.f. power amplifier.

621.396.621 2683
Radio Data Sheet 335.—(*Radio Craft*, May 1946, Vol. 17, No. 8, p. 547.) Servicing data for RCA Victor receivers 54B1, 54B1-N, 54B2 & 54B3.

621.396.621 : 621.396.662 2684
Practical Radio Course : Part 43.—A. A. Ghirardi. (*Radio News*, April 1946, Vol. 35, No. 4, pp. 46..107.) Automatic frequency control of receivers. For previous parts of the series see 1646 of June.

621.396.621 : 621.396.9 + 621.396.61 2685
Radar Technique.—F. L. D. (*See* 2751.)

621.396.621.004.67 2686
Radio Servicing.—(*Elect. Rev., Lond.*, 11th Jan. 1946, Vol. 138, No. 3555, p. 70.) Summary of I.E.E. discussion led by R. C. G. Williams. The need for more test equipment and for a recognized qualification for servicemen was stressed.

621.396.621.029.56/.58 2687
The "Super-3".—(*Radio News*, June 1946, Vol. 35, No. 6, pp. 76..88.) Constructional details of a three-valve regenerative receiver for 1.7–14.5 Mc/s.

621.396.621.029.6 : 621.317.1 2688
V.H.F. Receiver [selectivity] **Measurements.**—Gordon & George. (*See* 2621.)

621.396.621.029.62.004.67 2689
100 Mc/s Receivers require New Servicing Techniques.—D. W. Gunn. (*Radio News*, May 1946, Vol. 35, No. 5, pp. 36..68.) Brief survey of the complications which a service-engineer will encounter in f.m. receivers for frequencies over 50 Mc/s.

621.396.621.5 2690
Super-Regenerative Receivers.—"Cathode Ray". (*Wireless World*, June 1946, Vol. 52, No. 6, pp. 182–186.) A simple explanation of the manner of operation prompted by the recent successful application of the principle in military equipments.

621.396.621.5 2691
The Super-Regenerative Detector : an Analytical and Experimental Investigation.—F. R. W. Strafford. (*J. Instn elect. Engrs*, Part I, April 1946, Vol. 93, No. 64, p. 192.) Summary of 1032 of April.

621.396.621.54 2692
Practical Radio Course : Parts 44 & 45.—A. A. Ghirardi. (*Radio News*, May & June 1946, Vol. 35, Nos. 5 & 6, pp. 48..123 & 55..70.) An account of frequency conversion in superheterodyne receivers. For previous parts in the series see 2684 above.

621.396.621.54.029.62 2693
Miniature Tubes in a Six-Meter Converter.—R. W. Houghton. (*QST*, June 1946, Vol. 30, No. 6, pp. 18–21.) Constructional details of a set for frequency conversion to 10.5 Mc/s.

621.396.622 2694
Audio - Modulated Detection : an Improved Method for Reception of C.W. Signals.—D. A. Griffin & L. C. Waller. (*QST*, July 1946, Vol. 30, No. 7, pp. 13–15, 124.) Two diodes are connected in opposite senses, in parallel, with a common load resistance. One diode is biased by a square-wave a.f. signal from a special generator. This provides a.f. modulation of c.w. signals, and also an upper-level limiting action. Bias applied to the other diode provides low-level limiting. The advantage in signal/noise ratio is considerable, especially if an a.f. output filter is used.

621.396.82 2695
Analysis of Radio Interference Phenomena.—(*Radio News*, June 1946, Vol. 35, No. 6, p. 54.) A table showing character, cause, type of receivers affected, where prevalent, and suggested service remedies, for 11 types of interference.

621.396.822 2696
A Theory of Valve and Circuit Noise.—N. R. Campbell & V. J. Francis. (*J. Instn elect. Engrs*, Part I, April 1946, Vol. 93, No. 64, p. 190.) Summary of 1037 of April.

621.396.822 : 621.315.59 2697
Voltage Fluctuations in Electronic Semi-Conductors.—Davidov. (*See* 2603.)

621.396.822 : 621.396.13 2698
Theoretical Signal-to-Noise Ratios.—J. E. Smith. (*Electronics*, June 1946, Vol. 19, No. 6, pp. 150–152,

154.) The sources of noise are described. Signal/noise ratios are derived in terms of the frequency bands of the signal and of the transmitted radio carriers for single or double a.m. or f.m., as used in u.h.f. multiplex relay systems. The relative advantages of the systems are shown in tabular form.

621.396.822 : 621.396.671 **2699**

Fluctuation Noise in a Receiving Aerial.—R. E. Burgess. (*Proc. phys. Soc.*, 1st May 1946, Vol. 58, No. 327, pp. 313–321.) "The factors which determine the signal/noise ratio at the terminals of a receiving aerial are discussed. The aerial noise considered is the random fluctuation type, consisting of (i) thermal noise associated with the loss resistance of the aerial, and (ii) noise associated with the radiation resistance which is induced by the surroundings. The effective temperature of the radiation resistance is expressed in terms of the temperature distribution of the surroundings and the distribution of power dissipation when the aerial is transmitting. Radiation from the sun and from the Milky Way are briefly discussed, and it is shown that the detection of solar radiation at radio frequencies requires the use of highly directional aerials. The limitations imposed by the receiver noise on (*a*) the sensitivity of the reception of signals, and (*b*) the accuracy of measurement of aerial noise, are discussed and the results presented graphically.

"The conclusions are of most practical interest at the higher radio frequencies (above about 30 Mc/s) where atmospheric noise is negligible."

621.396.823 **2700**

Interference from Industrial Electronic Apparatus.—(*Wireless World*, June 1946, Vol. 52, No. 6, p. 198.) Report of an I.E.E. discussion led by M. R. Gavin.

STATIONS AND COMMUNICATION SYSTEMS

621.396 **2701**

Engineers study F.M.—(*Electronic Industr.*, May 1946, Vol. 5, No. 5, pp. 66–70.) Extracts from papers read at the sixth Broadcasting Conference, held at Columbus, Ohio, in March 1946, covering other subjects besides f.m. See also 2125 of August and cross references.

621.396.1 **2702**

The Conference on [allocation of] **Radio Frequencies for Civilian Services in the Liberated Countries of Europe.**—L. B. de Cléjoulx. (*Onde élect.*, Jan. 1946, Vol. 26, No. 226, pp. 45–46.) A short account of the conference held in London in Sept. 1945.

621.396.619.018.41 **2703**

Frequency Modulation: Parts 1 & 2.—P. Besson. (*Onde élect.*, Jan.-May 1946, Vol. 26, Nos. 226–230, pp. 6–25, 74–91, 107–129, 155–172 & 194–214, with other parts to follow.) A comprehensive review of existing knowledge. Part 1 is in nine chapters as follows: introduction; history; principles of f.m.; production of f.m. oscillations; spectra of f.m. oscillations; reception; comparison of f.m. and a.m. reception; propagation of f.m. waves; comparative advantages and disadvantages of f.m. and a.m. Part 2, dealing with technique and applications, is in four chapters as follows: transmitters; receivers; measurements; applications. The instalments to which references are given include part 1, and chapters 1, 2 and part of chapter 3 of part 2.

621.396.619.018.41 **2704**

Spectrum of a Phase- or Frequency-Modulated Wave.—R. E. Burgess. (*Wireless Engr*, July 1946, Vol. 23, No. 274, pp. 203–204.) If the carrier frequency is an integral or half-integral multiple of the modulation frequency, each sidewave is a doublet and, in general, the mean square value of the wave is not equal to half the square of the amplitude. See also 2637 of 1944 (Colebrook).

621.396.619.018.41 : 621.396.61 **2705**

Frequency Modulated Transmitters for Police and Similar Services.—Fairbairn. (*See* 2750.)

621.396.619.16 + 621.396.61.029.64 **2706**

Multi-Channel Pulse Modulation.—(*Wireless World*, June 1946, Vol. 52, No. 6, pp. 187–192.) Details of the British Army Signalling Equipment No. 10, operating on a wavelength in the 6–7 cm band, and using pulse modulation.

"Narrow pulses recurrent at 9 kc/s are width-modulated by the a.f. signals which are to be transmitted. The pulses are narrow with an average duration of 3.5 μs. Each of the eight speech channels modulates a separate train of pulses of the same recurrence frequency, but all the trains are staggered in time so that they can fit together without interference. A synchronizing pulse is also included to control the receiving apparatus which sorts out the pulses, demodulates them, and routes the a.f. signals to the proper output circuits." See also 470 of February.

621.396.619.16 **2707**

Pulse Modulation.—F. F. Roberts & J. C. Simmonds. (*Wireless Engr*, July 1946, Vol. 23, No. 274, p. 204.) Continuation of correspondence on 183 of January (Roberts & Simmonds) following 1676 of June (Shepherd).

621.396.7 (058.7) **2708**

Radio Stations.—(*Electronic Industr.*, May 1946, Vol. 5, No. 5, pp. 89–92.) Directory giving the addresses of U.S. broadcasting stations for a.m., f.m. and television, and names of the chief engineers.

621.396.712.004.5 **2709**

Preventive Maintenance for Broadcast Stations.—C. H. Singer. (*Communications*, June 1946, Vol. 26, No. 6, pp. 22–28, 52.) First of a series of papers dealing with the methods and equipment used. The need for proper care of tools is emphasized. Descriptions are given of simple tools used in the maintenance of relay contacts and commutators.

621.396.931 **2710**

Railroad F.M. Satellite System.—W. S. Halstead. (*Communications*, May 1946, Vol. 26, No. 5, pp. 17–21 . . 55.) See also 2325 of August (Halstead).

621.396.931 **2711**

Railroad Radio — from F.C.C. to I.C.C.—J. Courtney. (*Electronics*, June 1946, Vol. 19, No. 6, pp. 92–94.) A general discussion of the advantages of a railway radio service compared with existing manual block systems.

621.396.931.029.62 **2712**

V.H.F. Communication Equipment.—(*Wireless World*, June 1946, Vol. 52, No. 6, pp. 180–181.) A short general description of some of the equipment for the police radio system described in 1357 of May and 2326 of August.

SUBSIDIARY APPARATUS

621.314.2 2713

Equivalent Capacitances of Transformer Windings.—Duerdoth. (*See* 2493.)

621.314.632 2714

The Effect of Temperature Gradient on the Rectifying Action of Copper Oxide Rectifiers.—Amirkhanoff. (*See* 2602.)

621.316.86 : 546.281.26 2715

Silicon Carbide [non-ohmic] **Resistors.**—Ashworth, Needham & Sillars. (*See* 2606.)

621.316.974 : 621.318.4.017.31 2716

Power Loss in Electromagnetic Screens.—C. A. Siocos. (*Wireless Engr*, July 1946, Vol. 23, No. 274, p. 202.) A letter in which the method of Davidson *et al.* (1077 of April) is extended to the calculation of the eddy current density produced by a coil in an infinite shielding plane.

621.317.755 2717

A Simple Oscilloscope: using the Mains as a Time Base.—F. P. Williams. (*Wireless World*, June 1946, Vol. 52, No. 6, p. 206.) Deflecting and focusing coils are made nearly to follow television practice. The "straight" middle portion of the sine wave from the mains is used to provide a timebase.

621.318.22/.23 2718

Modern Hard Magnetic Materials.—Hoselitz. (*See* 2607.)

621.318.24 2719

Capacitor Discharge Magnetizer for Plant Shops.—W. L. Porta. (*Electronics*, June 1946, Vol. 19, No. 6, pp. 168..188.) Description of an equipment for magnetizing small permanent magnets. The essential components are a source of d.c. voltage, a capacitor, and a special transformer. The transformer, which must have a low leakage-reactance factor, carries a 1-millisecond power impulse. The capacitor discharge into the primary is initiated through a manually controlled ignitron tube.

621.318.4.013.22 2720

On a System of Coils producing a Uniform Magnetic Field for a Narrow Wilson Chamber.—E. Nageotte. (*C. R. Acad. Sci., Paris*, 16th April 1945, Vol. 220, No. 16, pp. 557–559.) The coils are each wound on a rectangular former of which the long sides are curved in a direction normal to the plane of the undistorted rectangle. The coils are arranged so that the convex sides of the formers are together. The magnetic field shows variations of 0.1% over a space in which the Helmholtz arrangement would show variations of 0.5%.

621.318.44 2721

The Technique of Toroidal Winding.—F. E. Planer. (*Electronic Engng*, July 1946, Vol. 18, No. 221, pp. 199–203.) A brief description of the advantages of toroidal coils and an account of the principle of a toroidal winding machine. Various methods are given for maintaining constant tension in the wire, illustrated by photographs of representative types of machine.

621.318.57 2722

A Simple Electronic Relay.—F. A. P. Maggs. (*J. sci. Instrum.*, April 1946, Vol. 23, No. 4, pp. 85–86.) Simple valve circuits for operating electromagnetic and thermal types of relay.

621.319.42 2723

Capacitor Machine.—(*Radio, N.Y.*, April 1946, Vol. 30, No. 4, pp. 10, 12.) Illustrated note on a U.S. Dept. of Commerce OPB report PB421 by F. E. Henderson, describing a German (Bosch Co.) method of making paper capacitors, in which the usual metal foil is replaced by a film of zinc or cadmium deposited on the paper dielectric.

621.383 2724

Cooling Photosensitive Cells [by the use of solid CO_2].—E. F. Coleman. (*Electronics*, June 1946, Vol. 19, No. 6, pp. 220 . . 224.)

621.383.2 : 621.314.6 2725

Experimental Behaviour of a Photoelectric Cell under the Influence of an Alternating Potential of Very High Frequency.—D. Charles. (*C. R. Acad. Sci., Paris*, 29th Oct. 1945, Vol. 221, No. 18, pp. 495–497.) It has previously been shown (1948 of 1941—Geest) that over particular frequency ranges and for particular voltages the cell gives abnormally high rectified currents, even when not illuminated. Special cells were made to investigate this phenomenon, using a number of different photoelectric materials. Potential differences of 0–300 V were applied at frequencies of 60–120 Mc/s. The following results were obtained: (*a*) Ni and Al cathodes do not show the phenomenon; (*b*) for a given frequency, the rectified current increases to a maximum, then drops abruptly to zero as the h.f. voltage is increased; (*c*) for a given voltage the output rises to a maximum and falls again as the frequency is increased; (*d*) positive bias increases, and negative bias decreases the output; (*e*) the phenomena are unaffected by illumination; (*f*) cooling to −140°C neither stops the phenomenon nor prevents it from starting; (*g*) modulation at 10 Mc/s has no effect; (*h*) a magnetic field normal to the axis stops the phenomenon. See also 2726 below.

621.383.2 : 621.314.6 2726

Theory of the Behaviour of a Photoelectric Cell under the Influence of an Alternating Potential of Very High Frequency.—D. Charles. (*C. R. Acad. Sci., Paris*, 2nd Jan. 1946, Vol. 222, No. 1, pp. 65–67.) For experimental work see 2725 above. Electron trajectories were calculated by numerical integration of the equations of motion. For a fixed h.f. voltage and frequency a group of electrons with initial velocities lying within a small range are concerned in a resonant phenomenon. Complete calculation of the response curves as a function of frequency for different values of h.f. voltage for a caesium cell gives results in good agreement with experiment.

621.385.18 : 621.396.9 2727

The Gas-Discharge Transmit-Receive Switch.—Samuel, Clark & Mumford. (*See* 2783.)

621.389 : 535-1 2728

Night Vision with Electronic Infrared Equipment.—(*See* 2660.)

621.389 : 623.555.2 : 535-1 2729

Electronic Night Sight.—W. MacD. (*See* 2661.)

621.394.624 **2730**
Electronic Code Translator.—H. W. Babcock. (*Electronics*, June 1946, Vol. 19, No. 6, pp. 120–122.) Code signals (*e.g.* Morse) from a receiver are fed into a discriminator in which five or more pairs of thyratrons produce a unique voltage determined by each group of dots and dashes. This triggers a gas-filled tube. Illumination from the tube causes corresponding letters and numerals to appear on a moving fluorescent screen.

621.396.66 : 621.395.63 **2731**
Selective Calling System.—J. K. Kulansky. (*Electronics*, June 1946, Vol. 19, No. 6, pp. 96–99.) A description of a system for positive control of remote communication equipment based on the transmission of suitably pulsed audio signals. The coding and decoding devices can be added to existing transmitter and mobile receivers respectively, enabling a control operator to call any of 84 stations by dialling a four-digit number.

621.396.662.2.029.6 **2732**
V.H.F. Coil Design.—A. H. Meyerson. (*Communications*, June 1946, Vol. 26, No. 6, pp. 46–47, 50.) The influences of coil shape and dimensions on the *Q* and on stability with temperature are examined experimentally for frequencies in the range 60–120 Mc/s. See also 66 of January (Meyerson).

621.396.682 **2733**
Type 1261-A Power Supply.—E. E. Gross. (*Gen. Radio Exp.*, March 1946, Vol. 20, No. 10, pp. 4–6.) Description and circuit diagram of an a.c.-operated power supply for instruments which use the U.S. Signal Corps BA48 Battery Block (Burgess Type 6TA60).

621.396.689 : 621.362 **2734**
Thermoelectric Generator for Portable Equipment.—J. M. Lee. (*Electronics*, May 1946, Vol. 19, No. 5, pp. 196..202.) Banks of chromel-P/constantan thermocouples embedded in ceramic material and heated by a gasoline burner give power outputs up to 20 W at 12 V. The couples have a useful life of about 2 000 hours, and the generator weighs about 2½ lb per watt output.

771.448.1 : 778.39 **2735**
A High-Power Stroboscope.—D. A. Senior. (*J. Sci. Instrum.*, April 1946, Vol. 23, No. 4, pp. 81–83.) Description of the circuit for producing, with a suitable discharge tube, 1 000 5-μs flashes a second for periods of several seconds, each with sufficient intensity to permit photography by reflected light of areas up to 50 sq. ft.

TELEVISION AND PHOTOTELEGRAPHY

621.397 **2736**
Practical Television.—R. A. Monfort. (*Radio News*, May 1946, Vol. 35, No. 5, pp. 38..120.) A semi-technical explanation of the principles and of the main parts of the equipment.

621.397 **2737**
4-Color Facsimile Transmission.—E. C. Thomson. (*Communications*, May 1946, Vol. 26, No. 5, pp. 32–34.) Four black and white pictures for use in a colour-superposition process are transmitted separately. Report of a partly successful experimental transmission by Cable & Wireless Ltd between England and Australia.

621.397 : 621.398 **2738**
Tele-Guided Missiles.—(*Electronic Industr.*, May 1946, Vol. 5, No. 5, pp. 62–65 . . 118.) A 325-line, 40-frames/sec television transmitter in a remotely controlled bomb sends a picture of the target to the bomb aimer. A "Vericon" tube is used in the transmitter, with stabilized supplies, and the iris of the lens system is controlled by the video signal to give a constant average signal intensity. Circuit diagrams are given.

621.397 : 621.398 **2739**
Television Equipment for Guided Missiles.—C. J. Marshall & L. Katz. (*Proc. Inst. Radio Engrs, W. & E.*, June 1946, Vol. 34, No. 6, pp. 375–401.) "A brief history of the technical problems associated with the development of compact airborne television equipment is outlined. The system provides resolution, linearity, and stability which approaches that obtained from broadcast equipment. Technical difficulties which arose after the completion of the equipment design are described. The final solution of these and other problems resulting from its installation in guided missiles are discussed. Photographs taken from the receiver screen during experimental flights are shown."

621.397.5 **2740**
Contribution to the Study of a Video Standard.—Y. Angel. (*Onde élect.*, Feb. 1946, Vol. 26, No. 227, pp. 60–73.) The definition of a television image is no longer limited by technique, but by economic and practical considerations. As a preliminary to the large-scale development of television it is necessary to agree on standards for the transmitted signal. For this purpose it is necessary to determine the optimum definition. The article gives a detailed analysis of the technical and subjective factors involved, and concludes that the standard should be (*a*) 1 200–1 300 lines with fourfold interlacing with the sequence 1-3-4-2 (or 1-3-2-4), or (*b*) 800–900 lines with twofold interlacing.

621.397.5 **2741**
1 015-Line Television Apparatus of the Compagnie des Compteurs, Montrouge.—P. Mandel. (*Onde élect.*, Jan. 1946, Vol. 26, No. 226, pp. 26–37.) A description of development work by the company; a theoretical and practical examination of the requirements for a television system with detail limited only by visual acuity. The work eventually realized a complete system, including the radio link, on 145 Mc/s. The development included apparatus permitting a wide range of scanning, both in lines and degree of interlacing. The final selection of interlaced scanning with 1 015 lines was based on tests over a wide range of values.

621.397.62 : 621.317.79 **2742**
A Television Signal Generator : Part 2—Monoscope and Video Circuits.—Hibberd. (*See* 2646.)

621.397.621 **2743**
Television Deflection Channels : Part 13.—E. M. Noll. (*Radio News*, May 1946, Vol. 35, No. 5, pp. 55 . . 147.) Description of circuits for generating and synchronizing the sweep signals, with particular reference to the G.E. Model 90 receiver. For other parts in this series on television circuits see 2349 of August (which should read part 14) and back references.

621.397.645.2 : 621.396.621.54 **2744**
I.F. Amplifiers in Television Receivers.—Kronenberg. (*See* 2534.)

621.397 + 621.396 **2745**
Modern Practical Radio and Television. [Book Review]—Quarrington. (*See* 2789.)

621.397 **2746**
Television, the Eyes of Tomorrow. [Book Review] —W. C. Eddy. Prentice-Hall, New York, 1945, 330 pp., $3.75. (*Electronics*, May 1946, Vol. 19, No. 5, p. 320.)

TRANSMISSION

621.396.61 **2747**
F.C.C. Approved A.M. Broadcast Transmitters.—R. G. Peters. (*Communications*, May 1946, Vol. 26, No. 5, pp. 26..34.) Some features, including circuit diagrams, of Collins, Gates and R.C.A. a.m. transmitters in the range 100 W–50 kW.

621.396.61 **2748**
Modern A.M. Transmitter Design. — W. E. Phillips & C. Probeck. (*Radio, N.Y.*, April 1946, Vol. 30, No. 4, pp. 30..54.) Some modern improvements, and a description (with circuit diagram) of a 250-W transmitter of quality comparable with that of the largest stations.

621.396.61 **2749**
High Power in Two Stages. — D. Mix. (*QST*, June 1946, Vol. 30, No. 6, pp. 13–17.) A crystal oscillator operating in the 3.5-, 7-, 14- and 28-Mc/s bands feeds an 800-W beam-tetrode transmitter. Shunt feed enables plug-in coils to be used with safety.

621.396.61 : 621.396.619.018.41 **2750**
Frequency Modulated Transmitters for Police and Similar Services.—E. P. Fairbairn. (*Electronic Engng*, July 1946, Vol. 18, No. 221, pp. 213–218.) The main advantage of f.m. is freedom from interference in dense traffic areas. Tests show that a small deviation at the transmitter gives widest service area at some expense in signal/noise ratio. A circuit diagram is given of a 10-W phase-modulated transmitter. Other illustrations show typical headquarters and mobile equipments. Crystal-controlled receivers are used. See also 2326 of August (Brinkley).

621.396.61+621.396.621] : 621.396.9 **2751**
Radar Technique.—F.L.D. (*Wireless World*, May 1946, Vol. 52, No. 5, pp. 154–156.) Review of papers on receivers and transmitters presented at the I.E.E. Radiolocation Convention.

621.396.61 : 621.396.933 **2752**
Unit-Type Multi-Channel Aircraft Ground Transmitter.—R. G. Peters. (*Communications*, June 1946, Vol. 26, No. 6, pp. 54–55.) Description of an equipment covering the ranges 200–540 kc/s, 2–20 Mc/s and 108–140 Mc/s, with circuit diagram of the 108–140 Mc/s, 200-W r.f. circuit.

621.396.61.029.5/.62 **2753**
1 000 Watt R.F. Amplifier for the Ham.—H. D. Hooton. (*Radio News*, May 1946, Vol. 35, No. 5, pp. 28..88.) Constructional details and performance. Uses two 4–125A tubes, and can be operated at frequencies up to 250 Mc/s.

621.396.61.029.56/.58 **2754**
A Beginner's Two-Stage Transmitter.—A. D. Middelton. (*QST*, July 1946, Vol. 30, No. 7, pp. 16–22, 126.) Constructional details of a mains- or battery-operated crystal-controlled c.w. circuit of very simple design for 3.5 and 7 Mc/s.

621.396.61.029.62 **2755**
A Mobile Rig for 50 and 28 Mc/s.—E. P. Tilton. (*QST*, June 1946, Vol. 30, No. 6, pp. 31–35, 110.) For economical operation from a car battery. A crystal-controlled "push-to-talk" transmitter uses midget valves with quick-heating filaments. A rotary converter is used for h.t.

621.396.61.029.62 **2756**
More Stations per Megacycle at Two Meters.—C. F. Hadlock & R. S. Hawkins. (*QST*, July 1946, Vol. 30, No. 7, pp. 61–66.) Constructional details of a 100-W push-pull crystal-controlled transmitter. A 5.4-Mc/s crystal is used with three frequency tripler stages and a power amplifier.

621.396.61 + 621.396.621].029.63 **2757**
Getting Started on 420 Mc/s.—W. F. Hoisington. (*QST*, June 1946, Vol. 30, No. 6, pp. 43–45.) Constructional details of a portable station comprising a half-wave-line transmitter feeding a 6-element array, a modulator, two power units, and a superregenerative receiver.

621.396.61+621.396.677] .029.63 **2758**
CQ 2 400 Mc/s : Transceivers and Antennas for the 13-Centimeter Band.—A. R. Koch & G. H. Floyd. (*QST*, July 1946, Vol. 30, No. 7, pp. 32–38.) A cavity-tuned 2C40 (lighthouse) tube is used as oscillator and superregenerative detector with a separate quench oscillator working at 100–250 kc/s. Constructional details are given for the tuned-plate tuned-grid cavity. Two aerials with parabolic reflectors are described; the smaller has a gain of about 25, the larger has a 7° beam and a power gain of about 200.

621.396.61.029.63 **2759**
A U.H.F. Ham Transceiver.—Queen. (*See* 2678.)

621.396.615 : 621.396.611.21 + 621.317.361 **2760**
Series-Resonant Crystal Oscillators.—Butler. (*See* 2524.)

621.396.615.17 **2761**
Power Pulse Generator.—M. Levy. (*Wireless Engr*, July 1946, Vol. 23, No. 274, pp. 192–196.) Pulses of the required amplitude are produced in a high impedance by a low-power pulse generator, then a succession of cathode followers increases the power and reduces the impedance, without materially reducing the amplitude. In the circuit given, 1 000-V pulses in 150 Ω with a peak pulse output power of 6.5 kW were obtained with an overall efficiency of about 20%.

621.396.619 **2762**
Features of Grid and Plate Modulation in New System.—(*Electronics*, May 1946, Vol. 19, No. 5, pp. 192..196.) Greater power output for size in portable transmitters is obtained by the use of grid modulation on negative half cycles and an additional r.f. amplifier as side-band generator on positive half cycles. A circuit diagram is given.

621.396.619 2763
Class "C" Grid Bias Modulation: Part 2.—W. W. Smith. (*Radio News*, May 1946, Vol. 35, No. 5, pp. 70..129.) Study of an inexpensive 100–125-W grid-modulated transmitter, based on 2379 of August.

621.396.619.018.41 2764
Frequency-Shift Keying.—G. G. (*QST*, June 1946, Vol. 30, No. 6, pp. 46–48.) While the system speeds up commercial traffic, the bandwidth required appears to be too large for its use in amateur communication.

621.396.619.018.41 2765
F.M. Carrier Stabilization: Part 1.—The General Electric and Federal Systems.—I. Queen. (*Radio Craft*, May 1946, Vol. 17, No. 8, pp. 537..549.) An illustrated simple account of the phasitron (see 1405 of May), and of the Miller-effect method.

621.396.619.018.41 2766
F.M. Transmitters using Phase Modulators.—N. Marchand. (*Communications*, June 1946, Vol. 26, No. 6, pp. 38–42.) Phase-modulator circuits used in commercial transmitters are analysed in relation to the principles described in 2767 below. Part 6 of a series.

621.396.619.018.41 2767
Phase to Frequency Modulation.—N. Marchand. (*Communications*, May 1946, Vol. 26, No. 5, pp. 36..58.) An analysis of the Armstrong method, and an account of the phasitron (see 1405 of May). Part 5 of a series. For other parts in this series see 2766 above, 2383 of August and back references.

621.396.619.018.41 2768
A New Angular-Velocity-Modulation System employing Pulse Techniques.—J. F. Gordon. (*Proc. Inst. Radio Engrs, W. & E.*, June 1946, Vol. 34, No. 6, pp. 328–334.) A two-valve multivibrator is controlled by synchronizing pulses from a crystal oscillator applied to the grid of one valve, and the relative duration of the conducting periods of the two valves is controlled by the amplitude of an audio-frequency modulating voltage applied to the grid of the second valve. The multivibrator output is differentiated and the positive pulses clipped, leaving a series of negative pulses which operate a class-C amplifier. An r.f. voltage is therefore produced, of which the phase is controlled by the phase of the pulses, and thus by the amplitude of the modulating voltage. In an experimental transmitter, a multivibrator frequency of 200 kc/s was multiplied to 105.6 Mc/s. Phase deviation was constant within $\pm$ 4% for modulation frequencies between 50 and 20 000 c/s and harmonic distortion was small.

621.396.619.018.41 : 621.396.611.21 2769
Crystal-Controlled Frequency Modulation.—A. Preskinsky. (*Wireless World*, May 1946, Vol. 52, No. 5, p. 170.) A letter commenting on 479 of February (Lewer).

621.396.619.16 : 621.385.38 2770
Thyratron Pulser Tube for Industrial Microwaves.—(*Electronics*, May 1946, Vol. 19, No. 5, pp. 70..180.) Application of a hydrogen thyratron in a line modulator circuit to new methods of plastic manufacture, high-speed welding, and electroplating. The system with a 4C35 tube can provide 100-kW pulses at a rate of 1 500 per sec, with average power 1 kW.

621.396.645.3.029.56/.58 2771
A Conservative Kilowatt.—D. Mix. (*QST*, July 1946, Vol. 30, No. 7, pp. 54–56.) Description of a push-pull amplifier for four amateur bands.

621.396.645.3.029.58 2772
Long Leads aren't Necessary.—G. W. Shuart. (*QST*, June 1946, Vol. 30, No. 6, pp. 55–57.) Parasitic oscillations are avoided in a push-pull tetrode amplifier for 28 Mc/s by careful arrangement of components and by neutralizing.

621.396[.65 + .71 + .812.3 2773
Radio Communication Developments.—A. H. Mumford. (*Nature, Lond.*, 19th Jan. 1946, Vol. 157, No. 3977, p. 83.) Summary of the inaugural address of the Chairman of the Radio Section, I.E.E. For full paper see 1102 of April.

VALVES AND THERMIONICS

621.385 2774
Radio Design Worksheet: No. 49.—Perveance.—(*Radio, N.Y.*, June 1946, Vol. 30, No. 6, p. 29.)

621.385 2775
The Electron-Optical Theory of Ultra-High-Frequency Oscillators.—P. Golubkoff. (*Zh. eksp. teor. Fiz.*, 1944, Vol. 14, Nos. 7/8, pp. 289–306.) There are three types of u.h.f. oscillators: retarding-field type, magnetrons, and electron-beam type. Various theories of these oscillators have been developed, but they are usually based on a study of the movement of a single (isolated) electron, and each applies to one type of oscillator only. The author proposes a new theory based on the following considerations. (1) The electronic processes and the mechanisms for sustaining oscillations in all types are identical. The primary mechanism is a continuous electron stream in which a process of phase focusing takes place, and this establishes the necessary interaction between the stream and the elements of the oscillatory system. Thus all types of oscillators can be regarded as electron-beam devices and interpreted by a single theory. (2) From the point of view of kinematics the physics of ultra-high frequencies can be regarded as a development as well as a practical application of electron optics.

From these considerations the author has developed a general method of investigation in which the conception of a moving focus in an electron stream is introduced, and the movement of the focus studied. It is possible, with the aid of this method, to interpret the main characteristics of u.h.f. oscillators of all types. A brief survey of an extensive theoretical and experimental investigation by the author is presented to support this claim. The following are the main points of the survey. The movement of an electron stream in the absence of a retarding field is examined, and the conclusions reached are applied to the case of klystrons. An analogy between klystrons and Barkhausen oscillators is established. The focusing of the stream in constant and in alternating retarding fields is discussed, and the main features of the Barkhausen circuit, such as the discreteness of the regions of oscillations, the position of the centres of these regions, the appearance of "dwarf" waves, etc., are explained. For a general case of the triode when the electron stream is acted upon by two electrodes at the boundaries of the focusing space (and not by one—the phase lens—as in the previous cases), a law is formulated governing the distribution of the centres of the regions of oscillations in the plane V_A, V_G. The focusing of the electron stream in a magnetic field is also considered, and a theoretical interpretation of the operation of the magnetron with a whole anode is given.

621.385 **2776**

Contribution to the Study of Reflex Velocity-Modulation Oscillators.—M. Kuhner & A. M. Gratzmuller. (*Onde élect.*, Jan. 1946, Vol. 26, No. 226, pp. 38–44.) A simple account of the theory of the reflection klystron, and a description of a number of types developed at the L.M.T. laboratories in 1943–1944. These embody glass-metal disk seals and external tunable cavities, tuning either by the screwing in of pistons, or by single or double rectangular plungers moving in rectangular cavities. One model of the latter type gave a maximum output of 200 mW at 149 mm wavelength, with an efficiency of 1%, and had a tuning range of 95 to 156 mm.

621.385 **2777**

Electron Ballistics in High-Frequency Fields.—A. L. Samuel. (*Bell Syst. tech. J.*, Jan. 1946, Vol. 25, No. 1, Supplementary page.) Corrections to 1396 of May.

621.385 : 621.396.9 **2778**

Radar Technique.—M. G. S. : H. B. D. (*Wireless World*, May 1946, Vol. 52, No. 5, pp. 146–151.) Review of papers on valves presented at the I.E.E. Radiolocation Convention.

621.385.1.032.216 : 537.583 **2779**

On a New Method of Measuring the Intensity of the Saturation Current in an Oxide Cathode.—R. Champeix. (*C. R. Acad. Sci., Paris*, 23rd May 1945, Vol. 220, No. 21, pp. 736–738.) A capacitor is discharged through the valve under test, once per second. The resulting impulsive current is passed through a non-inductive resistor of small value, and the potential drop so produced is used to trigger a thyratron which carries an adjustable grid bias. The bias is set to the maximum negative value at which the impulses will fire the discharge, and its value is used as a measure of the current in the valve under test.

Curves of electron current against applied voltage have the same general shape as those for pure tungsten, though higher voltages are necessary to produce saturation. The phenomenon of increasing saturation current with increase of applied voltage gradient at the cathode (Schottky's law) is much more marked for oxide cathodes than for pure metals.

621.385.16 **2780**

Recent Developments in Magnetron Technique.—G. Goudet. (*Onde élect.*, Feb. 1946, Vol. 26, No. 227, pp. 49–59.) A short account of the basic theory of the simple magnetron and of the multiple-cavity tube. The frequency and phase relationships of the latter are derived in terms of a closed ring of filter cells, and the field conditions by application of Maxwell's equations to an approximately equivalent system of plane parallel electrodes with equi-spaced gaps in one of them. The mechanism of oscillation is explained in terms of the interaction between the emitted electrons and the high-frequency fields in the annular interelectrode space.

621.385.16.029.64 **2781**

The Cavity Magnetron.—J. T. Randall. (*Proc. phys. Soc.*, 1st May 1946, Vol. 58, No. 327, pp. 247–252.) An address delivered to the Physical Society giving an historical outline of the development.

621.385.16.029.64 **2782**

An Introduction to Multi-Resonator Magnetrons.—R. Latham, A. H. King & L. Rushforth. (*Engineer, Lond.*, 5th & 12th April 1946, Vol. 181, Nos. 4708 & 4709, pp. 310–312 & 331–333.)

621.385.18 : 621.396.9 **2783**

The Gas-Discharge Transmit-Receive Switch.—A. L. Samuel, J. W. Clark & W. W. Mumford. (*Bell Syst. tech. J.*, Jan. 1946, Vol. 25, No. 1, pp. 48–101.) A detailed account of the purpose, design, construction, testing, method of application, and performance, of gas-filled resonant-cavity tubes used to protect radar receivers from damage due to the high-power transmitted pulse, and to prevent energy received in the dual-purpose aerial from being absorbed by the quiescent transmitter instead of by the receiver. The tubes operate by presenting a virtual short-circuit when discharged by the transmitted pulse, and an open-circuit when undischarged. A mathematical analysis of an idealized system is developed, and an account is included of the parameters of the coupling holes in the associated waveguides.

MISCELLANEOUS

621.3.078 **2784**

On Increasing the Stability of Self-Regulation by Means of Back Coupling.—V. Peshkoff. (*Zh. eksp. teor. Fiz.*, 1944, Vol. 14, No. 12, pp. 514–518.) Very often self-oscillations appear in automatic regulating systems. In many cases this can be prevented by introducing back coupling between the regulating indicator and the regulated mechanism. A detailed mathematical analysis is given as an example, confirmed by experiments, of Bancroft's thyratron thermostat (2253 of 1942) in which a transformer provides the required back coupling.

621.38/.39] (058.7) **2785**

1946 Electronic Engineering Directory.—(*See* 2610.)

621.396 **2786**

Engineers study F.M.—(*See* 2701.)

621.396.621.004.67 **2787**

Radio Servicing.—(*See* 2686.)

621.396.7 (058.7) **2788**

Radio Stations.—(*See* 2708.)

621.396 + 621.397 **2789**

Modern Practical Radio and Television. [Book Review]—C. A. Quarrington. W. H. Date (Ed.). The Caxton Pub. Co., London, 3 vols., 70s. (*Electronic Engng*, July 1946, Vol. 18, No. 221, p. 227.)

621.396 **2790**

Principles of Radio for Operators. [Book Review]—R. Atherton. Macmillan, New York, 1945, 344 pp., $3.75. (*Science*, 22nd March 1946, Vol. 103, No. 2673, p. 374.) "... outgrowth of the author's experience in training Navy men and women ... The general plane of instruction is excellent."