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CONTENTS

EDITORIAL. The Lorentz Transformations applied to the Problem of Two Electrons	155
SERIES-RESONANT CRYSTAL OSCILLATORS By F. Butler, B.Sc.	157
EQUIVALENT CAPACITANCES OF TRANSFORMER WINDINGS By W. T. Duerdoth, B.Sc. ..	161
PROPAGATION CHARACTERISTICS OF A UNIFORM LINE By I. F. Macdiarmid and H. J. Orchard	168
IRON-CORED LOOP RECEIVING AERIAL By R. E. Burgess, B.Sc.	172
CORRESPONDENCE	178
WIRELESS PATENTS	180
ABSTRACTS AND REFERENCES Nos. 1422—1743	A107-A132

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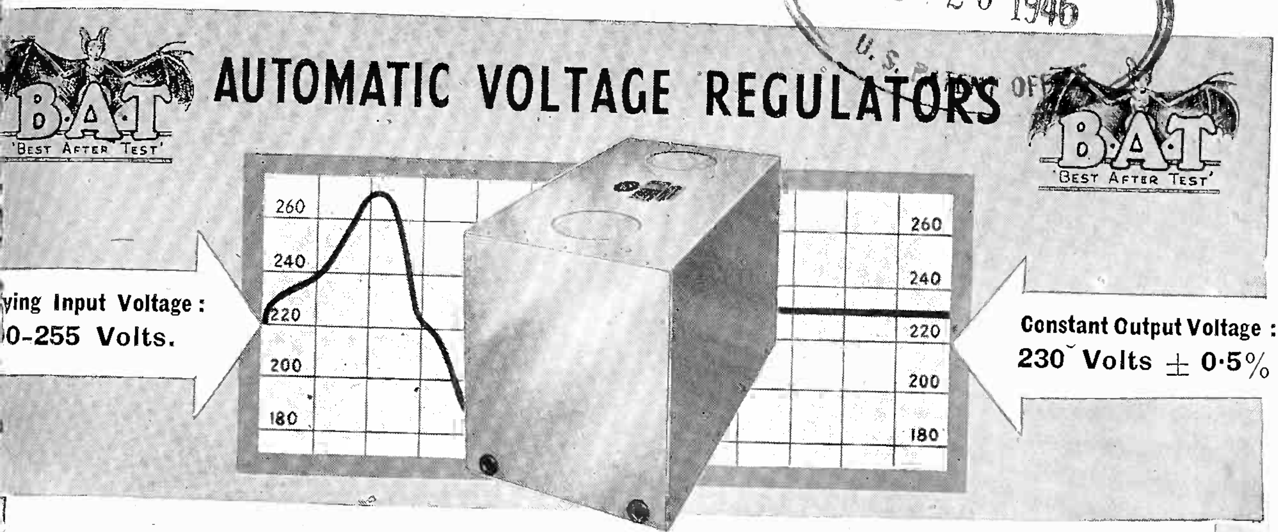

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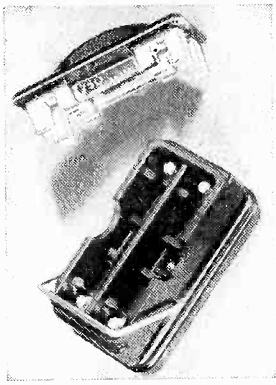

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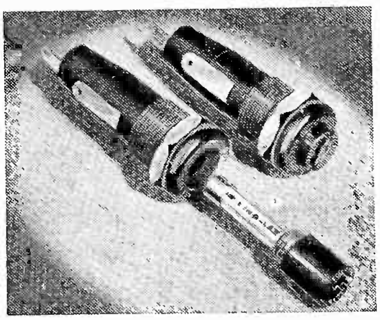

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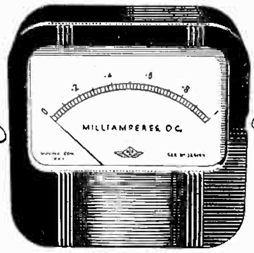

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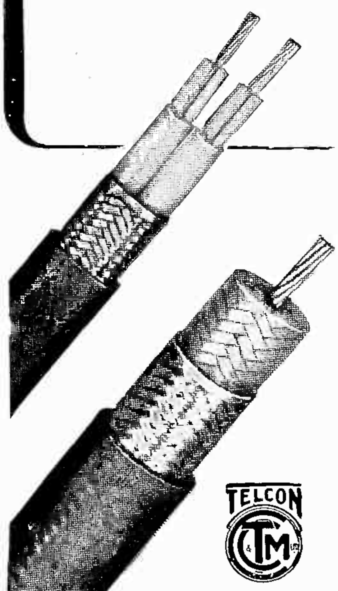

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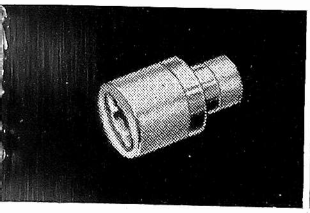

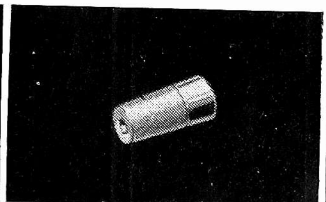

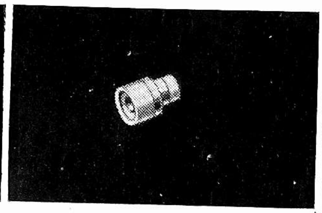

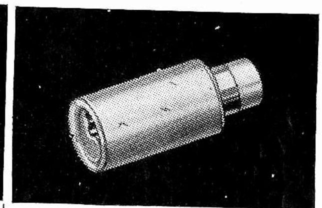

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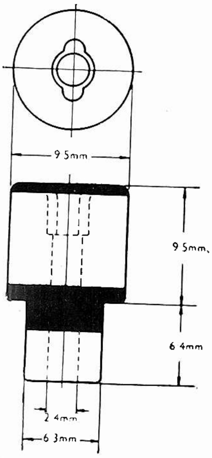

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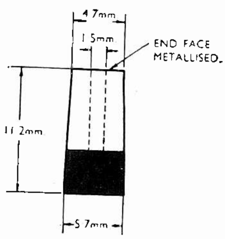

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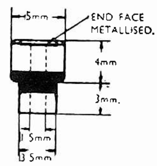

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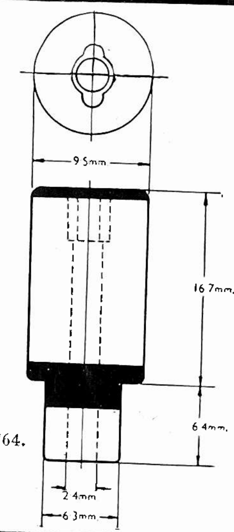

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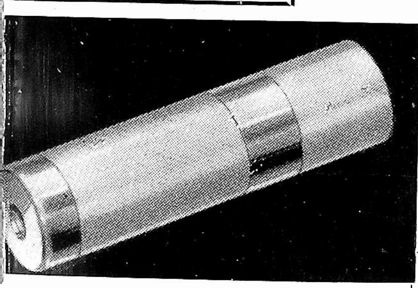

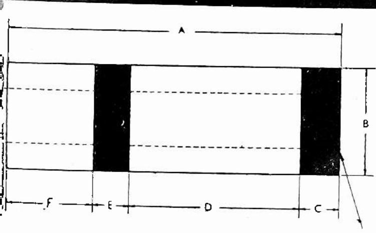

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R. 50769	GCTBC03	20.3	15.2	3.8	6.4	4.6	5.5	6.4
R. 50770	GCTBC04	38.1	10.2	5.1	15.7	6.4	10.9	4.1
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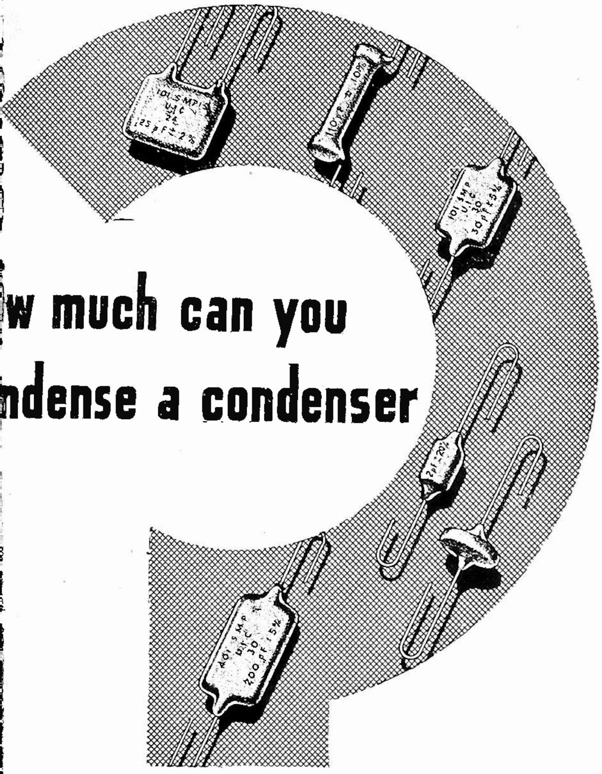

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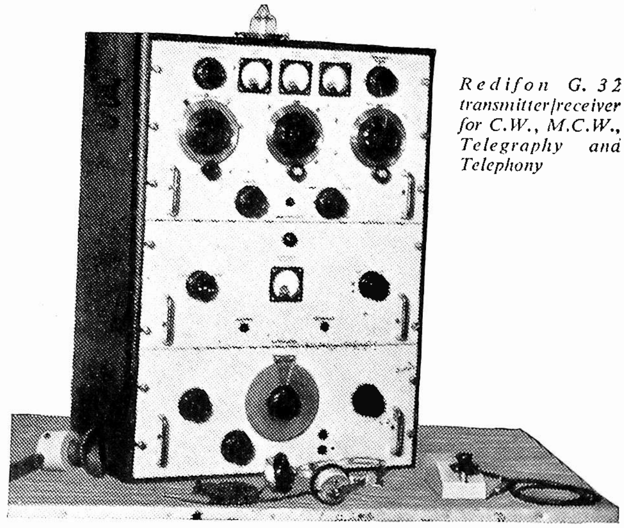

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JUNE 1946

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EDITORIAL

The Lorentz Transformations applied to the Problem of Two Electrons

IN the January Editorial we referred to the time-honoured question of two electrons moving side by side at high speed, which had been recently raised by a correspondent in the American journal, *Electrical Engineering*. On page 178 of this issue we publish a letter from R. E. Burgess of the National Physical Laboratory in which he refers to the principle of special relativity and applies the Lorentz transformations to the problem. We doubt very much, however, whether this resolution will satisfy the propounder of the question. As Mr. Burgess states, the ordinary physical laws are invariant, that is to say, the acceleration of the one electron away from the other is equal to the force divided by the mass after you have applied the Lorentz transformations to all the quantities concerned.

If v is the velocity of the two electrons relative to the observer, $k = (1 - v^2/c^2)^{-\frac{1}{2}}$ is slightly greater than 1.0 and the effective mass is increased in this ratio, whilst the resultant force due to the electric and magnetic fields is decreased in the same ratio. At first sight the increased mass and the decreased resultant force would suggest a decrease in the ratio of $1/k^2$ in the acceleration with which the electrons move apart, but apparently this is exactly what the acceleration would appear to be to the stationary observer. Lengths transverse to the direction of motion are unaltered but time intervals are shortened in the ratio $1/k$ and therefore the observed acceleration is modified in the ratio $1/k^2$.

In the *Philosophical Magazine* of June, 1945, there is a paper on the "Derivation of the Lorentz Transformations," by H. E. Ives of the Bell Telephone Laboratories, in which he attacks the very principle that Mr. Burgess invokes, viz., the principle of special relativity. He says, "The Lorentz transformations were obtained by Lorentz as a succession of *ad hoc* inventions to reconcile Maxwell's theory with the results of experiments on moving bodies. By Einstein they were derived after a discussion of the nature of simultaneity and the adoption of a *definition* of simultaneity which violates the intuitive and common-sense meaning of that term."

He then says: "It is proposed here to show that these transformations can be derived by imposing the laws of conservation of energy and of momentum on radiation processes as developed by Maxwell's methods. To do this we shall study the interaction of radiation and matter, specifically the impact of a train of radiation on a perfectly reflecting particle. The energy and momentum of the radiation will be obtained from electromagnetic wave theory; the resultant momentum and energy of the particle will be obtained by requiring the conservation of momentum and energy. It will be found that there is an apparent discrepancy in the equations; the resolution of this discrepancy demands that the *mass* of the particle vary with velocity. Next, the same impact experiment will be considered with the system in uniform motion. Again, a discrepancy will be encountered, and the resolution of this demands that the *lengths* and *time intervals* also vary with velocity . . . From these variations of dimensions the Lorentz transformations can be derived. *The space and time concepts of Newton and Maxwell are retained without alteration.*"

He finds that the mass M when moving is equal to $M_0(1 - v^2/c^2)^{-\frac{1}{2}}$ as stated above. He then assumes that the experiment is carried out on a platform *moving with respect to the wave transmitting medium*. This is presumably our old friend the ether, but Dr. Ives gives no indication as to the method of determining whether a body is moving with respect to it or not. We are reminded of the Kelvin Lecture given in 1925 by Sir J. H. Jeans on 'Electric Forces and Quanta'; in moving a vote of thanks Sir Oliver Lodge —who was a great believer in mediums— remarked that although the lecturer had *expressed* the facts of nature in a wonderful manner without referring to a medium, he had not *explained* them.

We cannot help feeling that Dr. Ives's argument is seriously weakened by this statement without any explanation. As the result of his calculations he finds that frequencies and lengths in the direction of motion are reduced in the ratio $1/k$, in fact, he always gets the right answer, which leads one to suspect that by 'moving with respect to the medium,' he means 'moving with respect to the observer,' and that every observer has his own medium, but it is not at all clear.

In discussing his method, Dr. Ives says: "The derivation of the Lorentz transformations here given has for its central feature the confronting of the Maxwell electromagnetic picture of radiation with the laws of conservation of energy and momentum. The derivation here given is strictly according to the tenets of classical physics. . . . No hybrid 'space-time' is invoked. . . . The indeterminances and impotences by which the 'Restricted Theory of Relativity' has been widely publicized . . . all follow as consequences of our resolution of the apparent conflict between the laws of the conservation of energy and momentum on the one hand and the laws of light propagation on the other, our resolution being in favour of the conservation laws. . . . The fact that the Lorentz transformations with all their consequences are deducible from these, classes the 'Special Principle of Relativity' as a superfluous hypothesis."

It is certainly very reassuring to know that Dr. Ives arrives at exactly the same results; it makes one wonder whether he is really adopting a different line of attack or merely clothing it in different words. This suspicion is strengthened by the vagueness of his references to movement relative to the medium.

G. W. O. H.

The Permeability of Iron-Dust Cores

IN this number we publish an article by R. E. Burgess which deals with the permeability of iron-dust cores. The author says "If the mass core contains uniform spherical particles of iron of permeability μ_0 insulated from each other, the effective permeability of the core is given by

$$\mu = \frac{(\mu_0 + 2) + 2p(\mu_0 - 1)}{(\mu_0 + 2) - p(\mu_0 - 1)} = \frac{1 + 2pM_0}{1 - pM_0}$$

where p is the fractional volume of the iron and $M_0 = (\mu_0 - 1)/(\mu_0 + 2)$. This equation can also be written in the form

$$\frac{\mu - 1}{\mu + 2} = p\frac{\mu_0 - 1}{\mu_0 + 2} = pM_0$$

In *Wireless Engineer* of January 1933 we considered this question and developed a formula on the assumption that the iron particles were cubes arranged symmetrically with parallel faces. The formula obtained was

$$\mu = \frac{\mu_0 - 2\delta(\mu_0 - 1)/3}{1 + (\mu_0 - 1)\delta/3}.$$

where δ is the fraction of the space occupied by insulating material, i.e. $\delta = 1 - p$.

Out of curiosity we calculated several points on Fig. 1 of Mr. Burgess's article to see how the values calculated on the cubical assumption compared with those calculated on the spherical assumption. To our surprise the values agreed so exactly that we suspected that the two formulae were really one and the same, and so it turned out. Putting $p = 1 - \delta$ we have

$$\mu = \frac{1 + 2pM_0}{1 - pM_0} = \frac{1+2(1-\delta)(\mu_0-1)/(\mu_0+2)}{1-(1-\delta)(\mu_0-1)/(\mu_0+2)}$$
$$= \frac{(\mu_0 + 2) + 2(1 - \delta)(\mu_0 - 1)}{(\mu_0 + 2) - (1 - \delta)(\mu_0 - 1)}$$
$$= \frac{3\mu_0 - 2\delta(\mu_0 - 1)}{3 + \delta(\mu_0 - 1)} = \frac{\mu_0 - 2\delta(\mu_0 - 1)/3}{1 + (\mu_0 - 1)\delta/3}$$

Hence whether the particles are assumed to be spherical or cubical, the approximations that are made lead to the same final formula.

Although in 1933 we thought that the cubical assumption was original and that all previous writers had assumed spherical or spheroidal particles, we subsequently discovered that Doebke, to whose work we gave a reference, had already in 1930 calculated the permeability for the cubical particles and had obtained a formula which was really the same as we obtained in 1933.

G. W. O. H.

SERIES-RESONANT CRYSTAL OSCILLATORS*

By F. Butler, B.Sc., A.M.I.E.E.

SUMMARY.—After a discussion of the equivalent circuit of a vibrating quartz crystal, associated with external series or parallel reactances, a number of new oscillator circuits are described. These all employ series resonance of the quartz crystal. Their frequency of operation is found to be slightly affected by associated external series reactance, in contrast with oscillators of the type using parallel crystal resonance, which are affected by reactances in parallel with the crystal. It is finally suggested that the same fundamental circuit principles can be employed in the design of crystal filters.

Introduction

ANALYSIS of the equivalent circuit of a vibrating quartz crystal shows that it is capable of simulating a series-tuned circuit at one particular frequency and that it exhibits parallel resonance at a slightly higher frequency. Most of the early crystal oscillators employed the parallel-resonant mode, the frequency of which is dependent on associated parallel reactances, e.g. the holder capacitance. More recently, oscillators have been developed using series resonance. These include the Meacham bridge-stabilized oscillator, and a two-valve circuit previously described by the writer. The frequency of these oscillators is scarcely affected by impedance in parallel with the crystal, but is sensitive to the effect of reactance in series with it. The attainable precision of frequency with the series-resonant oscillator is higher than that possible with the parallel-resonant alternative. The circuits to be described do not compare in frequency stability with the Meacham oscillator, but they deliver a very much larger power output.

* MS. accepted by the Editor, January 1946.

Equivalent Circuit of Quartz Crystal

Fig. 1 shows the electrical equivalent of a piezo-electric crystal in which :—

L = effective inductance of the crystal.
C = effective capacitance.
C_1 = holder capacitance.
X_s = series reactance (air-gap effects, wiring inductance or inserted reactance).

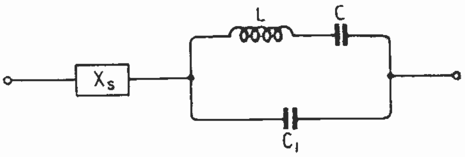

Fig. 1. *Equivalent circuit of a piezo-electric crystal.*

Neglecting resistances, analysis of this circuit shows that parallel resonance occurs at the frequency given by :—

$$\omega^2 = \frac{1}{L}\left(\frac{1}{C} + \frac{1}{C_1}\right)$$

It is clearly affected by the holder capacitance. The condition for series resonance is given by the following expression :—

$$\left(j\omega L + \frac{1}{j\omega C}\right) = \frac{-X_s/j\omega C_1}{X_s + 1/j\omega C_1}$$

In the particular case when X_s is due to an inductance L_1, $X_s = j\omega L_1$, and the condition becomes :—

$$\left(\omega L - \frac{1}{\omega C}\right) = \frac{L_1/C_1}{\omega L_1 - 1/\omega C_1}$$

The true series resonant frequency of the crystal alone is, therefore, only attained if L_1 is zero or if the inductive reactance is cancelled by tuning out with additional series capacitance.

If X_s is a capacitive reactance, given by $1/j\omega C_2$ then :—

$$\left(\omega L - \frac{1}{\omega C}\right) = \frac{1}{\omega(C_1 + C_2)}$$

The series-resonance frequency is therefore more closely approached as the external series inductance tends to zero or as the external capacitance is indefinitely increased.

Influence of Associated Circuits

In some of the oscillator circuits to be described, the crystal will be placed in series with the untuned primary of an R.F. transformer of which the secondary is tuned. It is of interest to calculate the input impedance and to estimate the probable effect on the series-resonant frequency of the crystal-transformer combination.

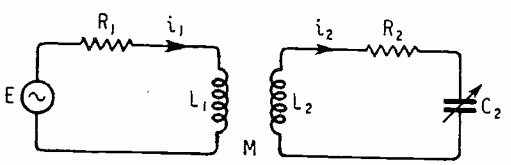

Fig. 2. *Equivalent circuit of an R.F. transformer.*

Writing down and solving the network equations, using the symbols shown in Fig. 2, the input impedance is given by :—

$$E/i_1 = Z_i = \left\{R_1 + \frac{\omega^2 M^2}{R_2^2 + (\omega L_2 - 1/\omega C_2)^2} \cdot R_2\right\} + j\left\{\omega L_1 - \frac{\omega^2 M^2}{R_2^2 + (\omega L_2 - 1/\omega C_2)^2} \cdot (\omega L_2 - 1/\omega C_2)\right\}$$

For secondary resonance, the equation simplifies to :—

$$Z_i = R_1 + \frac{\omega^2 M^2}{R_2} + j\omega L_1$$

The real component does not affect the frequency. The reactive term lowers the frequency below that of the crystal. Slight mistuning of the secondary can be used to cancel the inductive reactance $j\omega L_1$, this condition being indicated by maximum R.F. crystal current. To estimate the probable effect of the reactance $j\omega L_1$ on the final frequency, the electrical constants of the crystal must be known. For a high-frequency crystal, the effective inductance is commonly of the order of one henry. The inductance L_1 is unlikely to exceed about 10 μH, so that the frequency change can be expected to be about five parts in one million. Mistuning of the secondary puts larger transferred reactances in series with the crystal and causes proportionately greater frequency changes, particularly when the transferred reactance is inductive.

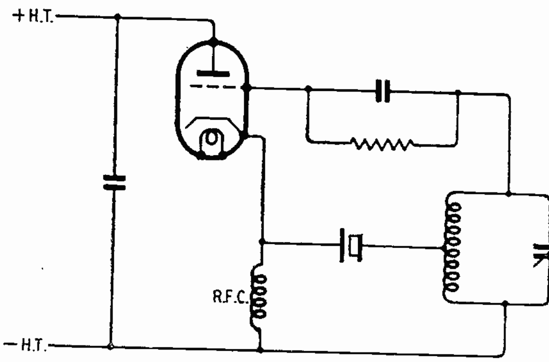

Fig. 3. *"Earthed anode" type of Hartley oscillator.*

Practical Series-Resonant Oscillator Circuits

The three basic circuits to be discussed are shown in Figs. 3, 4 and 5. All involve the use of a cathode-driven amplifier, connected regeneratively to ensure self-oscillation. They are merely variations of the Hartley circuit, with the inclusion of a quartz crystal between the cathode tap or coupling coil and the valve cathode. The input impedance of the amplifier and the output impedance of the tapped coil or R.F. transformer are both low in value. At series resonance, the crystal appears as a low resistance and permits the maintenance of oscillations. At off-resonance frequencies the crystal simulates a high reactance linking the low-impedance source and load. Attenuation and phase shift occur and oscillations cease.

The radio-frequency choke in each cathode lead serves merely to provide a D.C. path for the valve anode current without short-circuiting the R.F. driving voltage. Its value is not critical. In low-power oscillators, a resistance of the order of 20,000 ohms may be substituted provided that the grid leak is in this case returned to the cathode so as to avoid over-biasing the valve.

As they stand, all three circuits are only suitable for use with low-frequency crystals. Subject to this limitation, they give a satisfactory performance, and oscillate readily even with poor quality crystals, difficult to operate in standard circuits. Figs. 3 and 4 are convenient for use with symmetrically-mounted crystals of the reversible plug-in type. Fig. 5 is best for use with precision crystals of which the holders are usually designed to have one side earthed.

At high frequencies it is found that the holder capacitance (C_1 in Fig. 1) exercises a shunting effect and oscillations occur at frequencies uncontrolled by the crystal. There are two standard methods of dealing

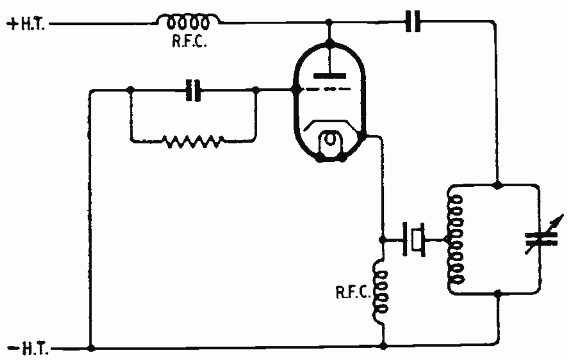

Fig. 4. *"Earthed grid" Hartley oscillator.*

with difficulties of this kind. Neutralization may be employed, or alternatively, the undesired capacitance can be tuned out by parallel inductance, in which case the capacitive reactance is removed, and a high resistance substituted. In the present case, the neutralization technique will be employed.

Figs. 6 and 7 show practical circuits. In the first a neutralizing winding is coupled to the main tuned circuit, and is preferably wound on the same coil former. The number of turns should be equal to those included between the crystal tap and earth. In this case, stability is ensured at all frequencies, other than at the desired series resonance frequency of the crystal, when the neutralizing capacitor setting is equal to the crystal holder capacitance. Over-neutralization gives a measure of control of the R.F. oscillator excitation voltage. Fig. 7 shows a circuit which has the practical advantage that one side of both crystal and neutralizing capacitor can be earthed. In both cases, the neutralizing-bridge balance is upset at the series-resonance frequency of the crystal. At this frequency the crystal resistance is very low and regenerative feedback occurs,

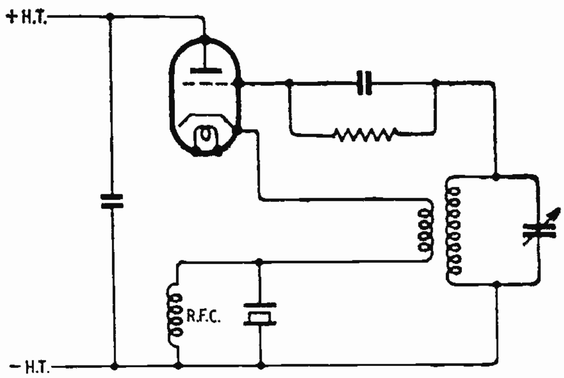

Fig. 5. *"Earthed anode" oscillator with cathode-circuit reaction coil.*

sufficient to start oscillations. In Fig. 6 the crystal-tapping point on the main inductance should be positioned to include one-fifth to one-tenth of the total turns on the main winding. The corresponding coupling coil in Fig. 7 should be similarly proportioned. In this circuit, an earthed electrostatic screen may be provided between the two windings.

A feature of all the oscillator circuits so far described is that the correct tuning point is indicated, with leak and condenser bias, when the valve anode current falls to a

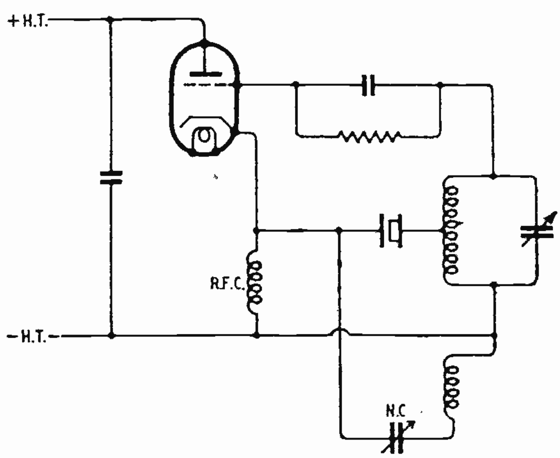

Fig. 6. *A neutralized Hartley oscillator of the "earthed anode" type.*

minimum. This is in contrast with conventional crystal circuits, which become unstable as this condition is approached. The dip in anode current is pronounced and sharply defined, though the narrow crevasse is not quite symmetrical. Measurements show that minimum anode current does not exactly correspond to maximum radio-frequency current through the crystal. The latter is the correct setting, but the difference is of no practical consequence, except when the highest precision is required from the oscillator.

It is possible to use in the oscillators a number of parallel-connected crystals of different resonant frequencies, provided that these frequencies are separated by at least a few parts in one thousand. Each crystal is selected in turn, merely by tuning the main variable capacitor.

Switching is not required, as there is no apparent interaction. As many as five crystals have been employed at once. For each operating frequency, there is a corresponding dip in anode current. An increase

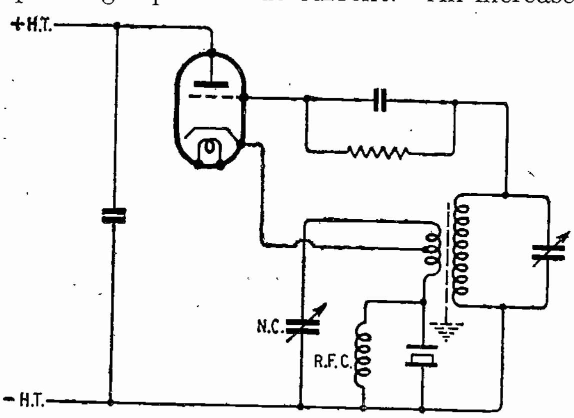

Fig. 7. *With a cathode-circuit reaction coil neutralization of the crystal holder capacitance can be effected in the manner shown here.*

in the neutralizing capacitance is required to compensate for the effect of each added crystal. In actual practice, perfect stability is maintained when two or three crystals are inserted in or removed from the oscillator.

Oscillators and Harmonic Amplifiers

The circuits of Figs. 6 and 7 can be conveniently adapted for harmonic operation. Tetrode valves can be employed in place of the triodes shown, a circuit tuned to the desired harmonic being connected in the anode circuit. With a 6F6 valve operating at an anode potential of 400 volts, with a 200-volt screen supply, a grid leak of 30,000 ohms and a grid capacitor of 100 micro-micro-farads the radio-frequency crystal current is 80 mA. These figures were obtained with a high-grade crystal of 2,600 kc/s, the third harmonic being selected in the anode circuit. Under these conditions, the D.C. anode current was 25 mA.

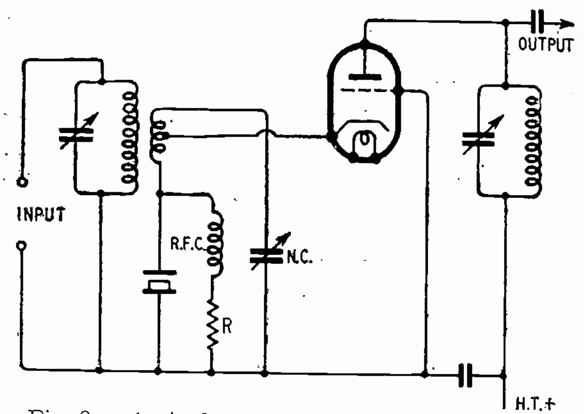

Fig. 8. *A single-peaked filter using a cathode-input stage can take this form.*

Crystal-Filter Circuits

No experimental work has been done on quartz filters but two circuits are suggested in Figs. 8 and 9 which show some promise. The first is a single-peaked filter ; the second has band-pass characteristics. The low input impedance of the earthed-grid triode amplifier seems likely to give a satisfactory impedance match without the use of auxiliary transformers. In Fig. 9 the coil centre-tap and the cathode connections may be interchanged to permit earthing one side of each crystal.

The use of an earthed-grid amplifier makes it possible to employ a triode valve with the filter. From the point of view of impedance matching and amplifier gain, a pentode may

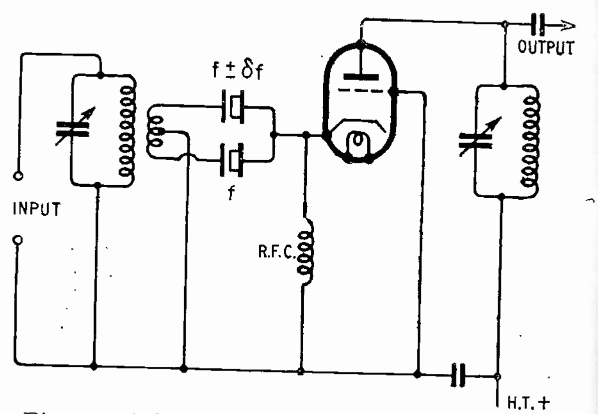

Fig. 9. *A band-pass circuit with two crystals does not require neutralizing if the shunt capacitances on the two crystals are identical.*

be preferred. Cathode bias may be used, the correct value being obtained, if necessary, by the inclusion of a resistance *R* to supplement that of the R.F. choke.

EQUIVALENT CAPACITANCES OF TRANSFORMER WINDINGS*

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SUMMARY. The paper shows that the distributed capacitances between windings, or windings and screens, of transformers may be represented by lumped capacitances provided that the magnetic coupling between the turns of a winding is perfect. Expressions have been obtained for the equivalent capacitances of a number of different arrangements including windings in layers, sections and with screens.

1. General

A KNOWLEDGE of the equivalent capacitance of a transformer winding is essential if the performance of the transformer is to be predictable. This knowledge is particularly necessary when the transformer has to pass a wide band of frequencies, as often occurs in telecommunication work. It is sometimes desirable to design the transformer as a filter section giving low-pass or band-pass characteristics as may be required, in which case the leakage inductance, shunt inductance and self-capacitances of the windings can be employed together with added components to form the elements of the filter. By these means, a maximum band-width can be passed without frequency distortion. This paper, however, considers only the equivalent capacitances of the winding.

Capacitance between transformer windings, and between windings and screens, consists of distributed capacitances along the whole windings and the effect of these capacitances depends upon the voltages between the corresponding points on the windings or screens.

It is possible to express the distributed capacitance as a lumped capacitance across the whole winding and relate this lumped capacitance to a total capacitance which can be easily measured at a low frequency where the inductance of the winding can be neglected.

Capacitances between turns on the same layer of a winding also exist but these are usually negligible in comparison with inter-layer and layer to screen capacitances.

If it is assumed that the windings or screens form coaxial cylinders, then it becomes necessary to estimate the capacitance C between such cylinders and in this analysis the following expressions have been employed.

$$C = \frac{0.0885kSD}{\Delta} \mu\mu F \quad \ldots \quad \ldots \quad (1)$$

Where k = permittivity of the dielectric
S = mean circumference (cm)
D = length of cylinder (cm)
Δ = distance between cylinders (cm)

This is true when the dimensions are such that the capacitor may be regarded as parallel plates extending to infinity but will be in error in cases where D or S are not large compared with Δ in which case the following will be more accurate.

$$C = \frac{1.1kD}{2\log_e r_2/r_1} \mu\mu F \quad \ldots \quad \ldots \quad (2)$$

This is true for two coaxial cylinders of radii r_1 and r_2 extending to infinity.

Both (1) and (2) are in error when finite capacitors are considered, since a modified field occurs at the boundaries of the plates and gives rise to an increased capacitance; this effect is usually referred to as "fringing," but will not be of importance when DS is large compared with Δ.

However, this limitation must be borne in mind when practical measurements are compared with the theoretical expressions obtained below.

2. Equivalent Circuit for an Unbalanced Single-layer Winding with Screen

Fig. 1 represents a single-layer winding AB with one end B connected to an adjacent screen. When connection B is broken, the total winding-to-screen capacitance C_0 can be measured at a low frequency. Alternatively, C_0 may be calculated using expressions (1) or (2).

An equivalent circuit can be developed

* MS accepted by the Editor, January 1946.

representing the transformer. The following symbols will be employed :

L = total length of wire in winding
x = length of wire between section δx and B
l = length of one turn
N = number of turns
R = a constant, proportional to the reluctance of the magnetic circuit.
L_0 = inductance = N^2/R.

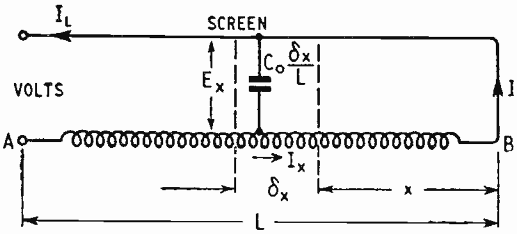


Fig. 1 (above). *A single-layer winding AB has the end B connected to a screen.*

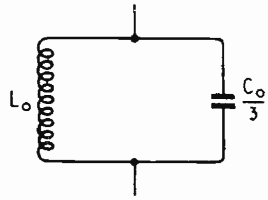


Fig. 2 (right). *In the equivalent circuit the effective capacitance is one-third of the total.*

Assume that all turns are perfectly coupled, i.e., the same flux is linked with all turns or the field strength is uniform throughout the coil. Then the voltage falls linearly along the winding and

$$E_x = E_L\, x/L \quad \ldots \quad \ldots \quad \ldots \quad (3)$$

Then flux produced by a small length of wire $\delta x = \frac{I_x \delta x}{Rl}\, 10^8$ maxwells.

Therefore, total flux

$$= \frac{1}{Rl}\int_0^L I_x dx \times 10^8 \text{ maxwells} \quad \ldots \quad (4)$$

Thus for sinusoidal excitation

$$\text{Applied e.m.f.} = E_L = j\omega \frac{N}{Rl}\int_0^L I_x\, dx \qquad (5)$$

Considering a small section

$$\frac{dI_x}{dx} = E_x \frac{j\omega C_0}{L} \quad \ldots \quad \ldots \quad \ldots \quad (6)$$

Then from (3), (5) and (6)

$$\frac{dI_x}{dx} = \frac{j\omega C_0 x E_L}{L^2}$$

$$\therefore \quad I_x = \frac{j\omega C_0 x^2 E_L}{2L^2} + K \quad \ldots \quad \ldots \quad (7)$$

where K is a constant of integration.

Substitute (7) in (5)

$$E_L = \frac{j\omega N}{Rl}\int_0^L \left(j\omega \frac{C_0 E_L x^2}{2L^2} + K\right)dx$$

$$= \frac{j\omega N}{Rl}\left[\frac{j\omega C_0 x^3}{6L^2} E_L + Kx\right]_0^L$$

$$= \frac{j\omega L_0}{Nl}\left(\frac{j\omega C_0 L}{6} E_L + KL\right)$$

$$= jwL_0\left(\frac{j\omega C_0}{6} E_L + K\right)$$

$$\therefore \quad K = -E_L\left(\frac{j\omega C_0}{6} - \frac{1}{j\omega L_0}\right)$$

$$\therefore \quad I_x = \frac{j\omega C_0 x^2}{2L^2} E_L - E_L\left(\frac{j\omega C_0}{6} - \frac{1}{j\omega L_0}\right)$$

$$= \left(\frac{j\omega C_0}{3} + \frac{1}{j\omega L_0}\right) E_L$$

$$Z_L = \frac{E_L}{I_L} = \frac{j\omega L_0}{1 - \omega^2 L_0 C_0/3} \quad \ldots \quad (8)$$

The equivalent circuit is shown in Fig. 2 and the effective capacitance is $\frac{1}{3}$ of the total capacitance.

We also have

$$\frac{I_0}{E_L} = -j\omega\frac{C_0}{6} + \frac{1}{j\omega L_0} \quad \ldots \quad \ldots \quad (9)$$

Where I_0 is the current at B.

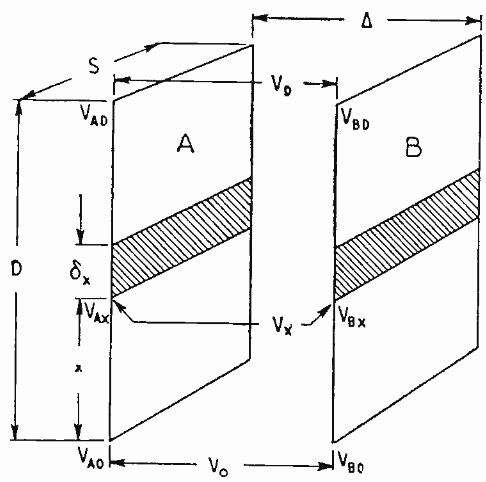


Fig. 3. *Adjacent screens or windings can be represented by plates A and B.*

We can thus regard the winding as a parallel capacitance and inductance, provided that the coupling between each turn of the winding is perfect. However, at a very high frequency the leakage flux and the capacitance between adjacent turns become f importance, and the impedance of the vinding passes through a number of series nd parallel resonances.

In the above treatment, we have proved hat the distributed capacitance is equivalent o a lumped capacitance across the whole vinding. Only the simple case of a winding vith one end connected to the screen has een treated, but capacitances between vindings at different alternating potentials ould be treated in a similar way. However, we assume that all cases are reducible to a mple lumped capacitance, then the following known method based on energy conideration only is more suitable for determining its magnitude.

. Calculation of Equivalent Lumped Capacitance

Consider two plates A and B representing wo adjacent screens or windings of area) $\times S$, as shown in Fig. 3.

$$\text{Total capacitance} = 0.0885\frac{kSD}{\Delta} = C_0\,\mu\mu\text{F} \quad \ldots \quad (10)$$

Energy stored $= W = \dfrac{C_0V^2}{2}$ joules, when he potential difference is V over the whole rea.

Consider the case where the potential ifference V_x is a linear function of x.

Then the energy stored in δx is given by

$$\delta W = \frac{0.0885kS}{\Delta} \cdot \frac{\delta x}{2} \cdot V_x^2 \times 10^{-12}\text{ joules}$$

Total energy

$$= W = \frac{0.0885kS}{2\Delta}\int_0^D V_x^2\,dx \times 10^{-12}\text{ joules}$$

$$V_x = V_{Bx} - V_{Ax} = \left[V_{B0} + (V_{B_D} - V_{B0})\,x/D\right] - \left[V_{A0} + (V_{A_D} - V_{A0})\,x/D\right] \ldots \quad (11)$$

(This assumes that the voltage changes nearly as assumed to be the case in the ast section.)

$$\therefore \quad V_x = V_0 + (V_D - V_0)\,x/D$$

$$\int_0^D V_x^2\,dx = \int_0^D [V_0 + (V_D - V_0)\,x/D]^2\,dx$$

$$= (D/3)\left[V_0^2 + V_0V_D + V_D^2\right]$$

$$\therefore \quad W = \frac{0.0885kS}{2\Delta} \cdot (D/3)\left[V_0^2 + V_0V_D + V_D^2\right] 10^{-12}\text{ joules}$$

If one of the windings has an applied voltage of V_p then the equivalent lumped capacitance is given by

$$C_p = \frac{2W}{V_p^2}10^{12}\,\mu\mu\text{F}$$

$$= \frac{0.0885kS}{\Delta V_p^2} \cdot D/3\left[V_0^2 + V_0V_D + V_D^2\right]\mu\mu\text{F}$$

$$C_p = \frac{C_0}{3V_p^2}(V_0^2 + V_0V_D + V_D^2) \quad \ldots \quad (12)$$

When $V_0 = 0$ and $V_D = V_p$ as in the previous section, then $C_p = C_0/3$.

4. Equivalent of Interlayer Capacitance of Single-layer Transformers

In this section the following additional symbols are employed :—

V_p = voltage on primary winding

V_s = voltage on secondary winding

N = V_s/V_p

C_p = equivalent lumped capacitance across primary

C_s = equivalent lumped capacitance across secondary

C_0 is the low-frequency capacitance measured between the primary winding and screen or secondary windings with ends disconnected.

Case A. Unbalanced windings without screen, earth connections at similar ends; Fig. 4(*a*).

$$\text{From equation (12) } C_p = \frac{C_0}{3}\frac{(+V_s - V_p)^2}{V_p}$$

$$C_p = (N - 1)^2\,C_0/3$$

If secondary is wound in same direction as primary, as shown, then sign of V_s is positive and vice versa.

If N is small then $C_p \approx C_0/3$

If N is large then $C_p \approx N^2C_0/3$

or $C_s \approx C_0/3$

If N is unity then $C_p = 0$ if the secondary is wound in the same direction as the primary, which is self evident since only capacitance between layers is considered.

Case B. Unbalanced windings without screen, earth connected to opposite ends; Fig. 4(*b*).

$$C_p = [V_0^2 + V_0V_D + V_D^2]\,C_0/(3V_p^2)$$

$$C_p = [1 + N + N^2]\,C_0/3$$

If the secondary is wound in the same direction as the primary, then the sign of V_s is negative, as shown.

If N is small then $C_p \approx C_0/3$

If N is large then $C_p \approx N^2C_0/3$

or $C_s = C_0/3$

If N is unity then $C_p = C_0$

if the windings are in the same direction.

Case C. Unbalanced windings with screen, earths at similar ends ; Fig. 4(*c*).

Advantage can be obtained by winding both windings in same direction.

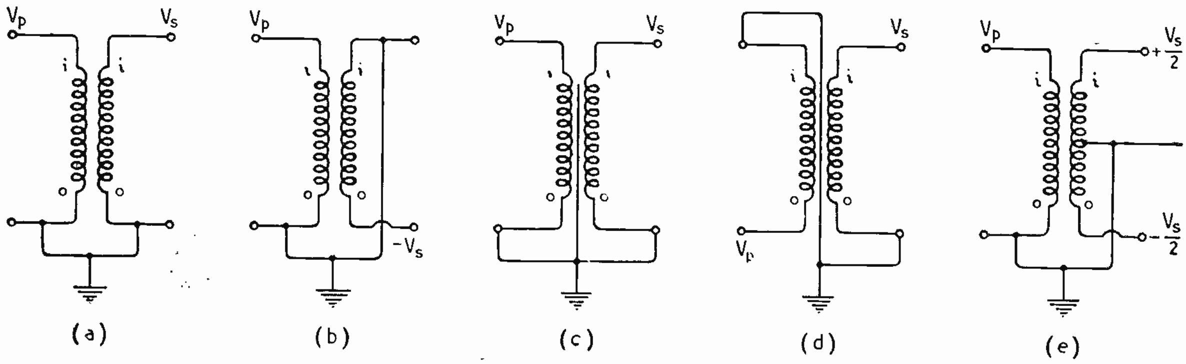
Case F. Balanced to unbalanced windings with screen ; Fig. 4(*f*).

$$C_p = (1 + N^2/4)\, C_0/3$$

Case G. Balanced to balanced windings unscreened ; Fig. 4(*g*).

$$C_p = (1 - N)^2\, C_0/12$$

With appropriate choice of sign $C_p = 0$ when $N = 1$.



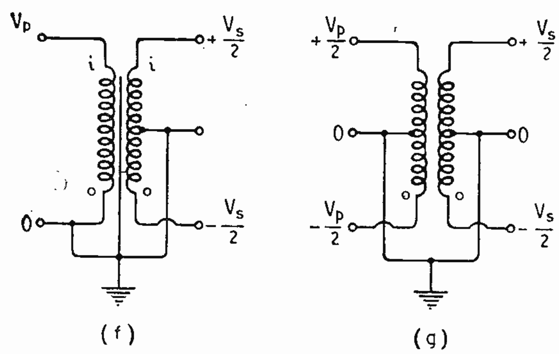


Fig. 4. *Various arrangements of winding connections and screens are shown : (a) Windings unscreened and unbalanced ; similar ends earthed ; (b) windings unscreened and unbalanced, opposite ends earthed ; (c) windings screened and unbalanced ; similar ends earthed; (d) windings screened and unbalanced; opposite ends earthed ; (e) windings unbalanced to balanced ; no screen ; (f) windings unbalanced to balanced ; with screen ; (g) windings balanced to balanced ; no screen.*

Assuming the screen to be equidistant from both windings.

$$C_p = [1 + N^2]\, C_0/3$$

If N is small then $C_p = C_0/3$

If N is large then $C_p \approx N^2C_0/3$

or $C_s = C_0/3$

If N is unity then $C_p = 2C_0/3$

Case D. Unbalanced windings with screen, earth to opposite ends of windings ; Fig. 4(*d*) Identical to *Case C.*

Case E. Unbalanced to balanced windings without screen ; Fig. 4(*e*).

$$C_p = [1 - N/2 + N^2/4]\, C_0/3$$

When N is small $C_p = C_0/3$

When N is large $C_s = C_0/12$

or $C_p = C_0N^2/12$

When N is unity $C_p = C_0/4$

5. Multilayer Coils

5.1 *Multilayer Coil without a Screen.*

Consider a coil, as in Fig. 5, consisting of n layers each separated by Δ_1. Area of each layer $= SD$. Let $C_0 = L.F.$ capacitance between any two adjacent layers when the interconnection is broken.

The total equivalent lumped capacitance may be estimated by summing the energies stored between each pair of layers.

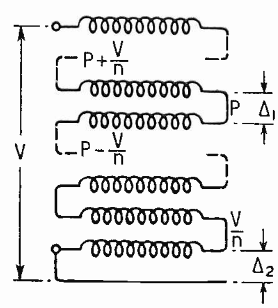


Fig. 5. *Capacitance of a multilayer coil.*

$$W = \frac{2}{3n^2} \cdot \frac{KV^2}{\Delta_1}$$

Where $K = 0.0885SDk$ (The value V_D of Fig. 3 is $2V/n$ in this case.)

There are $(n - 1)$ pairs of layers

$$\therefore \quad \Sigma W = \frac{2(n-1)}{3n^2} \cdot \frac{KV^2}{\Delta_1}$$

and since $C_0 = K/\Delta_1$

The equivalent lumped capacitance C is given by

$$C = \frac{2\Sigma W}{V^2}$$

$$= \frac{4C_0}{3n^2}(n-1) \quad \ldots\ldots\ldots \text{ from } (12)$$

5.2 *Multilayer Coil with Screen connected to the Low-potential end.*

Consider the case where a screen spaced Δ_2 from the first layer is connected to the low potential end of the winding.

Energy stored between first layer and screen

$$W_1 = \frac{1}{6n^2} \cdot \frac{KV^2}{\Delta_2}$$

Energy stored between the layers

$$\Sigma W_2 = \frac{2(n-1)}{3n^2} \cdot \frac{KV^2}{\Delta_1}$$

$$\therefore \quad C = C_0\left[\frac{4(n-1)}{3n^2} + \frac{1}{3n^2} \cdot \frac{\Delta_1}{\Delta_2}\right]$$

If $\quad \Delta_1 = \Delta_2$

$$C = \frac{4n-3}{3n^2} C_0$$

6. Multilayer Coils with Sections

6.1 *Without Screen*

Where there are G sections, each of n layers, but no screen, then by a similar method we have :—

Energy stored between pair of layers

$$= \frac{2KV^2}{3(Gn)^2\Delta_1}$$

Total energy for each section

$$= \frac{2KV^2(n-1)}{3(Gn)^2\Delta_1}$$

Total energy for G sections

$$= \frac{2KV^2(n-1)G}{3(Gn)^2\Delta_1}$$

$$\therefore \quad C = \frac{4(n-1)}{3Gn^2} C_0$$

6.2. *With a Screen*

A general expression for a coil wound on to a screen and divided into sections each containing several layers is very involved, but the particular case shown in Fig. 6 of four sections with n layers each, is treated below :—

Energy stored between screen and

$$\text{section A} = \frac{KV^2}{6\Delta_2} \cdot \frac{1}{16n^2}$$

Energy stored between screen and

$$\text{section B} = \frac{KV^2}{6\Delta_2} \cdot \frac{3n^2+3n+1}{16n^2}$$

Energy stored between screen and

$$\text{section C} = \frac{KV^2}{6\Delta_2} \cdot \frac{12n^2+6n+1}{16n^2}$$

Energy stored between screen and

$$\text{section D} = \frac{KV^2}{6\Delta_2} \cdot \frac{27n^2+9n+1}{16n^2}$$

Total energy stored between screen

$$\text{and coil} = \frac{KV^2}{6\Delta_2} \cdot \frac{3(14n^2+6n+1)}{16n^2}$$

Total energy stored between layers

$$\text{of coil} = \frac{KV^2}{6\Delta_1} \cdot \frac{n-1}{n^2}$$

$$\text{Total energy} = \frac{KV^2}{6}\left[\frac{n-1}{\Delta_1 n^2} + \frac{3(14n^2+6n+1)}{\Delta_2 16n^2}\right]$$

$$\therefore \quad C = \frac{C_0}{3}\left[\frac{n-1}{n^2} + \frac{3(14n^2+6n+1)\Delta_1}{16n^2\Delta_2}\right]$$

If $\quad \Delta_1 = \Delta_2$

$$C = C_0 \frac{42n^2+34n-13}{48n^2}$$

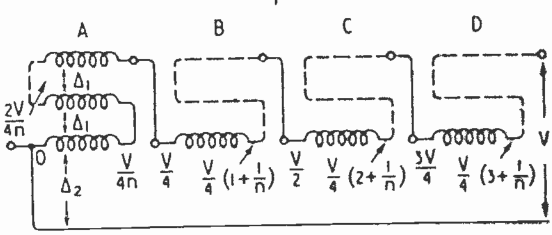


Fig. 6. *Capacitance of a coil of 4 sections with a screen.*

7. Disposition of Balanced Winding with respect to a Screen

In order to compare the various methods of winding balanced coils, C_0 has been chosen as the capacitance to a single screen of the whole winding laid out a distance Δ_2 from the screen. Thus, if we assume a constant screen area, the value of C_0 for a winding of n layers will be n times that of a single layer winding.

7.1 *Single Screen Only.*

Case A. Fig. 7(*a*).

This case has been dealt with in section 3.

$$C = C_0/12$$

Case B. Fig. 7(*b*).

Let Δ_2 be spacing between adjacent windings

Then energy, coil to screen $= \dfrac{KV^2}{48\Delta_2}$

Energy between layers in coil $= \dfrac{KV^2}{12\Delta_1}$

Total energy $= \dfrac{KV^2}{12}\left[\dfrac{1}{4\Delta_2} + \dfrac{1}{\Delta_1}\right]$

$$\therefore \quad C = \frac{K}{6\Delta_2}\left[\frac{1}{4} + \frac{\Delta_2}{\Delta_1}\right]$$

$$\text{or} \quad C = \frac{C_0}{24}\left[\frac{1}{4} + \frac{\Delta_2}{\Delta_1}\right] \quad \ldots\ldots$$

Since $C_0 = \dfrac{4K}{\Delta_2}$

If $\Delta_1 = \Delta_2$ $\quad C = \dfrac{5}{96} C_0$

Case C. Fig. 7(*c*).

Energy, coil to screen $= \dfrac{KV^2}{48\Delta_2}$

Energy, between layers $= \dfrac{13KV^2}{48\Delta_1}$

Total energy $= \dfrac{KV^2}{48}\left[\dfrac{1}{\Delta_2} + \dfrac{13}{\Delta_1}\right]$

$$\therefore \quad C = \frac{K}{24\Delta_2}\left[1 + 13\frac{\Delta_2}{\Delta_1}\right]$$

$$\text{or} \quad C = \frac{C_0}{96}\left[1 + 13\frac{\Delta_2}{\Delta_1}\right] \ldots$$

Since $C_0 = \dfrac{4K}{\Delta_2}$

If $\Delta_1 = \Delta_2$ $\quad C = \dfrac{7}{48} C_0$

7.2 *Screen on Both Sides of Winding.*

In this section, a fixed amount of winding space is assumed, i.e. the space between two screens distance Δ apart and having a capacitance $C_1 = \dfrac{0.0885kSD}{\Delta}$ $\mu\mu$F.

Case A. Fig. 8(*a*).

Single-layer winding $C = C_1/3$.

Case B. Fig. 8(*b*)

Two layer winding.

Energy, coil to screen $= \dfrac{KV^2}{12\Delta_2}$

Energy between layers $= \dfrac{KV^2}{6\Delta_1}$

Total energy $= \dfrac{KV^2}{6}\left[\dfrac{1}{2\Delta_2} + \dfrac{1}{\Delta_1}\right]$

$$\therefore \quad C = \frac{K}{3}\left[\frac{1}{2\Delta_2} + \frac{1}{\Delta_1}\right]$$

$$\text{or} \quad C = \frac{1}{3}\left[\frac{1}{2\Delta_2} + \frac{1}{\Delta_1}\right]\Delta C_1$$

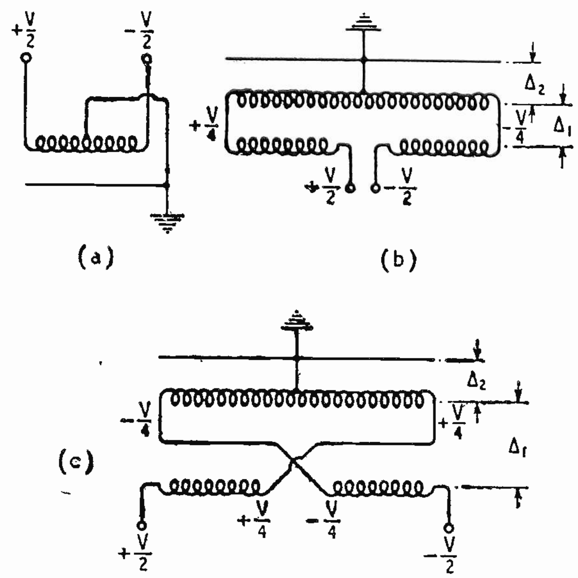


Fig. 7. *The capacitance of single-screened balanced windings: (a) plain winding; (b) sectionalized winding; (c) cross-connected sectionalized winding.*

If Δ_1 and Δ_2 are varied with Δ constant, then a minimum capacitance occurs when $\Delta_1 = \Delta/2$ and $\Delta_2 = \Delta/4$

When $C_{min.} = 4C_1/3$

Case C. Fig. 8(*c*).

Two-layer winding

Energy between layers $= \dfrac{KV^2}{12\Delta_1}$

Energy between layers and screen (Δ_2)

$$= \frac{KV^2}{48\Delta_2}$$

Energy between layers and screen (Δ_3)

$$= \frac{7KV^2}{48\Delta_3}$$

$$\therefore \quad C = \frac{K}{6}\left[\frac{1}{\Delta_1} + \frac{1}{4\Delta_2} + \frac{7}{4\Delta_3}\right]$$

$$\text{or} \qquad C = \left[\frac{4}{\Delta_1} + \frac{1}{\Delta_2} + \frac{7}{\Delta_3}\right]\frac{C_1\Delta}{48}$$

$$\text{Thus} \quad C = \frac{C_1\Delta}{48}\left[\frac{4}{\Delta_1} + \frac{1}{\Delta_2} + \frac{7}{\Delta - \Delta_1 - \Delta_2}\right]$$

since $\Delta = \Delta_1 + \Delta_2 + \Delta_3$.

C will be a minimum when $\dfrac{\partial C}{\partial \Delta_1} = \dfrac{\partial C}{\partial \Delta_2} = 0$

provided that $\dfrac{\partial^2 C}{\partial \Delta_1^2} \cdot \dfrac{\partial^2 C}{\partial \Delta_2^2} > \dfrac{\partial^2 C}{\partial \Delta_1 \partial \Delta_2}$.

$$\text{Then} \quad \frac{\partial C}{\partial \Delta_1} = -\frac{4}{\Delta_1^2} + \frac{7}{(\Delta - \Delta_1 - \Delta_2)^2} = 0$$

$$\text{and} \quad \frac{\partial C}{\partial \Delta_2} = -\frac{1}{\Delta_2^2} + \frac{7}{(\Delta - \Delta_1 - \Delta_2)^2} = 0$$

$$\therefore \quad \Delta_2 = \frac{\Delta_1}{2} \text{ and } \frac{\Delta_3}{\Delta_1} = \frac{\sqrt{7}}{2}$$

$$\text{also} \quad \Delta_3 = \Delta - \frac{3\Delta_1}{2}$$

$$\therefore \quad \Delta = \Delta_1\,(\sqrt{7}/2 + 3/2)$$

$$\therefore \quad \Delta_1 = 0.354\Delta$$

$$\Delta_2 = 0.177\Delta$$

$$\Delta_3 = 0.467\Delta$$

$$\therefore \quad C_{\min} = 0.665C_1$$

Case D. Fig. 8(*d*).

Three-layer winding.

Energy stored between layers (1) and (2)

$$= \frac{KV^2}{9\Delta_3}$$

Energy stored between layers (2) and (3)

$$= \frac{4KV^2}{9\Delta_2}$$

Energy stored between layer (1) and screen

$$= \frac{KV^2}{108\Delta_4}$$

Energy stored between layer (3) and screens

$$= \frac{19KV^2}{108}$$

$$\text{Total energy} = \frac{KV^2}{18}\left[\frac{19}{6\Delta_1} + \frac{8}{\Delta_2} + \frac{2}{\Delta_3} + \frac{1}{6\Delta_4}\right]$$

$$\therefore \quad C = \frac{\Delta C_1}{18}\left[\frac{19}{6\Delta_1} + \frac{8}{\Delta_2} + \frac{2}{\Delta_3} + \frac{1}{6\Delta_4}\right]$$

$$\text{If} \quad \Delta_1 = \Delta_2 = \Delta_3 = \Delta_4 = \frac{\Delta}{4}$$

$$\text{then} \qquad C = \frac{80}{27}C_1$$

8. Conclusion

A number of winding arrangements have been examined and compared by a method which can be applied to any arrangement

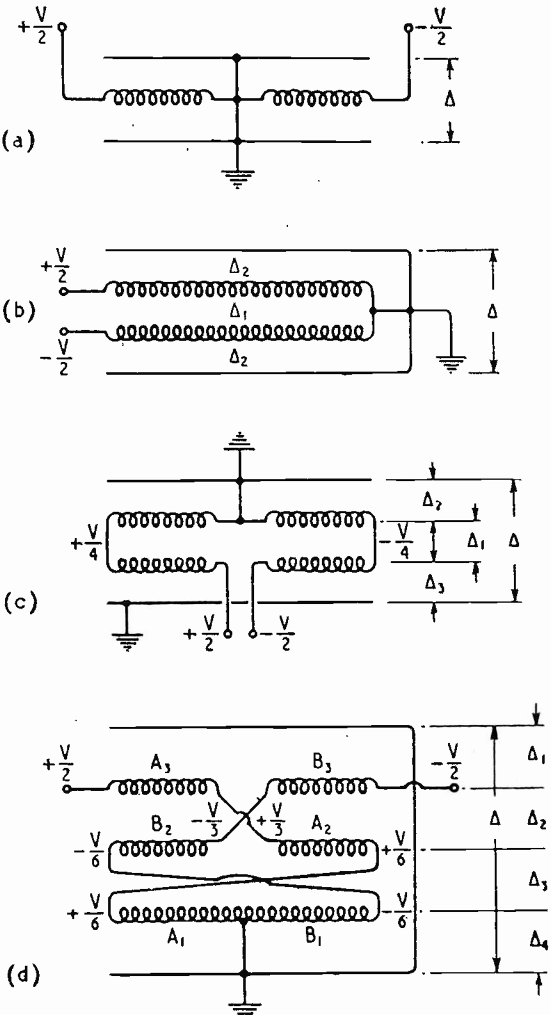


Fig. 8. *The capacitance of totally-screened balanced windings: (a) single-layer winding; (b) two-layer winding; (c) two-layer winding; (d) three-layer winding.*

where the turns of a winding can be regarded as perfectly coupled. This condition is usually applicable within the working band of a transformer but at very high frequencies a further analysis is necessary in order to predict the performance of the transformer.

PROPAGATION CHARACTERISTICS OF A UNIFORM LINE*

By I. F. Macdiarmid, A.M.I.E.E., and H. J. Orchard, Assoc. Brit. I.R.E.

(Engineer-in-Chief's Training School, Post Office Engineering Department)

SUMMARY.—A simple method is given of developing expressions which appear to be hitherto unpublished for the attenuation and phase-change coefficients of a uniform line in terms of the primary constants and the real parts of the characteristic impedance and admittance. The expressions are thought to give a clearer physical picture than is provided by the well-known expressions.

1. Introduction

EXPRESSIONS for the attenuation and phase-change coefficients of a uniform line in terms of the primary constants of the line are well known. While these expressions, or approximations derived from their general form, provide a satisfactory means of calculating the attenuation and phase-change coefficients in particular cases, they do not provide the student who is commencing the subject with a very clear idea of the causes of attenuation and phase-change and how these are affected by changes in one or more of the primary constants.

The purpose of this note is to show one simple method of deriving an alternative form of expression for attenuation and phase-change coefficients which helps to give a clearer understanding of the factors affecting attenuation and phase-change.

2. List of Symbols

α = attenuation coefficient of the uniform line.

β = phase-change coefficient of the uniform line.

γ = $\alpha + j\beta$ = propagation coefficient of the uniform line.

Z_0 = characteristic impedance of the uniform line.

Y_0 = $1/Z_0$ = characteristic admittance of the uniform line.

ψ = angle of Z_0.

R_0 = real part of Z_0.

G_0 = real part of Y_0.

R = resistance per unit length of the line (ohms).

L = inductance per unit length of the line (henries).

C = capacitance per unit length of the line (farads).

G = leakance per unit length of the line (mhos).

Z = $R + j\omega L = |Z| \underline{|\theta} $ = total series impedance per unit length.

Y = $G + j\omega C = |Y| \underline{|\phi}$ = total shunt admittance per unit length.

* MS. accepted by the Editor, November 1945.

3. The Propagation Coefficient

Considering a very short length of a uniform line, δx units long, with a single-frequency alternating current being propagated under reflectionless conditions, as in Fig. 1, and using the notation of this figure, we have :—

$$V_1/I_1 = V_2/I_2 = V/I = Z_0 \quad .. \quad (1)$$

and $$I_1/V_1 = I_2/V_2 = I/V = Y_0 \quad .. \quad (2)$$

The decrease in voltage over the length $\delta x \approx IZ\delta x$.

$$\therefore \quad \delta V \approx - IZ\delta x$$

whence $$\partial V/\partial x = - IZ \quad .. \quad .. \quad (3)$$

and similarly

$$\partial I/\partial x = - VY \quad .. \quad .. \quad (4)$$

But $$V = V_0\epsilon^{-\gamma x} \quad .. \quad .. \quad .. \quad (5)$$

where V_0 is the voltage at the sending end of the line and x is the distance from the sending end to the element.

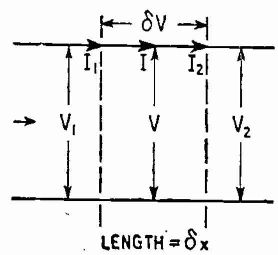


Fig. 1. *Section of a uniform line.*

$\therefore$ by differentiating (5)

$$\partial V/\partial x = - \gamma V$$

and similarly

$$\partial I/\partial x = - \gamma I$$

substituting this in (3) and (4)

$$\gamma = \frac{Z}{V/I} \quad \text{and} \quad \gamma = \frac{Y}{I/V} \quad .. \quad (6)$$

But $V/I = Z_0$ and $I/V = Y_0$

$$\therefore \quad \gamma = Z/Z_0 = Y/Y_0 \quad .. \quad .. \quad (7)$$

Therefore since $Y_0 = 1/Z_0$

$$Z_0 = \sqrt{Z/Y} \quad .. \quad .. \quad .. \quad (8)$$

Therefore from (7) and (8)

$$\gamma = \sqrt{ZY} \quad .. \quad .. \quad .. \quad .. \quad (9)$$

Which is the well known relationship.

Expression (7) can obviously be obtained directly from equations (8) and (9) if the latter are assumed to be known, but the direct method of obtaining (7) given above has been included for the sake of completeness.

3.1. Discussion of Propagation Coefficient

The quantities appearing in equation (7) can all be represented by vectors in a complex plane. Vectors representing the directions of

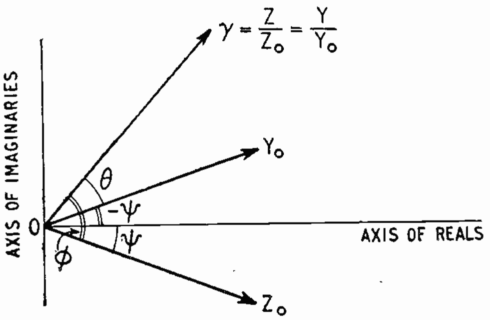


Fig. 2. *Vector representation of the quantities of Equation (7).*

γ, Z_0 and Y_0 are drawn in Fig. 2, where the lengths have no particular significance.

Since $Y_0 = 1/Z_0$

the angle of Y_0 = − (the angle of Z_0).

From (7)

(the angle θ of Z) = (the angle of γ) + (the angle of ψ of Z_0)

and

(the angle of ϕ of Y) = (the angle of γ) + (the angle $-\psi$ of Y_0)

The above angles are all indicated in Fig. 2.

4. The Attenuation and Phase-Change Coefficients

The information contained in Fig. 2 can be used to find the values of the real and imaginary parts of γ, i.e. α and β. To do this the vector representing γ must be drawn to scale so that its length represents the magnitude of γ. This is done in Fig. 3 where the figure is completed by drawing perpendiculars to the Z_0 and Y_0 vectors and to the axis of reals, as shown in the figure.

Using the notation shown in Fig. 3,

$$\angle EOB = \angle BOH = \psi$$

Whence and from the geometry of the figure

$$\angle GAB = \angle BAD = \psi$$

$$\therefore \quad HB = BE \qquad \ldots \qquad (10)$$

$$\text{and} \quad GB = BD \qquad \ldots \qquad (11)$$

Since OA represents γ and OB is the real part of OA

$$\alpha = OB = \tfrac{1}{2}(OG + OD) \qquad \ldots \qquad (12)$$

$$\text{Similarly } \beta = AB = \tfrac{1}{2}(AH + AE) \qquad \ldots \qquad (13)$$

$$\text{Since} \quad OG = OF \cdot \frac{1}{\cos\psi}$$

$$\text{and} \quad OF = AO \cdot \cos\phi$$

$$= \frac{|Y|}{|Y_0|} \cdot \cos\phi \text{ because } |\gamma| = \frac{|Y|}{|Y_0|}$$

$$\therefore \quad OG = \frac{|Y| \cos\phi}{|Y_0| \cos\psi}$$

$$= G/G_0 \quad \text{as } \cos\psi = \cos(-\psi) \qquad \ldots \qquad (14)$$

$$\text{Since} \quad OD = OC \cdot \frac{1}{\cos\psi}$$

$$\text{and} \quad OC = OA \cdot \cos\theta$$

$$= \frac{|Z|}{|Z_0|} \cdot \cos\theta \text{ because } |\gamma| = \frac{|Z|}{|Z_0|}$$

$$\therefore \quad OD = \frac{|Z| \cos\theta}{|Z_0| \cos\psi}$$

$$= R/R_0 \qquad \ldots \qquad (15)$$

Therefore from (12), (14) and (15)

$$\alpha = \tfrac{1}{2}\left(\frac{R}{R_0} + \frac{G}{G_0}\right) \qquad \ldots \qquad (16)$$

In the same way it can be shown from (13) that

$$\beta = \tfrac{1}{2}\left(\frac{\omega L}{R_0} + \frac{\omega C}{G_0}\right) \qquad \ldots \qquad (17)$$

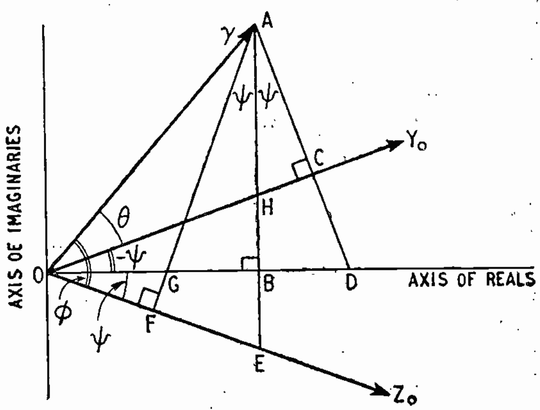


Fig. 3. *Vectorial derivation of $\alpha = OB$ and $\beta = AB$.*

Approximate expressions, having some relationship with (16) have been given by W. L. Everitt[1] for two special cases, viz.:

$$\text{For unloaded cables } \alpha \approx 0.707 \frac{R}{|Z_0|}$$

For loaded cables $\alpha \approx \frac{1}{2}\left(\frac{R}{Z_0} + \frac{GZ_0}{1}\right)$

These expressions are derived independently, from the two usual approximations for unloaded cables and loaded cables at voice frequencies (Equations (19) and (21) in this note). Being derived from approximations they only hold numerically where the approximations apply. The expression given in (16) being derived without any approximation is, however, exact under all circumstances.

4.1. Discussion of Attenuation Coefficient

Examination of (16) and (17) brings out a number of interesting points about the behaviour of lines. For example the attenuation of a line of given characteristic impedance depends solely on the resistive parts of Z and Y, whereas the phase change is determined by the reactive parts of Z and Y. Frequency will affect the attenuation of a line only in so far as it affects the real parts of the characteristic impedance and characteristic admittance, unless, as happens at high frequencies, it also alters the values of R and G.

The attenuation coefficient of a line depends upon the fraction of the transmitted power which is dissipated in the line and appears as heat in the conductors and the dielectric; that is upon the I^2R and V^2G losses. Now the expression for the attenuation coefficient given by (16) is composed of two terms, one directly proportional to R and the other to G. The term $\frac{1}{2} \cdot (R/R_0)$ therefore represents that part of the attenuation coefficient which is due to the I^2R losses and the term $\frac{1}{2} \cdot (G/G_0)$ likewise for the V^2G losses. This fact is shown more clearly in the alternative derivation of (16) given in the Appendix.

In any particular transmission line, where R and G may be considered fixed by economic reasons, it will not necessarily happen that the primary constants R, C, L and G are so related to one another that the sum of the I^2R and V^2G losses is a minimum. An underground telephone cable pair used at voice frequencies provides one such example. Here L and C are so related to R and G that the I^2R losses are much greater than the V^2G losses (i.e. $\frac{1}{2} \cdot (R/R_0)$ is much greater than $\frac{1}{2} \cdot (G/G_0)$), and also that Z_0 has a negative angle of approximately 45° (i.e. $\psi = -45°$), as will be shown in Section 5.1.

If now some means can be found of increasing R_0 then the I^2R losses (i.e. $\frac{1}{2} \cdot (R/R_0)$) will be reduced, resulting in a reduction in the attenuation coefficient. Such an increase in R_0 may not be possible without a proportionate decrease in G_0, but the net effect would still be a reduction in the attenuation coefficient up to the point where the I^2R and V^2G losses became equal.

R_0 can be increased either by decreasing ψ or by increasing $|Z_0|$. Both of these can be effected by increasing the ratio L/C. This can be done most easily by increasing L with inductance added continuously or in discrete quantities at intervals, i.e. by loading.

Since $R_0 = |Z_0| \cdot \cos\psi$

and $G_0 = \frac{1}{|Z_0|} \cdot \cos(-\psi)$

we have $R_0 \cdot G_0 = \cos^2\psi$

and so the increase in R_0 and the decrease in ψ caused by adding inductance must make the change in G_0 less than inversely proportional to the change in R_0.

Hence adding inductance will make the attenuation coefficient smaller within the limitations given above.

5. Approximations for the Attenuation and Phase-Change Coefficients

The usual well known approximations for the attenuation and phase-change coefficient can be derived readily from (16) and (17).

5.1. *Unloaded Cable at Voice Frequencies*

At the frequencies under consideration, usually

$$R \gg \omega L \quad \text{and} \quad \omega C \gg G$$

Then $Z_0 = \sqrt{\frac{R}{j\omega C}} = \sqrt{\frac{R}{\omega C}}\,\lfloor\overline{45°}$

whence

$$R_0 = \sqrt{\frac{R}{2\omega C}} \quad \text{and} \quad G_0 = \sqrt{\frac{\omega C}{2R}} \qquad (18)$$

$\therefore$ From (16)

$$\alpha = \frac{1}{2}\left\{\sqrt{2\omega CR} + G \cdot \sqrt{\frac{2R}{\omega C}}\right\}$$

As G is very small the second term is normally negligible.

$$\therefore \quad \alpha \approx \sqrt{\frac{\omega CR}{2}} \quad \ldots \quad \ldots \quad \ldots \quad (19)$$

and from (17) and (18)

$$\beta = \frac{1}{2}\left\{\frac{\omega L}{1} \cdot \sqrt{\frac{2\omega C}{R}} + \sqrt{2\omega CR}\right\}$$

and since ωL is very small

$$\beta \approx \sqrt{\frac{\omega CR}{2}} \quad \ldots \quad (20)$$

5.2. *Loaded Uniform Cable and Unloaded Cable at High Frequencies*

At frequencies such that $\omega L \gg R$ and $\omega C \gg G$ it can be shown[2] that

$$Z_0 \approx \sqrt{\frac{L}{C}} \,\Big|\, \tfrac{1}{2}(G/\omega C - R/\omega L)$$

The angle is normally only a few degrees

$\therefore$ $R_0 \approx \sqrt{L/C}$ and $G_0 \approx \sqrt{C/L}$

$\therefore$ From (16)

$$\alpha \approx \frac{R}{2}\sqrt{C/L} + \frac{G}{2}\sqrt{L/C} \quad \ldots \quad (21)$$

and from (17)

$$\beta \approx \tfrac{1}{2}\{\omega\sqrt{L\,.\,C} + \omega\sqrt{L\,.\,C}\}$$
$$\approx \omega\sqrt{L\,.\,C} \quad \ldots \quad (22)$$

Conclusions

Expressions which are believed to have been hitherto unpublished have been derived for the attenuation and phase-change coefficients of a uniform line. They are :—

$$\alpha = \tfrac{1}{2}(R/R_0 + G/G_0)$$
$$\beta = \tfrac{1}{2}(\omega L/R_0 + \omega C/G_0)$$

The method of deriving the expressions which has been given is not the only one possible but has been chosen because it yields the expressions for both α and β, and can readily be followed by a student with limited mathematical equipment. The latter is generally advantageous as it permits the student to concentrate on the physics of the problem rather than the mathematics. An alternative method of obtaining the expression for α which is equally simple, if not simpler than that already used, is given in the Appendix. This method, however, cannot be readily used to develop the expression for β.

The main advantages of the expressions developed above appear to be in the teaching of principles of transmission lines, since they do not seem to have any advantage over the conventional expressions for calculating numerical values of attenuation and phase-change coefficients.

APPENDIX

Alternative Derivation of Equation (16)

The power entering the element of line in Fig. 1 is $P_1 = V_1\,.\,I_1 \cos\psi$ where V_1 and I_1 are R.M.S. values of voltage and current respectively and since the length of the element is approaching zero, the power leaving the element is

$$P_2 \approx P_1 - I_1^2\,.\,R\,.\,\delta x - V_1^2\,.\,G\,.\,\delta x$$

$\therefore$ Power Ratio $P_2/P_1 = \epsilon^{-2\alpha\delta x}$

$$\approx \frac{V_1\,.\,I_1\cos\psi - (I_1^2R + V_1^2G)\delta x}{V_1\,.\,I_1\,.\cos\psi}$$
$$\approx 1 - \frac{(I_1^2R + V_1^2G)\delta x}{V_1\,.\,I_1\cos\psi} \quad \ldots \quad (23)$$

But $\epsilon^{-2\alpha\delta x} = 1 - \frac{2\alpha\delta x}{1!} + \frac{(2\alpha\delta x)^2}{2!} - \frac{(2\alpha\delta x)^3}{3!} + \ldots$

$\therefore$ From (23) when $\delta x \to 0$

$$2\alpha\delta x = \frac{(I_1^2R + V_1^2G)\delta x}{V_1\,.\,I_1\cos\psi}$$

$$\therefore \quad 2\alpha = \frac{R}{V_1/I_1\,.\cos\psi} + \frac{G}{I_1/V_1\,.\cos\psi}$$

$$\therefore \quad \alpha = \tfrac{1}{2}\{R/R_0 + G/G_0\} \quad \text{since } V_1/I_1 = |Z_0|$$

REFERENCES

[1] W. L. Everitt, "Communication Engineering," second edition, 1937. (McGraw-Hill Book Co.) p. 116 and p. 119.

[2] E.g., see Ref. 1, p. 118.

Book Review

Problèms de propagations guidées des ondes électromagnétiques.

By Louis de Broglie. Pp. 114 and 14 Figs. Gauthier-Villars, 55, Quai des Grands-Augustins, Paris. 160 francs.

This is a monograph by a well-known French scientist who is Professor of Physics at Paris. It was published in 1941. De Brogle's work on "La Mécanique Ondulatoire des Systèmes de Corpuscules" is well known and we were somewhat surprised to find that he had become interested in wave-guides. He says in the preface that he was attracted by the great theoretical and practical importance of the transmission of ultra short waves in tubes and horns and was led to study the published works on the subject, particularly those of Clavier and Brillouin. He claims no originality, but has merely written a résumé of these works in a form which he hopes will be helpful to physicists and radio-engineers. It is divided into six chapters, the first of which deals with the Maxwell equations and the Hertzian vector potentials. In the following chapters the formulae are developed for the various types of wave-guides and for the various modes of propagation, resonance conditions in cavities, loss of energy in the walls and the resulting attenuation. Chapter V deals with the propagation in horns, i.e. where the cross-section of the guide is gradually increasing, and Chapter VI with the diffraction of the waves at the mouth of the guide or horn. There is a bibliography of eighteen items, giving references to the papers by Barrow, Southworth, and others.

The work is, of course, rather mathematical, as it must be, but it is developed step by step and all set out very clearly. It is a carefully prepared review of the subject written by a theoretical physicist of world-wide repute.

G. W. O. H.

IRON-CORED LOOP RECEIVING AERIAL*

By R. E. Burgess, B.Sc.

(Communication from the National Physical Laboratory)

SUMMARY.—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.

1. Introduction

IT is well known that the insertion of a ferromagnetic core into a coil increases its self-inductance by virtue of the increase of flux. In this case the field producing the flux is due to the current in the coil, but if it were due to an external source an increase of flux would still be obtained. Thus, the use of an iron core to increase the pick-up of a loop is at once suggested. At radio frequencies the losses in a solid core would be prohibitive, and it is necessary to use a mass core consisting of finely-divided iron particles, insulated from each other and compressed into a suitable form.

Apart from increasing the signal pick-up of the loop, the core will produce an increase of inductance and effective resistance, and thus it is necessary to examine to what extent the latter effect offsets the increased pick-up. There appears to be no previously published work on this problem, and the object of the present paper is to discuss these considerations in detail.

The object of an iron-cored loop is to increase the sensitivity for a given size, or conversely, to reduce the size for a given sensitivity. Thus it has particular application where space is limited, e.g. on aircraft where the greater weight of the iron-cored loop will be offset, at least partially, by the reduced drag for the smaller size. Such an iron-cored loop has already been used in aircraft and W. J. Polydoroff has been granted a patent for this invention (British Patent 522,492).

* MS. accepted by the Editor, February 1946.

2. The Magnetic Properties of a Mass Core

If the mass core contains uniform spherical particles of iron of permeability μ_0 and conductivity σ (e.m.u.) insulated from each other, the effective permeability of the core is given[1, 2, 5] by

$$\mu = \frac{(\mu_0 + 2) + 2p\,(\mu_0 - 1)}{(\mu_0 + 2) - p\,(\mu_0 - 1)} = \frac{1 + 2pM_0}{1 - pM_0} \quad \ldots \quad \ldots \quad (1)$$

where p is the fractional volume of the iron and $M_0 = (\mu_0 - 1)/(\mu_0 + 2)$. This equation can also be written in the form

$$\frac{\mu - 1}{\mu + 2} = p\frac{\mu_0 - 1}{\mu_0 + 2} = pM_0 \quad \ldots \quad (2)$$

showing the additive nature of the term $(\mu_0 - 1)/(\mu_0 + 2)$ since the corresponding term for the insulator is zero.

The value of p for a cubical arrangement of spheres in contact is $\pi/6 = 0.52$ while for the most closely packed formation it is $\sqrt{2}\pi/6 = 0.74$. Thus in cores having a higher value of p than 0.74 the particles must be non-spherical and in general their irregularities will be randomly distributed in direction, so that the core is still magnetically isotropic. When the iron content is high ($p > 0.9$) the equation (1) based on simple statistical considerations is no longer accurate and permeabilities some 50 per cent. higher than those given by this equation have been observed.[5]

The relation between μ and p for $\mu_0 = 50$, 100 and infinity is shown in Fig. 1. It is seen that as the permeability of the iron particles is increased indefinitely, the effective permeability of the core tends to an upper limit given by

$$\mu_{\max} = \frac{1 + 2p}{1 - p} \quad \ldots \quad (3)$$

on account of a lower limit to the magnetic reluctance being set by the non-magnetic gaps between the particles.

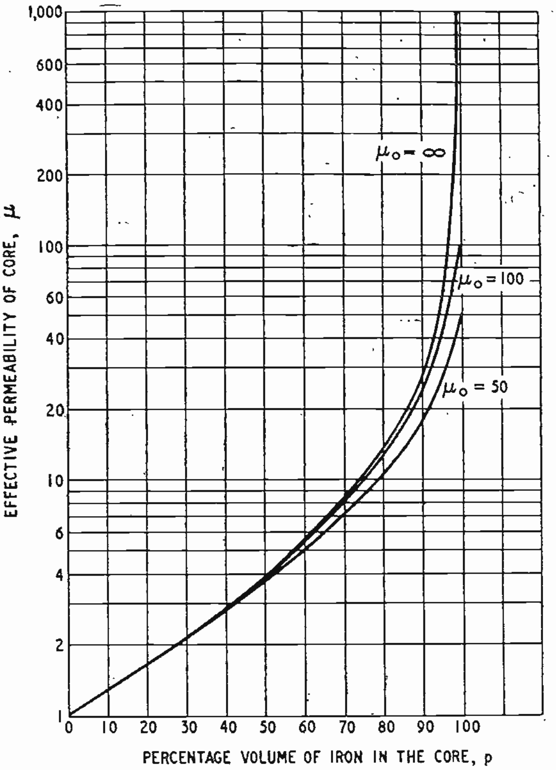


Fig. 1. *The variation of the effective permeability of the core with the volume proportions for various particle permeabilities.*

The equations (1), (2) and (3) are only strictly valid for magnetostatic fields. In alternating fields, eddy currents are set up in the particles which introduce a power loss depending upon the particle size and frequency, and the effective permeability of the core assumes a complex value $(\mu - j\mu')$. A parameter α can be defined[3] which determines the behaviour of the μ and μ' with frequency:

$$\alpha = r\sqrt{2\pi\omega\sigma\mu_0} \quad \ldots \quad (4)$$

where r is the particle radius, and it may be said that below the "bounding frequency" f_g at which $\alpha = 1$ the effect of the eddy currents on the real part of the permeability is small. Thus

$$f_g = \frac{1}{4\pi^2\mu_0\sigma r^2} \quad \ldots \quad (5)$$

For example if $\mu_0 = 100$, $\sigma = 10^{-4}$ e.m.u. (for iron) and $r = 5 \cdot 10^{-4}$ cm it is found that $f_g = 10$ Mc/s.

From Divilkovsky's analysis[3] it may be shown that in order to take account of the eddy currents in the particles the expression $M_0 = (\mu_0 - 1)/(\mu_0 + 2)$ in Equation (2) should be replaced by

$$M = \left[\frac{\mu_0 - 1}{\mu + 2} - \frac{3}{175}\frac{\mu_0(\mu_0 + 9)}{(\mu_0 + 2)^3}\alpha^4\right] - j\frac{3}{5}\frac{\mu_0}{(\mu_0 + 2)^2}\alpha^2 \quad \ldots \quad (6)$$

when $\alpha^2 \ll 1$, that is at frequencies well below the bounding frequency which is the condition in practical iron-cored systems.

On making this substitution it is found that the imaginary part of the permeability is given by

$$\mu' = \frac{18\pi}{5}\frac{p\mu_0^2\,\omega r^2\sigma}{[(\mu_0 + 2) - p(\mu_0 - 1)]^2} \quad \ldots \quad (7)$$

Eddy-current loss in the core is the most important, and in a well-designed iron-cored loop would represent the major part of the loss due to the core.

In the case where the core completely fills the loop without magnetic leakage, the inductance is increased μ-fold from L_0 to L while the eddy-current loss resistance R_e is related to the imaginary component μ' of the effective permeability :—

$$R_e + j\omega L = j\omega L_0(\mu - j\mu') \quad \ldots \quad (8)$$

or

$$R_e = \omega L_0\mu' = \omega L\,\mu'/\mu \quad \ldots \quad (9)$$

In the case where magnetic leakage exists, that is $L < \mu L_0$, which in practice always holds, it is *not* correct to write

$$R_e + j\omega L = j\omega L_0 k(\mu - j\mu')$$

which implies that $R_e/\omega L$ always equals μ'/μ.

In Appendix I the relation between the increase of resistance and inductance of a solenoid due to a spheroidal core is deduced. It is shown that whatever the relative size of the core and coil

$$\frac{R_e}{\omega(L - L_0)} = \frac{\mu'}{(\mu - 1)[(\mu - 1)g + 1]} \quad \ldots \quad (10)$$

where g is the demagnetization coefficient of the spheroid. In the case of an infinitely elongated core ($g = 0$) this equation agrees with (9) but in other cases the latter does not hold.

It appears from the equations developed

in Appendix I that there is no general equivalent circuit for an iron-cored coil that will represent its impedance irrespective of the configuration of core and coil. It may, however, be shown that if there is no magnetic leakage, i.e. the inductance is increased μ-fold by the core, then the "eddy current coefficient"[4] defined as R_e/f^2L is given by

$$W \equiv \frac{R_e}{f^2L} = \frac{8\pi^3}{5}\sigma r^2\mu \quad \ldots \quad \ldots \quad (11)$$

Thus the eddy current loss can be reduced either (a) by very fine subdivision of the particles or (b) by reducing the conductivity of the particles, which is possible by the use of certain nickel-iron alloys. In practice the first method is the most usual, the particles then being of pure iron.

Just as Equ. (1) for the permeability gives too low a figure for materials of large iron content ($p > 0.9$) so Equ. (11) tends to underestimate the eddy current losses owing to proximity losses and the increased tendency for electrical contact between the particles.

3. Increase of Pick-up due to the Core

Let the core have effective permeability μ and let the loop be wound directly on the surface of the core. The increase of pick-up due to the core is then equal to the ratio of the flux in the core to that when the core is absent. This increase will not in general be μ-fold since the induced magnetization gives rise to a secondary field which is in opposition to the original field.

If the original field H_0 is uniform and the induced magnetization is I, then the secondary field is written as $-4\pi gI$ where g is termed the demagnetization coefficient and is a function of the shape of the core and its orientation with respect to the field.

The effective magnetizing force in the core is thus

$$H = H_0 - 4\pi gI$$

although it is clear that these must be statistical values in the case of a mass core on account of the discrete magnetic and non-magnetic regions.

Now by definition of μ

$$I = KH = \frac{\mu - 1}{4\pi} H$$

where K is the susceptibility, and thus

$$H = \frac{H_0}{1 + g(\mu - 1)} \quad \ldots \quad \ldots \quad (12)$$

The flux density in the core is

$$B = \mu H = \frac{\mu}{1 + g(\mu - 1)} H_0$$

so the step-up in flux and hence of e.m.f. due to the core is

$$m \equiv \frac{B}{H_0} = \frac{\mu}{1 + g(\mu - 1)} \quad \ldots \quad \ldots \quad (13)$$

The only general shape for which the induced magnetization is uniform when H_0 is uniform is an ellipsoid; the expression for g of an ellipsoid is complicated unless it is an ellipsoid of revolution (spheroid), and therefore this shape will be considered now, for it is sufficiently general to cover the forms of practical interest. It is naturally assumed that the core and loop rotate together and that the usual cosine polar diagram is thus obtained.

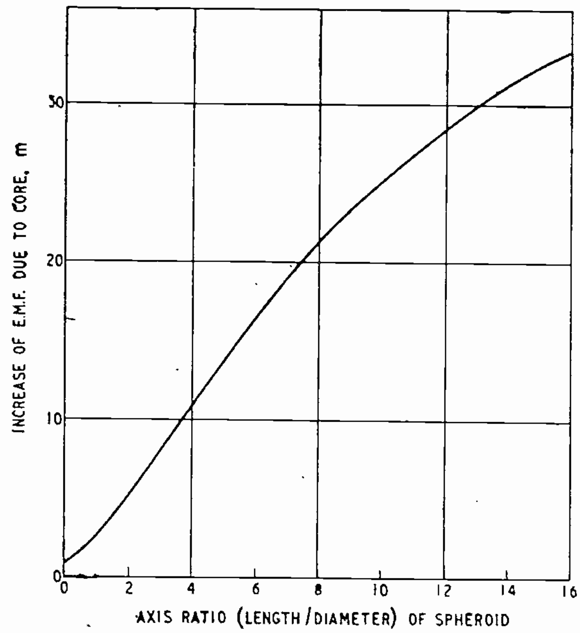


Fig. 2. *Increase of e.m.f. in a loop due to the insertion of a spheroid of permeability* 50.

It is seen from Equ. (13) that g should be as small as possible, which means that the core should be elongated in the direction of the field so that the "poles" are removed as far as possible from the centre. For this reason we will consider the case of a prolate spheroid with its major axis parallel to the field, for which

$$g = \frac{1 - e^2}{e^2}\left(\frac{1}{2e}\log_e\frac{1 + e}{1 - e} - 1\right) \quad \ldots \quad (14)$$

where e is the eccentricity which is related to the axis ratio a/b by

$$e = \sqrt{1 - \frac{b^2}{a^2}} \quad \ldots \quad (15)$$

In the case of the sphere for which $e = 0$ we find that $g = \frac{1}{3}$. To illustrate these equations Fig. 2 shows the variation of the step-up m with the axis ratio a/b when the core has a permeability of 50.

It is important to note that in general the increase of e.m.f. and the increase of inductance due to the core are not the same (as is assumed in the Patent mentioned earlier). These must both be smaller than μ, and only when the coil produces a uniform field (as in a long solenoid) will they be equal.

It is of interest to examine the effect of hollowing the core on the increase of e.m.f. which it gives. In Appendix II the case of a hollow spheroid is analysed, and it is found that the step-up m does not fall appreciably below the value for a solid spheroid until the shell of magnetic material is quite thin. For example Fig. 3 shows the variation of m with the ratio γ of the cross-sectional area of the hollow space to the total cross section for a permeability of 50 and an external axis ratio of 5 : 1. It is seen that γ may be as large as 0.8 before m is reduced to half the value for the solid spheroid.

The analysis shows that a hollow spheroid gives the same step-up m as the solid spheroid of the same outer dimensions but smaller permeability μ_e, that is

$$m = \frac{\mu_e}{1 + g_2(\mu_e - 1)} \quad \ldots \quad (16)$$

where g_2 is the demagnetization coefficient for the outer surface. The "equivalent permeability" μ_e is given to a close approximation by the mean of the product (area × permeability) for the cross section, viz.:

$$\mu_e \approx \gamma + (1 - \gamma)\mu \quad \ldots \quad (17)$$

The advantage of hollowing the core is clearly to save weight and material and the extent to which it is carried out is determined by the desirable compromise between these factors and the step-up m which governs the sensitivity.

If the core is of high permeability the field external to it and in its immediate vicinity will be considerably smaller than the original field H_0 and thus if the loop is spaced from the core, as will probably be necessary to avoid excessive dielectric loss, the step-up will be reduced to a value given approximately by

$$m' = m\frac{A_0}{A} \quad \ldots \quad (18)$$

where A_0 is the cross-sectional area of the core and A is that of the loop.

By way of illustration we may take a fairly typical case corresponding approximately to that of the loop used in German aircraft which was described by C. P. Edwards of the Royal Aircraft Establishment[6].

With core permeability $\mu = 50$

and ratio of radii $\frac{a}{b} = 4$

then the demagnetization coefficient $g = 0.075$ giving for the increase of pick-up $m = 10.6$.

If, however, the core is hollow with the hollow area being 40 per cent. of the whole area at the central cross section

$$\gamma = 0.4$$

giving $\mu_e = 34$

for which the increase of pick-up will now be $m = 10.0$, showing that the hollowing produces but little reduction of step-up.

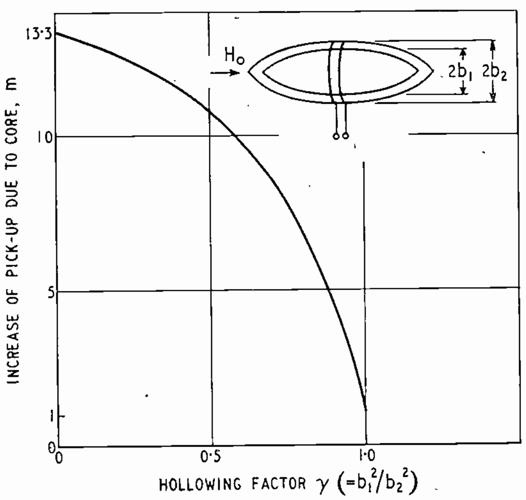


Fig. 3. *Increase of pick-up for a hollow spheroidal core of external axis ratio* 5 : 1 *and permeability* 50.

If the loop is spaced from the core so that the latter's outer cross-sectional area occupies only 60 per cent. of the loop area, the resultant step-up with the hollow core will be

$$m' = 0.6\, m = 6.0.$$

4. Conclusions

The main features of the iron-cored loop have been treated, and it may be concluded:—

(a) the core should have a high permeability ;

(b) the core should be elongated in the direction of the axis of the loop ;

(c) the particle size should be as small as possible to minimize W ;

(d) the core can be hollow to save weight and material and although this reduces the step-up m, a hollowing ratio of about 50 per cent. could usually be adopted with advantage ;

(e) the loop should be spaced from the core just sufficiently to prevent excessive dielectric loss, remembering that the step-up will be reduced ;

(f) the loop conductor should be designed to have minimum copper losses *with the core inserted* ; since the proximity loss will usually be relatively small owing to the reduced magnetic field at the conductor, the main consideration is to minimize the skin effect, implying that stranded wire will usually be advantageous.

5. Acknowledgements

The work described above was conducted as part of the programme of the Radio Research Board to whom this paper was circulated as a confidential paper in November 1941. The co-operation of the Royal Aircraft Establishment in providing loop systems from German aircraft is gratefully acknowledged. This paper is published by permission of the Department of Scientific and Industrial Research.

APPENDIX I

The relation between the eddy current loss resistance and increase of inductance of a solenoid due to a spheroidal core.

Divilkovsky[3] has discussed the case of a spherical mass core in a solenoid and the analysis will now be extended to the case of a spheroidal core. The solenoid is assumed to be sufficiently long for the field within it to be substantially uniform (H_0) in the absence of the core.

If the demagnetization coefficient of the spheroid is g the mean (or macroscopic) field within the core will be given by

$$H = \frac{H_0}{1 + g(\mu - j\mu' - 1)}$$

where $(\mu - j\mu')$ is the complex effective permeability of the core for the frequency concerned.

Now it may be shown that the field in which each particle is situated, i.e. the field observed when the particle is removed, is

$$H' = \frac{H}{1 - pM}$$

where M is the complex effective value of $(\mu_0 - 1)/(\mu_0 + 2)$ defined by equation (6). Bearing in mind that

$$\mu - j\mu' = \frac{1 + 2pM}{1 - pM}$$

it is found that

$$H' = \frac{H_0}{1 - pM(1 - 3g)}$$

Thus only for the simple case of a sphere in which $g = 1/3$ will $H' = H_0$.

Knowing the field H' at the particles, their magnetic moment can be calculated and the value of the magnetic vector potential at the solenoid winding found. From this, the electric intensity is deduced and thus the e.m.f. due to the core is found ; the component of e.m.f. in phase with the current gives the eddy current loss resistance while the quadrature component gives the increase of inductance.

In the present case of a spheroid it is finally found that

$$R_e + j\omega L = j\omega L_0\left[1 + \frac{3pM}{1 - pM(1 - 3g)}\,\frac{V_0}{V}\right]$$

where V_0 and V are the volumes of the core and coil respectively. If we use the relation

$$\frac{\mu - j\mu' - 1}{\mu - j\mu' + 2} = pM$$

we find that

$$R_e + j\omega L = j\omega L_0\left[1 + \frac{\mu - j\mu' - 1}{(\mu - j\mu' - 1)g + 1}\,\frac{V_0}{V}\right]$$

giving to a very close approximation as $\mu' \ll \mu$ in practice

$$L = L_0\left[1 + \frac{\mu - 1}{(\mu - 1)g + 1}\,\frac{V_0}{V}\right]$$

and

$$\frac{R_e}{\omega L_0} = \frac{\mu'}{[(\mu - 1)g + 1]^2}\cdot\frac{V_0}{V}$$

We may thus eliminate $\frac{V_0}{V}$ and obtain a relation which is independent of the relative volumes of the core and coil :

$$\frac{R_e}{\omega(L - L_0)} = \frac{\mu'}{(\mu - 1)[(\mu - 1)g + 1]} \qquad (10)$$

Alternatively we may eliminate g and obtain a relation which is not dependent on the ratio of length to diameter of the core.

$$\frac{R_e L_0}{\omega(L - L_0)^2} = \frac{\mu'}{(\mu - 1)^2}\,\frac{V}{V_0}$$

$$= \frac{2\pi}{5}\,\frac{\sigma\omega r^2}{p_0}\left(\frac{\mu_0}{\mu_0 - 1}\right)^2\frac{V}{V_0}$$

Thus if the ratio of the total volume pV_0 of the iron particles to the volume V of the coil is denoted by p_0 we have in general

$$\frac{R_e L_0}{\omega^2(L - L_0)^2} = \frac{2\pi}{5}\,\frac{\sigma r^2}{p_0}\left(\frac{\mu_0}{\mu_0 - 1}\right)^2 \approx \frac{2\pi}{5}\,\frac{\sigma r^2}{p_0} \text{ since } \mu_0 \gg 1$$

If the core is long or toroidal ($g = 0$) and fills the coil ($V_0 = V$) we have $L = \mu L_0$ and the " eddy current coefficient " is given by

$$W = \frac{R_e}{f^2L} = \frac{8\pi^3}{5}\sigma r^2 \frac{(\mu - 1)^2}{p\mu}\left(\frac{\mu_0}{\mu_0 - 1}\right)^2$$

$$\approx \frac{8\pi^3}{5}\sigma r^2 \mu \qquad \ldots \qquad \ldots \quad (11)$$

the latter approximation being very close for $\mu \gg 2$.

APPENDIX II

Increase of flux due to hollow core bounded by confocal prolate spheroidal surfaces.

The form of core which will be investigated is that bounded by confocal prolate spheroidal surfaces since this case gives a fairly simple analysis while it is sufficiently general to be applied to a wide variety of cores and in particular gives useful information regarding the effect of a hollow core.

The loop is assumed to be wound at the central circular cross-section of the core while the field is perpendicular to the plane of the loop, i.e. along the major axis of the core. The total flux threading the loop will be calculated and thus the step-up m due to the core is found; the results can also be conveniently expressed in terms of the equivalent permeability which a solid spheroid of the same outer dimensions must have to give the same increase m in flux.

Let the surfaces bounding the core have equations

$$x^2/a_1^2 + r^2/b_1^2 = 1 \text{ and } x^2/a_2^2 + r^2/b_2^2 = 1$$

where $r^2 = y^2 + z^2$ and $a_2^2 - b_2^2 = a_1^2 - b_1^2$ since the surfaces are confocal.

Let the core material have permeability μ and the original external field be H_0 parallel to the x-axis. The potential of this field is

$$V_0 = -H_0 x$$

Now it is known that the additional potential due to the core has the form

$$V_1 = (c_1 g_1 + c_2 g_2) H_0 x$$

in the hollow interior, while in the shell it has the form $V_2 = (c_1 g + c_2 g_2) H_0 x$

where $$g_1 = \frac{a_1 b_1^2}{2}\int_0^\infty \frac{d\lambda}{(a_1^2 + \lambda)^{3/2}(b_1^2 + \lambda)}$$

$$= \frac{1 - e_1^2}{e_1^2}\left(\frac{1}{2e_1}\log_e\frac{1 + e_1}{1 - e_1} - 1\right)$$

in which

$$e_1 = \frac{1}{a_1}\sqrt{a_1^2 - b_1^2}$$

is the eccentricity of the inner surface of the core

$$g_2 = \frac{a_2 b_2^2}{2}\int_0^\infty \frac{d\lambda}{(a_2^2 + \lambda)^{3/2}(b_2^2 + \lambda)}$$

$$= \frac{1 - e_2^2}{e_2^2}\left(\frac{1}{2e_2}\log_e\frac{1 + e_2}{1 - e_2} - 1\right)$$

in which $e_2 = \frac{1}{a_2}\sqrt{a_2^2 - b_2^2} = \frac{a_1}{a_2}e_1$ is the eccentricity of the outer surface, while

$$g = \frac{a_1 b_1^2}{2}\int_u^\infty \frac{d\lambda}{(a_1^2 + \lambda)^{3/2}(b_1^2 + \lambda)}$$

$$= e\left(\frac{1 - e_1^2}{e_1^3}\right)\left(\frac{1}{2e}\log_e\frac{1 + e}{1 - e} - 1\right)$$

where $e = \sqrt{\frac{a_1^2 - b_1^2}{a_1^2 + u}}$ is the eccentricity of the confocal surface corresponding to the parameter u which is the positive root of

$$\frac{x^2}{a_1^2 + u} + \frac{r^2}{b_1^2 + u} = 1.$$

The constants c_1 and c_2 are determined by the boundary conditions and it is found that

$$-\frac{\mu - 1}{c_1}$$

$$= \frac{(\mu - 1)[\mu - g_1(\mu - 1) - \beta(1 - g_2)(\mu - 1)]}{c_2}$$

$$= [\mu - g_1(\mu - 1)][1 + g_2(\mu - 1)] - \beta g_2(1 - g_2)(\mu - 1)^2$$

where $\beta = \frac{a_1 b_1^2}{a_2 b_2^2}$

is the ratio of the volume of the hollow to the total volume. It will be recognised that g_1 and g_2 are the demagnetization coefficients for spheroids of eccentricities e_1 and e_2 respectively, cf. equation (14).

The field in the hollow space is uniform and is given by

$$H_{1x} = -\frac{\partial}{\partial x}(V_0 + V_1) = H_0(1 - c_1 g_1 - c_2 g_2)$$

while the x-component of field inside the shell is given by

$$H_{2x} = -\frac{\partial}{\partial x}(V_0 + V_2)$$

$$= H_0\left[1 - c_1\frac{\partial}{\partial x}(gx) - c_2 g_2\right]$$

As the loop is wound around the central circular section of the core, its equation is $x = 0$ and $r = b_2$. Thus the relevant value of H_{2x} is that in the plane $x = 0$:—

$$H_{2x} = H_0(1 - c_1 g - c_2 g_2)$$

which is clearly non-uniform since g is a function of u which in turn is a function of r, namely at $x = 0$

$$r^2 = b_1^2 + u$$

Thus the total magnetic flux perpendicular to the plane of the loop is

$$\Phi = \pi b_1^2 H_{1x} + \mu\int_{b_1}^{b_2} H_{2x} 2\pi r \,.\, dr$$

Hence the step up in flux due to the core is

$$m = \Phi/\pi b_2^2 H_0 = \frac{b_1^2}{b_2^2}\left(1 - c_1 g_1 - c_2 g_2\right)$$
$$+ \mu\left(1 - \frac{b_1^2}{b_2^2}\right)\left(1 - c_2 g_2\right) - \frac{\mu c_1}{b_2^2}\int_{b_1}^{b_2} 2gr \,.\, dr$$

which after some reduction gives

$$m = \frac{\mu}{\mu - 1}c_2$$

$$= \mu\frac{\mu - g_1(\mu - 1) - \beta(1 - g_2)(\mu - 1)}{[\mu - g_1(\mu - 1)][1 + g_2(\mu - 1)] - \beta g_2(1 - g_2)(\mu - 1)^2}$$

In the limiting case of a solid spheroid ($\beta = 0$, $g_1 = 0$) we have as an upper limit to m the familiar expression

$$m = \frac{\mu}{1 + g_2(\mu - 1)}$$

while for an infinitely thin shell ($\beta = 1$, $g_1 = g_2$) the lower limit is naturally

$$m = 1$$

For any degree of hollowness between these two limits, it is seen that β lies between 0 and 1 while g_1 lies between 0 and g_2.

A convenient means of expressing the "hollowing effect" is the equivalent permeability μ_e which a solid spheroid of the same outer dimensions as the hollow core (and hence same g_2) must have in order to give the same m. This is evaluated by putting

$$m = \frac{\mu_e}{1 + g_2 (\mu_e - 1)} \quad \ldots \quad (16)$$

and gives

$$\mu_e = \mu \frac{\mu - g_1 (\mu - 1) - \beta (1 - g_2)(\mu - 1)}{\mu - \mu g\ (\mu_1 - 1) + \beta g_2 (\mu - 1)}$$

$$= \mu - \frac{\beta (\mu - 1)}{1 - (g_1 - \beta g_2)(1 - 1/\mu)}$$

If the ratio of the area of the hollow to that of the whole central cross section is

$$\gamma = \frac{b_1^2}{b_2^2}$$

the equivalent permeability is closely given by the mean of the product of area and permeability namely

$$\mu_e \approx \gamma + \mu (1 - \gamma) = \mu - (\mu - 1)\gamma \quad \ldots \quad (17)$$

for a core having an external axis ratio of greater than about 3 : 1.

To test the validity of this simple formula we consider the case of

$$\frac{a_2}{b_2} = 5 \qquad \mu = 50 \qquad \text{for which } \mu_1 = 13.3 \text{ in}$$

a solid spheroid.

Fig. 3 shows the variation of the step-up m with the area ratio γ as deduced from the exact equation, but the exact and approximate values of m agree to better than 1 per cent. everywhere.

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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.

Problem of Two Electrons

To the Editor, "Wireless Engineer."

SIR,—The apparent paradox discussed in your Editorial of January, 1946, is readily resolved by reference to the principle of special relativity according to which physical laws are invariant with respect to the Lorentz transformation.

In the case considered let the "stationary" frame of reference be S and let the frame S' move with the electrons in the x direction with velocity u. The co-ordinates of the electrons in S' are $(x', y', 0)$ and $(x', y' + r, 0)$ so that in S' the y-component of electrostatic repulsion is the only electromagnetic field component : i.e.,

$$Y' = \frac{e}{r^2}$$

If the Lorentz transformation is applied to Y' to give the fields observed in S there is the electric field

$$Y = KY' \quad \left[K = \left(1 - \frac{u^2}{c^2}\right)^{-\frac{1}{2}}\right]$$

and in addition a magnetic field in the z-direction

$$\gamma = K \frac{u}{c} Y'$$

which for u/c small corresponds to the usual formula for the magnetic field generated by crossing an electric field with uniform velocity u.

The resultant force between the electrons as observed in S is

$$eY - \frac{e\gamma u}{c} = KeY'\left(1 - \frac{u^2}{c^2}\right) = \frac{eY'}{K}$$

Since in passing from S' to S the electronic mass is transformed in the ratio K, the y component of acceleration in the ratio $1/K^2$, and the charge is invariant the equation of motion in the frame S is found to have the same form as in S'.

It is noted that the ratio of the attractive force to the repulsive force in S is u^2/c^2, and since this is always less than unity the resultant force appears to be a repulsion in any frame of reference.

This problem serves to illustrate that the resolution of an electromagnetic field into its electric and magnetic components is entirely relative to the observer and that the Lorentz transformation provides a set of relations which is useful in considering the electrodynamics of moving systems.

R. E. BURGESS.

National Physical Laboratory, Teddington.

Newton's Third Law

To the Editor, "Wireless Engineer"

SIR,—I have followed with considerable interest your Editorials relating to the forces on current elements and Newton's third law. As a consequence I have endeavoured to track down the history of the development of the Biot-Savart-Ampère formulas and have succeeded reasonably well with the exception of one link. In "A Treatise on Electricity and Magnetism" by Mascart and Joubert (translated by Atkinson), Thomas de la Rue & Co., London, 1883, to which you refer, I find on p. 441 :

"V. *The action of a magnet on an element of current is applied to the element.*—This results from the following experiment, due to M. Liouville. Part of the rectilinear current is made movable about its axis; with this object, its ends dip in two small mercury cups by which the current enters. The rectilinear element does not rotate at all, in whatever manner the magnet is presented to it."

If you or any of your readers can direct me to the account of that experiment I shall be much obliged. The experiment apparently was to test for the existence of a couple acting on a current element under the influence of a magnetic pole, the idea of which was so annoying to Ampère. However, whether this experiment actually entered into Ampère's argument or was made sometime after his work was done is a point on which I am not satisfied. I get the impression from the presentation of arguments by Mascart and Joubert that they outline the way Ampère might have reasoned with the hindsight available in 1883. I cannot directly trace all of the Mascart-Joubert logic in Ampère's papers. The reference they make to Liouville does not clearly say whether he performed the experiment or merely suggested it. Joseph Liouville (who presumably is meant) appears never to have done any experimental work, only pure mathematical analysis, according to his listed papers, and his one paper (1829) dealing with current elements does not suggest the experiment. The few others of the name Liouville who have works listed prior to 1883 do not deal with the subject, as far as I can ascertain.

R. H. Frazier.

Associate Professor, Massachusetts
Institute of Technology, U.S.A.

Preferred Numbers and Filter Design

To the Editor, "Wireless Engineer."

Messieurs,—Dans le numéro 265 de votre Revue (Octobre 1945), j'ai relevé un article de M. H. Jefferson, qui a retenu toute mon attention.

Cet article en effet fait ressortir les avantages importants que peuvent procurer l'application des séries géométriques décimales de nombres pour caractériser les éléments normalisés. Toutefois, l'auteur suggère des séries comportant respectivement, 6, 12 et 24 termes entre 2 puissances successives de 10, c'est-à-dire de raisons $\sqrt[6]{10}$, $\sqrt[12]{10}$ — $\sqrt[24]{10}$:

10 — 15 — 22 — 33 — 47 — 68 — 100
10 — 12 — 15 — 18 — 22 — 27 — 33 — 39 — 47 — 56 — 68 — 82 — 100
10-11-12-13-15-16-18-20-22-24-27-30-33-36-39-43-47-51-56-62-68-75-82-91-100

Ayant eu l'honneur de présider les réunions du Comité International de l'I.S.A. qui avait été chargé, en 1932, d'étudier la normalisation des nombres abstraits, je me permets de vous signaler les résultats des études qui avaient été entreprises avec le concours des Associations de Normalisation de 19 nations[1] :

Prenant comme base les remarquables travaux effectués en 1877 par le Français RENARD, alors capitaine du Génie, l'accord international de Stockholm (1934) consacre les séries géométriques décimales dites "Séries RENARD" comportant respectivement 5-10-20 & 40 termes entre deux puissances successives de 10, c'est-à-dire de raisons :

$$\sqrt[5]{10} \quad \sqrt[10]{10} \quad \sqrt[20]{10} \quad \sqrt[40]{10}$$

Le document ci-joint[2] donne les valeurs correspondantes, ainsi que les valeurs calculées des nombres des séries géométriques.

Les nombres normaux (Preferred Numbers) ainsi définis ont d'ailleurs été extraits :

- tantôt de la norme française de Renard utilisée depuis des dizaines d'années, en construction aéronautique principalement ;
- tantôt de la norme allemande établie en 1923 par la DIN*.

En fait, ils ont été choisis pour serrer de plus près les valeurs exactes (les écarts maximum sont − 1,01% et + 1,26%) ; ils possèdent évidemment les mêmes avantages que ceux que Monsieur Jefferson signale dans son article, mais ils permettent en outre des facilités de calcul qui sont exposées en détail en page 6 du document joint.

J'ajoute que dès 1936, l'A.S.A. (AMERICAN STANDARDS ASSOCIATION) a publié les mêmes séries géométriques décimales que celles de l'I.S.A.. complètées par des séries analogues pour le système de numération correspondant au pouce, à ses multiples et à sous-multiples.

J'estime donc qu'il y aurait un intérêt de premier order à ce que tous les travaux de normalisation faisant appel aux séries géométriques décimales s'appuient sur les séries RENARD internationales et je serais heureux de recevoir à ce sujet votre avis ainsi que celui de votre collaborateur.

P. Blanchet.

Compagnie Francaise Thomson-Houston,
Paris.

To the Editor, "The Wireless Engineer."

Sir,—I am grateful to M. Blanchet for drawing my attention to the work of Colonel Renard. The Renard series clearly obeys the basic definition given on page 484 of my paper and is particularly convenient in that the logarithms of the "Renard numbers" are all multiples of 0.025. The artifice of using logarithms to a base of 1.47 is therefore avoided.

The series used in my paper is, of course, that adopted during the war for the rationalization of British components. The object of my paper was to arrive at a solution in this series: the method is, as M. Blanchet points out, equally applicable to any scheme of preferred numbers.

H. Jefferson.

London, S.W.1.

[1] Les Associations nationales de normalisation allemande, autrichienne, belge, danoise, finlandaise, française, hollandaise, norvégienne, polonaise, suédoise, suisse, tchécoslovaque, U.R.S.S., avaient adhéré à l'accord international de 1934. L'Association japonaise s'était déclarée favourable par lettre. Les Associations espagnole, hongroise et italienne, sans faire obstacle au projet, avaient simplement réservé leur accord définitif.

[2] Fascicule XI-oo d'Août 1942, de l'Association Française de Normalisation (AFNOR).

* DIN (Deutsche Industrie Normen.)—[Ed.]

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.

TRANSMITTING CIRCUITS AND APPARATUS

573 957.—Oscillator-unit, comprising a grid-controlled valve coupled to a wave-guide, for generating centimetre waves.

The General Electric Co. Ltd. and E. G. James. Application date 11th February, 1943.

573 981.—Cathode-circuit modification of a short-wave oscillator comprising two push-pull valves feeding a concentric-line resonator.

The General Electric Co. Ltd. and M. R. Gavin. Application date 5th November, 1940.

574 051.—Reflectionless plug-and-socket connection for concentric lines carrying centimetre waves.

The General Electric Co. Ltd. and R. J. Clayton. Application date 7th January, 1942.

574 127.—Plug-and-socket joint, forming a half-wave section, for connecting a pair of concentric cables carrying centimetre waves.

The General Electric Co. Ltd. and R. J. Clayton. Application date 7th January, 1942.

SIGNALLING SYSTEMS OF DISTINCTIVE TYPE

573 900.—Signalling by a beam of light or infra-red rays through an arrangement of modulating shutters, which are made to overlap cyclically under the control of Morse or speech.

H. S. Molyneux-Ffennell, E. T. J. Tapp, F. G Charlton, and Vacuum-Science Products, Ltd. Application date 25th August, 1941.

574 133.—Delay-network for generating pulsed signals which are predetermined time-derivatives of an input or triggering pulse, and are of given shape and amplitude.

D. Blumlein (legal representative of A. D. Blumlein). Application dates 17th June and 21st July, 1942.

CONSTRUCTION OF ELECTRONIC-DISCHARGE DEVICES

573 901.—Electrode-system for focussing to a straight line the electron-stream of a cathode-ray or like discharge-tube.

O. E. H. Klemperer (commonly known as O. Klemperer). Application date 24th March, 1942.

573 999.—Making glass-to-glass seals, in the manufacture of thermionic valves, by radio-frequency heating.

Standard Telephones and Cables Ltd., P. K. Chatterjea, and S. J. Powers. Application date 22nd October, 1943.

574 009.—Amplifier or oscillator valve in which the input circuit forms part of a hollow resonator which is arranged symmetrically about the grid of the valve.

C. S. Bull. Application date 21st June, 1941.

574 056.—Electrode lens-system for focussing an electron-stream about a straight line.

O. E. H. Klemperer (commonly known as O. Klemperer). Application date 8th April, 1942.

574 109.—Multiple anode for generating a desired spectrum of X-rays, particularly for use in diffraction analysis.

The British Thomson-Houston Co., Ltd. Convention date (U.S.A.) 11th January, 1943.

574 237.—Process of "gettering" electron-multipliers on which metallic antimony or rubidium is used to remove excess alkali metal from the secondary-emission electrodes.

Cinema-Television Ltd., and A. Sommer. Application date 19th January, 1944.

574 248.—Metal-to-glass seal for an electron discharge tube.

Standard Telephones and Cables Ltd., and S. J. Powers. Application date 21st January, 1944.

574 383.—Construction of cathode core for an electron-discharge tube, designed to prevent damage to the sensitized coating in the process of assembly.

The M-O Valve Co., Ltd., G. F. Klepp, and J. A. Smyth. Application date 2nd April, 1943.

SUBSIDIARY APPARATUS AND MATERIALS

573 660.—Radio-frequency screening-device made of rubber mixed with a high carbon-content, to reduce its electrical resistance.

J. J. Davis. Application date 29th November, 1940.

573 679.—Continuously-adjustable choke-coil, incorporating a stationary magnetic circuit and a movable winding.

Akt. Brown, Boveri et Cie. Convention date (Switzerland) 23rd December, 1942.

573 717.—Transformer with two saturable paths, and a relay, for automatically generating a sub-harmonic of the triggering frequency.

Automatic Electric Laboratories Inc. Convention date (U.S.A.) 6th May, 1943.

573 793.—Amplifier, for use with a photo-electric cell or otherwise, comprising a saturated reactance coil and a crystal rectifier providing negative feed-back for voltage-control.

Standard Telephones and Cables Ltd. (communicated by International Standard Electric Corporation). Application date 6th November 1943.

573 936.—Electrical processing of a dry-contact rectifier to build-up its reverse resistance.

Standard Telephones and Cables Ltd., B. B. Grace, L. J. Ellison, H. S. Leman, C. W. Leng, and A. S. Bridge. Application date 15th February, 1943.

573 959.—Circuit arrangement for giving a continuous indication of the reverse resistance of a dry-contact rectifier during the final processing of the rectifier.

The British Thomson-Houston Co., Ltd., and J. Dyson. Application date 29th March, 1943.

574 345.—Logarithmic ready-reckoner for electrical calculations, comprising three coaxial rotary discs and a cursor.

K. M. Bancroft and Ionic Laboratories Ltd. Application date 25th January, 1944.

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.

	PAGE A.
Acoustics and Audio Frequencies	107
Aerials and Transmission Lines	109
Circuits	110
General Physics	113
Geophysical and Extraterrestrial Phenomena ...	115
Location and Aids to Navigation	115
Materials and Subsidiary Techniques	117
Mathematics	119
Measurements and Test Gear	120
Other Applications of Radio and Electronics ...	122
Propagation of Waves	124
Reception	125
Stations and Communication Systems	127
Subsidiary Apparatus	128
Television and Phototelegraphy	130
Transmission	130
Valves and Thermionics	131
Miscellaneous	132

ACOUSTICS AND AUDIO FREQUENCIES

534.21 1422

Propagation of Sound Waves in a Liquid Layer between Two Absorbing Half-Spaces.—L. M. Brekhovskikh. (*C. R. Acad. Sci. U.R.S.S.*, 1945, Vol. 48, No. 6, pp. 397–400. In English.) Theoretical analysis of a problem in sound wave propagation which "may find application in geophysics (propagation through an oil layer), in architectural acoustics and in hydroacoustics (propagation of sound through the sea)".

534.213 : 621.395.623.7 1423

Wave Propagation and Outdoor Field Tests of a Loudspeaker System.—F. L. Hopper & R. C. Moody. (*J. Soc. Mot. Pict. Engrs*, Feb. 1946, Vol. 46, No. 2, pp. 115–123.) Propagation tests at frequencies between 100 c/s and 8 kc/s are described, indicating an inverse square law attenuation up to a distance of 400 wavelengths for heights of transmitter and receiver of $7\frac{1}{2}$ ft and 3 ft respectively. The effects of snow, wind and pressure are also discussed.

534.321.9 : 620.179 1424

Supersonic Inspection of Strip Materials.—(*Electronics*, March 1946, Vol. 19, No. 3, pp. 166..170.) See also 1149 of May.

534.41 + 534.781 1425

Visible Speech.—R. K. Potter. (*Bell Lab. Rec.*, Jan. 1946, Vol. 24, No. 1, pp. 7–11.) An account in general terms of the system described in 823 of April.

534.43 1426

Spotting and Repeating Record Player.—R. H. Bailey. (*Radio News*, Feb. 1946, Vol. 35, No. 2, pp. 29..128.) Brief general account of a device to enable accurate location of any particular passage on a disk record, and to enable the accurate setting of the pickup to repeat any passage of which the position has been noted.

534.6 : 621.395.623.7 1427

On the Maximum Possible Reinforcing of Sound under Outdoor Conditions: and in an Enclosure.—G. M. Suharevsky. (*C. R. Acad. Sci. U.R.S.S.*, 1940, Vol. 26, No. 9, pp. 885–891, 892–899. In English.) The degree of "sound reinforcing" is defined as 10 log (I_a/I_o) where I_a is the average level of intensity at the point of audition and I_o is the average level of intensity at the microphone. This quantity is limited by acoustical feedback, and by associated frequency distortion. Very detailed formulae are given, depending on the polar diagram and location of the loudspeaker and the constants of the boundaries of the sound field. The acoustic feedback limitation is very severe in an enclosure. One means of raising the possible level of reinforcement is to "mix up" the standing wave systems by revolving the loudspeaker so that its axis describes a cone; or by means of a revolving screen in the horn or at the mouth of the loudspeaker. A gain of 6–8 db can be realized in this way.

534.62 : 061.6 1428

Acoustic Laboratory.—(*Electronic Industr.*, March 1946, Vol. 5, No. 3, p. 78.) Description of the Harvard "Anechoic Chamber": reflection coefficient at walls 0.1%. Development of sound insulating assemblies: investigation of speech intelligibility at high altitudes. See also 1429 below.

534.62 : 061.6 1429

Achievements of Harvard's Electro-Acoustic Laboratory.—(*Electronics*, March 1946, Vol. 19, No. 3, pp. 312..323.) A review of the work done in sound-proofing aeroplane cabins and in improving the system of communications between the members of the crew; includes a description of the construction of an echo-free room. See also 1428 above.

534.845 : 677.521 **1430**
Forms, Properties and Functions of Fibrous Glass Acoustical Materials.—W. M. Rees. (*Communications*, Jan. 1946, Vol. 26, No. 1, pp. 36, 38.) The materials are provided as wool, blankets, etc., or bonded with resin to form boards. A table gives absorption data. Graphs show that for layers of wool, etc., up to about 2 inches thick the absorption increases equally with density throughout the frequency spectrum. At greater thicknesses absorption is little affected by density at frequencies above 500 c/s. Encasing boards in paint, cellophane, etc., causes a small increase in absorption at low frequencies and a large decrease at high frequencies. See also 1156 of May (Rees & Taylor).

534.88 + 621.395.625.3 **1431**
I.R.E. Winter Technical Meeting January 1946.—(*Waves and Electrons*, Feb. 1946, Vol. 1, No. 2, pp. 80..93.) Abstracts are given of the following papers read at the meeting. Magnetic Recorder as an Adjunct to the Home Receiver.—S. J. Begun (Title only). Basic Principles of Underwater Sound-Equipment Design.—R. Bennett. For other abstracts, see *Electronics*, March 1946, Vol. 19, No. 3, pp. 92..108, and *Electronic Industr.*, March 1946, Vol. 5, No. 3, pp. 62..74. For titles of other papers read, see other sections.

534.88 : 533.6.013.22 **1432**
On the Propagation of Sound in Turbulent Atmosphere.—V. Krassilnikov. (*C. R. Acad. Sci. U.R.S.S.*, 1945, Vol. 47, No. 7, pp. 469–471. In English.) The phase fluctuations of a sound wave propagated in a turbulent atmosphere are calculated, and applied to the problem of locating a source of sound by spaced microphones. The inaccuracy of sound rangers in conditions of strong wind is attributed to the effect of turbulence on the mean velocity of sound along the path.

621.395.42 : 534.7 **1433**
Infrasonic Switching.—Montani. (*See* 1667).

621.395.614 **1434**
A Unidirectional Crystal Microphone.—A. M. Wiggins. (*Communications*, Jan. 1946, Vol. 26, No. 1, pp. 20–22.) A unidirectional microphone is obtained by placing a damped, dead diaphragm of correct mechanical impedance behind the crystal. The formulae for calculating the correct resistance and mass of this diaphragm are given. The response rises at the rate of 6 db per octave, and this is counteracted by an RC filter of complementary characteristics.

621.395.615 **1435**
Electronic Microphone.—J. Rothstein. (*Electronics*, March 1946, Vol. 19, No. 3, pp. 230..232.) A flexible diaphragm is mechanically coupled to the grid of a parallel-plane electron tube. Motion of the diaphragm changes the position of the grid relative to the anode and cathode and so changes the amplification factor of the valve. Summary of U.S. Patent 2 389 935.

621.395.623.7 : 621.392.53 **1436**
Reduction of Loudspeaker Distortion.—F. C. Jones. (*Radio, N.Y.*, Jan. 1946, Vol. 30, No. 1, pp. 26–28.) Harmonic distortion at low frequencies, which is serious at high inputs, arises from the low resonant frequency of the moving-coil system. It may be substantially reduced by using a "bass reflex" cabinet enclosure with an air vent adjacent to the speaker cone. High-frequency distortion, arising from frequency modulation of the high frequencies by a low frequency, may be reduced by using two speakers in a dual system, one for the high and the other for the low frequencies. The design of cross-over filter networks for use in such a system is discussed and practical constructional data are given. The circuits described are for 6-ohm speech coils and give cross-overs at about 1 500 c/s. See also 529 of March.

621.395.623.75 **1437**
Design of Compact Two-Horn Loudspeaker.—P. W. Klipsch. (*Electronics*, Feb. 1946, Vol. 19, No. 2, pp. 156–159.) "For room-corner locations, where walls can be utilized to produce reflections that multiply the mouth area of the woofer sufficiently for efficient propagation of sound waves down to 40 c/s. Companion tweeter gives wide-angle radiation." Based on 830 of April (Klipsch) and back references.

621.395.623.8 **1438**
Massive Speaker Cabinet.—C. A. Volf. (*Electronics*, March 1946, Vol. 19, No. 3, pp. 206..214.) Another account of the system described in 531 of March and 1162 of May.

621.395.623.8 **1439**
High-Power Military Sound Systems.—(*Electronics*, Feb. 1946, Vol. 19, No. 2, pp. 216..222.) An outline of the device described in 532 of March, and of a giant loudspeaker system installed in a bomber, for addressing an entire city from the air.

621.395.645 **1440**
Hi-Fi Amplifier Contest.—J. W. Straede. (*Radio Craft*, Jan. 1946, Vol. 17, No. 4, pp. 249..272.) Details of the four best amplifiers in an Australian contest, limited to amateurs, in which the fidelity was judged by a public audience, and the technical details by a special panel.

621.395.92 **1441**
The Testing of Deaf Aids.—T. H. Turney. (*J. sci. Instrum.*, March 1946, Vol. 23, No. 3, pp. 58–59.) Describes the measurement of the gain by a two-voltmeter method and by an artificial ear method. Substantial agreement is obtained.

621.395.92 : 534.771 **1442**
Historic Firsts : The Audiometer.—(*Bell Lab. Rec.*, Feb. 1946, Vol. 24, No. 2, pp. 57–58.) An instrument for investigating hearing ability at any frequency between 30 and 16 000 c/s.

621.396.611.21 + 621.317.361 **1443**
Duplex Crystals.—Lane. (*See* 1582.)

621.396.97 **1444**
Radio Sound Effects. [Book Review]—J. Cramer & W. B. Hoffman. Ziff-Davis, New York, 1945, 71 pp., $1.50. (*Electronics*, Feb. 1946, Vol. 19, No. 2, p. 360.) "Not a text, but a syllabus of study for broadcasters, sound technicians . . . [it] makes no pretence of teaching its readers in one easy lesson."

AERIALS AND TRANSMISSION LINES

621.317.79 : [621.315.212.1.029.62/.63 1445
Measuring Coaxials at Ultra-High Frequencies.—Fleming. (*See* 1597.)

621.392 1446
The Capacity of Twin Cable—II.—J. W. Craggs & C. J. Tranter. (*Quart. appl. Math.*, Jan. 1946, Vol. 3, No. 4, pp. 380–383.) An extension of 660 of March (Craggs & Tranter) to the case in which the dielectric sheaths enclosing the two conductors are separated and surrounded by another dielectric medium. Two methods are given, leading to series solutions, one converging more rapidly than the other, according to the ratio of the dielectric constants of the two dielectrics involved.

621.392 1447
Waves and Wave Guides: Part III.—G.G. (*QST*, March 1946, Vol. 30, No. 3, pp. 61..134.) Nomenclature of modes of guided waves, and determination of cut-off frequencies. A practical consideration of guide dimensions for the amateur bands. For previous parts see 546 of March and 284 of February.

621.396.67 1448
I.R.E. Winter Technical Meeting January 1946.—(*Waves and Electrons*, Feb. 1946, Vol. 1, No. 2, pp. 80..93.) Abstracts are given of the following papers read at the meeting. Antenna for Frequency-Modulation Station WGHF.—A. Alford. Broad-Band Antennas and Direction-Finding Systems for Very High Frequencies.—A. Alford, J. D. Kraus, A. Dorne & J. Christensen. Metal Lens Antennas.—W. E. Kock. Design Considerations in Broadside Arrays. — J. Ruze. Ultra-High-Frequency Television Transmitters and Antennas.—R. Serrell. Model Aircraft-Antenna Measurements.—G. Sinclair, E. W. Vaughan & E. C. Jordan. Beam-Shaping Methods in Antenna Design.—L. C. Van Atta. Directional Couplers.—W. W. Mumford. From Wiring to Plumbing.—E. M. Purcell. Equivalent Circuits for Wave-Guide Structures.—J. Schwinger. For other abstracts see *Electronics*, March 1946, Vol. 19, No. 3, pp. 92..108, and *Electronic Industr.*, March 1946, Vol. 5, No. 3, pp. 62..74. For titles of other papers, see other sections.

621.396.67 1449
The Effective Length of a Half-Wave Dipole.—G.W.O.H. (*Wireless Engr*, April 1946, Vol. 23, No. 271, pp. 95–96.) The equivalence of a dipole with sinusoidal current distribution, and an aerial λ/π long carrying a uniform current equal to that at the centre of the dipole, is not exact for points off the equatorial plane, and the field deduced therefrom is shown to be up to 5 or 6% in error. See also 1802 of 1945 (G.W.O.H.).

621.396.67 1450
Some Experiments with Linear Aerials.—J. S. McPetrie & J. A. Saxton. (*Wireless Engr*, April 1946, Vol. 23, No. 271, pp. 107–114.) A method of determining the polar diagram of a linear aerial based on the summation of the effect of current waves induced in the aerial by an incident electromagnetic field gives satisfactory agreement between theoretical and experimental polar diagrams for aerials an integral number of halfwaves in length. An experimental investigation of front-to-back ratios of a halfwave dipole and parasite shows that, as a reflector system, the optimum length and spacing of the parasite were 0.5 and 0.1 of the wavelength respectively, giving 11 db ratio, and as a director system, the corresponding values were 0.47 and 0.05, giving 20 db ratio.

621.396.67 1451
Antenna Construction.—A. Alford & M. Fuchs. (*Radio, N.Y.*, Jan. 1946, Vol. 30, No. 1, pp. 43, 58.) A dipole consists of two metal sleeves end to end, insulated from and surrounding a coaxial tubular metal mast which projects beyond the upper sleeve. The aerial is substantially balanced, and is fed from a transmission line carried through the interior of the mast. Summary of U.S. Patent 2 385 783.

621.396.67 : 621.396.82 1452
QRM—The Electronic Life Saver: Part II [Aerials for radar countermeasures].—Robbiano. (*See* 1527.)

621.396.671 1453
The Cylindrical Antenna: Current and Impedance.—R. King & D. Middleton. (*Quart. appl. Math.*, Jan. 1946, Vol. 3, No. 4, pp. 302–335.) A theoretical analysis of an idealized case, with premises similar to those postulated in earlier investigations (see Hallén, 2763 of 1939, and King & Harrison, 817 of 1944). The aim of the paper is solely analytical improvement of the solution of the problem. This has been achieved by introducing parameters different from those used by Hallén (*loc. cit.*) in order to fit the actual current distribution to an analytical form. A comparison of first-order solutions for the current distribution in Hallén's and the present work shows that "The new, more exact theory leads to a distribution with somewhat greater relative amplitudes nearer the outer parts of the antenna, and with a somewhat larger component in phase with the driving potential difference." The impedance, on the new theory, differs in detail but not in any major manner from that given by the previous work. In general, resistances are smaller at antiresonance, and are greater at resonance: these differences are most significant for large values of antenna radius. A comparison is made between the new theory and unpublished experimental determinations, by D. D. King, of the impedance characteristics of such an antenna. The agreement is good for all quantities in the second-order theory, but only approximate in the first order.

621.396.671 1454
The Efficiency of a Short Transmitting Antenna.—V. J. Andrew. (*Communications*, Jan. 1946, Vol. 26, No. 1, pp. 52–53.) The inefficiency of aerials of lengths less than 0.18 λ is due to the fact that they have a low radiation resistance and a large negative reactance, resulting in relatively high losses in the loading inductor. The adverse effect of the lead-in capacitance can be reduced by placing the loading inductor at the base of the aerial.

621.396.673 1455
The Quadrant Aerial.—N. Wells. (*Marconi Rev.*, Jan./March 1946, Vol. 9, No. 80, pp. 21–23.) This is a horizontal aerial system with an omnidirectional pattern. It can be used over a wide frequency band for both transmission and reception. The system comprises two aerials with figure-of-eight diagrams set at right angles to each other and in phase quadrature. The omnidirectional character

is well maintained over a frequency range of 2 to 1, as is the vertical directivity pattern, which is governed by the height of the aerial. A group of four poles carrying four aerials can together cover the band 2–30 Mc/s.

621.396.677 **1456**

A Generalised Radiation Formula for Horizontal Rhombic Aerials.—H. Cafferata. (*Marconi Rev.*, Jan./March 1946, Vol. 9, No. 80, pp. 24–35.) Limitations of the earlier calculations of the radiated field are pointed out and a more general treatment is given. In particular, the positions where mixed polarized waves are received is considered, and account is taken of current attenuation in the conductors. "The source considered is that of a multiple array of horizontal rhombic elements arranged *n* in cascade and *m* cascades in parallel, and all contained in the same horizontal plane. The formula allows for arbitrary phase relations as between cascades and between elements in each cascade. This permits of "steering" the main lobe of radiation and control of the interference pattern in the XZ plane." To be continued.

621.396.677.029.6 **1457**

High-Gain Microwave Antennas.—W. G. Tuller. (*QST*, March 1946, Vol. 30, No. 3, pp. 34–40, 122.) A survey of antenna arrays developed for radar during the war, and of their possible uses in amateur communication. Types discussed include a dipole array with a plane reflector, a parabolic dish fed by a dipole, and a cut dish fed by a waveguide and radiating horn. Types of feeder include the solid-dielectric coaxial line, the rectangular waveguide, and the stub-supported coaxial line. Design and performance data for representative arrays for 200 Mc/s and 3 000 Mc/s are given. Brief mention is made of many possible variations.

CIRCUITS

621.317.733 **1458**

Note on the Helmholtz Make-and-Break Theorem and an Application to the Wheatstone Net.—Freeman. (*See* 1593.)

621.318.572 **1459**

Pulse-Integrating Circuit.—W. N. Tuttle. (*Radio, N.Y.*, Jan. 1946, Vol. 30, No. 1, p. 42.) The pulses to be counted are passed through a valve with zero plate voltage to a capacitor with a suitable leak resistance. The d.c. potential difference created across the capacitor is proportional to the rate of arrival of the pulses, and is measured with a triode voltmeter. Summary of U.S. Patent 2 374 248.

621.318.572 **1460**

Geiger-Muller Counter Technique for High Counting Rates.—C. O. Muehlhause & H. Friedman. (*Phys. Rev.*, 1st/15th Jan. 1946, Vol. 69, Nos. 1/2, p. 46.) A G-M counter coupled to a 4-Mc/s-wide video amplifier followed by a scaling circuit and a triggered-sweep oscilloscope (see 1859 of 1942—Stever) enabled counting rates up to 10^5 sec^{-1} to be obtained. Summary of Amer. Phys. Soc. paper.

621.318.572 : 621.317.361 **1461**

Decade Counting Circuits.—Regener. (*See* 1581.)

621.392 + 621.396 [.622 + .64 **1462**

I.R.E. Winter Technical Meeting January 1946.—(*Waves and Electrons*, Feb. 1946, Vol. 1, No. 2, pp. 80..93.) Abstracts are given of the following papers read at the meeting. Capacitance-Coupled Wide-Band Intermediate-Frequency Amplifiers.—M. J. Larsen & L. L. Merrill. Directional Couplers.—W. W. Mumford. From Wiring to Plumbing.—E. M. Purcell. Equivalent Circuits for Wave-Guide Structures.—J. Schwinger. Discriminators for Frequency-Modulation Receivers.—S. W. Seeley. Stagger-Tuned Wide-Band Amplifiers.—H. Wallman. For other abstracts see *Electronics*, March 1946, Vol. 19, No. 3, pp. 92..108, and *Electronic Industr.*, March 1946, Vol. 5, No. 3, pp. 62..74. For titles of other papers read, see other sections.

621.392.5 : 621.396.619.018.41 **1463**

The Transmission of a Frequency-Modulated Wave Through a Network.—W. J. Frantz. (*Proc. Inst. Radio Engrs, N.Y.*, Part I, March 1946, Vol. 34, No. 3, pp. 114–125.) The paper was prompted by the increasing use of frequency-modulated signal generators in conjunction with a cathode-ray-tube display for investigating the bandwidth, etc., of amplifiers. The possible errors arising in such testing procedure are investigated in detail, but the results are of general application. "A practical [and simple] method for calculating the effect of a four-terminal network upon a frequency-modulated wave being transmitted through it is developed and demonstrated... [It is] equally accurate and practical for large and small values of modulation index and for any physical network." The amplitude and phase of the output voltage is expressed as a function of time or of instantaneous input frequency. A simple expression is derived giving the necessary correction to the output-voltage/frequency characteristic determined under steady-state conditions. Practical worked examples are included.

621.392.52 **1464**

Analysis of a Resistance-Capacitance Parallel-T Network and Applications.—A. E. Hastings. (*Proc. Inst. Radio Engrs, N.Y.*, Part I, March 1946, Vol. 34, No. 3, pp. 126–129.) "... to find the conditions for a null in output and the transfer characteristic. Its use in the return circuit of a feedback amplifier is considered, and the bandwidth of the resulting tuned amplifier is found. The requirements for stability of the amplifier and for its use as an oscillator are discussed." The discussion deals with three arrangements, and is mainly concerned with networks in which the resistive and capacitive elements respectively, comprising the horizontal arms of the two T's, are equal. For such a network interposed in the feedback link of an amplifier, the maximum available "effective *Q*" of the amplifier is a quarter of the voltage gain of the unit before feedback is applied.

621.392.52 **1465**

Bridged-T Null Networks.—Z. Bryl. (*Proc. Instn Radio Engrs, Aust.*, Nov. 1945, Vol. 6, No. 5, pp. 8–9.) Mathematical derivation of conditions necessary for zero transmission.

621.392.52 **1466**

Bridged-T Circuit.—R. B. Essex. (*Radio Craft*, Feb. 1946, Vol. 17, No. 5, p. 316.) The use in amplifier design is described. The circuit can be arranged either to suppress one particular frequency or to suppress all but one frequency. Numerical data for computing the network parameters are given.

621.392.52 : 512.831 1467

Applications of Matrix Algebra to Filter Theory.—P. I. Richards. (*Proc. Inst. Radio Engrs, N.Y.*, Part I, March 1946, Vol. 34, No. 3, pp. 145–150.) "After a brief introduction to matrix notation, methods are presented for the derivation of design equations for filter sections with special attention to symmetrical types. Finally, insertion and mismatch loss formulas, obtainable directly from the matrices, are given." The matrix notation for each of the circuit components normally encountered in filter design is given. It is shown that the matrix for the whole filter is formed by multiplying, in the order of connexion, the individual matrices of the circuit components. From a knowledge of the elements in the complete filter matrix it is shown how to derive the properties of the filter: conversely, the method may be used to design a filter having specified properties.

621.394/.397].645 1468

On Maximum Gain—Band Width Product in Amplifiers.—W. W. Hansen. (*J. appl. Phys.*, Feb. 1946, Vol. 17, No. 2, p. 109.) Correction to 3810 of 1945 (Hansen).

621.394/.397].645 1469

Radio Design Worksheet: No. 44—The Cathode Follower Circuit.—(*Radio, N.Y.*, Jan. 1946, Vol. 30, No. 1, p. 44.)

621.394/.397].645 1470

Network Design Using Electrolytic Tanks.—R. W. Kenyon. (*Electronic Industr.*, March 1946, Vol. 5, No. 3, pp. 58–60.) "Every amplifier has specific complex frequencies at which the gain is infinity [poles], and other complex frequencies at which the gain is zero [zeros]." In the electrolytic tank the locations of the poles and zeros (which uniquely determine the shape of the frequency/gain curve) are represented by electrodes suitably placed and energized. A probe electrode connected to a valve voltmeter enables the gain function to be determined. A short account of the design, and an outline of typical applications of the tank in amplifier design are given.

621.394/.397].645 : 621.396.822 1471

Noise Factor of Valve Amplifiers.—N. R. Campbell, V. J. Francis & E. G. James. (*Wireless Engr*, April 1946, Vol. 23, No. 271, pp. 116–121.) Conclusion of 1191 of May. A definition of noise factor is given, and on the assumption that all the quantities concerned are either zero or a constant value, formulae are derived and applied to transformer-coupled stages and to a common-cathode pentode stage. The effect on the noise factor of the addition of a pre-amplifier before the first stage of an amplifier is shown to be the same as if the amplifier produced no noise but the load conductance of the pre-amplifier were increased.

621.394/.397].645.34 1472

Selective Amplifiers.—B. M. Hadfield. (*Electronic Industr.*, March 1946, Vol. 5, No. 3, p. 104.) ". . . a frequency-selective amplifier or oscillator of the feedback type . . . which permits frequency control by adjustment of one potentiometer in the feedback circuit." Summary of U.S. Patent 2 386 892.

621.395.645 1473

Telephone Amplifier.—H. K. van Jepmond. (*Electronics*, March 1946, Vol. 19, No. 3, p. 139.) Simple two-way amplifier for use on privately serviced telephone lines.

621.395/.397].645 : 621.396.619.018.41 1474

Carrier-Frequency Amplifiers: Transient Conditions with Frequency Modulation.—C. C. Eaglesfield. (*Wireless Engr*, April 1946, Vol. 23, No. 271, pp. 96–102.) The "transient modulation ratio" is defined as the ratio of the modulation at the output to that at the input, and the general equations for it are obtained assuming an instantaneous change of amplitude or frequency at the input. The equations are simplified by considering small modulation values only, and by making the carrier frequency coincide with the central frequency of the amplifier. For "staggered" circuits a simple relation between modulation ratios for a.m. and f.m. is obtained, and graphical results for representative amplifiers are shown. With modification, the results can be used for coupled circuits. See also 1196 of May and 68 of January (Eaglesfield).

621.395.645 : 621.395.44 1475

A New Linear Amplifier Circuit.—S. T. Fisher. (*QST*, Feb. 1946, Vol. 30, No. 2, pp. 21–26.) A less detailed account of the system described in 877 of April.

621.395.645 : 621.395.8 1476

Hum Elimination.—J. C. Hoadley. (*Radio-Craft*, Feb. 1946, Vol. 17, No. 5, pp. 313..356.) A review of possible sources of mains-voltage hum in amplifiers together with design hints for its elimination. Some of the less obvious causes of hum, such as multiple grounding to chassis and emission from the exposed parts of cathode heater elements, are dealt with.

621.395.645 : 621.395.8 1477

The Effect of Negative Voltage Feedback on Power-Supply Hum in Audio-Frequency Amplifiers.—G. Builder. (*Proc. Inst. Radio Engrs, N.Y.*, Part II, March 1946, Vol. 34, No. 3, pp. 140–144.) Analysis leading to the conclusions (*a*) Under conditions of simple negative feedback, the signal/hum ratio at constant signal output is improved by the gain-reduction factor $(1-\beta M)$. This factor is, in general, complex and frequency-dependent. (*b*) "Failure to achieve the improvement in signal/hum ratio thus predicted may be due to the feedback voltage's including voltage other than the fraction β of the output voltage required for simple negative feedback . . ." (*c*) "In general, hum balancing within the amplifier is independent of the feedback only when the conditions for simple negative feedback are satisfied." (*d*) With an amplifier employing feedback, it is not generally valid to calculate the output hum voltage by simple potential division between the load impedance and the effective anode impedance of the valve if the latter is taken to be $R_a/(1-\mu\beta)$. A particular case is considered, however, when this procedure is permissible. The case of an amplifier having transformer coupling to the load is discussed in detail, with extensions to cover other types of output arrangements.

621.395.645.3 : 621.385.2 1478

A Voltage Amplifier Using a Pre-Saturation Diode as Load.—A. M. I. A. W. Durnford. (*Canad. J. Res.*, Sept. 1944, Vol. 22, Sec. A, No. 5, pp. 67–76.) "Pre-saturation characteristics for various types of diodes indicate that for constant plate voltage, the product of plate current and plate resistance is constant for each diode. $(i_d r_{pd} = K)$. For different diodes, the value of K varies from 200 to 700 V for a plate voltage of 125 V.. Tubes with thoriated tungsten cathodes have the highest K values with K proportional to plate voltage which renders the logarithmic characteristics linear and parallel. The pre-saturation diode when used as a load provides a high resistance that does not require an exceedingly large supply voltage. If the quiescent plate voltage, E_{bo}, is kept constant, the effective load lines converge at a common point, the voltage of which represents a virtual supply voltage equal to $E_{bo} + K$. Voltage gain and maximum distortionless output are both greater than when a fixed resistance is used as load. Variation of the filament current of the diode, or of a self-bias resistor, provides a distortionless gain control. The gain is almost constant for frequencies less than 300 c.p.s. but for high frequencies drops more rapidly."

621.395.645[.33 + .35 **1479**
A Constant Time Interval Reference Potential Indicator for Use with R-C Coupled Amplifiers.—E. W. Kammer. (*Rev. sci. Instrum.*, March 1946, Vol. 17, No. 3, pp. 102–106.) A device used for the observation of low-frequency signals such as those produced in the recording of dynamic strain in structures. The input signal is chopped rapidly to form a carrier which can then be amplified. The electronic switch, consisting of two cathode followers and a differential amplifier, is operated by a pulse with recurrence frequency 1 000 per sec and 10^{-4} sec duration.

621.395.645.34 + 621.396.63 **1480**
[Frequency-] **Selective Control Circuit.**—E. S. Purington. (*Radio, N.Y.*, Jan. 1946, Vol. 30, No. 1, p. 59.) A double-triode amplifier circuit with selective feedback designed to operate a relay when excited with a predetermined frequency of about 10 c/s. Summary of U.S. Patent 2 382 097.

621.396/.397].645.31 **1481**
Some Considerations in the Design of Wide-Band Radio-Frequency Amplifiers.—J. E. Cope. (*J. Instn elect. Engrs*, Part I, Feb. 1946, Vol. 93, No. 62, p. 109.) Summary of 576 of March.

621.396.611 : 621.396.615.14.029.63 **1482**
Tunable Microwave Cavity Resonators.—J. J. Guarrera. (*Electronic Industr.*, March 1946, Vol. 5, No. 3, pp. 80..122.) Design of coaxial resonators and associated components for use with disk-sealed triode oscillators, with particular reference to tuning, feedback, and shorting-plug problems. The design and testing of assemblies for operation under Service conditions are briefly discussed. See also 1483.

621.396.611 : 621.396.615.14.029.63 **1483**
Cavity Oscillator Circuits.—A. M. Gurewitsch. (*Electronics*, Feb. 1946, Vol. 19, No. 2, pp. 135–137.) Outline of the design of re-entrant type cavity circuits for use with disk-sealed triodes (lighthouse tubes) in the decimetre band of wavelengths. A number of different mechanical arrangements are described and illustrated: typical dimensions are given for wavelengths in the 9–30 cm region together with an indication of operating voltages and currents under c.w. and pulsed conditions.

621.396.611.2.029.62/.63 **1484**
Tuned Circuit Design for U.H.F.—M. Apstein & M. Joffe. (*QST*, Feb. 1946, Vol. 30, No. 2, pp. 13–16, 106.) Photographs and drawings of an LC circuit with three interchangeable coils, for use with an acorn triode to oscillate over the range 140–450 Mc/s.

621.396.611.3 : 621.396.615.1 **1485**
Bimodal Oscillator.—S. Lubkin. (*Electronics*, Feb. 1946, Vol. 19, No. 2, pp. 242..248.) Various examples are quoted of circuits having two closely-spaced natural frequencies. The advantages to be expected from using such circuit elements in the construction of beat-frequency oscillators, together with methods for varying the frequency separation, are discussed. Possible circuits are described; one suggested arrangement is to pulse-excite the double-peaked circuit, using the output beat frequency to synchronize the exciter.

621.396.615.17 **1486**
Rectangular-Pulse Generator.—R. K. McCombs & F. C. Walz. (*Electronics*, March 1946, Vol. 19, No. 3, pp. 222..228.) Long summary of 4136 of 1945.

621.396.615.17 : 578.088.7 **1487**
Nerve Stimulator.—Weiss. (*See* 1609.)

621.396.615.17 : 621.397.335 **1488**
Synchronizing Generators for Electronic Television.—A. R. Applegarth. (*Proc. Inst. Radio Engrs*, Part II, March 1946, Vol. 34, No. 3, pp. 128–139.) "The system of electronic circuits employed to generate the complex wave forms required as a base for television picture transmission is described. It comprises four principal sections which are: (1) source of accurately timed pulses; (2) frequency-divider chain; (3) components-generating circuits; and (4) signal-synthesis circuits. The various means for accomplishing these functions are briefly discussed, and illustrated by circuits which have been used successfully for such purposes in practical applications."

621.396.619 **1489**
Carrier Wave Modulation.—W. R. Bennett. (*Radio, N.Y.*, Jan. 1946, Vol. 30, No. 1, p. 42.) A circuit for reducing second-harmonic distortion in a ring-type balanced modulator when the signal frequency is too close to the sideband frequency to be conveniently separated by filters. Summary of U.S. Patent 2 377 858.

621.396.619 : 621.396.933.2 **1490**
Modulation Circuit.—Alford & Patterson. (*See* 1548.)

621.396.619.018.41 : 621.396.622 **1491**
Frequency Discriminator.—G. G. Bruck. (*J. appl. Phys.*, March 1946, Vol. 17, No. 3, p. 213.) Letter giving preliminary notice, with circuit diagram, of a discriminator that is insensitive to amplitude. "The circuit . . . is essentially an oscillator which falls into step with an incoming signal. The differential plate-current of the two halves of the [double triode] tube is the discriminator output."

621.396.621.53 **1492**
Calculation of the Output from Non-Linear Mixers.—H. Stockman. (*J. appl. Phys.*, Feb. 1946, Vol. 17, No. 2, pp. 110–120.) A detailed critical survey of existing methods of determining the mixer performance of non-linear conductors from an assumed knowledge of the current/voltage charac-

teristic. The methods are compared for consistency by application to a typical diode characteristic, and the discrepancies range from − 25% to + 35%. It is concluded that there is no "best" method and that "all methods require simplifying assumptions that frequently lead the converter designer to consider practical measurements or special methods such as the frequency-conversion diagram technique". The latter is a compromise between direct measurement and complete computation, due to Chaffee, and is described in detail in a number of papers by H. Stockman (*e.g.* 778 of 1945 and 408 of February).

621.396.645 + 621.396.61 **1493**

Analysis of Parasitic Oscillations in Radio Transmitters.—Jackson. (*See* 1714.)

621.396.645 **1494**

A 4.3-Mc. F.M./A.M. I.F. and Audio Amplifier.—J. W. Brannin. (*QST*, March 1946, Vol. 30, No. 3, pp. 51–54.) Circuit and constructional details. A beat oscillator is included for c.w. reception.

621.396.645.029.62 **1495**

A V.H.F. Amplifier Using the 829.—G. G. (*QST*, March 1946, Vol. 30, No. 3, pp. 55–56.) Constructional details. 50 W output at 144 Mc/s.

621.396.645.34 **1496**

Selective Amplifier or Oscillator.—B. M. Hadfield. (*Radio, N.Y.*, Jan. 1946, Vol. 30, No. 1, pp. 42–43.) A feedback amplifier with a narrow pass band that can be varied over several octaves by a single potential divider. It may also be used as an oscillator over the same frequency range. Summary of U.S. Patent 2 386 892.

621.396.662.32 **1497**

The Ideal Low-Pass Filter in the form of a Dispersionless Lag Line.—M. J. E. Golay. (*Proc. Inst. Radio Engrs, N.Y.*, Part I, March 1946, Vol. 34, No. 3, pp. 138–144.) "Some theoretical and practical aspects of the design . . . are considered. It is shown, in particular, that an artificial line made up of series inductances and shunt capacitances, with 18 per cent aiding mutual inductance between adjacent coils and 8 per cent shunt capacitance between alternate tie points, forms, for many purposes, a sufficiently good approximation of a dispersionless lag line. The mathematical study of the function $f(n) = \int_0^\pi (1 - \cos\phi/\phi^2) \cos n\phi d\phi$, which is of pertinent interest in lag-line theory, is given in an appendix." The modification of the values of line components to take account of resistive losses, and the effective characteristic impedance of the line are briefly examined.

621.396.662.34 **1498**

H.F. Band-Pass Filters: Part III.—H. P. Williams. (*Electronic Engng*, March 1946, Vol. 18, No. 217, pp. 89–93.) Examination of the response of coupled dissimilar circuits with unequal damping and staggering. The position and separation of the peaks, the gain on tune and the response well off tune are derived, and the results applied to interstage and aerial coupling circuits. An approximate treatment of the signal/thermal-noise ratio is given. For previous parts see 895/896 of April.

621.396.9 **1499**

Loran Indicator Circuit Operation.—Davidson. (*See* 1529.)

621.397.335 **1500**

Synchronizing and Separation Circuits: Part 12.—Noll. (*See* 1710.)

621.397.645 **1501**

Compensating Amplifier.—Gillespie. (*See* 1712.)

621.392.4 : 621.3.015.33 **1502**

Pulsed Linear Networks. [Book Review]—E. Frank. McGraw-Hill Book Co., New York, 267 pp., $3.00. (*Proc. Inst. Radio Engrs, N.Y.*, Part II, March 1946, Vol. 34, No. 3, pp. 159–160.) The networks examined are limited to simple types, but the treatment (by the classical method only) is clear and thorough.

GENERAL PHYSICS

532.517.4 **1503**

A Contribution to the Statistical Theory of Turbulence.—V. G. Nevzgljadov. (*C. R. Acad. Sci. U.R.S.S.*, 1945, Vol. 47, No. 7, pp. 466–468. In English.)

535.215 : 621.383 **1504**

Investigation of the Surface Photoelectric Effect of Metallic Films under the Influence of Strong Electrostatic Fields.—V. P. Jacobsmeyer. (*Phys. Rev.*, 1st/15th Jan. 1946, Vol. 69, Nos. 1/2, p. 50.) Photoelectric current from Bi films and a "liquid bright Pt" surface measured in fields up to 1.4×10^5 V/cm give a "Schottky line" which is consistent with the theoretical predictions of Guth & Mullin (1909 and 2766 of 1941). Summary of an Amer. Phys. Soc. paper.

535.3 **1505**

Transmission of Light by Water Drops 1 to 5 μ in Diameter.—R. Ruedy. (*Canad. J. Res.*, May/July 1944, Vol. 22, Sec. A, Nos. 3/4, pp. 53–66.) The size of growing drops of water formed as water vapour condenses can be determined by application of Mie's theory. As the drop size increases, the intensity of the transmitted light passes through a series of maximum and minimum values. This results in coloured light, and from observation of the cycles in changes of colour, the radius of the growing drops can be determined.

535.43 **1506**

Scattering of Light by Small Drops of Water.—R. Ruedy. (*Canad. J. Res.*, Dec. 1943, Vol. 21, Sec. A, No. 12, pp. 99–109.) Scattering of light of wavelength λ obeys Rayleigh's law for drop diameters up to λ/4. For diameters between λ/4 and λ, back radiation decreases almost to zero, and forward radiation obeys approximately a sixth power law. At the same time the colour of the scattered light changes gradually from blue, through the spectrum, to red.

536.2 **1507**

A Cylinder Cooling Problem.—S. A. Schaaf. (*Quart. appl. Math.*, Jan. 1946, Vol. 3, No. 4, pp. 356–360.) The problem of an infinitely long circular cylinder at an initial temperature above zero which is instantaneously immersed in an infinite medium initially at zero is solved by means of a Laplace transformation, leading to a solution in terms of Bessel functions.

536.2 : 621.315.2.017.7 **1508**

On a Modification of Forchheimer's Problem.—I. A. Charny. (*C. R. Acad. Sci. U.R.S.S.*, 1945, Vol. 48, No. 1, pp. 27–30. In English.) A mathematical paper on the determination of the heat losses from a length of tube buried in the ground, when the coefficient of heat transfer from the surface of the ground to the air has a finite value.

537.122 : 538.3 **1509**

Some Criticisms of the Theory of Point Electrons.—T. Lewis. (*Phil. Mag.*, Aug. 1945, Vol. 36, No. 259, pp. 533–541.) Dirac's analyses of the behaviour of point electrons (1822 of 1942 and previous work) contain a serious mathematical error, leading to false conclusions. Some of his results are invalid in so far as they are supposed to contain a term corresponding to radiation damping.

537.525 **1510**

On the Theory of the Varying Electric Discharge in Gases.—V. L. Granovsky. (*C. R. Acad. Sci. U.R.S.S.*, 1940, Vol. 26, No. 8, pp. 876–880. In English.) Varying discharge phenomena may be classified into (a) those associated with the electrical inertia of the discharge, (b) those associated with the thermal inertia of the gas and the electrode system. The paper deals with (a), with the following limitations :—1. The plasma is the main part of the discharge. 2. Displacement current small compared with conduction current. 3. Pressure not too low. 4. Plasma quasi-neutral (*cf.* Schottky's theory of ambipolar diffusion). 5. Maxwellian electron velocity distribution. 6. Energy of the electron gas derived from the electric field and lost by collision with gas molecules. 7. The normal atoms are directly ionized. 8. Tube connected to a source of e.m.f. through a resistance. Formulae are given for the balance of ions, the equation of ionization, the balance of energy in the electron gas, the equation of mobility and the current. See 1224 of May (Granovsky) for a sequel to this paper.

537.525 : 621.3.029.64 **1511**

Initiation of High Frequency Gas Discharges.—Holstein. (*See* 1726.)

537.525.82 **1512**

Computation of the Positive Column Characteristics.—B. Klarfeld. (*C.R. Acad. Sci. U.R.S.S.*, 1940, Vol. 26, No. 9, pp. 873–875. In English.) Deals with the calculation of the discharge characteristics from the atomic properties of the gas. The theory, now freed from previous simplifying assumptions which limited the pressure range, has been extended to cover the whole range within which the low-pressure plasma theory remains valid, and has been confirmed by experiments with mercury vapour. (See V. Granovsky, *Bull. Acad. Sci. U.R.S.S.*, Sér. phys., 1938, No. 4, p. 419.)

537.581 **1513**

An Explanation of Anomalous Thermionic Emission Current Constants.—N.-T. Sun & W. Band. (*Proc. Camb. phil. Soc.*, Feb. 1946, Vol. 42, Part 1, pp. 72–77.) Anomalously large or small values of the constant A in the thermionic current formula for metals, ($I = AT^2e^{-\chi/kT}$), are explained by taking into account the sharing of free electrons by two competing overlapping energy bands. Cases where one overlapping energy band is nearly full and where both overlapping bands are nearly empty are considered, and are used to explain the observed values of A for nickel and hafnium.

538.1 **1514**

Resonance Absorption by Nuclear Magnetic Moments in a Solid.—E. M. Purcell, H. C. Torrey & R. V. Pound. (*Phys. Rev.*, 1st/15th Jan. 1946, Vol. 69, Nos. 1/2, pp. 37–38.) The absorption of r.f. energy by a solid material in a strong magnetic field due to changes of orientation of nuclear spin has been observed. The experimental method of determining the proton magnetic moment (using the transition relation $h\nu = 2\mu H$) is described. A resonant cavity (frequency 29.8 Mc/s) filled with paraffin wax was placed in a magnetic field and the r.f. power transmitted through the cavity balanced by a direct signal in antiphase. The magnetic field was varied until the very sharp resonance absorption (about 10 oersteds wide) was observed at 7 100 oersteds, giving 2.75 nuclear magnetons for the proton moment. The relaxation time to establish thermal equilibrium between the spins and the lattice was apparently less than a minute.

538.3 **1515**

Initial Boundary Problems of Electrodynamics.—J. N. Feld. (*C.R.Acad.Sci.U.R.S.S.*, 1945, Vol. 48, No. 3, pp. 172–174. In English.) A formal solution of the electromagnetic field within a space v bounded by a closed surface s when the tangential components of the electric and/or magnetic vectors are given as arbitrary functions of time. Equations (6) and (7) give the electric and magnetic fields at the point of observation assuming zero conductivity within v and zero tangential field at the surface s.

538.3 **1516**

A Study of Stationary Electromagnetic Modes for Region Between Parallel Perfectly Conducting Planes and Application to Electron Accelerator.—E. S. Akeley. (*Phys. Rev.*, 1st/15th Jan. 1946, Vol. 69, Nos. 1/2, p. 50.) Theoretical evaluation of the TM modes with circular symmetry and their relation to the design of the surfaces and the energy dissipation in an electron accelerator. Summary of an Amer. Phys. Soc. paper.

538.3 **1517**

Reciprocal Electric Force.—F. W. Warburton. (*Phys. Rev.*, 1st/15th Jan. 1946, Vol. 69, Nos. 1/2, p. 40.) From the assumed potential energy of two charges e and e' with relative velocity u the reciprocal force between them is obtained in a form involving certain undetermined coefficients and the relative acceleration of the charges. When the radius vector, velocity and acceleration are parallel, the usual mass-energy relation is found. The formulae are applied to calculating the change of magnetization and torque of a rod by a longitudinal current. It is suggested that the treatment provides a unified electromagnetic theory which is more complete than the conventional theory with the necessary relativity corrections.

553.631 : 621.3.011.2 **1518**

Effect of Transverse Pressure on the Steady-State Electrical Conductivity of Rocksalt.—Hamtil. (*See* 1557.)

621.314.632 **1519**

A Method for Measuring Effective Contact E.M.F. between a Metal and a Semi-Conductor.—W. E. Stephens, B. Serin & W. E. Meyerhof. (*Phys. Rev.*, 1st/15th Jan. 1946, Vol. 69, Nos. 1/2, pp. 42–43.) Bethe's theory for the current in a rectifier formed by the potential barrier between a metal and a semiconductor gives $j = j_0 \exp\left(-\frac{e\phi}{kT}\right)$ $[\exp\left(\frac{eV}{kT}\right) - 1]$ where j_0 is the available current, ϕ the effective contact e.m.f., and V the applied voltage across the contact. If the resistance of the contact at zero applied voltage is R at temperature T, and $\log(R/T)$ is plotted against $1/T$, the effective contact e.m.f. ϕ can be deduced. [Reference is made to Fig. 1 which has been omitted.] ϕ is of importance in the operation of rectifiers and thermistors, and by correcting it for image force and the tunnel effect, the true difference of work function can be estimated.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.5 : 621.396.11 **1520**

Meteoric Impact Ionization Observed on Radar Oscilloscopes.—O. P. Ferrell. (*Phys. Rev.*, 1st/15th Jan. 1946, Vol. 69, Nos. 1/2, pp. 32–33.) Report on short-duration echoes observed on about 40 Mc/s and 100 Mc/s and attributed to ionic clouds formed by meteoric impact. On 105 Mc/s the echoes lasted for about ½–3 sec, and occurred at ranges of 30–125 km, with maximum rate of occurrence in the early morning (0100–0700) peaking at about 0400 hours.

523.72 **1521**

Radio Noise from the Sun.—Department of Scientific and Industrial Research. (*Electronic Engng*, March 1946, Vol. 18, No. 217, p. 98.) A brief account of the phenomenon described in 323 of February (Appleton).

551.515.43 : 621.396.9 **1522**

Spotting Hurricanes and Thunderstorms by Radar.—Winters. (*See* 1541.)

531.437 : 621.3.011.2 **1523**

The Use of Cumulative Resistance in Earth-Resistivity Surveys.—R. Ruedy. (*Canad. J. Res.*, July 1945, Vol. 23, Sec. A, No. 4, pp. 57–72.) When the soil consists of layers having different resistivities, an almost linear relation is obtained between the cumulative resistance and the electrode spacing until the distance between electrodes is equal to the thickness of the upper material. Earth-resistivity and cumulative-resistance curves are interpreted for some typical terrains.

551.5 (021) **1524**

Dynamic Meteorology. [Book Review] — J. Holmbe, G. E. Forsythe & W. Gustin. John Wiley & Sons, New York, $4.50, Chapman & Hall, London. (*Curr. Sci.*, Feb. 1946, Vol. 15, No. 2, p. 54.) "... an ideal and up-to-date text-book for the advanced student."

LOCATION AND AIDS TO NAVIGATION

621.383 **1525**

Sensory Aid for the Blind.—Cranberg. (*See* 1520.)

621.396.62 + 621.396.9 **1526**

Germany's UHF Tubes [Radar Camouflage].—Combined Intelligence Sub-Committee. (*See* 1648.)

621.396.82 : 621.396.67 **1527**

QRM—The Electronic Life Saver: Part II.—Robbiano. (*QST*, Feb. 1946, Vol. 30, No. 2, pp. 27–35.) Description of receiving equipment and of aerial systems used in radar countermeasures, including a cone aerial for 300–3 000 Mc/s, and a "fish hook" aerial for 500 Mc/s. For part I dealing with jamming transmitters, see 1060 of April.

621.396.9 **1528**

I.R.E. Winter Technical Meeting January 1946.—(*Waves and Electrons*, Feb. 1946, Vol. 1, No. 2, pp. 80 . . 92.) Abstracts are given of the following papers read at the meeting. Considerations of Frequency, Power, and Modulation for a Long-Range Radio Navigation System.—P. R. Adams. Broad-Band Antennas and Direction-Finding Systems for Very High Frequencies.—A. Alford, J. D. Kraus, A. Dorne & J. Christensen. Naval Airborne Radar.—L. V. Berkner. Airborne Radar Equipment for Aircraft Interception.—F. L. Holloway, R. P. Burrows & J. E. Keto. The Theory and Application of the Radar Beacon.—R. D. Hultgren & L. B. Hallman, Jr. Enemy Radio and Radar Equipment.—E. L. Luke & J. C. Link (Title only). Application of Radar Techniques to Aircraft Fire-Control Systems.—E. A. Massa, I. Paganelli & F. A. Best, Jr. The Role of Electronics in Antiaircraft Gun-Fire Control.—F. B. MacLaren. Aircraft Automatic Position Plotter.—A. C. Omberg & W. L. Webb. Radar Model XAF.—R. M. Page. An Introduction to Hyperbolic Navigation.—J. A. Pierce. A Frequency-Modulated Altimeter for Meter and Light Indication and the Automatic Altitude Control of Aircraft.—R. C. Sanders, Jr., W. R. Mercer, J. Wolff & J. C. Smith (Title only). An Automatic Visual-Indication Radio Direction Finder.—A. Scandurra & S. Stiber. Ground Controlled Approach.—E. Storrs, W. Devitt & B. Green. Radar Aspects of Naval Fire Control.—D. P. Tucker. A Pulse Altimeter for High Accuracy at High Altitudes.—I. Wolff, W. D. Hershberger, G. W. Leck & R. R. Welsh. (Title only). For other abstracts see *Electronics*, March 1946, Vol. 19, No. 3, pp. 92 . . 108, and *Electronic Industr.*, March 1946, Vol. 5, No. 3, pp. 62 . . 74. For titles of other papers read, see other sections.

621.396.9 **1529**

Loran Indicator Circuit Operation.—D. Davidson. (*Electronic Industr.*, March 1946, Vol. 5, No. 3, pp. 84 . . 132.) Brief discussion of propagation factors in Loran, leading to the following requirements in the design of the receiver-indicator:—"(1) a *slow* time base . . . to examine the entire recurrence cycle (40 000 μs); (2) a *fast* . . . time base (say, 200 to 2 500 μs) to enable identification of pulse train components and to permit a coincidence measurement of master and slave pulses; (3) a *delay* circuit . . . so that a coincidence measurement can be made [that] will result in a time difference reading equal to the delay; (4) an *amplitude balance* scheme so that the two pulses . . . may be made identical in shape during measurement; (5) an effective method for sweep calibration in order that the time difference can be read with the required precision and rapidity; (6) a method of splitting the time base so that the pulse from one station may be brought under that of its mate for superposition; (7) a method of identifying the recurrence rates of the station pairs in a positive manner; (8) ample sensitivity in the receiver . . . ; (9) the receiver bandwidth should be wide enough to allow satisfactory discrimination of sky and ground wave components and yet narrow enough to keep adjacent channel interference at a minimum." The manner in which these requirements have been met in the DAS–1 equipment for shipboard use is discussed in detail. The design of the airborne model is briefly mentioned, together with other variants of the basic design, in existence or contemplated.

621.396.9 **1530**

Loran Transmitting Stations.—D.G.F. (*Electronics*, March 1946, Vol. 19, No. 3, pp. 109–115.) A description of the circuits and operating functions of the transmitter, timer, and synchronizer. The transmitter, generating 40-μs pulses at 1.7–2.0 Mc/s at a peak power of 75–150 kW, consists of a self-excited tuned-grid, tuned-anode, push-pull oscillator, pulse modulated in the cathode circuit. The modulator is driven by two exciters, so that two pulse sequences at slightly different rates can be generated simultaneously when necessary. The timing unit consists of a 50-kc/s quartz-crystal oscillator-circuit and six divider-circuits. The timer repeats the pulses, at 25 or $33\frac{1}{3}$ pulses per second, so precisely that the accumulated timing error does not exceed 1 μs over a period of several minutes. To synchronize the pulse sequence of the slave station with that of the master station, the remote and local signals are displayed on a cathode-ray indicator and the phase of the transmitted keying system is adjusted until the two signals are superimposed. The phase adjustment may be made manually, or automatically, by switching in the synchronizing unit which holds the two sequences in synchronism to an accuracy of 1 μs. This paper concludes a series. For previous parts see 605/606 of March.

621.396.9 **1531**

Loran—the Latest in Navigational Aids : Part III. Navigators' Equipment and Summary.—A. A. McKenzie. (*QST*, Feb. 1946, Vol. 30, No. 2, pp. 62 . . 124.) For previous parts see 926 of April.

621.396.9 **1532**

Loran — Radio Navigation Aid.—E. F. Brissie. (*Radio Craft*, Jan. 1946, Vol. 17, No. 4, pp. 236 . . 292.) A simple account of the system.

621.396.9 **1533**

Technical and Tactical Features of Radar.—J. H. DeWitt, Jr. (*J. Franklin Inst.*, Feb. 1946, Vol. 241, No. 2, pp. 97–123.) Military uses of ground radar include aircraft warning, gun laying, searchlight control, and ground-controlled interception. The basic principles are common to all these systems, the essential components being a timer, transmitter, antenna system, receiver, indicator, and power-supply unit. These components, and the operational use of a radar system, are discussed in general terms, and details are given of sets designed for various specific requirements. Precision must often be sacrificed in operational sets to obtain simplicity and to reduce weight.

621.396.9 **1534**

The [AN/] MPG-1 Radar.—H. A. Straus, L. J. Rueger, G. A. Wert, S. J. Reisman, M. Taylor, R. J. Davis & J. H. Taylor. (*Electronics*, March 1946, Vol. 19, No. 3, pp. 140–147.) Details of the timing, p.p.i., B-indicator, and remote-B-indicator systems. Crystal-controlled sweeps, together with a special quadrant-type capacitive phase-shifter, enable a static range accuracy of $\pm$ 3 yd to be obtained. When calibrated by transmitted pulses, the accuracy is about $\pm$ 20 yd. For previous parts of this 3-part series, see 610 of March and 1250 of May.

621.396.9 **1535**

Radar on 50 Centimeters.—H. A. Zahl & J. W. Marchetti. (*Electronics*, Feb. 1946, Vol. 19, No. 2, pp. 98–103.) Details of the transmitter, receiver and display systems of the AN/TPS–3 equipment of which other parts were described in 1249 of April. The transmitter comprises four triodes in parallel push-pull within a common envelope which also contains the grid and anode resonant lines. Tuning is effected externally by operation of a shorting bar on the filament lines. Power is taken by direct coupling to the anode line. The valve is modulated by 1.5-μs pulses from an artificial-line circuit discharged 200 times per second by a rotary spark gap : peak output power is 200 kW at 25–30% efficiency. The receiver comprises two stages of signal-frequency amplification, a crystal first detector, 6 i.f. stages (bandwidth 1.25 Mc/s), a diode second detector, and separate video stages for the A-scope and p.p.i. The noise figure of the receiver is about 10 db. Timebase lengths equivalent to 20, 60 and 120-mile ranges are provided for both display tubes. Separate timebase units, each triggered from the transmitted pulse, are provided, the timebase length in each case being controlled by variation of the characteristics of an asymmetric multivibrator. In the case of the p.p.i. the linear sweep voltage is distributed to the deflector coils by a rotary transformer (goniometer), the search coil of which rotates in synchronism with the aerial system, and carries the sweep currents. Range markers at 10-mile intervals are provided on the A-scope and p.p.i., use being made of a 9.3-kc/s oscillator keyed by the transmitted pulse.

621.396.9 **1536**

Lightweight Radar for Early Warning.—W. C. Hendricks. (*Communications*, Jan. 1946, Vol. 26, No. 1, p. 54.) Some details of the radar set AN/TPS–3. For a longer account see 1535 above.

621.396.9 **1537**

The SCR-584 Radar [Part III].—D.G.F. (*Electronics*, Feb. 1946, Vol. 19, No. 2, pp. 110–117.) " Details of the circuits for timing echoes to 0.01 microsecond, the deflection system for the plan position indicator, and the automatic antenna positioning gear are presented in this final installment." See 608/609 of March for earlier parts. A crystal oscillator at 81.95 kc/s is used to develop circular traces (on separate cathode-ray tubes) with circumferences equivalent to ranges of 32 000 yd and 2 000 yd respectively, and also a trigger at 1 707 c/s for actuating the transmitter. Signals appear as radial deflexions. A cursor on the 32 000-yd screen can be rotated to cover the selected echo. The 2 000-yd display is " gated " so that the trace is illuminated only for the equivalent of $\pm$ 250 yd on each side of the echo selected by the cursor on the long-range tube. A cursor on the 2 000-yd tube enables the range to be measured accurately. The p.p.i. circuits are arranged to provide magnetically deflected sweeps of either 70 000 or 35 000-yd range. Precautions are taken to ensure that the spot is at the centre of the tube at the beginning of each radial sweep. Intensity modulation is provided by the following signals (a) a signal to ensure that the tube is illuminated only during the outward sweep of the spot to the edge of the tube, (b) echoes from targets, (c) 10 000-yd marker pulses derived from a 16.4-kc/s oscillator shock-excited by the transmitted pulses, (d) an azimuth-marker pulse derived from the cursor setting on the long-range dial. Automatic target-following is provided.

621.396.9 **1538**

Elements of Radar: Parts II & III.—J. McQuay. (*Radio Craft*, Jan. & Feb. 1946, Vol. 17, Nos. 4 & 5, pp. 246..287 & 317..337.) Brief general descriptions of oscillators, aerials and waveguides, and of receivers and display systems. For part I see 930 of April.

621.396.9 **1539**

I.F.F.—(*Radio Craft*, Feb. 1946, Vol. 17, No. 5, p. 332.) A simple explanation of the principles of the system used during the war for distinguishing between enemy and friendly aircraft.

621.396.9 : 518.5 **1540**

Post Office Equipment for Radar.—P. A. Marchant & K. M. Heron. (*P.O. elect. Engrs' J.*, Jan. 1946, Vol. 38, Part 4, pp. 117–120.) The solution is given of the problems concerning the rapid interpretation of radar observations to a form suitable for teleprinter transmission to a control centre. The technique is an extension of multichannel selection circuitry.

621.396.9 : 551.515.43 **1541**

Spotting Hurricanes and Thunderstorms by Radar.—S. R. Winters. (*Radio News*, March 1946, Vol. 35, No. 3, pp. 45..104.) An illustrated account of the plotting of the path of a hurricane. See also 918 of March.

621.396.9 : 621.3.089.6 **1542**

Artificial Radar Target.—(*Electronics*, Feb. 1946, Vol. 19, No. 2, pp. 214, 216.) The radar pulse received on a dipole is heterodyned to a lower frequency, converted by a piezoelectric transducer to a pulse of mechanical vibration which is delayed by transmission along a glass rod, is reconverted by the transducer, and retransmitted at carrier frequency from the dipole to simulate a radar echo. The device is set up about 20 yards from the radar set, and, on account of the time delay, the echo appears as if from a distant target. It is used for test purposes. A German device.

621.396.9 : 623.26 **1543**

Land Mine Locators.—West. (See 1626.)

621.396.9 : 623.454.25 **1544**

Proximity Fuzes for Artillery.—Selvidge. (See 1627.)

621.396.931/.933].22.029.54/.56 **1545**

Twin Bearing DF Unit.—(*Electronic Industr.*, March 1946, Vol. 5, No. 3, pp. 79, 124.) Brief description of a compact, rotating, cross-loop, twin-channel d.f. with crossed-pointer indication—the Simon Radioguide, a variant of SCR-503-A. Frequency range 0.1 to 3.0 Mc/s in two units. Messages may be read while bearings are being taken.

621.396.933 **1546**

The Teleran Proposal.—P. J. Herbst, I. Wolff, D. Ewing & L. F. Jones. (*Electronics*, Feb. 1946, Vol. 19, No. 2, pp. 124–127.) Teleran (Television-Radar Air Navigation) is "a system of navigation and traffic control, utilizing existing television and radar techniques to present visual information directly to the pilot . . ." It comprises, in each aircraft, a television receiver, and a transponder beacon coupled to the altimeter. The transponder is interrogated by ground radar equipments located near each aerodrome, and coding introduced by the altimeters in the respective aircraft enables the position and track of the machines within given height limits to be displayed on p.p.i. tubes particular to the respective strata. These pictures are televised, together with superposed information (map of neighbourhood, wind velocity, visibility, etc.), to the aircraft. The pilot receives the information appropriate to his height; he thus sees a map of the aerodrome neighbourhood, his own position and the position and course of other aircraft at roughly his own height. The system is designed also to assist the aircraft in landing. The properties of Teleran are compared with those of other navigation systems.

621.396.933.1/.2 **1547**

Radio and Radar Aids to Aerial Navigation.—R. L. Rod. (*Radio, N.Y.*, Vol. 30, No. 1, pp. 35..60.) An outline of systems, including the radio range, v.h.f. direction finder, Loran, the radar p.p.i., and Racon (radar beacon).

621.396.933.2 : 621.396.619 **1548**

Modulation Circuit.—A. Alford & G. K. Patterson. (*Radio, N.Y.*, Jan. 1946, Vol. 30, No. 1, pp. 41–42.) A circuit to produce an ordinary modulated carrier, and, simultaneously from different terminals, the sidebands without the carrier. Its use is proposed for a beacon system, in which the carrier is applied to one member of an aerial array, and sidebands only are applied to the other members, so that the modulation of the received signal is directional, but a constant carrier is available for a.v.c. Summary of U.S. Patent 2 383 456.

621.396.029.64 **1549**

Radiation Laboratory Technical Series. [Book Notice]—(*J. appl. Phys.*, Feb. 1946, Vol. 17, No. 2, pp. 105–106.) The results of 5 years' wartime work on radar ("20 000 technical man-years") are to be embodied in a series of 28 books on the physics and engineering of microwave radio. "For the first time the technical literature of a large subject is being created all at once on a uniform basis." The book will be prepared by staff of the Radiation Laboratory, with British collaboration, and will be published by the McGraw-Hill Book Company. See also *Electronics*, Feb. 1946, Vol. 19, No. 2, pp. 254..262.

MATERIALS AND SUBSIDIARY TECHNIQUES

531.788+533.5 **1550**

Audio Aid for Vacuum-Leak Hunting.—V. Wouk. (*Electronics*, Feb. 1946, Vol. 19, No. 2, pp. 138–141.) For use with vacuum systems equipped with gauges which provide a variation of output voltage when there is a change in partial pressure of one or more of the gases in the system. The outside of the vacuum system is sprayed at the doubtful points with the appropriate gas; when a leak is discovered the change in d.c. output voltage (amplified if necessary) is applied to the grid of a thyratron relaxation oscillator, thus causing a frequency change. See also 1551 below.

531.788+533.5 **1551**

Frequency Modulated Oscillator for Leak Hunting.—W. M. Brubaker & V. Wouk. (*Rev. sci. Instrum.*, March 1946, Vol. 17, No. 3, pp. 97–98.) Leaks representing pressure rises of less than 4×10^{-8} mm have been detected. The circuit can also be used for general monitoring purposes. See also 1550 above.

531.788.6 **1552**

A Pirani Gauge for Use at Pressures up to 15 mm.—E. S. Rittner. (*Rev. sci. Instrum.*, March 1946, Vol. 17, No. 3, pp. 113–114.) Thin tungsten wire is supported inside pyrex capillary tubing, and used in a constant-resistance bridge circuit. Pressures up to 15 mm Hg can be recorded with an accuracy of ± 2.5% or better.

533.5 **1553**

Iron-Nickel-Cobalt Alloy for Sealing to Glass.—G. D. Redston & J. E. Stanworth. (*J. sci. Instrum.*, March 1946, Vol. 23, No. 3, pp. 53–57.) A report on some Kovar-type alloys for sealing to borosilicate glass. Expansion curves are shown, and the effect of composition is discussed. An alloy containing 29 ± 0.5% Ni, 17% Co, 0.3% Mn, 0.15% Si is proposed. Impurities affect the expansion, so the alloy is specified by comparing with a molybdenum rod. The differential expansion coefficient should be zero at 25° C, and the curve should pass through $(3 \pm 1) \times 10^{-4}$. The stresses in sandwiches of alloy and glass are discussed, using measured stress/temperature curves. It is shown that seals can be made with very low stresses at all temperatures.

533.5 : 621.791.3 **1554**

A Simple [laboratory] **Method of Sealing Gas- or Vacuum-Packed Tins.**—W. A. Bryce & H. Tessier. (*Canad. J. Res.*, Sept. 1945, Vol. 23, Sec. F, No. 5, pp. 304–305.) ". . . The heating element is a coil of resistance wire supported over a hole in a flat surface of the tin. When the heating circuit is closed, a small piece of solder previously hung in the upper end of the coil is melted and drops on the area about the hole and thereby produces an effective seal."

535.87 : 539.234 **1555**

High-Reflexion Films.—K. M. Greenland. (*J. sci. Instrum.*, March 1946, Vol. 23, No. 3, pp. 48–50.) An account of single- and multiple-layer films for neutral or coloured filters and beam splitters operating on the interference principle, and having a high optical efficiency. They are made by high-vacuum evaporation.

539.232 : 621.317.794 **1556**

Production and Properties of Nickel Bolometers.—F. G. Brockman. (*J. opt. Soc. Amer.*, Jan. 1946, Vol. 36, No. 1, pp. 32–35.) Filaments as thin as 0.1 micron are obtained by nickel-plating copper foil. Ribbons of foil are soldered to a platinum frame and the copper dissolved in potassium cyanide solution, leaving nickel ribbons with a temperature coefficient of resistance of 0.005 per deg, and a time constant of 5 millisec. An equation is derived relating resistance to the ambient temperature and current. Bismuth ribbons have been similarly prepared.

553.631 : 621.3.011.2 **1557**

Effect of Transverse Pressure on the Steady-State Electrical Conductivity of Rocksalt.—C. N. Hamtil. (*Phys. Rev.*, 1st/15th Jan. 1946, Vol. 69, Nos. 1/2, p. 50.) Transverse pressure of 33 kg/cm² produced 7% increase of d.c. conductivity (at 100 V) in the temperature range 300–317°C, attributed to the reduction of the polarization counter-e.m.f. Summary of an Amer. Phys. Soc. paper.

62 "1945" **1558**

Progress in Engineering Knowledge during 1945: Materials Engineering.—P. L. Alger, J. Stokley, C. F. Scott, H. B. Marvin, J. L. Tugman & K. W. Given. (*Gen. elect. Rev.*, Feb. 1946, Vol. 49, No. 2, pp. 9–19.) A review, with an extensive bibliography, covering many materials for mechanical and electrical use.

621.315.613.1 **1559**

Some Physical Properties of Mica.—P. Hidnert & G. Dickson. (*Bur. Stand. J. Res.*, Oct. 1945, Vol. 35, No. 4, pp. 309–353.) Samples of mica from different sources have widely different physical and electrical properties. The coefficients of thermal expansion of samples have been measured in a direction normal to the cleavage plane under a pressure of 30 lb/sq. inch and at temperatures up to 700° C. In some cases coefficients are only a few parts in 10^6 per deg C; in others they are 5% per deg over small temperature ranges. Large changes between initial and final dimensions may result from a heating and cooling cycle, and successive cycles may produce quite different effects. Power factors of raw samples, measured at 100 kc/s and 1 000 kc/s lie between 0.03% and 1%, but may be considerably different at the two frequencies. In general, a heating and cooling cycle causes a substantial increase in power factor. Heating also produces changes in opacity and colour, and X-ray diffraction photographs show that the overall physical changes are often associated with changes in fine structure.

621.315.616 **1560**

Synthetic Rubbers and Plastics: XI (Part I) Water and the High Polymer.—F. T. White. (*Distrib. Elect.*, Apr. 1946, Vol. 18, No. 162, pp. 107–110.) The water absorption characteristics of various organic materials are discussed in relation to molecular structure, with particular reference to the effect of the presence of the hydroxyl (–OH) group. The absence of this group in a polymer chain generally means low absorption, and *vice versa*. For part X, see 585 of March.

621.315.616.011.2 (213) **1561**

Some Wartime Problems with Electrical Insulating Materials.—S. A. Prentice. (*J. Instn Engrs Aust.*, Oct./Dec. 1945, Vol. 17, Nos. 10/12, pp. 197–204.) The chief conditions under which the performance of insulating materials is impaired are temperature extremes, humidity, and fungus growth. Curves are given for some phenolic resins showing the decay of insulation resistance with time under various conditions of temperature and humidity. The adverse effect of a period of storage at high humidity is illustrated. A cyclic humidity change also produces gradual deterioration. Ceramics are free from these defects, but surface conductivity is often serious. Some thermoplastic materials, *e.g.* polyvinyl chloride, are little affected by humidity, but most are subject to distortion at temperatures above about 70° C. Humidity effects are minimized by treatment with varnishes or waxes. Conditions of high humidity are conducive to fungus growth, which leads to surface leakage, directly, and also by assisting condensation. The required properties and applications of a number of base materials and protective coatings used in insulation are tabulated, and testing techniques described.

621.315.616.011.2 (213) 1562

The Effect of High Humidity and Fungi on the Insulation Resistance of Plastics.—J. Leutritz, Jr. & D. B. Herrmann. (*ASTM Bull.*, Jan. 1946, No. 138, pp. 25–32.) A description of experiments determining the effect of prolonged exposure to fungi and 97% humidity on the insulation resistance of methyl methacrylate, glass-bonded mica, glass mat laminate phenolic, phenon fabric, phenol fibre, and wood-flour-filled phenol plastic. Lowering of resistance is produced, reaching a steady value in a period varying from a few hours to several weeks, according to the material. Fungus appears on all specimens during this period, but even in its presence, resistance recovery eventually occurs on reduction of humidity to 52%. Re-exposure to 97% humidity causes a rapid drop in insulation resistance, even in the case of methyl methacrylate, which, on initial exposure in the absence of fungus, shows only a very gradual decrease. In all cases, original insulation can be recovered by cleaning and drying, and deterioration under humid conditions can be retarded by surface varnishing. Water adsorption and absorption determine insulation degradation, the additional effect of fungus being negligible. A comprehensive bibliography is given.

621.357.5/.6] : 621.396.69 1563

Electroforming Microwave Components.—F. Hassell & F. Jenks. (*Electronics*, March 1946, Vol. 19, No. 3, pp. 134–138.) Use is made of electroforming techniques to overcome the difficulty of obtaining accurately machined surfaces on the insides of waveguides and other microwave circuit components. Where the shape is such that the mandrel can be withdrawn from the finished component, use is made of differential expansion on heating to secure separation of plating and mandrel. Alternatively, a separating film of tin is first deposited on the mandrel and the film is melted before the mandrel is withdrawn. Where the shape prevents withdrawal, the mandrel is made of fusible alloy, wax, or of a material easily dissolved. The electroformed products are stress-free. The process is "noted for precision, if not for economy".

621.385.1.032.2 + 533.5 + 539.234 1564

Fine Wires in the Electron-Tube Industry.—A. Espersen. (*Proc. Inst. Radio Engrs, N.Y.*, Part II, March 1946, Vol. 34, No. 3, pp. 116–120.) ". . . Some fundamental basic properties which confront the wire manufacturer are briefly discussed. Design formulas, including a nomograph, are given for electron-tube filaments. The use of coatings of gold, platinum, and zirconium on metals of the refractory group have assisted in the reduction of grid emission. A unique method of utilizing zirconium, both to accelerate the vacuum exhaust process and to serve as a continuous "getter", is described. A novel method of securing a uniform rate of evaporation of thin films of metals is discussed." The nomograph gives the length and diameter of tungsten wire needed to obtain a desired operating temperature under various conditions of voltage and current.

621.385.8.032.7 1565

Glass Problems in the Manufacture of Miniature Tubes.—H. J. Miller. (*Waves and Electrons*, Feb. 1946, Vol. 1, No. 2, p. 87.) Abstract of a paper read at the I.R.E. Winter Technical Meeting, January 1946.

679.5 1566

Frictional Heat.—C. A. Breskin. (*Sci. Amer.*, Jan. 1946, Vol. 174, No. 1, pp. 20–22.) A new process for welding, bonding, and moulding thermoplastics by generation of frictional heat. Bonding to non-plastics is also possible.

679.5 : 53.084 1567

Plastics and [plastic] Cements: Applications to Scientific Instruments.—J. C. Swallow. (*J. sci. Instrum.*, March 1946, Vol. 23, No. 3, pp. 44–48.) A discourse delivered at the Physical Society's thirtieth annual exhibition.

778 : 5 + 6 1568

Photography in Research and Development.—W. H. Banyard. (*Distrib. Elect.*, Apr. 1946, Vol. 18, No. 162, pp. 115–118.)

MATHEMATICS

512.831 : 621.392.52 1569

Applications of Matrix Algebra to Filter Theory.—Richards. (*See* 1467.)

517.93 + 518.12 1570

Solution of Linear and Slightly Nonlinear Differential Equations.—S. A. Schelkunoff. (*Quart. appl. Math.*, Jan. 1946, Vol. 3, No. 4, pp. 348–355.) A method based on the idea that solutions of linear differential equations may be regarded as distorted or "perturbed" sinusoidal or exponential functions, similar to the Rayleigh-Schrödinger treatment. Better results are obtained than by Picard's method which regards the solutions as perturbed straight lines. The method would be suitable for numerical solution of at least a certain class of differential equations, though it was originally developed to obtain convenient analytical approximations to a number of problems in wave theory.

517.947.4 1571

A Method of Solution of Field Problems by means of Overlapping Regions.—H. Poritsky & M. H. Blewett. (*Quart. appl. Math.*, Jan. 1946, Vol. 3, No. 4, pp. 336–347.) "In problems involving the determination of fields, it often happens that the region R for which the field is to be determined is difficult to handle directly, but can be broken up into several overlapping regions R_1, R_2, . . . for each of which the field can be determined by standard methods. We suppose that the breaking up is carried out in such a manner that every point of the region R falls into *at least one* of the regions R_1, R_2, . . ." The procedure is illustrated by consideration of the problem of propagation of a transverse electromagnetic wave round a corner formed by the junction at right angles of two regions bounded by infinite parallel plates.

517.948 **1572**

Some Integral Equations of Potential Theory.—H. Bateman. (*J. appl. Phys.*, Feb. 1946, Vol 17, No. 2, pp. 91–102.) Deals with integral equations of the type $f(x) = 1/\pi \int_{-\infty}^{\infty} \frac{yF(t)dt}{(x-t)^2 + y^2}$ where y is a "fixed positive quantity", $f(x)$ is an analytic function supposed to be determined from observation, and $F(t)$ is the function to be determined. Such integrals arise from inverse potential problems. The function $F(t)$ is required to be such that the integral exists and represents a potential $V(x, y)$ regular for $y > 0$, which is the solution of the Dirichlet problem for the half-plane.

518.61 **1573**

Calculation of the Magnetic Field in Dynamo-Electric Machines by Southwell's Relaxation Method.—H. Motz & W. D. Worthy. (*J. Instn elect. Engrs*, Part I, Feb. 1946, Vol. 93, No. 62, pp. 108–109.) Summary of 658 of March.

519.2 **1574**

On a Problem Connected with Purely Discontinuous Random Processes.—V. M. Dubrovsky. (*C. R. Acad. Sci. U.R.S.S.*, 1945, Vol. 47, No. 7, pp. 459–461. In English.)

518.2 **1575**

Table of Arc Sin X. [Book Review]—Mathematical Tables Project. Columbia Univ. Press, New York, 1945, 124 pp., $3.50. (*Proc. Inst. Radio Engrs, N.Y.*, Part II, March 1946, Vol. 34, No. 3, p. 160.) Gives values in radians to 12 decimal places, with 0.000 1 intervals from 0 to 0.989 0 and 0.000 01 intervals from 0.989 00 to unity. Interpolation aids are included.

MEASUREMENTS AND TEST GEAR

621.317 **1576**

I.R.E. Winter Technical Meeting January 1946.—(*Waves and Electrons*, Feb. 1946, Vol. 1, No. 2, pp. 80..93.) Abstracts are given of the following papers read at the meeting. Test Equipment and Techniques for Airborne Radar Field Maintenance.—E. A. Blasi & G. C. Schutz. A Spectrum Analyzer for Microwave Pulsed Oscillators.—F. J. Gaffney. Electrical Characteristics of Quartz-Crystal Units and Their Measurement.—W. D. George, M. C. Selby & R. Scolnik. A Three-Beam Oscillograph for Recording at Frequencies up to 10,000 Megacycles.—G. M. Lee (see 1692). Microwave Power Measurement.—T. Moreno & O. C. Lundstrom. A New High-Speed Recording Potentiometer.—V. L. Parsegian. Electronic Frequency Stabilization of Microwave Oscillators.—R. V. Pound. Metallized-Glass Attenuators for Radio-Frequency Applications.—E. Weber. The New "Speedomax" Power-Level Recorder.—A. J. Williams, Jr. & W. R. Clark. For other abstracts see *Electronics*, March 1946, Vol. 19, No. 3, pp. 92..108, and *Electronic Industr.*, March 1946, Vol. 5, No. 3, pp. 62..74. For titles of other papers read, see other sections.

621.317.31.082.5 : 621.362 **1577**

Electric Precision Measurements by Means of Incandescent Lamp Used as Indicator of Current Equality in Two Circuits.—A. I. Fürstenberg. (*C. R. Acad. Sci. U.R.S.S.*, 1945, Vol. 48, No. 1, pp. 23–26. In English.) "In an incandescent lamp with a metal filament the appearance of a barely visible light emission is a very fine criterion for visual estimation of the heating current. A circumstance of interest for metrology is that under conditions of a vanishingly small light emission by the filament, slight increments in the current passing through the lamp have a considerable effect upon brightness." Experiment shows that in the neighbourhood of this threshold light emission, an increase of $\pm$ 0.5% in current results in $\pm$ 50% increase in brightness. The possible use of the method for determining the equality of the currents in circuits is considered, with a brief discussion of the errors involved in using it for comparing inductors and capacitors. Errors are limited to a few parts in ten thousand.

621.317.33 : 621.317.755 **1578**

Impedance Measurements with the Cathode-Ray Oscilloscope.—W. Vissers, Jr. (*Radio, N.Y.*, Jan. 1946, Vol. 30, No. 1, pp. 23..62.) A source of voltage of the required frequency is connected to the unknown impedance in series with a known impedance. The voltage across the unknown is applied to the X-deflexion plates of a cathode-ray oscilloscope, and the total voltage is applied to the Y-plates. Measurements of the resulting ellipse enable the magnitude and phase of the unknown impedance to be found.

621.317.334/.335 **1579**

Simple Capacitance and Inductance Measurements.—T. A. Gadwa. (*QST*, March 1946, Vol. 30, No. 3, pp. 71..136.) The frequency-variation method is used.

621.317.361 + 531.76 **1580**

WWV Schedules.—(*QST*, Feb. 1946, Vol. 30, No. 2, p. 41.) Details of the standard-frequency transmissions by the National Bureau of Standards (station WWV) at 2.5, 5, 10 and 15 Mc/s. The accuracy of the carrier and modulation frequencies and of the time signals is better than 1 part in 10^7.

621.317.361 : 621.318.572 **1581**

Decade Counting Circuits.—V. H. Regener. (*Phys. Rev.*, 1st/15th Jan. 1946, Vol. 69, Nos. 1/2, p. 46.) Ten miniature pentodes (6AK6) in a ring with 30 resistors and 10 capacitors, or ten twin-pentodes (12L8–GT) with 20 resistors and no capacitors form counters suitable for frequency measurement up to 100 kc/s with cathode-ray indication. Summary of Amer. Phys. Soc. paper.

621.317.361 + 621.396.611.21 **1582**

Duplex Crystals.—C. E. Lane. (*Bell Lab. Rec.*, Feb. 1946, Vol. 24, No. 2, pp. 59–62.) Two quartz crystal plates bonded together enable the lowest natural flexure frequency of a free bar to be excited and give resonance frequencies in the range 1 kc/s –10 kc/s. Methods of bonding and of inducing the required vibrations in the crystals are given, and their uses as oscillators and filters are described in general terms.

621.317.4 **1583**

A New Magnetomotive Force Gauge and Magnetic Field Indicator.—W. B. Ellwood. (*Rev. sci. Instrum.*, March 1946, Vol. 17, No. 3, pp. 109–111.) A compact instrument in which a magnetically operated reed switch is closed by the field being measured. Current through an external winding re-opens the switch, and the value of current required to do so is a measure of the field.

621.317.7 : 621.316.93 1584

An Electronic Bypass for Measuring Purposes.—L. A. Finzi. (*Electronics*, Feb. 1946, Vol. 19, No. 2, pp. 196..202.) A simple gas-tube circuit "for protection of . . . measuring instruments [against transient overloads] . . . where it is necessary to restore the normal instrument operation as soon as the current falls back within the limits of the instrument range."

621.317.714 + 621.317.725].089.6 1585

Production Testing of [d.c.] Panel Meters.—R. Ammon. (*Electronics*, Feb. 1946, Vol. 19, No. 2, pp. 170..178.) A 6N7 is used with variable grid bias to provide a smoothly variable rheostat; with a d.c. voltage source the rheostat is used to adjust the voltage across a standard 100-V voltmeter connected in parallel with the meter under test, which is itself arranged as a 100-V voltmeter by the use of a suitable series resistor.

621.317.714 1586

Multi-Range Milliammeter.—R. P. Turner. (*Radio News*, Feb. 1946, Vol. 35, No. 2, pp. 49..142.) Conversion of a commercial 0–1-mA meter to read 0–1 A d.c. in four decade ranges. The circuit is arranged so that standard-value resistors can be used.

621.317.725 1587

An Integrating Meter for Measurement of Fluctuating Voltages.—H. E. Haynes. (*J. Soc. Mot. Pict. Engrs*, Feb. 1946, Vol. 46, No. 2, pp. 128–133.) The meter integrates the voltage over a chosen interval, *e.g.* 0.5–5.0 sec, and indicates the average value that has existed. A combined voltage-amplifier and phase-inverter is followed by a full-wave rectifier and capacitor-charging circuit. Timing is accomplished by a resistance-capacitance circuit and a thyratron relay. The frequency response is flat from 50 c/s to 15 kc/s, and the instrument can withstand considerable overload.

621.317.725 : 621.314.632 1588

A.C. Voltage Measurements.—O. E. Carlson. (*Radio Craft*, Jan. 1946, Vol. 17, No. 4, pp. 250..275.) Summary of the properties of copper-oxide rectifiers in relation to their use in a.c. voltmeters.

621.317.725 : 621.385.2 1589

A Stable Diode Voltmeter.—Furzehill Laboratories, Ltd. (*Electronic Engng*, March 1946, Vol. 18, No. 217, p. 94.) A television-type diode feeding a triode d.c. amplifier in a bridge circuit with a microammeter for peak voltage measurements. The arrangement of the circuit to give a stable zero is described.

621.317.725 : 621.385.3 1590

High-Resistance D-C Voltmeter.—D. L. Waidelich. (*Electronics*, March 1946, Vol. 19, No. 3, pp. 158–160.) The design of a reflex-type valve voltmeter that has maximum errors of 8% and 5% when measuring voltages of sources having internal resistances of 1 000 MΩ and 100 MΩ respectively. Over-voltage protection is provided, and no electrical zero adjustment is required. Voltage ranges 0–50, 100, 250 V.

621.317.725.029.3/.62] : 621.385.2 1591

R.F. Probe Design.—D. F. McAvoy. (*Radio News*, March 1946, Vol. 35, No. 3, pp. 35..135.) Constructional details of an r.f. diode probe used as an accessory for a d.c. valve voltmeter. A suitable circuit is described.

621.317.73.029.54/.58 1592

A Radio-Frequency Capacitance and Conductance Bridge.—R. F. Proctor & E. G. James. (*J. Instn elect. Engrs*, Part I, Feb. 1946, Vol. 93, No. 62, p. 103.) Summary of 676 of March.

621.317.733 1593

Note on the Helmholtz Make-and-Break Theorem and an Application to the Wheatstone Net.—G. F. Freeman. (*Phil. Mag.*, Aug. 1945, Vol. 36, No. 259, pp. 541–546.) In bridge networks it is often advantageous to consider a change in impedance as the independent variable, and to use the galvanometer as a deflexional instrument. Examples relating to the Wheatstone network are given.

621.317.761.029.63 1594

V.H.F. Heterodyne Frequency Meter.—A. A. Goldberg. (*Radio News*, March 1946, Vol. 35, No. 3, pp. 32..128.) A split-concentric-tuned oscillator, using an acorn valve, covers the range 280–430 Mc/s. Harmonic operation is possible up to 3 000 Mc/s, and the oscillator may be pulsed or square-wave modulated.

621.317.763.029.63/.64 1595

Types and Applications of Microwave Frequency Meters.—W. J. Jones. (*Radio, N.Y.*, Jan. 1946, Vol. 30, No. 1, pp. 29–34.) Three instruments commonly used are the coaxial-line, cylindrical-cavity, and transition types, the latter being a combination of the other two. The principles and methods of operation are described, and their accuracies briefly discussed.

621.317.763.029.63/.64 1596

A Resonant-Cavity Wavemeter.—J. McQuay. (*Radio News*, Feb. 1946, Vol. 35, No. 2, pp. 36..78.) Description of a waveguide fitted with an adjustable plunger and a neon indicator.

621.317.79 : [621.315.212.1.029.62/.63 1597

Measuring Coaxials at Ultra-High Frequencies.—C. C. Fleming. (*Bell Lab. Rec.*, Jan. 1946, Vol. 24, No. 1, pp. 2–5.) Methods of measuring cable attenuation by insertion loss are briefly described. Measurements at frequencies from 120 to 420 Mc/s can be made directly and can be extended to 3 000 Mc/s by a heterodyne system. Characteristic impedance is obtained at 100 Mc/s by a resonance method using a Q meter and standard calibrating resistors.

621.317.79 : 621.385 1598

A Method of Measuring Grid Primary Emission in Thermionic Valves.—A. H. Hooke. (*Electronic Engng*, March 1946, Vol. 18, No. 217, pp. 75–80.) In most valves the grid becomes contaminated and raised in temperature due to proximity to the cathode. The resulting primary emission is of importance, though not readily measurable because of the grid current due to gas and to secondary emission. The test circuit described supplies periodic voltages consecutively to the grid and anode, followed by the application of a high negative potential to the grid, during which time the grid primary emission is measured. The cycle is repeated at 60-millisecond intervals. The details of the circuit and its operation are given, and the interpretation of the meter readings discussed. Typical curves showing the grid emission as a function of grid dissipation, anode power and grid material are given. Two appendices give theoretical analyses of (i) the conversion of the measured average current and voltage to mean power for a linear and a 3/2-power law and (ii) the relation between the anode and grid dissipations for a given grid primary emission.

621.317.79 : 621.396.619.16 **1599**
Measuring Pulse Characteristics.—A. Easton. (*Electronics*, Feb. 1946, Vol. 19, No. 2, pp. 150–154.) Outline description of methods and apparatus required for measuring pulse spacing, pulse length, time of rise and time of decay. Long bibliography.

621.317.79 : 621.396.82 **1600**
Effectiveness of Conduit as R-F Shielding.—S. L. Shive. (*Electronics*, Feb. 1946, Vol. 19, No. 2, pp. 160..166.) Equipment for measuring the screening properties of conduits in the frequency range 0.1–150 Mc/s is described, with typical results. A screening box contains a transmitting coil at the end of a polystyrene rod, over which the conduit is slipped; a receiving loop mounted outside the conduit feeds a receiver external to the box. A standard-signal generator energizes the transmitting loop: the figure of merit for the conduit is the ratio of voltages (db) applied to the transmitter with and without the conduit, for a constant output from the receiving loop.

621.317.791 **1601**
Universal Test Instrument.—M. Silver. (*Radio News*, Feb. 1946, Vol. 35, No. 2, pp. 32..118.) A detailed description of a d.c./a.c. instrument with 51 ranges for voltage, current, resistance, and attenuation measurement. The voltmeter can be used at frequencies up to 100 Mc/s.

621.317.794 : 539.232 **1602**
Production and Properties of Nickel Bolometers.—Brockman. (*See* 1556.)

621.395.92 **1603**
The Testing of Deaf Aids.—Turney. (*See* 1441.)

621.396.615 : 621.317.79 **1604**
Wide-Range Electronic Sweeper.—A. D. Smith, Jr. (*Communications*, Jan. 1946, Vol. 26, No. 1, pp. 24..31.) Detailed description of an h.f. wobbulator for aligning television receivers. Range of mean frequency 500 kc/s–110 Mc/s, extensible to 220 Mc/s by harmonic operation, frequency sweep adjustable up to 10 Mc/s. Output independent of frequency to ± 10%. The signal is derived by a heterodyne process using a reactance tube to produce the sweep. 1-Mc/s and 10-Mc/s crystals are included to provide frequency markers.

621.396.615.12 **1605**
Signal Generator Covers All Bands.—B. White. (*Radio Craft*, Jan. 1946, Vol. 17, No. 4, pp. 243–244.) Constructional details of a mains-operated generator with continuous coverage from 65 kc/s–34 Mc/s, with plug-in coils. Internal modulator range 24–20 000 c/s.

621.396.615.12 : 621.317.79 **1606**
[TF.867] **Signal Generator.**—L. Hauser. (*J. sci. Instrum.*, March 1946, Vol. 23, No. 3, p. 63.) Letter correcting a statement on provisional operating data given in 1289 of May.

621.317.7 : 621.396 **1607**
Radio Test Instruments. [Book Review]—R. B. Turner. Ziff-Davis, New York, 1945, 228 pp., $4.50. (*Electronic Industr.*, March 1946, Vol. 5, No. 3, p. 154.) "Of particular interest to amateur radio operators and experimenters, this book describes how to construct the great majority of instruments commonly used . . ."

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

538.3 **1608**
A Study of Stationary Electromagnetic Modes for Region Between Parallel Perfectly Conducting Planes and Application to Electron Accelerator.—Akeley. (*See* 1516.)

578.088.7 : 621.396.615.17 **1609**
Nerve Stimulator.—W. I. Weiss. (*Electronics*, Feb. 1946, Vol. 19, No. 2, p. 155.) "Thyratron-type relaxation oscillator with plate-supply keying [foot-operated, or panel switch] eliminates high initial pulse, provides 0.3 c/s–30 kc/s pulses for biological research and medical therapy." Circuit diagrams are given.

615.84 : 621.3.029.62 **1610**
Shortwave Diathermy.—J. M. Oxley. (*Radio Craft*, Jan. 1946, Vol. 17, No. 4, pp. 245..281.) Brief discussion of the biological basis of electro-therapeutic effects, with an outline of the principal methods used.

621.317.39.083.7 : 629.13 **1611**
Fuel Consumption Indicator.—D. W. Moore, Jr. (*Electronics*, March 1946, Vol. 19, No. 3, pp. 152–153.) "Electronic servo transmits the position of an aircraft fuel-flow valve to a remote indicator. A mechanical integrator combines this information with impulses from a 100 c/s time standard to totalize the liquid flow. The device is applicable to industrial telemetering problems."

621.317.39.087.4 + 621.3.078 **1612**
Instruments for the Automatic Controlling and Recording of Chemical and other Processes.—Prinz. (*See* 1691.)

621.36/.39 **1613**
I.R.E. Winter Technical Meeting January 1946.—(*Waves and Electrons*, Feb. 1946, Vol. 1, No. 2, pp. 80..93.) Abstracts are given of the following papers read at the meeting. Electronics in Naval Warfare.—R. Bennett. Navy Radio and Electronics in World War II.—J. B. Dow. The Role of Electronics in Antiaircraft Gun-Fire Control.—F. B. MacLaren. A New System of Radio Telemetering.—D. W. Moore, Jr. & F. G. Willey. Duplex Operation of Independent High-Power Oscillators for Induction Heating.—W. C. Rudd. A Frequency-Modulation Altimeter for Meter and Light Indication and the Automatic Altitude Control of Aircraft.—R. C. Sanders, Jr., W. R. Mercer, I. Wolff & J. C. Smith (Title only). One-Millionth-of-a-Second Radiography and its Applications.—C. M. Slack & D. C. Dickson. For other abstracts see *Electronics*, March 1946, Vol. 19, No. 3, pp. 92..108, and *Electronic Industr.*, March 1946, Vol. 5, No. 3, pp. 62..74. For titles of other papers read, see other sections.

621.365.52 **1614**

Induction Heating of Hollow Metallic Cylinders.—A. Gemant. (*J. appl. Phys.*, March 1946, Vol. 17, No. 3, pp. 195–200.) "The rigorous expression for the heat input, valid for the whole frequency range, is developed . . . It is shown in an example how the equation is used for numerical computations. Existing approximate formulas for the low frequency and the high frequency ranges are checked by means of the rigorous equation; the maximum deviation between the rigorous and approximate equations is about 10 per cent."

621.365.52 : 621.385.1 **1615**

Induction Heating in Radio Electron-Tube Manufacture.—E. E. Spitzer. (*Proc. Inst. Radio Engrs, N.Y.*, Part II, March 1946, Vol. 34, No. 3, pp. 110–115.) The following applications are considered:—degassing, getter-flashing, vacuum-firing, metal-to-glass sealing, brazing, and welding. Frequencies in the range 200–500 kc/s and powers of 2–15 kW are usually employed. The theory of this method of heating metallic elements is considered.

621.365.52 : 621.791.3 **1616**

R-F Soldering of Metal-to-Glass Seals.—R. A. Ammon. (*Electronics*, March 1946, Vol. 19, No. 3, pp. 120–121.) 30-Mc/s heating generators, distributed throughout the plant and fed from a central power supply, are used in the production of sealed meters. Increased efficiency over ordinary techiques, such as the use of soldering irons, is obained in soldering edge-metallized glass windows and glass-insulated feed-through terminals to the metal case of the instrument, in mounting polepieces on instrument magnets, and in soldering metallized glass jewel bearings in position. The generator circuit is given.

21.365.92 : 678.028 **1617**

High-Frequency Heating Developments.—(*Elecronics*, March 1946, Vol. 19, No. 3, pp. 170..178.) Features of a 125-kW 13.6-Mc/s unit developed by Westinghouse for curing and drying a new spongeubber product. See also 698 of March.

21.365.92 : 678.028 **1618**

Electronic Rubber Preheater.—E. Mittelmann & . P. Bosomworth. (*Electronics*, March 1946, Vol. 19, No. 3, pp. 128–130.) A brief general escription of the use of a 20-kW 27-Mc/s heating nit for preheating rubber parts, in readiness for uring.

21.38 : 62 **1619**

Economics of Electronics.—J. Markus. (*Sci. mer.*, Dec. 1945, Vol. 173, No. 6, pp. 349–351.) discussion on recent electronic developments, cluding references to ingenious but uneconomical evices.

1.383 **1620**

Sensory Aid for the Blind.—L. Cranberg. (*Elecnics*, March 1946, Vol. 19, No. 3, pp. 116–119.) detailed account of the device noted in 1309 of ay.

1.385.833 **1621**

Filmless Sample Mounting for the Electron croscope [Airborne particles supported on thin ss fibres].—J. H. L. Watson. (*J. appl. Phys.*, b. 1946, Vol. 17, No. 2, pp. 121–127.)

621.386.1 : 620.179 **1622**

Production Control with 2,000,000-volt X-rays.—D. Goodman. (*Electronics*, Feb. 1946, Vol. 19, No. 2, pp. 146–149.) "Details of super-voltage installation using conventional X-ray tube construction and resonant transformer, and description of continuous industrial radiographic setup used during war for inspection of powder charges in large loaded shells and bombs."

621.386.1 : 621.396 **1623**

Notes on Radiography in the Wireless Industry.—R. M. Mitchell. (*Marconi Rev.*, Jan./March 1946, Vol. 9, No. 80, pp. 13–20.) A simple introduction to the subject is followed by a description of modern industrial X-ray plant. After a survey of the factors governing the making of satisfactory radiographs, practical applications, chiefly relating to physical examination of valves, are illustrated.

621.43 : 621.317.39 **1624**

Pressure-Time Curves in Electronic Observation of Engines.—W. F. Brown. (*Electronics*, Feb. 1946, Vol. 19, No. 2, pp. 168, 170.) Comprises a 100-kc/s bridge in which one of the arms includes a diaphragm-operated capacitor which is subjected to the pressure changes. An amplifier and detector system associated with the bridge enables either the full 100-kc/s output, the output with negative half-waves suppressed, or the envelope, together with timing marks, to be displayed on a synchronized oscillograph.

621.9 : 621.38 **1625**

Contouring Control for Machine Tools.—(*Electronics*, Feb. 1946, Vol. 19, No. 2, p. 178.) A stylus guides itself round a template at constant speed, or is drawn round a diagram like a pencil, and its motion controls the cutting operation through electronic mechanisms. "The new control . . . is capable of a variety of intricate cutting operations. . . ."

623.26 : 621.396.9 **1626**

Land Mine Locators.—S. S. West. (*Electronic Engng*, March 1946, Vol. 18, No. 217, pp. 69–74.) Models I, II and III are based on the Felici bridge in which equal and opposite mutual inductances give zero output. Two overlapping circular coils are adjusted to have zero coupling so that, when brought near a metallic object, an unbalance due to external coupling is produced. This unbalance is indicated by feeding one coil from an a.f. oscillator and amplifying the output from the other to headphones. A balancing circuit enables the initial stray reactive and resistive coupling to be annulled. A detailed description of the construction and of the circuit is given. A novel and improved locator (Model IV) gives discriminatory detection. The coils are connected respectively to the input and output of a 3-stage amplifier having negligible phase shift; the coupling between the coils provided by a nearby object causes a.f. self-oscillation. By introducing a controllable phasing network, the detector can be made to have greatest sensitivity to a ferrous object, which produces a 90° phase shift.

623.454.25 : 621.396.9 **1627**

Proximity Fuzes for Artillery.—H. Selvidge. (*Electronics*, Feb. 1946, Vol. 19, No. 2, pp. 104–109.) A description of the problems encountered, and the methods used in overcoming them. General details of a proximity fuse suitable for shells of 75 mm calibre and upwards are given. A satisfactory fuse must withstand the high accelerations experienced when the shell is in the barrel (although it is not then operative), and the high spin accelerations throughout its flight; its battery must not deteriorate when stored under adverse conditions; its electrical characteristics must be reproducible, and the polar diagram of the aerial should "roughly match the fragmentation pattern of the projectile". The fuse described comprises a self-quenched superregenerative oscillator which acts as transmitter and also as detector for the receiver, the further stages of the latter comprising two RC-coupled pentodes feeding the grid of a thyratron. The latter is normally in the quiescent state, but when an audio-frequency signal of appropriate amplitude is applied to its grid (due to the interaction at the detector of the transmitted signal and that received back from the target) the thyratron fires and sets off the detonating cap. The battery is of the "reserve" type, the electrolyte being kept out of contact with the plates until the instant of firing. The mechanical construction of the fuse is discussed, together with the design of the rugged valves. Methods used in testing the valves and fuses under production conditions are indicated. For a description of a fuse suitable for low-acceleration projectiles, see 624 of March (Huntoon & Miller).

629.13 : 621.38 **1628**

Future of Electronics in Aviation.—J. D. Goodell & D. J. Coleman. (*Radio News*, Feb. 1946, Vol. 35, No. 2, pp. 25..135.) Some applications of wartime developments, *e.g.* radar, control of cabin temperature, autopilots, etc., are discussed. The principle of the ceilometer, a photoelectric device for measuring cloud height, is described.

629.13 : 621.398 **1629**

Radio Target Planes.—(*Radio Craft*, Jan. 1946, Vol. 17, No. 4, pp. 242..279.) A short general account. See also 1323 of May.

639.2 : 621.383 **1630**

Photoelectric Fish Counter.—L. V. Whitney & A. D. Hasler. (*Electronics*, March 1946, Vol. 19, No. 3, pp. 178..182.) The beam from a light source falling on a photocell is interrupted by fish swimming across the beam. The interruptions are observed on a microammeter in a simple amplifier connected to the photocell. The size of catch correlates with the rate of count.

639.3.06 : 621.38 **1631**

Electronic Control of Fish Fence.—(*Electronics*, March 1946, Vol. 19, No. 3, p. 164.) Rapid pulses of special waveform used with an electric fence turn back both small and large fish without harming them.

771.36 : 621.317.39 **1632**

Electronic [camera-] **Shutter-Testers.**—R. F. Redemske. (*Electronics*, Feb. 1946, Vol. 19, No. 2, pp. 128–134.) The first device is a recorder which plots aperture area as a function of time for an iris shutter, and the screen speed at the beginning, middle and end of the traverse for a focal-plane shutter. The delay in functioning of light-operated shutters may also be measured. Teledeltos paper is used for recording, and in the case of the iris shutter, for example, the record consists of ten straight lines drawn parallel to each other and to the direction of uniform motion of the recording paper. Each line corresponds to a definite shutter area, and the length of the line represents the time for which the particular shutter area is exceeded. The curve enclosing the extremities of the lines gives aperture area as a function of time. The result is achieved by a photoelectric cell, voltage dividers and associated amplifiers, the design of which is discussed. The second device gives on a meter the percentage departure of actual shutter speed from rated speed. Here, the photoelectric current is arranged, through suitable amplifiers, to charge a capacitor through a series pentode. The voltage so developed is used as a measure of the time for which the shutter is open.

621.385.833 : 54 **1633**

Major Instruments of Science and Their Applications to Chemistry. [Book Review]—R. E. Burk & O. Grummitt (Eds.). Interscience Publishers Inc., New York, 1945, 151 pp., $3.50. (*Electronic Industr.*, March 1946, Vol. 5, No. 3, p. 153.) Vol. IV in the series "Frontiers in Chemistry". Contains "Electron Diffraction" by H. Germer, "The Electron Microscope" by L. Marton, "X-Ray Diffraction" by M. L. Huggins, and articles on spectroscopy by W. R. Brode and R. R. Barnes.

PROPAGATION OF WAVES

517.947.4 **1634**

A Method of Solution of Field Problems by means of Overlapping Regions.—Poritsky & Blewett. (*See* 1571.)

621.396[.11 + .81 **1635**

Forecasting Long-Distance Transmission.—W. R. Foley. (*QST*, Feb. 1946, Vol. 30, No. 2, pp. 36–41.) Description of a method using maximum-usable-frequency charts for the determination of optimum frequencies and times for long-distance transmission.

621.396.11 **1636**

I.R.E. Winter Technical Meeting January 1946.—(*Waves and Electrons*, Feb. 1946, Vol. 1, No. 2, pp. 80..93.) Abstracts are given of the following papers read at the meeting. Field Intensities Beyond Line of Sight at 45.5 and 91 Megacycles.—C. W. Carnahan, N. W. Aram & E. F. Classen. Development in Radio Sky-Wave Propagation Research and Applications during the War.—J. H Dellinger & N. Smith. The Role of Atmospheric Ducts in the Propagation of Short Radio Waves.—J. E. Freehafer. Three- and Nine-Centimetre Propagation Measurements in Low-Level Ocean Ducts.—M. Katzin & R. W. Bauchman. Microwave Propagation: Part I. The Effect of Rain Upon the Propagation of Waves in the One- and Three-Centimetre Region.—S. D. Robertson & A. P. King. Microwave Propagation: Part II. Propagation of Six-Millimetre Waves.—G. E. Mueller. Measurement of the Angle of Arrival of Microwaves.—W. M. Sharpless. For other abstracts see *Electronics*, March 1946, Vol. 19, No. 3, pp. 92..108, and *Electronic Industr.*, March 1946, Vol. 5, No. 3, pp. 62..74. For titles of other papers read, see other sections.

621.396.11 **1637**

Curved Earth Geometrical Optics.—G. Millington. (*Marconi Rev.*, Jan./March 1946, Vol. 9, No. 80, pp. 1–12.) "In this paper the effect of the earth's curvature in the geometric-optical treatment of propagation within the visual range is presented as a correction to the flat-earth geometry. Factors to be applied to the flat-earth values of the angle of elevation at the point of reflection and of the path difference between the direct and reflected waves are derived by a single graphical process, and also the divergence factor arising from the increased divergence on reflection from the convex surface of the earth is obtained. Asymptotic values are given for points very near to the horizon."

621.396.11 **1638**

On a Modification of the Interference Method of Investigating the Propagation of Radio Waves.—L. I. Mandelstam & N. D. Papalexi. (*C. R. Acad. Sci. U.R.S.S.*, 1940, Vol. 26, No. 8, pp. 775–779. In English.) The interference method referred to is the measurement of the phase difference between emitted waves and those returned to the sender from a distant "reflector" or re-radiator which transforms the frequency in a simple integral or fraction ratio. The distance of the re-radiator can be found in terms of an unknown whole number of wavelengths and a phase-constant of the apparatus. These unknowns must be determined separately, *e.g.* by a small change of wavelength. It is pointed out that the method is capable of greater accuracy if applied to the determination of small changes of distance, the wavelength being constant, *e.g.* with the re-radiating station on a boat or car. The method was used to determine the mean velocity of radio waves over water (Black Sea, Lake Ilmen, and White Sea) and over land, the distance changes being independently measured by ordinary surveying methods. In the Black Sea measurements the velocity was found to be $(2.999 \pm 0.005)10^{10}$ cm/sec. The land experiments (130–195 m) were carried out to determine the phase structure of the field both in the immediate neighbourhood of the transmitter and at gradually increasing distance from it. Theoretical expectations were confirmed. There was a sharp distortion of the phase structure near a deep ravine. This work is to be described in a separate publication.

621.396.11 **1639**

Influence of the Earth's Surface upon Phase Structure of the E.M. Field of a Radiating Aerial.—L. Alpert & V. V. Migulin. (*C. R. Acad. Sci. U.R.S.S.*, 1940, Vol. 26, No. 9, pp. 881–884. In English.) An experimental study of the upward extent of the influence of the earth's surface on the propagation of radio waves. A transmitter on the surface radiated two coherent oscillations with frequencies in the ratio 3/2, (wavelengths 120–180 m and 300–450 m) and observations of the phase difference $\Delta\phi = \phi_1 - 3\phi_2/2$ were made on receivers carried by captive balloons. It was found that the effect of the earth's surface, as shown by the magnitude of $\Delta\phi$, gradually diminishes with increase of height above the earth's surface and practically ceases at a height of about 4λ. The distances at which this effect was observed were about $25–50\lambda$ on the shorter wavelengths and $–20\lambda$ on the longer, and the total variation in $\Delta\phi$ was about 30° on the shorter wavelengths and 18° on the longer.

621.396.11 **1640**

Irregularities in Radio Transmission: Part 2.—P. Ferrell. (*Radio, N.Y.*, Jan. 1946, Vol. 30, No. 1, pp. 25..63.) Description of several instances of anomalous radar reflections, from clouds, hot air, birds, and auroral ionized regions. For part 1 see 1333 of May.

621.396.11 : 551.51.053.5 **1641**

On the Connection between the Anomalies of Polarization of Half-light and the State of Ionization.—V. M. Bovcheverov, A. V. Mironov, I. M. Mikhailine, V. M. Morozov, Z. L. Ponizovsky, S. P. Sokolov & I. A. Khvostikov. (*C. R. Acad. Sci. U.R.S.S.*, 1940, Vol. 26, No. 9, pp. 900–903. In French.) According to Rayleigh's theory of molecular scattering, the degree of polarization of half-light (dawn and twilight) should decrease uniformly with ϕ the zenithal distance of the sun, according to $p = (1 - \cos^2\phi)/(1 + \cos^2\phi)$, but in fact the polarization shows a well marked minimum. This may be attributable to effects of the ionized layers. The experiments described (simultaneous observations of polarization and of critical frequency of pulsed radio transmissions) were carried out to find whether there was any correlation between ionization and the degree of anomalous polarization. A comparison between the observed decreases of polarization from the theoretical values and the corresponding critical frequencies show that there is such a correlation, *i.e.* the maximum decrease is associated with the highest critical frequencies.

621.396.11.029.62 **1642**

Need There be Line-of-Sight?—E. P. Tilton. (*QST*, March 1946, Vol. 30, No. 3, pp. 47–50.) A line-of-sight path is not always necessary for good communication in the 144-Mc/s band. Aerial systems are described which have led to improved reception under conditions where the transmitter has been obscured by hills.

631.437 : 621.3.011.2 **1643**

The Use of Cumulative Resistance in Earth-Resistivity Surveys.—Ruedy. (*See* 1523.)

RECEPTION

621.395.8 : 621.395.645 **1644**

Audio Distortion in Radio Reception.—J. Minter. (*Proc. Radio Cl. Amer.*, Jan. 1946, Vol. 23, No. 1, pp. 1–5.) Discussion of transient response and cross-modulation distortion. Cross-modulation between high and low audio frequencies can be reduced by suitable transformer design.

621.396.611.1 : 621.396.621.54 **1645**

Practical Radio Course: Part 41. Analysis of the various causes of drift in the resonance frequency of tuned circuits and in the frequency of the superheterodyne oscillator.—A. A. Ghirardi. (*Radio News*, Feb. 1946, Vol. 35, No. 2, pp. 50..98.) Methods of reducing the effects of temperature, humidity and age, on coils and capacitors. For part 40 of the series see 1342 of May.

621.396.611.1 : 621.396.621.54 **1646**

Practical Radio Course: Part 42. Covering various easily applied methods of minimizing and compensating for oscillator frequency drift in superheterodyne receivers.—A. A. Ghirardi. (*Radio News*, March 1946, Vol. 35, No. 3, pp. 46..100.) For previous parts see 1645 above.

621.396.619.018.41 : 621.396.622 **1647**

Frequency Discriminator.—Bruck. (*See* 1491.)

621.396.62 + 621.396.9 **1648**
Germany's UHF Tubes.—Combined Intelligence Sub-Committee. (*Electronic Industr.*, March 1946, Vol. 5, No. 3, pp. 162..168.) A continuation of 1399 of May, dealing very briefly with "radar camouflage" (materials of low radio-reflecting power), and u.h.f. receivers of various types.

621.396.62 **1649**
I.R.E. Winter Technical Meeting January 1946.—(*Waves and Electrons*, Feb. 1946, Vol. 1, No. 2, pp. 20..93.) Abstracts are given of the following papers read at the meeting. Design of Communication Receivers for the Naval Service with Particular Consideration to the Very-High-Frequency and Ultra-High-Frequency Ranges.—T. McL. Davis. Microwave-Converters.—C. F. Edwards. Tunable Receivers for Very High Frequencies.—G. E. Hulstede, J. M. Pettit, H. E. Overacker, K. Spangenberg & R. R. Buss. Ultra-High-Frequency Television Receivers. — H. T. Lyman. Noise Spectrum of Crystal Mixers.—P. H. Miller. Discriminators for Frequency-Modulation Receivers.—S. W. Seeley. Theory of Impulse Noise in Ideal Frequency-Modulation Receivers.—D. B. Smith & W. E. Bradley. Crystal Rectifiers in Heterodyne Receivers.—H. C. Torrey. For other abstracts, see *Electronics*, March 1946, Vol. 19, No. 3, pp. 92..108, and *Electronic Industr.*, March 1946, Vol. 5, No. 3, pp. 62..74. For titles of other papers read, see other sections.

621.396.62 **1650**
CAA Alaskan Diversity Receiving System: Part 1.—Ivers. (*See* 1669.)

621.396.62 **1651**
New Svenskradio Wireless Receivers.—C. Fredin. (*Ericsson Rev.*, 1945, Vol. 22, No. 4, pp. 187–193.) Includes illustrations and a comparative table of details of types 1453, 1454, 1457 and 1458 V(LV).

621.396.62 **1652**
Wide-Range Converter-Receiver.—B. E. Hargrove. (*Radio News*, Feb. 1946, Vol. 35, No. 2, pp. 40–42, 140.) Constructional details of a superregenerative receiver for 112 Mc/s and a 14, 21, 28, 50 Mc/s converter for use with a lower-frequency receiver. High signal/noise ratio, good stability, and freedom from image troubles are claimed.

621.396.62 : 621.317.755 **1653**
Panoramic Reception, 1946.—J. R. Popkin-Clurman & B. Schlessel. (*QST*, March 1946, Vol. 30, No. 3, pp. 22–27.) Description of a receiver adaptor for displaying signals received within a wide frequency band, using a cathode-ray tube with X-deflexion representing signal frequency and Y-deflexion representing signal strength. Pre-war development and wartime applications are reviewed.

621.396.621 : 621.396.619.018.41 **1654**
Frequency Modulation Receiver.—M. Ziegler. (*Radio, N.Y.*, Jan. 1946, Vol. 30, No. 1, p. 41.) The receiver incorporates a feedback arrangement to allow a narrow-band i.f. amplifier to be used, and a frequency-counter circuit is used instead of the conventional discriminator. Summary of U.S. Patent 2 383 359.

621.396.621.029.62 **1655**
144-Mc Radio.—I. Queen. (*Radio Craft*, Jan. 1946, Vol. 17, No. 4, pp. 239..289.) Constructional details of a superregenerative receiver.

621.396.621.029.62 **1656**
A Non-Radiating Superregenerative Receiver for Two Meters.—E. P. Tilton. (*QST*, Feb. 1946, Vol. 30, No. 2, pp. 53..108.) The circuit comprises a tuned r.f. stage, superregenerative detector, and two a.f. stages, all using miniature valves. The receiver is battery-operated, with low current consumption.

621.396.621.029.62 **1657**
Service Considerations in Megacycle Bands.—C. J. Sheridan. (*Radio News*, March 1946, Vol. 35, No. 3, pp. 38..96.) An outline of v.h.f. and f.m. reception techniques.

621.396.621.53 **1658**
Calculation of the Output from Non-Linear Mixers.—Stockman. (*See* 1492.)

621.396.621.54 **1659**
Single Signal C.W. Reception and Crystal Filters.—B.G. (*QST*, March 1946, Vol. 30, No. 3, pp. 59–61.) If the beat-oscillator frequency is set to one side of a narrow i.f. response curve, the heterodyne whistle on one side of the zero-beat position is suppressed.

621.396.621.54 **1660**
Oscillatorless Superheterodyne.—R. W. Woods. (*Electronics*, Feb. 1946, Vol. 19, No. 2, pp. 224..234.) Comprises two mixers, each fed with the signal. The load of one is tuned to intermediate frequency, and is connected to a second input terminal of the other mixer; the load circuit of the latter is tuned to (signal ± intermediate) frequency and is connected to a second input terminal of the first mixer. Output at intermediate frequency is taken from the load circuit of the first mixer. An analogous arrangement may be used as a beat-frequency "oscillator" to follow the i.f. stages of a c.w. morse receiver. Theory and properties of these arrangements are given, with suggested additional applications.

621.396.621.54.029.58 **1661**
A 28-Mc. Receiver/Converter.—B. Goodman. (*QST*, Feb. 1946, Vol. 30, No. 2, pp. 17–20, 106.) Constructional details of an oscillator and frequency converter giving an i.f. of about 1.5 Mc/s, intended to precede an ordinary broadcast receiver.

621.396.622.7 **1662**
Superregenerative Detector Selectivity.—A. Easton. (*Electronics*, March 1946, Vol. 19, No. 3, pp.154–157.) "Experimental determination of characteristics . . . indicates that selectivity and sensitivity increase and noise decreases with decreasing quench frequency. Selectivity decreases with increasing quench amplitude."

621.396.622.7 **1663**
The Application of Modulation-Frequency Feedback to Signal Detectors.—G. Builder. (*Proc. Inst. Radio Engrs, N.Y.*, Part I, March 1946, Vol. 34, No. 3, pp. 130–137.) "Positive modulation-frequency feedback to the load circuit of a signal detector may be adjusted to make the effective modulation-frequency admittance of the load circuit equal to its direct-current conductance, thus eliminating peak clipping and improving the efficiency of detection. Conversely, negative feedback to the detector tends to increase peak clipping. Incidental effects in the amplifier from which the

positive feedback voltage is derived may correspond to positive or negative, voltage or current, feedback, depending on the general circuit arrangement, Design formulas and some simple equivalent circuits are given and the major design considerations are outlined. Typical examples include a detector arrangement which provides automatic-volume-control voltages and has a high input impedance and a very low output impedance, as well as Varrell's arrangement which is in agreement with the design procedure outlined in this paper. Attention is also drawn to the great care necessary in the design of detector-amplifier circuits, using multipurpose valves . . . , to avoid distortion due to incidental negative modulation-frequency feedback to the detector." Methods previously suggested for avoiding the type of detector distortion dealt with in the paper are discussed.

621.396.66 **1664**

A.S.C. Radio.—E. Aisberg. (*Radio Craft*, Jan. 1946, Vol. 17, No. 4, pp. 240..263.) Automatic selectivity control is obtained by use of parallel i.f. channels of different bandwidths for the upper and lower audio ranges, the effect being enhanced by crossing the a.v.c. lines.

621.396.822 **1665**

Noise Figures of Microwave Receivers.—W. G. Hawkins. (*Radio, N.Y.*, Jan. 1946, Vol. 30, No. 1, pp. 21..64.) An elementary survey. The noise figure is defined, and its measurement is explained, with numerical examples for typical microwave receivers.

STATIONS AND COMMUNICATION SYSTEMS

621.39 **1666**

I.R.E. Winter Technical Meeting January 1946.—(*Waves and Electrons*, Feb. 1946, Vol. 1, No. 2, pp. 88..93.) Abstracts are given of the following papers read at the meeting. Phase and Frequency Modulation—A New Method.—R. Adler, F. M. Bailey & H. P. Thomas. Some Technical Developments in Light-Wave Communications.—J. M. Luke & N. E. Porter. A New System of Angular Velocity Modulation Employing Pulse Techniques.—J. F. Gordon. Naval Warfare Communications Problems.—J. O. Kinert. Two Multichannel Microwave Radio-Relay Equipments for the U.S. Army Communication Network.—R. E. Lacy. Frequency Allocations.—P. D. Miles. For other abstracts, see *Electronics*, March 1946, Vol. 19, No. 3, pp. 92..108, and *Electronic Industr.*, March 1946, Vol. 5, No. 3, pp. 62..74. For titles of other papers read, see other sections.

621.395.42 : 534.7 **1667**

Infrasonic Switching.—A. Montani. (*Electronics*, March 1946, Vol. 19, No. 3, pp. 214..222.) It is shown that the ear perceives as continuous a sound which has been suppressed for 0.033 second at a repetition rate of 15 c/s. A system is outlined, utilizing this phenomenon, to give simultaneous two-way communication over a single-channel link.

621.396.1 **1668**

Diversity System.—C. W. Hansell. (*Electronic Industr.*, March 1946, Vol. 5, No. 3, p. 104.) A two-carrier system in which the channels are derived at the transmitter by beating a strong carrier with a weak one (on which the intelligence modulation is superposed) of slightly different frequency. The amplitude modulation due to the mixing process is removed and the resultant phase-modulated carrier (having two principal intelligence-carrying sidebands separated from the carrier by the beat frequency) is fed to the aerial through amplifiers and frequency multipliers. Summary of U.S. Patent 2 388 053.

621.396.1 **1669**

CAA Alaskan Diversity Receiving System: Part 1.—J. Ivers. (*Communications*, Jan. 1946, Vol. 26, No. 1, pp. 40..46.) Technical description, with circuit diagrams, of an 8-channel system using four frequencies between 100 kc/s and 20 Mc/s. A common a.v.c. is used so that a signal received in one channel suppresses the noise from others. The system provides high-speed automatic tape recording up to 250 w.p.m., radio signal relays, or aural reception.

621.396.61 **1670**

Three-Channel 25-Watt Radiotelephone System for Ship-to-Shore.—D. A. Heisner. (*Communications*, Jan. 1946, Vol. 26, No. 1, pp. 32..44.) The system operates on any one of three crystal-controlled frequencies between 2 and 3 Mc/s. The "push to talk" operation is avoided by rectifying part of the speech voltage and using it to operate a relay that activates the transmitter. The transmitter-receiver occupies a space 20" × 18" × 9½", and the power supply 12" × 10" × 7½". It can operate from 6, 12, 32, or 110 V. The circuits are described.

621.396.619.018.12 + .41 **1671**

"A Review of Frequency Modulated Systems".—A. E. Murphy. (*Proc. Instn Radio Engrs, Aust.*, Nov. 1945, Vol. 6, No. 5, p. 10.) Extract only, dealing with phase modulation.

621.396.619.018.12 **1672**

Further Note on the Phase Modulator and Two Applications.—R. A. Wooding. (*Proc. Instn Radio Engrs, Aust.*, Nov. 1945, Vol. 6, No. 5, pp. 5–8.) A supplement to 2943 of 1945 dealing with a circuit modification, in which grid phase-shift networks are replaced by cathode-circuit reactances, giving phase deviation linear up to a theoretical limit of ± 53 degrees. Application to Armstrong and Chireix systems is described.

621.396.619.018.41 **1673**

Extract from "A Review of Frequency Modulated Systems".—A. E. Murphy. (*Proc. Instn Radio Engrs, Aust.*, Jan. 1946, Vol. 7, No. 1, p. 13.) Non-mathematical outline of the theory of the phase discriminator, using vector diagrams. The use of pre-emphasis and de-emphasis for improving signal/noise ratio is mentioned. Long extract only.

621.396.619.018.41 **1674**

Probable Fallacies and Truths about Frequency Modulation.—E. G. Beard. (*Proc. Instn Radio Engrs, Aust.*, Feb. 1946, Vol. 7, No. 2, pp. 3–14.) Discussion of frequency and phase modulation giving detailed methods and claims of early experimenters. It is shown by diagrams that f.m. gives no advantage over a.m. respecting interference represented by an amplitude-modulated carrier. It is suggested that the high signal/noise ratios of f.m. reception can be achieved by a.m. reception using wide bandwidths and suitable circuits. Non-mathematical explanations are given of five methods of phase-modulating a stabilized frequency.

621.396.619.018.41 1675
Fundamental Relationships of F-M Systems.—N. Marchand. (*Communications*, Jan. 1946, Vol. 26, No. 1, pp. 56..61.) Equations for frequency-modulated waves are derived and compared with those for amplitude and phase modulation. This is the first of a series of papers on the operation and design of f.m. transmitters.

621.396.619.16 1676
Pulse Modulation.—R. B. Shepherd. (*Wireless Engr*, April 1946, Vol. 23, No. 271, pp. 114–115.) A letter criticizing statements by Roberts and Simmonds in 183 of January. "A chain of rectangular pulses is a succession of discrete phenomena to which the idea of continuous variation cannot be applied." A mathematical method of obtaining the Fourier series of the modulated chain of pulses is outlined.

621.396.619.16 1677
PPM — New Technique.—F. Shunaman. (*Radio Craft*, Feb. 1946, Vol. 17, No. 5, pp. 314..359.) An outline of the system of pulse-position modulation with particular reference to the AN/TRC-6 system described in 1056 of April.

621.396.931 1678
New Radio Dispatching System.—J. E. Hubel. (*Radio News*, March 1946, Vol. 35, No. 3, pp. 68, 70.) Short general description of the Milwaukee police and fire departments' radio control system.

621.396.933.42 1679
The Martin Aircraft H-F Test Network.—F. Albrecht. (*Communications*, Jan. 1946, Vol. 26, No. 1, pp. 15–19, 45.) A general description of the radio equipment used at a small airport and testing ground, including the circuit diagram of a 75-W, 6-channel h.f. radio-telephone transmitter and receiver.

621.396.1 1680
Two-Way Radio. [Book Review]—S. Freedman. Ziff-Davis, Chicago, 1946, 506 pp., $5.00. (*Electronic Industr.*, March 1946, Vol. 5, No. 3, pp. 153–154.) Deals with the systems and equipment used in two-way or point-to-point communications.

SUBSIDIARY APPARATUS

531.788 + 533.5 1681
Audio Aid for Vacuum-Leak Hunting.—Wouk. (*See* 1550.)

535.215 : 621.383 1682
Investigation of the Surface Photoelectric Effect of Metallic Films under the Influence of Strong Electrostatic Fields.—Jacobsmeyer. (*See* 1504.)

621.3.087.64 1683
Electrodynamic Direct-Inking Pen. — H. B. Shaper. (*Electronics*, March 1946, Vol. 19, No. 3, pp. 148–151.) "Mathematical analysis, choice of drive, constructional details, and performance curves of a high-speed unit capable of inking wave forms of transients directly on a moving paper chart. Response is essentially flat from zero up to 100 c/s." A compact design permits easy construction of multichannel recorders.

621.314.1 1684
High-Voltage Rectified Power Supply Using Fractional-Mu Radio-Frequency Oscillator.—R. L. Freeman & R. C. Hergenrother. (*Proc. Inst. Radio Engrs, N.Y.*, Part II, March 1946, Vol. 34, No. 3, pp. 145–147.) A specially designed triode is arranged in a self-biased oscillator circuit operating at a few hundred kc/s, in which the unidirectional voltage built up across the grid leak is many times the applied anode voltage. ". . . the oscillator circuit should have a large voltage step-up ratio from anode to grid and the [valve] amplification factor . . . should be . . . less than unity . . . The transconductance of the tube at small grid voltages should exceed the total circuit conductance." Various experimental valve types are described. Performance figures quoted include :—4.9 kV across a leak of 23.6 MΩ for an anode current of 21.5 mA at 312 V, and 7.75 kV across 12.6 MΩ at 43.0 mA, 485 V anode input.

621.314.22/.23 1685
The Impedances of Multiple-Winding Transformers: Part I.—S. A. Stigant. (*Beama J.*, Feb. 1946, Vol. 53, No. 104, pp. 70–74.) Practical formulae used for the predetermination of power-transformer leakage reactances have fundamental academic shortcomings. The reactances may be measured however, and a mathematical analysis shows how they affect the current and voltage equations.

621.314.632 1686
A Method for Measuring Effective Contact E.M.F. between a Metal and a Semi-Conductor.—Stephens, Serin & Meyerhof. (*See* 1519.)

621.314.632.029.62 : 546.289 1687
Germanium Crystal Diodes.—Cornelius. (*See* 1728.)

621.316.722.078.3 : 621.396.682 1688
Voltage-Regulated Power Supplies : Part III.—G. E. Hamilton & T. Maiman. (*Communications*, Jan. 1946, Vol. 26, No. 1, pp. 48–49.) Further details of the equipment described in 1076 of April.

621.316.74 1689
Analysis of Operation of a Thermostat with Contact Thermo-Regulator.—Z. Jelonek. (*Proc. Camb. phil. Soc.*, Feb. 1946, Vol. 42, Part 1, pp. 62–72.) Theoretical analysis for the case in which heater current starts and stops suddenly in the form of rectangular pulses.

621.316.86 1690
High Stability Carbon Resistors.—G. V. Planer & F. E. Planer. (*Electronic Engng*, March 1946, Vol. 18, No. 217, pp. 66 . . 97.) Description of the manufacture and properties of the pyrolytic or cracked-carbon resistor. The particle size and the choice of a suitable ceramic carrier give a low temperature-coefficient of resistance (0.02–0.07% per deg C). The resistance value is adjusted by grinding a helical groove in the carbon deposit, but this increases the self-inductance. Carbon composition resistors are briefly discussed, and the difficulty of controlling their values is stressed. A new technique of carbon-coating fine carrier particles (*e.g.* quartz) gives improved controllability, but the process is elaborate.

621.317.39.087.4 + 621.3.078 1691
Instruments for the Automatic Controlling and Recording of Chemical and other Processes.—D. G. Prinz. (*J. sci. Instrum.*, March 1946, Vol. 23,

No. 3, pp. 41–44.) Report of a joint conference of the Institute of Physics, the Institution of Chemical Engineers, and the Chemical Engineering Group (Society of Chemical Industry) held in London on 19th October 1945.

621.317.755 1692

A Three-Beam Oscillograph for Recording at Frequencies up to 10,000 Megacycles.—G. M. Lee. (*Proc. Inst. Radio Engrs, N.Y.*, Part II, March 1946, Vol. 34, No. 3, pp. 121–127.) ". . . A description is given of a [continuously evacuated] three-beam high-speed micro-oscillograph which extends the range of application of single-sweep oscillographic recording by a factor of approximately 10 in frequency over previous limits imposed by transit-time distortion. The reduction in deflection sensitivity attributable to transit-time effect is calculated to be 4 per cent at 3 000 megacycles and 40 per cent at 10 000 megacycles. Single-sweep oscillograms of 3 000 and 10 000 megacycle oscillations and breakdown transients with fronts of the order of 10^{-9} second duration are shown." Electromagnetic focusing is used (spot diameter 0.01 mm). Provision is made to prevent fogging of photographs due to light from the filament. Enlargement of the recorded trace by 100 diameters is used and, at this magnification, the sensitivity is 0.1 mm/V. For recording in the microwave region an accelerating voltage of 50 kV is used. At this voltage the total beam current is 5 mA, of which about 2% strikes a triggering probe in the deflexion chamber. The tube is biased to cut off in the stand-by position. For examination of single-stroke phenomena the bias is restored to normal for 1 millisecond; the front edge of the resulting current pulse received by the probe is used to trigger the sweep, the phenomenon to be examined, or both. "The instrument in its present state of development opens up entirely new fields of research. . . ." See also 1970 of 1940 (von Ardenne).

621.317.755.087.5 1693

High Speed Photography of the Cathode-Ray Tube.—H. Goldstein & P. D. Bales. (*Rev. sci. Instrum.*, March 1946, Vol. 17, No. 3, pp. 89–96.) Techniques are described for the periodic recording of single traces at rates up to 4 000 per second. Writing speeds as high as 70 cm/μsec can be obtained without sacrifice of sensitivity, using commercial tubes and films. High-speed 16-mm cameras are also described.

621.327.4 1694

Gaseous Discharge Tubes and Applications.—Mulliard. (*See* 1729.)

621.352.12 1695

Characteristics of Mercury Type Batteries.—(*Electronic Industr.*, March 1946, Vol. 5, No. 3, p. 74.) Short statement of the design and properties of the Ruben-Mallory cell. High ampere-hour-capacity/volume ratio: capacity is the same under continuous and intermittent operation: long shelf life: good performance at high temperatures: low resistance.

621.352.7 1696

Dry Battery Characteristics and Applications.—M. Potter. (*Proc. Instn Radio Engrs, Aust.*, Jan. 1946, Vol. 7, No. 1, pp. 3–11.) The theory and construction of dry cells is outlined, with mention of the newer flat forms of construction using duplex plates and rubber or plastic seals between cells. This form of construction saves up to 50% in volume and 33% in weight. The measurement of voltage, short-circuit current, and internal resistance is discussed. "The test that best represents any particular service is that which most nearly represents the rate of energy output of the battery when in actual use. There is no direct relation between the results of continuous tests and intermittent tests of longer duration." Curves show the effect of discharge conditions on service life and ampere-hour capacity. Decrease in temperature increases the shelf life but increases the internal resistance, and the cells become useless at about −21° C, although no permanent damage is done. Applications of cells, and design considerations for battery-operated receivers are reviewed.

621.383.5 1697

An Apparatus for Use in the Selection of Barrier-Layer Photocells.—J. A. Hall. (*J. sci. Instrum.*, March 1946, Vol. 23, No. 3, pp. 59–60.) 12 cells mounted in a wheel are in turn balanced in a bridge circuit against a standard cell, all the cells being illuminated by the same source for some time before the comparison is made.

621.385.832 : 621.316.578.1 1698

Tube-Seasoning Timer.—M. Silverman. (*Electronics*, Feb. 1946, Vol. 19, No. 2, p. 145.) For applying "seasoning" voltages to batches of about 20 cathode-ray tubes during manufacture. Comprises time switches connected in the electrode circuits, which operate at various times, up to a maximum of 55 min, after the commencement of the seasoning operation.

621.394/.397].645 1699

Network Design Using Electrolytic Tanks.—Kenyon. (*See* 1470.)

621.396 1700

I.R.E. Winter Technical Meeting January 1946.—(*Waves and Electrons*, Feb. 1946, Vol. 1, No. 2, pp. 88..93.) Abstracts are given of the following papers read at the meeting. Electrical Characteristics of Quartz Crystal Units and Their Measurement.—W. D. George, M. C. Selby & R. Scolnik. Linear Servo-Theory.—R. E. Graham. A Three-Beam Oscillograph for Recording at Frequencies up to 10 000 Megacycles.—G. M. Lee (See 1692.). Problems Associated with the Standardization of Quartz-Crystal Units for Military Equipment.—C. J. Miller, Jr. Directional Couplers.—W. W. Mumford. High-Frequency Plated Quartz-Crystal Units for Control of Communications Equipment.—R. A. Sykes. Hermetically Sealed Metal Holder for Crystal Units.—A. W. Zeigler. For other abstracts, see *Electronics*, March 1946, Vol. 19, No. 3, pp. 92..108, and *Electronic Industr.*, March 1946, Vol. 5, No. 3, pp. 62..74. For titles of other papers read, see other sections.

621.396.662.2 1701

Wire Length of Universal Coils.—A. W. Simon. (*Electronics*, March 1946, Vol. 19, No. 3, p. 162.) Procedure for calculating outside diameter and total wire length in terms of number of turns, form diameter, gear ratio and wire size. See also 1085 of April (Simon).

621.398 1702
Self-Synchronous Transmission System.—E. Hansen. (*Radio News*, Feb. 1946, Vol. 35, No. 2, pp. 38..143.) The use of the selsyn motor for remote control. Simple diagrams illustrate the basic principles.

TELEVISION AND PHOTOTELEGRAPHY

621.383.8 1703
The Image Orthicon.—(*Radio, N.Y.*, Jan. 1946, Vol. 30, No. 1, pp. 10..16.) See also 1376 of May.

621.397 1704
I.R.E. Winter Technical Meeting January 1946.—(*Waves and Electrons*, Feb. 1946, Vol. 1, No. 2, pp. 88..93.) Abstracts are given of the following papers read at the meeting. Television in the Ultra-High Frequencies.—P. C. Goldmark. Ultra-High - Frequency - Television Receivers.—H. T. Lyman. Television-Studio Equipments.—J. J. Reeves. The Image Orthicon, A Sensitive Television Pickup Tube.—A. Rose, P. K. Weimer & H. B. Law. Electro-optical Characteristics of Television Systems.—O. H. Schade. Sight and Sound on One Carrier.—K. Schlesinger. A Kinescope for Home Projection-Type Television Receivers.—L. E. Swedlund. Color-Television Transmitter.—N. H. Young. For other abstracts, see *Electronics*, March 1946, Vol. 19, No. 3, pp. 92..108, and *Electronic Industr.*, March 1946, Vol. 5, No. 3, pp. 62..74. For titles of other papers read, see other sections.

621.397 1705
A Writing Cathode-Ray Tube.—H. Lineback. (*Radio News*, Feb. 1946, Vol. 35, No. 2, pp. 30, 31, 126.) A device for making a c.r.t. spot follow a predetermined path of arbitrary shape, *e.g.* trace a written word. The required path is translated by a graphical process into an opaque pattern on a transparent disk, something like a variable-width film sound track. The pattern is scanned by two photocells, the outputs from which control the deflector voltages of the c.r. tube.

621.397 1706
Television Must Sound Right.—R. Hubbell. (*Radio News*, Feb. 1946, Vol. 35, No. 2, pp. 46..137.) The need for new and spacious studios is stressed. The main points for their design are outlined.

621.397 1707
CBS Shows its Color.—(*Electronic Industr.*, March 1946, Vol. 5, No. 3, pp 75..125.) Demonstration of colour television (using rotating colour disk) with sound transmitted by frequency modulation during the fly-back period. A short description of some items of the equipment used.

621.397.3 1708
Highlights of Dr. P. C. Goldmark's Paper on Color Television.—R. G. Peters. (*Communications*, Jan. 1946, Vol. 26, No. 1, pp. 50–51.) A brief account of the Baird tri- and di-chromatic systems, a comparison between colour photography and television, and a discussion of the present C.B.S. tri-chromatic system.

621.397.335 1709
Facsimile Synchronizing Methods.—D. Schulman. (*Electronics*, March 1946, Vol. 19, No. 3, pp. 131–133.) The required accuracy of synchronization of transmitter and receiver drives is investigated, and methods for obtaining this accuracy are described, ranging from simple manual control at the receiving end, to electronic methods. Self-contained frequency-control systems operate on local standard-frequency sources (tuning fork, or crystal), the phase of receiver and transmitter being matched by a transmitted pulse. Other methods have been devised in which the same audio-frequency tone drives both transmitter and receiver motors, or where each line in the receiver scan is synchronized by a transmitted pulse. Descriptions are given of several devices aimed at stabilization of drive frequencies.

621.397.335 1710
Synchronizing and Separation Circuits : Part 12.—E. M. Noll. (*Radio News*, March 1946, Vol. 35, No. 3, pp. 54 . . 60.) For parts 10 and 11 see 782 of March and 1381 of May. (The series does not *begin* with 782 as stated in 1381.)

621.397.335 : 621.396.615.17 1711
Synchronizing Generators for Electronic Television.—Applegarth. (*See* 1488.)

621.397.645 1712
Compensating Amplifier. — C. N. Gillespie. (*Electronics*, March 1946, Vol. 19, No. 3, pp. 232.. 234.) To compensate for the distortion produced by the nonlinearity of the recording method and paper of a facsimile system, an equal but opposite nonlinearity is introduced into the amplifier. Summary of U.S. Patent 2 378 999.

621.397.5 1713
Television Programming and Production. [Book Review]—R. Hubbell. Murray Hill Books, New York, 1945, 203 pp., $3.00. (*Electronic Engng*, March 1946, Vol. 18, No. 217, p. 96.) " . . . I have not seen a better, or simpler introduction to the peculiar problems which beset the television producer or his team. . . ." See also 785 of March.

TRANSMISSION

621.396.61 + 621.396.645 1714
Analysis of Parasitic Oscillations in Radio Transmitters.—J. S. Jackson. (*Radio News*, Feb. 1946, Vol. 35, No. 2, pp. 68 . . 88.) The occurrence and suppression in class-C r.f. amplifiers.

621.396.61 1715
I.R.E. Winter Technical Meeting January 1946.—(*Waves and Electrons*, Feb. 1946, Vol. 1, No. 2, pp. 80 . . 93.) Abstracts are given of the following papers read at the meeting. Generation of Continuous-Wave Power at Very High Frequencies.—W. G. Dow, J. N. Dyer, W. W. Salisbury & E. A. Yunker. Electronic Frequency Stabilization of Microwave Oscillators.—R. V. Pound. Ultra-High-Frequency Television Transmitters and Antennas.—R. Serrell. Color-Television Transmitter.—N. H. Young. For other abstracts, see *Electronics*, March 1946, Vol. 19, No. 3, pp. 92 . . 108, and *Electronic Industr.*, March 1946, Vol. 5, No. 3, pp. 62 . . 74. For titles of other papers, see other sections.

621.396.61 : 621.396.619.018.41 1716
Twelve-Tube F-M Handie-Talkie.—J. M. Lee. (*Electronics*, March 1946, Vol. 19, No. 3, pp. 194 . . 200.) Weight 8 lb, frequencies 47–55.4 Mc/s, range 1 mile in open country and 300 yd in dense

jungle. A U.S. Signal Corps equipment, type AN/PRC-6.

621.396.61 : 621.396.933 1717

Low Power Aircraft Transmitter.—A. B. Kaufman. (*Radio News*, March 1946, Vol. 35, No. 3, pp. 29 . . 88.) A 3-Mc/s crystal-controlled transmitter using a suppressor-modulated output valve, has low current consumption and reasonable voice quality.

621.396.61.029.56/.58 1718

A Band-Switching V.F.O. Exciter Unit.—W. E. Bradley. (*QST*, March 1946, Vol. 30, No. 3, pp. 29–33.) With calibrated bandspread for 3.5, 7, 14 and 28 Mc/s. Special attention is given to the switching device and mechanical construction. Performance data are given.

621.396.61.029.58 1719

350 Watt — 5 Band Transmitter.—H. S. Brier. (*Radio News*, March 1946, Vol. 35, No. 3, pp. 40 . . 125.) Constructional details of a self-contained crystal-controlled transmitter with switched coils for amateur bands from 3.5 to 29.7 Mc/s.

621.396.61.029.58 1720

Home Built Signal Shifter.—S. Heytow. (*Radio News*, March 1946, Vol. 35, No. 3, pp. 48 . . 117.) Constructional details of a variable-frequency oscillator for amateur transmitters, giving 20 W output at 20- or 80-metre wavelengths.

621.396.61.029.58 1721

A Low Power 28-Mc. 'Phone-C.W. Transmitter.— . Mix. (*QST*, March 1946, Vol. 30, No. 3, pp. 18–21.) Circuit and constructional details. The power supply unit can be used for either the transmitter or a receiver, by switching. A separate modulator is provided for telephone operation. "Complete equipment for the beginner."

621.396.61.029.62 : 621.396.619.018.41 1722

Design of F-M Transmitter for 88-108 Mc.— L. Sack. (*Electronics*, March 1946, Vol. 19, No. 3, pp. 184–194.) A high-fidelity transmitter, output 250 W, incorporates balanced modulators, push-pull primary oscillator, and linear circuit elements in the anode tank circuits. Stable centre-frequency control is obtained by the variation of the primary oscillator tuning capacitance by a motor controlled by the difference in frequency between the primary oscillator and a standard crystal oscillator.

621.396·615.12.029.62 1723

144-Mc Transmitter.—I. Queen. (*Radio Craft*, Feb. 1945, Vol. 17, No. 5, pp. 319 . . 363.) Simple description of self-oscillator circuits, using standard receiver-type valves.

621.396.619.2 1724

Premodulation Speech Clipping and Filtering.— W. Smith. (*QST*, Feb. 1946, Vol. 30, No. 2, pp. 46–50.) By incorporating amplitude limitation, clipping, and low-pass filtering in an amplitude-modulated transmitter, it is possible to realize an effective power gain of nearly 100 times. The intelligibility of weak signals is greatly increased with a slight loss in quality.

621.396.664 1725

Instantaneous Program Switching.—J. Zelle. (*Electronics*, Feb. 1946, Vol. 19, No. 2, pp. 142–144.) Simple circuit for feeding a given wire channel in succession from a number of input channels. "The arrangement has three advantages: circuits can be preset in the order in which they are to be used, so that switching becomes virtually errorless and semi-automatic; secondly the switching will be practically instantaneous . . . ; and thirdly, the circuits are interlocked, therefore only one circuit can be put on the outgoing channel at a time."

VALVES AND THERMIONICS

537.525 : 621.3.029.64 1726

Initiation of High Frequency Gas Discharges.— T. Holstein. (*Phys. Rev.*, 1st/15th Jan. 1946, Vol. 69, Nos. 1/2, p. 50.) The microwave sparking field in a gas is investigated theoretically as a function of electrode spacing, pressure, electron temperature and Townsend's coefficient α. Summary of an Amer. Phys. Soc. paper.

537.581 1727

An Explanation of Anomalous Thermionic Emission Current Constants.—Sun & Band. (*See* 1513.)

621.314.632.029.62 : 546.289 1728

Germanium Crystal Diodes.—E. C. Cornelius. (*Electronics*, Feb. 1946, Vol. 19, No. 2, pp. 118–123.) Construction, theory of operation, properties, and typical applications of the rectifier type 1N34. The unit is enclosed in a small metal cylinder with leads for soldering it into circuit. It consists of a tungsten wire touching an optically polished surface of germanium mixed with a small quantity of tin. The equivalent circuit, and the current/voltage characteristics are discussed in relation to the energy levels of the free electrons in the crystal and the tungsten. It is shown that the slope resistance becomes negative when the steady voltage across the device exceeds a particular value that depends on the polarity of the voltage, the temperature, and the particular crystal unit used. The 1N34 is suitable for use as a rectifier for frequencies up to 100 Mc/s, and with low load resistances compares well with a conventional thermionic diode. "Preliminary tests show that no failure or deterioration has occurred for more than 1 000 hours of continuous operation [under adverse conditions]." Other applications, *e.g.* voltage regulation and for relaxation oscillators (up to 500 kc/s) are briefly mentioned. For brief accounts see *Electronic Industr.*, Dec. 1945, Vol. 4, No. 12, p. 80, and *Electronics*, Jan. 1946, Vol. 19, No. 1, p. 252.

621.327.4 1729

Gaseous Discharge Tubes and Applications.— R. C. Hilliard. (*Electronics*, March 1946, Vol. 19, No. 3, pp. 122–127.) Details are given of a robust and reliable modulator glow lamp suitable for portable sound-on-film recording systems. The lamp is of the crater type in which the ionization of the gas excites the cathode material, giving a high-intensity point source of light. The neon-filled strobotron designed as a medium-intensity light source for stroboscopic applications is also described. It can be operated either as an arc discharge tube with short-duration high-peak anode currents (300–400 A), or as a glow-discharge tube with low currents, as required in control devices. Other types of this tube have been developed for special features of operation such as high hold-off voltage (2kV), high repetition rate (1 kc/s), low time-jitter, and high light output, and for use in systems requiring the generation of short current pulses of high intensity (*e.g.* welding control). For photographic analysis of motion and for extremely short exposure photography, a series of externally excited discharge tubes with high light output and short flash-duration have been developed. The design of the discharge circuit is discussed. See also 2351 of 1937 (Edgerton, Germeshausen, Nottingham & White).

621.38 **1730**

I.R.E. Winter Technical Meeting January 1946.—(*Waves and Electrons*, Feb. 1946, Vol. 1, No. 2, pp. 80..93.) Abstracts are given of the following papers read at the meeting. Design of Small Size High-Voltage Rectifier Type 1Z2.—G. Baker. Two New Miniature Tubes for Frequency-Modulation Conversion.—R. M. Cohen, R. C. Fortin & C. M. Morris. Microwave Magnetrons. — G. Collins. Improved Cathode-Ray Tubes with Metal-Backed Luminescent Screens.—D. W. Epstein & L. Pensak. A Medium-Power Triode for Frequencies about 600 Megacycles.—S. Frankel, J. J. Glauber & J. Wallenstein. Microwave Triodes Adapted to Modern Usage.—E. M. Goodell. Cascade Amplifier Klystrons.—E. C. Levinthal. Secondary-Emission Cathodes for Magnetrons.—J. W. McNall, H. C. Steele, Jr. & C. L. Shackelford. Glass Problems in the Manufacture of Miniature Tubes.—H. J. Miller. Joint Electron Tube Engineering Council.—O. W. Pike. Magnetron Cathodes.—M. A. Pomerantz. The Image Orthicon, A Sensitive Television Pickup Tube.—A. Rose, P. K. Weimer & H. B. Law. For other abstracts, see *Electronics*, March 1946, Vol. 19, No. 3, pp. 92..108, and *Electronic Industr.*, March 1946, Vol. 5, No. 3, pp. 62..74. For titles of other papers read, see other sections.

621.383.8 **1731**

The Image Orthicon.—(*Radio, N.Y.*, Jan. 1946, Vol. 30, No. 1, pp. 10..16.) See also 1376 of May.

621.385 **1732**

The Resnatron.—W. W. Salisbury. (*Electronics*, Feb. 1946, Vol. 19, No. 2, pp. 92–97.) Constructional details and properties of a very-high-power tunable tetrode suitable for c.w. operation in the decimetre region either as a self-oscillator or as an amplifier. The valve has tuned cavities between directly heated cathode and control grid, and between anode and screen respectively. The tungsten filament requires 2V, 1800 A. Typical plate operating conditions are 8 A, 17½ kV. Under these conditions about 85 kW output may be obtained. The valve described is continuously pumped, and extensive use is made of water cooling. For use in countermeasures operations during the war, see 1356 of May.

621.385 : 621.317.79 **1733**

A Method of Measuring Grid Primary Emission in Thermionic Valves.—Hooke. (*See* 1598.)

621.385.1 **1734**

Reflex-Klystron Oscillators.—E. L. Ginzton & A. E. Harrison. (*Proc. Inst. Radio Engrs, N.Y.*, Part I, March 1946, Vol. 34, No. 3, pp. 97–113.) "A comprehensive analysis of reflex klystrons is developed by considering the electrons as particles acted upon by forces which modify their motion. The analysis is similar to earlier explanations of electron bunching in a field-free drift space and predicts a similar current distribution when bunching takes place in a reflecting field. The effect of the bunched electron beam is treated qualitatively by considering the effect of the beam admittance upon a simple equivalent circuit. A quantitative mathematical analysis based upon oscillator theory is also derived and the results are presented in a series of universal curves which are used to explain the operating characteristics of these tubes. Power output, efficiency, starting current, electronic tuning, and modulation properties are discussed. Some general remarks on reflex-oscillator design considerations are also included." The effect of transit time across the gap of the resonator is considered. The equivalent circuit mentioned above comprises a parallel LC circuit representing the resonator, connected in shunt with the equivalent admittance of the bunched beam and with resistances to simulate circuit losses and loading. The design criteria for reflex klystrons designed to have wide-band electronic tuning are discussed relative to considerations of power output and efficiency.

621.385.1 : 537.122 : 538.3 **1735**

"Electron Transit Time in Time-Varying Fields".—L. A. Ware & H. B. Phillips: A. B. Bronwell. (*Proc. Inst. Radio Engrs, N.Y.*, Part I, March 1946, Vol. 34, No. 3, p. 151.) Comments on 3884 of 1945 (Bronwell).

621.385.1 : 621.365.52 **1736**

Induction Heating in Radio Electron-Tube Manufacture.—Spitzer. (*See* 1615.)

621.385.1.032.2 + 533.5 + 539.234 **1737**

Fine Wires in the Electron-Tube Industry.—Espersen. (*See* 1564.)

621.385.2 : 621.395.645.3 **1738**

A Voltage Amplifier Using a Pre-Saturation Diode as Load.—Durnford. (*See* 1478.)

621.385.832 : 621.316.578.1 **1739**

Tube-Seasoning Timer.—Silverman. (*See* 1698.)

MISCELLANEOUS

061.22 : 621.396(054) **1740**

Waves and Electrons.—The journal which appeared with this title in Jan. and Feb. 1946, is continued as "*Proc. Inst. Radio Engrs, N.Y.*, Part II," from March 1946. See 1122 of April.

621.38/.39] (038) **1741**

Electronics Dictionary. [Book Review]—N. M. Cooke & J. Markus. McGraw-Hill, New York, 1945, 433 pp., $ 5.00. (*Electronic Industr.*, Dec. 1945, Vol. 4, No. 12, p. 162. Also *Electronics*, Jan. 1946, Vol. 19, No. 1, pp. 344 . . 348.)

621.38 (075.8) **1742**

Electronics Laboratory Manual. [Book Review]—R. R. Wright. McGraw-Hill Book Co., New York, 77 pp., $ 1.00 (*Waves and Electrons*, Feb. 1946, Vol. 1, No. 2, p. 103.)

621.396.029.64 **1743**

Radiation Laboratory Technical Series. [Book Notice]—(*See* 1549.)