WIRELESS ENGINEER

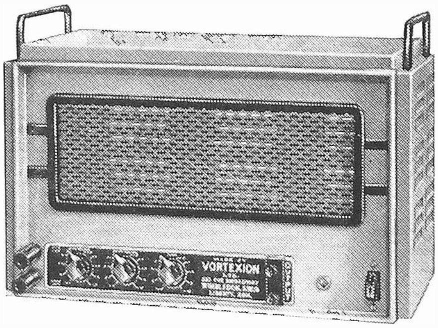
THE JOURNAL OF RADIO RESEARCH & PROGRESS

JUNE 1947

VOL. XXIV TWO SHILLINGS AND SIXPENCE · · No. 285


VORTEXION LTD.
WIMBLEDON.

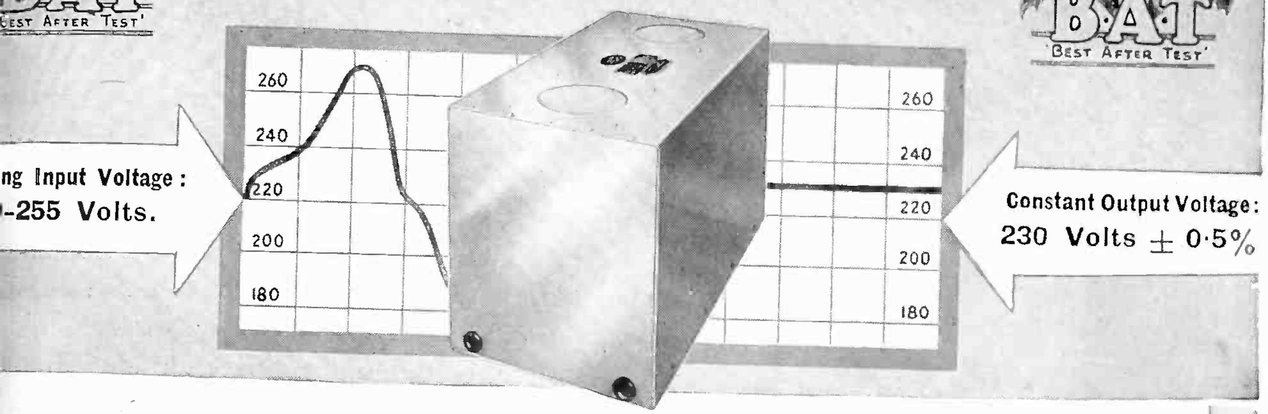

VOLUME
TONE
MAINS


VORTEXION

B.A.T 'BEST AFTER TEST'

AUTOMATIC VOLTAGE REGULATORS

B.A.T 'BEST AFTER TEST'



General Characteristics

Constant A.C. Output

Example : 230 volts ± 0·5%—50-cycles/sec.—single phase. Any output voltage may be ordered (see below).

Wide A.C. Input Voltage Limits

Example : 190-255 volts, 50-cycles, 1-phase. Other single-phase voltages or frequencies can be dealt with, on special orders.

Entirely Automatic—Quick Action

There are no moving parts. No adjustments need ever be made and no maintenance is required. The regulating action is virtually instantaneous, the time required for adjustment to a new voltage, or load condition being so short that it is quite imperceptible by ordinary means.

Load Rating

Eight standard nominal ratings are carried in stock as listed below. Others can be built, including models giving (example) 115 v. ± 1% on 190-255 v. input : or multiple outputs, all regulated. The regulators also stabilize well under all load conditions, from no-load to 100% load.

General Advantages and Uses, etc.

Constant A.C. Input voltage is essential for the effective operation of many electrical devices, both industrial and laboratory patterns. Examples : X-ray apparatus, incandescent-lamp light sources (photometers, photo-printing, colour comparators, photo-electric cell applications, spectrography, etc.), laboratory test-gear (vacuum tube volt meters, amplifiers, oscillators, signal generators, standards of frequency, etc.) : the larger patterns for stabilizing a complete laboratory room or test-bench : the smaller units as integral components of equipment.

Priorities, Etc.

No Priority is now required. "M" Certificates (Iron and Steel Control) are also not now required.

Complete Data

Please request Bulletin VR 10744.

EIGHT STOCK MODELS ARE OFFERED

Type	Watts	A.C. Input Voltage	Output Voltage	Net Wt.	Price
VR-10	10	190-255 50~ 1-phase	230 v. ± 0·5 per cent. Or, as ordered (see text above)	3 lbs.	£5 - 15
VR-20	20			7 lbs.	£8 - 0
VR-60	60			17 lbs.	£10 - 10
VR-150	150			42 lbs.	£13 - 10
VR-300	300			62 lbs.	£22 - 10
VR-500	500			68 lbs.	£29 - 10
VR-1000	1000			120 lbs.	£47 - 10
VR-2500	2500			450 lbs.	£175 - 0

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180, Tottenham Court Road, London, W.1 and 76, OLDHALL ST. LIVERPOOL, 3, LANCS.

The AvoMeter is one of a useful range of "Avo" electrical testing instruments which are maintaining the "Avo" reputation for an unexcelled standard of accuracy and dependability—in fact, a standard by which other instruments are judged.

The Model 7 Universal AvoMeter is th world's most widely used combination elec trical measuring instrument. It provide 50 ranges of readings on a 5″ scale and i guaranteed accurate to B.S. first grad limits on D.C. and A.C. from 25c/s to 2kc/ It is self-contained, compact and portabl simple to operate and almost impossible t damage electrically. It is protected by a automatic cut-out against damage throug severe overload, and is provided wit automatic compensation for variations i ambient temperature.

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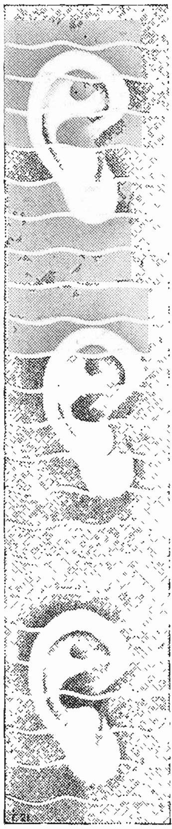

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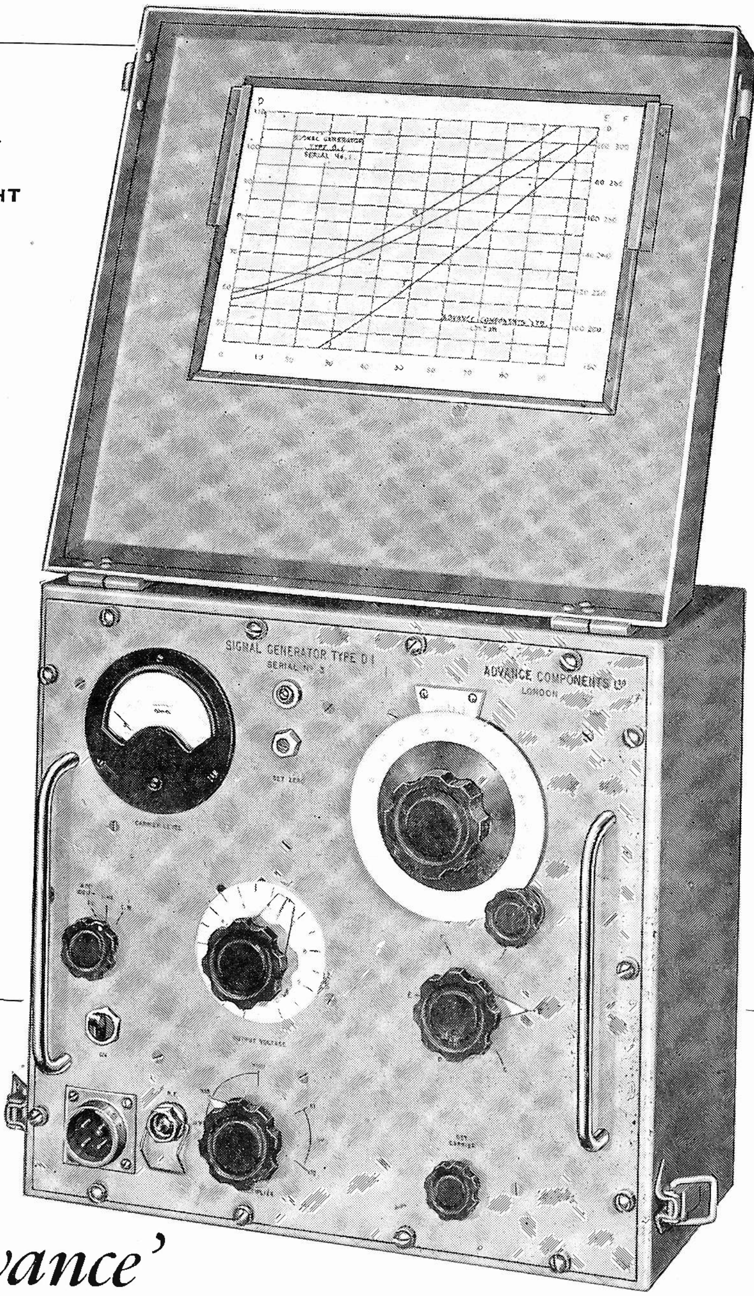

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and pulsed
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at 1,000∼

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★ **FREQUENCY CALIBRATION**
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The instrument is totally enclosed in a grey enamelled steel case with a detachable hinged lid for use during transport.

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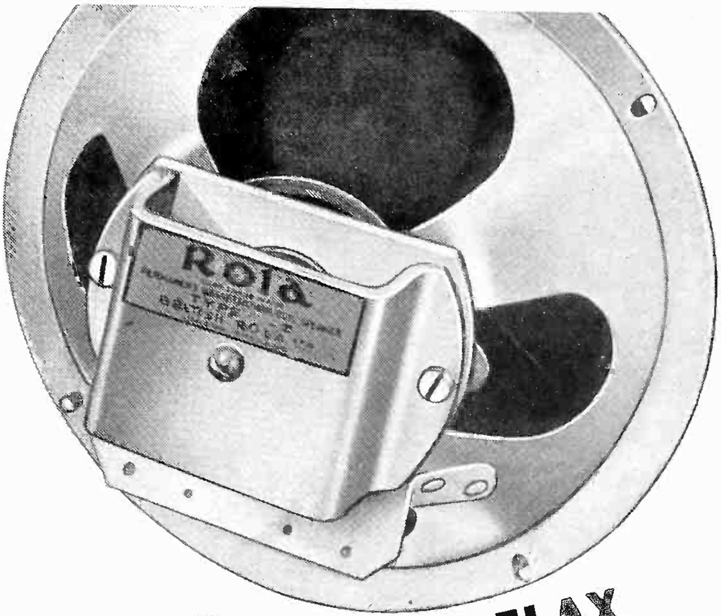
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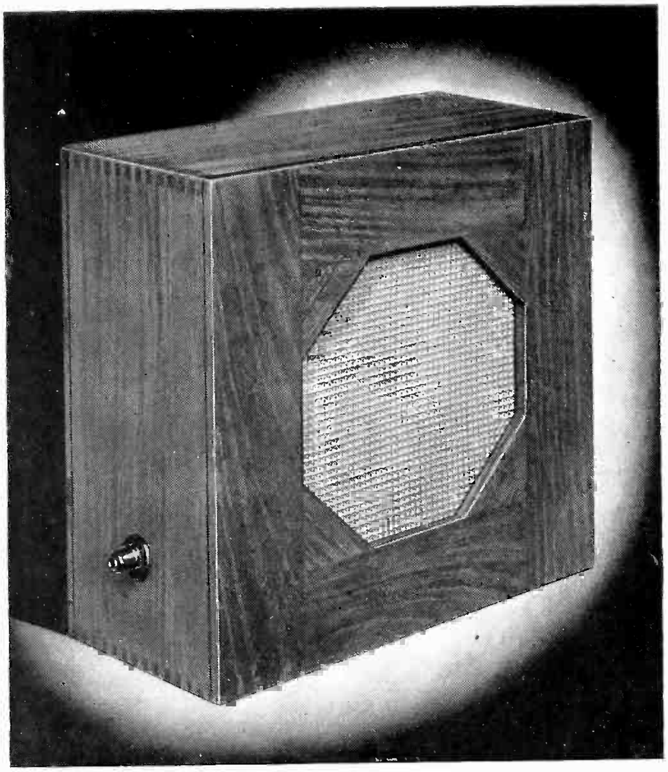

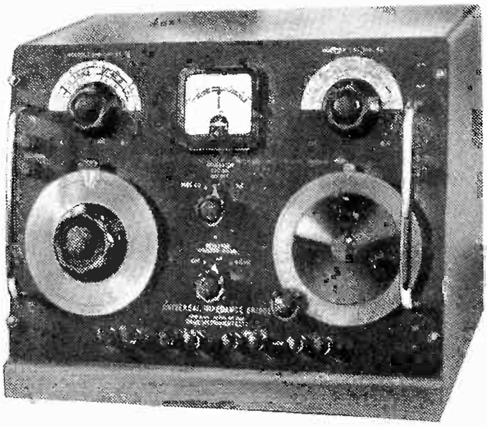

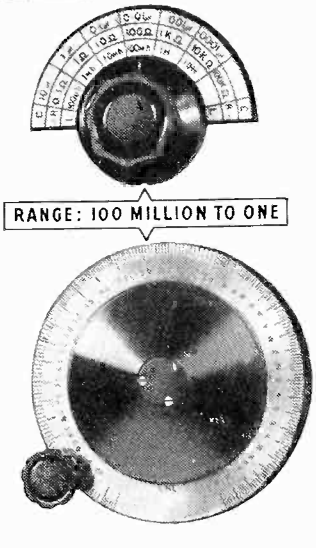
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"We solved it — successfully — by going to our Regional Appointments Office. I don't know a better or easier way. They're linked by teleprinter with all the other Regions, so they cover the whole country. They interview, 'screen' and classify men and women for every kind of job above a certain level. Result: you tell them your needs, and in two or three days you have a short list of candidates for interview — carefully sifted in advance, so that each one is a 'possible' for the post. It's as simple as that. Your *only* difficulty is picking the best.

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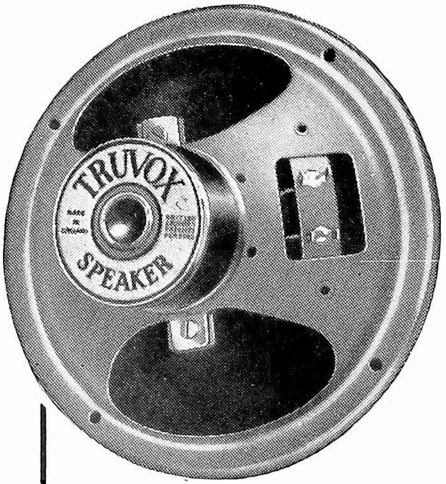
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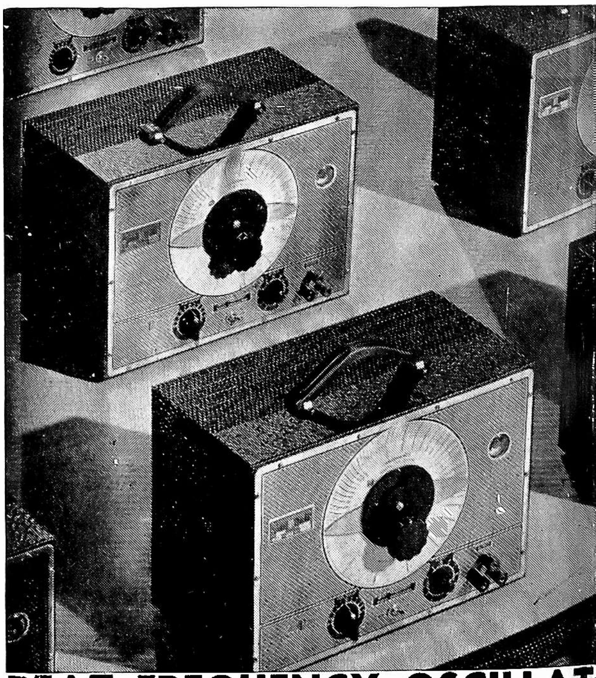

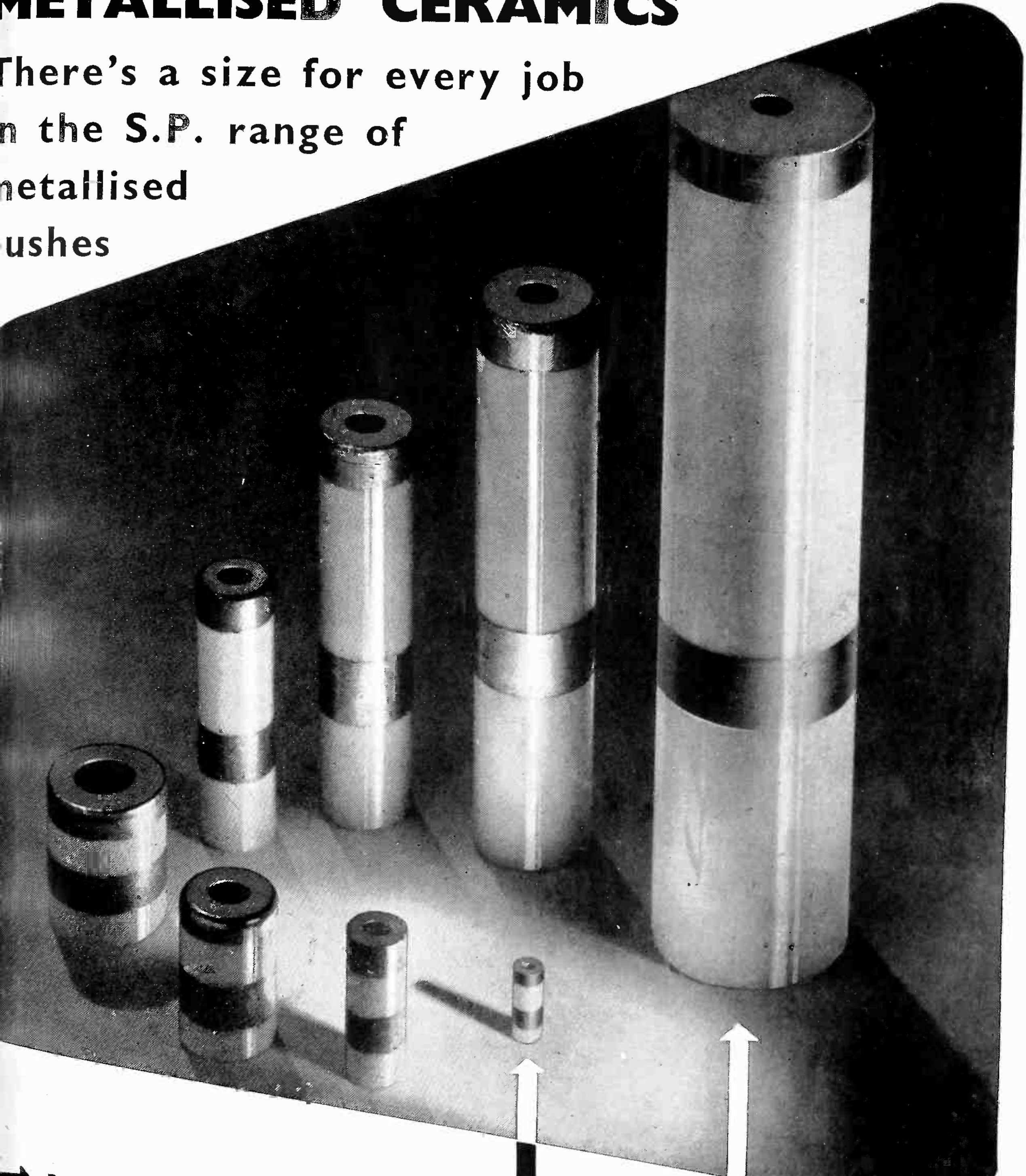


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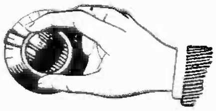

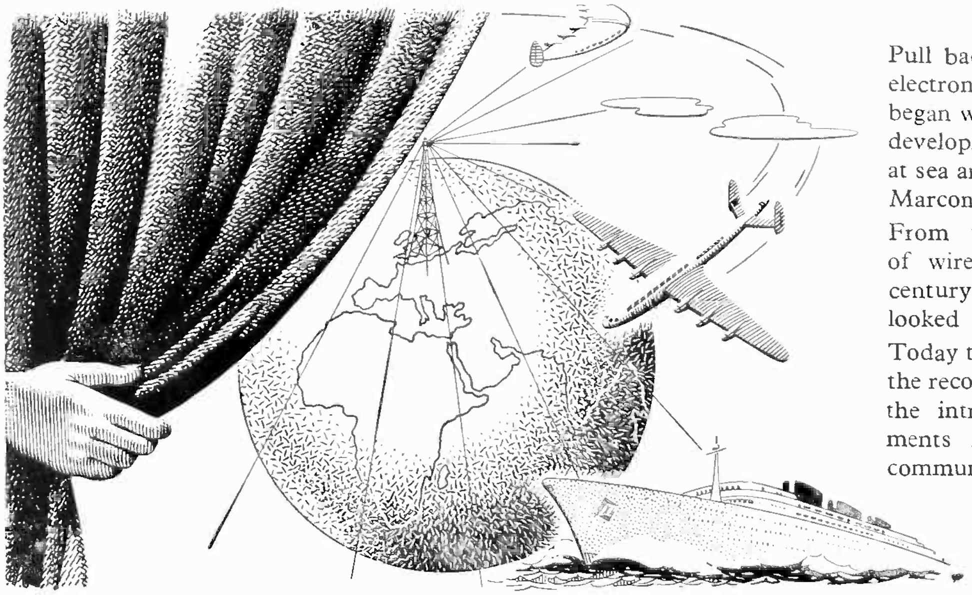

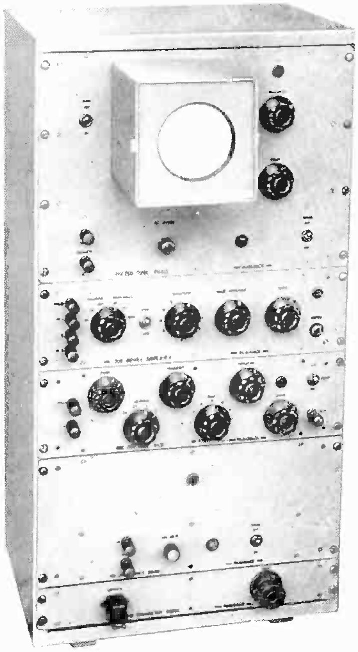


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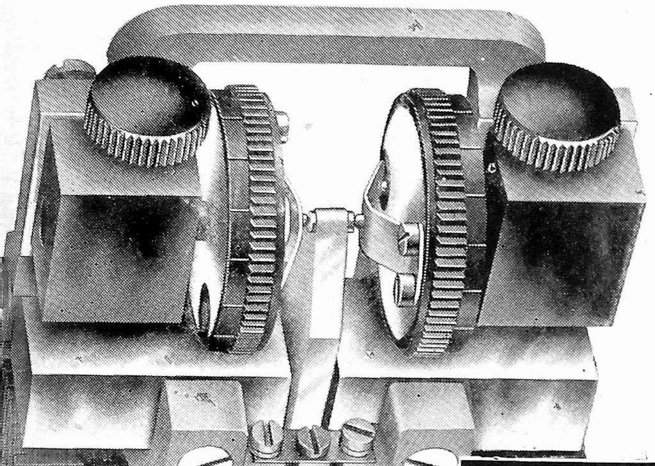
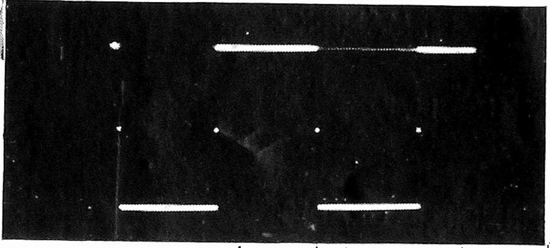


The Carpenter Relay in its standard adjustment reproduces, with a 5 AT input, square pulses from less than 2 milli-seconds upwards with a distortion of 0.1mS, i.e., 5% for 2mS pulses or 1% at 10mS.

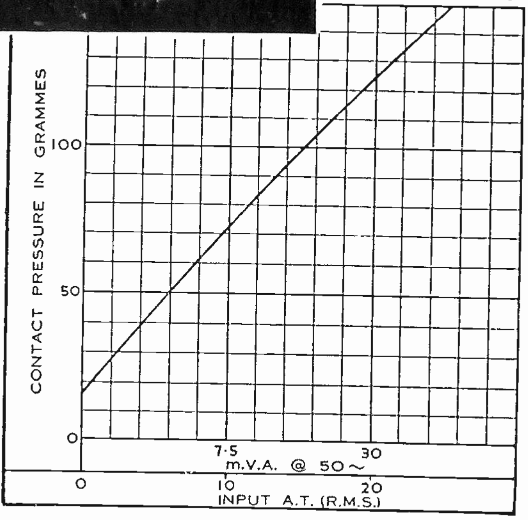
This unequalled performance is due to inherent features of the design of the relay, ensuring short transit time, high sensitivity and low hysteresis.

● (Above) *Contact mechanism of Relay showing damped compliant mountings of side contacts.*

● (Right) *Unretouched photograph (3 sec. exposure)*
f oscillogram showing contact performance of Relay
n special adjustment for a measuring circuit; coil
nput 18 AT (25 mVA) at 50 c/s.



(Below) *Graph showing contact pressures developed at 50c/s against mVA and ampere turns input for type 3E Carpenter Relay.*



There is complete absence of contact rebound
t any input power and contact pressures
re exceptionally high *(see graph).* Adjust-
nent can be made with great ease. Moreover,
ince the armature is suspended at its centre
f gravity, the relay has high immunity from
ffects of mechanical vibration and there
s no positional error. Effective screening is
rovided against external fields. Because of
hese characteristics, the Carpenter Relay
as many applications in the fields of measure-
nent, speed regulation, telecontrol and the
ke, in addition to the obvious use in telegraph
ircuits; details of models suitable for such
urposes will be supplied willingly on request.

MENSIONS IN COVER: $2\frac{3}{8} \times 1\frac{5}{16} \times 4\frac{1}{4}$. WEIGHT with standard socket: 22 ozs.

Ask for Booklet 1017 W.E.

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JUNE 1947

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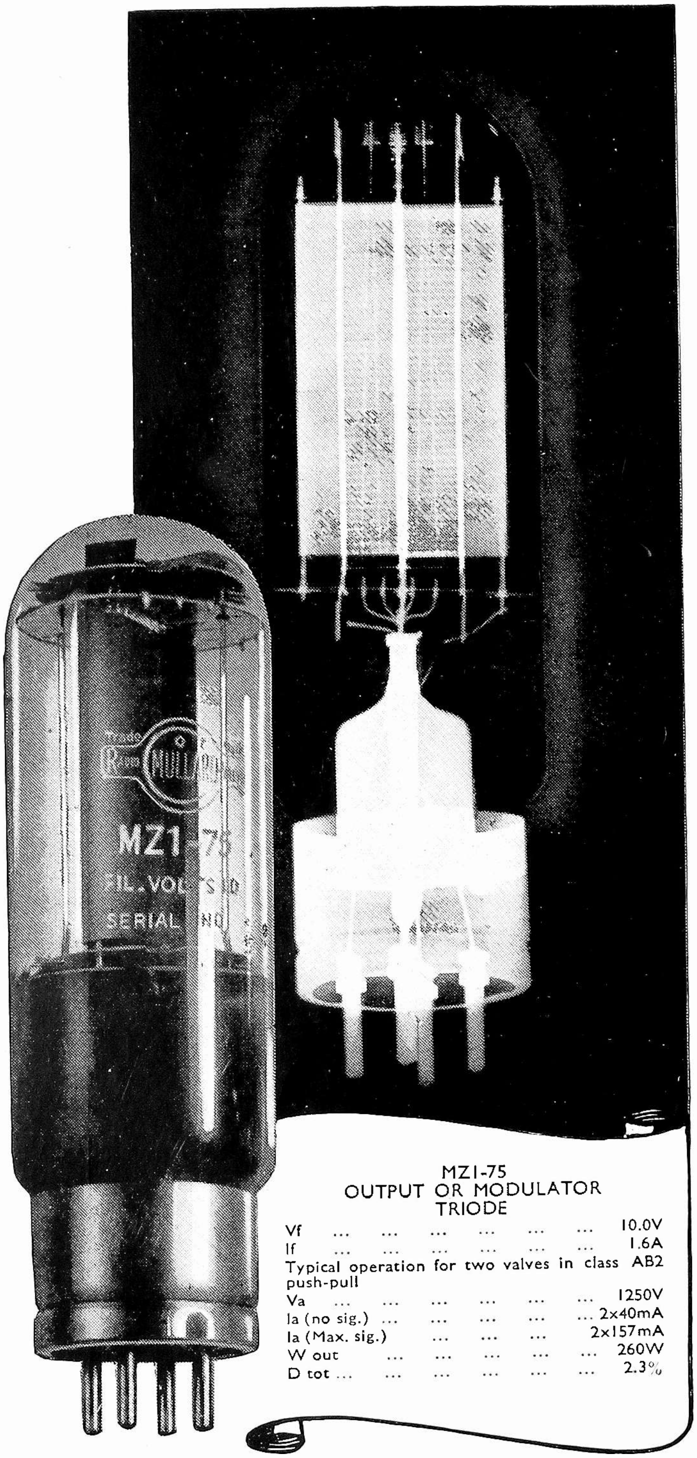

MZ1-75
OUTPUT OR MODULATOR
TRIODE

Vf	10.0V
If	1.6A

Typical operation for two valves in class AB2 push-pull

Va	1250V
Ia (no sig.)	2x40mA
Ia (Max. sig.)	2x157mA
W out	260W
D tot	2.3%

For modulation and L.F. amplification

HERE'S A BETTER VALV

All things being equal, the real measure a valve's performance is the length of tim it will operate with undiminished efficienc In this respect the MZ1-75 scores heavil Like all Mullard transmitting and industri valves, it has been designed with an eye cost per life-hour.

Modern manufacturing methods ensure reliability and offer the additional adva tage that all valves of the same type a characteristically consistent. The long-li oxide-coated filament is unaffected fluctuations in the mains supply, and al permits operation at an economical ano voltage.

A glance at the data panel with the factors in mind will explain why the MZ 75 wins the approval of communicatio engineers who want a better valve modulation and L.F. Amplification.

For further developments watch Mullar
THE MASTER VALV

Mullard

Technical data and advice on the applications of the MZ1-75 Triode can be obtained from
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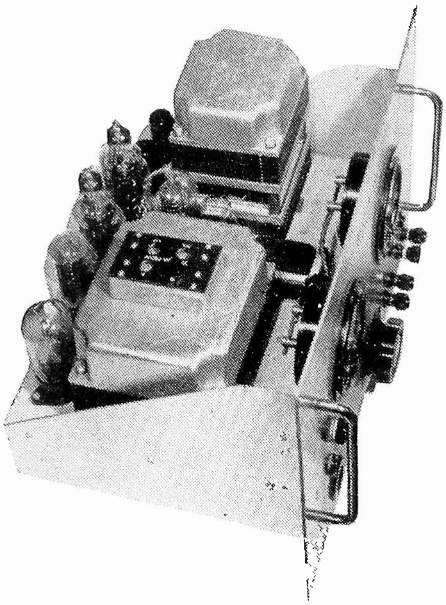

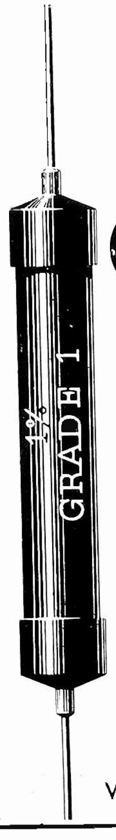
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GRADE 1

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The absolute sensitivity of a radio frequency amplifier is usually expressed as the number of microvolts input required to raise the output a specified number of decibels above the valve and circuit noise.

In practice this input may be less than one microvolt. Most research laboratories associated with manufacturers are usually located on the factories premises so that a high level of radio interference generated by electric motors, welders, relays, etc., is invariably experienced. Measurements, such as previously described, are usually quite impossible unless a well-designed screened compartment is available.

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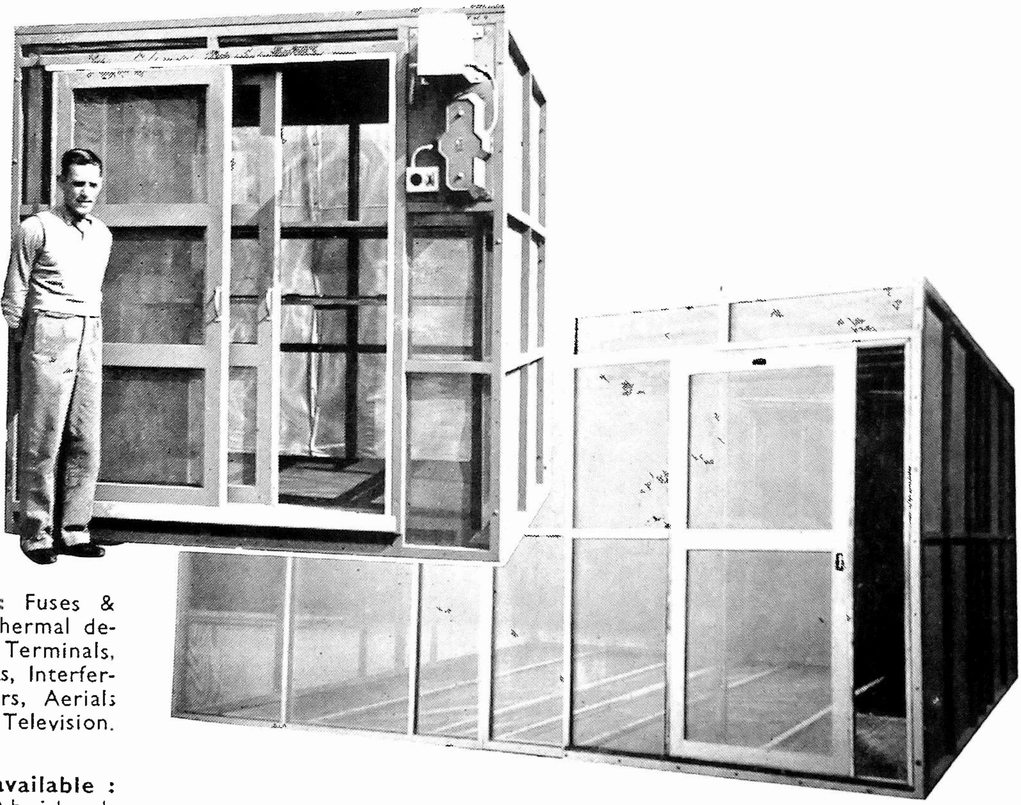

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★ TIME BASE 0.2 c/s to 150 Kc/s. Variable through X amplifier 0.2 to 5 screen diameters. Singl sweep available

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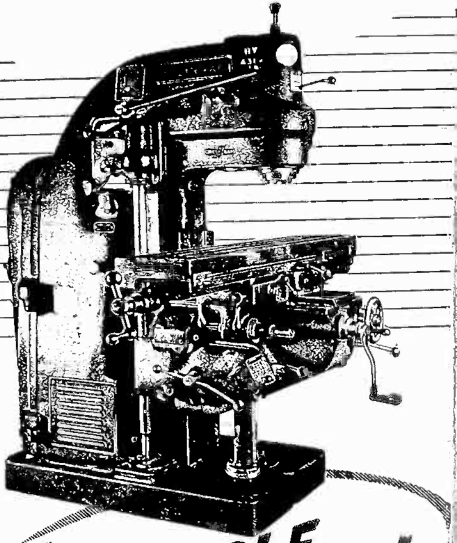

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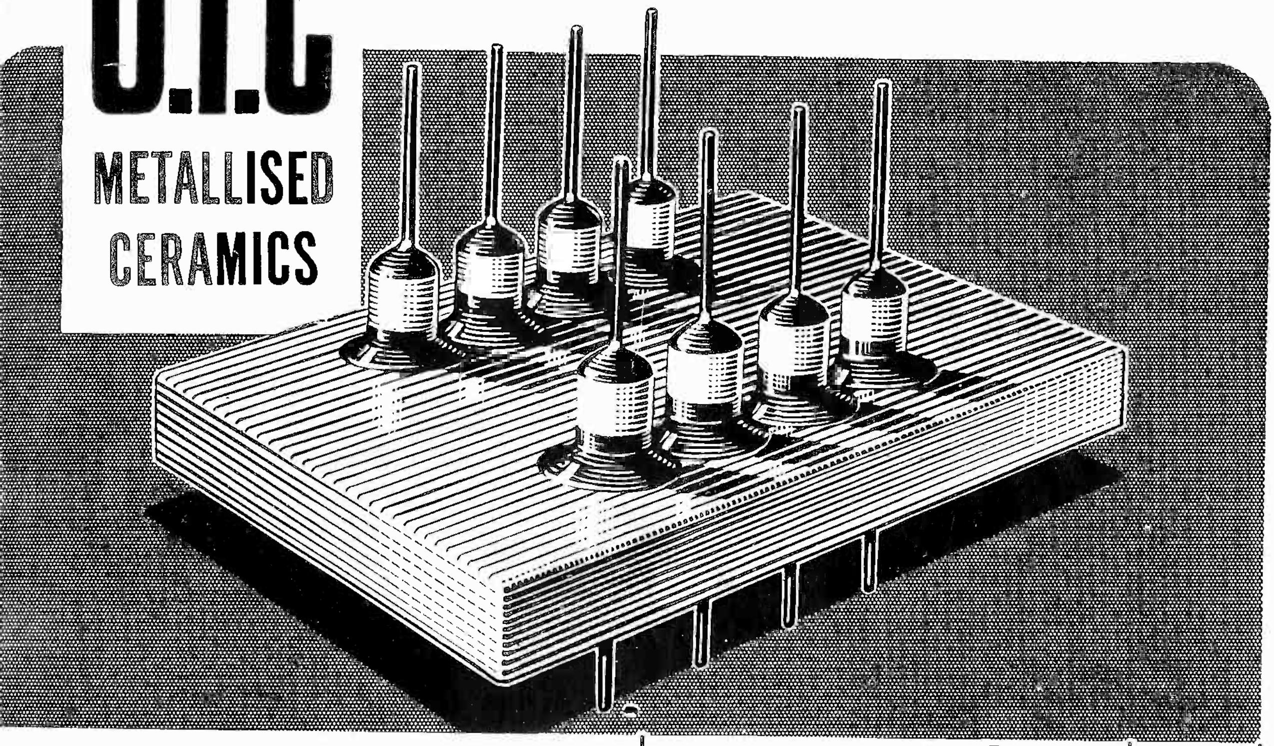

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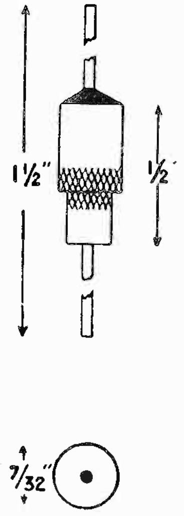

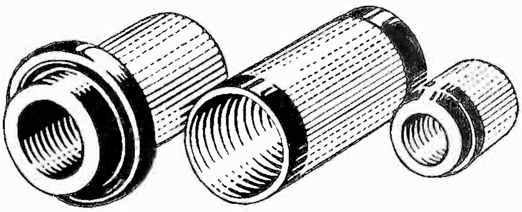

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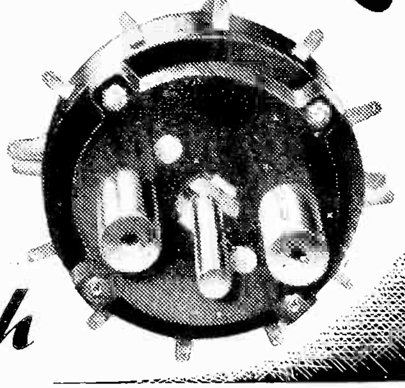

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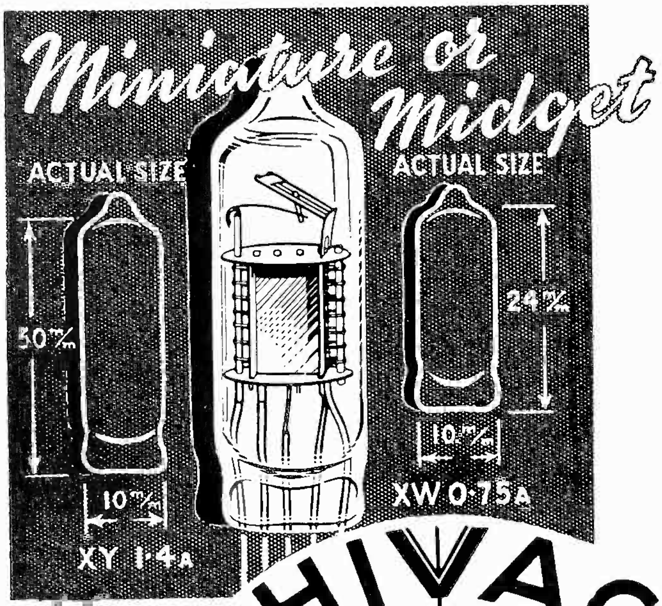
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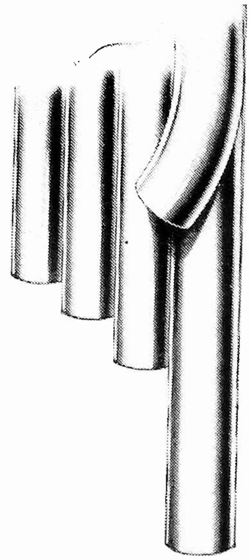
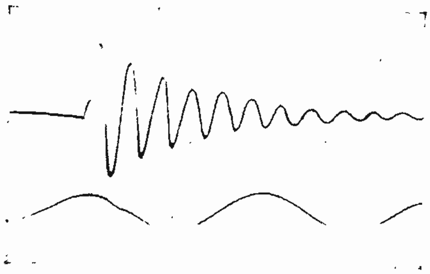
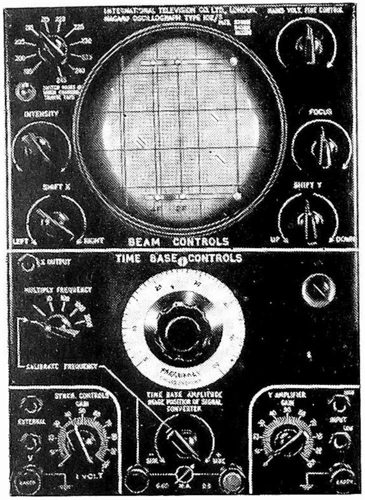
50m/m
10m/m
XY 1·4A
ACTUAL SIZE
24m/m
10m/m
XW 0·75A



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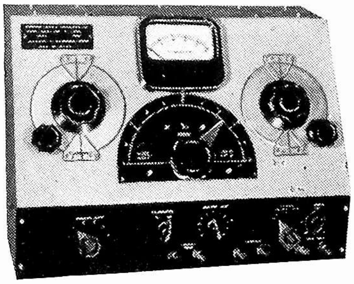

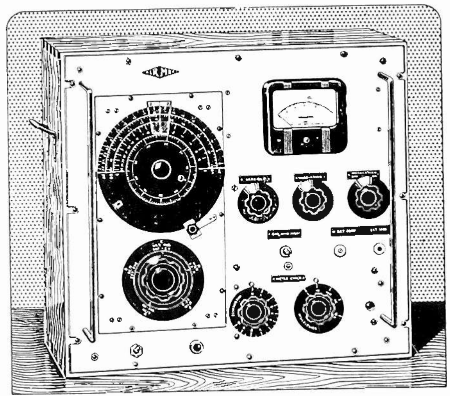

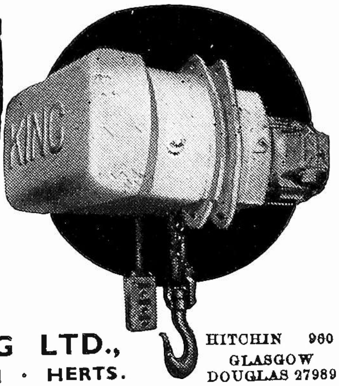

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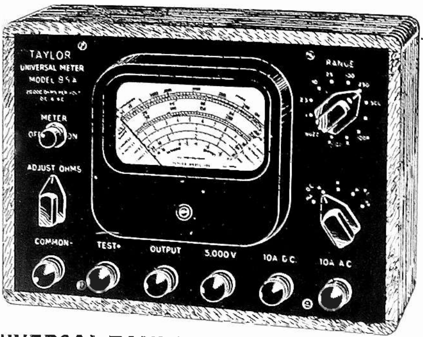

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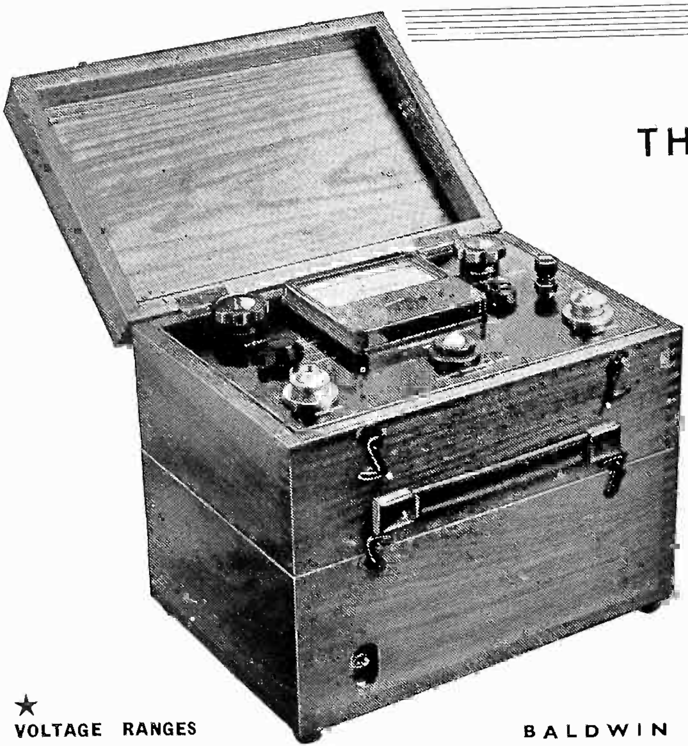

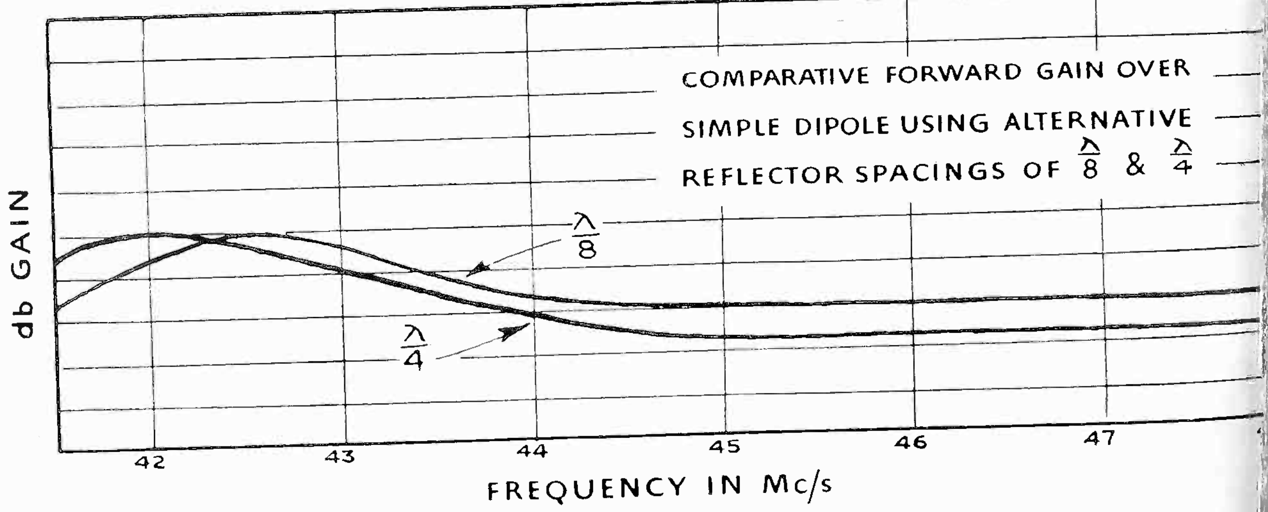
COMPARATIVE FORWARD GAIN OVER SIMPLE DIPOLE USING ALTERNATIVE REFLECTOR SPACINGS OF λ/8 & λ/4
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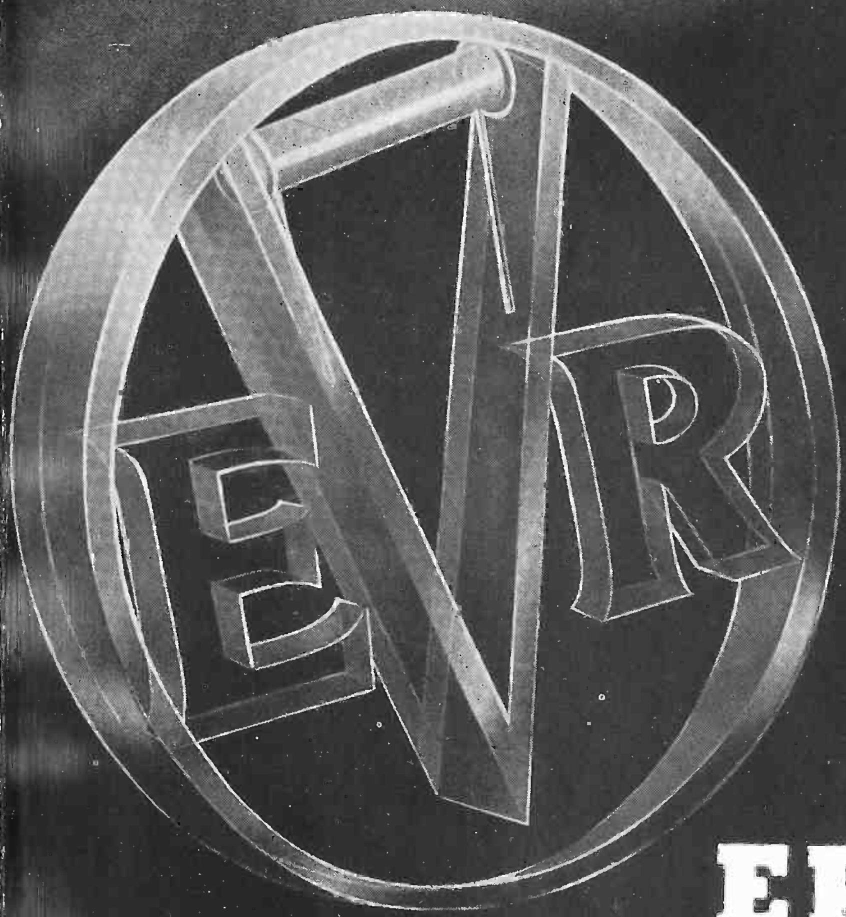
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WIRELESS ENGINEER

Vol. XXIV. JUNE 1947 No. 285

EDITORIAL

The Experimental Basis of Electromagnetism

AN interesting article with this title, by Norman R. Campbell and L. Hartshorn, was published in the November number of the *Proceedings of the Physical Society*. It is stated to be the first of several sections and it is limited to the consideration of the direct-current circuit. Its object is to develop the working principles of electromagnetism from real experimental facts and not from imaginary experiments such as those with unit poles and point charges, which are " either quite impracticable or incapable of being performed with an accuracy that would be regarded as significant to-day."

The authors differentiate between the familiar definitions, such as that involving the work done in moving unit charges, upon which the mathematical reasoning is based, and the definitions of the magnitudes involved in experimental science, which are capable of measurement to a far higher degree of accuracy. They start with ammeters and voltmeters rather than with " the traditional amber, catskin and lodestone " which are " far less amenable to precise observation." By means of ammeters with blank scales in suitable circuits they verify the addition and subtraction of currents and Kirchhoff's first law, the measurement of resistance by substitution, and the fact that resistance is independent of current, all without any reference to electromotive force or potential difference.

Conductance is also considered before any mention is made of " voltage," a term that the authors appear to prefer to the more expressive " potential difference." Fortunately they stick to the expressive terms " current " and " resistance " and do not introduce " amperage " and " ohmage." By means of voltmeters with blank scales and suitable circuits they then verify the law of addition and subtraction and pass to a consideration of Ohm's law. In the section dealing with electromotive force one is surprised to read that " the well-known Weston standard cell can be regarded as a standard of either e.m.f. or voltage, *though voltage is unquestionably the more fundamental property*." The italics are not in the original but we feel that the statement merits them. We suspect that the word " unquestionably " was introduced to drown any scruples that the authors may have felt, but we may be misunderstanding what is meant by " fundamental." Potential difference is due to distribution of electric charges, and surely the electricity-moving-force, which is the *fons et origo* of this distribution, is more fundamental than the resulting potential difference. The fact that the e.m.f. is determined by measuring the potential difference produced by it does not suggest that it is any less fundamental, but rather more so.

The paper is given a somewhat archaic appearance by the retention of the accent in " ampère " when used as the name of the unit. This is not only contrary to B.S.I. usage but will not be found in books such as Silvanus Thompson's "Electricity and Magnetism," published last century. Having dropped the final "ay" from Faraday and the final "a" from Volta, one need have no scruples about dropping the accent from Ampère; it is a minor mutilation. In wireman's slang the process is carried a step further and the ampere becomes an "amp".

In subsequent sections it is proposed to deal with varying currents and voltages, inductance and capacitance, the principles of magnetism and electrostatics, and electromagnetic waves. The authors say that they hope to remove from electrostatics and magnetism many of the difficulties that frequently disturb the experimentalist. In conclusion they say, "To the mathematician there are many possible ways of developing the relations of electromagnetism, and they are all exactly equivalent in logic and differ only in elegance. The experimentalist has less latitude; his practice is dictated by stubborn facts that defy representation in equations. The one course he is compelled by circumstances to follow is more important to him than the more elegant courses he might have followed in a better world. This is the excuse we offer to those who find our treatment clumsy".

It is not only the experimentalist who is disturbed by the difficulties in electrostatics and magnetism, and we look forward with interest to the subsequent sections of this paper. G. W. O. H.

DOPPLER EFFECT IN PROPAGATION*

By H. V. Griffiths

(Engineer-in-Charge, B.B.C. Receiving Station)

THE Doppler effect, whereby a change of frequency (and wavelength) of a radiation may occur if the relative positions of source and observer do not remain constant, is a phenomenon which is well known in optics and acoustics. The effect can be used to calculate the velocity of a moving emitter of radiations or of a moving body which is reflecting radiations from a relatively stationary source, the reflected radiation being received by an observer who may be either at rest or in motion with a known velocity.

The Doppler effect has been observed with radio waves which have been reflected from meteors [1] and from rocket projectiles such as the German V2†, on occasions when a relatively "direct" ray was also present. The Doppler effect on the reflected ray caused a detectable heterodyne or difference frequency between the waves arriving by the different paths.

Thus, the above observations—together with optical and astronomical observations and familiar acoustic examples—present examples of the effect in which source and observer are in relative motion. A moving reflector, such as the V2 projectile, presents a case where a twofold effect is produced as, in the primary consideration, the reflector is a moving observer and then "retransmits" the wave as a moving source.

In this respect, it is interesting to note that the equations for moving source and moving observer are not, as might at first be supposed, identical, although they will give approximately identical results if the velocity of relative motion is small compared with the wave-velocity.

A less familiar example of Doppler effect, which is quite commonly noted in long-distance radio transmission, is produced by movement of the reflecting "layer" in the ionosphere, when indirect-ray transmission is involved.

At Tatsfield, in November 1943, the standard frequency transmissions of Washington WWV, which transmits a 15-Mc/s carrier frequency with an error not greater than 1×10^{-8} reference absolute frequency, was commonly received at around 1,600 G.M.T. with an error of $2 - 7$ parts in 10^8, the average being about -3×10^{-8} reference absolute frequency (i.e., $-$ 0.5 c/s at 15 Mc/s.)

The following notes discuss the relative

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

† At the B.B.C. Receiving Station at Tatsfield and elsewhere.

mechanics and trigonometry in simple form, as it is believed that published discussions on this subject are not usually accessible easily to radio engineers and that the information needful has not previously been assembled in one paper.

The Doppler Effect

(*a*) *Source in Motion ;*

A source of radiation, of wave-velocity c and frequency f_s has wavelength λ_s, so that $\lambda_s = \frac{c}{f_s}$ and the observed frequency f_0 when source and observer are relatively stationary will be equal to f_s, assuming a homogeneous medium of propagation.

At the end of the first cycle of radiation, the initial positive peak of the wave will have travelled a distance λ away from the source. But if the source be in motion with velocity v, say towards the observer, then in this period of 1 cycle duration, the source will have moved a distance $\frac{v}{c}\lambda_s$ and the wavelength as measured by the stationary observer will have been decreased by this amount ; i.e., to $\lambda_s(1 - v/c)$.

The number of waves passing the observer each second (i.e., f_0) will have increased,

$$\text{as } f_0 = \frac{c}{\lambda_0} = \frac{c}{\lambda_s\left(1 - \frac{v}{c}\right)}$$

$$\text{or } f_0 = \frac{f_s}{1 - \frac{v}{c}} \quad \ldots \quad \ldots \quad \ldots \quad (1)$$

The negative sign in the denominator will apply if the relative motion of the source is toward the observer. If the direction of v be reversed, the sign will be reversed.

(*b*) *Observer in Motion :*

Consider the conditions at the head of the previous paragraph when the source and observer were both at rest. Now imagine that the observer begins to move at a velocity v per second toward the source. More waves per second will be received, equal to the number of wavelengths in the distance covered per second ; i.e., an increase of $\frac{v}{\lambda_s}$.

Hence, the observed frequency f_0 will now be $f_s + \frac{v}{\lambda_s} = f_s + f_s\frac{v}{c} = f_s\left(1 + \frac{v}{c}\right) \ldots \quad (2)$

The positive sign in the brackets applies to motion towards the source and the sign will be reversed if the direction of motion is reversed.

Now, it will be seen that the two equations for moving source and moving observer, although similar, are not identical. But, if the velocity v is small compared with the wave velocity c, then higher powers of $\frac{v}{c}$ can be neglected. In this case, the two equations will be approximately equal and also f_d, the difference frequency between f_s and f_0, will be approximately equal to

$$f_d = f_s \cdot \frac{v}{c} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (3)$$

whether the relative motion is from source or observer.

Equations (1) and (2), as shown, differ, although they may be equivalent to a first approximation in the limiting case when v is small compared with c. The reason why the cases differ depending upon whether observer or source is in motion, is at first sight difficult to imagine.

Some assistance in visualizing the difference may be obtained by imagining that, if the source is moving, a wave crest ejected at a particular instant of time has a distance to travel which is " fixed," in that the distance is not changing during the time of transit of the wave crest although the source which ejected it is in motion. On the other hand, in the case of the crest ejected from a stationary source which is to be received by an observer in motion, the distance to be travelled by the wave is constantly changing during its transit. Relative to the wave direction, there is also a difference in the sign of the direction of motion.

In the case of the moving source, the crest emitted at time $t = 0$ will arrive at time $t = \frac{d}{c}$ and a crest emitted at time $t = 1$ (one second later) will arrive at time $t = 1 + \frac{d - v}{c}$. So if the source emits f_s waves per second these f_s waves will be received in $1 + \frac{d - v}{c} - \frac{d}{c}$ seconds $= 1 - \frac{v}{c}$ seconds, and the received frequency will be $f_0 = \frac{f_s}{1 - \frac{v}{c}}$ which is equation (1). But, in the case of the observer being in motion, the wave crest ejected at one instant of time will not have a fixed distance d to travel. A crest emitted at time $t = 0$ arrives at time $t_1 = \frac{d_1}{c}$ where d_1 represents the distance between source and observer at the time-instant t_1. If the distance at time $t = 0$ was d_0, then $d_1 = d_0 - v\,t_1 = d_0 - vd_1/c$

Thus $d_1 = \dfrac{d_0}{1 + \frac{v}{c}}$ and $t_1 = \dfrac{d_0}{c + v}$

A crest ejected at time $t = 1$ (i.e., one second later) will be received at time $t_2 = 1 + \frac{d_2}{c}$, where d_2 represents the distance at the time-instant t_2; then $d_2 = d_0 - vt_2 =$

$$d_0 - v\left(1 + \frac{d_2}{c}\right) \text{ and } d_2 + \frac{vd_2}{c} = d_0 - v\,;$$

so $d_2 = \dfrac{d_0 - v}{1 + \frac{v}{c}}$ and $t_2 = \dfrac{d_0 - v}{c + v}$

Therefore, the number of crests ejected in one second is received in time

$$t_2 - t_1 = 1 + \frac{d_0 - v}{c + v} - \frac{d_0}{c + v} = 1 - \frac{v}{c + v}$$

and so f_0 (the received frequency)

$$= \frac{f_s}{1 - \dfrac{v}{c + v}} = f_s\left(1 + \frac{v}{c}\right)$$

which is Equation (2).

Having demonstrated and explained the difference between the two cases of moving source and moving observer, it is interesting to consider what will happen when the velocity of motion in either case becomes almost equal to the wave-velocity c. It will be evident, by substituting $v = c$ in Equations (1) and (2), that when the observer moves toward the source the received frequency will be *twice* the transmitted frequency, whereas with the observer stationary and the source in motion toward him, then the received frequency will tend toward infinity.

Thus, the simplified equation (3) which is a safe approximation to use in either case when v is small compared with c will be increasingly inaccurate as v becomes comparable with c. Fortunately, this approximation can safely be used when the wave under consideration is an electromagnetic or light wave, but with waves of low velocity (e.g., sound) the difference between the two conditions could not be neglected.

(c) *Reflection by a moving object*

In the case of radiations from a fixed source, which are reflected by an object in relative motion, the reflector can be considered first as a moving observer, and then as a moving source re-transmitting the "first observed" frequency. The Doppler effect will thus change the source frequency in two steps and both equations (1) and (2) will operate together.

Calling the final observed frequency still f_0, then

$$f_0 = f_s \frac{\left(1 \pm \frac{v}{c}\right)}{1 \pm \frac{v}{c}} = f_s\left(1 \pm \frac{v}{c}\right) \cdot \frac{c}{c \mp v} \quad \ldots \quad \ldots \quad (4)$$

So $f_0 = f_s \dfrac{c + v}{c - v}$ if the relative motion is toward the final observer, the signs being reversed if the motion is in the opposite direction.

The difference frequency, $f_d = f_s \dfrac{c + v}{c - v} - f_s$

but, when $v \ll c$, this can be approximated to

$$f_d = \frac{2v}{c} \cdot f_s \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (5)$$

With this approximation,

$$v = \frac{f_d\, c}{2\, f_s} = \frac{\lambda_s}{2} \cdot f_d \quad \ldots \quad \ldots \quad \ldots \quad (6)$$

and this is the equation for velocity of a moving reflector body, such as a meteor or V2, which is considered to be directly approaching the observer. The same approximation in terms of f_d will apply if the motion be reversed.

Washington Standard Frequency Transmissions

The mean observed f_d of the 15-Mc/s signal at 16.00 G.M.T. in December 1945, was minus 0.5 c/s, by measurement, whereas the transmitted frequency was 15 Mc/s exactly.* A close approximation of what this change of frequency represents in terms of change of path length can be obtained if the geometry of the path taken by the Washington signal can be calculated. Furthermore, it will be possible to relate this measurement to the path in such a way as to determine the rate of change of virtual height of the reflecting layer. Before ob-

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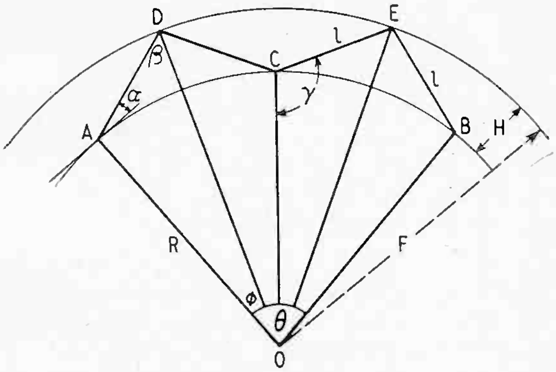
* The possible error limits quoted by Washington are $\pm\, 2 \times 10^{-8}$ but this transmission error has been neglected here.

taining this information from the measured value of f_d, the trigonometry of the reflected ray path between Tatsfield and Washington must first be considered.

Propagation Trigonometry

In the figure, A and B are the points of transmission and reception lying along an arc of the great circle ACB. The angle θ is subtended by the points A and B at the earth's centre, R being the earth's radius. The outer arc joining points D and E is in the ionosphere at the virtual height of reflection, which must be taken as being at a mean distance F from the centre of the earth, although it will not actually be constant either in time or in space, because the ionic density will vary over a long path of this nature, particularly if there is a large difference of longitude between D and E.



The path length L will be the sum of the distances AD, DC, CE, EB, etc., which will vary in number and position with the number of hops (reflections) which the signal undergoes, a two-hop path being illustrated. Thus, there are $2n$ identical triangles involved with a signal taking n hops to reach the observer, each (plane) triangle having an angle at O equal to $\frac{\theta}{2n}$

The radiation and reception angle α is subtended with the tangent line which is at right angles to R. Thus, the angle DAO $= (90 + \alpha)^\circ$ and the angle $ADO =$

$$\left(90 - \alpha - \frac{\theta}{2n}\right)^\circ$$

If H is the mean height of virtual reflection above the earth's surface, then

$$H = F - R = R\,\frac{\sin\,(90 - \alpha)^\circ}{\sin\left(90 - \alpha - \frac{\theta}{2n}\right)^\circ} - R = R\left[\frac{\sin\,(90 - \alpha)^\circ}{\sin\left(90 - \alpha - \frac{\theta}{2n}\right)^\circ} - 1\right] \quad \ldots \quad \ldots \quad (7)$$

Similarly, if the total path length $= L$ and the length of one limb of the path $= l$, then the distance $AD = l = \frac{L}{2n}$

$$\text{Hence } \frac{L}{2n} = R\,\frac{\sin\frac{\theta}{2n}}{\sin\left(90 - \alpha - \frac{\theta}{2n}\right)^\circ} \quad \ldots \quad (8)$$

$$\text{and} \quad L = 2nR\left[\frac{\sin\frac{\theta}{2n}}{\sin\left(90 - \alpha - \frac{\theta}{2n}\right)}\right]$$

Now, in the figure there are six dimensions, of which three must be known in order to solve the others. R and θ are known from geodesy, H can be measured for vertical incidence or for other angles with greater difficulty, by radio-sondage technique. Measurements of the vertical angle of reception (α) can be obtained most easily with a pulsed signal, such measurements having been carried out in India on special B.B.C. pulse transmissions sent out for that purpose in February, 1942. Chamanlal, in India, has also carried out vertical angle measurements of an ordinary carrier signal, using spaced horizontal dipoles, in 1940-42 and 1943[2, 3].

The probable degree of error of measurement of both vertical angle and virtual height of reflection is not negligibly small, although the angle measurements with a pulsed signal are fairly accurate and the number of hops taken by the predominant signal can be determined with some certainty in this case. With continuous carrier signals errors may be larger. Virtual height measurements will have some error compared with the true height—even when measured at vertical incidence, because the retardation of the wave in the layer, and the penetration cannot be estimated accurately. We may, however, find it profitable to consider the example of the Doppler effect given earlier, in conjunction with the figure showing propagation path details, which may be combined with hypothetical or calculated values of n, H and α in worked examples, the results from which should be of the right order of magnitude even if they may not be strictly accurate.

To solve the geometry of the propagation path, we shall need to have a measured value of either H or α, as stated above, α being the more convenient because the number of hops taken by the signal will be immediately obvious from the measured value of α. Unfortunately, measurements of α—although possible (e.g., by using spaced-horizontal dipoles connected to the X and Y plates of a cathode-ray tube)—are liable to considerable error. On the other hand, the measured value of virtual height was available for the period in question, so the problem is here presented in the form where H, θ, and of course R, are the known elements.

According to Bennington, the minimum height at 16 00 G.M.T. in November 1945, was approximately 230 km on the western half of the path and 250 km on the eastern half, to which must be added about 30 km for penetration, to obtain virtual height H. In the diagram, H is shown as constant and this simplification will be continued by taking the mean value of $H = 270$ km.

For the solution of this oblique triangle, the equations (7) and (8) already considered are not suitable because the value of α is not known, but other trigonometrical ratios which can be used instead are quite as simply derived.

Putting the angles $\frac{\theta}{2n}$ or $AOD = \phi$, $DAO = \gamma$, and $ADO = \beta$, then there are two methods of solving for γ and β and for l which is the length of the limb between A and D. In the first method, which is the more convenient :

$$\tfrac{1}{2}(\gamma + \beta) = 90^\circ - \tfrac{1}{2}\phi \quad \ldots \quad \ldots \quad (9a)$$

$$\tan \tfrac{1}{2}(\gamma - \beta) = \frac{F - R}{F + R} \cdot \tan \tfrac{1}{2}(\gamma + \beta) \quad (9b)$$

Having solved (9a) and, from it, (9b), then

$$l = F + R \, . \frac{\cos \frac{1}{2}(\gamma + \beta)}{\cos \frac{1}{2}(\gamma - \beta)} \quad \ldots \quad (10)$$

follows logically.

In the second method :

$$l = R^2 + F^2 - 2\,RF \cos \phi \quad \ldots \quad (11)$$

and from the equation

$$l = R \frac{\sin \phi}{\sin \beta}$$

then $\sin \beta = \frac{R \sin \phi}{l}$ .. .. (12)

Putting $H = 270$ km, then $R = 6{,}370$ km and $F = 6{,}640$ km, and $\phi = \frac{35}{} = 13^\circ\,15'$, *for two hops.*

Solving Equations (9)

$$\tfrac{1}{2}(\gamma + \beta) = 90^\circ - 6^\circ\,37\tfrac{1}{2}' = 83^\circ\,22\tfrac{1}{2}'$$

$$\tan \tfrac{1}{2}(\gamma - \beta) = \frac{6{,}640 - 6{,}370}{6{,}640 + 6{,}370} \tan 83^\circ\,22\tfrac{1}{2}'$$

$$= 0.02075 \times 8.54$$

$$= 0.177$$

$$\tfrac{1}{2}(\gamma - \beta) = 10^\circ\,02'$$

$$\tfrac{1}{2}(\gamma + \beta) = 83^\circ\,22\tfrac{1}{2}'$$

then $\gamma = 93^\circ\,24\tfrac{1}{2}'$ $\therefore \alpha = 3^\circ\,24\tfrac{1}{2}'$

and $\beta = 73^\circ\,14\tfrac{1}{2}'$

And Equation (10)

$$l = 6{,}640 + 6{,}370 \frac{\cos 83^\circ\,22\tfrac{1}{2}'}{\cos 10^\circ\,02'}$$

$$= 13{,}010 \times 0.1172$$

$$= \text{antilog } 3.183 = 1{,}324 \text{ km}$$

$$\therefore L = 4l = 6{,}096 \text{ km.}$$

So, assuming that the Washington signal had only two reflections, the angle of arrival of the wave α

$$\alpha = \gamma - 90^\circ = 93^\circ\,24\tfrac{1}{2}' - 90^\circ = 3^\circ\,24\tfrac{1}{2}'$$

If the Washington signal were received by a path having *three reflections*, the corresponding solutions of equations (9) and (10) would become :

$y = 100^\circ\,37\tfrac{1}{2}'$ $\therefore \alpha = 10^\circ\,37\tfrac{1}{2}'$

$\beta = 70^\circ\,32\tfrac{1}{2}'$

$\phi = 8^\circ\,50'$

$l = 1{,}038$ km $\therefore L = 6l = 6{,}228$ km.

Now, reconsidering the Doppler effect on the Washington 1-Mc/s signal, it is obvious that, from Equations (3) or (6), we can calculate the change of path length per second which will produce the observed difference in frequency between the Washington signal as transmitted and as received.

From (3) $f_d = f_s \cdot \frac{v}{c}$, so $v = c\frac{f_d}{f_s} = \lambda f_d$, when λ = wavelength of f_s in metres. .. (15)

In the case of two hops f_d will be one-quarter of the total observed difference frequency because we are considering what happens over the path length l, which is one-half of one hop. Similarly, if the signal were propagated over a three-hop path, then f_d would be one-sixth of the total observed f_d.

Thus the change of path length $v = \lambda f_d =$

$\frac{0.5}{4} \times 20 = 2.5$ metres/second for the 1,524 km (two hops) limb,

$\frac{0.5}{6} \times 20 = 1.6$ metres/second for the 1038 km (three hops) limb.

These two different velocities, 2.5 and 1.6 metres/second represent the change of length of the limb l, in the conditions of propagation considered, this change being caused by the increase in the virtual height of reflection during the time under review. They do not, however, represent the actual rate of change of the virtual reflection-height itself. To obtain this figure from the rate of change of l, the ratio $\dfrac{\text{rate of change of } H}{\text{rate of change of } l}$ must be calculated, this ratio being itself a variable, dependent upon the variation of the angles γ or β. Rather complex calculations would be required in order to obtain the exact ratio, in the circumstances of several dependent variables being present, the value of f_d and thus of the rate of change of l being, in reality, dependent upon the primary variable, which is the rate of change of H. However, a fair approximation of the ratio can be obtained by inspection; the ratio $\dfrac{\Delta H}{\Delta l}$ is found to be approximately 2 in the three-hop case and between 2 and 3 in the two-hop case. Applying this ratio, we obtain values for the rate of change of the virtual height of reflection H of *between* 3.2 *and about* 6 *metres per second*, due to the increase in ionic density at about 16.00 G.M.T. in December 1945, calculated from the observed signals on 15 Mc/s. Considerably greater changes of observed frequency have been measured on some occasions: the difference taken in the above example has frequently been exceeded, and the difference-frequency has progressed through zero to the opposite sign in daylight and darkness periods, corresponding with changes in ionic density. For this reason, the assumption made above, that the measured divergence of WWV was not wholly or largely due to differences in frequency standards, is not unreasonable. To confirm this point, simultaneous measurements have been carried out recently, on frequency bands in which propagation paths and conditions were dissimilar. Variations in the percentage divergence so measured were of the expected order of magnitude and of sign.

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NEW BOOKS

Introduction to Electron Optics.

By V. E. COSSLETT. Pp. 272+x. Oxford University Press, Amen House, Warwick Square, London, E.C.4. Price 20s.

This book is the outcome of a series of lectures given to students taking an Honours Course in Physics at the University of Oxford.

In the past books on this subject have been either too compendious, or else too limited in size or treatment. Moreover the progress has been so rapid that a book has been out of date almost before it was published. In recent years, however, the subject has shown signs of consolidation, only the practical applications advancing at any great speed. Thus the publication of this book is timely, and due to the concise style of writing, its size is sufficient.

After the introduction there is a chapter on the electrostatic field. Various methods of finding the field shape both theoretically and practically are discussed. Mention is made of the relaxation method of calculation, also of the usual rubber sheet and electrolytic-tank techniques.

There follow chapters on electrostatic and magnetic focusing and one on aberrations of the image. The treatment here is largely trigonometrical, but the logical exposition of the subject is not spoilt by a maze of mathematics, the Hamiltonian method being relegated to the Appendix.

The emission of electrons is covered in 23 pages. Starting with energy states in metals, thermionic-, field- and photo-emission are dealt with, and the chapter concludes with a description of some Image Tubes and Multipliers.

The cathode-ray tube and its derivatives (the Iconoscope, Orthicon, etc.) are grouped together.

A chapter on electron diffraction and the electron microscope has some very fine diagrams and photographs. Under the title "Other Applications: Cylindrical Fields," the broad principles of operation of a number of devices are described, including the Magnetron, Cyclotron, Betatron, and Beam Power Valve. A final chapter covers velocity-modulated electron beams and their applications.

The bibliography is in the form of "Further Reading" at the end of each chapter, and the references given are rather more international than in some previous works on this subject. There is a list of symbols.

Altogether a most excellent book, readable, and useful alike to students and to technicians who require a book of reference.

B. C. F-W.

R.C.A. Technical Papers Index. Vols. I & II (a)

Pp. 143 + vi and 21 + iv. Obtainable free on application to Radio Corporation of America, R.C.A. Laboratories Division, Princeton, N.J., U.S.A.

In this index to papers by R.C.A. authors the arrangement is first, in chronological order; secondly, in alphabetical order; and there are, in addition, author and subject indexes. Vol. I covers the years 1919-1945 and Vol. II (a) deals with 1946.

Principles of Radio for Operators.

By RALPH ATHERTON, M.S. Pp. 344 + x. Published by Macmillan & Co. Ltd., St. Martin's Street, London, W.C.2. Price 20s.

VERY-WIDE BAND RADIO-FREQUENCY TRANSFORMERS*

By D. Maurice, Ing., E.S.E. (Paris), A.M.I.E.E. and R. H. Minns, A.M.I.E.E.

(British Broadcasting Corporation)

SUMMARY.—The salient features of such transformers are discussed, and as far as possible, quantitative expressions are given which enable the best use to be made of them.

THIS paper attempts to give sufficient information on the design of low-power very-wide band radio-frequency transformers for engineers at present unfamiliar with the necessary technique to acquire a working knowledge of it and to be in a position to initiate transformer designs.

CONTENTS

Introduction

The technique of wide-band r.f. transformer design has advanced considerably in the past few years due largely to the stimulant of war-time requirements. Bandwidths of the order of 200 to 1 are now quite normal requirements (150 kc/s to 30 Mc/s). This width covers the normal range of broadcast frequencies. Recently the upper frequency limit has been raised to cover application to the very high frequencies. A recent design transforms a balanced 100-ohm circuit into an unbalanced 100-ohm circuit at all frequencies between 100 kc/s and 150 Mc/s with a loss of less than 2 db.

The wide variety of applications is indicated by the following few examples :—

800-ohm rhombic aerial to coaxial 100-ohm feeder for a broadcast monitoring service.

100-ohm balanced feeder to 3.33-ohm bus-bars for simultaneous distribution of radio frequencies to a large number of receivers. 100 kc/s to 25 Mc/s. This also forms part of a monitoring service.

Coupling the anode of a KT.44 tetrode to an 80-ohm line and obtaining and feeding 5 watts into the line over the frequency range 200 kc/s to 1,500 kc/s. This is part of the drive equipment of a variable-frequency transmitter.

100 ohms to 3,000 ohms, 40 Mc/s to 60 Mc/s. This was a receiver input transformer used instead of a bandpass-coupled circuit. If carefully designed, the transformer gives greater coupling efficiency than ordinary coupled circuits due to various practical considerations.

100 ohms unbalanced to 500 ohms, unbalanced, covering the range 50 kc/s to 27 Mc/s. This transformer couples a 100-ohm coaxial cable to the aerial and earth terminals of an all-wave receiver. Many such transformers are in service at monitoring centres.

An admittance-measurement bridge-transformer covering the frequency range 100 kc/s to 60 Mc/s.

All these examples deal only with low powers. The application of transformer design principles to high-power equipment remains a development problem, which it is considered, will present no great fundamental difficulties other than the question of insulation breakdowns due to high potential gradients per turn.

2. General

A transformer may be regarded as a device for :—

(*a*) Connecting together impedances of different values.

(*b*) Converting circuits which are unsymmetrical with respect to earth to circuits which are so symmetrical.

(*c*) Isolating circuits which carry direct current from circuits which do not.

(*d*) Obtaining accurate voltage and current ratios.

Most of the above qualities are possessed

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* MS. accepted by the Editor, October 1946.

by ordinary coupled circuits, but transformers differ from these in that their frequency pass range is much greater than that of the coupled circuits. It is this fact which makes transformer theory so very different from coupled-circuit theory even though transformers are, of course, a special case of coupled circuits. In fact, a transformer might be called a coupled circuit in which the coupling factor is practically unity and having such a wide pass-band that the per-performance at each end of the range is determined by entirely different physical and electrical constants. As it is this performance which we wish to control we shall sub-divide this paper into two sections. We shall deal first with the factors affecting performance at the lower-frequency limit of the pass range and secondly with the upper-frequency limit.

A simple two-winding transformer, Fig. 1 (*a*), may be represented in greater detail by (*b*), in which R_s and L_s are the resistances and leakage inductances of the windings referred to primary and secondary sides. R_p and L_p are the shunt resistance and shunt inductance referred to the primary side and due to the presence of the core ; C_1 and C_2 are the shunt capacitances due to the distributed inter-turn capacitances and any overall capacitance across input and output.

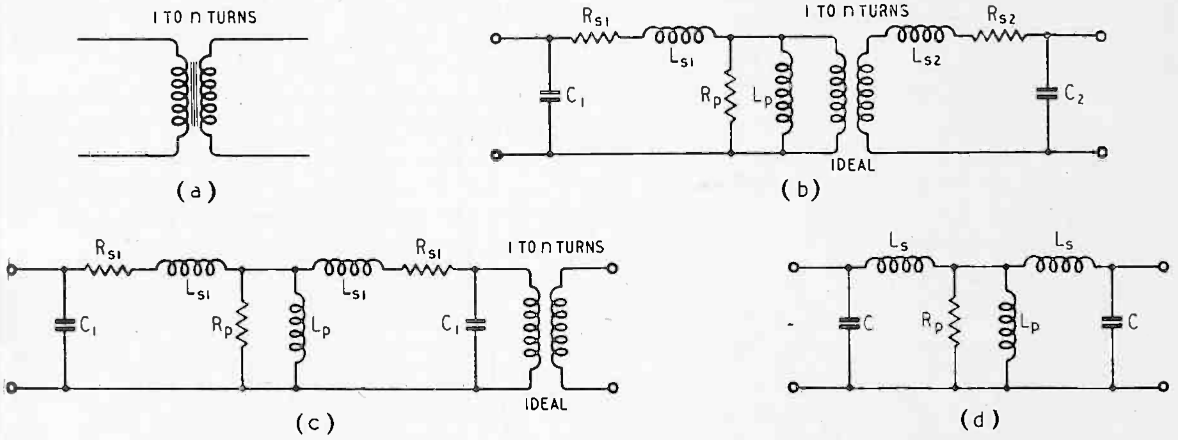


Fig. 1. *A transformer* (*a*) *has the equivalent circuit* (*b*) *which can be reduced to the form* (*c*). *When the effect of* R_s *is negligible a further simplification to* (*d*) *can be made.*

The purpose of the ideal transformer is evident. It might be added that an ideal transformer is one without leakage and shunt admittance. Fig. 1 (*b*) changes to the form (*c*) if we move the ideal transformer to the right, and assume that the windings are sufficiently similar for the secondary series elements, when referred to the primary side, to be equal to the primary series elements.

In all the applications with which this paper deals the resistance R_s is negligible compared with the reactance of L_s. Fig 1 (*c*) thus simplifies to (*d*). We shall henceforth omit the ideal transformer and consider only one-to-one ratio transformers. In most applications the transformer is designed to work between pure resistances which we shall call R_0.

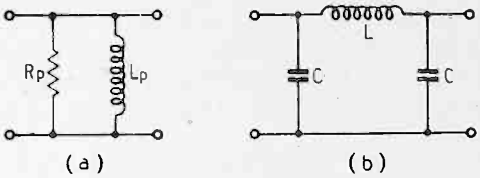


Fig. 2. *At low and high frequencies respectively the transformer equivalent circuit can be reduced to the forms* (*a*) *and* (*b*).

The lower and upper frequency limits of the transformer pass-band will be called f_1 and f_2, respectively.

Now in all wide-band transformers, defined as having an upper cut-off frequency at least five times the lower ($f_2 > 5f_1$), we find $L_s \ll L_p$. Such transformers are so designed that at f_1 the reactance of L_s and the susceptance of C are negligible compared with R_0, whilst the transformer insertion loss is caused entirely by the admittance of the parallel combination of R_p and L_p. Thus at low frequencies Fig 1 (*d*) becomes Fig. 2 (*a*). In the case of wide-band radio-frequency transformers, particularly those using metallic alloy cores, R_p and the reactance of L_p rise with increasing frequency and become such that their admittance is negligible at f_2.

As the frequency increases, however, the reactance of L_s and the susceptance of C both rise and become predominant at f_2. At f_2, therefore, Fig 1 (*d*) becomes Fig. 2 (*b*) in which $L = 2L_s$.

3. Low-frequency Parameters

3.1. *Shunt Losses.*

The transformer complete with its feed and load, considered at the frequency f_1, is pictured in Fig. 3 (*a*). It has been found in practice that for r.f. work an insertion loss of 2 db is quite permissible. By definition the insertion loss represents the difference in loss between (*a*) and (*b*) of Fig. 3. As the loss of $\frac{e_2}{e_1}$ is 6 db in Fig. 3 (*b*) and we are to admit an insertion loss of 2 db in (*a*), it is clear that the overall voltage gain in (*a*) must be $\frac{e_2}{e_1} = -8$ db. If Q be the ratio of R_p to the reactance of L_p at f_1 it is possible to obtain a relationship between the modulus, $|Z|$, of the open-circuit impedance due to R_p and L_p on the one hand and Q on the other; the entire relation is determined by the fact that the total loss of Fig. 3 (*a*) is 8 db. This expression is

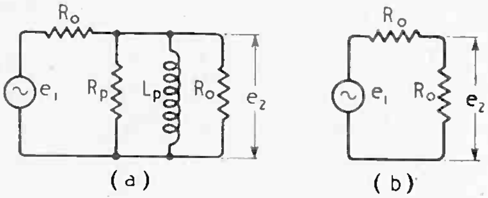


Fig. 3. *The equivalent of the transformer at low frequencies with its generator and load resistances is shown at (a), and the generator and load without the transformer at (b).*

$$|Z| = \frac{2R_0}{9\sqrt{1+Q^2}}\left(4+\sqrt{25+9Q^2}\right) \quad (1)$$

When a given core shape has been decided upon, equation (1) can be used as a specification for the maintenance of good and regular core performance. This may be conveniently done by assuming for test purposes a one-ohm to one-ohm transformer ($R_0 = 1$). The tester should measure R_p and $2\pi f_1 L_p$ and form their ratio, Q. Putting this value of Q into equation (1) gives the required value of $|Z|$. The actual open-circuit impedance per turn squared, as measured, is

$$Z = \frac{R_p}{n^2\sqrt{1+Q^2}} \quad \dots \quad \dots \quad \dots \quad (2)$$

where n is any convenient number of turns wound on the core for test purposes. The number of turns required to give the 2 db insertion loss is :

$$N = \sqrt{\frac{|Z|}{Z}} \quad \dots \quad \dots \quad \dots \quad \dots \quad (3)$$

and a suitable upper limit may be placed on this.

This assumes a 1-ohm to 1-ohm transformer. Any impedance level R_0 can be catered for simply by multiplying N by $\sqrt{R_0}$. Equation (1) is plotted in Fig. 4 for ready reference. It should be noted that the method of test allows the greatest possible latitude to the core manufacturer as it does not tie him down to specific values of shunt resistance and reactance, but only to a necessary overall combination of the two. The values of N for different core materials, but constant core shape, plotted against frequency, not only tell the designer how many turns are needed for a given load resistance and a given core material, but also enable him to choose the best core material for the particular value of f_1 that his design requires. Curves of N against frequency for various core materials are shown in Fig. 5. For example, it would be immaterial in a transformer in which the lower-frequency limit was 170 kc/s whether Rhometal or Permalloy C were used, but it would be better to use Mumetal because if, for instance, it was a 100-ohm to 100-ohm transformer then 16 turns would be necessary on a Rhometal or Permalloy C core while only 10 turns would be necessary on the Mumetal core. The lower the number of turns required, the better the core, as this is only used as a turn reducer.

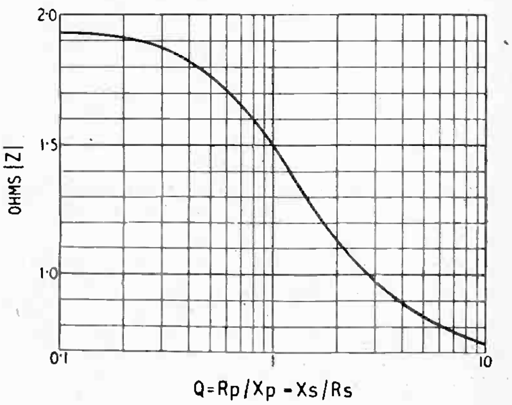


Fig. 4. *Permissible open-circuit impedance of a 1-Ω to 1-Ω transformer for an insertion loss of 2 db.*

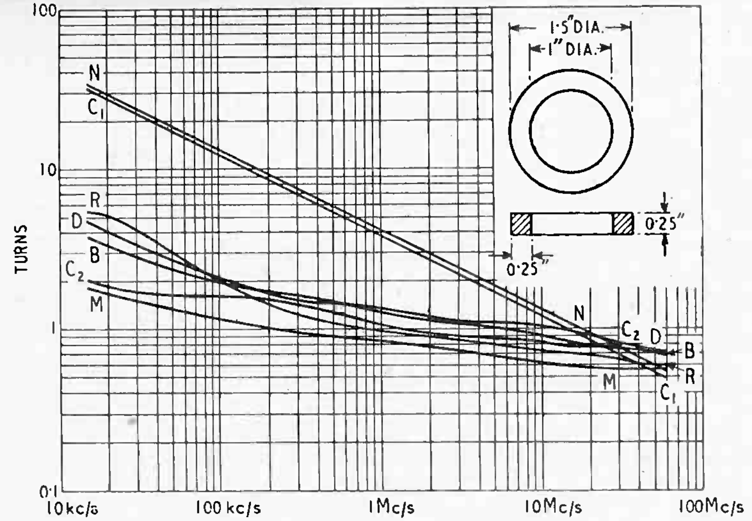


FIG. 5

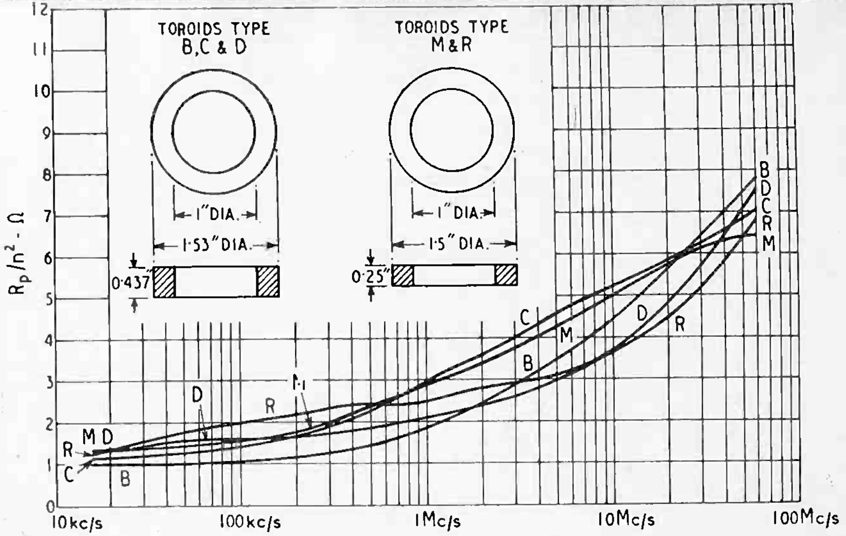


FIG. 7

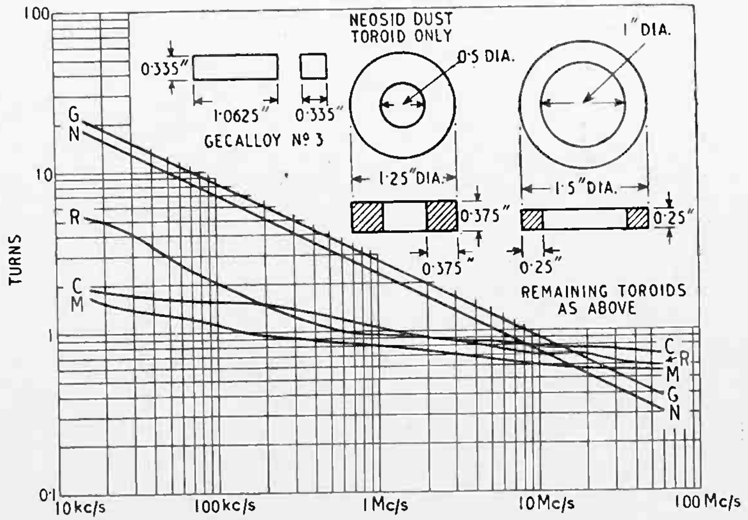


FIG. 6

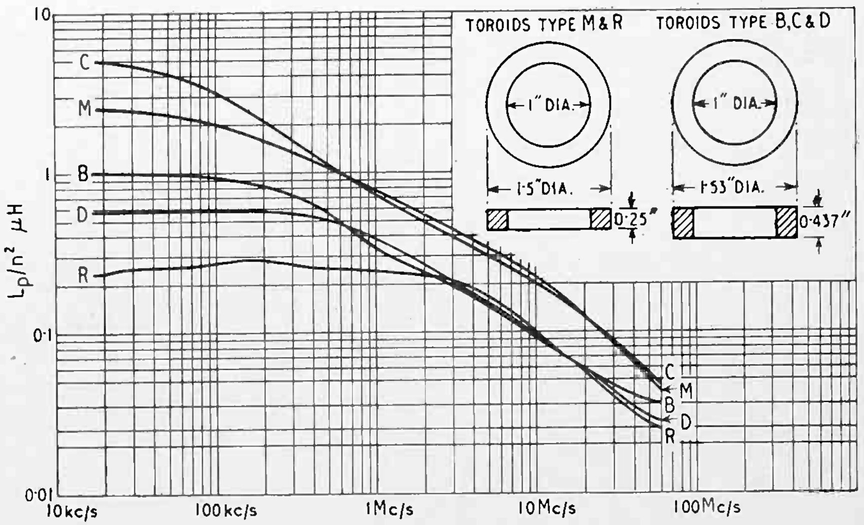


FIG. 8

Figs. 5, 6, 7 and 8. *Comparison of core materials in turns per* 1-Ω *winding for* 2-*db loss due to shunt R and X. In the following the figures in brackets refer to the tape thickness in thousandths of an inch. B, Permalloy "B"* (2) *: C*$_1$*, Dust, Grade C Post Office ; C*$_2$*, Permalloy "C"* (2) *; D, Permalloy "D"* (2) *; M, Mumetal* (1.5)*; N, Neosid Dust ; R, Rhometal* (2).

Further, as will be shown later, the leakage inductance is proportional to the square of the number of turns, and an examination of Fig. 2 (*b*) shows that the lower the leakage, L, the higher will be f_2. Fig 5 shows that for v.h.f. transformers dust cores are better than alloy-strip cores. Fig. 6 shows a similar comparison between core materials, but this time the issue is complicated by the fact that only actual marketed cores are shown. It may be observed that the usefulness of dust begins at a lower frequency than in Fig. 5. Little has so far been said about the variation of R_p and L_p with increasing frequency. In practice, though L_p falls with increasing frequency, this increase is less rapid than hyperbolic with the result that R_p and $2\pi f L_p$ actually increase with frequency, the phase angle remaining very approximately constant. Though this increase is slow it is none the less sufficient to reduce the 2-db insertion loss to a quite negligible quantity when $f > 5f_1$ approximately. There is one case in which the cause of insertion loss is of importance. This is when the transformer has to handle higher powers than is usual. A 2-db loss may still be admitted at f_1, but this should be caused in the main by L_p and not R_p as R_p is an indication of the running temperature of the core. The core shape and size dealt with in Fig. 5 will, in the case of curves M and C_2, handle about 1 watt. For powers exceeding this some other material should be chosen such as dust or Rhometal. The same type of core test as described above may be used but higher values of Q should be used than is normal for lower powers. Some curves showing the variations of shunt resistance and inductance against frequency, for various metals and core shapes are shown so that the power handling problem can be tackled more quantitatively. Fig. 7 shows resistance and Fig. 8 inductance. Dust has such a high Q that for our purpose it may be taken as purely reactive.

The measurements made for Figs. 5 to 8 show the characteristics of the cores only. This entailed the use of a rather special type of winding, as most of the cores were toroidal in shape. Consider a winding of bunched turns covering only a small proportion of the total metallic flux path, Fig. 9. The dotted lines indicate leakage flux paths. For a given current in the coil, the ratio of the core flux to the leakage flux will vary as the permeability. This in turn is found to vary more or less inversely as the frequency. If the permeability is high, practically all the flux will travel through the core, and so the inductance L_p of the coil will not vary with the amount of the core covered by the winding. When the permeability is low, as occurs at the higher frequencies, this inductance may well increase with diminishing utilization of the winding space as the inductance of the winding may in large measure be due, not to the core, but to the turns acting as an air-cored solenoid. As is well known, the inductance of such a solenoid becomes greater as the turns are bunched closer together. R_p would increase with turn bunching because it is the core which introduces the losses. In order to eliminate the effects due to different types of windings, therefore, a single turn of metal covering the entire toroid was used. This was split into two discs each having an annular groove the shape of the toroid and of depth half the width of the toroid, Fig. 10. These two discs were arranged to make contact round their outer edges and to have minimum spacing between the central portions when screwed

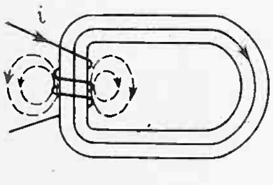


Fig. 9. *Illustrating the leakage flux paths in an iron-cored coil.*

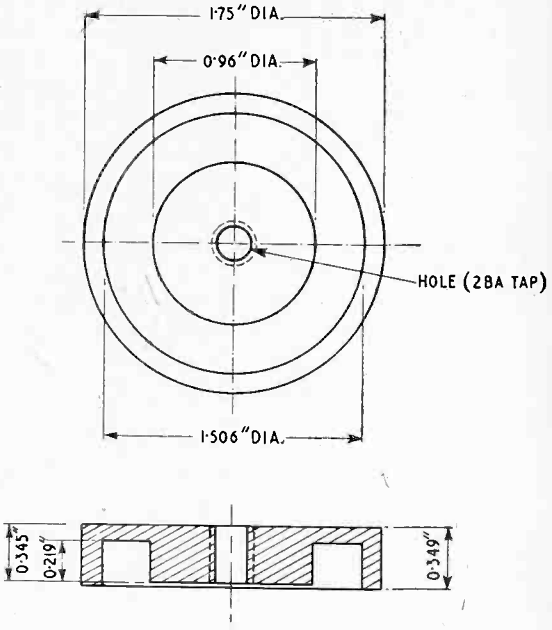


Fig. 10. *Single-turn toroid used in the measurements.*

together by a single central screw which threaded into the farther disc and was insulated from the nearer disc. Connections to this single turn were made by two tag-like washers, one of which made contact with the nearer disc while the other made contact with the under edge of the screw head, the screw being connected with the farther disc and insulated from the nearer. As the discs formed a large mass of high conductivity and unit permeability which closely fitted the toroid, there was little flux outside the core. That flux which contributes to the inductance of the coil, hence to the measurement of the permeability, has a direction tangential to the circular centre line of the core material. Any components of flux which are at right angles to this tangent do not contribute to the measured permeability and are therefore wasted. When the brass discs are present these right-angular components of flux will set up eddy currents which in turn will produce opposing fluxes, the result being an effective increase of permeability.

The above remarks show that the core characteristics are not necessarily fully used unless quite special precautions are taken, precautions which are quite impracticable when it comes to actual designs. It can now be seen that for a given core shape and material there will be a certain value of the frequency below which the permeability will be so high that the form of winding does not matter. Above that frequency, skill and experience are necessary in deciding how much of the core winding space should be covered, and if the number of turns is small, what strip width should be used for the windings. Fig. 11 shows the ratio of the open-circuit inductance of a single turn of rather fine wire to that of the single solid turn, as described above, against varying values of the latter obtained by varying the frequency of measurement. The core used for this measurement was the one marked M in Figs. 5, 6, 7 and 8. It may be observed from Fig. 11 that for inductances of the order of 0.2 μH per turn-squared there is a definite advantage in arranging for the winding to cover a large portion of the available winding space. For inductances less than 0.05 μH per turn-squared the core may be regarded electrically as if it were a short stub core. The large values attained by the ratio in Figs. 11 for low values of $\frac{L_p}{n^2}$ of the solid turn, do not indicate that the core is useless, but do show that distributing the winding round it is not always good practice.

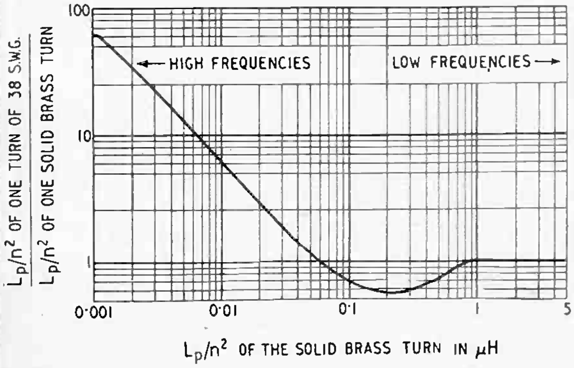


Fig. 11. *Increase in the fraction of the flux not coupling to the core as L_p/n^2 diminishes (with increasing frequency.)*

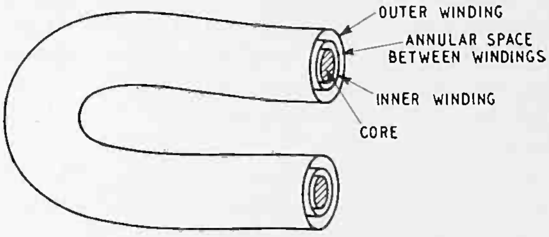


Fig. 12. *Two coaxial single-layer windings are shown.*

All the above discussion may in many cases be quite vitiated by considerations of leakage inductance which will be dealt with later. For example it may, in spite of the foregoing, be desirable to use a spiral winding of flat strip on account of the singularly low-leakage inductances obtained, and to admit of the few extra turns necessary because the winding space is not being fully utilized. It will have been observed that a good deal of the discussion has hinged around cores of toroidal shape. We shall now discuss the question of optimum shape.

3.2. *Core Shape*

Any inductance is inversely proportional to the reluctance offered by the path to the flux passing through it. Inductance is therefore proportional to the cross-sectional area and inversely proportional to the path length. Now consider two single-layer windings, one exactly overlapping the other and entirely filling the winding space of a closed magnetic path, Fig. 12.

A section of core and windings has been removed from Fig. 12 so as to show the general internal configuration. Both the shunt inductance and the leakage inductance path lengths are the same, namely the total core length, therefore the ratio of shunt to leakage inductance is independent of core length. Now the main linking flux passes through the core cross-section while the leakage flux may be regarded as that part of the total flux which links the inner winding but misses the outer winding. The area through which it passes is therefore that of the annular space between the two windings. The shunt inductance is thus proportional to the square of the core radius (cross-sectional area of core) while the leakage inductance is proportional to the cross-sectional area of the annular space between the windings, therefore proportional to the core radius. The ratio of shunt to leakage inductance is, therefore, proportional to the core radius. Assume for simplicity that we are dealing with a core of Q somewhat greater than unity so that it is the shunt inductance alone which determines f_1. We may write

$$2\pi f_1 L_p \approx R_0, \text{ the load} \quad \ldots \quad \ldots \quad (4)$$

Further, it is evident from Fig. 2 (*b*) that

$$2\pi f_2 L \approx R_0 \quad \ldots \quad \ldots \quad \ldots \quad (5)$$

because the leakage reactance is only limited by how much loss can be tolerated at the upper frequency limit. Dividing equation (4) by (5)

$$\frac{f_2}{f_1} \approx \frac{L_p}{L} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (6)$$

Now replace $\frac{L_p}{L}$ by core radius in (6) and we see that the ratio of the pass-band limits is proportional to the core radius. We therefore want a core with as big a cross-sectional diameter as possible but having any mean length that may be convenient. The above argument applies to cores without derived magnetic paths and wherein the windings cover the whole core length. If the windings cover only a fraction of the total flux-path length then cores with as small a path length as possible should be employed as the path length for the leakage flux will be equal to the winding length, whereas that of the linking flux will be equal to the total core length. The ratio $\frac{f_2}{f_1}$ will only be maximum when the two path lengths are equal.

The most convenient shape for cores fulfilling the above requirements is evidently a toroid of convenient overall diameter and with as large a core diameter as possible. For concentric spiral windings of thin wide copper strip the best core shape is the double-window style usually formed by "E" and "I" laminations or interleaved "E" laminations. The reason that this style is not more popular is the difficulty of obtaining such laminations in very thin (0.002in) material. Considerations relating to shunt capacitance do have an influence on the foregoing, but are not decisive as they usually depend more on the form of the windings than on the actual winding diameter.

Those transformer constants relevant to the performance at f_1 have now been dealt with, and we turn to those which affect performance at f_2.

4. High-frequency Parameters

4.1. *Leakage Inductance.*[1]

It is evident from inspection of Fig. 2 (*b*) that the lower we can make L and C the higher will be f_2, all other things being equal. The number, n, of turns required has been fixed for us by equation (3) and the required working load, R_0. In this section we shall show how to make L as small as possible. Much of the technique which follows is only possible because n is a small number, and this should be constantly borne in mind. It is not proposed to go into the details of calculation of leakage inductance, but by means of figures to illustrate qualitatively, and by algebraic results to give quantitative information useful in transformer design. Consider a simple two-winding transformer, Fig. 13. The dotted lines represent leakage flux. If we neglect the no-load current, as was done in Fig. 2 (*b*), we see that were it not for leakage flux the inductance seen between the terminals 1 would be zero if the terminals 2 were short-circuited. As there is always leakage flux this inductance L is always present. It is calculated by equating $\frac{1}{2}Li^2$ to the total magnetic energy present in the two windings and in the insulation space between them

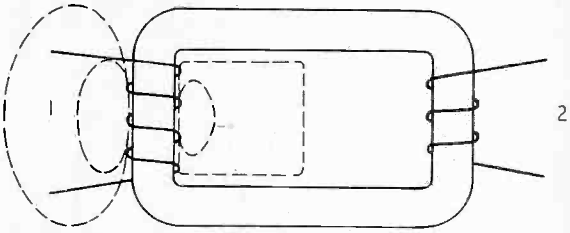


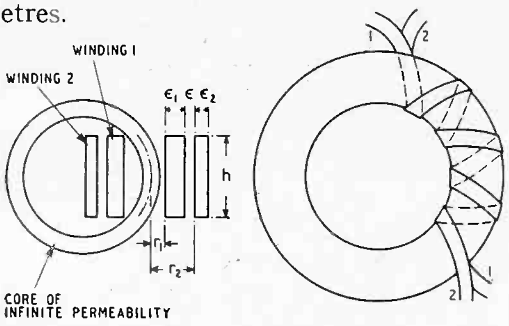
Fig. 13. *Simple two-winding transformer. The dotted lines represent leakage flux.*

when one of the windings is short-circuited and the other carries the current i. Consider now one or two practical cases.

Fig. 14 (*a*) shows two concentric helical or spiral windings. If the windings are spiral, then wide strip must be assumed for them. It is assumed that h, the winding width, is large compared with the radial distances between the ends of the windings and the core. In other words the length of the flux paths outside the core is considered to be equal to h only. If n is the number of turns in either winding

$$L = \frac{8\pi^2 n^2}{1{,}000h}\left[\left(\frac{r_1}{3}+\frac{\epsilon_1}{4}\right)\epsilon_1+\left(r_2-\frac{\epsilon}{2}\right)\epsilon+\left(r_2+\frac{\epsilon_2}{4}\right)\frac{\epsilon_2}{3}\right] \quad \ldots \quad \ldots \quad \ldots \quad (7)$$

wherein the units are microhenrys and centimetres.



(a) (b)

Fig. 14. *Two concentric windings are shown at (a) and a double winding of copper tape at (b).*

If the windings are wound with infinitely thin, wide strip then $\epsilon_1 = \epsilon_2 = 0$, provided that the number of turns on each is small and that the inter-turn insulation is very thin. Equation (7) then becomes

$$L = \frac{4\pi n^2}{1{,}000}\cdot\frac{2\pi\left(r_1+\frac{\epsilon}{2}\right)\epsilon}{h} \quad \ldots \quad \ldots \quad (8)$$

The second factor in (8) is simply the cross-section of the annular space between windings divided by its length. Equation (8) will be much used hereafter as it is very simple and sufficiently accurate for many spiral-strip type windings.

Consider now the case wherein the two windings are interleaved turn by turn instead of consisting of two separate blocks as shown in Fig. 14 (*a*). For this case we have

$$L = \frac{8\pi^2 \epsilon}{1{,}000h}\left[n\left(r_1+\frac{\epsilon}{2}\right)+\epsilon\,(n+2)(n-1)\right] \quad \ldots \quad \ldots \quad \ldots \quad (9)$$

wherein the units are in micohenrys and centimetres.

Again the windings are assumed to be wide copper strip of zero thickness separated by inter-turn insulation of thickness ϵ.

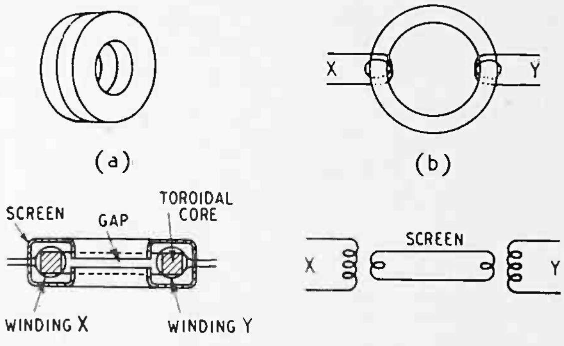
If the windings consist of two single concentric layers of helically wound infinitely thin, wide copper strip with spaced turns, Fig. 14 (*b*)

$$L = \frac{8\pi^2 n\epsilon}{1{,}000\,h}\left(r_1+\frac{\epsilon}{2}\right) \quad \ldots \quad \ldots \quad (10)$$

wherein the units are in microhenrys and centimetres.

If the region of applicability is carefully adhered to; formulae (8), (9) and (10) give accuracies usually better than 20 per cent, which is adequate. It is now evident that to reduce L to a minimum the winding should be as long as is feasible (h as large as possible), the core and inter-turn insulations should be as thin as possible, as should be the winding strip; and finally the number of turns should be restricted to no more than is essential to satisfy the 2-db shunt losses at f_1.

There is one more way in which L may be reduced and that is by the use of a copper screen between the windings. This applies particularly to toroidal types in which the windings cannot conveniently be made of strip in which corresponding turns fit over one another. For example, in cases where a rather large number of turns is required and the turns ratio is very different from unity. The action of the split screen may easily be seen from Fig. 15. Here (*a*) shows the two half-screens soldered together round the edges of their outer peripheries. The two windings X and Y are shown at opposite ends of a major diameter, but in fact the



(a) (b) (c) (d)

Fig. 15. *The two half-screens of a split screen (a) surround a core with windings X and Y (b) A section is shown at (c) and the equivalent circuit at (d).*

argument also holds if they are spread all round the toroid. The windings may both be inside or outside of the screen or one may be inside. It can immediately be seen from (*d*) that the screen acts as a mutual winding coupled to both X and Y. The screen may also be used to obtain greater coupling of a single winding to the core as indicated by Fig. 10. Screens may also be used to isolate windings from each other but this will be dealt with later.

So far we have dealt with only one of the two major causes of loss in performance at f_2. The other is shunt capacitance, C, Fig. 2 (b).

4.2. *Shunt Capacitance*[2]

This is very dependent upon the type and form of the windings adopted. Formulae will be given which give the overall shunt-capacitance, C, in terms of the component capacitances of which it is made up. Expressions for the individual component capacitances will be given, but these may be subject to large errors for a variety of reasons. Most of the formulae will give results to an accuracy within about 50 per cent, which, though somewhat disappointing, is none the less of great use when exploring paper designs before actually making a try-out. The formulae below were obtained by equating the total dielectric energy $\frac{1}{2}CV^2$ to the sum of the partial dielectric energies.

The case of a single-layer winding on a metal core is shown in Fig. 16 (*a*). The start of the winding is assumed to be at the same potential as the core, that is, the winding is unbalanced with respect to earth. Then

$$C = \frac{n-1}{n^2}C_2 + \frac{(n+1)(2n+1)}{6n}C_1 \qquad (11)$$

If $C_1 = C_2$

$$C = \left(\frac{n}{3} + \frac{1}{2} + \frac{7}{6n} - \frac{1}{n^2}\right)C_1 \quad \ldots \quad (12)$$

If n is large

$$C = \frac{n}{3}C_1 \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (13)$$

With a single-layer balanced winding on a

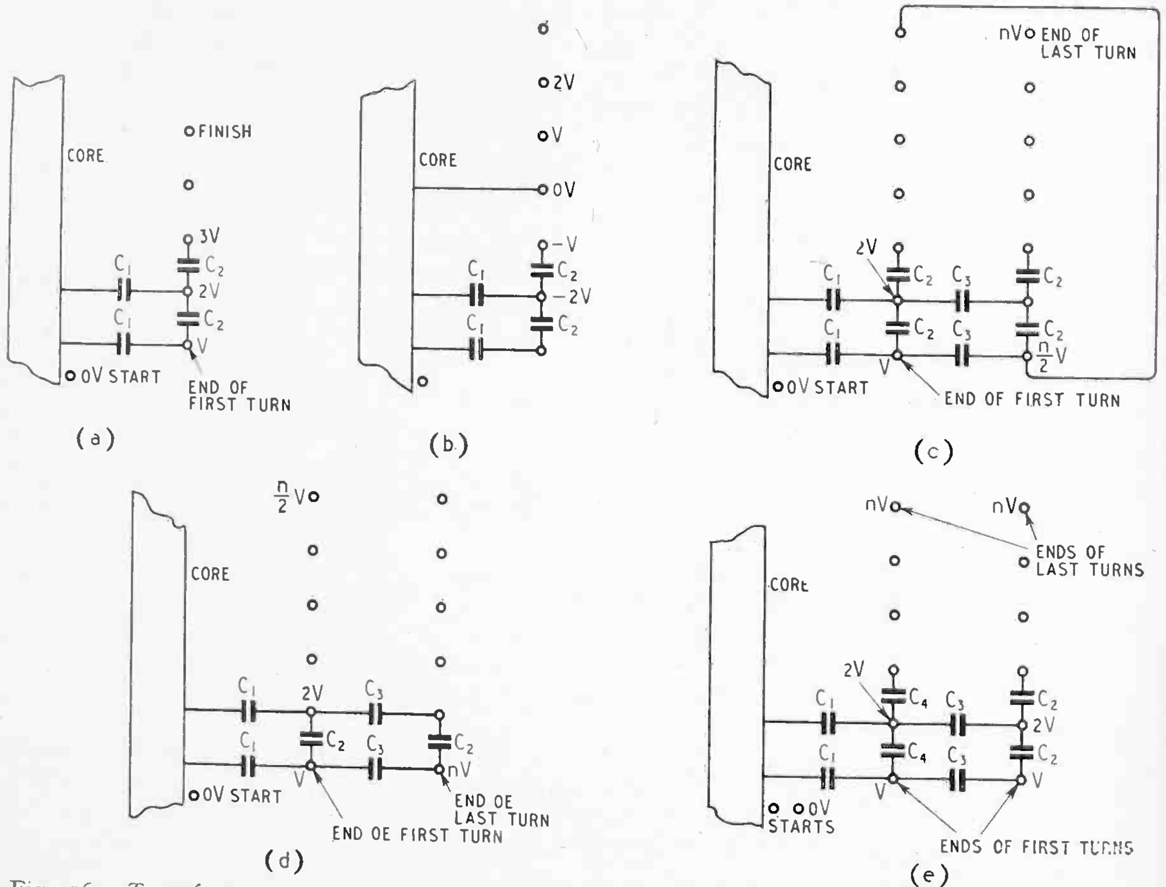


Fig. 16. *Transformer capacitances with a metal core:—(a) single-layer winding; (b) single-layer balanced winding; (c) double-layer winding; (d) double-layer unbalanced winding; (e) two concentric single-layer helical unbalanced windings.*

metal core, Fig. 16 (*b*), the centre of the winding is assumed connected to the core which may be earthed or floating. Then

$$C = \frac{n-2}{n^2} C_2 + \frac{(n+2)(n+1)}{6n} C_1 \qquad (14)$$

If $C_1 = C_2$

$$C = \left(\frac{n}{6} + \frac{1}{2} + \frac{7}{6n} - \frac{2}{n^2}\right) C_1 \quad .. \quad (15)$$

If n is large

$$C = \frac{n}{6} C_1 \quad .. \quad .. \quad .. \quad .. \quad (16)$$

A double-layer winding on a metal core is shown in Fig. 16 (*c*). The winding is assumed to be unbalanced; that is, to start at core potential.

$$C = \frac{\left(\frac{n}{2}+1\right)(n+1)}{12n} C_1 + \frac{n-2}{n^2} C_2 + \frac{\left(\frac{n}{2}-1\right)^2}{2n} C_3 \quad .. \quad (17)$$

If $C_1 = C_2 = C_3$

$$C = \left(\frac{n}{6} - \frac{3}{8} + \frac{19}{12n} - \frac{2}{n^2}\right) C_1 \quad .. \quad (18)$$

If n is large

$$C = \frac{n}{6} C_1 \quad .. \quad .. \quad .. \quad .. \quad (19)$$

A double-layer unbalanced winding on a metal core, in which the second-layer is wound back over first layer is shown in Fig. 16 (*d*). For this

$$C = \frac{\left(\frac{n}{2}+1\right)(n+1)}{12n} C_1 + \frac{n-2}{n^2} C_2 + \left[\frac{1}{3n}\left(\frac{n}{2}+1\right)(n-2) + \frac{1}{n^2}\right] C_3 \quad .. \quad (20)$$

If $C_1 = C_2 = C_3$

$$C = \left(\frac{5}{24} n + \frac{1}{8} + \frac{5}{12n} - \frac{1}{n^2}\right) C_1 \quad .. \quad (21)$$

If n is large $C = \frac{5}{24} n$ .. .. (22)

Two concentric single-layer helical unbalanced windings on a metal core, Fig. 16 (*e*) have a capacitance when both windings are in phase of

$$C = \frac{n-1}{n^2}\left(C_2 + C_4\right) + \frac{(n+1)(2n+1)}{6n} C_1 \quad .. \quad .. \quad .. \quad (23)$$

If $C_1 = C_2 = C_3 = C_4$

$$C = \left(\frac{n}{3} + \frac{1}{2} + \frac{13}{6n} - \frac{2}{n^2}\right) C_1 \quad .. \quad (24)$$

In these cases C is the shunt capacitance to earth of either winding.

If n is large $C = \frac{n}{3} C_1$ .. .. (25)

The same construction with windings in anti-phase has a capacitance

$$C = \frac{n-1}{n^2}\left(C_2 + C_4\right) + \frac{(n+1)(2n+1)}{3n}\left(\frac{C_1}{2} + 2C_3\right) \qquad (26)$$

If $C_1 = C_2 = C_3 = C_4$

$$C = \left(\frac{5}{3} n + \frac{5}{2} + \frac{17}{6n} - \frac{2}{n^2}\right) C_1 \quad .. \quad (27)$$

If n is large

$$C = \frac{5}{3} n C_1 \quad .. \quad .. \quad .. \quad .. \quad (28)$$

The last winding to be considered is a single spiral of flat strip. This capacitance may be obtained immediately as a special case of a single-layer winding by putting $C_1 = 0$ in equation (11).

$$C = \frac{n-1}{n^2} C_2 \quad .. \quad .. \quad .. \quad (29)$$

It is now necessary to give expressions for the component capacitances C_1, C_2, C_3 and C_4. It should be stated, however, that the principal use of formulae (11) to (29) is in showing the way in which the number of turns and the component capacitances affect the overall shunt capacitance.

To determine C_1, we assume the core to have a circular cross-section for simplicity.[3] Let ϵ_0 be twice the distance between the wire and the core surfaces and assume $\epsilon_0 \ll r$ where r is the core cross-sectional radius in inches. Let d be the wire diameter and K the dielectric constant of the core insulation.

$$C_1 = 0.96 \frac{Kr}{\log_{10} \dfrac{\sqrt{\epsilon_0^2 + 2d\,\epsilon_0} + \epsilon_0}{\sqrt{\epsilon_0^2 + 2d\,\epsilon_0} - \epsilon_0}} \quad .. \quad (30)$$

where C_1 is in picofarads.

For C_2, C_3 and C_4, we let r be the winding radius in inches and assume $r \gg \epsilon'_0$ where ϵ'_0 is the distance between turns (between wire surfaces).

$$C_2 = C_3 = C_4 = 1.9 \frac{Kr}{\log_{10} \dfrac{\sqrt{\epsilon_0'^2 + 2d\,\epsilon'_0} + \epsilon'_0}{\sqrt{\epsilon_0'^2 + 2d\,\epsilon'_0} - \epsilon'_0}} \quad .. \quad .. \quad (31)$$

where C_2, C_3 and C_4 are in picofarads.

(*To be continued*)

NON-LINEAR REGENERATIVE CIRCUITS*

Frequency and Amplitude Discrimination

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1. Introduction

AMPLITUDE discrimination in non-linear circuits is a well-known phenomenon; it must often have been realized that if two tones of unequal level are applied at suitably high levels to a non-linear circuit, then the tone of higher level can considerably reduce the output of the other. However, the effect is not very useful because it is not easy to predict quantitatively with precision, and because it is accompanied by the production of various harmonic and intermodulation frequencies. The application of positive feedback to such a circuit modifies the effect somewhat, and if the feedback is large enough to permit of free oscillation in the absence of an input signal then the amplitude discrimination may be very large at very small levels of input signal and is accompanied by the production of only small amounts of other frequencies. Under these conditions it is a very useful effect. It is necessary to ensure, of course, that no free oscillation does actually occur in the presence of the input signal, and for this reason, the circuit must take the form of a synchronized oscillator. This involves a frequency-discriminating circuit, and the resultant response is therefore a function of this as well as of amplitude discrimination.

In this paper the problem is analysed as that of a synchronized oscillator with an unwanted signal mixed with the injected locking tone. In order to discuss at first only the amplitude discrimination, the unwanted signal is assumed to be of a frequency very close to that of the locking tone; this allows any frequency-dependent effect to be neglected. Later on, the effect of frequency selectivity is briefly discussed.

Before proceeding to the analysis, some previously published discussion[1] of the subject must be referred to. The discrimination of a synchronized oscillator against unwanted signals mixed with the injected synchronizing tone has been shown to be in general a function of two main factors, (1) non-linearity of the amplifier circuit, and (2) the effect of the tuned positive feedback. The previous work considered three frequency ranges, thus :—

(1) Where the unwanted signal has a frequency well removed from the oscillator frequency, so that the effect of the tuned positive feedback can be neglected. The only effect is then non-linearity, and it is shown that the amplification of the unwanted signal is itself linear (because of the small relative amplitude, due to the absence of feedback), but entirely dependent on the amplitude of the forced oscillation at the synchronized frequency. Over ordinary ranges of amplitude, an increase in wanted output causes a reduction of unwanted output.

(2) Where the unwanted signal has a frequency within the range of operation of the tuned positive feedback, but not too close to the oscillator frequency; this last restriction is defined arbitrarily by requiring the unwanted frequency to be outside the range of frequencies over which the oscillator will synchronize. Here, in addition to the non-linear effect mentioned above, which leads to a linear amplification of the unwanted signal, there is a frequency selective response, such that frequencies nearer the synchronized frequency receive some amplification by the tuned positive feedback effect, and are not, therefore, provided with as much discrimination as frequencies further away.

(3) Where the unwanted frequency lies close to the wanted (synchronized) frequency (it is arbitrarily defined as lying within the frequency range over which synchronization occurs), and in consequence both frequencies have comparable amounts of feedback and non-linearity. This case is really the general case, where no real analysis seems practicable. However, by restricting the level of the unwanted signal and assuming it is so close in frequency to the synchronizing frequency that the selective

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* MS. accepted by the Editor, July 1946.

effect produced by tuning is negligible, it becomes possible to effect an analysis, and this is, of course, the condition described briefly in the first paragraph. An application which is of importance in radio work for automatic tuning and in other equipment now being developed by the Post Office Research Branch, is the use of a synchronized oscillator to reduce the depth of modulation of a test or carrier tone with a modulated envelope. The modulation may be, say, 30 c/s, or at the most, a few hundred cycles/sec, on a carrier of a few megacycles/sec. For convenience of expression, the analysis which follows is given in terms of the suppression of the modulation in such a case. Some experimental work on this subject has been published by Byard and Eccles,[2] but it does not attempt to explain the results or analyse the problem. The effect is in no way dependent on the unwanted frequencies being sidebands of an amplitude-modulated wave; the result applies for any frequency close to the synchronizing frequency.

2. Analysis

The problem to be analysed may be restated thus :—

A feedback oscillator, as in Fig. 1, has injected into its grid circuit a tone of frequency very near its natural frequency, modulated to a small depth by a very low frequency such that the sidebands produced by the modulation lie so close in frequency to the tone itself that it can be assumed that the feedback in the oscillator circuit is of substantially constant amplitude response and zero phase for all the frequencies concerned. It is then required to determine by how much the depth of modulation (or the relative amplitude of the sidebands) is reduced in the output by passage through the oscillator circuit.

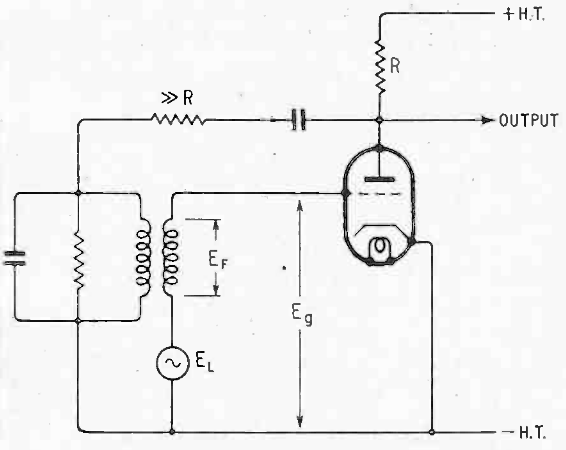

Fig. 1. *In this typical oscillator circuit E_F = feedback voltage, E_L = input, synchronizing or locking voltage and E_g = grid voltage. Suffixes 1 and 2 are used in the text to indicate components of angular frequency ω_1 and ω_2, while the suffix* 0 *indicates the free oscillation value.*

By small depth of modulation is meant one such that twice the square of the sideband amplitude is small relative to the square of the carrier amplitude (see later). The closeness of the sideband frequencies to the tone frequency is less readily defined, but little error is likely to be produced if the phase angle of the tuned circuit at the sideband frequency does not exceed say, 0.1 radian ; this defines the maximum modulating frequency as a proportion of about $\frac{0.05}{Q}$ of the tone frequency, where Q is the magnification factor for the oscillator tuned circuit itself.

2.1 *Non-linearity in the Valve as an Amplifier.*

Consider two signals $\hat{E}_{g1} . \sin \omega_1 t$ and $\hat{E}_{g2} . \sin \omega_2 t$ applied in series to an amplifier whose instantaneous output voltage e_m is related to its instantaneous input voltage e_i by a power series, thus :—

$$e_m = \sum_{n=0}^{\infty} a_n e_i^n \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (1)$$

For limited ranges of amplitude, a sufficiently close representation is obtained by considering only terms up to the third power. If we do this, and substitute the input signal* $\hat{E}_{g1} \sin \omega_1 t + \hat{E}_{g2} \sin \omega_2 t$ for e_i then we find that the output contains several components, including d.c., sum and difference products, harmonics, and the two input frequencies represented by

$$(a_1 + \tfrac{3}{4} a_3 \hat{E}_{g1}^2 + \tfrac{3}{2} a_3 \hat{E}_{g2}^2)\, \hat{E}_{g1} \sin \omega_1 t + (a_1 + \tfrac{3}{4} a_3 \hat{E}_{g2}^2 + \tfrac{3}{2} a_3 \hat{E}_{g1}^2)\, \hat{E}_{g2} \sin \omega_2 t \quad \ldots \quad \ldots \quad (2)$$

It will be appreciated that a_3 is negative, so that the amplification for these components is reduced as the amplitudes increase.

If we now make the stipulation that $2\hat{E}_{g2}^2 \ll \hat{E}_{g1}^2$

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*If this signal is applied to a grid circuit with a bias voltage E_b, then instead of a_1 in equation (2) we should write $A_1 = a_1 + 2a_2E_b + 3a_3E_b^2$. As this term is independent of the signal voltages, the presence of bias in no way affects the working and conclusions of this paper, except that wherever a_1 is written, it should be replaced by A_1.

$$\left.\begin{array}{l}\text{we see that the amplification to}\\ \omega_1 \text{ is } \quad a_1 + \tfrac{3}{4} a_3 \hat{E}_{g1}^{\,2}\\ \text{and the amplification to}\\ \omega_2 \text{ is } \quad a_1 + \tfrac{3}{2} a_3 \hat{E}_{g1}^{\,2}\end{array}\right\} \quad \ldots \quad (3)$$

The amplification to ω_2 is thus linear ; i.e., independent of its own amplitude.

We can see that this working represents the case of the modulated tone, except that we have considered only one sideband. The sidebands evidently have negligible influence on each other, so the reduction in depth of modulation caused by the passage through the amplifier is by the ratio

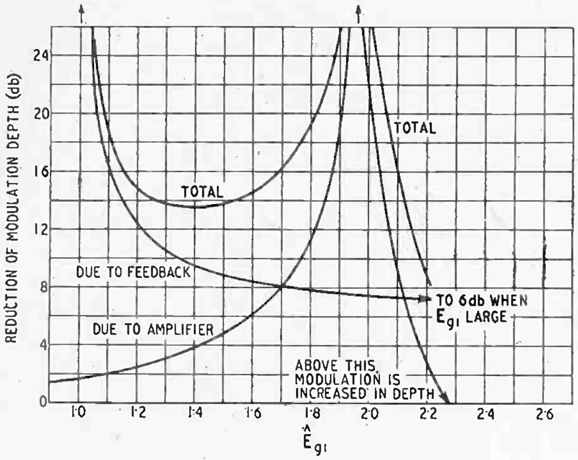

Fig. 2. *The reduction of the modulation depth of the injected signal as a function of grid voltage* $(a_1/a_3 = -5{\cdot}75,\ \hat{E}_{g0} = 1.0)$.

$$R_A = \frac{a_1 + \frac{3}{4} a_3 \hat{E}_{g1}^{\,2}}{a_1 + \frac{3}{2} a_3 \hat{E}_{g1}^{\,2}} \quad \ldots \quad \ldots \quad (4)$$

where $\hat{E}_{g1}$ is the peak amplitude of the tone when unmodulated. The absolute values of a_1 and a_3 are of no significance ; it is their ratio which determines R_A.

To see the effect of this it is convenient to take numerical values. For a small r.f. pentode a typical value of a_1/a_3 is $-5{\cdot}75$. For this value the ratio R_A is plotted (in decibels) against the peak tone voltage in Fig. 2. It will be seen that the reduction in depth of modulation increases as the voltage increases, up to an infinite point where the amplification to the sideband is zero ; above this the reduction in depth decreases, until after a certain point the depth is actually increased. We must not, however, overlook the fact that this considers only those components of the output which are at the input frequencies ; other frequencies are produced by the non-linear action, and are discussed in Section 2.4.

2.2 *Positive Feedback*

The circuit of the oscillator is shown in Fig. 1, and it is clear that in the equilibrium condition for any frequency the grid voltage E_g is the sum of the input voltage E_L and the fed-back voltage E_F ; i.e.,

$$E_g = E_L + E_F \quad \ldots \quad \ldots \quad \ldots \quad (5)$$

Now for the tone at frequency ω_1 (considering the fundamental component only)

$$\hat{E}_{F1} = (a_1 + \tfrac{3}{4} a_3 \hat{E}^2_{g1})\, \hat{E}_{g1}$$

$$\text{i.e., } \hat{E}_{L1} = \hat{E}_{g1} - \hat{E}_{F1}$$

$$= \hat{E}_{g1}\,(1 - a_1 - \tfrac{3}{4} a_3 \hat{E}_{g1}^{\,2}) \quad \ldots \quad (6)$$

where a_1 and a_3 can now be specifically defined as the loop-gain coefficients. Similarly for the sideband at frequency ω_2

$$\hat{E}_{F2} = (a_1 + \tfrac{3}{2} a_3 \hat{E}_{g1}^{\,2})\, \hat{E}_{g2}$$

$$\text{i.e., } \hat{E}_{L2} = \hat{E}_{g2} - \hat{E}_{F2}$$

$$= \hat{E}_{g2}\,(1 - a_1 - \tfrac{3}{2} a_3 \hat{E}_{g1}^{\,2}) \quad \ldots \quad (7)$$

The input modulation ratio is $\dfrac{2\hat{E}_{L2}}{\hat{E}_{L1}} = m_L$

$\therefore$ From (6) and (7),

$$\frac{2\hat{E}_{g2}\,(1 - a_1 - \frac{3}{2} a_3 \hat{E}_{g1}^{\,2})}{\hat{E}_{g1}\,(1 - a_1 - \frac{3}{4} a_3 \hat{E}_{g1}^{\,2})} = m_L \quad \ldots \quad (8)$$

The modulation ratio on the grid, as a result of the feedback, is

$$\frac{2\hat{E}_{g2}}{\hat{E}_{g1}} = m_g \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (9)$$

Thus from (8) and (9), the ratio of reduction in depth of modulation due to the feedback effect is

$$R_F = \frac{m_L}{m_g} = \frac{1 - a_1 - \frac{3}{2} a_3 \hat{E}_{g1}^{\,2}}{1 - a_1 - \frac{3}{4} a_3 \hat{E}_{g1}^{\,2}} \quad \ldots \quad (10)$$

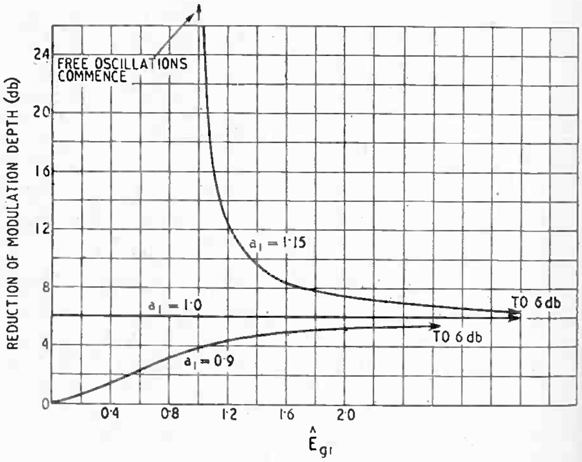

Fig. 3. *The reduction of modulation depth due only to feedback for different amounts of feedback* $(a_1/a_3 = -5{\cdot}75)$.

It is clear that in this case the actual magnitudes of a_1 and a_3 matter, and not only their ratio. A numerical case is shown in Fig. 2 for values of a_1 and a_3 of 1.15 and −0.2 respectively. These values give the same ratio of a_1/a_3 as was used for illustrating the previous effect, and give a free oscillation amplitude (i.e., the amplitude of the self-oscillation when there is no input) at the grid of $\hat{E}_{g0} = 1.0$. It will be seen that the reduction in depth of modulation is very large when the grid amplitude only slightly exceeds the free oscillation value (i.e., when the input voltage is small), but decreases to an ultimate value of 6 db as the grid voltage (and also the input voltage) is increased.

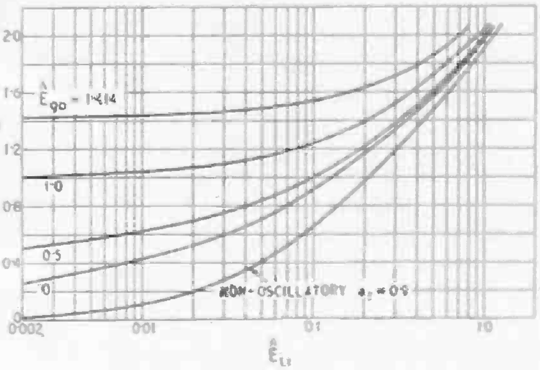

Fig. 4. *The relation between the input, forced, and free oscillation amplitudes* ($a_1/a_3 = -5.75$).

When the output is taken from the anode circuit of the oscillator, the two effects described above are additive (in decibels), as shown in Fig. 2. The minimum reduction in depth of modulation in this example is about 14 db except at very high amplitudes which would not normally be met with in practice on this type of valve.

In Fig. 3 the feedback effect is illustrated for the case of the circuit being *just* capable of self-oscillation ($a_1 = 1.0$), and of being incapable of self-oscillation ($a_1 = 0.9$). The former case gives a constant reduction of 6 db, while the latter gives a reduction rising from zero to 6 db.

2.3 *Relations between Input and Grid Voltages*

So far, for ease in analysis, the effects have been expressed in terms of grid voltage. It is generally more convenient for practical applications to use relationships in terms of input voltage. It is simple enough to relate input and grid voltages, and also to determine the amplitude $\hat{E}_{g0}$ of free or self-oscillation.

We obtain the value of $\hat{E}_{g0}$ as the particular value of $\hat{E}_{g1}$ in Equ. (6) when the input $\hat{E}_{L1} = 0$;

$$\text{i.e., } \hat{E}_{g0} = \sqrt{\frac{1 - a_1}{\frac{3}{4}a_3}} \qquad \ldots \quad \ldots \quad (11)$$

Therefore, for any specified value of $\hat{E}_{g0}$ and given value of a_1/a_3, we can determine the corresponding values of a_1 and a_3. Then by Equ. (6) we can relate the input and grid voltages. This has been shown graphically in Fig. 4, where a family of curves relating $\hat{E}_{L1}$ and $\hat{E}_{g1}$ is given for a number of values of $\hat{E}_{g0}$. A curve is also given for a case where the loop gain is insufficient to maintain self-oscillation (i.e., $a_1 < 1.0$). It will at once be observed that when $\hat{E}_{g0}$ is large, and consequently the valve is being worked very non-linearly, the input voltage, if small, has practically no influence on the grid voltage. This leads to a simple physical explanation of the reduction in depth of modulation; variations in the envelope amplitude of the input signal, produced by the modulation, cause little corresponding variation in the output envelope amplitude.

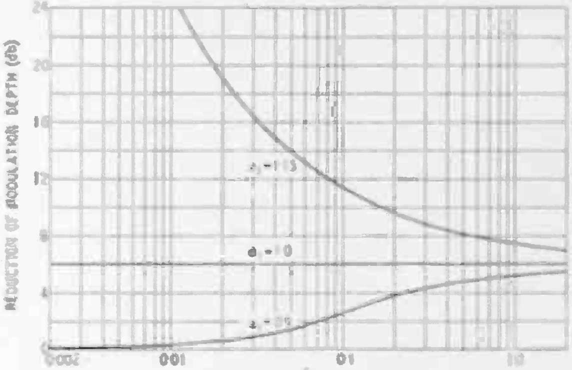

Fig. 5. *The reduction of modulation depth due to feedback expressed as a function of the injected voltage* ($a_1/a_3 = -5.75$).

In Fig. 5 the reduction in depth of modulation due to the feedback effect alone is plotted against input voltage for the three conditions, circuit capable, just capable, and not capable of self-oscillation.* In Fig. 6, the overall reduction (i.e., the sum of the amplifier non-linearity and feedback effects) is similarly plotted against input voltage. It will be seen how satisfactory the synchronized

---

* It should be noted that if the feedback were negative, or "degenerative," the effect would be opposite for the smaller values of input voltage (i.e., the difference in level between the two tones would be reduced), although to a relatively small extent. At larger input voltages, the effect is similar whether the feedback is positive or negative, and approaches a final value of discrimination of 6 db.

oscillator can be in producing a relatively pure tone from a modulated one, which can be of very low level, or from one mixed with some other unwanted frequency, providing the amplitude of free oscillation is high enough to indicate a reasonable margin of loop gain.

2.4 *Other Frequency Components*

In the foregoing sections only the components at the fundamental frequencies have been considered. A number of other components is produced by the non-linear action, comprising harmonics of the input frequencies, various sum and difference products, and particularly products such as $2\omega_1 - \omega_2$ and $2\omega_2 - \omega_1$. Of all these, only those of the last type fall anywhere near the oscillator frequency, and it is reasonable to assume the tuned circuit of the oscillator reduces the amplitudes of all the others to negligible proportions. Assuming the cubic law, the only frequencies falling near the oscillator frequency are $2\omega_1 - \omega_2$ and $2\omega_2 - \omega_1$. The amplitudes are $-\frac{3}{4} a_3 \hat{E}_{g1}{}^2 \hat{E}_{g2}$ and $-\frac{3}{4} a_3 \hat{E}_{g1} \hat{E}_{g2}{}^2$ respectively.

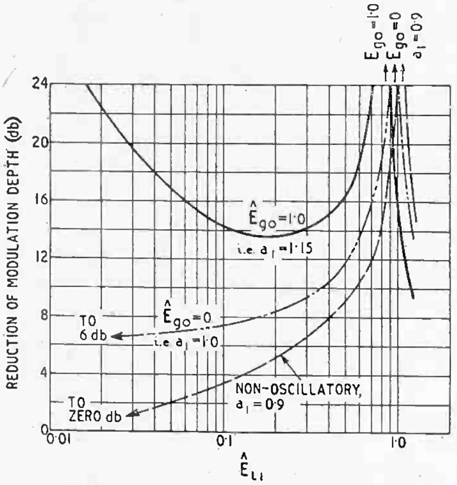

Fig. 6. *Overall reduction of modulation depth for various injected voltages* $(a_1/a_3 = -5{\cdot}75)$.

Since we know $\hat{E}_{g2}{}^2 \ll \hat{E}_{g1}{}^2$, the second is obviously very small, leaving the component $-\frac{3}{4} a_3 \hat{E}_{g1}{}^2 \hat{E}_{g2} \sin(2\omega_1 - \omega_2)\,t$ as the only one of any significance.

It is evident, then, that the ratio of the amplitude of this component to that of the sideband frequency ω_2 is $-\frac{3}{4} a_3 \hat{E}_{g1}{}^2$. For a large proportion of practical cases, this ratio will be considerably less than unity, and the third-order component will be of little importance. For large values of $\hat{E}_{g1}$, the amplitude of the new frequency will exceed that of ω_2, and this may be serious; but it must be observed that such cases correspond to large injected voltages. In practice, such large injected voltages are unlikely to be used, as they do not take advantage of the feedback effect and serve no useful purpose.

2.5 *Discrimination over the Whole Frequency Range*

Having shown how the synchronized oscillator provides discrimination on a basis of amplitude non-linearity to an unwanted signal, even when this is of frequency very close to the synchronized frequency, it will be useful to consider the discrimination over the whole frequency range. As pointed out in Section 1, when the unwanted frequency ω_2 is remote from ω_1, then it receives no positive feedback although ω_1 has the full amount. There is thus a frequency discrimination against ω_2, coupled with a smaller amount of non-linear amplitude discrimination if the amplitude $\hat{E}_{L2}$ is less than $\hat{E}_{L1}$. As ω_2 approaches more nearly to ω_1 it receives an increasing amount of feedback, and the discrimination against it is reduced. Finally, when ω_2 is very close to ω_1, it has only amplitude discrimination.

A numerical example will make this clear. Taking $a_1/a_3 = -5{\cdot}75$, $\hat{E}_{g0} = 1.0$, and $Q = 15$, we have, assuming $\hat{E}_{L2}{}^2 \ll \hat{E}_{L1}{}^2$:

Case (1). If $\hat{E}_{L1} = 0.1$, then from Fig. 4, $\hat{E}_{g1} = 1.24$. The gain which ω_1 receives due to feedback is therefore 21.9 db. There is also, from Fig. 2, 2.7 db discrimination against ω_2 due to non-linearity. The total discrimination when $\omega_2 \ll \omega_1$ or $\omega_2 \gg \omega_1$ is therefore 24.6 db. When $\omega_2 \approx \omega_1$ the discrimination is 14.2 db, from Fig. 6. In the intermediate range the gain provided by the feedback effect to the frequency ω_2 is given [1, 3] by

$$\left|\frac{\hat{E}_{g2}}{\hat{E}_{L2}}\right| = \sqrt{\left[\frac{1 + Q^2(1 - x^2)^2}{(1 - x)^2 + Q^2(1 - x^2)^2}\right]} \qquad (12)$$

where $x = \omega_2/\omega_1$. This ignores the fact that the feedback loop is non-linear, since the consideration of this factor makes the equations practically insoluble.[1]

On plotting these relations, it is easy to join the three ranges smoothly together, giving an approximate overall response curve as shown in Fig. 7, which must necessarily give a close approach to the curve in spite of the difficulty just mentioned.

Case (2). If $\hat{E}_{L_1} = 0.03$, then $\hat{E}_{g1} = 1.09$. Thus when $\omega_2 \ll \omega_1$ or $\omega_2 \gg \omega_1$ the discrimination is $31.2 + 2.0$ db. When $\omega_2 \approx \omega_1$ it is 19.6 db (from Fig. 6). The overall response is also shown in Fig. 7.

It will be seen, therefore, that the response is throughout dependent on the actual voltages in the circuit, but if the unwanted signal is of considerably lower amplitude than the synchronizing signal in the input, then the discrimination is dependent only on the voltages of the synchronizing frequency. The important conclusion is that there is everywhere some discrimination, a property not possessed by any linear frequency-selective circuit.

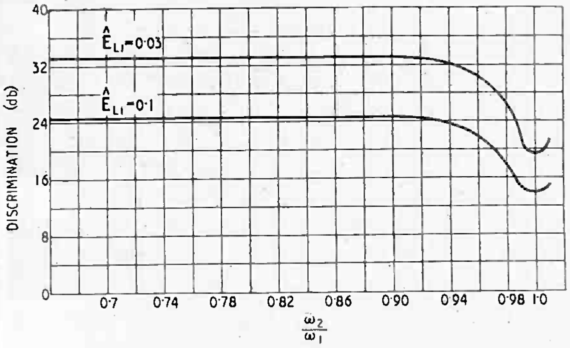

Fig. 7. *Discrimination over the whole frequency range* ($\hat{E}_{g0} = 1.0$, $a/a_3 = -5.75$, $Q = 15$).

2.6 *Output from the Tuned Circuit*

We have so far considered the output as being taken from the anode of the valve, with the feedback resistance large compared with the anode resistance, as shown in Fig. 1. If the output were taken across the tuned circuit, as is quite common practice, the performance would be slightly modified, in that the frequency response of the tuned circuit itself must be added to the discrimination shown in Fig. 7.

2.7 *Practical Design of the Circuit*

The analysis has shown how the circuit behaves; in practice, we have to consider how best to use this behaviour for meeting specified requirements. To reduce the depth of modulation to a specified extent, it will be necessary to obtain from a family of curves such as Fig. 6 suitable values of $\hat{E}_{g0}$ and $\hat{E}_{L_1}$ (assuming, of course, that some particular valve is already decided upon). The value of $\hat{E}_{g0}$ should not be unnecessarily high, as this might lead to difficulty with the third-order products, as discussed in Section 2.4. The value of $\hat{E}_{L_1}$, in relation to $\hat{E}_{g0}$ and the Q-factor of the tuned circuit, determines over what frequency range the oscillator will remain synchronized; if the input is unmodulated, the relationship for small values of $\hat{E}_L$ is as follows[3] :—

$$\hat{E}_L = 2Q \, . \, \hat{E}_{gp} (1 - x_p) \qquad \ldots \qquad \ldots \qquad (13)$$

where $\hat{E}_{gp}$ is the grid voltage at the pull-out point (this is somewhat less than $\hat{E}_{g0}$, by an amount not exceeding about 30%), and x_p is the ratio of the pull-out frequency to the frequency of free oscillation.

If the input is modulated, the value of $\hat{E}_L$ used must be that at the troughs of the modulated envelope.

The synchronous range must cover the normal range of variation of frequency of the oscillator, and thus the value of $\hat{E}_{L_1}$ chosen must satisfy this condition also. The Q-value of the tuned circuit does not influence this requirement, because the stability of the oscillator is approximately proportional to Q, but it does affect the discrimination against unwanted frequencies at some parts of the frequency range, as discussed in Section 2.5.

3. Byard and Eccles' Experimental Results

In view of the previous publication of experimental observations on this subject by Byard and Eccles,[2] it seems only just to consider whether their results are consistent with the theory developed here. The experimental results will be discussed in terms of the diagram numbers of the published paper. It should be observed that Byard and Eccles use a somewhat different circuit from that considered in the present paper, and their " output " is taken from the grid circuit. This means that only the feedback effect of Section 2.2 is involved.

Fig. 2.—An increase in the depth of modulation is indicated as the oscillator is tuned to a frequency less close to that of the input tone. This is quite reasonable, since the amount of feedback is thereby reduced, thus decreasing the reduction in depth of modulation.

Fig. 3.—The linear relation between output and input depths of modulation is to be expected from equation (3), provided the depth of modulation is not excessive. It is surprising that the linear relationship appears to hold so closely even up to 100% modulated input. The increase in modulation of the output as the input level is increased is to

be expected from a consideration of Fig. 5 of the present paper.

Figs. 4 and 5.—It is only to be expected that the synchronous frequency range will be determined by the troughs of the modulated input signals, and the linear relation is in agreement with the present author's general theory of synchronization[3].

Fig. 6.—The increase in modulation of the output as the input level is increased is to be expected (as mentioned above under Fig. 3); the fact that this figure shows a linear relationship suggests that the range of input voltage is restricted. Since the actual range of input voltage shown is from zero to a value slightly in excess of the free oscillation voltage, it must be concluded that the valve used had a different (i.e., lower) ratio a_1/a_3 from that considered in the examples in the present paper.

Fig. 7.—The decrease of the depth of modulation as the feedback is increased with constant input voltage correctly corresponds to the theory, as can be seen from Fig. 5 of the present paper.

It can thus be concluded that Byard and Eccles' experimental results are consistent with the theory here presented.

4. The Non-linear Response of Actual Valves

It has been assumed above that the valve response can be adequately represented by a cubic relationship between output and input voltage. In practice, of course, the relationship is much more complicated, and it is difficult to match the response even by a fifth-power expression. But the general nature of the behaviour of the circuit is quite accurately demonstrated by the use of the cubic, and the author has, in fact, found little serious discrepancy between theory and experiment due to using the cubic law as far as the feedback effect is concerned. The direct effect (see Section 2.1) gives poorer agreement. The best plan is to determine a suitable cubic law covering the usually small range of voltage values used in practice. For instance, the ratio of $a_1/a_3 = -5{\cdot}75$, which has been used in the numerical examples above, represents approximately a valve type SP41 (Mazda) used with an H.T. voltage of 150 volts and a grid voltage (peak a.c.) range from 0 to about 1.5 volts.

5. Conclusions

Amplitude discrimination in non-linear circuits, whereby a signal of high level can reduce the output of a signal of lower level, can be made a useful and readily utilizable effect by the application of positive feedback sufficient to produce self-oscillation in the absence of an input signal. The system then becomes a synchronized or " locked-in " oscillator, which is shown to provide discrimination against signals mixed with the locking tone not only when they are of frequency lying outside the synchronous range (an effect due chiefly to frequency-selectivity), but also when they lie extremely close to the synchronizing frequency (an effect due entirely to non-linearity coupled with positive feedback). Thus, if a modulated tone is injected into an oscillator as a synchronizing signal, then the output of the oscillator will contain a greatly reduced depth of modulation. This system provides a means of obtaining a relatively pure tone from a signal which may have other frequencies mixed with it which cannot be separated by ordinary frequency-selective circuits. The only restriction needed is that the unwanted signal component in the input must be smaller than the wanted component. The effect is not dependent on the unwanted frequencies being sidebands of an amplitude-modulated wave.

Previously published experimental work by Byard and Eccles is shown to be consistent with the theory developed here.

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[1] D. G. Tucker, " The Synchronization of Oscillators, Part IV :—The Discrimination of a Synchronized Oscillator against Unwanted Signals mixed with Control Tone," *Electronic Engineering*, 1943, Vol. 16, p. 114.

[2] S. Byard and W. H. Eccles, " The Locked-in Oscillator," *Wireless Engineer*, 1941, Vol. 18, p. 2.

[3] D. G. Tucker, " Forced Oscillations in Oscillator Circuits, and the Synchronization of Oscillators," *Journ. Instn elect. Engrs*, 1945, Vol. 92, Part III, p. 226.

CORRESPONDENCE

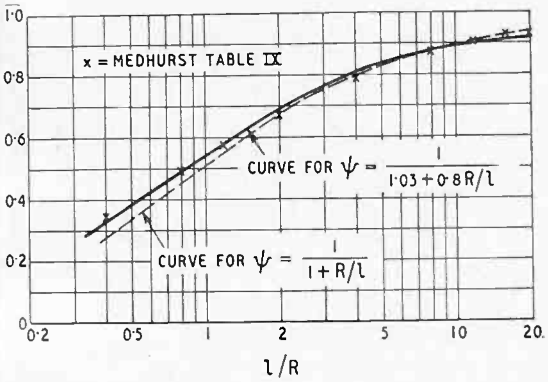
Letters of technical interest are always welcome. In publishing such communications the Editors do not necessarily endorse any technical or general statements which they may contain.

Q of Solenoid Coils

To the Editor, "Wireless Engineer"

SIR,—From Mr. Medhurst's paper in *Wireless ngineer* for March, one concludes that the results f his careful measurements on solenoid coil Q ıould be taken as superseding Butterworth's ormulæ which have for so long been taken by ngineers as a practical basis for coil design.

Mr. Medhurst's results for the Q of coils wound ith the optimum gauge of wire are contained in ıs formula $Q = 0.15R\psi\sqrt{f}$ and his curve for ψ ıown in Fig. 13. It is interesting to note that he latter curve agrees to an accuracy of a few per ent with the simple empirical equation $\psi = 1/(1.03$ $+ 0.8R/l)$ as shown in the graph. Thus we arrive the following simple and practical formula for ıe Q of a properly designed solenoid, using plain opper wire, at high radio frequencies :—



$$Q = \frac{\sqrt{f}}{(6.9/R + 5.4/l)}$$

here R and l are the radius and length of the coil ı centimetres. In a majority of practical cases e can use the even simpler formula

$$Q = 0.15\sqrt{f}/(1/R + 1/l)$$

hich follows the data to a few per cent, provided $> R$.

The range of conditions under which this formula pplies is the same as that to which Mr. Medhurst's ata refer : in particular—

(*a*) The ratio of wire diameter to wire spacing must approximate to the optimum (i.e., this ratio should lie between 0.5 and 0.7 for short coils ($l<2R$) or 0.6 to 0.8 for l order of $4R$, and 0.75 to 0.9 for very long coils).

(*b*) The formulæ apply strictly only for very high frequencies, but (from Sect. 10) the accuracy will be better than ± 10 per cent provided $z>7$; i.e., provided frequency in Mc/s exceeds 0.5 divided by (wire diameter in mm)^2.

Solenoids are chiefly used in current practice on frequencies above about 3 Mc/s, and here Litz wire is of little or no advantage ; the following table, giving the thinnest gauge for which the formula applies to ± 10 per cent, shows that most practical solenoids will be covered.

freq.	1 Mc/s	4 Mc/s	16 Mc/s
wire ..	22 S.W.G.	28 S.W.G.	37 S.W.G.

(*c*) The formulæ do not hold for coils of very few turns (or extremely short coils). Mr. Medhurst gives us little guidance as to how far Butterworth's correction factor for coils of few turns can be relied on. Further experimental work seems indicated in view of the practical importance of coils of very few turns on v.h.f. and u.h.f.

(*d*) Dielectric loss is not, of course, allowed for. This is unlikely to be material except where the coil has a rather poor dielectric (bakelite or worse) and is used in a circuit having a low parallel tuning capacitance.

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Absolute Rotation and a Rotating Magnet

To the Editor, "Wireless Engineer"

Sir,—There are two aspects of the old rotating magnet problem : one is the absolute property of rotation, and the other is the electromagnetic reaction on the magnet itself. Various suggestions have been put forward by different authors to explain absolute rotation and from them to interpret the electromagnetic reaction upon the magnet.[1]

The problem is, however, as fundamental as it is hoary. We should especially notice that the questions of absolute acceleration and absolute rotation, which were sanctioned in the special relativity and which signify a return to the Newtonian Hypothesis of absolute space and absolute time, served, in fact, as the genesis of Einstein's theory of general relativity. Einstein's "principle of equivalence" asserts that, for the description of physical processes, we may take either the view-point of an observer A, at rest with an inertial coordinate system or that of another observer B, on a constantly accelerated coordinate system who measures the physical processes by means of the coordinates of observer A and, in addition, assumes the presence of a homogeneous gravitational field.[2] This principle, properly applied here for any particular instant, gives us a simple way of explaining so-called "absolute rotation". The electromagnetic reaction upon a rotating magnet and the resultant electric field distribution in it can then be obtained from Maxwell's equations modified for the relativity effect.

1. *Absolute Rotation*

Consider an inertial system Σ, in which the position-vector of any point is given by

$$\mathbf{r} = \mathbf{i}x + \mathbf{j}y + \mathbf{k}z \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (1)$$

and another coordinate system Σ', with its $\mathbf{k}'$-axis and origin coinciding with the $\mathbf{k}$-axis and origin of the first system, and rotating about the $\mathbf{k}$-axis with constant angular velocity ω. The position-vector for the same point referred to the latter system is then

$$\mathbf{r} = \mathbf{i}'x + \mathbf{j}'y + \mathbf{k}'z \quad \ldots \quad \ldots \quad \ldots \quad (2)$$

Denoting differentiation with respect to the rotating system by a prime ('), then the acceleration of the position-vector for both systems is given by the following relation[3]:

$$\frac{d^2\mathbf{r}}{dt^2} = \frac{d'^2\mathbf{r}}{dt^2} + 2\left[\boldsymbol{\omega} \times \frac{d'\mathbf{r}}{dt}\right] + [\boldsymbol{\omega} \times (\boldsymbol{\omega} \times \mathbf{r})] \quad (3)$$

The second term on the right-hand side is the Coriolis acceleration, which is present only when the particle moves in the rotating system. Let m be the mass of the particle considered, then Newton's second law of motion gives

$$\mathbf{F} = m\frac{d^2\mathbf{r}}{dt^2} = m\frac{d'^2\mathbf{r}}{dt^2} + 2m\left[\boldsymbol{\omega} \times \frac{d'\mathbf{r}}{dt}\right] + m[\boldsymbol{\omega} \times (\boldsymbol{\omega} \times \mathbf{r})] \quad \ldots \quad \ldots \quad \ldots \quad (4)$$

The force referred to the moving system is

$$\mathbf{F}' = m\frac{d'^2\mathbf{r}}{dt^2} = \mathbf{F} - 2m\left[\boldsymbol{\omega} \times \frac{d'\mathbf{r}}{dt}\right] - m[\boldsymbol{\omega} \times (\boldsymbol{\omega} \times \mathbf{r})] \quad \ldots \quad \ldots \quad \ldots \quad (5)$$

This is zero since **r** is held constant both in magnitude and in direction with respect to the rotating coordinate system. The Coriolis force also disappears for the same reason. Thus

$$\mathbf{F} = m[\boldsymbol{\omega} \times (\boldsymbol{\omega} \times \mathbf{r})] = m\omega^2 r\mathbf{n} = m\mathbf{a} \quad \ldots \quad (6)$$

represents the force which is necessary to hold the particle in rotation and which is constantly changing in direction, and **a** gives the centripetal acceleration as shown in Fig. 1.

Now an observer in the rotating coordinate system Σ' with a centripetal acceleration, but unconscious of the force **F**, will report the existence of a gravitational field corresponding to the familiar centrifugal force

$$\mathbf{F}' = -m[\boldsymbol{\omega} \times (\boldsymbol{\omega} \times \mathbf{r})] = -m\omega^2 r\mathbf{n} \quad \ldots \quad (7)$$

This is nothing but Einstein's principle of equivalence applied to a rotating system for the observer B at any instant. The observer B in the rotating system Σ' will then use the coordinate language of his comrade observer A in the rest inertial system Σ for the description of the physical processes in addition to the presence of this gravitational field **F'**, and ascribe a constant tangential velocity

$$|\mathbf{v}| = |\boldsymbol{\omega} \times \mathbf{r}| = |\omega r\mathbf{t}_1| = \omega r \quad \ldots \quad \ldots \quad (8)$$

to the particle at any instant. In fact, the observer B will attribute his gravitational field **F'** to the cause of rotation with constant angular velocity ω if he uses the same physical and mathematical language as we use. This explains therefore the so-called "absolute property of rotation."

2. *Electromagnetic Reaction on a Rotating Magnet*

Let the observer B mounted on the rotating magnet (Σ'-system) be armed with a unit positive electric charge e. Since he is moving with a constant linear velocity ($v = \omega r$) in a stationary magnetic field produced by the magnet, he will experience an "electric force" at any instant given by[4].

$$\mathbf{F}'_e = e\mathbf{E}' = \frac{1}{\sqrt{1 - \left(\frac{v}{c}\right)^2}}[\mathbf{v} \times \mathbf{B}] \quad \ldots \quad (9)$$

in a direction perpendicular to the magnetic flux density **B** and to his direction of motion at that instant. As shown in Fig. 1, at the instant considered, the rotating co-ordinate-axes x, y, z, coincide with those of the inertial system x, y, z and the rotating system is turning about the **k**- or **k'**-axis with a constant angular velocity ω. Let the observer armed with the unit positive charge be at point P on the y-axis. Then his velocity will only have a component along the negative x-axis, that is

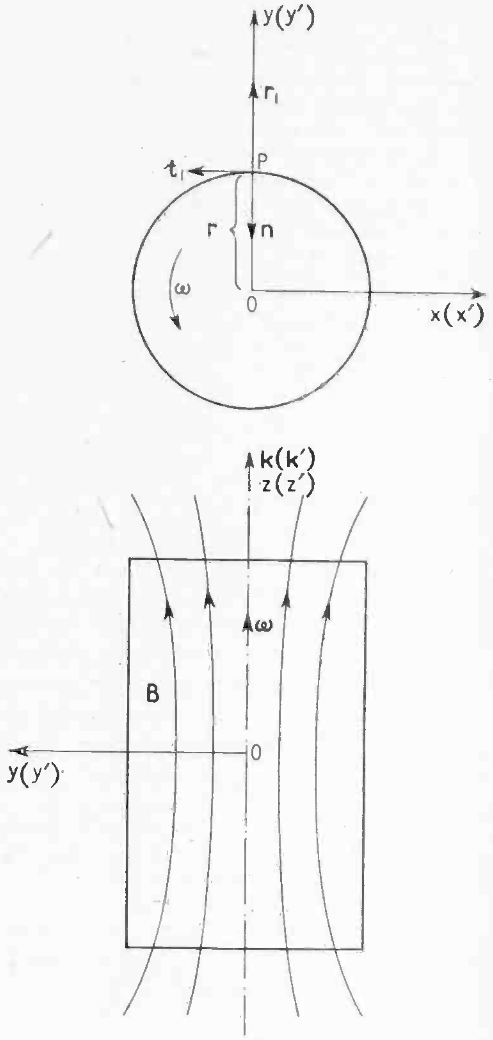

Fig. 1

$$\mathbf{V} = -v\mathbf{i} + 0\mathbf{j} + 0\mathbf{k} \quad \ldots \quad \ldots \quad \ldots \quad (10)$$

The magnetic field **B**, on the other hand, can have no component along the x-axis from circular symmetry, that is

$$\mathbf{B} = 0\mathbf{i} + B_y\mathbf{j} + B_z\mathbf{k} \quad \ldots \quad \ldots \quad \ldots \quad (11)$$

Substituting relations (10) and (11) into (9), we obtain the components for the electric field experienced by the observer B: viz.,

$$E'_x = 0,\; E'_y = \frac{vB_z}{\sqrt{1 - \left(\frac{v}{c}\right)^2}},\; E'_z = \frac{-vB_y}{\sqrt{1 - \left(\frac{v}{c}\right)^2}} \quad \ldots \quad \ldots \quad \ldots \quad (12)$$

Since the magnetic field is practically parallel to the z-axis, E'_z is negligible compared with E'_y. Thus the movement of charges inside the magnet is for the most part due to E'_y, while the small E'_z

component is effective only at both end-edges. The charge distribution thus arising can easily be calculated if we make the reasonable assumptions, to wit

$v/c \ll 1$, $B_y = 0$ and $B_z =$ constant (inside the magnet).

Relations (12) then simplify to

$$E'_x = 0,\ E'_y = vB_z = \omega r B_z,\ E'_z = 0 \quad \ldots \quad (12a)$$

From the evident circular symmetry, the field quantities are best expressed in terms of cylindrical coordinates; that is

$$\mathbf{B} = B_z\mathbf{k}\ ;\ \mathbf{E}' = E'_r\,\mathbf{r}_1 = vB_z\mathbf{r}_1 = \omega r B_z\,\mathbf{r}_1 \quad \ldots \quad \ldots \quad \ldots \quad (13)$$

The charge density produced inside the rotating magnet is then given by the divergence of the dielectric displacement; viz,

$$\rho = \text{div.}\ \mathbf{D}' = \text{div.}\ (\epsilon\mathbf{E}') = \epsilon\left[\frac{1}{r}\frac{d}{dr}(rE'_r)\right]$$

$$= \frac{e}{r}\frac{d}{dr}(\omega^2 r^2 B_z) = 2\epsilon\omega B_z \quad \ldots \quad \ldots \quad (14)$$

We thus have constant space charge density of positive charges in the magnet and a line-charge of negative charges along the axis of rotation. The existence of the positive space charge in a rotating magnet is susceptible to experimental verification by its action upon a test charge nearby. At the same time the positive space charge, rotating with the magnet, will produce a convection current strengthening the magnetic field of the magnet.

3. *Forces acting upon a Rotating Test-Charge outside a rotating Magnet*

From previous considerations it is evident that, if a test-charge $+q$ outside the magnet is rotating together with the magnet with the same angular velocity ω, it will be subjected to three electromagnetic forces, namely,

(*a*) An electrostatic Coulomb force upon the test-charge $+q$ caused by the space-charge distribution inside the magnet which itself arises from the magnet's rotation, the space-charge at any point being given by Equ. (14). This force is directed radially outward and is denoted by $\mathbf{F}_{1q}$ in Fig. 2.

(*b*) Part of the magnetic force upon the test-charge convection-current $+q\mathbf{v}_q$ due to the magnetic field produced by the rotating space-charge in the rotating magnet. This is directed radially inward and is shown in Fig. 2 as $\mathbf{F}_{2q}$.

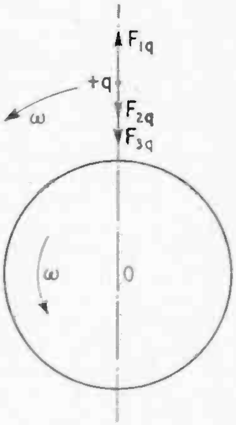

Fig. 2

(*c*) Part of the magnetic force upon the test-charge convection-current $q\mathbf{v}_q$ due to the original stationary magnetic field produced by the magnet at rest. This force is denoted by $\mathbf{F}_{3q}$ in Fig. 2 and is also directed radially inward.

The above simple considerations tend to throw some light upon the complexion of an electron moving along its orbit about the nucleus in an atom. In addition to the Coulomb force, each electron is definitely subjected to a sort of magnetic reaction due to the magnetic field produced inside an atom by other moving electrons and also the nuclear motion, especially for heavy atoms.

National Central University, Nanking.
CHANG-PEN HSÜ
(Professor of Electrical Engineering.)

REFERENCES

[1] G. W. O. Howe: *Wireless Engineer*, Vol. XXIII, Oct. 1946.
[2] Haas: "Introduction to Theoretical Physics", Vol. II.
[3] Joos: "Theoretical Physics", Chapter X.
[4] In M.K.S units; see Stratton: "Electromagnetic Theory."

Permeability of Dust Cores

To the Editor, "Wireless Engineer"

SIR,—I have read with interest the correspondence and Editorials on this subject and I believe that there is a difference of principle between toroidal cores of high permeability and toroidal cores made from the material used for high-frequency cores. The latter contain much more insulation and binder, and are pressed with less than 10% of the tonnage required for the high-permeability cores; it can be said in general that the particles in such cores will not be deformed.

Having made microscopic examinations of high-permeability toroidal cores on many occasions I have, as yet, never seen the aligned particles which Mr. Bardell claims, nor have I been able to observe his slab formation with the dimensional ratio of 10 : 2 : 1. The idea itself is not new and is contained in the British Patent Specification No. 342,999 of 1929 which claims the magnetic adjustment of oblong particles inside the mould before pressing. I assume that Polydoroff's application of the brush-production technique to cores as described in the British Patent Specification No. 557,838 may also be applicable. Originally suggested for long cores, there may be difficulties with toroids.

It has, on the other hand, occurred to me that there may be another cause for the observed increase of permeability. All formulae are based on the assumption that the particles have a reasonably regular shape and that the airgaps are evenly distributed. Every microscopic examination has so far shown two facts:

1. The very soft (annealed) material will "flow" under pressure and the boundary airgap round the now very irregular shaped particle will be about half a micron;

2. Relatively large airpockets can be observed on every section of the toroidal core under examination.

The second point probably needs some explanation. It is due to the effect known as "bridging" in powder metallurgy. The insulated powder has to be shaken into the mould, a vessel with limited space. In this process the powder is liable to form arches. The space under the arches is void of powder and the arches themselves resist pressure, retaining an airpocket in the core. The bridging depends partly on the particle size and shape, but also on the insulation and atmospheric conditions. As they are distributed unevenly, some airpockets will lie across some lines of force while others will not be affected. The effect will be that the particles will be separated by a smaller airgap than that calculated from the percentage of magnetizable material present in the core and its weight. These smaller airgaps should not only have an effect on the permeability but also on the hysteresis coefficient. This assumption seems to agree with the thread-like alignment suggested elsewhere.

There is another point : Instead of going through the airpocket the affected lines will go round it, and thus increase the induction in the particles adjacent to such pockets. An increased induction, however, (we are far from the saturation point) will create locally all those phenomena normally associated with an increased field. I wonder therefore whether this and the aforementioned consideration will not explain both the increased permeability and part of the residual loss as far as this is due to inaccuracies in our methods for loss separation.

Dr. Bardell's judgment on Legg & Given's figures stands on very weak feet. His own figures can be considered as arbitrary as he does not mention a basis on which his figures and the other ones can be compared. If the intrinsic permeability is measured with direct current it is certainly necessary to link the permeability to the current. Speed & Elmen in their 1921 paper on "Magnetic Properties of Compressed Powdered Iron" have published interesting tables on measurements made on toroids from uninsulated powder, pressed under different tonnage. They show how the permeability of such cores is affected by the magnetic test conditions (due to the considerable wider hysteresis loop) and make it unnecessary to repeat figures.

It is, however, doubtful whether either of the authors has actually measured the true intrinsic permeability in this way. All powders are annealed in a reducing atmosphere. Even if they were supplied in containers filled with inert gas (and they are not), the powder with its relatively large surface area is exposed to air during the weighing, filling of the die, levelling and closing. This is long enough, and the material is sufficiently stirred up to be oxidized. It will depend on the oxide formed and its magnetic and mechanical properties how near the measured figures are to the permeability of the powder particle. The oxide may have magnetic properties or not, it may be brittle or it may be tough enough to resist a certain pressure. It may be pierced and it may only be dented, thus altering the conductivity as compared with the particles which form direct contact. That these considerations are important is shown by the fact that oxidization by steam and water can result in reasonable insulation (Speed & Elmen).

Ferrocart A which was mentioned in Professor Howe's Editorial has very quickly been replaced by moulded grades and it is doubtful whether anything more than samples have reached this country. The replacement took place probably in 1933.

Manchester. E. R. FRIEDLAENDER.

The Equivalent Diode

To the Editor, "Wireless Engineer"

SIR,—The article by G. B. Walker on the theory of the equivalent diode in your January issue, and particularly his remark on the difficulty of obtaining experimental verification, has led me to make a comparison of the expression which he derives, that obtained by Fremlin,[1] and one due to Tellegen,[2] with the intention of finding what conditions are necessary for a critical experiment.

The expressions may all be written in the form

$$I = 2.34 \times 10^{-6}\frac{(V_a + \mu V_g)^{\frac{3}{2}}}{g^2(1 + \mu)^{\frac{3}{2}}B}$$

where, with $s = a - g$,

$$B_W = \left(1 + \frac{s}{g\,(1 + \mu)}\right)^2$$

$$B_F = \left(1 + \frac{\left(1 + \frac{s}{g}\right)^{\frac{4}{3}} - 1}{1 + \mu}\right)^{\frac{3}{2}}$$

$$B_T = \left(1 + \frac{4}{3}\frac{s}{g\,(1 + \mu)}\right)^{\frac{3}{2}}$$

Since $s/g\mu$ may be written $p \log (\coth \pi\, d/p)/2\pi g$ where p is the pitch of the grid and d the diameter of its wire, and since p/g must be less than unity to avoid "Inselbildung" $s/g\,(1 + \mu)$ depends mainly on d/p, is larger for widely-spaced fine wires, and will be less than unity for any reasonable structure. Hence B_T may be expanded by the binomial theorem

$$B_T = 1 + \frac{2s}{g\,(1 + \mu)} + \frac{2}{3}\left(\frac{s}{g\,(1 + \mu)}\right)^2 - \frac{4}{27}\left(\frac{s}{g\,(1 + \mu)}\right)^3 + \frac{2}{27}\left(\frac{s}{g\,(1 + \mu)}\right) - \ldots$$

$$B_W = 1 + \frac{2s}{g\,(1 + \mu)} + \left(\frac{s}{g\,(1 + \mu)}\right)^2$$

These expansions are remarkably similar, and differ by 10 per cent only when $s/g(1 + \mu)$ is nearly unity, so that it would be very difficult to choose between the formulæ of Walker and Tellegen experimentally.

It is not possible to expand B_F in terms of integral powers of s/g if s/g is greater than unity, but by taking particular cases it may be seen that the difference between B_F and B_W is greater than that between B_T and B_W, and that it could be made large enough to permit of experimental discrimination with reasonably fine grid wires and a reasonable pitch. The structure used by Fremlin, with $d = 0.0089$ cm, and $p = 0.175$ cm, has $B_F/B_W = 1.12$, and this ratio could be increased to 1.2 by using wire of half the diameter and increasing s. Unfortunately Fremlin's very careful experiment is vitiated, because he neglected the factor[3] $[1 + 0.024T^{\frac{1}{2}}\,(V - V_m)^{-\frac{1}{2}}]$, where T is the absolute temperature of the cathode, V is the anode voltage of the equivalent diode, and V_m is the voltage of the potential minimum, which corrects the three-halves law for the existence of initial electron velocities, and with the voltages which he used this had an average value of about 1.25. (The agreement obtained between theory and experiment was probably due to a compensating error introduced by the incorrect method of allowing for contact potential.) With an apparatus essentially similar to that used by Fremlin, but using finer grid wire and with an anode voltage of about 1,000, it would be possible to reduce the initial velocity corrections below 10 per cent for a range of grid volts and thus determine which formula best fits the facts. Special methods of cooling the anode would be needed. The fact that such a highly unusual triode would be necessary s an indication that the matter is of academic interest only.

A point of theoretical interest arises from the above work. Walker tests the validity of the formulæ he discusses in the extreme cases when μ tends to infinity or zero. Tellegen's formula fails in the latter case, though it does not tend to zero. A similar but slightly more complicated test is to consider the anode removed to a great distance. Since μ is proportional to s, $s/g(1+\mu)$ remains finite, but

$$\left(1+\frac{s}{g}\right)^{\frac{4}{3}}/(1+\mu)$$

tends to infinity.

Thus the current predicted by Walker and Tellegen tends as we would expect to something a little less than

$$2.34 \times 10^{-6} V_g/^{\frac{3}{2}}g^2,$$

whereas that given by Fremlin tends to zero.

It may be concluded that an experimental discrimination between the formulæ of Walker and Tellegen is not possible, that an experimental choice between these two and Fremlin's though possible, s not worth the trouble, but that Walker's approach is to be preferred, because it is theoretically sounder, because it is not dependent on the three-halves law and can thus be used in calculations where initial velocities are taken into account, and because it has the important virtue of simplicity.

J. EASTABROOK,
Standard Telephones and Cables, Ltd.
Ilminster, Somerset.

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[1] *Philosophical Magazine*, 1939, Vol. 27, p. 709.
Electrical Communication, July 1939.

[2] *Physica*, 1925, Vol. 5, p. 301.

[3] Langmuir I, *Physical Review*, 1923, Vol. 21, p. 419.

I.E.E. RADIO SECTION

The annual dinner of the Radio Section of the Institution of Electrical Engineers was held at the Café Royal, Regent Street, London, on 30th April.

Prof. Willis Jackson, Chairman of the Section, announced that Mr. H. J. Nunn was relinquishing his position as secretary of the Radio Section and that it would be taken over by Mr. K. W. T. Brown. A presentation was made to Mr. Nunn in appreciation of his services and it was pointed out that he had not only been secretary of the Section for fifteen years but had been associated with the Institution for forty.

INSTITUTION OF ELECTRONICS

The second exhibition of electronic apparatus will be held at the Great Hall, College of Technology, Manchester, on 22nd July at 2.30–9 p.m. and on 23rd July at 10 a.m.–9 p.m. Admission is by ticket and non-members should apply to A. Coates, 16, Didsbury Park, Manchester, 20.

NEW E.R.A. SECTIONAL COMMITTEE

A new sectional committee of the Electrical Research Association has been set up with as terms of reference the studying and developing of the electrical equipment of automotive systems with particular reference to electrical ignition systems, radio interference suppression devices and electrical accessories. Under the chairmanship of Dr. E. A. Watson, the committee comprises representatives of the Ministry of Supply, the Admiralty, the Motor Industry Research Associations, the British Internal Combustion Engine Research Association, the Department of Scientific and Industrial Research and most of the leading manufacturers of automobile electrical components.

WIRELESS PATENTS

A Summary of Recently Accepted Specifications

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

ACOUSTICS AND AUDIO-FREQUENCY CIRCUITS AND APPARATUS

581 125.—Radiating intense waves or beams of audible frequency from a plate-resonator carrying standing-waves created by a piezo-electric or magnetostrictive drive.

Marconi's W.T. Co. Ltd. and H. J. Round. Application date 7th November, 1941.

581 510.—Microphone amplifier, with a readily-adjustable reed, suitable for a deaf-aid set.

Fortiphone Ltd. and L. M. Seath. Application date 10th August, 1944.

AERIALS AND AERIAL SYSTEMS

580 569.—Aircraft aerial comprising a sphere or hemisphere of dielectric material with metallic dipoles or slot radiators.

Standard Telephones and Cables Ltd. and E. O. Willoughby. Application date 21st April, 1944.

580 750.—Directive aerial system comprising two elements, each with two diverging conductors, driven in phase-quadrature.

Marconi's W.T. Co. Ltd. (assignees of C. W. Hansell). Convention date (U.S.A.) 27th January, 1943.

581 570.—Eccentric driving-gear for rotating a parabolic reflector so that the axis of the radiated beam generates a small-angled cone in space.

E. C. Cork and M. Bowman-Manifold. Application date 15th February, 1943.

DIRECTIONAL AND NAVIGATIONAL SYSTEMS

580 481.—Rotary-drum keying arrangement for the aerials of a blind-approach radio-navigational system using ultra-short waves.

Standard Telephones and Cables Ltd. and E. O. Willoughby. Application date 11th August, 1941.

580 482.—Radiolocation equipment with means for the automatic correction of marginal errors made when following a moving target.

The British Thomson-Houston Co. Ltd. and T. H. Mackenzie. Application date 3rd April, 1942.

580 640.—Blind-landing system in which a tangential glide-path is formed by radiators located fore and aft of the landing point.
Standard Telephones and Cables Ltd. and E. O. Willoughby. Application date 5th March, 1943.

580 687.—Scanning system for televising the indications received by radiolocation equipment to a remote point.
B. J. Edwards and Pye Ltd. Application date 13th August, 1943.

580 975.—Radiolocation set in which the transmitting and receiving aerials scan the field-of-view in quadrature so as to determine two angular coordinates of the target.
Sperry Gyroscope Co. Ltd. and R. H. Nisbet. Application date 8th November, 1940.

RECEIVING CIRCUITS AND APPARATUS

(*See also under Television*)

580 402.—Reducing " shot noise " in an electronic tube utilizing velocity-modulation for superheterodyne reception.
C. S. Bull. Application date 2nd September, 1942.

580 815.—Piezo-electric discriminator circuit for detecting phase-modulated carrier-signals.
Marconi's W.T. Co. Ltd. (assignees of M. G. Crosby). Convention date (U.S.A.), 16th February, 1943.

580 905.—Time-controlled relay for switching a wireless receiver on and off at predetermined times, and a local tone-generator which allows the device to be used as an alarm-clock.
J. Carter. Application date 3rd October, 1944.

TELEVISION CIRCUITS AND APPARATUS

For Transmission and Reception

580 413.—Television receiver of the kind in which a scanning-beam varies the opacity of a fluid layer, thereby controlling the passage of light through it from an external lamp.
Gesell. zur Forderung & Technischen Hochschule. Convention date (Switzerland) 5th June, 1943.

580 860.—Cathode-ray tube designed to secure a more sensitive control of the electron beam for line-scanning than for frame-scanning, say in television.
L. F. Broadway. Application date 9th March, 1944.

580 887.—Television system in which pulsed voltages are applied to the c.r. signal-developing tube, to increase its operating efficiency.
H. G. Lubszynski. Application date 16th March, 1944.

SIGNALLING SYSTEMS OF DISTINCTIVE TYPE

580 738.—Generating a frequency-modulated signal from a time-modulated train of pulses, the carrier-wave being developed from an harmonic of the train.
Standard Telephones and Cables Ltd. and C. T. Scully. Application date 23rd April, 1943.

580 769.—Generating pulse-trains with time-phase and time-duration modulation, wherein the unmodulated interval is short compared with the repetition period of the pulses.
Standard Telephones and Cables Ltd., P.K. Chatterjea and C. T. Scully. Application date 2nd May, 1944.

580 801.—Generating pulses at a frequency of say 15 kc/s, each pulse occupying only 0.3 per cent of the repetition time.
Standard Telephones and Cables Ltd. and W. A. Beatty. Application date 27th February, 1940.

580 843.—Secret signalling-system based upon the rise of an amplitude-modulated train of pulses in combination with a masking signal.
Standard Telephones and Cables Ltd., P. K. Chatterjea and L. W. Houghton. Application date 18th December, 1941.

580 930.—Multiple wired-wireless system in which each transmitter is tuned, to serve as a wave-trap to shut-out intruding signals.
The British Thomson-Houston Co. Ltd. Convention date (U.S.A.) 27th July, 1943.

CONSTRUCTION OF ELECTRONIC-DISCHARGE DEVICES

580 164.—Electrode arrangement and deflecting means for generating very short waves in a valve of the beam type.
Marconi's W.T. Co., Ltd., and C. S. Franklin. Application date 13th March, 1941.

580 730.—Cathode-ray tube in which the electron beam is arranged to produce an illuminated area with a dark centre-spot to increase the accuracy of definition.
The General Electric Co. Ltd. and L. C. Jesty. Application date 14th September, 1942.

580 964.—Hollow resonator, as used for velocity-modulation, provided with a tuning device in the form of a rotary vane.
N. C. Barford. Application date 7th September, 1943.

SUBSIDIARY APPARATUS AND MATERIALS

579 940.—Rotary transformer and balanced-bridge arrangement for stabilizing the frequency of an electric power-supply system.
F. R. Milsom, Furzehill Laboratories, Ltd. and S. Smith & Sons (Motor Accessories), Ltd. Application date 14th March, 1944.

580 044.—Valve relay circuit operated by the movement of a magnetic core ; e.g., for controlling liquid levels or pressures.
Igranic Electric Co., Ltd., J. M. Bedford, J. A. Field and S. P. Maynard. Application date 24th January, 1944.

580 377.—Construction of a flexible waveguide of rectangular cross-section.
C. E. Fenwick and C. S. Wright. Application date 28th March, 1944.

580 527.—Integrating circuit in which negative reaction is utilized to increase the normal time-constant : applicable for generating saw-tooth voltages.
A. D. Blumlein. Application date 5th June, 1942.

580 824.—Crystal-controlled oscillation-generator in which an auxiliary inductance serves to offset the normal falling-off in amplitude of the higher harmonics.
Marconi's W.T. Co. Ltd. and G. L. Grisdale. Application date 8th May, 1944.

ABSTRACTS AND REFERENCES

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

ACOUSTICS AND AUDIO FREQUENCIES

4.213.4 + 621.392.029.64 **1655**

Acoustic and Electromagnetic Wave Guides of nplicated Shape.—Krasnooshkin. (*See* 1666.)

4.756 + 621.39 **1656**

Theory of Communication.—D. Gabor. (*J. Instn ct. Engrs*, Part I, Jan. 1947, Vol. 94, No. 73, 58.) Summary of 1057 of April.

.85 + 621.395.625 **1657**

ound Recording and Reproduction.—R. V. thev. (*Proc. Instn Radio Engrs, Aust.*, Dec. 6, Vol. 7, No. 12, pp. 4–7.)

.314.2.029.3 **1658**

esponse of Audio-Frequency Transformers at h-Frequencies.—Webb. (*See* 1688.)

.395.61 **1659**

icrophones and Receivers.—L. C. Pocock. *Brit. Instn Radio Engrs*, June/Aug. 1943, 3, No. 5, pp. 197–215. Discussion, pp. 215–.) A comprehensive paper dealing with (*a*) surement of speech by means of a speech rgy meter, which integrates electrical speech er by storing it as heat in a valve cathode, thus nging anode current, (*b*) the speech power to delivered to a loudspeaker to produce a given reading on a noise meter in a standard position, (*c*) various kinds of microphones and their frequency characteristics, and the disturbance of free field due to the introduction of a microphone or of different listeners, (*d*) coupling between transducers and amplifying equipment, (*e*) distortion, including that due to a wrong level of reproduction, (*f*) equivalent networks, allowing for the dependence of acoustic impedances upon air density and pressure, (*g*) force factor, which is a measure of the coupling between the acoustic or mechanical side of the apparatus and the electrical or electromagnetic side, (*h*) future developments.

621.395.625 : 621.396.611.1 : 621.396.619 **1660**

Resonant Circuit Modulator for Broad-Band Acoustic Measurements.—Hull. (*See* 1702.)

621.395.625.3 **1661**

Magnetic Tape Recorder.—(*See* 1832.)

621.396.645.029.3 **1662**

High-Impedance Input Circuits for A.-F. Service.—Parry. (*See* 1715.)

AERIALS AND TRANSMISSION LINES

621.315.2.029.5 : 621.317.333.4 **1663**

New Methods for Locating Cable Faults, particularly on High-Frequency Cables.—F. F. Roberts. (*J. Instn elect. Engrs*, Part I, Jan. 1947, Vol. 94, No. 73, pp. 61–63.) Summary of 1002 of April.

621.315.212 **1664**

Determination of the Optimum Ratio of Conductor Diameters in Coaxial Cables.—K. O. Schmidt. (*Elektrotech. Z.*, 4th May 1944, Vol. 65, Nos. 17/18, pp. 170–173.) Methods in common use require too much copper. Expressions are derived which permit accurate calculation of the diameter ratio. Sets of curves show the correctness of the formulae used and enable the optimum ratio to be found for particular cables when the attenuation per km is given. The saving in copper, as compared with the usual methods, may be of the order of 20–30%.

621.315.212.2 : 621.317.336 **1665**

New Method of Measuring the Impedance Errors of Concentric Pairs.—Fuchs. (*See* 1848.)

621.392.029.64 + 534.213.4 **1666**

Acoustic and Electromagnetic Wave Guides of Complicated Shape.—P. Krasnooshkin. (*J. Phys., U.S.S.R.*, 1946, Vol. 10, No. 5, pp. 434–445.) Waves in hollow pipes of rectangular section and complicated shape are discussed. An equivalence theorem is established relating a pipe of complicated shape to a straight cylindrical pipe filled with a non-uniform medium. Pipes having parabolic, elliptic and toroidal forms are discussed and reference is made to tunnel effect, band-pass effect and 'clinging phenomenon'.

621.392.029.64 **1667**

The Effect of the Curvature of a Waveguide on Propagation.—M. Jouguet. (*C. R. Acad. Sci., Paris*, 26th Aug. 1946, Vol. 223, No. 9, pp. 380–381.) To a first approximation the curvature of a waveguide does not affect the phase velocity. For relatively great curvatures, an expression is here derived for the change of propagation constant in the case of $H_{n,o}$ waves in a rectangular curved guide and from this the change of phase velocity, group velocity and wavelength are at once obtained. The expression has no meaning if the frequency is equal to the cut-off frequency of the guide for zero curvature. Formulae for the field components and cut-off frequency are given which should be used near the cut-off frequency, in place of those previously given for the general case. See also 1320 of May and back references.

621.392.029.64 **1668**

The Effect of a Curvature Discontinuity on Propagation in Waveguides.—M. Jouguet. (*C. R. Acad. Sci., Paris*, 23rd Sept. 1946, Vol. 223, No. 13, pp. 474–475.) The case is considered of an $H_{p,o}$ wave propagated in a guide of rectangular cross-section whose axis is curved in a plane perpendicular to the electric field. At a point where the curvature changes suddenly, parasitic waves and oscillations, all of the form $H_{n,o}$, occur in the two portions of the guide. Formulae are given for the ratio of the complex amplitudes of the perturbations in the two parts of the guide (*a*) for the general case of any frequency and (*b*) where the frequency is very near the cut-off frequency of the wave $H_{q,o}$, $(q + p)$ being odd. A loss factor, the ratio between the energy of the parasitic waves and that of the wave $H_{p,o}$, can be calculated. Owing to the abnormal excitation of the parasitic waves near the cut-off frequency, the loss factor shows a sharp rise. See also 1667 above.

621.392.029.64 **1669**

Rectangular Waveguide Systems.—N. Elson. (*Wireless Engr*, Feb. 1947, Vol. 24, No. 281, pp. 44–54.) The mismatch at a discontinuity in a waveguide (expressed in terms of the percentage amplitude reflection coefficient) is determined for a chopped-off corner of various angles in the E and H planes at λ 10.8 cm, and an optimum design is deduced. Curves for the s.w.rs of various bends at λ 3.2 ± 0.1 cm are given. An analysis of the equivalent circuit of T- and Y-junctions is given and the results presented graphically. An appendix treats theoretically the location of the plane of minimum transverse current in a curved waveguide to determine the best position for a cut.

621.392.029.64 **1670**

Some Applications of the Principle of Variation of Wavelength in Wave Guides by the Internal Movement of Dielectric Sections.—G. E. Bacon & J. C. Duckworth. (*J. Instn elect. Engrs*, Part IIIA, 1946, Vol. 93, No. 1, p. 56.) Summary of 1321 of May.

621.392.029.64 : 538.3 **1671**

Quasi-Stationary Field Theory and Its Application to Diaphragms and Junctions in Transmission Lines and Wave Guides.—G. G. Macfarlane. (*J. Instn elect. Engrs*, Part IIIA, 1946, Vol. 93, No. 1, pp. 63–64.) Summary of 1323 of May.

621.392.029.64 : 621.318.572 **1672**

The Rhumbatron Wave-Guide Switch.—A. Maclese & J. Ashmead. (*J. Instn elect. Engrs*, Part IIIA, 1946, Vol. 93, No. 1, p. 65.) Summary of 1325 of May.

621.392.029.64 : 621.396.662.3 **1673**

Filtering of Guided Waves.—J. Ortusi. (*Bull. Soc. franç. Élect.*, Nov. 1946, Vol. 6, No. 63, pp. 589–596.) A theoretical treatment is given of simple cavity resonators and of the effect in a waveguide of an obstacle formed by a conducting plate with a window. Two such windows in a waveguide constitute the simplest band-pass filter; a band-stop filter is obtained by means of a resonator coupled to the guide through an aperture in the wall. Band-pass filters are usually formed by a number of elementary cells with suitable coupling. Examples are given of the results obtained when four or more windows are used, with a description of the apparatus used to verify the theory.

621.392.029.64.091 **1674**

Calculation of Attenuation in Wave Guides.—S. Kuhn. (*J. Instn elect. Engrs*, Part IIIA, 1946, Vol. 93, No. 1, pp. 61–63.) Summary of 1328 of May.

621.396.67.029.62 **1675**

A Stacked Array for 6 and 10 [metres].—E. P. Tilton. (*QST*, Feb. 1947, Vol. 31, No. 2, pp. 38–41 . . 130.) Structural details and tuning procedure are given for a three-element array for 10 m, matched by means of a T section, and for a four-element 6-m array using a folded dipole, both being fed by 300-Ω lines.

621.396.671 : 621.396.611.33 **1676**

Matching Ranges [plages d'adaptation] **of Transmitters.**—Glazer & Familier. (*See* 1958.)

621.396.677 **1677**

Slotted-Cylinder Antenna.—E. C. Jordan & W. E. Miller. (*Electronics*, Feb. 1947, Vol. 20, No. 2, pp. 90–93.) A discussion of the properties with polar diagram and impedance data.

621.396.677 **1678**

Rhombic Aerials and Matching Circuits for Reception.—T. S. Rangachari, B. H. Paranjpye & G. S. Deshpande. (*Electrotechnics*, Dec. 1946, No. 19, pp. 19–31.) The practical design of these aerials to meet various common requirements and restrictions is discussed. The construction of a non-inductive weatherproof terminating resistor and of a matching transformer is described, with the transformer adjustment procedure required to eliminate reflection.

621.396.677 : 621.317.7 **1679**

Radio-Frequency Measurements on Rhombic Antennae.—Christiansen, Jenvey & Carman. (*See* 1857.)

621.396.677.029.58 **1680**

The Double Triplex Beam.—J. A. Biggs. (*CQ*, Jan. 1947, Vol. 3, No. 1, pp. 27–30.) Constructional details and performance of a directive array for use on long range amateur radio-telephony in the 20 m wavelength band.

621.396.677.029.63 **1681**

A Switched-Beam Directive Aerial on 600 Mc/s.—R. V. Alred. (*J. Instn elect. Engrs*, Part IIIA, 1946, Vol. 93, No. 2, pp. 411–422.) The aerial has a cylindrical, parabolic reflector fed by a line of dipoles parallel to its axis. Particular problems in design for horizontal polarization, ease of production and installation aboard ships are discussed. A phasing system and mechanism for rapid switching of the direction of the beam are described, with an electrical method for correcting mechanical and electrical variations in production. A summary of this paper is given in Part IIIA, 1946, Vol. 93, No. 1, pp. 322–323.

621.396.677.029.64 **1682**

A Dielectric-Lens Aerial for Wide-Angle Beam Scanning.—F. G. Friedlander. (*J. Instn elect. Engrs*, Part IIIA, 1946, Vol. 93, No. 1, pp. 53–54.) Summary of 1358 of May.

621.396.677.029.64 **1683**

A Detailed Experimental Study of the Factors nfluencing the Polar Diagram of a Dipole in a 'arabolic Mirror.—E. G. Brewitt-Taylor. (*J. Instn lect. Engrs*, Part IIIA, 1946, Vol. 93, No. 1, p. 57.) ummary of 1357 of May.

21.396.677.029.64 : 621.392.029.64 **1684**

Directive Couplers in Wave Guides.—M. Surdin. (*J. Instn elect. Engrs*, Part IIIA, 1946, Vol. 93, No. 1, . 66.) Summary of 1359 of May.

21.396.677.029.64 : 621.392.029.64 **1685**

Resonant Slots.—W. H. Watson. (*J. Instn elect. ngrs*, Part IIIA, 1946, Vol. 93, No. 1, p. 67.) ummary of 1360 of May.

21.396.679.4 : 621.315.24 **1686**

A Six-Wire Transmission Line Application.—C. Wadsworth. (*Proc. Instn Radio Engrs, Aust.*, ec. 1946, Vol. 7, No. 12, pp. 7–11.) The required aracteristics for feeding r.f. energy to a vertical rial are obtained by a six-wire unbalanced transission line, which has also the advantages of a ncentric line.

1.396.96 : 621.396.82 **1687**

Radar Reflections from Long Conductors.—F. och, M. Hamermesh & M. Phillips. (*J. appl. ys.*, Dec. 1946, Vol. 17, No. 12, pp. 1015–1020.) hen rolls of metallized strip are to be released aircraft to give rise to strong back-scattering, e effective cross-section is of importance. The lection pattern of a straight cylindrical conctor, of great length and small diameter comred to λ, has a sharp lobe perpendicular to its gth. In practice a rope-like conductor deviates m this ideal condition and the effective crosstion becomes nearly independent of the inntaneous shape and orientation of the rope. oss-sections have been derived assuming the ape to be a series of (*a*) helical turns (*b*) straight tions whose parameters are such that the ttering adds incoherently. The two assumptions e similar results.

The cross-section of a thin twisted ribbon contor whose width d exceeds 0.75λ is almost the same for both parallel and transverse polarization. As d is decreased below 0.75λ, the cross-section for transverse polarization passes through a small maximum at $d = 0.5\lambda$ and then falls away rapidly.

CIRCUITS AND CIRCUIT ELEMENTS

621.314.2.029.3 **1688**

Response of Audio-Frequency Transformers at High-Frequencies.—E. K. Webb. (*Proc. Instn Radio Engrs, Aust.*, Jan. 1944, Vol. 5, No. 3, pp. 3–12.) The response of loaded and unloaded audio frequency transformers at high audio frequencies is analysed and nomograms for numerical calculations are given. Constructional characteristics and the causes and effects of losses present are examined. Measurements for an interstage transformer show the order of accuracy to be expected in practice.

621.314.3 **1689**

Hum in High-Gain Amplifiers.—P. J. Baxandall. (*Wireless World*, Feb. 1947, Vol. 53, No. 2, pp. 57–61.) The hum level in an a.c. mains-operated audio amplifier can be made negligible in comparison with the Johnson noise, for an input grid circuit having an impedance of the order of 50 kΩ. Causes of hum and practical means of eliminating it are discussed.

621.316.726.078.3 : 621.396.615.14 **1690**

Electronic Frequency Stabilization of Microwave Oscillators.—R. V. Pound. (*Rev. sci. Instrum.*, Nov. 1946, Vol. 17, No. 11, pp. 490–505.) Frequency control is achieved by feeding back to the oscillator a control voltage derived from an external cavity resonator coupled to the circuit by 'Magic Tee' waveguide sections to form a frequency discriminator; the cavity has a frequency-dependent reflection coefficient going through zero at resonance. Two particular control circuits of this type are described: the first uses two 'Magic Tee' sections and the reflected wave controls the relative outputs from two crystals, one in each section: the outputs are applied to the oscillator via a suitable low-pass amplifier. The second circuit uses a single 'Magic Tee' in which the reflected wave from the cavity goes to a crystal having an applied i.f. voltage. This crystal reflects amplitude-modulated sidebands with i.f. spacing on either side of the signal frequency. These mix with the signal frequency in a second crystal, using the i.f. oscillator as reference, and give a voltage which is amplified and applied to the microwave oscillator.

621.317.727 **1691**

Potentiometers.—L. A. Nettleton & F. E. Dole. (*Rev. sci. Instrum.*, Oct. 1946, Vol. 17, No. 10, pp. 356–363.) Discusses methods used to improve performance, including the use of 'Paliney' (an alloy of Pt, Pd, Au, Ag, Cu and Zn) wire for the contactor.

621.318.323.2.042.15 **1692**

Permeability of Dust Cores.—G.W.O.H. (*Wireless Engr*, Feb. 1947, Vol. 24, No. 281, pp. 33–34.) The high effective permeability of dense dust cores can be attributed to the alignment of non-spherical particles in the direction of the magnetic field, though the cause of this alignment is obscure.

621.318.323.2.042.15 **1693**

Permeability of Dust Cores.—P. R. Bardell.

(*Wireless Engr*, Feb. 1947, Vol. 24, No. 281, p. 63.) Measured permeabilities can be explained by assuming that the particles are uniformly coated rectangular slabs (dimension ratio 10 : 2 : 1) of permeability 1 000 and oriented with the long axis parallel to the magnetic field.

621.318.371.011.2/.4 **1694**

H.F. Resistance and Self-Capacitance of Single-Layer Solenoids.—R. G. Medhurst. (*Wireless Engr*, Feb. & March 1947, Vol. 24, Nos. 281 & 282, pp. 35–43 & 80–92.) Results of measurements on 40 coils wound with copper wire on grooved distrene formers. The resistances agree well with the values given by Butterworth's theory for widely spaced turns but are considerably lower for closely spaced turns. The magnification of a coil of mean radius R at frequency f can be represented by $Q = 0.15\,R\,\Psi\sqrt{f}$, where Ψ is a tabulated function of the length/diameter and spacing ratios. The self-capacitance is given by $C = HD$ μμF where D is the mean coil diameter and H is a tabulated function of the length/diameter ratio.

621.318.4.042.15 **1695**

Dust Ring-Core Coils and Their Possible Applications.—E. Ganz. (*Brown Boveri Mitt.*, Aug. 1946, Vol. 33, No. 8, pp. 219–221.) Describes some of the stock sizes of Brown Boveri dust rings and pots and discusses their use for frequencies up to 2 or 3 Mc/s.

621.392.52 + 537.228.1 **1696**

Progress in the Construction of Crystal Filters.—H. Biefer, H. Keller & B. Matthias. (*Brown Boveri Mitt.*, Aug. 1946, Vol. 33, No. 8, pp. 214–218.) The temperature coefficient of crystals of potassium and ammonium phosphate limits their application, but mixed crystals obtained by replacing ammonium by rubidium or thallium have a much lower temperature coefficient at room temperature. Methods of using length, thickness and bending vibrations for various frequency ranges are described and the properties of bridge-type filters illustrated. The simultaneous use in such filters of potassium and ammonium phosphate crystals gives very wide pass bands with sharp cut-off.

621.392.52 **1697**

Insertion Characteristics of Filters.—J. B. Rudd. (*A. W. A. tech. Rev.*, Dec. 1946, Vol. 7, No. 2, pp. 145–176.) Expressions are obtained for the insertion loss L and phase shift B of multi-sectioned low-, high- and band-pass filters of 'constant k' form, terminated in the design resistance. The 'unit current' method is used, and it is shown that with proper choice of the frequency variable K, the expressions for L and B are functions of K identical for all three types of filter. The values of L and B at cut-off frequencies are given and the general behaviour of these quantities with respect to K is tabulated. Values of L are plotted as a function of K for filters containing up to five sections and compared with an experimental curve for a five-section low-pass filter. Phase shift curves for one-, two- and four-section filters are also shown. Two methods of examining the effects of dissipation in the filter are considered. The preferred method assumes that all inductor and capacitor elements have equal dissipation factors, and gives the dissipation at any frequency at which the slope of the phase shift characteristic is known.

621.392.52.094 **1698**

Distortion of Frequency-Modulated Signals by Band-Pass Filters.—P. Güttinger. (*Brown Boveri Mitt.*, Aug. 1946, Vol 33, No. 8, pp. 185–187.) Quasi-stationary methods are applied to the calculation of the distortion due to passage through two-stage band-pass filters. A practical formula is given for the noise factor.

621.394.652 : 621.394.141 **1699**

The Electroplex — a New Automatic Key.—J. T. Dixon. (*Radio News*, Jan. 1947, Vol. 37, No. 1, pp. 38–39 . . 151.) A system comprising a timing circuit, control circuit, a.f. monitoring oscillator and power supply enabling dots and dashes of a given length to be formed automatically for radio-telegraphy.

621.396.611.1 + 534.112 **1700**

Self-Maintenance of Several Oscillations on the Same Wire.—Jouty & Rocard. (*See* 1728.)

621.396.611.1 **1701**

Duality of the Mechanisms of Self-Oscillation.—Y. Rocard. (*C. R. Acad. Sci., Paris*, 23rd March 1942, Vol. 214, No. 12, pp. 601–603.) The introduction of a negative resistance into an oscillatory circuit is not the only means of producing self-oscillation. An alternative process is termed "confusion of the two natural frequencies". An example is that of two resonating circuits coupled by mutual induction. The voltage at the terminals of the second circuit capacitor feeds an amplifier which, through a very small resistance, supplies in the first circuit an e.m.f. in proportion but in opposition. Self-oscillation is obtained in this case by adjustment of the amplifier gain.

621.396.611.1 : 621.396.619 : 621.395.625 **1702**

Resonant Circuit Modulator for Broad Band Acoustic Measurements.—G. F. Hull, Jr. (*J. appl. Phys.*, Dec. 1946, Vol. 17, No. 12, pp. 1066–1075.) The theory is outlined and an experimental recorder described having a uniform response between 0.5 and 1 000 c/s. Apart from a 5-db resonance peak at 7 500 c/s, the response between 0.025 and 10 000 c/s is constant to within ± 3 db.

621.396.611.1.013.62 **1703**

On Self-Excitation of Electric Systems with Distributed Parameters.—S. Gvosdover. (*J. Phys., U.S.S.R.*, 1946, Vol. 10, No. 5, pp. 481–488.) Formulae are derived for determining the conditions for self-excitation and the amplitude and frequency of the steady state oscillations in terms of the constants of the system.

621.396.611.21 : 621.396.615.029.5 **1704**

Crystal-V.F.O. Mixing.—W. A. Sparks. (*Short Wave Mag.*, Nov. 1946, Vol. 4, No. 9, pp. 554–555.) Details of a circuit for mixing the output of a 6 040 kc/s crystal oscillator with that of a variable frequency oscillator covering the range 1 000–1 500 kc/s to obtain a variable frequency in the range 7 040–7 540 kc/s.

621.396.611.21 : 621.396.62 **1705**

Receiver with Automatic Tuning and Quartz Control.—Martin. (*See* 1912.)

621.396.611.4 **1706**

Resonant Frequencies of the Nosed-In Cavity.—E. Mayer. (*J. appl. Phys.*, Dec. 1946, Vol. 17, No. 12, pp. 1046–1055.) The resonant frequency of a cylindrical cavity with a smaller coaxial re-entrant cavity at one end is studied theoretically using the Ritz variational method. Results on an experimental model differed from theoretical values by about 2.5%, because a slight departure from the theoretical shape was necessary for tuning purposes. Results for several sets of parameters are shown graphically.

21.396.611.4 : 535.214 **1707**

Perturbations and Radiation Pressure in Electromagnetic Cavities.—T. Kahan. (*C. R. Acad. Sci., Paris*, 13th Nov. 1946, Vol. 223, No. 20, pp. 785–786.) From the formulae previously given (2521 of 1946) the perturbed natural frequency of a circular cylindrical cavity is calculated. The general xpression is used to find the effect of inserting axially a small metal piston and also the radiation ressure exerted on a small metal diaphragm placed oaxially at the centre of the cavity.

21.396.611.4 : 621.384.6 **1708**

The Study of a Certain Type of Resonant Cavity nd Its Application to a Charged Particle Accelerator.—E. S. Akeley. (*J. appl. Phys.*, Dec. 1946, Vol. 17, o. 12, pp. 1056–1060.) A linear accelerator consting of a number of cavity resonators placed end end, with the inside end plates removed, is iscussed. To determine possible shapes for each vity, the stationary TM_{01} modes between two arallel plates are found. When only one mode is xcited, the radius of the cavity becomes infinite the plates when the phase velocity is less than that light. By exciting simultaneously two modes aving suitable relative amplitudes, the radius can made finite everywhere.

21.396.615.029.5 **1709**

H.F. Beat-Frequency Oscillator.—R. Lemas. *Radio en France*, 1947, No. 1, pp. 4–8.) An oscillator a fixed frequency F_1, which may be modulated ther in amplitude or frequency, is coupled to the id g_1 of a mixer and a second c.w. oscillator, of riable frequency F_2, is coupled to the grid g_3 of e mixer. The difference frequency $F_2–F_1$ is lected and its voltage measured by a valve voltmeter. The amplitude of the output is directly oportional to the amplitude applied to g_1, which controlled by an attenuator. Details of a fivege instrument are given.

1.396.615.17
1.317.39 : 531.76 **1710**

Some Precision Circuit Techniques used in Waverm Generation and Time Measurement.—B. ance. (*Rev. sci. Instrum.*, Oct. 1946, Vol. 17, 10, pp. 396–415.) An outline of the characteres and uses of nonlinear circuit elements in tems developed for distance measurement, nputation and timing in radar circuits. The neration of sinusoidal and other waveforms is scribed, with diagrams of the circuits employed, the development of electronic switch circuits modulation and demodulation of rapidly varying veforms is discussed.

.396.615.17 : 621.317.755 **1711**

A Linear Time Base of Wide Range.—D. F. Gibbs W. A. H. Rushton. (*J. sci. Instrum.*, Nov. 1946, Vol. 23, No. 11, pp. 270–271.) Circuit details for an oscillograph timebase having sweep times from 0.6 ms to 60 sec and delivering a sweep voltage of 350 V. The timebase can also be set for single sweeps.

621.396.619.13 **1712**

Frequency Modulator.—Bruck. (*See* 1960.)

621.396.645 **1713**

A Stabilized 813 Amplifier.—R. M. Smith. (*QST*, Feb. 1947, Vol. 31, No. 2, pp. 23–27, 128.) Uses an 813 beam tetrode with neutralization. Readjustment of the neutralizing capacitor is not necessary when changing frequency bands.

621.396.645 : 535.215 **1714**

Amplification of Very Feeble Photoelectric Currents.—A. Blanc-Lapierre. (*C. R. Acad. Sci., Paris*, 30th March 1942, Vol. 214, No. 13, pp. 660–662.) A brief discussion of various methods.

621.396.645.029.3 **1715**

High-Impedance Input Circuits for A.-F. Service.—C. A. Parry. (*Proc. Instn Radio Engrs, Aust.*, Dec. 1940, Vol. 4, No. 5, pp. 73–75.) The effects of an unbypassed cathode resistor on the input impedance of an RC amplifier are analysed and formulae derived. Noise voltages developed in the resistor and methods of neutralizing hum pickup are discussed.

621.396.645.36 **1716**

150 Watts Push-Pull.—L. H. Thomas. (*Short Wave Mag.*, Dec. 1946, Vol. 4, No. 10, pp. 598–602.) Construction and operational details of a power amplifier of simple design, using pentode valves, for the 14 and 28 Mc/s amateur bands.

621.396.645.37.012.3 **1717**

Three Feedback Amplifier Charts.—J. S. Wells. (*Proc. Instn Radio Engrs, Aust.*, Nov. 1946, Vol. 7, No. 11, pp. 4–7.) These give the gain, phase shift and improvement in gain stability with feedback, as functions of A for various values of θ, A being the magnitude and θ the phase angle of the gain from the input to the feedback terminals.

621.396.662.21.076.2 **1718**

Coaxial Coils for F.M. Permeability Tuners.—W. J. Polydoroff. (*Radio, N.Y.*, Jan. 1947, Vol. 31, No. 1, pp. 9–10..32.) Requirements of coil design and windings for adaptation of permeability tuning to the new f.m. band.

621.396.662.3 : 621.392.029.64 **1719**

Filtering of Guided Waves.—Ortusi. (*See* 1673.)

621.396.662.34 **1720**

On the Extension of Band-Pass Effect at High Frequencies.—Miss Rajeswari & S. P. Chakravarti. (*Electrotechnics*, Dec. 1946, No. 19, pp. 52–63.) For h.f. wave-filters terminated in thermionic negative impedances. See also 1721 below.

621.396.662.34 : 621.396.611.21 **1721**

On Calculations relating to Band-Pass Effect in Crystal Resonator associated with Thermionic Negative Impedance Element.—S. P. Chakravarti & Miss Rajeswari. (*Electrotechnics*, Dec. 1946, No. 19, pp. 32–40.) The series and parallel connexions of crystal and thermionic element in both the tuned and the detuned conditions are considered theoretically. Previous experimental results are confirmed (1032 and 3267 of 1941 and back references, and 1720 above).

621.396.665 : 518.4 **1722**

A.V.C. Calculations : Graphical Methods of estimating the Performance of Circuits.—S. W. Amos. (*Wireless World*, Feb. 1947, Vol. 53, No. 2, pp. 46–50.)

621.396.69 + 621.396.621 **1723**

New Methods of Radio Production.—Sargrove. (*See* 1913.)

621.396.69 **1724**

Midget Electronic Equipment.—F.R. (*Electronics*, Feb. 1947, Vol. 20, No. 2, pp. 84–89.) Design data, available types of components and possible applications are given for equipment small enough to be carried in a suit pocket.

621.392.52 + [548.0 : 538.3 **1725**

Wave Propagation in Periodic Structures: Electric Filters and Crystal Lattices. [Book Review]—Brillouin. (*See* 1907.)

621.396.611.4.029.62 **1726**

Der Frequenzstabile Schwingtopf-Generator. [Thesis]—A. Braun. Verlag A. G. Gebr. Leemann, Zürich, 79 pp., 7.50 Swiss francs. (*Wireless Engr*, Feb. 1947, Vol. 24, No. 281, p. 64.) " It deals with the use of a cavity resonator as the oscillatory circuit of a valve oscillator at frequencies of about 200 Mc/s, with special reference to the constancy of frequency and to the optimum design of the cavity resonator."

GENERAL PHYSICS

531.18 : 531.15 **1727**

Is Rotation Relative or Absolute ?—P. M. C. Lacey. (*Wireless Engr*, Feb. 1947, Vol. 24, No. 281, p. 63.) Further correspondence on 3564 of 1946 (G.W.O.H.). See also 390 of February.

534.112 + 621.396.611.1 **1728**

Self-Maintenance of Several Oscillations on the Same Wire.—R. Jouty & Y. Rocard. (*Rev. sci., Paris*, 15th Sept. 1946, Vol. 84, No. 3257, pp. 283–285.) A steel wire can be made to oscillate at several widely different frequencies simultaneously by means of an electromagnet whose coil is connected to an amplifier, the output from which passes through the wire. Frequency ratios of about 1 to 10 can be obtained, the higher frequency not being an exact harmonic of the lower. This result differs from van der Pol's for two coupled tuned circuits made self-oscillating by a triode, in which case gradual transfer from one natural frequency to the other was possible, but not simultaneous excitation of both.

534.756 + 621.39 **1729**

Theory of Communication.—D. Gabor. (*J. Instn elect. Engrs*, Part 1, Jan. 1947, Vol. 94, No. 73, p. 58.) Summary of 1057 of April.

535.23.08 + 621.317.794 **1730**

The Production of Film Type Bolometers with Rapid Response.—C. B. Aiken, W. H. Carter, Jr, & F. S. Phillips. (*Rev. sci. Instrum.*, Oct. 1946, Vol. 17, No. 10, pp. 377–385.) The construction is described of small, rapid-response bolometers consisting basically of a thin gold strip and blackening material deposited successively by evaporation on a cellulose nitrate film. Methods of testing during construction for noise and sensitivity are explained and it is shown that results are improved by the use of metal back plates and a gas pressure of 20 mm of nitrogen.

535.338 **1731**

The Molecular Beam Magnetic Resonance Method. The Radiofrequency Spectra of Atoms and Molecules.—J. B. M. Kellogg & S. Millman. (*Rev. mod. Phys.*, July 1946, Vol. 18, No. 3, pp. 323–352.)

535.343.4 : 538.56.029.64 **1732**

Rotational Spectra of Some Linear Molecules near 1-cm Wave-Length.—C. H. Townes, A. N. Holden & F. R. Merritt. (*Phys. Rev.*, 1st Jan. 1947, Vol. 71, No. 1, p. 64.) A technique is used similar to that noted in 1309 of May for ammonia. Measured frequencies and intensities of lines are tabulated for BrCN, ClCN, and OCS, and moments of inertia and nuclear bond distances are computed. The structure of these molecules is discussed in the light of the results obtained.

537 + 538].081 **1733**

Dimensions and Units of Electromagnetic Quantities.—G. J. Baker. (*Geophys.*, July 1946, Vol. 11, No. 3, pp. 373–384.) A historical discussion, with a conversion table for various systems. The adoption of the m.k.s. system of electromagnetic units by physicists and engineers is urged.

537.122 **1734**

The Characteristics of the Electron.—H. Ebrall. (*Proc. Instn Radio Engrs, Aust.*, April 1944, Vol. 5, No. 4, pp. 3-11.) A historical survey including brief reviews of quantum theory, work function, photoelectric and thermionic emission, shot effect and filament materials.

537.122 : 530.12 **1735**

The Motional Mass of the Electron.—C. A. Boddie. (*Elect. Engng, N.Y.*, Jan. 1947, Vol. 66, No. 1, pp. 45–60.) The interaction of a moving electron and a deflecting electric field is reconsidered *ab initio*. An electric field is shown to have lateral inertia and elasticity as well as the well-known longitudinal tension and lateral pressure ; a velocity of propagation identical with the velocity of light is deduced. The observed change in the ratio of electron charge to mass, previously attributed to change in mass, is shown to be due to a motional effect of the electron in the deflecting electric field ; electron mass is independent of velocity. The kinetic energy acquired by an electron accelerated in an electric field is very much less than the product of charge and potential, owing to the reduction in pull on an accelerated electron with increase in velocity. No practical advantage is to be gained by using potentials over 2 MV for electron acceleration.

537.133 : [546.217+546.621 **1736**

Theoretical Range-Energy Values for Protons in Air and Aluminum.—J. H. Smith. (*Phys. Rev.*, 1st Jan. 1947, Vol. 71, No. 1, pp. 32–33.) Range-energy values are tabulated for energies up to 10^4 MeV. For the theoretical derivation of the formula used see for example 1303 of 1943 (Rossi & Greisen) : restrictions on its validity and computational procedure are discussed.

537·311.31 + 537·311.33 **1737**

Thermomagnetic Nernst Effect in Semiconductors and Metals.—Pisarenko. (*See* 1815.)

537·311.33 **1738**

Mechanism of Luminescence of Alkali-Halogen Phosphors.—Antonov-Romanovskij. (*See* 1817.)

537·311.37 : 537·525 **1739**

Conductivity of Gases excited by H.F. Discharges.—P. Mesnage. (*C. R. Acad. Sci., Paris*, 8th April 1942, Vol. 214, No. 14, pp. 702–704.) Measurements at about 25 Mc/s give conductivities of 0.03–0.3 mho/cm for hydrogen, 0.2–0.4 mho/cm or more for neon and still higher values for a mixture of argon and cadmium vapour, the pressures being a few tenths of 1 mm Hg. The volt-ampere characteristics are of the descending type, as with arcs.

537·523·4 **1740**

Influence of Electric Fields on the Extinction of Electric Sparks.—O. Yadoff. (*C. R. Acad. Sci., Paris*, 8th July 1946, Vol. 223, No. 2, pp. 74–75.) Application of a radial electric field between a ylinder and the axial path of sparks maintained etween two Pt-Ir points has the effect of rogressively reducing the brightness of the parks as the field intensity is increased, until nally the sparks are completely extinguished. or sparks 10 cm long maintained by 51 kV the ritical extinction voltage for the cylinder was 54 kV. Reduction of the air pressure increased he extinction voltage. Application is envisaged high-voltage circuit breakers.

37·533 : 537·311.31 **1741**

Cold Emission from Plane Metallic Surfaces.— Bertein. (*C. R. Acad. Sci., Paris*, 23rd Sept. 946, Vol. 223, No. 13, pp. 475–478.) Comparison the cold emission current I from a nearly plane rface S with I_0, that from the geometrically ane surface obtained by projecting S on to its ean plane, shows that there may be a considerable screpancy, so that determination of the characteric constant A, in Fowler and Nordheim's formula, om the curve relating I and the applied voltage, ves results which are too high. This does not pply to the constant B, which is found from the ope of the curve, which is practically the same for and I_0. The case of microscopically rough rfaces is also discussed.

7·535·75 **1742**

Absorption of Secondary Electrons by Thin reens.—A. Saulnier. (*C. R. Acad. Sci., Paris*, h April 1946, Vol. 222, No. 15, pp. 876–878.) condary X rays from a sheet of lead 0.2 mm ck are used with a graded series of foils of minium or cellophane to obtain photometrically set of curves connecting foil thickness, photophic density and primary X-ray voltage.

3.12 **1743**

Note on Magnetic Energy.—G. H. Livens. (*Phys. v.*, 1st Jan. 1947, Vol. 71, No. 1, pp. 58–63.) ply to a recent article by E. A. Guggenheim 27 of 1946) referring to the author's earlier work 25 of 1945) and continuing the discussion of the rect formulation of the Lagrangian and Hamilian functions for a system of linear currents and manent magnetization.

538.23 **1744**

Calculation of the Coercive Field from the Theories of Becker and of Kersten.—L. Néel. (*C. R. Acad. Sci., Paris*, 17th July 1946, Vol. 223, No. 3, pp. 141–142.)

538.323 **1745**

The Force exerted by a Rectilinear Current on a Parallel Current.—É. Brylinski. (*C. R. Acad. Sci., Paris*, 16th Sept. 1946, Vol. 223, No. 12, pp. 453–455.) A supplement to 1747 below. The force between two parallel conductors carrying current is calculated rigorously, including the effects of the finite cross-section of the conductors and their internal permeability. The formulae provide a solid foundation for an absolute definition of the ampere in the m.k.s. system.

538.23 **1746**

Mathematical Representation of Hysteresis Cycles.—P. Bricout. (*Rev. gén. Élect.*, June 1945, Vol. 54, No. 6, pp. 183–191.) A mathematical relation between the magnetizing field and the induction is derived by means of an 'indicator curve' whose abscissae are inductions and whose ordinates are the logarithms of the slopes of the tangents to the cyclic curve. From the shape of the 'indicator curve' a representative function can be derived easily, which contains in general a hyperbolic term, an exponential term, and an added constant. The errors due to the use of this function are within those of experimental measurement.

538.31 **1747**

The Force exerted by a Magnetic Field on an Element of Current.—É. Brylinski. (*C. R. Acad. Sci., Paris*, 26th Aug. 1946, Vol. 223, No. 9, pp. 378–380.) A demonstration, by a new argument, that the permeability which figures in the classical formula for the force exerted by a magnetic induction field on a current element is that of the medium and not that of the conductor. See also 1745 above.

548.0 : 547.476.3-162 **1748**

Structure and Thermal Properties associated with Some Hydrogen Bonds in Crystals: Part 7—Behaviour of KH_2PO_4 and KH_2AsO_4 on Cooling.—A. R. Ubbelohde & I. Woodward. (*Proc. roy. Soc. A*, 11th Feb. 1947, Vol. 188, No. 1014, pp. 358–371.) Part 6 was noted in 3256 of 1946.

537·53 : 535·42 **1749**

The Diffraction of X Rays and Electrons by Free Molecules. [Book Review]—M. H. Pirenne. Cambridge University Press, 1946, 160 pp., 12s. 6d. (*Nature, Lond.*, 11th Jan. 1947, Vol. 159, No. 4028, pp. 45–46.)

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.16 **1750**

Induction of Fast Charged Particles Currents by Rotating Magnetized Cosmic Bodies.—J. Terletzky. (*J. Phys., U.S.S.R.*, 1946, Vol. 10, No. 4, pp. 377–382.) Computation of the motion of charged particles in the electromagnetic field of a rotating magnetized cosmic body whose magnetic and geographic poles do not coincide. The energy up to which the particles can be accelerated in such a field is calculated.

523.165 + 550.37 + 550.38 + 551.510.535] : 061.6 **1751**

Summary of the Year's Work, to June 30, 1946, Department of Terrestrial Magnetism, Carnegie Institution of Washington.—J. A. Fleming. (*Terr. Magn. atmos. Elect.*, Dec. 1946, Vol. 51, No. 4, pp. 517–529.) An account of investigations completed or in progress, with Prof. S. Chapman's comments on the programme. Items are described under the following headings: geomagnetic investigations, cosmic relations, terrestrial·electricity, ionosphere, nuclear physics, and observatory and field work. A review of war applications is given and the more important achievements of the various service departments are listed.

523.5 : 621.396.82 **1752**

Radar Detection of Meteor Trails.—E. V. Appleton & R. Naismith. (*Nature, Lond.*, 28th Dec. 1946, Vol. 158, No. 4026, pp. 936–938.) The diurnal and seasonal variation of transient radio echoes from the upper atmosphere have been observed since 1932, and have been ascribed to the incidence of sporadic meteors. Normally, a few tens of these echoes are observed in the evening, rising to a hundred or more in the early morning. On the night of 9th–10th Oct. 1946, their number rose sharply to about 1 500 for a short period at the time of the Giacobinid meteor shower. Though visual observation was impossible on this occasion, the shape of the curve giving rate of echo occurrence agrees remarkably well with that obtained visually by F. G. Watson during the last Giacobinid shower in 1933. It is concluded that the meteoric origin of these transient radio echoes is established.

523.5 : 621.396.82 **1753**

Meeting of the Royal Astronomical Society: Observations of the Giacobinids, 1946.—(*Observatory*, Feb. 1947, Vol. 67, No. 836, pp. 1–8.) The minutes of a meeting of the Society held on 13th Dec. 1946, which included the description and discussion of visual observations of the 1946 Giacobinid shower by J. P. M. Prentice and radio observations by A. C. B. Lovell and J. S. Hey. The ionization density and velocity of the meteor trails were deduced by radar methods. The values obtained—particle radii 0.01–0.03 cm and average velocity 22.9 km/s—are in good agreement with other evidence.

523.7 + 550.385] " 1946 " **1754**

Magnetic Storms and Solar Activity 1946.—H. W. Newton. (*Observatory*, Feb. 1947, Vol. 67, No. 836, pp. 37–39.) Outstanding solar phenomena in 1946 were the largest and second largest sunspots ever recorded at Greenwich (in February and July) with associated brilliant flares and the highest prominence ever observed. The correlation of this activity with radio fade-out and magnetic storms is demonstrated.

523.7 + 550.385] " 1946.07/.09 " **1755**

Solar and Magnetic Data, July to September, 1946, Mount Wilson Observatory.—S. B. Nicholson & E. S. Mulders. (*Terr. Magn. atmos. Elect.*, Dec. 1946, Vol. 51, No. 4, pp. 561–562.)

523.7 **1756**

On the Distribution of the Solar General Magnetic Field and Remarks concerning the Geomagnetism and the Solar Rotation.—C. Walén. (*Ark. Mat. Astr. Fys.*, 6th Feb. 1947, Vol. 33, Part 3, Section A, No. 18, 63 pp. In English.) A general mathematical paper. The convective motions in the core are studied in section 2 without any regard to magnetism. Although certain ' convective convulsions ' are shown to exist, only a qualitative demonstration of convection within the core is given. The magnetic field in the core is studied in section 3. Calculations are difficult but the conclusion that the general field in the core is reproduced by the convection is, at any rate, certain. The rotation of the sun is discussed in section 4 together with the elastic torsional vibrations resulting from the recurrent ' convulsions '. The conditions at the solar surface (non-uniform rotation, sunspots and the corona) are tentatively discussed. Concluding sections of the paper deal with the general magnetic field in the radiative envelope and with the electric drift-current in the granulated layer.

523.72 + 537.591] : 621.396.822.029.5 **1757**

Radio-Frequency Investigations of Astronomical Interest.—G. Reber & J. L. Greenstein. (*Observatory*, Feb. 1947, Vol. 67, No. 836, pp. 15–26.) Critical survey of the experimental and theoretical work on cosmic and solar radiations at r.f. The dependence of cosmic noise intensity on frequency and on galactic coordinates and the mechanism of its production are discussed, but on present evidence it cannot be decided whether interstellar particles or the stars themselves are the radiant sources. The abnormally intense solar radiation at $\lambda > 1$ m originates in the corona and the frequency dependence is determined by the variation of absorption and temperature with height. The marked increase (1 000 times) of intensity during periods of sunspot activity may be related to gyromagnetic effects caused by the field of the sunspot.

523.72 : 621.396.822 **1758**

On Radio-Frequency Emission from the Sun.—J. V. Garwick. (*C. R. Acad. Sci., Paris*, 24th Feb. 1947, Vol. 224, No. 8, pp. 551–553.) Discussion based on the gyromagnetic theory of solar radio-frequency emission shows that for the observed radiation the Döppler effect is of little importance. The radiation intensity was previously found to be proportional to $H^{\frac{8}{3}}$. A revised value of the index is 0.9.

523.72 : 621.396.822.029.62 **1759**

Study of the Conditions of Emission of Metre Radio Waves by the Solar Atmosphere.—J. Denisse. (*Rev. sci., Paris*, 15th Sept. 1946, Vol. 84, No. 3257, pp. 259–262.) The possible emission of short waves by electrons moving in the magnetic field of a sunspot is considered. As the waves are absorbed by the very medium which produces them, it is suggested that a layer of thickness z_0 is solely responsible for the emission. In the lower corona z_0 is of the order of several kilometres for a magnetic field of 1 000 gauss, corresponding to a wavelength of 10 cm. Such waves could only emerge if there is a considerable gradient of the magnetic field in the layer of thickness z_0: during the growth of a sunspot, high field gradients appear to be possible. The waves could be transmitted through the corona if generated in its lowest layers, where there is strong emission and feeble absorption. At the surface of the earth it is calculated that a radiation of about 10^{-16} W/cm² per megacycle of bandwidth should be obtained for wavelengths from several centimetres to several metres.

523·72 : 621.396.822.029.63 **1760**

Solar Radiation at 480 Mc/s.—G. Reber. (*Nature, Lond.*, 28th Dec. 1946, Vol. 158, No. 4026, p. 945.) Discussion of results of daily automatic recording at Wheaton, Illinois. During a partial eclipse the intensity decreased and variable activity was observed during a radio storm.

523·746 : 538.12 **1761**

Magnetic Field of Sunspots : Part 1.—L. Gurevich & A. Lebedinsky. (*J. Phys., U.S.S.R.*, 1946, Vol. 10, No. 4, pp. 327–332.) The magnetic field of sunspots is explained in terms of a self-excitation process related to the hydrodynamic circulation inside a sunspot. Calculations on this hypothesis give fields of several thousand gauss in the outer layers of sunspots and also show that the magnetic fields in the components of bipolar groups of sunspots are oppositely directed.

523·746 " 1946.07/.09 " **1762**

Provisional Sunspot-Numbers for July to September, 1946.—M. Waldmeier. (*Terr. Magn. atmos. Elect.*, Dec. 1946, Vol. 51, No. 4, p. 500.)

537·591 + 523.165 **1763**

Measurements of the Absorption of Cosmic Rays at an Altitude of 3 050 m above Sea Level.—A. Alichanian & A. Weissenberg. (*J. Phys., U.S.S.R.*, 1946, Vol. 10, No. 3, pp. 293–294.) Continuation of work described in 73 of 1946 (Alichanow & Alichanian). Of two drops observed in the absorption curve one is attributed to the probable presence of protons in the soft component ; the other is as yet unexplained. See also 1423 of May.

537·591 + 523.165 **1764**

On Narrow Showers.—A. Alichanian & A. Alexandrian. (*J. Phys., U.S.S.R.*, 1946, Vol. 10, No. 3, pp. 296–297.) From an investigation of cosmic ray showers at heights of 960 m and 3 250 m it is concluded that there exist : (*a*) Auger showers of radius about 100 m, (*b*) narrow showers of radius about 50 cm, the radius decreasing with increase of altitude, and (*c*) dense penetrating showers of undetermined radius.

537·591 **1765**

Highly Ionizing Particles in the Cosmic Radiation.—W. Wechsler, N. Dobrotin & V. Khvoles. (*Zh. eksp. teor. Fiz.*, 1946, Vol. 16, No. 7, pp. 553–556. In Russian, with English summary.) At an altitude of 3 860 m the number of particles with ionizing power 3–4 times that of fast mesotrons was found to be less than 0.5% of the total of penetrating particles. For English version see *J. Phys., U.S.S.R.*, 1945, Vol. 9, p. 277.

537·591 **1766**

On the Space Correlation of Particles in Cosmic Rays : Part 2 — Correlation between Electrons and Photons.—V. Berestetzky. (*Zh. eksp. teor. Fiz.*, 946, Vol. 16, No. 8, pp. 665–671.) English version oted in 1435 of May.

37·591.15 **1767**

Ionization Bursts created by Mesotrons.—S. Belenky. (*Zh. eksp. teor. Fiz.*, 1946, Vol. 16, No. 6, pp. 465–473. In Russian, with English summary.) Bursts created by δ-electrons due to mesotrons re considered. The spectrum of mesotrons is aken into account. The expression obtained in his way for the number of bursts differs considerably from that given by other authors." English version in *J. Phys., U.S.S.R.*, 1946, Vol. 10, No. 2, pp. 144–150.

537·591.5 **1768**

Variation with Altitude of the Cosmic Ray Showers. Photons in the Showers.—S. Gorodetzky, P. Chanson & H. Denamur. (*C. R. Acad. Sci., Paris*, 16th Feb. 1942, Vol. 214, No. 7, pp. 310–312.)

550·38 " 1946.01/.03 " **1769**

Five International Quiet and Disturbed Days for January to March, 1946.—W. E. Scott. (*Terr. Magn. atmos. Elect.*, Dec. 1946, Vol. 51, No. 4, p. 560.)

550·38 " 1946.07/.09 " **1770**

American Magnetic Character-Figure, C_A, Three-Hour-Range Indices, K, and Mean K-Indices, K_A, for July to September, 1946.—W. E. Scott. (*Terr. Magn. atmos. Elect.*, Dec. 1946, Vol. 51, No. 4, pp. 505–508.)

550·384 **1771**

Periodicity of Geomagnetic Activity.—A. Ogg. (*Terr. Magn. atmos. Elect.*, Dec. 1946, Vol. 51, No. 4, pp. 543–546.) The results are given of a harmonic analysis of three-hour-range magnetic indices at Hermanus from 1940 to 1946. Periods of magnetic activity which were half and one-third of the solar rotation period were found, in addition to the full period of 26.875 days.

550·385 " 1946.09 " **1772**

Geomagnetic Disturbance of September 16–23 [1946].—(*Nature, Lond.*, 5th Oct. 1946, Vol. 158, No. 4014, pp. 477–478.) A description of the geomagnetic and associated phenomena of the disturbance, which coincided with marked sunspot activity. At Abinger the extreme ranges were $2^\circ\ 16'$ in D, 925γ in H and 450γ in V. A short radio fade-out was reported during the storm and during this fade-out a small solar flare was observed at Greenwich. Reprinted in *Terr. Magn. atmos. Elect.*, Dec. 1946, Vol. 51, No. 4, pp. 577–578.

551·508.1 : 621.396.9 **1773**

The British Radio-Sonde.—P. A. Sheppard. (*Quart. J. R. met. Soc.*, April/July 1946, Vol. 72, Nos. 312/313, pp. 169–173.) A description is given of the apparatus and procedure used, and the magnitude of errors in measurement is discussed. Report of a discussion at a Royal Meteorological Society meeting. For the original account of the apparatus see 4263 of 1938 (Thomas).

551·510.535+621.396.812 **1774**

Echoes on Short Waves at Low Altitude.—R. Jouaust. (*C. R. Acad. Sci., Paris*, 16th Feb. 1942, Vol. 214, No. 7, pp. 309–310.) During the daily ionosphere measurements carried out in Paris about midday in 1938 and 1939, echoes were observed at altitudes lower than that of the E layer. Comparison with records of abnormal long-wave (11 000 m) propagation between Paris and Casablanca and Tunis, shows that on three occasions when the short-wave echoes were sharpest, almost simultaneous strengthening of the 11 000 m signals occurred. The observed heights of the reflecting layer were of the order of 70 to 100 km.

551·510.535 **1775**

Early History of the Ionosphere.—A. L. Green.

(*A.W.A. tech. Rev.*, Dec. 1946, Vol. 7, No. 2, pp. 177–228.) A detailed survey, with extensive bibliography, of early work on the ionosphere.

551.510.535 **1776**

The Origins of the E Layer of the Ionosphere.—R. Jouaust. (*C. R. Acad. Sci., Paris*, 2nd March 1942, Vol. 214, No. 9, pp. 441–442.) Ionization in the E layer is shown to be due not to normal atoms, but to atoms excited to a metastable state. Study of the light from the night sky leads to the conclusion that certain metastable atoms should exist at a height of about 100 km and these can be excited by radiation of wavelength 1 323 Å. The mechanism proposed by Cabannes and Aynard to explain the production of these atoms also explains the low altitude of the E layer.

551.510.535 : 523.78 (485) " 1945.07 " **1777**

Chalmers Solar Eclipse Ionospheric Expedition 1945.—O. E. H. Rydbeck. (*Chalmers tekn. Högsk. Handl.*, 1946, No. 53, 44 pp. In English.) Observations were made for a month, centred on 9th July 1945 the day of the eclipse, near Umeå in northern Sweden. Effects in the E and F_1 layers were regular and symmetrical: apparent recombination coefficients were $\alpha_E = 1.1 \times 10^{-8}$ for the E layer and $\alpha_{F_1} = 0.6 \times 10^{-8}$ for the F_1 layer. The electron-density/time curves for the E and F_1 layers during the eclipse period had practically the same shape and gave a ratio of recombination coefficients equal to 1.8. Calculations on the basis of constant α-values indicate that the ionizing radiation from the corona cannot be neglected. The results of the observations agree with those of previous workers and add further proof to the theory that the primary ionizing force of the E and F_1 layers is ultra-violet light from the sun.

F_2-layer results were more complex. The recombination coefficient α_{F_2} was of the order of 3×10^{-9}; there was a definite ultra-violet light eclipse effect, and there may be an eclipse temperature effect resulting in a contraction of the layer. No definite indications of a corpuscular eclipse were noticed. High latitude F_2 eclipse effects seem to show a definite character which differs from that observed in equatorial regions; this may be due to geomagnetic control.

The paper includes a general theoretical discussion of eclipse effects upon the time-variation of ionospheric electron densities. Simple formulae have been obtained for the calculation of E and F_1 minimum solar eclipse electron densities; these formulae can be used for rapid determination of the recombination coefficient.

The influence of horizontal diffusion upon the F_2 eclipse electron density variation is also discussed theoretically and is shown to be unimportant even in regions of high electron temperature.

551.510.535 : 523.78(485) " 1945.07.09 " **1778**

Records of the Ionosphere during the Total Eclipse in the North of Sweden on July 9, 1945.—W. Stoffregen. (*Terr. Magn. atmos. Elect.*, Dec. 1946, Vol. 51, No. 4, pp. 495–499.) Measurements made with a portable automatic recorder showed variations of ionization nearly symmetrical with respect to the eclipse, and agreeing with earlier observations. The F_2 region showed a marked increase in height, while towards the end of the eclipse the abnormal-E ionization disappeared.

551.510.535 : 621.396.812 **1779**

Radio Waves and the Ionosphere.—G. W. O. Howe. (*J. Brit. Instn Radio Engrs*, Jan./Feb. 1947, Vol. 7, No. 1, pp. 36–42.) A short historical survey is followed by a discussion of the mechanism which causes radio waves entering the ionosphere to return to the earth. Reference is made to experimental results by Appleton and others.

551.594 : 546.212-14/-16 **1780**

Electrical Effects associated with a Change of State of Water.—J. E. Dinger & R. Gunn. (*Terr. Magn. atmos. Elect.*, Dec. 1946, Vol. 51, No. 4, pp. 477–494.) A laboratory investigation shows that freezing and melting give rise to changes in contact potential at the air-water interface. The results are of interest in explaining the charges on atmospheric precipitation and the separation of electricity in clouds.

551.594.11(65) **1781**

A Year's Records of the Atmospheric Electric Potential Gradient at Tamanrasset (Hoggar).—J. Lacaze. (*Rev. sci., Paris*, 15th Sept. 1946, Vol. 84, No. 3257, pp. 286–289.)

551.594.12 **1782**

The Ionisation in the Lowest Regions of the Atmosphere.—J. A. Chalmers. (*Quart. J. R. met. Soc.*, April/July 1946, Vol. 72, Nos. 312/313, pp. 199–205.) Assuming the rate of ionization at the earth's surface to be 5 times that at a height of 1 m, it is possible to obtain a variation of conductivity in general agreement with Hogg's results without requiring a space charge large enough to alter the field appreciably.

551.594.21 **1783**

A Theory of the Fundamental Phenomena of Atmospheric Electricity.—J. Frenkel. (*Bull. Acad. Sci. U.R.S.S., sér. géogr. géophys.*, 1944, Vol. 8, No. 5, pp. 244–272. In Russian, with English summary.) The principles of colloidal electrochemistry are applied to the thunder-cloud. In the dispersive medium formed by the weakly ionized air, negative electrification of the water drops or ice crystals is deduced.

551.594.221 : 621.396.821 **1784**

The Nature and Variation of Atmospherics caused by Lightning Discharges.—H. Norinder & W. Stoffregen. (*Ark. Mat. Astr. Fys.*, 6th Feb. 1947, Vol. 33, Part 3, Section A, No. 16, 44 pp. In English.) Preliminary report of research in progress. The electric field changes associated with atmospherics and the variation of waveform with distance from the source are examined. Waveforms are photographically recorded by various oscillographic techniques and preliminary discharges have been observed up to 0.5 sec before the main discharge in addition to the usual 'leader' strokes. Further measurements will be made by taking simultaneous recordings at various distances from the source.

551.594.52(481) " 1911/44 " **1785**

Frequency of 12,330 Measured Heights of Aurora from Southern Norway in the Years 1911-1944.—C. Störmer. (*Terr. Magn. atmos. Elect.*, Dec. 1946, Vol. 51, No. 4, pp. 501–504.)

LOCATION AND AIDS TO NAVIGATION

621.396.9 : 551.594.6 **1786**

Atmospherics and Their Location.—C. Clarke. (*J. Instn elect. Engrs*, Part I, Jan. 1947, Vol. 94, No. 73, pp. 54–55.) The nature of the discharge process associated with a lightning flash is briefly considered. A direction-finder used for locating lightning flashes must have a selective tuned receiver for an operating band of about 10–30 kc/s, a linear output for a wide range of input voltages, and a higher overall gain than would be necessary for c.w. signals of similar field strength. A c.r.t. with a persistent screen must be used. Brilliance modulation has marked advantages. Atmospherics are located by simultaneous observation of their direction of arrival at two well-separated stations. The main source of inaccuracy is polarization error. Summary of an I.E.E. Students' section paper.

621.396.932 **1787**

Decca Navigator.—(*Wireless World*, Feb. 1947, Vol. 53, No. 2, p. 67.) Now approved by the Ministry of Transport for general marine navigation, after extensive trials.

21.396.932/.933].24 **1788**

Consol.—(*Wireless World*, Feb. 1947, Vol. 53, o. 2, p. 67.) The first Consol d.f. station, situated Bush Mills, Co. Antrim, Northern Ireland, now rovides a 24-hour service for civil aircraft. For a escription of the system, see 2912 of 1946 (Clegg).

21.396.933 **1789**

Gee : A Radio Navigational Aid.—R. J. Dippy. . *Instn elect. Engrs*, Part IIIA, 1946, Vol. 93, o. 2, pp. 468–480.) A 'master' station transmits μs pulses at a repetition frequency of 500 per sec ; vo 'slave' stations also transmit 6-μs pulses at a petition frequency of 250 per sec, and are locked the 'master' so that the interval between naster' and 'slave' pulses remains constant. ne signals are displayed on the airborne receiver so at the path difference from the aircraft to the naster' and each 'slave' station can be read rectly. Charts are provided with a system of perbolae of constant path difference plotted for e 'master' and each 'slave' station : the craft's position is then determined as an interction of a hyperbola of each system. The airborne and ground-station equipments are scribed in detail. A later development was the roduction of a third slave station to give allund coverage, and the design of light transportable ound stations. Future possibilities of the system briefly discussed. A summary of this paper is en in Part IIIA, 1946, Vol. 93, No. 1, pp. –345.

.396.933 **1790**

boe : A Precision Ground-Controlled Blindmbing System.—F. E. Jones. (*J. Instn elect. grs*, Part IIIA, 1946, Vol. 93, No. 2, pp. 496–511.) craft flying at 30 000 ft could be controlled from ground stations up to a range of nearly 300 es. Bombs were released automatically by a nal sent from one of the ground stations ; the ial bombing error was of the order of 100 yards, ying with aircraft height, range and speed. two ground stations transmit pulses on the e carrier frequency but different pulse recurrence frequencies. The tracking station guides the pilot on to a circular track passing over the target while the releasing station determines the ground speed of the aircraft along this track and hence the bomb release point. The airborne and ground station equipments are described in detail. Possible future applications are mentioned. A summary of this paper is given in Part IIIA, 1946, Vol. 93, No. 1, pp. 345–346.

621.396.933 **1791**

200-Mc/s Radar Interrogator-Beacon Systems.—K. A. Wood. (*J. Instn elect. Engrs*, Part IIIA, 1946, Vol. 93, No. 2, pp. 481–495.) Extensive wartime use of radar beacon systems is described. Particular problems encountered in developing both ground and airborne equipment are discussed. The development of the i.f.f. identification system is outlined, with the modifications which later made possible its use as a simple navigational aid. The series of equipments known as 'Rebecca-Eureka' enabled aircraft not fitted with radar search equipment to use the beacon system. Aircraft approach and landing by the Beam Approach system are considered. Special developments of the beacon system to meet operational requirements of all three Services are outlined and trends in development for future use are reviewed. A summary of this paper is given in Part IIIA, 1946, Vol. 93, No. 1, pp. 347–348.

621.396.933.23 **1792**

The C.A.A. Instrument Landing System.—C. E. Planck. (*Inter-Avia*, Nov./Dec. 1946, Vol. 1, Nos. 8/9, pp. 63–66.) The system introduced by the U.S. Civil Aeronautics Administration consists of a runway localizer (110 Mc/s), a straight-line glide path (330 Mc/s) and marker beacons (75 Mc/s). These elements are briefly described, and details and photographs of the ground and airborne equipments are given. Advantages over the G.C.A. radar system are claimed in cheapness and in automatic devices for instrument landing and control tower monitoring.

621.396.96 **1793**

H_2S : An Airborne Radar Navigation and Bombing Aid.—C. J. Carter. (*J. Instn elect. Engrs*, Part IIIA, 1946, Vol. 93, No. 2, pp. 449–467.) A radar device giving a visual plan of the terrain beneath so that accurate navigation and bombing become possible from above clouds. Development of the system since 1939 is traced, and its operational use by the R.A.F. is briefly considered. The functions of the various units which make up the complete equipment are outlined ; particular attention is paid to the correction of slant-range distortion to give a true plan position indication, and to problems of display and aerial design peculiar to this equipment. A summary of this paper is given in Part IIIA, 1946, Vol. 93, No. 1, pp. 343–344.

621.396.96 **1794**

Principles and Applications of Radar.—L. Bouthillon. (*Bull. Soc. franç. Élect.*, Nov. 1946, Vol. 6, No. 63, pp. 563–578.) An outline of the basic principles of radar and of the technique of determining direction and range, together with some military applications.

621.396.96 **1795**

Questions concerning Different Radar Systems.—G. Guanella. (*Inter-Avia*, Nov./Dec. 1946, Vol. 1, Nos. 8/9, pp. 48–54.) Comparison of the short-pulse method and the beat method, in which the transmitter frequency is changed uniformly during a pulse of relatively long duration, shows that theoretically the two methods are comparable in accuracy. The detection of moving objects by Döppler techniques is discussed and the practical embodiment of the various systems is described.

621.396.96 (44) **1796**

French Contributions to Radar.—M. Ponte. (*Bull. Soc. franç Élect.*, Nov. 1946, Vol. 6, No. 63, pp. 579–588.) A general description of experimental work on the detection of obstacles by ultra short waves, carried out in France between 1934 and 1942. See also 1099 of April and back reference.

621.396.96 : 371.3 : 534.321.9 **1797**

H_2S Trainer.—G. W. A. Dummer. (*Wireless World*, Feb. 1947, Vol. 53, No. 2, pp. 65–66.) Ultrasonic waves propagated through water on to submerged relief maps enabled H_2S operational conditions to be reproduced on a scale of 1 : 200 000.

621.396.96 : 531.55 **1798**

Naval Fire-Control Radar.—J. F. Coales, H. C. Calpine & D. S. Watson. (*J. Instn elect. Engrs*, Part IIIA, 1946, Vol. 93, No. 2, pp. 349–379.) Development is reviewed from the beginning as a simple range finder to the present accurate and complex form. Aerial arrays are described, with polar diagrams, and the installation of the equipment aboard ships is shown diagrammatically.

Methods of display and the problem of discriminating between targets are discussed. Such narrow beam widths are required, and the actual size of the equipment is so limited by the available space, that very high frequencies are used. The development of equipment in the 600, 3 000 and 10 000 Mc/s bands is described and methods of using radar to locate shell splashes to correct gun laying are explained.

A summary of this paper is given in Part IIIA, 1946, Vol. 93, No. 1, pp. 323–325.

621.396.96 : 531.55 **1799**

Precision Ranging Systems for Close-Range Weapons.—H. W. Pout. (*J. Instn elect. Engrs*, Part IIIA, 1946, Vol. 93, No. 2, pp. 380–394.) The development of close-range sets is described and also the reasons for the successive changes. The prediction of aircraft future position is discussed, including analysis of 'rate aiding' as a method of rate determination. Two sets are described, with block diagrams. The first of these, Type 282, did not give the required accuracy. The second, Type 285 incorporating Panel L22, is described in detail and the problems associated with the use of Panel L22 in conjunction with the auto-barrage unit are discussed. An account is given of a completely self-contained automatic set including auto-ranging and auto-aiming of the guns. The limitations of such a system are outlined. A summary of this paper is given in Part IIIA, 1946, Vol. 93, No. 1, pp. 326–327.

621.396.96 : 531.55 **1800**

The Application and Design of Medium-Precision Ranging Equipment.—H. A. Prime. (*J. Instn elect. Engrs*, Part IIIA, 1946, Vol. 93, No. 2, pp. 395–410.) A description of Ranging Outfit R.T.A. for use on small vessels, with diagrams of parts of the circuit. The modified version (Outfit R.T.C.) of this set, used for control of short-range anti-submarine weapons, is also described.

The design of medium-precision ranging equipment for use with p.p.i. display is discussed, reference being made to Ranging Outfit R.T.D., Display Outfit J.P. and the applications of such equipment to torpedo control, navigation and shore bombardment.

The limitations of this type of display led to the development of equipment known as Ranging Outfit R.T.B. which combines a sector display with R.T.A.

A summary of this paper is given in Part IIIA, 1946, Vol. 93, No. 1, pp. 327–328.

621.396.96 : 531.55 **1801**

Engineering Design of Ship-Borne Gunnery Radar Panels.—T. C. Finnimore & W. D. Mallinson. (*J. Instn elect. Engrs*, Part IIIA, 1946, Vol. 93, No. 2, pp. 441–447.) The design of radar equipment to operate under the conditions experienced aboard ship is described with particular reference to two types developed during the war. A note on future trends in design is added. A summary of this paper is given in Part IIIA, 1946, Vol. 93, No. 1 pp. 329–330.

621.396.96 : 531.55 **1802**

Naval Gunnery Radar.—(*J. Instn elect. Engrs*, Part IIIA, 1946, Vol. 93, No. 2, pp. 447–448.) A discussion led by Cmdr. C. G. Mayer, U.S.N.R., on 1798 to 1801 above, and 1803 and 1804 below.

621.396.96 : 531.55 **1803**

Checking the Angular Accuracy of Precision Fire-Control Radar.—G. H. Beeching. (*J. Instn elect. Engrs*, Part IIIA, 1946, Vol. 93, No. 2, pp. 519–526.) The principal sources of error are incorrect alignment of the mechanical axis of the aerial system and of the electrical axis of the radar equipment, and incorrect installation with respect to predictor, guns, etc. Methods are described for checking and correcting these errors in typical Army equipment.

621.396.96 : 531.55 : 621.396.611.21 **1804**

A Precision-Ranging Equipment using a Crystal Oscillator as a Timing Standard.—C. A. Laws. (*J. Instn elect. Engrs*, Part IIIA, 1946, Vol. 93, No. 2, pp. 423–440.) A detailed account, with practical difficulties experienced, of the development of a system for naval armament fire-control. One period of the crystal is equal to a radar range of 1 000 yards and the intervals are continuously subdivided by a phase-shifting transformer in the oscillator output, one revolution giving a linear phase-change of 360°, so that it can be calibrated in terms of radar range. The form of display of the radar and timing signals on a cathode-ray tube gives a very high measurement accuracy.

Recent applications of the system are described, with an appendix on the factors governing the phase-shifting transformer accuracy.

A summary of this paper is given in Part IIIA, 1946, Vol. 93, No. 1, pp. 325–326.

621.396.96.004.11 **1805**

Airborne Radar Specifications.—(*Electronics*, Feb.

1947, Vol. 20, No. 2, p. 132.) Brief technical details of various U.S. Army radar equipments recently declassified. Further information can be obtained from the Superintendent of Documents, Government Printing Office, Washington, D.C.

621.396.96.088.2 **1806**

The Measurement of the Zero Error of Range in Radar Equipments.—H. I. S. Allwood, J. G. Bartlett & G. T. Davies. (*J. Instn elect. Engrs*, Part IIIA, 1946, Vol. 93, No. 2, pp. 513–518.) A brief description of the precautions necessary for eliminating errors, and details of the measurement of the zero error by means of an electronic calibrator employing an artificial target. Ranges can usually be determined to within ten yards of the true value. A summary of this paper is given in Part IIIA, 1946, Vol. 93, No. 1, p. 127.

MATERIALS AND SUBSIDIARY TECHNIQUES

35.215.6 : 546.281.26 **1807**

A New Spectral Effect and a Method for determining the Long Wave-Length Limit of the Rectifier Photoeffect in Carborundum Monocrystals.—O. V. osev. (*Bull. Acad. Sci. U.S.S.R., sér. phys.*, 941, Vol. 5, Nos. 4/5, pp. 494–498. In Russian ith English summary.) In the 'active layer' f carborundum monocrystals a special component f the rectifier photoeffect may be observed when n accelerating voltage is applied. The ratio of his component to the short-circuit photocurrent ncreases steadily if the wavelength of the monoromatic light is decreased. This fact is used for the etermination of the long wavelength limit of the ctifier photoeffect.

35.37 **1808**

Luminescence of (Zn, Be)$_2$SiO$_4$: Mn and Other anganese-Activated Phosphors.—J. H. Schulman. . *appl. Phys.*, Nov. 1946, Vol. 17, No. 11, pp. 902–98.)

5.37 **1809**

Luminescence Extinction in Complex Molecules.— A. Tumerman. (*Bull. Acad. Sci. U.R.S.S., r. phys.*, 1945, Vol. 9, Nos. 4/5, p. 328. In ssian.) Discusses the effect of temperature and the istence of a 'dark pause'.

5.37 : 537.311.33 **1810**

Contemporary Investigations of the Mechanism of minescence of Semi-Conductors.—V. L. Levšin . L. Lewschin]. (*Bull. Acad. Sci. U.R.S.S., r. phys.*, 1941, Vol. 5, Nos. 4/5, pp. 510–522. Russian with English summary.) Luminescent id phosphors can be divided into three groups. e first are activated by complex organic moleses; their absorption and emission spectra are sely connected. Emission consists of a momeny process lasting approximately 5 × 10^{-9} sec ether with an exponential process lasting about sec due to metastable levels of the activator lecules. The second group includes luminescent e inorganic substances and isomorphic crystals, which absorption and emission spectra are also sely connected. Luminescence consists estially of an exponential process lasting about ı sec. The third group includes crystalline sphors activated by impurities of heavy metals. Emission consists of a momentary process lasting less than 10^{-5} sec, a short exponential process lasting about 0.01 sec and a prolonged hyperbolic recombination process. The momentary process provides a fraction of the total radiation negligible at low intensity, and increasing with intensity. Curves obtained by using a phosphoroscope with a rotating mirror are given, illustrating these processes for ZnS.Mn- and ZnS.Cu-phosphors. Investigation of isotherms of decay shows that the character of the process changes little with temperature in the range 90°–670°K; increase in recombination probability with temperature is compensated by a decrease in the number of excited electrons present when excitation ceases. The character of isotherms taken when the number of electrons remains approximately constant is very sensitive to temperature changes, as are also the light sums accumulated during excitation. From the relationship between light sum and temperature, the energy required to liberate a localized electron can be deduced.

535.37 : 546.65 **1811**

Luminescence of Solutions of Salts of the Rare Earths.—A. N. Zaidel. (*Bull. Acad. Sci. U.R.S.S., sér. phys.*, 1945, Vol. 9, Nos. 4/5, pp. 329–334. In Russian.)

535.37 : 548.0 **1812**

The Luminescence of Crystalline Substances.—V. L. Levshin. (*Bull. Acad. Sci. U.R.S.S., sér. phys.*, 1945, Vol. 9, Nos. 4/5, pp. 355–368. In Russian.) A new scheme to explain the mechanism. Main features are the assumption of three types of discontinuous radiation, and an explanation of the nature of adhesion levels.

535.37 : 666.1 **1813**

Luminescence of Glasses.—V. V. Vargin & T. I. Weinberg. (*Bull. Acad. Sci. U.R.S.S., sér. phys.*, 1945, Vol. 9, Nos. 4/5, pp. 563–574. In Russian.) Critical survey of existing knowledge.

535.371 **1814**

Some Problems of the Synthesis of Zinc-Sulphide Phosphors with Long Afterglow.—V. M. Gougel. (*Bull. Acad. Sci. U.R.S.S., sér. phys.*, 1945, Vol. 9, Nos. 4/5, pp. 539–542. In Russian.) Discusses the effect of impurities on the intensity and colour of the luminescence.

537.311.31 + 537.311.33 **1815**

Thermomagnetic Nernst Effect in Semi-Conductors and Metals.—N. L. Pisarenko. (*Bull. Acad. Sci. U.R.S.S., sér. phys.*, 1941, Vol. 5, Nos. 4/5, pp. 417–421. In Russian with English summary.) The magnitude as well as the sign of this effect may be explained by the mode of dependence of the time of the electron free path on the velocity, and by the fact that in semiconductors in which the conductivity is of a complex nature both types of charge carriers are present.

537.311.33 + 621.315.59 **1816**

Experimental Investigation of the Metal-Semiconductor Contact.—V. I. Ljašenko, G. A. Fedorus & S. P. Felvašnikova. (*Bull. Acad. Sci. U.R.S.S., sér. phys.*, 1941, Vol. 5, Nos. 4/5, pp. 434–441. In Russian with English summary.) An investigation for cuprous oxide and selenium of the relation between metal-semiconductor contact potential difference and field strength, current density, and conductivity. The potential difference at first increases almost linearly with current density, then attains a maximum, and finally falls to zero. The smaller the conductivity of the specimen, the larger the potential difference.

The explanation suggested is that the flow of current reduces the 'holes' concentration in the layer of chemically uniform semiconductor near a contact. The experimental data agree fairly well with S. I. Pekar's theoretical results (1819 below).

537.311.33 **1817**

Mechanism of Luminescence of Alkali-Halogen Phosphors.—V. V. Antonov-Romanovskij. (*Bull. Acad. Sci. U.R.S.S., sér. phys.*, 1941, Vol. 5, Nos. 4/5, pp. 523–531. In Russian with English summary.) Discussion of decay phenomena leads to the conclusion that the Bloch-Wilson representation of semiconductors, if applied to phosphorescence, must take account of the interaction between the electron and the ionized centre, long before their recombination, and also of the fact that the average displacement of the thermal electron, in the time interval between captures, is relatively small.

537.311.33 **1818**

Electrical Conductivity of Semiconductors in Strong Electric Fields.—B. I. Davydov & I. M. Šmuškevič. (*Bull. Acad. Sci. U.R.S.S., sér. phys.*, 1941, Vol. 5, Nos. 4/5, pp. 399–408. In Russian with English summary.) Different mechanisms which may increase the conductivity of semiconductors in strong fields are considered. The magnitude of the field at which these mechanisms become significant is evaluated. The deviations from Ohm's law in the case of a semiconductor with an ionic lattice in a strong field are examined in detail. When solving the kinetic equation, not only the interaction of electrons with the optical vibrations of the lattice, but also inelastic collisions, *i.e.* the ionization, must necessarily be taken into account. In contrast to the semiconductor with an atomic lattice, the mobility of electrons in that with an ionic lattice increases in strong fields. The dependence of the mobility on the field and the temperature has different forms according to whether kT is greater or less than $h\omega_0$, ω_0 being the limiting frequency of optical vibrations of the crystal. The mobility of electrons, the number of ionizing collisions and the resulting concentration of free electrons are found for both limiting cases; the intermediate case may be interpolated.

537.311.33 **1819**

The Metal-Semiconductor Contact and the Contact Potential Drop.—S. I. Pekar. (*Bull. Acad. Sci. U.R.S.S., sér. phys.*, 1941, Vol. 5, Nos. 4/5, pp. 422–433. In Russian with English summary.) A theory of the metal-semiconductor contact is considered which differs from those previously given by Mott and Johnson by taking into account the redistribution of the concentration of conducting electrons in a semiconductor caused by the passage of the current. Contact potential differences and their dependence on current density could thus be calculated. The experimental data of Ljašenko, Fedorus & Felvašnikova (1816 above) confirm the theory.

537.311.33 : 546.281.26 **1820**

On the Mechanism of the Electric Conductivity of Silicon Carbide.—G. Busch & H. Labhart. (*Helv. phys. Acta*, 18th Dec. 1946, Vol. 19, Nos. 6/7, pp. 463–492. In German.)

537.323 : 546.817.221 **1821**

The Thermoelectric Effect in Lead Sulphide.—E. D. Devjatkova, J. P. Maslakovec & M. S. Sominskij. (*Bull. Acad. Sci. U.R.S.S., sér. phys.*, 1941, Vol. 5, Nos. 4/5, pp. 409–416. In Russian with English summary.) Theoretical expressions are given for the dependence of thermoelectric force on the concentration of carriers of electricity. The temperature variations of electrical and thermal conductivities and of thermoelectric force are investigated in lead sulphide having electronic as well as 'hole' conductivity. The results obtained show that in lead sulphide the concentration of carriers of electricity equals 10^{18}–10^{19}. The electrical conductivity is determined mainly by the temperature variation of the mobility of the carriers.

538.22 : [669.155 + 669.245.5 **1822**

Magnetic Properties and Chemical Nature of Solid Solutions of Weak Magnetic Elements in Nickel and Iron.—J. Dorfman. (*Zh. eksp. teor. Fiz.*, 1946, Vol. 16, No. 4, pp. 349–360. In Russian, with English summary.) The nature of the elementary nickel magnet can be understood if numerical results of gyromagnetic investigations are taken into account as well as experimental magnetic data.

The magnetic properties, above the Curie point, of solid solutions of different metals in nickel (at low concentrations of the alloying metal) are determined not only by the valency of the foreign atoms, but also by the individual chemical peculiarities of the elements. The theory of zones is not fully applicable to nickel *d*-electrons.

There is a close analogy between the magnetic properties of nickel and iron alloys, but the peculiar structure and filling of *d*-levels of metallic iron explains some magnetic properties of iron alloys.

The difference between the chemical bonds of 'included atoms' and 'replacement atoms' explains why carbon behaves differently from other elements when dissolved in iron.

538.221 **1823**

Time Effects in Materials containing Iron under the Influence of Mechanical and Magnetic Forces.—J. L. Snoek. (*Schweiz. Arch. angew. Wiss. Tech.*, Jan. 1947, Vol. 13, No. 1, pp. 9–14.) A discussion of dispersion and relaxation phenomena, both elastic and magnetic, based on Bloch's theory of magnetism. Long- and short-time after-effects are found, the first concerned with the diffusion of material particles, while the second is as yet unexplained, though possibly connected with the *s*- and *d*-electrons.

546.28 + 621.383.4 **1824**

A New Bridge Photo-Cell employing a Photo-Conductive Effect in Silicon. Some Properties of High Purity Silicon.—Teal, Fisher & Treptow. (*See* 1961.)

549.514.51 **1825**

Growing Quartz Crystals.—(*Radio, N.Y.*, Jan. 1947, Vol. 31, No. 1, p. 30.) An abstract of "Report of Investigations in European Theater; PB-28897" describing a German method in which a seed crystal of quartz is suspended in a mixture of finely ground glass and water and is heated in an autoclave to 375° C.

620.197(213) : 621.396.6 **1826**

Deterioration of Radio Equipment in Damp Tropical Climates and Some Measures of Prevention.—C. P. Healy. (*A. W. A. tech. Rev.*, Dec. 1946, Vol. 7, No. 2, pp. 103–129.) Reprint of 3502 of 1946.

621.3.011.5.029.5/.6] : 631.437 **1827**

Dielectric Properties of Indian Soils at High and Medium Radio-Frequencies.—Khastgir, Ray & Banerjee. (*See* 1900.)

621.315.59 **1828**

Papers on Semiconductors.—(*Bull. Acad. Sci. U.R.S.S., sér. phys.*, 1941, Vol. 5, Nos. 4/5, pp. 546–554.) Summaries in English and Russian of the following papers are given: A Complex Investigation of the Mechanism of Semi-Conductor's Electrical Conductivity, by V. A. Davidenko. A New Method of Investigating the Electrical Conductivity of Semi-Conductors and Results of Application of this Method to the Investigation of Conductivity of Alundum, by A. R. Šulman. Concentration Phenomena in Semi-Conductors, by B. I. Davydov. Galvanomagnetic Effects in Semi-Conductors, by I. D. Rožanskij. A Contact between a Semi-Conductor and a Metal, by A. V. Ioffe & A. F. offe. Solid Rectifiers, by P. V. Šaravskij. The nvestigation of Sulphide Rectifiers, by J. A. Dunajev & B. V. Kurčatov. Inner Photoelectric Effect in Sulphur and Electronic Energy Levels, by P. S. Tartakovsky & G. I. Rekalova. On the Photo-lectrically Inactive Absorption of Light by Some emi-Conductors, by F. F. Volkenštein. Spectral Distribution of Rectifier and Inner Photoeffect in elenium, by D. I. Arkadiev.

21.315.6 + [621.39 : 371.31 **1829**

The War-Time Education and Training of Radio ersonnel and Recent Developments in Dielectric aterials.—Jackson. (*See* 1973.)

21.315.616 : 621.317.33.011.5 **1830**

Dielectric Investigations on Polymeric Fluids.—. Goldschmidt. (*Schweiz. Arch. angew. Wiss. ech.*, Jan. 1947, Vol. 13, No. 1, pp. 21–27.) escribes the method of measurement of the rmittivity, loss angle, insulation resistance, scosity and expansion coefficient of oils and arious mixtures. The results are given graphically, hotographs show the development, during cooling, pseudo-crystalline patterns in mixtures of ocerite and resin.

1.357.8 **1831**

Electropolishing.—C. L. Faust. (*Metal Ind., Lond.*, th Dec. 1946, Vol. 69, No. 25, pp. 512–513.) scusses the status of electropolishing as a metal ishing process.

1.395.625.3 **1832**

Magnetic Tape Recorder.—(*Radio, N.Y.*, Jan. 47, Vol. 31, No. 1, p. 7.) Uses a paper tape ¼ inch de coated with a new metallic paint having gnetic properties that are claimed to approach se of Alnico III.

621.775.7 : 669.3 **1833**

Copper-Base Powder Metallurgy Parts.—H. Chase. (*Materials & Methods*, Dec. 1946, Vol. 24, No. 6, pp. 1439–1444.)

669.14 : 621.396.69 **1834**

A New Radar Transformer Steel.—G. H. Cole & R. S. Burns. (*Materials & Methods*, Dec. 1946, Vol. 24, No. 6, pp. 1457–1460.) A steel, containing 3 to 3.5% silicon, capable of being rolled to a thickness of 0.002 inch, and suitable for radar pulse-transformer laminations.

679.5 : 621.3 **1835**

Plastics in Electrical Industry.—T. J. Fielding. (*Electrician*, 1st–15th Nov. 1946, Vol. 137, Nos. 3570–3572, pp. 1196–1199, 1281–1285 & 1370–1374.) A general account of the properties of various bakelite materials, in laminated forms, as anti-tracking materials, and as resins, varnishes and cements, together with their electrical applications.

679.5 : [621.315.616 + 539.4 **1836**

Standard Tests for Thermosetting Plastics.—(*See* 1865.)

678.02 **1837**

Advances in Colloid Science, Vol. 2. [Book Review]—H. Mark & G. S. Whitby (Eds.). Interscience Publishers Inc., New York, 1946, 453 pp., $7. (*Nature, Lond.*, 28th Dec. 1946, Vol. 158, No. 4026, pp. 924–926.) A much-needed co-ordination of recent developments in rubber science.

MATHEMATICS

517.941.9 : 53 **1838**

A New Method of Solution of Certain Boundary Problems for Equations of Mathematical Physics permitting of a Separation of Variables.—G. Grünberg. (*J. Phys., U.S.S.R.*, 1946, Vol. 10, No. 4, pp. 301–320.)

518.5 **1839**

Concerning "Computer for Solving Linear Simultaneous Equations".—C. C. Eaglesfield: C. E. Berry. (*J. appl. Phys.*, Dec. 1946, Vol. 17, No. 12, p. 1125.) Suggested modifications to the design described in 1927 of 1946 to make it easier and quicker to use, and Berry's reply, stating that the same mathematical method had been suggested several months previously by W. M. Bleakney.

519.283 : 621.318.572 **1840**

On the Distribution of Counts in a Counting Apparatus.—Hole. (*See* 1877.)

530.162 : 621.394/.397].822 **1841**

Study of Fluctuations produced by the Shot Effect in Amplifiers.—A. Blanc-Lapierre. (*Rev. sci., Paris*, June/July 1946, Vol. 84, No. 3254, pp. 75–94.) A review of the analytical methods used for calculating the influence of linear amplifiers and of non-linear detectors on a random voltage. The properties of the correlation function of a random variable are related to its spectral density; for Bernamont's earlier work on this subject see 1715 of 1937 and back references. The relations between the characteristic function, the probability distribution and moments of a variable are established. In practice the Laplace-Gauss (normal) distribution is valid since the number of elementary events occurring within the time constant of the apparatus is very large, and its general formulation for K variables is derived. The results are applied to the case of detectors having the law $y = x^{2p}$ and to a linear detector (single- or full-wave). The significance of departures from the normal distribution is discussed and it is shown that if an arbitrary fluctuation voltage is applied to an amplifier of large time constant, the small fluctuations appearing at the output depend only on the amplifier and have a normal distribution. Cf. similar work by Fränz (3026 and 3027 of 1941, 2124 of 1943, and 443 of 1944). An abbreviated version of this paper appears in *Bull. Soc. franç. Élect.*, Nov. 1945, Vol. 5, No. 53, pp. 343–351.

518.3 : 621.3.012.3.081.4 **1842**

DB Gain-Loss Calculator.—L. A. Lohr. (*Electronics*, Feb. 1947, Vol. 20, No. 2, p. 134.) An abac for converting ratios to decibels.

MEASUREMENTS AND TEST GEAR

538.569.4 + 621.396.812].029.64 : 551.51 **1843**

Atmospheric Absorption Measurements with a Microwave Radiometer.—R. H. Dicke, R. Beringer, R. L. Kyhl & A. B. Vane. (*Phys. Rev.*, 1st/15th Sept. 1946, Vol. 70, Nos. 5/6, pp. 340–348.) The absorption of microwave radiation in traversing the earth's atmosphere has been measured at wavelengths of 1.0 cm, 1.25 cm, and 1.50 cm in the neighbourhood of a water vapour rotational absorption line. A sensitive radiometer was used to detect thermal radiation from the absorbing atmosphere at these wavelengths. From the measured absorption, together with data from humidity soundings of the atmosphere, absorption coefficients under standard conditions (293°K and 1 015 mb) can be calculated. These are 0.011, 0.026 and 0.014 db per km for 1 gm of water vapour per cubic metre, at wavelengths of 1.00, 1.25 and 1.50 cm respectively.

620.179.1 **1844**

Non-Destructive Testing.—(*Elect. Times*, 30th Jan. 1947, Vol. 111, No. 2884, pp. 147–148.) Summary and discussion of an I.E.E. paper by H. C. Turner and E. M. Tomlin. Special instruments described include (*a*) a magnetic sorting bridge using the change of the hysteresis loop with variations of heat treatment, hardness or depth of carburation, (*b*) a vibration analyser with crystal pickup, (*c*) a crystal resonance apparatus for thickness measurement from one side only and (*d*) a development of (*c*) for the detection of flaws in sheet material.

621.317.31.014.33 **1845**

The Measurement of Current Transients in a Low Voltage Circuit.—B. T. Barnes, E. Q. Adams & D. D. Hinman. (*Rev. sci. Instrum.*, Oct. 1946, Vol. 17, No. 10, pp. 426–427.) Measurement of current transients of the order of 1 A/μs, using an oscillograph, is discussed. The construction of a special coaxial type of oscillograph shunt, which has the very low inductance necessary, is described.

621.317.33.011.5 : 621.315.616 **1846**

Dielectric Investigations on Polymeric Fluids.—Goldschmidt. (*See* 1830.)

621.317.333.4 : 621.315.2.029.5 **1847**

New Methods for Locating Cable Faults, particularly on High-Frequency Cables.—Roberts. *See* 1663.)

621.317.336 : 621.315.212.2 **1848**

New Method of Measuring the Impedance Errors of Concentric Pairs.—G. Fuchs. (*Rev. gén. Élect.*, March 1946, Vol. 55, No. 3, pp. 109–117.) A differential method for measuring the asymmetry of quadripoles is applied to the determination of impedance errors arising from dimensional and material irregularities in the construction of concentric cables. Comparison between this differential method and that of Kaden at the resonant frequencies shows good agreement, though both methods fail when similar irregularities are symmetrically located with respect to the mid-point of the cable. Tests carried out at the transmission laboratory of the Société anonyme de Télécommunications show good agreement between experimental and theoretical values for cables with artificial impedance errors.

621.317.336 : 621.317.372 **1849**

On the Measurement of Negative Impedances with a Q-Meter.—B. N. Prakash. (*Electrotechnics*, Dec. 1946, No. 19, pp. 69–73.) Measurements on a *Q*-meter are made (*a*) with a suitable coil connected to the *Q*-meter coil terminals, (*b*) with a tuned circuit, shunted by the impedance, connected to the *Q*-meter capacitor terminals, the impedance being (i) passive, and (ii) active; i.e., with battery supplies connected. The value of the impedance can then be calculated.

621.317.372 + 621.396.615 + 621.317.75 **1850**

Service Station Measurement Apparatus : Part 1— Beat-Frequency Oscillator ; Part 2—A New Q-Meter ; Part 3—Frequency Spectrum Analyser.—R. Aschen. (*TSF pour Tous*, Sept.–Nov. 1946, Vol. 22, Nos. 215–217, pp. 171–173, 195–197 & 219–222.) In the beat-frequency oscillator a variable voltage at a fixed frequency F_3 is applied to the control grid of a frequency changer and beats with a second signal of constant voltage but variable frequency F_2 to give an output of variable frequency F_1 whose voltage can be varied from zero up to about 0.05 V. By modulating F_3 either by a.m. or f.m., F_1 is similarly modulated. For *Q*-measurements, the circuit is loosely coupled by a small capacitor to the generator, the resonance frequency F_0 determined and the voltage V measured by a valve voltmeter. F_3 is then varied by a known amount ΔF to give a 30% reduction of V, when $Q = F_0/2\Delta F$. Modification of this simple method gives a direct reading instrument. The frequency analyser, whose circuit diagram is given, displays the components of a signal on the 50 cm × 50 cm screen of a separate c.r.o., the horizontal scale giving the frequency and the vertical the voltage of the various components.

621.317.372 **1851**

A New Q-Meter.—L. Liot. (*Télévis. franç.*, Dec. 1946, No. 20, Supplement *Électronique*, pp. 6–7.) Describes the principles and operation of the Aschen meter. For the original account see 1850 above.

621.317.39 : 531.76 **1852**

The Measurement of Small Time Intervals by Electrical Bridge Methods.—L. U. Hibbard. (*Proc. Instn Radio Engrs, Aust.*, Dec. 1946, Vol. 7, No.

12, pp. 18–19.) Abstract of a paper read before the Physics Section of the 25th Congress of the Australian and New Zealand Association for the Advancement of Science. Wartime developments for radar equipments included a coarse and fine potentiometer with an accurately exponential law and a balanced amplifier for the bridge circuit output.

21.317.431— **1853**

B-H Meter for Samples of Small Cross-Sectional Area.—E. C. Crittenden, Jr, C. S. Smith, Jr, & . O. Olsen. (*Rev. sci. Instrum.*, Oct. 1946, Vol. 17, No. 10, pp. 372–374.) Displays on a c.r.o. the ysteresis loop of samples with cross-sections of he order of 10^{-5} cm².

21.317.6 : 621.396.619 **1854**

Direct-Reading Modulation Meter.—D. W. tchley, Jr, & R. E. Fricks. (*QST*, Feb. 1947, ol. 31, No. 2, pp. 55–57 . . 146.) The meter makes se of the 1N34 germanium crystal diode, thus voiding the necessity of power supplies and iminating the triode amplifier stage. Uniformity these crystal diodes renders individual calibration nnecessary.

1.317.6 : 621.396.619.11 **1855**

On an Amplitude Modulation Meter for Re-iving Stations.—S. M. Dasgupta. (*Electro-hnics*, Dec. 1946, No. 19, pp. 64–68.)

1.317.7 **1856**

Trend and Development in Meters and Instru-nts.—L. J. Matthews. (*J. Instn elect. Engrs*, rt I, Jan. 1947, Vol. 94, No. 73, pp. 39–44.) mmary of the inaugural address by the Chairman the I.E.E. Measurements Section. Several types instrument for measurement or control are luded, though the main subject is electricity pply meters. For other accounts see *Electrician*, Nov. 1946, Vol. 137, No. 3570, pp. 1225–1226, d *Engineer, Lond.*, 8th Nov. 1946, Vol. 182, No. 9, pp. 414–415.

.317.7 : 621.396.677 **1857**

adio-Frequency Measurements on Rhombic ennae.—W. N. Christiansen, W. W. Jenvey & D. Carman. (*A.W.A. tech. Rev.*, Dec. 1946, 7, No. 2, pp. 131–144.) Two instruments described for investigating performance. One impedance meter and the other is for measuring r.f. current flowing in an elevated horizontal ductor. The use of these instruments is de-bed and some typical results given.

317.71/.72] : 621.396.694.015.33 **1858**

ulse-Type Tester for High-Power Tubes.—C. Easton & E. L. Chaffee. (*Electronics*, Feb. , Vol. 20, No. 2, pp. 97–99.) For measuring ultaneously peak voltage and current for both e and grid of a triode, when the grid current ists of very short pulses.

317.72.027.7 **1859**

eter for High Voltage Measurement.—H. én & S. Eklund. (*Rev. sci. Instrum.*, Oct. 1946, 17, No. 10, pp. 353–356.) An instrument ring the range 2–100 kV and using a capacitance ge divider combined with a special automatic ing device.

621.317.725 **1860**

Measuring Complex Components of Voltage.—G. E. Pihl. (*Electronics*, Feb. 1947, Vol. 20, No. 2, pp. 128–131.) To the unknown voltage, assumed of sinusoidal waveform and reasonably good frequency stability, a relatively large reference voltage is added and the resultant is measured by an a.f. valve voltmeter arranged so that the component of the unknown voltage in phase with the reference voltage is measured directly. The phase of the reference voltage is then advanced by 90° and the imaginary component is measured similarly. Circuit details and operation are described. Full scale range can be 1, 5, 10 or 50 V, while the frequency range is 20–20 000 c/s.

621.317.725 **1861**

Super-Sensitive Voltmeter [WV-73A].—(*Radio, N.Y.*, Jan. 1947, Vol. 31, No. 1, pp. 28–29.) An R.C.A. instrument combining a high gain audio amplifier and diode rectifier in the meter circuit.

621.317.73.029.63 **1862**

Precision Impedance Measurement Apparatus for Decimetre Waves.—H. Klauser. (*Brown Boveri Mitt.*, Aug. 1946, Vol. 33, No. 8, p. 223.) An accurately constructed Lecher-wire system with sliding bridge is used for the measurement of the reflection coefficient, from which the unknown impedance is easily calculated.

621.317.794.029.6 **1863**

Apparatus for Radiation Measurements on Metre and Decametre Waves.—F. Carbenay. (*C. R. Acad. Sci., Paris*, 17th July 1946, Vol. 223, No. 3, pp. 143–144.) A local generator produces a current equal in value to that in the receiving aerial, coil or dipole to be measured, by means of a coupling circuit using toroidal coils. The wavelength of the coupling circuit is small compared with that of the wave to be measured, so that the mutual inductance between coupling circuit and aerial is effectively constant up to high frequencies. This mutual inductance is measured when the apparatus is calibrated. The coupling circuit includes a thermoelectric couple for current measurement. Operational details are discussed.

621.396.619.15.083 **1864**

Keying Monitors.—W. A. Roberts. (*Wireless World*, Feb. 1947, Vol. 53, No. 2, p. 73.) Circuit diagram and description of a keying monitor which is in effect an a.f. oscillator with the high voltage "obtained by self-rectification of the signals radiated from the transmitting aerial and picked up by the monitor".

679.5 : [621.315.616 + 539.4 **1865**

Standard Tests for Thermosetting Plastics.—(*Schweiz. Arch. angew. Wiss. Tech.*, Oct. 1944, Vol. 10, No. 10, pp. 323–330.) Test methods standardized by the Swiss Society of Machine Constructors for measuring dielectric breakdown strength, dielectric constant and loss factor, coefficient of linear expansion, density and modulus of elasticity.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

537.533.72 **1866**

The Study of Very Small Structures with the Electron Microscope.—L. de Broglie. (*C. R. Acad.*

Sci., Paris, 30th Sept. 1946, Vol. 223, No. 14, pp. 490–493.) Further discussion of the effect of the impact of particles from the source on the object (see 1545 of May) shows that when the particles are sufficiently rapid to enable details of atomic structure to be discerned, they are liable to detach the atom from the parent body and even to modify its internal structure by withdrawing electrons. It would therefore appear that, even with microscopes using very high voltage protons, it will be impossible to see the structure of atoms.

537.533.73 **1867**

Measurements of the Relative Intensity in Electronic Debye-Scherrer Diagrams.—S. v. Friesen & S. Lenander. (*Ark. Mat. Astr. Fys.*, 11th Feb. 1947, Vol. 33, Part 4, Section A, No. 20, 16 pp. In English.) Measured intensity distribution differs widely from theoretical predictions; the difference is in the same direction as that found by Ornstein.

538.691 : 539.163.2 **1868**

An Inhomogeneous Ring-Shaped Magnetic Field for Two-Directional Focusing of Electrons and Its Application to β-Spectroscopy.—N. Svartholm & K. Siegbahn. (*Ark. Mat. Astr. Fys.*, 11th Feb. 1947, Vol. 33, Part 4, Section A, No. 21, 28 pp. In English.) Dispersion and resolving power are included in a detailed account of the theory. Constructional details are given of a β-ray spectrograph using this focusing system.

539.16.08 **1869**

Types of Geiger-Müller Counters [Zählrohr und Spitzenzähler].—C. Brinkmann. (*Z. InstrumKde*, Jan./March 1944, Vol. 64, Nos. 1/3, pp. 46–64.)

539.16.08 **1870**

The Geiger-Müller Counter with a Hollow Anode.—T. Mikhaleva. (*J. Phys., U.S.S.R.*, 1946, Vol. 10, No. 3, p. 296.)

539.16.08 **1871**

Geiger-Müller Counter Characteristics in the Neher-Harper Circuit.—K. Siegbahn. (*Ark. Mat. Astr. Fys.*, 6th Feb. 1947, Vol. 33, Part 3, Section B, No. 7, 5 pp. In English.) The marked effect of the pentode grid potential upon the characteristics of tube counters is shown graphically, and must be examined whenever a new size and type of G.-M. counter is introduced into the circuit.

539.16.08 **1872**

The Theory of Misses in an Electromagnetic Numerator at the Output of a Dividing Scheme with a Geiger-Müller Counter.—E. Berlovich. (*Zh. eksp. teor. Fiz.*, 1946, Vol. 16, No. 6, pp. 547–552. In Russian, with English summary.)

539.16.08 **1873**

Statistics of Misses in Geiger-Müller Counters.—E. Berlovich. (*Zh. eksp. teor. Fiz.*, 1946, Vol. 16, No. 6, pp. 543–546. In Russian, with English summary.) A new formula for correcting for misses is derived, and existing theories are criticized.

550.83 : 621.317.49 **1874**

The Airborne Magnetometer.—G. Muffly. (*Geophys.*, July 1946, Vol. 11, No. 3, pp. 321–334.) A discussion of the problem of magnetic exploration for oil from an aircraft, with a detailed description of apparatus using an automatically-stabilized total-field magnetometer for continuous recording.

621.317.39 : 531.7 **1875**

Electric Amplification in Metrology.—J. Villey. (*Rev. sci., Paris*, Feb. 1945, Vol. 83, No. 3241, pp. 114–117.)

621.318.572 **1876**

Measurement of High Intensities with the Geiger-Mueller Counter.—C. O. Muehlhause & H. Friedman. (*Rev. sci. Instrum.*, Nov. 1946, Vol. 17, No. 11, pp. 506–510.) The maximum counting rate of a typical G.-M. counter may be extended to 100 000 per second by providing high pulse amplification. If a single large counter is replaced by a parallel combination of several small counters the limit of resolving power is set by the speed of the electronic counter.

621.318.572 : 519.283 **1877**

On the Distribution of Counts in a Counting Apparatus.—N. Hole. (*Ark. Mat. Astr. Fys.*, 6th Feb. 1947, Vol. 33, Part 3, Section B, No. 8, 8 pp. In English.) Continuation of 3336 of 1946, giving further calculations on the statistical problems arising in connexion with a counting apparatus receiving impulses with a known distribution of intervals.

621.318.572.015.33 **1878**

Final Stage of an Electronic Pulse Counter registering once for Any Number of Received Pulses.—T. Kahan & A. Kwartiroff. (*C. R. Acad. Sci., Paris*, 9th Dec. 1946, Vol. 223, No. 24, pp. 988–989.) Circuit arrangements with practical details are described whereby the received pulses, whatever their nature, are transformed into a series of identical pulses. An accumulation stage is then used to obtain a scale of the required order.

621.365 **1879**

Industrial Applications of High Frequency.—H. Baumgartner. (*Brown Boveri Mitt.*, Aug. 1946, Vol. 33, No. 8, pp. 204–210.) Curves show the changes in permittivity and loss factor of several rubbers during h.f. vulcanization, and of whitewood during h.f. drying. Methods of case-hardening gear teeth are described and the results illustrated.

621.365.92 **1880**

Radio-Frequency Heating, with Special Reference to Dielectric Heating.—J. H. Suker. (*J. Instn elect. Engrs*, Part I, Jan. 1947, Vol. 94, No. 73, pp. 55–56.) A general survey of the principles and methods of r.f. heating at present in use, and of the design of electronic r.f. power generators. Summary of an I.E.E. Students' section paper.

621.38 : 6 **1881**

Electronics in Industry.—H. A. Thomas. (*Elect. Rev., Lond.*, 24th Jan. 1947, Vol. 140, No. 3609, p. 154.) Summary of an I.E.E. paper. For another account see 1531 of May.

621.38 : 621.317.7 **1882**

Electronics in Measurements.—R. J. Kryter. (*Elect. Engng, N.Y.*, Jan. 1947, Vol. 66, No. 1, pp. 31–35.) The advantages and limitations of electronic methods are discussed and some typical applications to measurements not otherwise possible are described.

621.38.001.8 : 667.64 **1883**

Electrostatic Painting.—V. Zeluff. (*Sci. Amer.*, Dec. 1946, Vol. 175, No. 6, pp. 252–254.) Equipment and technique using electrostatic fields to control paint particles and so prevent waste and excess deposits on objects being sprayed.

621.384.6 **1884**

20-Million-Volt Betatron.—(*Mech. Engng, N.Y.*, Jan. 1947, Vol. 69, No. 1, pp. 34–35.) The first betatron to be used for industrial radiographic inspection has recently been installed at the Picatinny Arsenal, Dover, N.Y.

621.384.6 **1885**

Largest Cyclotron ready for Service.—(*Mech. Engng, N.Y.*, Jan. 1947, Vol. 69, No. 1, p. 84.) The 184-inch University of California cyclotron is now ready for service and is expected to produce protons with an energy of 3.5 × 10^8 eV, and α-particles of 4 × 10^8 eV.

621.384.6 : 621.317.311 **1886**

A Cyclotron Beam Current Integrator and Re- order.—E. A. Hamacher. (*Rev. sci. Instrum.*, Oct. 1946, Vol. 17, No. 10, pp. 364–368.)

21.384.6 : 621.396.611.4 **1887**

The Study of a Certain Type of Resonant Cavity nd Its Application to a Charged Particle Accelerator. —Akeley. (*See* 1708.)

21.385.83.032.29 **1888**

High Current Electron Guns.—Field. (*See* 1967.)

21.385.833 **1889**

Report of the Electron Microscope Society of merica's Committee on Resolution.—W. G. insinger, J. Hillier, R. G. Picard & H. W. Zieler. *. appl. Phys.*, Dec. 1946, Vol. 17, No. 12, pp. 9–995.)

21.386 **1890**

A Survey of Engineering Radiography.—V. E. ullin. (*Engineer, Lond.*, 10th Jan. 1947, Vol. 183, o. 4748, pp. 40–42.) A general account of the pplication of X rays to the inspection of light alloy stings, steel castings, fuses, welds and complex semblies.

1.392.029.63 : 535.33 **1891**

Microwave Spectroscope.—(*Radio, N.Y.*, Jan. 47, Vol. 31, No. 1, p. 26.) For spectrochemical alysis of vapours, using wavelengths from 1.2 to 5 cm. The vapour under test is sealed in a veguide excited by an oscillator swept in chronism with the horizontal sweep of a c.r.o. e output of the crystal detector gives vertical lexions for the various characteristic absorption nds.

1.396 : 539.172.4 **1892**

Electronics — Workhorse at Bikini. — (*Elect. gng, N.Y.*, Jan. 1947, Vol. 66, No. 1, pp. 6–10.) scription of instrumental work associated with Bikini atom bomb tests. In order to discover causes as well as the effects of damage to ndard military equipment, arrangements were de for (*a*) direct observation by television of the st and its effects, including wave heights, telemetering of air and water pressures and ioactivity, (*c*) study of electromagnetic wave cts including radio and radar monitoring, and precise measurement of the relative timing of losion phenomena. Drone boats were guided the contaminated area by remote control. Summary of papers presented by D. G. Fink and T. D. Hanscome at a joint meeting of the A.I.E.E. (New York section), the New York section of the I.R.E., and the Radio Club of America.

621.396.615.029.58 : 545.38 **1893**

Use of High-Frequency Oscillators in Titrations and Analyses.—F. W. Jensen & A. L. Parrack. (*Industr. Engng Chem.* (*Analyt. Edit.*), Oct. 1946, Vol. 18, No. 10, pp. 595–599.) Changes of the ionic or dipole content in ionized or un-ionized solutions are observed by means of the changes in anode current of a 15-Mc/s oscillator with tuned anode and tuned grid, the liquid being in a container placed inside the anode coil. The method gives sharp indication and has the advantage that no electrodes are inserted in the liquid.

621.396.66 : 371.3 : 534.321.9 **1894**

H_2S Trainer.—Dummer. (*See* 1797.)

621.398 : 621.396.933 **1895**

Electronic Commutation for Telemetering.—L. L. Rauch. (*Electronics*, Feb. 1947, Vol. 20, No. 2, pp. 114–120.) A method of time-division multiplexing wherein short samples of the various instrument or gauge readings are converted into pulse signals and transmitted in cyclic serial order over a f.m. radio link. 18 channels can each be sampled 952 times per second. A master pulse generator controls broken-ring trigger stages. The main application is to flight tests of new or pilotless aircraft.

621.791.76 **1896**

Conversion of Energy into Heat during the Transient Stage in Resonant Circuits. Application to Electric Welding.—D. Genkin. (*C. R. Acad. Sci., Paris*, 17th July 1946, Vol. 223, No. 3, pp. 140–141.) If the circuit has only resistance and inductance the quantity of heat released at a point weld in the first 3–5 cycles depends on the phase of the e.m.f. at the instant of closing, so that some welds are burnt while others are insufficiently welded. With a capacitor connected in series and tuned to resonance with the supply frequency the heat developed increases regularly with the time whatever the phase of the e.m.f. and hence much better control of spot welding is possible.

537.533.72 : 621.385.833 **1897**

The Electron Microscope. Book Review,—E. F. Burton & W. H. Kohl. Reinhold Publishing Corporation, New York, 2nd edn 1946, 325 pp., $4. (*Rev. sci. Instrum.*, Oct. 1946, Vol. 17, No. 10, pp. 441–442.) The second edition, with the text rearranged and many parts rewritten to produce a greatly improved volume. The book is addressed essentially to the non-technical reader.

621.386 : 620.11 **1898**

Engineering Radiography. Book Review,— Emmott & Co., Manchester, 57 pp., 2s. 6d. (*Electrician*, 10th Jan. 1947, Vol. 138, No. 3580, p. 134.) "Mechanical World" Monograph No. 26. An introduction to the application of X rays and gamma rays to analysis and inspection.

PROPAGATION OF WAVES

538.569.4 621.396.812].029.64 : 551.51 **1899**

Atmospheric Absorption Measurements with a Microwave Radiometer.—Dicke, Beringer, Kyhl & Vane. (*See* 1843.)

621.3.011.5.029.5/.6] : 631.437 **1900**

Dielectric Properties of Indian Soils at High and Medium Radio-Frequencies.—S. R. Khastgir, J. N. Ray & A. Banerjee. (*Indian J. Phys.*, Aug. 1946, Vol. 20, No. 4, pp. 119–147.) Dielectric properties of various soils and their variation with packing, moisture content, and frequency of alternating field are examined by differential transformer and oscillographic methods. Theory and experimental procedure are given. The constants of soils from various parts of India are compared.

621.396.11 **1901**

Solution of the Problem of Propagation of Electromagnetic Waves along the Earth's Surface by the Method of Parabolic Equation.—M. Leontovich & V. Fock. (*Zh. eksp. teor. Fiz.*, 1946, Vol. 16, No. 7, pp. 557–573.) Application of Leontovich's parabolic equation method to a plane earth leads to the well known Weyl–van der Pol formula; applied to a spherical surface it gives the formula Fock obtained by infinite summation of a more rigorous solution.

621.396.11 **1902**

On the Propagation of Radio Waves along an Imperfect Surface.—E. Feinberg. (*J. Phys., U.S.S.R.*, 1946, Vol. 10, No. 5, pp. 410–418.) Formulae are derived for the field perturbation and change of direction of propagation produced by a surface composed of sea, a linear shore slope, and land of finite conductivity. A term representing the effect of the transition zone must be added to equations previously used in the treatment of coastal refraction. Some relevant experiments are also discussed. For parts 1, 2, and 3 see 2529 of 1945 and 1962 of 1946.

621.396.81.029.64 **1903**

Millimetre Wave Propagation.—H. R. L. Lamont & A. G. D. Watson. (*Nature, Lond.*, 28th Dec. 1946, Vol. 158, No. 4026, pp. 943–944.) Propagation tests over sea at wavelengths of 5.81 mm and 6.35 mm are described and received signal/distance curves given. At λ 5.81 mm, the results indicate an atmospheric absorption of about 1.5 db/km. At λ 6.35 mm the mean level closely follows the expected inverse distance law. From laboratory experiments at 8.7 mm, the reflection coefficient of a plane sea surface for grazing angles is between 0.88 and 0.97. See also 3396 of 1946.

621.396.812 : 551.510.535 **1904**

Radio Waves and the Ionosphere.—Howe. (*See* 1779.)

621.396.812 + 551.510.535 **1905**

Echoes on Short Waves at Low Altitude.—Jouaust. (*See* 1776.)

621.396.812.4 : 551.510.535 **1906**

The Validity of Ionospheric Forecasts.—T. W. Bennington. (*B.B.C. Quart.*, Oct. 1946, Vol. 1, No. 3, 9 pp. Reprint.) A general article on the prediction and use of optimum working frequencies (O.W.F.), assessed as 15% below the maximum usable frequency (M.U.F.). M.U.F. predictions are based on normal ionospheric and solar variations, and are plotted as ionospheric contour charts on a Mercator projection, the surface of the earth being covered in three zones. Increased knowledge of the effects of sporadic-E ionization may lead to the provision of sporadic-E contour charts for use in conjunction with ionospheric contour charts. Sudden ionospheric storms are described, and the difficulty of forecasting them is pointed out.

621.392.52 + [548.0 : 538.3 **1907**

Wave Propagation in Periodic Structures : Electric Filters and Crystal Lattices. [Book Review]—L. Brillouin. McGraw Hill Book Co., New York & London, 1946, 247 pp., 20s. (*Nature, Lond.*, 28th Dec. 1946, Vol. 158, No. 4026, p. 926.) The propagation of waves in two- and three-dimensional lattices and the filtering properties of such systems are discussed. The application of the same mathematical treatment to problems of pure physics is also explained.

RECEPTION

621.385.3 + 621.396.694] : 621.396.645 **1908**

Grounded-Grid Technique.—Knowles. (*See* 1965.)

621.394/.397].822 : 530.162 **1909**

Study of Fluctuations produced by the Shot Effect in Amplifiers.—Blanc-Lapierre. (*See* 1841.)

621.396.6 **1910**

A Quiet Break-In System.—C. L. Robinson (*QST*, Feb. 1947, Vol. 31, No. 2, pp. 33–35.) A description of a smooth receiver-silencing system.

621.396.61/.62 **1911**

Operating the BC-645 on 420 Mc/s.—Ralph & Wood. (*See* 1957.)

621.396.62 : 621.396.611.21 **1912**

Receiver with Automatic Tuning and Quartz Control.—A. V. J. Martin. (*Radio en Franc*, 1947, No. 1, pp. 17–18.) A receiver circuit is described, with detailed diagram, using quartz control of the local oscillator frequency. The input circuit is tuned to different stations by means of pre-set capacitors.

621.396.621 + 621.396.69 **1913**

New Methods of Radio Production.—J. A. Sargrove. (*J. Brit. Instn Radio Engrs*, Jan./Feb. 1947, Vol. 7, No. 1, pp. 2–33.) Insulating plates are moulded into such a structural form that, when fully processed without manual aid, they contain the inductors, capacitors, resistors, potentiometer tracks, switch and other terminations, together with conductor paths, in an interrelated and interconnected manner, the plate material forming the dielectric of the capacitors. Full details are given of the methods adopted for making connexions through the plates for larger capacitors and inductances, fitting valve sockets, using the inside walls of the moulded cabinet for the larger resistors, etc. The manufacturing process is completely automatic. The plates, after infra-red drying, sandblasting and cleaning to remove grit, are metallized on both sides by spraying and then surface-milled to remove all projections, leaving the metal only in the grooves and depressions. The resistors are added by graphite spraying through stencils. Automatic timing ensures the accuracy of resistor values. Successive machines remove unwanted plastic, metal or graphite from holes, slots, etc., and insert metal sockets for valves, electrolytic capacitors, loudspeakers etc. The sockets are fixed permanently by a combined riveting, welding and soldering operation. After electrical overload tests and thermal aging the plates are lacquered, dried and passed to a conveyor belt where other parts are added, some manually. Automatic testing is applied at every stage of the process. Failure of any stage shuts down all previous ones. After a final radio signals test, the completed sets are automatically packed and sealed. The complete outfit is known as E.C.M.E.—Electronic—Circuit-Making Equipment—and has obviously wide applications to the speedy production of cheap, light-weight and uniform communication pparatus of many types.

621.396.621 **1914**

Ekco Model A28.—(*Wireless World*, Feb. 1947, Vol. 53, No. 2, pp. 54–56.) Test report on a table model receiver for a.c. mains.

621.396.621.5 : 573.312.62 **1915**

Demodulation by Superconductivity.—D. H. Andrews & C. W. Clark. (*Nature, Lond.*, 28th Dec. 1946, Vol. 158, No. 4026, pp. 945–946.) Demodulation of broadcast waves can be effected with superconducting bolometers of columbium itride operated within certain temperature limits. Quality is comparable with that of good standard adio reception.

21.396.621.54 **1916**

Home-Constructed Communications Superhet.—. B. Wright. (*Short Wave Mag.*, Oct. & Nov. 46, Vol. 4, Nos. 8 & 9, pp. 470–478 & 546–551.) onstructional details for a receiver of conventional esign suitable for amateur use.

1.396.621.54 **1917**

“Single-Span ”, a 4-Valve Single Range Superhet r A.C. Mains.—V. Stuzzi. (*Radio Welt*, Dec. 1946, ol. 1, No. 4, pp. 69–74.) The range is 200–2 000 m ithout switching. Constructional and circuit tails are given.

1.396.662 : 621.396.62 **1918**

Tuning Devices for Broadcast Radio Receivers.— C. G. Williams. (*J. Instn elect. Engrs*, Part I, n. 1947, Vol. 94, No. 73, pp. 59–61.) Summary 1194 of April.

1.396.8.015.33 : 621.396.82 **1919**

A Theoretical Comparison of the Visual, Aural, d Meter Reception of Pulsed Signals in the esence of Noise.—J. H. Van Vleck & D. Middle-. (*J. appl. Phys.*, Nov. 1946, Vol. 17, No. 11, 940–971.) The paper gives detailed theoretical dies of the dependence of the sensitivity of the ee methods on various parameters including the dth and shape of the i.f. response, the pulse gth, the pulse recurrence frequency and, in al or meter reception, the duration of the gate, width of the audio filter, and the time constant he meter.

.396.812.3 **1920**

iversity Reception.—Z. Jelonek, E. Fitch & H. H. Chalk. (*Wireless Engr*, Feb. 1947, Vol. 24, 281, pp. 54–62.) The diversity gain is a measure he permitted reduction of signal level for a given tional time loss of signal when more than one al is used. This is evaluated for two- and e-signal systems, assuming independent fading a Rayleigh or a square-law probability-integral tion. The bivariate Gaussian distribution is used to study the case of correlated two-signal fading. Experimental data on fading are given for ionospheric waves and for centimetre-wave propagation over long paths.

621.396.82 : 523.5 **1921**

Radar Detection of Meteor Trails.—Appleton & Naismith. (*See* 1752.)

621.396.821 : 551.594.221 **1922**

The Nature and Variation of Atmospherics caused by Lightning Discharges.—Norinder & Stoffregen. (*See* 1784.)

621.396.822.029.63 : 523.72 **1923**

Solar Radiation at 480 Mc/s.—Reber. (*See* 1760.)

621.396.828 : 061.3 **1924**

Radio Interference.—(*Wireless World*, Feb. 1947, Vol. 53, No. 2, p. 67.) Brief note of an international conference in London called by the British Standards Institution to consider radio interference suppression.

621.396.828 : 551.508 **1925**

Reduction of Radio Interference from Meteorological Installations.—R. L. Ives. (*Bull. Amer. met. Soc.*, Feb. 1946, Vol. 27, No. 2, pp. 59–61.) Interference is of three types, stray magnetic fields, l.f. oscillations and r.f. oscillations. Methods for its reduction are given.

621.396.828 : 621.396.645.029.3 **1926**

Hum in High-Gain Amplifiers.—Baxandall. (*See* 1689.)

STATIONS AND COMMUNICATION SYSTEMS

621.39 + 534.756 **1927**

Theory of Communication.—D. Gabor. (*J. Instn elect. Engrs*, Part I, Jan. 1947, Vol. 94, No. 73, p.58.) Summary of 1057 of April.

621.396.215 **1928**

Transmitter-Receiver Broadcasting System with Two Channels on a Single Carrier Wave.—J. Donnay. (*Radio en France*, 1945, No. 4, pp. 3–5 & 1946, No. 1, p. 18.) A rectangular signal with frequency of the order of 10 to 40 kc/s is applied to the transmitter; the positive and negative half-cycles cut out the audio modulation for the first and second channels respectively. At the same time the amplitude of the crests of the rectangular signal is modulated by the audio frequency to be transmitted on each channel. A signal of complex modulation is thus obtained which can be used for the transmission of two programmes. A suitable discriminating system in the receiver allows programme selection. In practice it is found preferable to separate the positive and negative crests by short quiescent periods.

621.396.24.029.63/.64 **1929**

Problems of Intelligence [speech or music] **Transmission on Microwaves.**—H. J. v. Baeyer. (*Brown Boveri Mitt.*, Aug. 1946, Vol. 33, No. 8, pp. 198–203.) A general discussion, with special treatment of the problems of generation, transmission and reception of wavelengths between 2 and 20 cm. Comparison is made between single channel a.m. and f.m., multichannel methods using frequency transposition for each channel, and pulse modulation systems.

The latter are particularly suitable when the number of channels is large.

621.396.44 **1930**

Development and Present Position of Multichannel Directional Technique.—E. Huber. (*Brown Boveri Mitt.*, Aug. 1946, Vol. 33, No. 8, pp. 182–185.) From a discussion of possible systems it is concluded that for a small number of channels (up to about 12) an adaptation of cable carrier-frequency technique gives the best results. A short description is given of two equipments, one providing two telephony and two telegraphy channels (a modification of the military equipment described in 1937 below) and the other six telephony channels, f.m. being used in both sets.

621.396.61/.62].029.64 **1931**

Dishing Out the Milliwatts on 10 kMc/s.—J. A. McGregor. (*QST*, Feb. 1947, Vol. 31, No. 2, pp. 58–61, 148.) The transmitter uses a type 723A/B valve and gives about 35 mW. With a 30-inch parabolic reflector and dipole aerial, very strong signals were obtained at a range of 2 miles, using a Hallicrafter S-29 portable receiver.

621.396.619.16 **1932**

The Application of Pulse Technique to Broadcasting.—H. L. Kirke. (*B.B.C. Quart.*, Oct. 1946, Vol. 1, No. 3, 6 pp. Reprint.) If the pulse amplitude in the receiver is limited, the signal/noise ratio is proportional to the steepness of the sides of the pulse provided the bandwidth passes the pulse without distortion. The pulse duration determines the ratio of the pulse-recurrence frequency to the highest audio frequency transmitted. A comparison with other systems shows that for a high-fidelity, low-noise service, pulse modulation is inferior to frequency modulation. The apparatus for a single-transmitter, multichannel system is simple but a greater bandwidth is required than for a frequency modulation system with separate transmitter. There is also a danger of cross-talk in the pulse-modulation system due to multi-path transmission.

621.396.619.16 **1933**

Pulse Time Modulation.—P. Güttinger. (*Brown Boveri Mitt.*, Aug. 1946, Vol. 33, No. 8, pp. 188–192.) The general principles of pulse modulation, and particularly of pulse time modulation, are explained. Broad frequency bands are required if this method is to be used to the greatest advantage, so that its main application is to microwaves. It also enables interference effects to be reduced considerably, as with f.m. A special feature of the system described gives a linear relation between the pulse lengths and the modulation values at fixed and equal time intervals which do not depend on the modulation process.

621.396.712 : 621.396.81 **1934**

Site Selection and Field Intensity Surveys for Broadcasting Stations.—R. Gill. (*Proc. Instn Radio Engrs, Aust.*, Nov. 1946, Vol. 7, No. 11, pp. 7–21.) The basic coverage requirements of a medium-wave broadcasting station are outlined and the factors governing the service area of such a station are examined. Survey procedure is described, with typical examples. The requirements for the siting of transmitters for f.m. v.h.f. operation are discussed and details given of a survey in the Sydney area using a mobile v.h.f. transmitter.

621.396.72.029.62 **1935**

First Steps on Five [metres].—J. Hum. (*Short Wave Mag.*, Nov. 1946, Vol. 4, No. 9, pp. 541–545.) Reconstruction of existing equipment to obtain a superheterodyne converter, a beam aerial, and a transmitter for 58-Mc/s amateur use.

621.396.72.029.63 **1936**

Simple Decimetre Directional Equipment.—R. Schüpbach. (*Brown Boveri Mitt.*, Aug. 1946, Vol. 33, No. 8, pp. 181–182.) The set comprises transmitter, receiver and supply units, works on a fixed frequency in the range 300–330 Mc/s and with a transmitter power of 1–2 W has a working range up to 10 km.

621.396.73.029.63 **1937**

Portable Decimetre Equipment for Military Directional Links.—R. Schüpbach. (*Brown Boveri Mitt.*, Aug. 1946, Vol. 33, No. 8, pp. 178–181.) The transmitter and receiver are of unit construction and operate in the range 330–355 Mc/s with a transmitter f.m. of ± 75 kc/s. With relay stations good communication is achieved at distances of several hundred kilometres.

621.396.75 **1938**

Directional Communication System.—H. Köhler. (*Radio Welt*, Dec. 1946, Vol. 1, No. 4, pp. 64–69.) A general account of basic principles, aerial systems, transmitting and receiving equipment.

621.396.9.029.63 **1939**

Interesting Applications of Decimetre Waves in the Communications and Remote Supervisory Control Fields.—R. Schüpbach. (*Brown Boveri Rev.*, Dec. 1945, Vol. 32, No. 12, pp. 453–456.) A survey of many possible uses of equipment at wavelengths in the region 0.8–1.5 m, which have definite advantages, particularly for beam transmission.

621.396.932 **1940**

Naval Radio Gear.—G. M. Bennett. (*Wireless World*, Feb. 1947, Vol. 53, No. 2, pp. 62–64.) Details of British naval communication systems used in large warships.

621.396.96 : 621.396.82 **1941**

Radar Reflections from Long Conductors.—Bloch, Hamermesh & Phillips. (*See* 1687.)

621.396.97(54) **1942**

Ultra Short Waves to replace Medium Waves for Regional Broadcasting in India.—S. P. Chakravarti. (*Electrotechnics*, Dec. 1946, No. 19, pp. 5–18.) A review of the broadcasting policy of India, and a comparison between medium and ultra-short waves is made. The problems of u.h.f. regional broadcasting are discussed and the advantages of ultra-short waves for dual programme transmission are stated. Finally, a scheme for India using 70 stations each of $\frac{3}{4}$ kW power working in the frequency range 60–150 Mc/s is suggested.

621.396.97.029.6 : 621.396.619 **1943**

Broadcasting at V.H.F.—T. Roddam. (*Wireless World*, Feb. 1947, Vol. 53, No. 2, pp. 70–71.) The relative merits of a.m., f.m., and pulse modulation ìor v.h.f. broadcasting are considered. In the present state of knowledge, no one of these systems should be adopted to the exclusion of the others.

SUBSIDIARY APPARATUS

ö21.3.032.53 : 621.51/.52 **1944**

Hermetic Seals and Bushes.—United Insulator ;o. (*J. sci. Instrum.*, Nov. 1946, Vol. 23, No. 11, ɔp. 272–273.) Ceramic bushes are metallized both round a shoulder, to allow soft-soldering to a metal ontainer, and on one end, for soldering a lead-in vire or metal insert. They will withstand vacuum nd pressures over 30 lb/sq. inch and have high ashover voltage and insulation resistance.

21.314.5 **1945**

Development of 150 Watts D.C.-A.C. Vibro-onverter of 50 Cycles per Second Fundamental requency.—R. N. Dewan. (*Electrotechnics*, Dec. 946, No. 19, pp. 81–88.) To convert 110/220 V .c. mains to 110/220 V single-phase a.c. of square-opped waveform having a fundamental frequency f 50 c/s. Extends work previously noted in 754 of 946.

21.314.6 + 621.319.4 + 621.383] : 669.018 **1946**

Light Alloys in Metal Rectifiers, Photocells, and ondensers.—Continuing the series in *Light Metals* entioned in 1593 of May and back references.

(xxii) July 1946, Vol. 9, No. 102, pp. 372–378. Finishing processes in the manufacture of fixed per capacitors, and certain more recent develop-ents in dielectric and impregnating media.

(xxiii) Aug. 1946, Vol. 9, No. 103, pp. 408–413. Aluminium and pigmented cellulose lacquers and amels, with special reference to corrosion resistance d durability tests.

(xxiv) Oct. 1946, Vol. 9, No. 105, pp. 517–524. Methods of identification marking of fixed paper pacitors, and inspection systems and testing chniques before dispatch.

(xxv) Dec. 1946, Vol. 9, No. 107, pp. 637–642. The use of plastic film in place of paper and mica the construction of fixed capacitors.

1.316.93 **1947**

Lightning Protection.—R. C. Cuffe. (*Elect. Rev., nd.*, 20th Dec. 1946, Vol. 139, No. 3604, p. 1011.) ɔstract of a paper presented at a meeting of the ansmission Section of the I.E.E.

1.317.755 **1948**

A Universal Oscillograph.—G. L. Hamburger. *lectronic Engng*, Jan. & Feb. 1947, Vol. 19, Nos. 227 228, pp. 7–10, 22 & 51–57.) A full account of the sign of a laboratory instrument of great versa-ty, with complete circuit diagram, table of com-ıents and constructional details. A direct sweep :h 5–30% flyback and a repetition rate of 0.5– 5 000 c/s is used and an elliptical timebase is also orporated. Voltage calibration of the trace is ect-reading from 100 mV to 200 V for both a.c. d d.c. The operator can calibrate adjustable dio frequencies up to about 10 kc/s in terms of ins frequency. Mechanical features include a ractable terminal panel and division of the trument into two main chassis, connected by a ge, so that all parts are easily accessible.

.385.832 **1949**

rigin of Trapezoidal Distortion in Cathode-Ray Oscillographs.—A. Cazalas. (*C. R. Acad. Sci., Paris*, 1st July 1946, Vol. 223, No. 1, pp. 27–28.) Experiments show that the distortion is due more particularly to stray lines of force from the ends of the plates which are located furthest from the source. Correction of the effect has been achieved by (*a*) making these plates circular, the centre of the circle being on the axis around which the beam of electrons turns when it is deflected by the other plates and (*b*) introducing between the two pairs of plates a cylindrical screen, concentric with the above circle, maintained at the final anode potential and having a window to allow passage of the beam.

621.394.652 **1950**

Telegraph Key Design.—(*Wireless World*, Feb. 1947, Vol. 53, No. 2, p. 61.) Abstract of article previously noted in 1232 of April.

621.396.682 : 621.316.722.078.3.029.6 **1951**

Power Supply for Microwave Equipment.—O. Hoag. (*Electronics*, Feb. 1947, Vol. 20, No. 2, pp. 110–113.) An electronically controlled servo motor adjusts an auto-transformer. Voltage can be varied between 300 and 5 000 V with less than 3 mV ripple in the d.c. output.

TELEVISION AND PHOTOTELEGRAPHY

621.397.5 : 621.385.83.032.29 **1952**

Electron Guns for Television Application.—Morton. (*See* 1968.)

621.397.62 **1953**

Television Receiver Construction : Part 2.—(*Wireless World*, Feb. 1947, Vol. 53, No. 2, pp. 40–42.) Construction and winding data for line deflector coils. For part 1 see 1245 of April.

621.397.621 **1954**

Image Quality and Definition.—P. Hémardinquer. (*Télévis. franç.*, Feb. 1946, No. 10, pp. 17–18 . . 22.) Discusses the optimum results to be expected from a 450-line system. Comparison is made with photographic reproduction.

TRANSMISSION

621.394.652 **1955**

Telegraph Key Design.—(*See* 1950.)

621.394.652 : 621.394.141 **1956**

The Electroplex — a New Automatic Key. — Dixon. (*See* 1699.)

621.396.61/.62 **1957**

Operating the BC-645 on 420 Mc/s.—J. T. Ralph & H. M. Wood. (*QST*, Feb. 1947, Vol. 31, No. 2, pp. 15–21.)

621.396.611.33 : 621.396.671 **1958**

Matching Ranges [plages d'adaptation] **of Transmitters.**—D. Glazer & V. Familier. (*Onde élect.*, Nov. 1946, Vol. 26, No. 236, pp. 430–437.) A transmitter is to be matched to an aerial of resistance R and reactance jX. For a particular frequency f, the possible pairs of values of R and X form an aggregate which is called the matching range (plage d'adaptation). In the general case where f, R, and X all vary, the matching range will be a skew surface when the coupling system has only one degree of freedom, and a volume when it has two degrees of freedom. A method is given, with numerical examples, for determining this matching volume for direct inductive coupling.

621.396.619.13 **1959**

Frequency Modulator.—G. G. Bruck. (*Electronics*, Jan. 1947, Vol. 20, No. 1, pp. 166..170.) Reactance tubes, double diode demodulator, and inverse feedback loop for linearity and mid-frequency control are here combined into a single unit.

621.396.645.36 **1960**

150 Watts Push-Pull.—Thomas. (*See* 1716.)

VALVES AND THERMIONICS

621.383.4 + 546.28 **1961**

A New Bridge Photo-Cell employing a Photo-Conductive Effect in Silicon. Some Properties of High Purity Silicon.—G. K. Teal, J. R. Fisher & A. W. Treptow. (*J. appl. Phys.*, Nov. 1946, Vol. 17, No. 11, pp. 879–886.) A method is described for making bridge type photocells by reaction of silicon tetrachloride and hydrogen gases at ceramic or quartz surfaces at high temperatures. The maximum photo-sensitivity occurs at 8 400–8 600 Å. The sensitivity is about equivalent to that of the selenium bridge; the silicon cell is far more stable and rapid in response. From measurements of the electronic conductivity of silicon as a function of temperature, it is concluded that the same electron bands are concerned in the photoelectric, optical and thermal processes. The low values of specific conductances found (1.8×10^{-5} mho/cm) are ascribed to high purity of the silicon, not polycrystalline structure.

621.383.5.032.21 : 537.533 **1962**

The Emission of Electrons by Active Semi-Conducting Cathodes.—N. D. Morgulis. (*Bull. Acad. Sci. U.R.S.S.*, *sér. phys.*, 1941, Vol. 5, Nos. 4/5, pp. 536–545. In Russian with English summary.) Survey of fundamental data concerning various properties of cathodes, used as thermo-emitters, photo-emitters and secondary electron emitters.

621.385 : 519.283 **1963**

Electron Tube Quality Control.—J. R. Steen. (*Radio, N.Y.*, Jan. 1947, Vol. 31, No. 1, pp. 16–19.) A survey of modern methods of control by sample testing.

621.385 : 621.317.723 **1964**

Improvements in the Stability of the FP-54 [Pliotron] **Electrometer Tube.**—J. M. Lafferty & K. H. Kingdon. (*J. appl. Phys.*, Nov. 1946, Vol. 17, No. 11, pp. 894–900.) Full paper, a summary of which was abstracted in 583 of February.

621.385.3 + 621.396.694] : 621.396.645 **1965**

Grounded-Grid Technique.—R. Knowles. (*Short Wave Mag.*, Nov. & Dec. 1946, Vol. 4, Nos. 9 & 10, pp. 536–540 & 603–608.) The uses of the grounded-grid triode as an r.f. or power amplifier, as an oscillator, or in a.f. amplifiers are discussed. Stage gain, matching, and circuit and constructional details are considered practically. Circuit values are given for a 58-Mc/s r.f. receiver amplifier. Characteristics of commercial valves suitable for grounded-grid working are tabulated.

621.385.3 : 621.396.694.012.8 **1966**

The Equivalent Diode.—W. E. Benham. (*Wireless Engr*, Feb. 1947, Vol. 24, No. 281, p. 62.) Comment on 949 of March.

621.385.83.032.29 **1967**

High Current Electron Guns.—L. M. Field. (*Rev. mod. Phys.*, July 1946, Vol. 18, No. 3, pp. 353–361.) The limitations in high current electron gun design include a maximum permissible current density, thermal velocities of emission, cathode poisoning, ion bombardment and anode heating. Single and multiple electrostatic guns, and combined electrostatic and magnetic field guns are described, with their relative advantages and limitations.

621.385.83.032.29 : 621.397.5 **1968**

Electron Guns for Television Application.—G. A. Morton. (*Rev. mod. Phys.*, July 1946, Vol. 18, No. 3, pp. 362–378.) A description of the iconoscope, orthicon, image orthicon, and the various electron guns used. A two-lens system is used in the iconoscope, and a low-velocity magnetically-focused beam in the orthicon systems. Diagrams are given for several types of commercial electron guns.

621.396.615.1 : 537.58 **1969**

Effect of Space Charge on the Frequency of Oscillation in Positive-Grid Oscillators.—C. Yeh. (*Chin. J. Phys.*, July 1946, Vol. 6, No. 2, pp. 79–84.) The effect of the non-uniformity of the inter-electrode field in a planar Barkhausen-Kurz tube due to space charge results in a frequency about 10–20% lower than in the absence of a space charge.

MISCELLANEOUS

001.4 **1970**

The Use of Standard Terms and Symbols.—G.W.O.H. (*Wireless Engr*, Feb. 1947, Vol. 24, No. 281, p. 34.) Several examples of confusion caused by slipshod use of symbols are given. See also 960 of March.

519.283 : 621 385 **1971**

Electron Tube Quality Control.—Steen. (*See* 1963).

621.3.016.25 **1972**

The Sign of Reactive Power.—E. W. Kimbark: H. K. Sels: R. H. Lindsay: G. A. Irland: W. V. Lyon. (*Elect. Engng, N.Y.*, Jan. 1947, Vol. 66, No. 1, pp. 107–110.) Further comment on 971 of March; see also 1289 of April. Agreement is expressed with the recommendation that the previously adopted sign of reactive power should be changed, but comment is made on some of the definitions suggested.

621.39 : 371.3] + 621.315.6 **1973**

The War-Time Education and Training of Radio Personnel and Recent Developments in Dielectric Materials.—W. Jackson. (*J. Instn elect. Engrs*, Part I, Jan. 1947, Vol. 94, No. 73, pp. 26–34, and Part III, Jan. 1947, Vol. 94, No. 27, pp. 2–10.) Details are given of the courses organized by the Wireless Personnel Committee for radio and wireless mechanics, radio officers, and of a post-graduate course in high vacua and electronics.

The structure of carbon, silicon, and titanium compounds, and its relation to their dielectric properties, is considered in detail.

Summaries of this paper were noted in 314 of January.