WIRELESS ENGINEER

THE JOURNAL OF RADIO RESEARCH & PROGRESS

JUNE 1953

VOL. 30 No. 6 THREE SHILLINGS AND SIXPENCE

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When purchasing a Valve Tester ask the following questions:—

Is it a simple Go/No Go instrument, or will it enable you to take measurements at any point on a characteristic curve?

The "AVO" Valve Characteristic Meter can be set up in a matter of seconds and used as a simple Go/No Go tester if required. It will, in addition, produce sufficient information to enable the **valve's static characteristic curves** to be plotted.

Does the instrument depend for its operation on pre-determined empirical data issued by its manufacturer?

The "AVO" Valve Characteristic Meter simulates normal working conditions for the valve under test and thus is capable of reproducing the **valve manufacturer's data.**

If you are called upon to select a pair of accurately matched valves, will the valve tester carry out the required checks, and maintain your reputation as an expert?

The "AVO" Valve Characteristic Meter enables the slope, and anode, screen, or grid current of a multi-electrode valve to be checked with any voltage between 0 and—100V on the control grid.

Before any information can be obtained about a valve, must the instrument be provided with a complex series of accessories which may become lost, mutilated, or are not available when an unusual or new type of valve has to be tested?

The "AVO" Valve Characteristic Meter is provided with two handbooks. The first gives detailed information on the technique of valve testing, full circuit diagrams and adequate operating instructions. The second is a quick reference Data Manual covering more than 3,000 British, American and Continental valves, and gives inter-service equivalents. The Valve Data Manual is issued for the convenience of the instrument user, but even without it **valves can, if necessary, be checked using data taken from the valve maker's data sheets.**

Will a high slope valve, when placed in the instrument, burst into spurious oscillation, thus giving rise to incorrect readings and possible damage to the valve?

The "AVO" Valve Characteristic Meter incorporates a specially designed panel layout and wiring system (prov. patent) which virtually eliminates spurious oscillation.

Does the instrument contain valves which may need replacing from time to time?

With the exception of a small protection diode, there are no internal valves to deteriorate or break down and cause misleading readings to be given. Thus, expensive periodical replacements are not required.

Will the instrument detect grid current and indicate its direction and magnitude?

The "AVO" Valve Characteristic Meter will indicate the direction of flow of grid current and give its magnitude in microamps.

Does the instrument test diodes and rectifiers under load?

The "AVO" Valve Characteristic Meter checks diodes and rectifiers under load conditions.

The "AVO" Valve Characteristic Meter measures inter-electrode insulation in megohms with valve cold or hot, also cathode/heater insulation with the valve hot, and indicates any breakdown below 10 megohms. It will carry out tests on small thyratrons, tuning indicators, etc. It is fitted with bases for most valves in current use and adaptors will be available to keep the instrument fully up-to-date should new bases come into use. A special form of polarised relay is incorporated to give protection against inadvertent overloads or valve failure.

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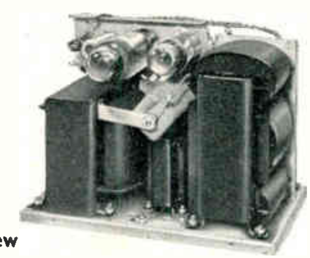
Rear View BAVR-200



Inside View BAVR-200

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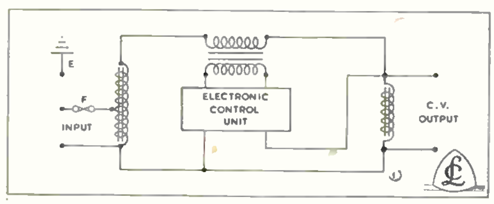

Functional Circuit BAVR

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Model BMVR-1725

* Full details now in active preparation, and will gladly be sent on written request.

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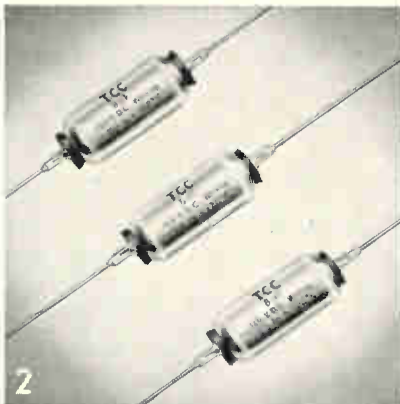

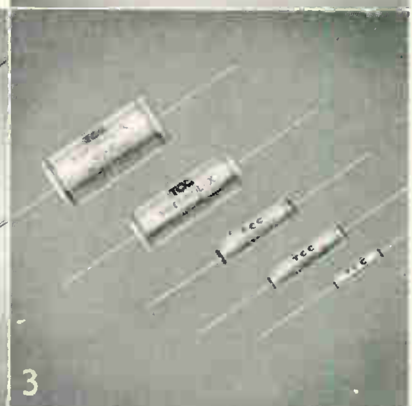

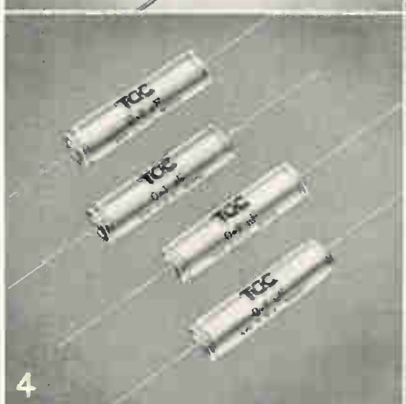

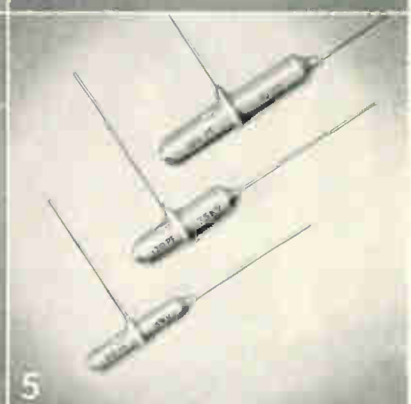

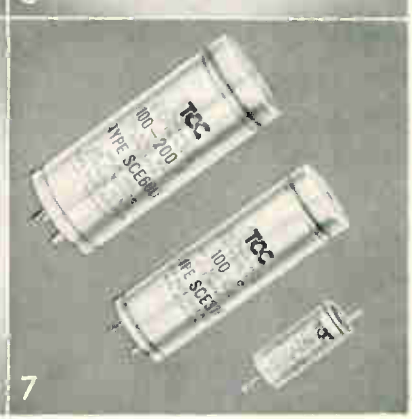

FOR the benefit of those who were unable to visit the recent R.E.C.M.F. Exhibition, and for those who had not the time available to inspect our Stand in detail, we illustrate here the range of T.C.C. latest developments displayed thereon.

These are indicative of the progress which is constantly being maintained, and which ensures that "T.C.C. Leadership in Condensers" is a tangible reality.

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For top end coupling in Band Pass Filters. Capacity range 0.5 pF to 5 pF at 500 v. D.C. Tolerances ±20% and ±10%.

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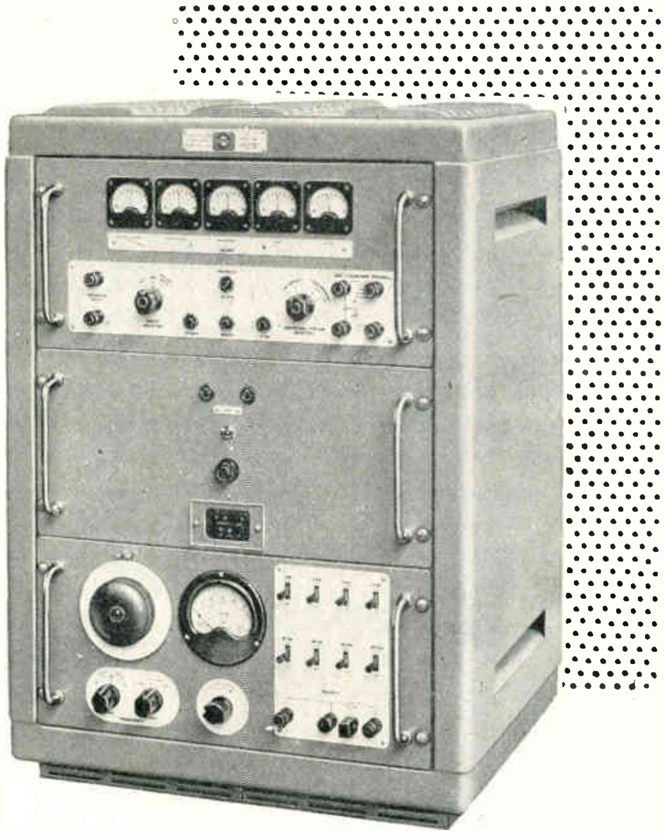

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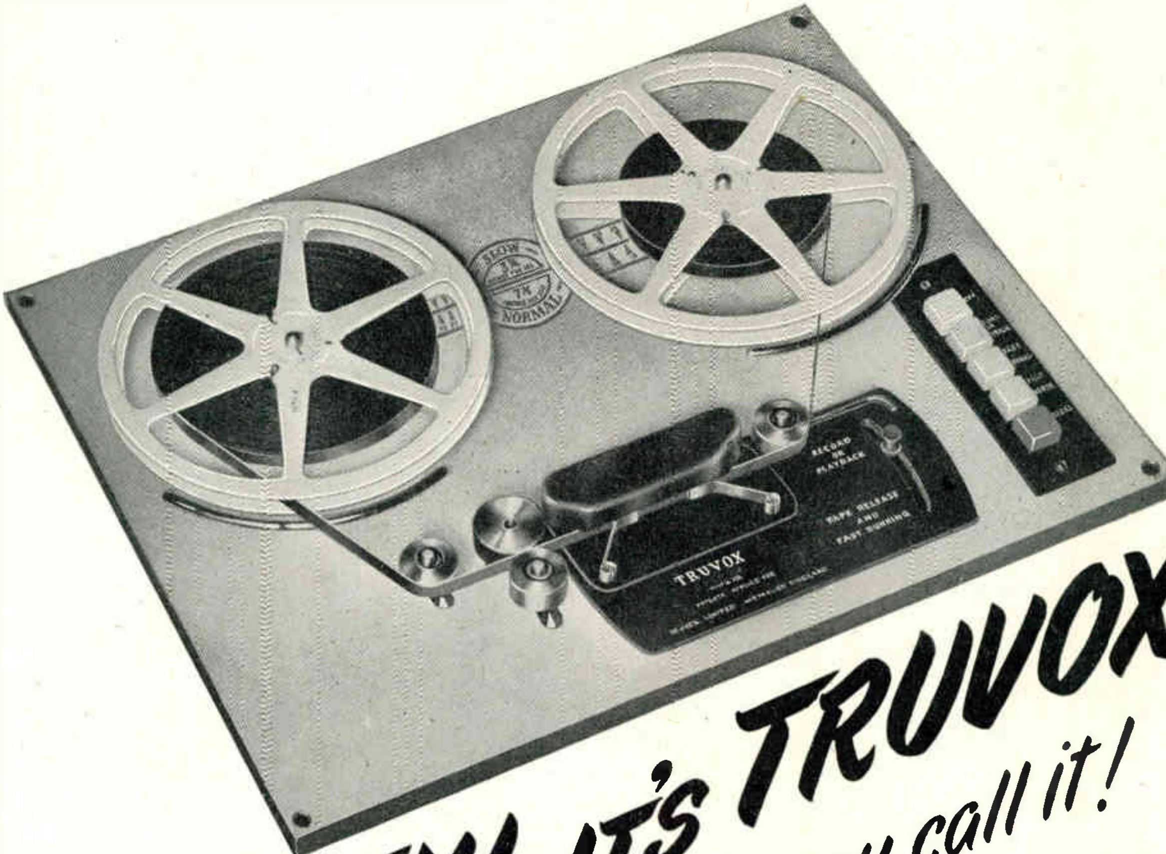

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MODEL TR2
200/250 volts
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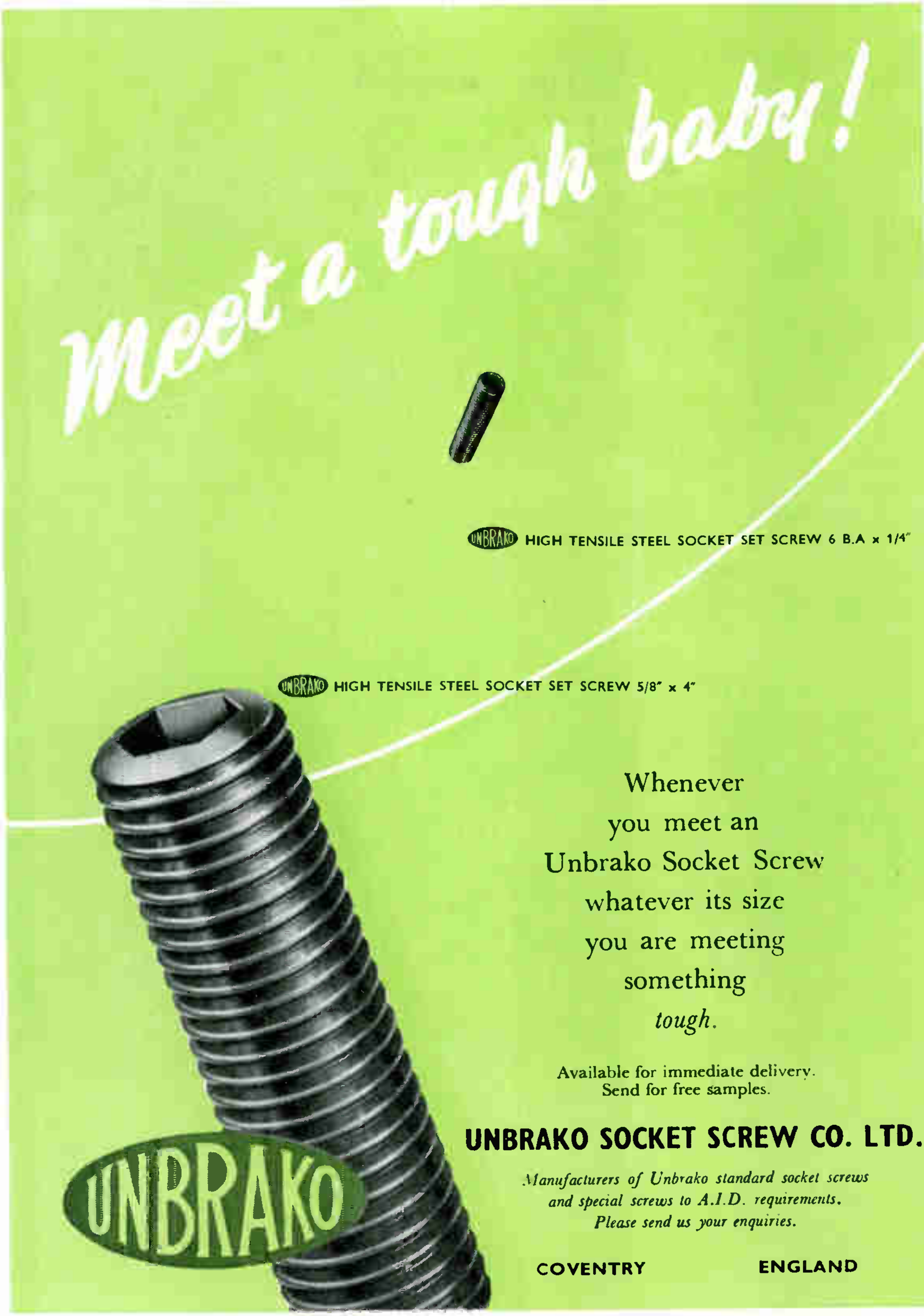
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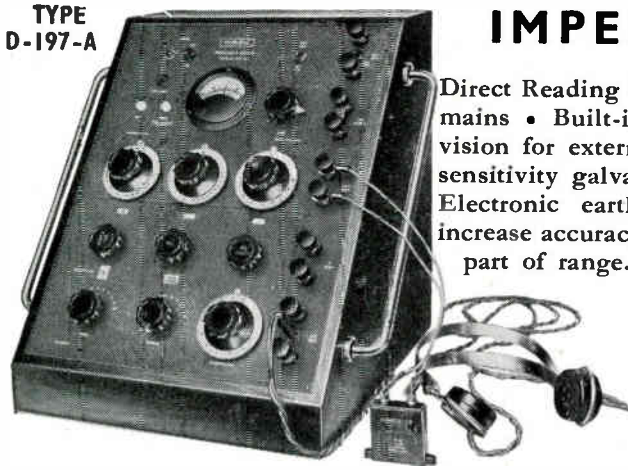

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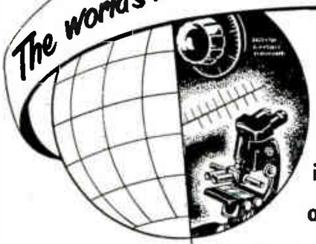

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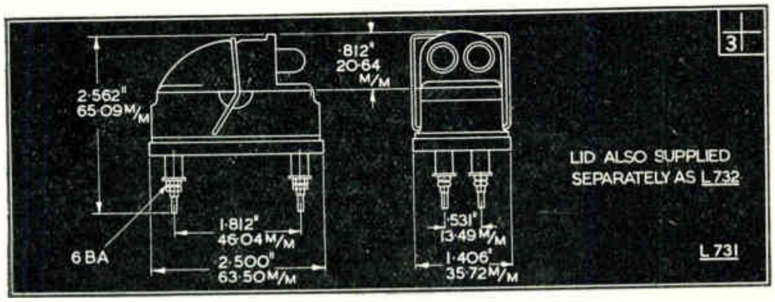
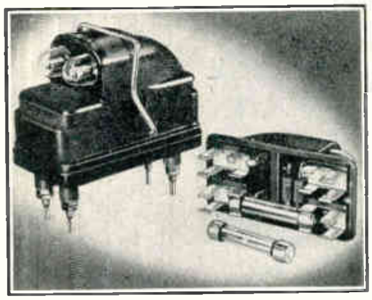
The "Belling-Lee" page for Engineers





L.732
TWIN NEON CARRIER WITH RETAINING CLIP

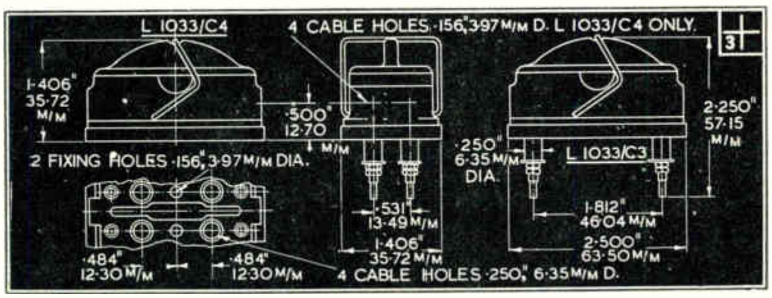
L.730
TWIN FUSEBOX WITH CARRIER AND RETAINING CLIP

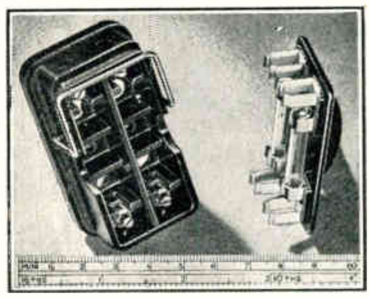
L.731
AS L.730 BUT WITH BACK CONNECTIONS

TWIN NEON INDICATING FUSEHOLDER

The twin neon indicating fuseholder L732, illustrated above top right, carries two neon lamps, and where required may be used to replace the normal fuse carrier on standard "Belling-Lee" twin fuseboxes L1033, shown below. The overall depth is thereby increased by $\frac{3}{16}$ in, but the panel area remains unchanged.

Two lamp-holders with resistors are built into the moulding, and are arranged in such a manner that the adjacent lamp gives visual indication when either fuse blows. The neons will glow satisfactorily over a voltage range of 180/250V. a.c. (r.m.s.), or 220/250V. d.c.





L.1033/C4
WITH CARRIER RETAINING CLIP

L.1033/C3
AS ABOVE, WITH BACK CONNECTIONS AND BUSHES

A replacement retaining clip is supplied with the carrier. This ensures that when the box is properly wired, the neons can only be inserted in the correct phase of the supply.

Supplied separately or complete with bases.

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In Canada, Montreal and Toronto already possess Marconi television studios installed by the Canadian Marconi Company. The new Marconi Transmitter for Ottawa is an important link in the television chain planned by Canadian Broadcasting Corporation to span the entire country.

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Equipment supplied includes:

- Marconi Image Orthicon Cameras
- 5 kW vision transmitters
- 3 kW sound transmitters
- High-gain aerial systems
- Associated monitoring and control equipment

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WIRELESS ENGINEER

The Journal of Radio Research and Progress

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Volume 30 · Number 6

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JUNE 1953

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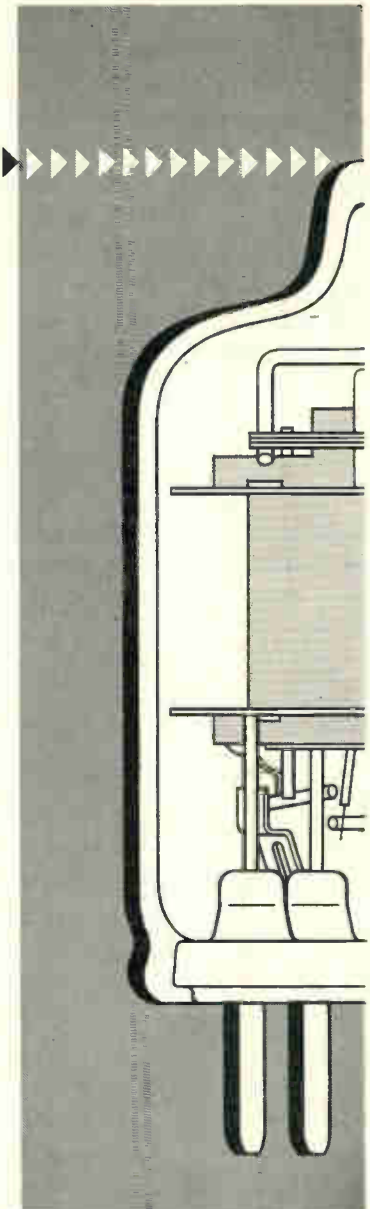
ACTUAL SIZE

EF95



A NEW R.F. PENTODE WITH

1. **Recommended Frequency Limit 400 Mc/s.**
2. **Mutual Conductance 5.1mA/V**
3. **Filament Consumption 175 mA.**

The Mullard R.F. Pentode, EF95, provides a better size to performance ratio than that previously obtainable from British valves of a similar class. It is constructed on the B7G miniature base and works efficiently at frequencies up to 400 Mc/s. Some of the more outstanding features of this valve include low input capacitance, low anode to grid capacitance, high mutual conductance and low heater consumption. These features, together with an operating voltage of 180 volts (120 volts under certain conditions), will particularly interest designers of compact communications equipment. In circuits involving a number of R.F. and I.F. stages, the EF95 may be used throughout, with a resultant marked saving in total heater consumption. Furthermore, the use of only one valve type in such applications enables maintenance problems to be reduced. The EF95 has similar electrical characteristics to the American 6AK5, and may be used as a direct replacement for it. Full technical information on this and other types in the Mullard range of communication valves is available on request.

TECHNICAL DATA

HEATER

V_h	-	6.3V
I_h	-	0.175 A

CAPACITANCES

c_{in}	-	4.0 µµF
c_{out}	-	2.8 µµF
$c_{a\text{-}g1}$	-	0.02 µµF

CHARACTERISTICS

V_a	-	180 V
V_{g2}	-	120 V
V_{g1}	-	−2.0 V
I_a	-	7.7 mA
I_{g2}	-	2.4 mA
g_m	-	5.1 mA/V
r_a	-	690 KΩ

LIMITING VALUES

V_a	max.	-	180 V
p_a	max.	-	1.7 W
V_{g2}	max.	-	140 V
p_{g2}	max.	-	0.5 W
I_k	max.	-	18 mA

BASE B7G

Mullard

MULLARD LTD., COMMUNICATIONS & INDUSTRIAL VALVE DEPT., CENTURY HOUSE, SHAFTESBURY AVENUE, LONDON, W.C.2

MVT133

WIRELESS ENGINEER

Vol. 30 JUNE 1953 No. 6

Double-Slug Transformers

THE use of double slugs either of metal or dielectric for tuning purposes in waveguides and coaxial lines has been known for some time. An interesting article on what the authors call "dielectric double-slug transformers" was published in the *Proceedings of the Institute of Radio Engineers* for December 1952.* The slugs act as an impedance transformer, their position in the line being adjusted until the field reflected from them cancels the field reflected from the mismatched load; their position then enables one to calculate the impedance of the load. The impedance of a load can thus be determined without making any electrical measurement, except the determination of the matched condition at the sending end of the line, which, as the authors say, is a relatively simple problem.

Since the method is based on the assumption that the length of each slug is equal to a quarter of the wavelength in the dielectric material of which the slug is made, the frequency is fixed. Previous analyses, to which the authors refer, have been based on this assumption, but one of the principal objects of the paper under consideration is to make the analysis more general and applicable to frequencies above and below this centre frequency. The results have been applied experimentally at frequencies from 0·5 to 1·5 times the centre frequency. The method has its limitations in that the load cannot be matched if it produces a voltage-standing-wave-ratio (v.s.w.r.) greater than ϵ^2, where ϵ is the dielectric constant of the slug, and this limiting value is reduced as the frequency departs from the centre value. As no probes are required, the slot need only be wide enough to enable the slugs to be moved; in a coaxial line the slugs act as supports for the centre conductor. In the opinion of the authors the principal disadvantage of the method is that the accurate measurement of the effective dielectric constant is essential. We shall confine ourselves to coaxial lines.

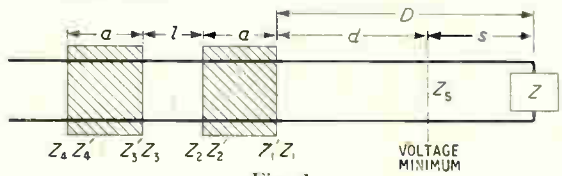


Fig. 1.

Fig. 1 shows diagrammatically a line with a terminating load of impedance Z and two dielectric slugs of length a separated by a distance l. As a result of reflection at the load there is a voltage minimum at a distance s; d is the distance from this minimum to the nearest slug and $D = d + s$. For the normalized impedance of the load (that is, in terms of the characteristic impedance) we have $Z = R + jX = \coth\theta = \coth(\rho + j\phi)$, and for that of the line from the point of voltage minimum to the load, $\coth k_0 s$ where $k_0 = \alpha_0 + j\beta_0$. Assuming the air-filled transmission line to be lossless, $\alpha_0 = 0$ and $k_0 = j\beta_0$; hence the combined impedance from the point of minimum voltage is $Z_s = \coth[\rho + j(\phi + \beta_0 s)]$.† At a point of minimum voltage $Z_s = 1/r$, where r is the v.s.w.r., and the phase angle $(\phi + \beta_0 s) = \pi/2$. Hence $Z_s = \coth(\rho + j\pi/2) = 1/\coth\rho = 1/r$. For the impedances at the first slug boundary we have

$$Z_1 = \coth\theta_1 = \coth[\rho + j(\pi/2 + \beta_0 d)].$$

* "The Determination of Impedance with a Double-Slug Transformer," by R. C. Ellenwood and E. H. Hurlburt p. 1690.

† We are here using the completely hyperbolic method as described in the last Editorial.

On passing into the dielectric, we have

$$Z_1' = \coth \theta_1' = \sqrt{\epsilon}\, Z_1 = \sqrt{\epsilon} \coth \theta_1$$

at the other end of the slug

$$Z_2' = \coth \theta_2' = \coth (\theta_1' + ka)$$

where $k = \alpha + j\beta$ is the propagation constant in the dielectric. Similarly

$$Z_2 = \coth \theta_2 = Z_2'/\sqrt{\epsilon} = (\coth \theta_2')/\sqrt{\epsilon}$$

$$Z_3 = \coth \theta_3 = \coth (\theta_2 + k_0 l)$$

$$Z_3' = \coth \theta_3' = \sqrt{\epsilon} Z_3 = \sqrt{\epsilon} \coth \theta_3$$

$$Z_4' = \coth \theta_4' = \coth (\theta_3' + ka)$$

$$Z_4 = \coth \theta_4 = Z_4'/\sqrt{\epsilon} = (\coth \theta_4')/\sqrt{\epsilon}$$

If l and d are adjusted until the input is matched, the normalized impedance $Z_4 = 1$.

By eliminating Z and θ from these equations one obtains the cumbersome equation

$$\frac{1 + r \coth k_0 d + \sqrt{\epsilon}\, r \coth ka + \sqrt{\epsilon} \coth ka \coth k_0 d}{\sqrt{\epsilon} r + \sqrt{\epsilon} \coth k_0 d + \coth ka + r \coth ka \coth k_0 d}$$
$$= \frac{\epsilon \coth ka \coth k_0 l - \sqrt{\epsilon} \coth k_0 l - \epsilon \coth ka + \epsilon\sqrt{\epsilon}}{\sqrt{\epsilon} \coth ka \coth k_0 l - \epsilon \coth k_0 l + 1 - \sqrt{\epsilon} \coth ka}$$

If it be assumed that both the line and the dielectric are lossless (i.e., that $\alpha_0 = 0$ and $\alpha = 0$) we can put $k_0 = j\beta_0$ and $k = j\beta$; then $\coth k_0 l = -j \cot \beta_0 l$, $\coth ka = -j \cot \beta a$ and $\coth k_0 d = -j \cot \beta_0 d$. By separating the real from the imaginary terms and eliminating $\cot \beta_0 d$ one obtains

$$r^2 - (b + 2)r + 1 = 0$$

in which

$$b = \{\tfrac{1}{2}(\epsilon - 1/\epsilon)(1 - \cos 2\beta a) \sin \beta_0 l$$
$$- \sqrt{\epsilon}(1 - 1/\epsilon) \sin 2\beta a \cos \beta_0 l\}^2.$$

The solution of this is

$$r = 1 + b/2 + \sqrt{b + (b/2)^2}.$$

This gives the v.s.w.r. in terms of ϵ, a, β_0, β and l; since $\beta_0 = 2\pi/\lambda_0$ and $\beta = 2\pi\sqrt{\epsilon}/\lambda_0$, these will depend on the frequency. To determine Z, the distance s from the load to the point of minimum voltage must also be known; this is obtained by measuring D, calculating d and putting $s = D - d$. The value of $\cot \beta_0 d$ which was eliminated from the equations to obtain r can also be calculated from them, and used to give the value of the distance d. Formulae are given in the article referred to, but we doubt the accuracy of some of them.

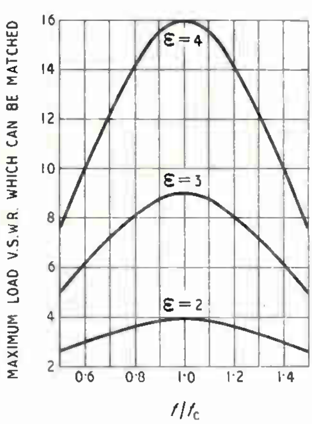


Fig. 2.

As we have already said, the device will only work if the v.s.w.r. does not exceed ϵ^2; if the frequency differs from the centre value, the limiting values of the v.s.w.r. are reduced as shown in Fig. 2. When using the device one begins with the slugs in contact and varies D to get the best result; one then separates the slugs by a small amount and readjusts D, and so on, step by step, until the best match is obtained. At the exact centre frequency the best result is obtained with $d = 0$; that is, with the side of the slug nearest the load at the point of minimum voltage.

The value of ϵ in all the formulae is the effective value of the dielectric constant of the slug as used, and not that of the material of which it is made; the difference is due to the fact that, as the slug must be capable of sliding in the line, it cannot be a tight fit between the inner and outer conductor, but must have an air film on either side.

It is possible to determine the effective value of ϵ from carefully made experiments in which the matched condition can be obtained with two different values of the separation l, but it involves the solution of a transcendental expression by successive approximation. The authors give some results obtained with a relatively crude apparatus and say that the accuracy was comparable with that of a precision slotted line with all its calibrated indicating system.

We mentioned at the beginning that the slugs can be of metal, in which case they only occupy a part of the cross-section as shown in Fig. 3.

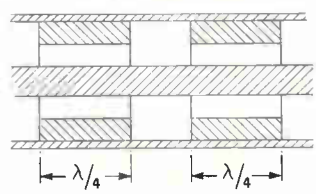


Fig. 3.

In the constricted part of the line L is decreased and C increased so that $Z_c = \sqrt{L/C}$ is decreased, as it is with the dielectric slugs in which C is increased but L unaffected. In this case, however, the wavelength is unaffected. We have made this reference to metal slugs because Dr. A. T. Starr in his recently published "Microwave and Short-Wave Techniques", after making a brief reference to dielectric slugs, says, "A good dielectric for this type of slug is mycalex (a sintered mixture of glass and mica); but one may just as well use metal." It will be interesting to see what the authors of the paper referred to, who are both at the American National Bureau of Standards and who have evidently devoted considerable time to research into the subject, have to say to this statement by Dr. Starr.

G. W. O. H.

HARMONIC DISTORTION AND NEGATIVE FEEDBACK

By R. O. Rowlands, M.Sc., A.M.I.E.E.

(B.B.C. Engineering Training Department)

SUMMARY.—The reduction due to the application of negative feedback in the harmonic distortion produced by an amplifier is examined by a more rigorous method than that usually employed. It is found that the distortion with feedback d_0' is given in terms of the distortion without feedback d_0 by $d_0' = d_0/(1 + \beta A_d)$ where A_d is not the linear gain of the amplifier but the slope of the curve in the region of the distortion. If a valve has a sharp cut-off, A_d will be quite small compared with A, resulting in no appreciable reduction of the distortion.

Introduction

IT is generally assumed that for a given output from an amplifier the harmonic distortion is reduced by negative feedback to $\frac{1}{1 + A\beta}$ of its original value where A is the amplification ratio without feedback and β is the fraction of the output fed back. This result is arrived at in the following manner. The harmonic distortion produced within an amplifier depends upon the magnitude of the signal voltage between grid and cathode and so upon the output level of the signal, since this is a direct multiple of the grid-to-cathode voltage irrespective of whether negative feedback has been employed or not. This harmonic distortion, designated d_0, appears in full in the output when no feedback is employed but with feedback it becomes d_0'

where $d_0' = d_0 - A\beta d_0'$

from which $d_0' = \dfrac{d_0}{1 + A\beta}$

This treatment of the problem is only approximately correct because it ignores the harmonics of higher order (viz., harmonics of the harmonics), which are produced as a result of feedback. This fact is admitted in some books* which say that the approximation is correct provided that the harmonic content is small. This is not always the correct criterion, however; e.g., if a valve has a sharp cut-off and the input voltage is just greater than the cut-off value no amount of negative feedback will reduce the distortion even though the distortion is only a small part of the signal.

A more rigorous treatment of the subject will now be undertaken in order to evaluate the extra distortion and to see under what conditions the approximate formula is sufficiently correct.

Theory

The simplest type of harmonic generation to deal with mathematically is that due to a parabolic amplifier characteristic and so the performance of this type of amplifier will be investigated first.

The instantaneous value of the output voltage V_0 may be expressed in terms of the instantaneous value of the input voltage V_1 by the equation,

$$V_0 = AV_1 + CV_1^2$$

where A and C are constant, depending upon the amplifier characteristics.

If a fraction β of the output is applied as negative feedback to the input,

then $V_0 = A(V_1 - \beta V_0) + C(V_1 - \beta V_0)^2$.

Multiplying throughout by β the equation becomes:

$$C\beta(V_1 - \beta V_0)^2 + A\beta(V_1 - \beta V_0) - \beta V_0 = 0$$

or $C\beta(V_1 - \beta V_0)^2 + (A\beta + 1)(V_1 - \beta V_0) - V_1 = 0$

and solving this quadratic in $V_1 - \beta V_0$ we get

$$V_1 - \beta V_0 = \{-(1 + A\beta) + \sqrt{(1 + A\beta)^2 + 4C\beta V_1}\}/2C\beta$$

the negative value of the root sign being ignored as $V_1 - \beta V_0$ is positive.

In order to be able to expand the expression under the root sign in terms of ascending powers of V_1,

$(1 + A\beta)^2$ must be greater than $4C\beta V_1$

Now $(1 - A\beta)^2 > 0$ for real values of A and β

i.e., $1 - 2A\beta + A^2\beta^2 > 0$

$\therefore\ 1 + 2A\beta + A^2\beta^2 > 4A\beta$

So $(1 + A\beta)^2 > 4A\beta > 4C\beta V_1$, if $A > CV_1$

The expansion is therefore possible provided that the harmonic content in the output without feedback does not exceed the fundamental.

With this condition

$$V_1 - \beta V_0 = \left\{-(1 + A\beta) + (1 + A\beta)\left[1 + \frac{2C\beta V_1}{(1 + A\beta)^2} - \frac{2C^2\beta^2V_1^2}{(1 + A\beta)^4} + \frac{4C^3\beta^3V_1^3}{(1 + A\beta)^6} + \ldots\right]\right\}\Big/2C\beta$$

$$= \frac{V_1}{1 + A\beta} - \frac{C\beta V_1^2}{(1 + A\beta)^3} + \frac{2C^2\beta^2V_1^3}{(1 + A\beta)^5}\ldots$$

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* See Bode, "Network Analysis and Feedback Amplifier Design," p. 79.

From which

$$V_0 = \frac{AV_1}{1 + A\beta} + \frac{CV_1^2}{(1 + A\beta)^3} - \frac{2C^2\beta V_1^3}{(1 + A\beta)^5} \cdots$$

If the input voltage is increased to $(1 + A\beta)V_1$ in order to give the same output of fundamental then:

$$V_0' = AV_1 + \frac{CV_1^2}{(1 + A\beta)} - \frac{2C^2\beta V_1^3}{(1 + A\beta)^2} \cdots$$

This shows that harmonics of higher order than the second are produced by negative feedback.

Their values may be calculated from the above equation using the relationships that if,

$$V_1 = v \sin \omega t$$

then $V_1^2 = v^2 \sin^2 \omega t = \frac{1}{2} v^2 (1 - \cos 2\omega t)$

and $V_1^3 = v^3 \sin^3 \omega t$

$= \frac{1}{4} v^3 (3 \sin \omega t - \sin 3\omega t)$ etc.

The solution in the general case could be obtained by writing an equation for the output voltage in terms of ascending powers of the input and proceeding in a similar manner but the mathematics would be very tedious and the final result would not be of much practical value.

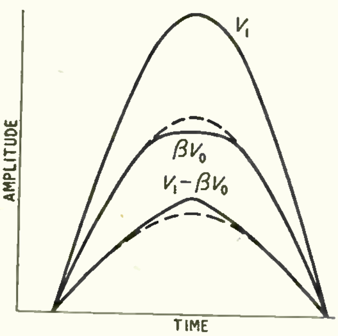

Fig. 1. *Predistortion of signal due to negative feedback. V_1 = input voltage, βV_0 = voltage fed back, $V_1 - \beta V_0$ = grid-cathode voltage. The broken-line curves are sinusoidal.*

General Solution

In order to obtain a general solution it is profitable to examine the physical process by which distortion is reduced by the application of negative feedback.

Let us suppose that the input wave is sinusoidal and the output flat topped. The result of subtracting a fraction of the output from the input will be to produce a peaky wave between grid and cathode. This is shown in Fig. 1. The application of negative feedback therefore has the effect of predistorting a signal in such a manner as to counteract the distortion due to the amplifier.

Now let us consider an amplifier which is non-linear. The graph of the instantaneous values of the output voltage against input voltage is shown in Fig. 2. The linear portion of the curve is given by:—

$$V_0 = AV_1$$

or $V_1 = \frac{1}{A} V_0$

and the full curve is given by

$$V_0 = AV_1 + F(V_1) \quad .. \quad (1)$$

or alternatively by $\quad V_1 = \frac{1}{A} V_0 + f(V_0) \quad .. \quad (2)$

Equation (1) gives the output voltage in terms of the input which is generally what we require to know in practice. $F(V_1)$ is a non-linear function of V_1 and gives the magnitude of the total harmonic distortion irrespective of frequency, which is present in the output. It is represented on the graph by the vertical deviation of the curve from linearity; e.g., by KJ.

Equation (2) is a theoretical equation which tells us by how much the input voltage must be distorted in order to produce an undistorted output. The required distortion in the input is represented by the non-linear term $f(V_0)$ and this is represented on the graph by the horizontal deviation of the curve from linearity; e.g., by HJ.

In order to construct a graph similar to Fig. 2 for the same amplifier with feedback, equation (2) will be used because the calculation is much simpler. With the same output voltage V_0 the grid–cathode voltage becomes $V_1 - \beta V_0$. Hence

$$V_1 - \beta V_0 = \frac{1}{A} V_0 + f(V_0)$$

giving $V_1 = \frac{1 + A\beta}{A} V_0 + f(V_0)$.

The significant fact emerging from this is that the amount of distortion $f(V_0)$ necessary in the input to produce an undistorted output does not depend upon the feedback. It is also seen that the V_0/V_1 curve corresponding to this equation has a linear portion similar to the original curve but with its slope decreased by the factor $1 + A\beta$. The actual curve is displaced from linearity in a horizontal direction by the same amount as the original curve; e.g., the points P and H correspond to the same value of V_0 and so PQ = HJ.

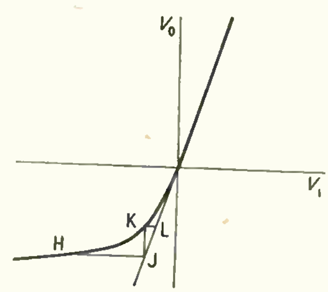

Fig. 2. *Graph of instantaneous values of output voltage against input for an amplifier.*

Now the question we are endeavouring to answer is by how much is the distortion which appears in the output reduced by the application of negative feedback. This we have seen is the vertical deviation of the curve from linearity.

It is represented by KJ in Fig. 2 which is the curve without feedback and by RQ in Fig. 3 which is the curve for the same amplifier with feedback. We shall therefore proceed to calculate the ratio of RQ to KJ from the geometry of the two figures.

In the curve of Fig. 2, HJ is the amount by which the input must be distorted to produce an undistorted output and is designated d_1, KJ is the distortion appearing in the output when the input is pure and is designated d_0. $\frac{\text{KJ}}{\text{KL}}$ is the slope of the linear portion of the curve and is equal to the amplification A.

So $$\text{KL} = \frac{d_0}{A}$$

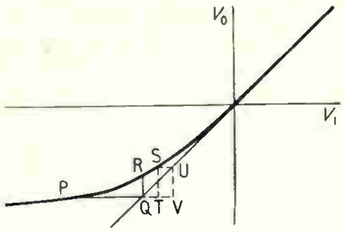

Fig. 3. *The same as Fig. 2 but with negative feedback.*

In Fig. 3 we have already seen that

$$\text{PQ} = \text{HJ} = d_1$$

Let S be a point on the curve corresponding to the same value V_0 as K in Fig. 2,

then $\text{UV} = \text{ST} = \text{KJ} = d_0$

And $\text{RQ} = \text{distortion with feedback} = d_0{}'$

Then $$\frac{\text{UV}}{\text{QV}} = \text{slope of curve} = \frac{A}{1 + A\beta}$$

So $$\text{QV} = \frac{(1 + A\beta)\, d_0}{A}$$

But $$\text{TV} = \text{SU} = \text{KL} = \frac{d_0}{A}$$

$$\therefore \quad \text{QT} = \text{QV} - \text{TV} = \frac{(1 + A\beta)}{A}\, d_0 - \frac{d_0}{A}$$
$$= \beta d_0$$

And $\text{PT} = \text{PQ} + \text{QT} = d_1 + \beta d_0$

The points P, R and S lie approximately on a straight line and so

$$\frac{\text{RQ}}{\text{PQ}} = \frac{\text{ST}}{\text{PT}}$$

i.e., $$d_0{}' = \frac{d_1 d_0}{d_1 + \beta d_0} = \frac{d_0}{1 + \beta \frac{d_0}{d_1}} = \frac{d_0}{1 + \beta A_d} \qquad (3)$$

where A_d is the slope of the amplifier curve without feedback in the region of the distortion, and is given by the ratio of the vertical to horizontal deviation of the curve from linearity.

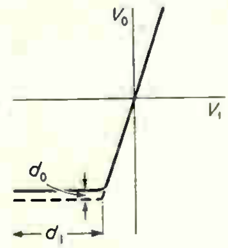

Fig. 4. *Amplifier with cut-off.*

This formula enables the effect of negative feedback to be calculated when the original distortion is not small and also where the distortion d_0 is small but the curvature of the valve is sharp. As an example, in the case of an amplifier with a cut-off, as shown in Fig. 4, if the input voltage slightly exceeds the critical value the distortion d_0 appearing in the output will be small but d_1 will be infinite and so from equation (3) it is seen that $d_0{}'$ will always be equal to d_0 whatever the value of β.

If *both* d_1 and d_0 are small the slope of the curve is approximately equal to that of the linear portion and

$$d_0{}' = \frac{d_0}{1 + \beta A}$$

Conclusions

It has been shown that in order that the usual formula for the reduction of harmonics by the application of negative feedback may be accurate, a necessary and sufficient condition is that the slope of the amplifier curve does not deviate appreciably from linearity over the working range.

A graphical method of calculating the total distortion in a general case has been given and a mathematical method of solving a particular case where the output may be expressed in terms of the input in an equation of low degree has been indicated.

Acknowledgments

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WIENER'S THEORY OF LINEAR FILTERING*

By D. A. Bell, M.A., B.Sc., M.I.E.E.

(*University of Birmingham*)

1. Introduction

THIS note is written in the belief that the significance of Wiener's theory of filtering (or 'smoothing') is hidden from many engineers by the combination of the complexity of the mathematics in Wiener's writing on the subject[1] and the novelty of the general idea. Bode and Shannon have made a very valuable contribution,[2] but by deliberately side-stepping Wiener's mathematical processes they have had to accept some limitations on the type of signal and noise which can be handled. An alternative approach is that the ability to handle Wiener's mathematics is less important—the solution of an equation can always be left to mathematicians—but that it is useful for the engineer to know the *purpose* of the mathematical operations involved. He can then come to terms with the mathematician on such questions as the type of function and degree of approximation which is permissible in finding an analytically-convenient expression for his power spectra, etc. On the other hand, the present discussion has been simplified by restricting it to two limiting cases of 'zero lag' and 'infinite lag' thus excluding the 'predicting' filters, which are prominently featured in the literature of the subject but are of very specialized practical application.

The novelty in this type of filter theory is the idea that there is an optimum linear filter, which can be quantitatively specified, for use when signal and noise occupy the same frequency band. (By 'linear filter' is meant a device which may have any desired frequency characteristic but is linear in the sense that its characteristic is independent of the amplitude of the signal.) Earlier experience of practical devices in radio communications had suggested that although one could use a linear filter to separate signal from interference on a neighbouring frequency, only non-linear devices, such as an amplitude-limiter for impulsive interference, 'rectifier discrimination' (otherwise known as 'the apparent demodulation of a weak signal by a strong one') or an f.m. system incorporating a limiter, would be effective when signal and interference occupy the same frequency band. It is true that the latter case is treated by a linear device in the gramophone scratch-filter, but the final judgment in that case is subjective.

Another example, however, may serve to show that there is sense in asking for a filter to improve the overall signal/noise when signal and noise are in the same frequency band. Suppose that a graphical representation of some function is being fed into a differential analyser or other calculating machine by the computor himself turning a handle in order to keep the input index of the machine in coincidence with the curve; then it is known that there is a very strong tendency for the human operator to produce small irregularities instead of turning the handle perfectly smoothly. Such irregularities are errors, or 'noise' in communication terminology; and the obvious step is to recognize that the system cannot receive all input frequencies in the presence of such noise and to insert a filter or 'smoothing circuit' between the shaft controlled by the operator and the input of the calculating machine. There would be little hesitation among mathematicians in saying that the obvious criterion to try first is to minimize the sum of the squares of the errors at all points, and the other important factors to know would be:

(*a*) The frequency-distribution of the noise.

(*b*) The average frequency distribution of the various types of curves which the system might have to follow from time to time.

(*c*) The way in which the total signal will be modified by the calculating machine: e.g., the effect of input irregularities is minimized if the system is organized to work in integrations rather than differentiations.

In the classical (Wiener) treatment of the filter problem, the criterion adopted is to minimize the mean-square error, but it must be remembered that this is not necessarily appropriate to all applications.

The limiting cases of infinite-lag and zero-lag have been chosen because they apply respectively to telecommunication and automatic-control systems of most kinds. For telecommunication work, the infinite-lag filter is usually an acceptable approximation, because a delay in the message equal to many times the period of any component frequency of the message waveform is still likely to be short on the time-scale of human interpretation and subsequent human action. The infinite-lag filter is then very easily computed according

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*This paper is based on Chapter 7 of the author's book, "Information Theory and its Engineering Applications," Pitman, London. 1953.

to the technique of Bode and Shannon,[2] provided there is no coherence between the different component frequencies in the Fourier analysis of the signal and of the noise. This will always be true of white noise, but will be only partially true of some kinds of signal.

In computing the contribution of error associated with a particular frequency ω, it must be remembered that since phase is ignored in the mean-square error criterion we may only work with power, intensity or squared amplitude and not with simple amplitude. The error is therefore taken as two uncorrelated parts, the noise as modified by the filter, and the difference between the message as modified by the filter and the original message, and the resultant squared error is obtained by summing the squares of these two parts:

$$E^2(\omega) = Y^2(\omega) . N(\omega) + [Y(\omega) - 1]^2 S(\omega) \quad \ldots \quad (1)$$

where $Y(\omega)$ is the amplitude attenuation factor of the filter at frequency ω, $N(\omega)$ is the noise power (squared amplitude) and $S(\omega)$ the message power at frequency ω. The mean-square error for the whole signal (all frequencies) is therefore

$$\overline{E^2} = \int_0^\infty \{Y^2(\omega) . N(\omega) + [Y(\omega) - 1]^2 S(\omega)\} d\omega \quad \ldots \quad (2)$$

Now the technique of minimizing a sum of squares of this type is employed by statisticians in constructing a 'regression line'; i.e., the straight line which fits a set of points so as to make the sum of the squares of the differences* between points and line a minimum.

The technique is to arrange (1) so that $Y(\omega)$ appears only in a squared term, which necessarily makes a positive contribution to the error, and the error is then least when this squared term vanishes. In our case the squared term is

$$\{[N(\omega) + S(\omega)]^{\frac{1}{2}} Y(\omega) - S(\omega) [N(\omega) + (S\omega)]^{-\frac{1}{2}}\}^2$$

so that the desired value for $Y(\omega)$ is

$$Y(\omega) = S(\omega)/[S(\omega) + N(\omega)] \quad \ldots \quad (3)$$

In formulating the problem in this way, it has been implied that the filter can be constructed with a specified amplitude-attenuation factor which is a simple number; i.e., the filter has no phase-shift. This is in fact impossible, because Bode's phase-integral theorem[3] states that the *minimum* difference in phase-shift of a network between two frequencies is a specified function of the difference in its attenuation between the two frequencies. But by connecting in series with the filter proper some suitable phase-equalizing networks (otherwise known as 'all-pass' networks, which are not minimum-phase networks) it is possible to arrange that the combination has the desired variation of attenuation with frequency and a phase characteristic which is equivalent to a constant delay. The complete transfer characteristic is then represented by

$$Y'(\omega) = \frac{S(\omega)}{S(\omega) + N(\omega)} e^{-j\beta\omega} \quad \ldots \quad (4)$$

and can always be constructed with desired amplitude function $|Y(\omega)|$ if we are prepared to make the delay time β long enough. This filter is called the 'infinite-lag' filter which is optimum for the given problem.

It is also essential to the method of computing $Y(\omega)$ that the various frequency components shall be independent, so that the total error is least when the errors at all frequencies are individually least. This condition applies to random noise but not to impulse interference or to many types of signal. Both for the infinite-lag filter with the most general type of signal and noise and for the zero-lag or specified-lag filter, it is necessary to use Wiener's method of analysis, working with time functions instead of with sinusoidal component frequencies. In automatic-control systems, in particular, delay is unacceptable and the filter must be designed for zero-lag. Leading or 'predicting' filters appear prominently in the literature of the subject, and strangely enough the leading filter is apparently more amenable to analysis than the lagging filter, but for most practical applications where a shift of time scale is significant the requirement is a zero-lag filter.

It should, perhaps, be mentioned also that in the application to automatic-control systems the Wiener filter is not an additional component to be added within the closed loop, as one might add a phase-shifting network. The requirement is that the frequency response of the complete system (i.e., the transfer function which relates output to input in working conditions) should comply with the specification for the optimum filter. This specification is partly dependent on the input signal/noise ratio, and if in fact the noise arises at some point in the servo system other than the signal-input point, it must be transformed into an equivalent noise at the input terminals. For example, in a remotely-controlled radar-aerial system the chief source of interference with the correct orientation of the aerial might be varying wind pressure on the aerial itself. In this case the 'noise' is introduced at the output member of the system and must be suitably transformed (according to the characteristics of the feedback path) into an equivalent input noise which can be compared with the input signal.

* The differences are measured parallel to one or the other of the two axes, *not* as the shortest distances between point and line.

2. Simple Example of Zero-Lag Filter

The mathematical complexity of Wiener's method of analysis places it beyond the scope of this paper, but something may be learned by examining the form of the result in a particular application.

The simple example which has been analysed by Wiener (loc.cit.) is of an input message having the power spectrum $1/(1 + \omega^2)$ with uniform random noise of which the power spectrum is merely a constant, ϵ^2. This can be translated into the more familiar terms of a signal/noise power ratio as follows:

Signal power at angular frequency ω

$S_\omega = 1/(1 + \omega^2)$

Noise power at angular frequency ω

$N_\omega = \epsilon^2$

Signal/noise ratio at angular frequency ω

$$(S/N)_\omega = 1/\epsilon^2 (1 + \omega^2)$$

$$\therefore \epsilon^2 = (N/S)_0 \qquad (5)$$

This will give the same form as was assumed above provided that we use the appropriate scale factors for ω and for ϵ^2, namely $\omega_0 = 2\mu$ for the 'scale frequency' and $2a^2/\mu$ for the unit of noise power per unit bandwidth. (The parameter μ enters into the latter, as well as a^2, because ϵ^2 as defined above is not a noise *power* but a noise *power-per-unit-bandwidth*).

Wiener[1] gives two formulae for the filter specification to meet this particular example of signal and noise spectra, one for a predicting filter [his equation (3.605) on p. 92] and another for an approximation to the ideal lagging filter [his equation (3·75) on p. 93]; but in the limit for zero lead or lag both reduce to

$$k(\omega) = \frac{1}{[\epsilon + N(1 + \epsilon^2)]\,[\sqrt{(1 + \epsilon^2)} + \epsilon j\omega]} \qquad (8)$$

where $k(\omega)$ is the transfer characteristic of the optimum filter (and $j = \sqrt{-1}$). By making use of equation (5) this can be expressed in the form:

$$k(\omega) = \frac{1}{\left[\sqrt{(N/S)} + \sqrt{\left(\frac{N+S}{S}\right)}\right]\left[\sqrt{\left(\frac{N+S}{S}\right)} + j\omega\sqrt{\frac{N}{S}}\right]} \qquad (9)$$

where $(N/S)_0$ is the value of noise/signal ratio at zero frequency. It is also obvious that ω may be regarded as a normalized frequency, equivalent to ω/ω_0, and the answer will be independent of the frequency scale-factor ω_0 provided the same scale (e.g., in radians/sec, c/s, kc/s, or Mc/s) is used for signal, noise and filter specification.

It will be convenient to divide both factors of the denominator by $\sqrt{(N/S)}$ so as to put (8) in the form $k(\omega) = C/(B + j\omega)$ where

$$\left.B = \sqrt{\left(\frac{N+S}{S}\right)} \text{ and } C = \frac{S/N}{1 + \sqrt{\left(\frac{N+S}{N}\right)}}\right\} \qquad (10)$$

Before proceeding with the filter calculation, it is worth noting that this signal spectrum corresponds to the so-called 'random telegraph signal,' which consists of a random sequence of positive and negative pulses which are all of the same amplitude but which have lengths distributed exponentially. Rice[4] has shown that if a is the amplitude of pulse and μ the average number of reversals per second (i.e., $1/\mu$ is the average length of pulse), the autocorrelation function is

$$\psi(\tau) = a^2 e^{-2\mu\tau}$$

and the power-spectrum in terms of frequency is

$$w(f) = \frac{2a^2\mu}{\pi^2 f^2 + \mu^2} \qquad (6)$$

Since $\omega = 2\pi f$, this can be re-arranged in the form

$$w(\omega) = a^2 \frac{8\mu}{\omega^2 + 4\mu^2}$$

$$= \frac{2a^2}{\mu}\,\frac{1}{1 + \omega^2/4\mu^2} \qquad (7)$$

It is to be understood that S and N refer to the signal and noise powers at zero frequency, and the quantities in (10) are now familiar: S/N is simply the signal/noise ratio, and $(N + S)/N$ is the quantity which enters into Shannon's expression for the communication rate of a channel, $W \log [(N + S)/N]$.

The output power, however, will be proportional to $|k(\omega)|^2$, and this is the quantity which is directly comparable with the input signal power. From (10),

$$|k(\omega)|^2 = \frac{(S/N)^2}{\left[1 + \sqrt{\left(\frac{N+S}{N}\right)}\right]^2}\,\frac{1}{\frac{N+S}{N} + \omega^2} \qquad (11)$$

As a check on this, examine the two limiting cases. If $S/N \to \infty$, $|k(\omega)|^2 \to 1$; i.e., if there is no noise there is no need for a filter. Alternatively, if $S/N \to 0$, $|k(\omega)|^2 \to 0$; i.e., if there is no signal the best thing is to disconnect the circuit entirely so as at least to keep out the noise. For a very small but not vanishing S/N we may approximate $(N + S)/N$ to unity, and then

$$|k(\omega)|^2 \approx \frac{(S/N)^2}{4\,(1 + \omega^2)} \quad \cdots \quad (12)$$

Before discussing particular values of S/N, the infinite-lag filter may be introduced with the characteristic[2]

$$|Y(\omega)| = \frac{S(\omega)}{S(\omega) + N(\omega)} = \frac{S/N}{S/N + 1 + \omega^2} \quad (13)$$

since in our example $S(\omega) = S/(1 + \omega^2)$, $N(\omega) = N$. The squared value is then

$$|Y(\omega)|^2 = \frac{(S/N)^2}{[S/N + 1 + \omega^2]^2} \quad \cdots \quad (14)$$

It may be noted that $k(\omega)$ is a complex function, so that it specifies phase as well as amplitude, but $|Y(\omega)|$ is a function of signal and noise powers only, so that it merely specifies an amplitude without a phase; and this difference appears to have the following physical significance. In the zero-lag filter any phase-change represents an error, but for a physically-realizable network there will be a certain minimum phase change for any specified amplitude characteristic (given by Bode's phase-integral theorem[3]); though in the infinite-lag filter, phase change may not represent a deterioration of performance, for provided that the phase change varies linearly with frequency it can be interpreted as a delay which is the same for all signals; i.e., part of the infinite lag. The minimum-phase network having the desired amplitude characteristic is unlikely to have precisely this phase characteristic, but at the cost of increasing the delay (already supposed infinite) one can always *add* phase-equalizing networks (which are non-minimum-phase) in order to get the desired phase characteristic without changing the amplitude characteristic.

low frequencies, compared with only 0·47 for the zero-lag filter. It will also be noted that the infinite-lag filter cuts off much more rapidly than the zero-lag filter, which is consistent with the point made above that the amplitude characteristic of the zero-lag filter is partly governed by consideration of phase characteristic, while that of the infinite-lag filter is not. Both approximate to the shape of the input spectrum when $S/N \to 0$ [equations (3) and (12)].

In considering the low value of transmission calculated for the zero-lag filter one must remember that the squared error is made up of the sum of two components: (*a*) the part of the noise which *is* transmitted, and (*b*) the part of the signal which *is not* transmitted. If we disregard any possible internal correlation between different frequency components of the signal (the assumption used in the simplified method of Bode and Shannon, loc. cit.) these two components may be represented by:

$$\text{Noise admitted} = \int_0^\infty |k(\omega)|^2\, N(\omega)\, d\omega \quad \cdots \quad (15)$$

$$\text{Signal excluded} \approx \int_0^\infty (1 - |k(\omega)|)^2\, S(\omega)\, d\omega \quad (16)$$

In (16) it is assumed that $k(\omega)$ has no phase angle, and this is therefore an approximation representing the lower limit of the error—an approximation which will be fairly good over the lower part of the frequency range. For the zero-lag

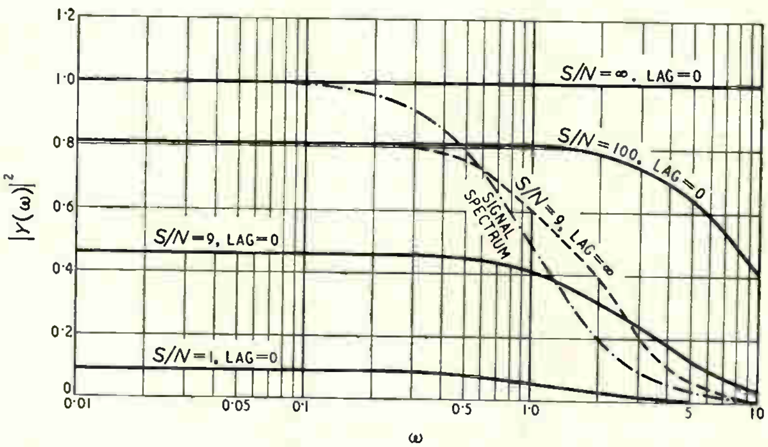

Fig. 1. *Representation for various signal/noise ratios of the squared transfer function of zero-lag filters. The dotted curve is for an infinite-lag filter.*

Fig. 1 is a graphical representation with logarithmic frequency scale and linear power scale of the signal spectrum $1/(1 + \omega^2)$ and of the squared transfer function of the zero-lag filters for the particular cases of $S/N = 1, 9, 100$ and ∞. The dotted curve shows the infinite-lag filter for $S/N = 9$. The overwhelming advantage of the infinite-lag filter (for applications in which delay can be tolerated) is shown by the fact that for $S/N = 9$ it transmits 0·81 of the input power at

filter of Fig. 1 with $S/N = 9$, the zero-frequency values of noise admitted and of signal rejected are of roughly the same order of magnitude, though certainly not equal. If, for the sake of argument, it was decided to make 'signal lost' equal to 'noise admitted' at zero frequency, we should have to increase $k(\omega)$ from 0·68 to 0·75 at $\omega = 0$ but it would still be well below the value of $Y(\omega) = 0{\cdot}9$ for the infinite-lag filter.

Fig. 2 shows a plot, *for our particular example,* of the relation between the signal transmission of the optimum zero-lag filter at zero frequency (expressed as an amplitude) and signal/noise ratio. Obviously, the error must always be greater than the proportion of signal excluded; and we therefore deduce that if a position-control system, for example, is to have an error of 1% when operating with our specfic forms of signal and noise spectra, the signal/noise ratio must be considerably better than 50:1 in amplitude; i.e., considerably better than 2,500:1 in power. (For large values of S/N, the formula for the curve of Fig. 2 is $[k(\omega)]_0 \approx 1 - \frac{1}{2} N/S$.)

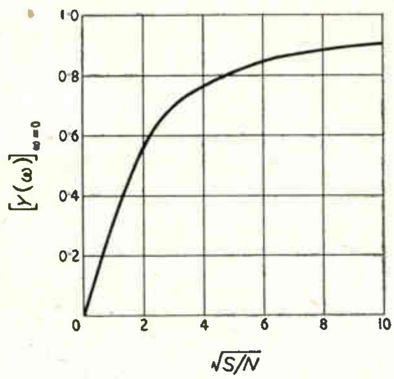

Fig. 2. *Relation between the signal transmission of an optimum zero-lag filter and signal/noise ratio.*

3. Effect of Departures from Optimum Filter Characteristic

It may be argued that it is frequently impracticable to conform exactly to the calculated optimum filter characteristic. Is it then worth while using it as the basis of design? For example, it will be said that with very high signal/noise ratio the calculated filter has an absurdly wide band. The answer to this is that if the band were limited without phase-shift in the pass-band, the resulting error would correspond to the part of the message spectrum so cut off; and it is therefore a sufficient approximation to the optimum filter to pass without distortion all frequencies for which the message power is significant. Since phase distortion is significant, the filter pass-band (in amplitude) will have to extend rather beyond the effective cut-off point of the signal.

Similarly, when the signal/noise ratio is so limited that the optimum filter attenuates parts of the message spectrum which are still of significant intensity, an upper limit to the effect of departures from the optimum characteristic can be estimated as follows:—

(1) If the actual transmission is greater than the optimum value over any given range of frequencies, the error will be *less* than that corresponding to the additional noise included. (Because some additional message will be transmitted.)

(2) If the transmission is less than optimum, the error will be less than that corresponding to the amount of signal excluded. (Because some noise will also be excluded.)

(3) The effect of a phase-error ψ can be taken as equivalent to the exclusion of a part $(1 - \cos \psi)^2$ of the message power.

On the whole, therefore, it seems that the performance will not be too sensitive to small deviations in the cut-off characteristic, and if the Wiener filter can be computed for a given application, it can reasonably be taken as the target for a practical approximation.

4. Applicability of the Wiener Filter

It remains to assess the value of this 'optimum' filter, and to consider what limitations on its application are inherent in its derivation. In the first place it is a least-squares solution, a characteristic which can be considered in two ways. In terms of frequency response (analysis of the output by normal Fourier methods) it means that *phase* is entirely disregarded in assessing the error—a characteristic which is largely possessed by the human ear. In terms of time-function response (e.g., using such concepts as 'rate-of-rise' and 'overshoot' in the response to a step function) it means that the system tends to avoid any very large errors but is rather insensitive to small errors. This least-squares criterion may or may not be appropriate to any individual case, but it is analytically attractive because it uses or is similar to well-known statistical techniques. It is relevant to communication systems *if ideal coding is used*, for it will be remembered that the discrimination between signal and noise is then in terms of squared amplitudes.

In some automatic process-control systems the critical factor is the maximum single error which persists for a significant time (compare the maximum-demand rating of electricity-supply services), because an error such as excessive temperature rise may cause irreparable damage to the product and perhaps to the plant as well. A similar argument would apply to the automatic guiding of an aircraft or other vehicle, where a momentary deviation beyond certain limits could cause a crash. In such cases no measure of mean or mean-square error is sufficient, and other mathematical techniques will have to be developed.*

In the Bode and Shannon treatment it had to be assumed that different frequency components

* For example, Westcott has suggested the use of a time-weighted-error criterion, but has also pointed out that it involves considerable mathematical difficulties. ("Proceedings of Symposium on Information Theory," Ministry of Supply, London 1950, p. 133.)

could be treated separately, and this required an absence of correlation between the component frequencies of the signal (and of the noise) as well as an absence of cross-correlation between signal and noise. In communication systems this requirement would be satisfied by ideal coding, but not usually by uncoded or imperfectly coded signals. Examples of the latter are that speech and music contain groups of harmonically related frequencies, teleprinter signals consist largely of harmonics of the fundamental pulse frequency (e.g., 25 c/s), and television signals contain harmonics of line and picture frequencies. But Wiener's more rigorous treatment does not impose this restriction, nor does it necessarily require the condition (which, however, will usually be satisfied in practice) that there should be no correlation between message and noise. Wiener's work has been centred on those cases in which both message and noise (and received signal, i.e., message plus noise) have auto-correlation functions and the latter have Fourier transforms. This double condition requires that message and noise should be free from any precisely periodic components which would have to be represented by discrete lines in the power spectrum. (The power spectrum is usually obtained as the Fourier transform of the auto-correlation function.) But Wiener points out that this is not a serious objection in principle because:

(*a*) An exactly periodic wave conveys no information—if it is interrupted or otherwise modulated in amplitude it ceases to be a discrete line in the spectrum.

(*b*) There are possible mathematical techniques for extracting any simply periodic components from the signal and handling their autocorrelation functions separately, so that their contributions may be combined with the others at an appropriate stage.

Wiener's method of solution is therefore very generally applicable to any type of message and any type of noise.

It should also be noted that Wiener's analysis relates to *linear* filters, which in practice means frequency-selective networks or 'electric-wave filters.' Wiener's optimum linear filter is better than any possible non-linear filter if the task to which it is applied satisfies two conditions:

(*a*) the mean-square error criterion is used, and

(*b*) the noise is random noise.

Non-linear filters have found practical application in cases where neither condition is met (e.g., 'noise-limiters' on the vision side of television receivers) and should be considered whenever the noise is not random; i.e., whenever the 'entropy power' of the noise is less than its physical power. If we know nothing about the noise, a Wiener filter is best; but if we know of some specific characteristic of the noise, some other form of filter may be preferable. In the case of impulse interference, for example, our knowledge of the noise characteristics can be expressed in several different forms, each of which suggests that the Wiener filter would not be applicable. We may merely say formally that its entropy power is low; or we may say that although it has the same power spectrum as random noise it differs from the latter in having specified phase relationships between the different frequency components, and therefore is not to be handled by a mean-square process which necessarily destroys all evidence of phase relationships: or we may say that its specific characteristic is a large amplitude which lasts for a short time, and therefore the filter should be designed on the amplitude-time plane rather than the amplitude-frequency plane.

Similarly, if something else is known about the signal, as well as its power spectrum, the Wiener technique may fail through not taking account of this other information. For example, if special weight is attached to the Heaviside step function as a possible message, it is possible to construct a non-linear filter which will improve the response of the system even in the absence of noise.[5]

Another point concerns the testing of Wiener filters. Such filters are explicitly designed to give the best performance for a certain ensemble of messages, and for any *single message* out of the ensemble it is always possible to design a better filter than the Wiener filter. To push this argument to its extreme, suppose that we are required to design a minimum-mean-square-error filter for a class (or ensemble) of signals in which all frequencies between 300 and 3,000 c/s are present, with a certain frequency-weighting. This might be relevant to a telephone channel, and the filter would obviously have a frequency characteristic roughly of the type of a band-pass filter in the region 300-3,000 c/s. Now suppose a technician decides to test this by setting up a variable-frequency oscillator to various selected frequencies in the band: it is obvious that for any single frequency he could do far better with a very narrow filter tuned to that frequency than with the Wiener filter. Now if any single message (of however great a frequency band) is completely specified, the phase relationship between its frequency components must be known and therefore it is in principle possible to use this knowledge to construct a filter which is better than the Wiener filter based on the omission of all phase information. Such a filter would, of course, be inferior on other messages out of the ensemble.

A closely related problem is that of 'coherent' versus 'incoherent' detection of messages which are transmitted as amplitude modulation of a carrier wave, a problem which has been discussed in detail in relation to radar systems.[6]

In the simpler types of receiver the output, obtained from an incoherent detector, is proportional only to the amplitude of the carrier wave. However, the regularity of phase of the carrier wave is one of the physical characteristics which distinguishes message from noise, and it therefore seems likely that some advantage could be obtained by using a coherent detector which takes note of the carrier phase of the received signal. Typical of such detectors are the homodyne and synchrodyne which work on the principle that a valve is made alternately conducting and non-conducting by a local oscillator which is synchronized with the carrier frequency of the received signal. When combined with an integrator (which averages the detector output over many cycles of carrier frequency) such a detector gives strong discrimination between the message associated with the desired carrier and any interference which does not have the same phase rate.

APPENDIX

The Significance of 'Factorization' in Wiener's Analytical Solution.

The main mathematical requirements for the solution by Wiener's method of analysis are (a) that the signal, noise and signal-plus-noise should all have power spectra which can be derived as Fourier transforms of the corresponding correlation functions; and (b) that it should be possible to 'factorize' the analytical expression for the signal-power spectrum in a certain way. The purpose of this mathematical device of factorization is to ensure that the characteristic deduced for the optimum filter shall be 'physically realizable,' a condition which we earlier associated with the existence of a minimum phase change associated with a given amplitude-attenuation characteristic. But a more direct criterion of physical realizability of a stable network is 'no output before the input has commenced,' and this is the one used by Wiener since he sets up his error criterion in terms of time functions (not frequency characteristics). But in the same way that the frequency analysis of a time function in pure sine waves can be obtained from a Fourier transform using real frequencies ω, a time-function representing either a signal or a network response can be translated into a Fourier transform using complex frequency $\omega' = \omega + j\alpha$ where α is interpreted as a damping coefficient. (Some confusion may arise here because the Fourier transform uses the multiplier $e^{j\omega' t}$ while the Laplace transform uses e^{pt} where p is a complex quantity $x + jy$. Then the real axis in the Laplace transform corresponds to the imaginary axis in the complex Fourier transform.) Then a physically realizable stable system is one in which the damping of all modes of oscillation is positive, and the transform of the time-function has no poles in the lower half of the transform-plane corresponding to negative damping. Then the time function vanishes for $t < 0$. The formula for the optimum filter is therefore rearranged so that it includes a factor derived from the signal which vanishes for $t < 0$, thus ensuring that the filter response vanishes for $t < 0$. This is the basis of the procedure of 'factorization.'

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CORRECTION

The data given in the caption to the Furzehill sensitive valve voltmeter in the report of the Physical Society's Exhibition (p. 126, May issue) is incorrect for the model illustrated and refers to an older instrument. The V.200 actually covers 1 mV to 1,000 V and has a frequency range of 10 c/s to 6 Mc/s. The meter scale is still logarithmic.

We are informed that the oscilloscope 1684 D/2, referred to on p. 127, has been superseded by the model 1684 D/2* and that this model has an amplifier response extending to 4 Mc/s.

ELECTRICAL OSCILLATIONS

A Physical Approach to the Phenomena

A. W. Gillies, B.Sc., Ph.D.,

(*Northampton Polytechnic*)

1. Introduction

THE laws governing the behaviour of any physical system are normally expressed in the form of mathematical equations, often differential equations. This is indeed necessary since the physicist or engineer is concerned with the quantitative relations between the variables specifying the state of the system at any instant. The behaviour of the system is then determined by the solution of the mathematical equations.

The majority of physicists and engineers find such a process of formulating the mathematical equations of the system and deriving the solution with prescribed initial conditions from the mathematical processes of solving the equations intellectually unsatisfying unless they can give alongside of the mathematical solution, a physical interpretation of the various relations in terms of, or by analogy with, simpler phenomena with which they are already sufficiently familiar to accept as obvious.

The phenomena connected with electrical oscillations are no exception to the above remarks. The simplest case of an electrical oscillator is governed by an equation of the form

$$\frac{d^2v}{dt^2} + \psi(v)\frac{dv}{dt} + \omega_0{}^2 v = 0 \quad \ldots \quad \ldots \quad (1)$$

Thus for the circuit shown in Fig. 1 we have

$$i = f(v) = -\frac{v}{R} - C\frac{dv}{dt} - \frac{1}{L}\int v dt \quad \ldots \quad (2)$$

giving

$$\frac{d^2v}{dt^2} + \frac{1}{C}\left\{\frac{1}{R} + f'(v)\right\}\frac{dv}{dt} + \frac{1}{LC}v = 0$$

which is of the above form with $\omega_0{}^2 = 1/(LC)$ and

$$\psi(v) = \frac{1}{C}\left\{\frac{1}{R} + f'(v)\right\}$$

If $\psi(v)$ is identically zero the equation has the solution

$$v = V\cos(\omega_0 t + \alpha)$$

a simple periodic function with amplitude and phase determined by the initial conditions.

If $\psi(v)$ is constant and equal to $2a$, say, then the equation has the solution

$$v = Ve^{-at}\cos(\omega t + \alpha)$$

MS accepted by the Editor, October 1952

where $\omega = \sqrt{(\omega_0{}^2 - a^2)}$ and V and α are again determined by the initial conditions. This is again a simple periodic function whose amplitude now decreases or increases exponentially according as a is positive or negative. The frequency will differ from ω_0 only by a second-order small quantity if a is small. A decreasing amplitude will be obtained if NLR is replaced by an ordinary resistance. An increasing amplitude will be obtained if NLR is a negative resistance of conductance numerically greater than $1/R$.

It is thus natural to interpret the term $\psi(v)$ as a resistance damping term whose magnitude in the general case depends on v. When the system is capable of self-oscillation, this resistance is negative for small values of v and positive for large values of v. Thus if v is initially zero and the system is given a small disturbance, it behaves as a negative resistance, NLR supplying more energy than is converted into heat in R. The amplitude therefore grows, the surplus energy being stored in the LC circuit, until ultimately $\psi(v)$ becomes positive at the crests of the oscillation. NLR then re-absorbs some of the energy which it supplies when v is small. Thus a steady amplitude of oscillation is reached when the net amount of energy supplied per cycle by NLR exactly equals the energy converted into heat in R per cycle.

We may thus describe the complete system as behaving like a negative resistance for small amplitudes and like a positive resistance for large amplitudes, the condition for a steady oscillation being that the average resistance over a cycle is zero.

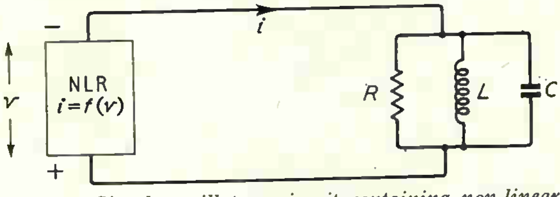

Fig. 1. *Simple oscillatory circuit containing non-linear resistance.*

The precise determination of the amplitude of oscillation, of course, requires the solution of the differential equation, but the above discussion explains the physical process which determines the amplitude of oscillation in terms of the simpler concept of resistance with which the engineer is already familiar. Indeed it suggests at once a procedure for determining the amplitude. When the stationary oscillation is established the energy absorbed per cycle in the resistance is $\int_0 \frac{v^2}{R} dt$ while the energy supplied by NLR is $-\int_0 vidt = -\int_0 vf(v)dt$ and equating these gives $\int_0 \frac{v^2}{R} dt = -\int_0 vf(v)dt$.

This is obtained from equation (2) by multiplying by v and integrating over a period. Assuming $v = V \cos(\omega t + \alpha)$, neglecting harmonics and taking a cubic characteristic $f(v) = a_1 v + a_2 v^2 + a_3 v^3$ gives, on evaluating the integrals,

$$\frac{V^2}{R} = -a_1 V^2 - \tfrac{3}{4} a_3 V^4$$

or

$$V^2 = -\frac{a_1 + 1/R}{\tfrac{3}{4}a_3} = \frac{|a_1| - 1/R}{\tfrac{3}{4}a_3}$$

since a_1 is negative in the case under consideration. This explanation of the phenomenon is the standard one which has been given in text-books since oscillators were first developed. Dealing as it does with the energy relations in the circuit it is fundamental to an understanding of the phenomenon. This approach does not, however, lend itself readily to the discussion of other aspects of oscillators such as the frequency deviation from the resonant frequency ω_0 of the linear circuit, and the phenomena connected with forced oscillations.

It was remarked earlier that a physical explanation of any phenomenon must be in terms of simple concepts, the relations between which the reader is prepared to accept without question. The resistance of the circuit in the above discussion was such a concept. Now in other non-linear devices such as detectors, frequency-changers and so on, the occurrence of 'modulation products' when simple periodic e.m.fs are applied to a non-linear circuit element forms the corner stone in the explanation of the phenomenon. Such devices are now so familiar to the communications engineer that the concept of 'modulation product' may be regarded as almost as familiar as that of resistance and an explanation of the phenomena connected with electrical oscillators in terms of modulation products may be as acceptable as the explanation of the amplitude determination given above in terms of the concept of resistance.

2. Modulation Products

Suppose a non-linear resistance has a characteristic

$$i = f(v)$$

and that a potential

$$v = V_1 \cos(\omega_1 t + \alpha_1) + V_2 \cos(\omega_2 t + \alpha_2) + \ldots + V_k \cos(\omega_k t + \alpha_k)$$

is applied to its terminals where

$$V_1 + V_2 + \ldots + V_k < V$$

so that the instantaneous value of v lies between $-V$ and $+V$ at every instant.

Now whatever the form of the function $f(v)$, provided only that it is continuous, it may, according to a theorem of Weierstrass, be represented within a given range to any required degree of accuracy by a polynomial. Thus we may find a polynomial

$$f_1(v) = a_1 v + a_2 v^2 + \ldots + a_n v^n$$

such that for $-V < v < V$ the difference between $f(v)$ and $f_1(v)$ is less than some pre-assigned arbitrarily small quantity. Thus we may with a negligible error replace $f(v)$ by $f_1(v)$ and assume

$$i = a_1 v + a_2 v^2 + \ldots + a_n v^n$$
$$= \sum_{r=1}^{n} a_r v^r$$

We may conveniently refer to $a_1 v$ as the first-order term of i, $a_2 v^2$ as the second-order term and so on.

Now let us suppose that v has the form given above; i.e.,

$$v = \sum_{h=1}^{k} V_h \cos(\omega_h t + \alpha_a)$$
$$= \sum_{h=1}^{k} \tfrac{1}{2} V_h \left\{ e^{j(\omega_h t + \alpha_h)} + e^{-j(\omega_h t + \alpha_h)} \right\}$$
$$= \sum_{h=-k}^{k} V_h e^{j\omega_h t}$$

where $V_h = \tfrac{1}{2} V_h e^{j\alpha_h}$

$$V_{-h} = \tfrac{1}{2} V_h e^{-j\alpha_h} = V_h^*$$

the asterisk following a letter denoting the conjugate complex number and

$$\omega_{-h} = -\omega_h.$$

With such a value for v the rth order terms of i will be

$$I(r) = a_r \left\{ \sum_{h=-k}^{k} V_h e^{j\omega_h t} \right\}^r$$

When this power is expanded the terms will occur in conjugate pairs which combine to give real cosine terms. A typical term in the expansion, consisting of a product of r factors each of which may be any one of the terms $V_h e^{j\omega_h t}$, will thus be of the form

$$a_r V_{h_1} V_{h_2} \ldots V_{h_r} \exp j(\omega_{h_1} + \omega_{h_2} + \ldots + \omega_{h_r})t,$$

and will combine with its conjugate to give a cosine term of angular frequency $|\omega_{h_1} + \omega_{h_2} + \ldots + \omega_{h_r}|$. Since each term in this sum is plus or minus one of the angular frequencies present in v we see that the rth order terms of i will contain every possible frequency which can be formed by a sum or difference combination of r of the frequencies present in v.

Thus, if only one frequency is present in v, the third-order terms of i will contain frequencies of the form $|\pm \omega_1 \pm \omega_1 \pm \omega_1|$ which gives two distinct frequencies ω_1 and $3\omega_1$. If two distinct frequencies are present in v the third-order terms of i will contain the frequencies $\pm \omega_h \pm \omega_k \pm \omega_l$ where each of h, k and l may be 1 or 2 giving ω_1, ω_2, $3\omega_1$, $3\omega_2$ $2\omega_1 - \omega_2$, $2\omega_1 + \omega_2$, $2\omega_2 - \omega_1$ and $2\omega_2 + \omega_1$, eight different frequencies.* Such frequency components are referred to as third-order modulation products.

The complex amplitude of any product is equal to the product of the coefficient of the appropriate power in the polynomial representation of the characteristic, the complex amplitudes of the frequency components in v which have been combined or their conjugates, and a multinomial numerical coefficient depending on the number of times that product occurs in the expansion of v^r.

With

$$v = V_1 \cos(\omega_1 t + \alpha_1) + V_2 \cos(\omega_2 t + \alpha_2)$$

the first-order (or linear) terms of i are

$$I^{(1)} = a_1 V_1 \cos(\omega_1 t + \alpha_1) + a_1 V_2 \cos(\omega_2 t + \alpha_2),$$

while the third-order terms are given by

$$I^{(3)} = a_3\{V_{-2}e^{-j\omega_2 t} + V_{-1}e^{-j\omega_1 t} + V_1 e^{j\omega_1 t} + V_2 e^{j\omega_2 t}\}^3$$

When the cube is multiplied out and conjugate terms combined this gives

$$\begin{aligned} I^{(3)} = \tfrac{1}{4}a_3 \{ & (3V_1^3 + 6V_1V_2^2)\cos(\omega_1 t + \alpha_1) + \\ & (3V_2^3 + 6V_1^2V_2)\cos(\omega_2 t + \alpha_2) \\ & + V_1^3 \cos(3\omega_1 t + 3\alpha_1) \\ & + V_2^3 \cos(3\omega_2 t + 3\alpha_2) \\ & + 3V_1^2V_2 \cos[(2\omega_1 + \omega_2)t + 2\alpha_1 + \alpha_2] \\ & + 3V_1^2V_2 \cos[(2\omega_1 - \omega_2)t + 2\alpha_1 - \alpha_2] \\ & + 3V_1V_2^2 \cos[(2\omega_2 + \omega_1)t + 2\alpha_2 + \alpha_1] \\ & + 3V_1 V_2^2 \cos[(2\omega_2 - \omega_1)t + 2\alpha_2 - \alpha_1]\} \end{aligned}$$

containing the eight different frequencies previously mentioned.

In the same manner the modulation products of other orders may be calculated.

The present argument does not depend on the polynomial form being the most convenient form by which to represent the non-linear characteristic. Whatever the form of the characteristic such modulation products will occur, and their production is an essential feature of a non-linear circuit, in contrast to a linear circuit in which no frequencies can occur other than those impressed on it by applied e.m.fs or those of its transient normal modes which are completely determined by the circuit parameters.

* ω_1 and ω_2 are each obtained from two combinations, thus $\omega_1 = \omega_1 + \omega_1 - \omega_1 = \omega_1 + \omega_2 - \omega_2$.

3. Negative-Resistance Oscillator

We return now to the consideration of the circuit of Fig. 1. Suppose first, that $\psi(0) = 0$ [i.e., $(1/C)\{1/R + f'(0)\} = 0$] giving $a_1 = -1/R$ since $f'(0) = a_1$. NLR is then a negative resistance, for small potentials, of magnitude equal to the linear circuit resistance. Let a small potential of frequency ω_0 exist at the terminals of the non-linear resistance, $v = V \cos(\omega_0 t + \alpha)$, V being sufficiently small for powers above the first in the characteristic $i = f(v)$ to be negligible. Then only the first-order term in the current will be appreciable and a current $i = a_1 V \cos(\omega_0 t + \alpha)$ will flow through NLR. But this current flowing through the linear circuit will produce exactly the potential v at the terminals of NLR. That is, the system will be self-maintaining and the current of this frequency and amplitude will continue to flow without external excitation.

If, secondly, $a_1 > -1/R$ (i.e., $|a_1| < 1/R$) then the flow of current in the linear circuit will be insufficient to maintain the potential at the NLR terminals. The amplitude will decrease and the current die away to zero unless maintained by the application of an external e.m.f.

On the other hand, if thirdly, $a_1 < -1/R$ (i.e., $|a_1| > 1/R$) the flow of current in the linear circuit will more than maintain the potential v at the NLR terminals. The amplitude will therefore grow unless opposed by a contrary external e.m.f.

In this third case with a growing amplitude the powers above the first in $f(v)$ can no longer be neglected and will have two effects which must be noted: (i) the production of modulation products in the current of harmonic frequencies $2\omega_0$, $3\omega_0$, . . .; and (ii) the production of higher-order modulation products of fundamental frequency ω_0.

Now the linear circuit will present a relatively small impedance to the higher-harmonic modulation products so that the flow of these currents through the linear circuit will produce only very small potentials at the NLR terminals. If for the moment we neglect these harmonic potentials we may regard the harmonic currents as usually inconvenient by-products.

On the other hand the higher modulation products of fundamental frequency will produce appreciable potentials at the NLR terminals and cannot be ignored. Now the falling characteristic of NLR can only exist in a limited range on either side of the operating point beyond which the gradient of the characteristic reverses. This means that higher odd powers of v in the polynomial must have positive coefficients. In the simplest case a_3 will be positive and we shall have in the current a third-order component of fundamental frequency $\frac{3}{4}a_3V^3 \cos(\omega_0 t + \alpha)$ which will increase relatively to the first-order component $a_1V \cos(\omega_0 t + \alpha)$ as V increases. Clearly this third-order current flowing in the linear circuit will produce a potential at the NLR terminals in exact anti-phase to the potential produced by the first-order current, which will therefore oppose the further increase of potential. If the fifth and higher powers are still negligible a stationary amplitude will be reached when the resultant potential produced by the flow of the first- and third-order currents through the linear circuit develops exactly the potential at the NLR terminals required to produce this current through NLR; i.e., when

$$-R[a_1V + \tfrac{3}{4}a_3V^3] \cos(\omega_0 t + \alpha) = V \cos(\omega_0 t + \alpha) \quad \ldots \quad (3)$$

giving

$$V^2 = -\frac{a_1 + 1/R}{\frac{3}{4}a_3} = \frac{|a_1| - 1/R}{\frac{3}{4}a_3} \quad \ldots \quad (4)$$

which gives the first approximation to the stationary amplitude. In practice, of course, the fifth- and higher-order modulation products may also contribute to the limitation of amplitude.

This analysis applies also to the first case since the powers of v above the first are never actually zero unless v is zero. The higher-order modulation currents of fundamental frequency will therefore produce a slow decay of amplitude, the final amplitude being zero as given by the above equation when $|a_1| = 1/R$.

4. Oscillator which is Stable at the Quiescent Point

The non-linear element of Fig. 1 may have a characteristic in which a_1 is positive, a_3 negative and a_5 positive. With a small potential applied to the NLR terminals the first-order current flowing in the linear circuit would produce an opposing potential and the amplitude would fall rapidly to zero unless externally maintained. On the other hand, the third-order current would produce a potential tending to maintain the current so that if a sufficiently large amplitude were produced initially for the effect of the third-order current to exceed sufficiently that of the first-order current, the amplitude would increase and the final limitation of amplitude would be produced by the fifth-order current of fundamental frequency.

We would then have a circuit which was stable for small disturbances, but which could be set into self-oscillation if a sufficiently large amplitude could be produced initially.

5. Frequency Determination

It is clear that all the modulation products considered in the previous sections are in phase or anti-phase with the potential and at the frequency ω_0 the linear circuit is resistive and produces no change of phase. A stationary state is thus possible at this frequency.

If a potential of any other frequency is applied to the NLR terminals the current flowing in NLR will be in phase or anti-phase with the potential, but this current flowing in the linear circuit will produce a potential of a different phase due to the reactive component of the linear circuit impedance, so that a stationary state is not possible. If the circuit is left to itself there will be a transient stage in which the frequency varies until a stationary oscillation at the frequency ω_0 is attained.

6. Frequency Deviation

It was assumed in the above discussion that the modulation currents of higher-harmonic frequencies flowing in the linear circuit produced negligible potential at the NLR terminals. In fact, of course, the impedance is never actually zero to such frequencies, so that small harmonic potentials will be produced. There will, for example, be a second-order modulation product of second-harmonic frequency $\frac{1}{2}a_2V^2 \cos(2\omega_0 t + 2\alpha)$. To this current the circuit will be capacitive so that the potential produced will lag by an angle β, say, giving a potential $V_2 \cos(2\omega_0 t + 2\alpha - \beta)$ say, at the NLR terminals.

We now have potentials of two different frequencies ω_0 and $2\omega_0$ at the NLR terminals and among the modulation products produced in the current will be a third-order difference frequency $2\omega_0 - \omega_0$ (i.e., a component of fundamental frequency given by $a_2V_2V \cos(\omega_0 t + \alpha - \beta)$ which flowing through the linear circuit will produce a potential $a_2RV_2V \cos(\omega_0 t + \alpha - \beta)$ lagging in phase by the angle β. This potential will be small but the phase shift prevents the attainment of a stationary amplitude at the frequency ω_0. To restore the correct phase relation a small frequency deviation from the frequency ω_0 is necessary so that the principal first- and third-order currents flowing through the linear circuit produce a potential slightly advanced in phase to compensate for the lag due to such intermodulation products formed between the fundamental and harmonic potentials at the NLR terminals. The linear circuit must, therefore, be slightly inductive at the oscillation frequency, and there will be again a transient stage in which the frequency falls and a stationary oscillation is produced at a frequency slightly less than ω_0.

There is thus in general a small frequency deviation below the resonant frequency of the linear circuit, which is intimately related to the amount of harmonics present and to the reactive component of the impedance at the harmonic frequencies. The exact expression for this frequency deviation was first obtained in its general form by Groszkowski.[1]

If we define the equivalent impedance of NLR at the fundamental frequency under the particular operating conditions to be the ratio of the fundamental components of potential and current, then at the frequency ω_0 the two are out of phase and the impedance must contain a capacitive component.

A condition of resonance (i.e., zero effective reactance of the complete circuit) is only obtained at a frequency slightly less than ω_0 when the linear circuit has an inductive component which cancels the capacitive component just mentioned. In this sense we may say that the free oscillation always takes place at a resonant frequency. This is, however, simply another way of saying that the combined result of the linear reactance and the intermodulation effects in NLR must be to produce zero loop phase shift, if a steady oscillation is to be maintained.

7. Forced Oscillations

If an e.m.f. $E \cos \omega_1 t$ is introduced into the circuit of Fig. 1 then we may expect potentials of two frequencies, ω_1 and the oscillation frequency ω, to be developed at the NLR terminals; and hence that the current will contain, in addition to these frequencies, the various modulation products; that is, harmonic and sum and difference combination frequencies.

Let us assume that none of these frequencies other than ω is near to ω_0. This means that ω_1 must not be near to any harmonic or sub-harmonic of ω_0. We may then assume that the linear circuit presents negligible impedance to these frequencies and that therefore no appreciable potentials other than those of frequencies ω_1 and ω are developed at the NLR terminals. We need then only concern ourselves with the modulation products in the current of fundamental frequency which arise from these potentials.

If

$$v = V_1 \cos (\omega_1 t + \alpha_1) + V \cos (\omega t + \alpha)$$

then the third-order current of frequency ω will be

$$\tfrac{3}{4} a_3 (V^3 + 2VV_1^2) \cos (\omega t + \alpha)$$

which remains in phase with the corresponding component of v so that the free oscillation frequency will still coincide, to this order of approximation, with ω_0, but its amplitude is increased by the intermodulation term $2VV_1^2$.

The stationary amplitude of the free oscillation will now be determined by the condition

$$- R[a_1 V + \tfrac{3}{4} a_3 (V^3 + 2VV_1^2)] \cos (\omega_0 t + \alpha)$$
$$= V \cos (\omega_0 t + \alpha)$$

instead of by equation (3).

This gives in addition to the solution $V = 0$

$$V^2 + 2V_1^2 = V_0^2$$

replacing (4) where V_0 is the amplitude of the free oscillations when no external e.m.f. is present.

If $2V_1^2 > V_0^2$ this equation gives no real value for V and the only solution is $V = 0$; that is, the free oscillation is absent. This means that if a potential of this frequency is produced, an intermodulation product of frequency $\omega_0 + \omega_1 - \omega_1 = \omega_0$ arises in the current which, flowing in the linear circuit, produces an opposing potential which causes this component to die away to zero, leaving only the forced oscillation. It will be noted that this opposing potential is of the first order in V and is therefore effective no matter how small the value of V.

Thus a sufficiently large applied e.m.f. will bring about the complete suppression of the free oscillation, the suppression being brought about by the effect of the intermodulation products in producing potentials which oppose the growth of any free oscillation which may be initiated.

We have neglected the higher-order modulation products so that the condition $2V_1^2 > V_0^2$ is only approximate. The general argument will, however, be unaffected and, if account is taken of the higher-order products, there will still be a definite critical value of V_1 above which the free oscillation is suppressed.

8. Amplitude of the Forced Oscillation

If we assume that the free oscillation is absent, we may proceed to form the equation for the amplitude of the forced oscillation after the manner in which equation (3) was obtained. Taking account of the third-order modulation current of fundamental frequency we now have the fundamental component of current as $(a_1 V_1 + \tfrac{3}{4} a_3 V_1^3) \cos (\omega_1 t + \alpha_1)$. This flows through a linear circuit, which has a reactive component at this frequency.

If ω_1 is near to ω_0 and we write $\omega_1 = \omega_0 + \Delta\omega$ the reactance will be

$$\left(\frac{dX}{d\omega}\right)_{\omega = \omega_0} . \Delta\omega = - 2CR^2 \Delta\omega.$$

The potential developed when this current flows through the linear resistance now has a quadrature component and equation (3) is replaced by

$$- R[a_1V_1 + \tfrac{3}{4}a_3V_1^3] \cos (\omega_1 t + \alpha_1)$$
$$- 2CR^2 \Delta\omega \, [a_1V_1 + \tfrac{3}{4}a_3V_1^3] \sin (\omega_1 t + \alpha_1)$$
$$+ E \cos \omega_1 t = V_1 \cos (\omega_1 t + \alpha_1) \quad \ldots \quad (5)$$

If we replace $E \cos \omega_1 t$ by

$$E \cos \{(\omega_1 t + \alpha_1) - \alpha_1\}$$
$$= E \cos \alpha \cos (\omega_1 t + \alpha_1) + E \sin \alpha \sin (\omega_1 t + \alpha_1)$$

and equate coefficients of $\cos (\omega_1 t + \alpha_1)$ and $\sin (\omega_1 t + \alpha_1)$ for the equation to be satisfied for all values of t, we obtain

$$\begin{cases} - R[a_1V_1 + \tfrac{3}{4}a_3V_1^3] + E \cos \alpha_1 = V_1 \\ - 2CR^2\Delta\omega \, [a_1V_1 + \tfrac{3}{4}a_3 V_1^3] + E \sin \alpha_1 = 0 \end{cases}$$

These two equations determine V_1 and α_1 for given values of E and $\Delta\omega$. They may be re-arranged in the form

$$\begin{cases} E \cos \alpha_1 = (1 + a_1R) V_1 + \tfrac{3}{4}a_3 RV_1^3 \\ E \sin \alpha_1 = 2CR^2 \Delta\omega \, [a_1V_1 + \tfrac{3}{4}a_3 V_1^3]. \end{cases}$$

Now other modulation products will only be negligible if E and V_1 are small and the negative resistance is small (i.e., $|a_1| - 1/R$ is small) so that $\tfrac{3}{4}a_3V_1^3$ is comparable with $(1 + a_1R) V_1$ in the first equation, but is negligible in comparison with a_1V_1 in the second equation. Neglecting $\tfrac{3}{4}a_3V_1^3$ in the second equation we now have

$$E \cos \alpha_1 = (1 + a_1R) V_1 + \tfrac{3}{4}a_3RV_1^3 \quad \ldots \quad (6)$$

$$E \sin \alpha_1 = 2CR^2 \Delta\omega a_1V_1. \quad \ldots \quad \ldots \quad (7)$$

Squaring and adding to eliminate α_1 gives

$$E^2 = V_1^2[(1 - |a_1|R + \tfrac{3}{4}a_3RV_1^2)^2 + (2CR^2a_1\Delta\omega)^2] \quad \ldots \quad \ldots \quad \ldots \quad (8)$$

an equation determining the amplitude of the forced oscillation near resonance.

To reduce this equation to a more manageable form we divide through by $V_0^2(|a_1|R - 1)^2$, where V_0 is the amplitude of free oscillation with no applied e.m.f. as given by equation (4); i.e.,

$$V_0^2 = \frac{|a_1| - 1/R}{\tfrac{3}{4}a_3}$$

Thus we obtain

$$\frac{E^2}{V_0^2(|a_1|R - 1)^2} = \frac{V_1^2}{V_0^2}\left[\left(1 - \frac{V_1^2}{V_0^2}\right) + \left(\frac{2CR^2a_1 \, \Delta\omega}{1 - |a_1|R}\right)^2\right]$$

If we now write

$$F^2 = \frac{E^2}{V_0^2(|a_1|R - 1)^2}$$

$$y = \frac{V_1^2}{V_0^2}$$

$$x = \frac{2CR^2|a_1|\Delta\omega}{|a_1|R - 1}$$

so that F is proportional to the amplitude of applied e.m.f., y to the square of the amplitude of forced oscillation, and x to the frequency difference from ω_0, the equation becomes

$$F^2 = y[(1 - y)^2 + x^2] \quad \ldots \quad \ldots \quad \ldots \quad (9)$$

The curves of y against x for fixed F give the resonance curves of the system. These curves (Fig. 2) were first given by Van der Pol. They are symmetrical about the y-axis; i.e., about the anti-resonant frequency ω_0 of the linear circuit. For large F or away from ω_0 they look very like the resonance curves of a linear oscillator, but for small F near to ω_0, ($x = 0$), they show three possible amplitudes of oscillation.

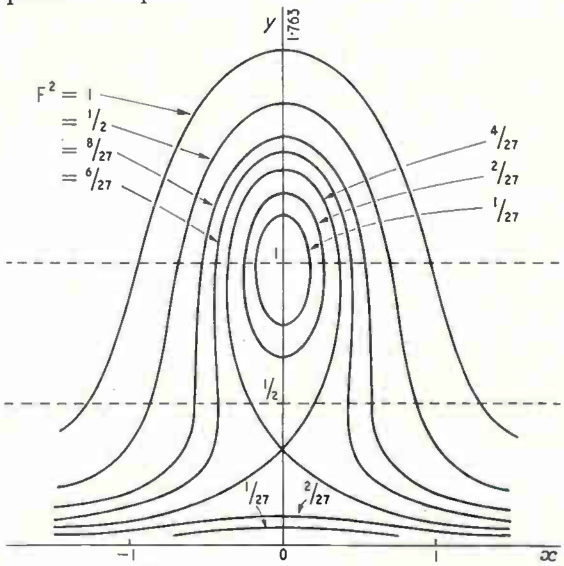

Fig. 2. *Van der Pol resonance curves.*

The reason for this may be seen if attention is limited to an applied e.m.f. of frequency ω_0; i.e., $\Delta\omega = 0$. Referring to the free oscillation alone in the first place we have seen that if a small potential of amplitude V exists at the terminals of NLR the current flowing in the linear circuit develops a greater potential in phase and, in consequence, the amplitude grows. The third-order modulation voltage in antiphase then comes into operation, and if the current became sufficiently large this third-order voltage would produce an opposing potential tending to reduce the amplitude, a stable oscillation being obtained when the first- and third-order components combine to give exactly the voltage at the NLR terminals required to maintain the current. This is shown in Fig. 3, in which the abscissa is the voltage required at NLR terminals to maintain the current, this also being the ordinate to the straight line OA. The curve ORSP represents the voltage returned to NLR terminals by the flow of current in the linear circuit, the stable oscillation being represented by the intersection at P.

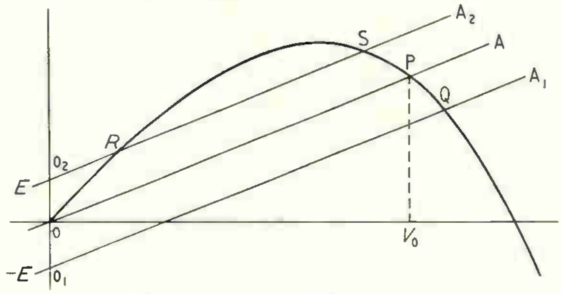

Fig. 3. *Determination of stationary amplitudes.*

Now imagine a small e.m.f. of amplitude E injected into the circuit, in phase with the current and therefore increasing the amplitude of the returned voltage. This will be represented by lowering the line OA by E to the new position O_1A_1 and the intersection will move to Q, a larger current having to flow before the third-order modulation voltage produces equilibrium. Clearly again a slightly smaller voltage will give a returned voltage tending to increase the amplitude, and the reverse for a slightly larger voltage, so that the oscillation represented by Q appears to be stable.

On the other hand balance can also be obtained if the injected voltage is in antiphase, represented by raising OA by E to O_2A_2. This gives two intersections at R and S if E is not too large. The lower one at R is evidently unstable since a slightly smaller voltage gives a returned voltage tending to decrease the amplitude and the reverse for a slightly larger voltage. If the voltage were slightly decreased it would therefore continue to decrease, passing through zero amplitude with phase reversal, bringing the injected e.m.f. into phase, the amplitude then increasing until the oscillation represented by Q is reached. If the amplitude were increased slightly on the other hand it would continue to increase until S was reached, the oscillation represented by S appearing to be stable. To show that the latter is in fact unstable, while the oscillation represented by Q is stable, we have to consider the effect of a slight variation of phase rather than amplitude. Suppose that in the situation represented by S the phase of the injected voltage is advanced slightly. Owing to the relation of phase opposition, the effect is to retard the phase of the voltage at NLR and therefore to increase the phase discrepancy, while in the case represented by Q the reverse effect occurs, and any slight disturbance of phase tends to be rectified. This is shown in Fig. 4, (a) referring to Q, and (b) to S.

Thus the two oscillations of smaller amplitudes are unstable. The middle one, although stable for variations of amplitude, is unstable for variations of phase. It is a type of equilibrium point known mathematically as a saddle point or col.

For other frequencies the situation is complicated by the phase shift produced by the linear circuit, the various voltages no longer having a simple 'in phase' or 'in anti-phase' relationship. More detailed analysis is then required and the lowest amplitude may become stable near the boundary of the frequency interval in which three amplitudes can occur.

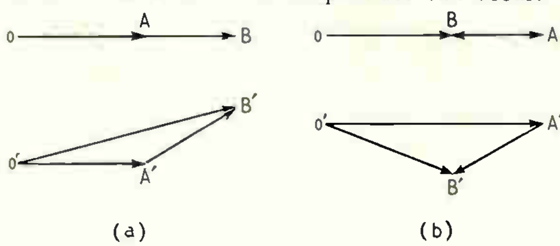

Fig. 4. *Showing effect of rotation of phase of injected voltage. OB represents voltage at NLR terminals; OA represents voltage developed by the flow of current in the linear circuit; AB represents the injected voltage; upper figures show stationary positions of vectors. In lower figures the phase of AB is slightly advanced. In (a), OB rotates so as to reduce the discrepancy giving stability, while in (b), OB rotates so as to increase the discrepancy giving instability.*

9. Effect on Forced Oscillation Amplitude of Other Modulation Products

It has been shown in connection with the free oscillation that when secondary modulation products of fundamental frequency are formed by intermodulation between harmonics, or between harmonics and fundamental, a phase shift will generally be introduced between the current and the voltage at NLR terminals, NLR then behaving as though it had a capacitive component. Zero phase shift is then obtained at a slightly lower frequency when the inductive reactance of the linear circuit neutralizes the phase shift of NLR.

Exactly the same effect occurs with the forced oscillation. At the frequency ω_0 a phase shift introduced in NLR has to be neutralized by a reverse phase difference between the returned voltage and the injected voltage, and in consequence the maximum voltage is not obtained at the NLR terminals. At a lower frequency when the capacitive phase shift of NLR is neutralized by the inductive phase shift of the linear circuit the returned voltage is brought into phase with the injected voltage, giving a resonance condition for the complete circuit. The maximum of the resonance curve is thus displaced towards the lower frequencies by an amount which increases with amplitude—since these secondary modulation products are relatively more important as the amplitude increases—and which is always greater than the deviation of the free oscillation frequency, approaching the latter as E tends to zero. When all modulation products of third order are taken into account the resonance curves take the form shown in Fig. 5.

When the impedance characteristic is unsymmetrical as in the case of a resistance-capacitance oscillator, asymmetry in the reverse direction is obtained. This is due to the increased in-phase component of returned voltage which can be obtained at a higher frequency. This cannot give an increased amplitude of free oscillation since the condition of zero phase shift cannot be satisfied at the higher frequency. When, however, the oscillator is synchronized to an injected e.m.f., zero phase shift can be achieved by a small phase difference between the returned voltage and the injected voltage, and an increased amplitude thus obtained until the effect of increasing phase shift begins to predominate and the amplitude falls off again. In such a circuit the maxima of the resonance curves may be displaced towards higher frequencies.

10. Effect of the Forced Oscillation on the Free Oscillation Frequency

It was seen earlier that when there is no external e.m.f. the free oscillation frequency coincided with ω_0 the resonant frequency of the linear circuit, provided no appreciable harmonic potentials were developed at the NLR terminals. But if account is taken of the harmonic potentials, then the intermodulation between these and the fundamental gives rise to a shift in phase of the current, which necessitates a deviation of frequency from the value ω_0 for a stationary oscillation to be possible.

The same is true when an external e.m.f. is present. So long as no appreciable potentials are developed at the NLR terminals other than those of the frequencies of the external e.m.f. and the free oscillation, then the free oscillation frequency will concincide with ω_0, as has already been shown.

On the other hand when account is taken of the harmonic and sum- and difference-frequency potentials developed at the NLR terminals further modulation currents of the free fundamental frequency will be produced, which will be out of phase with the potential, so that again a small deviation of frequency from the value ω_0 must occur in order to correct this phase shift. Some of these modulation currents will involve the forced oscillation; there will, for example, be a third-order current arising from the potentials of frequency ω_1 and $\omega_1 - \omega$ of frequency equal to the difference $\omega_1 - (\omega_1 - \omega) = \omega$. The amount of this frequency deviation will thus depend on the frequency and amplitude of the forced oscillation. The free oscillation frequency in the presence of an applied e.m.f. will therefore differ slightly from its value when no external e.m.f. is present.

This observation emphasizes the fact that a non-linear circuit, unlike a linear system, does not have fixed resonant and anti-resonant frequencies, which are determined solely by the circuit parameters and are independent of any external excitation.

The amount of this frequency variation depends, both as regards its magnitude and direction, in a rather complicated manner on the frequency and amplitude of the forced oscillation. It is small, and is only of importance when the frequency of the applied e.m.f. approaches that of the free oscillation, when it has a significant effect in the phenomenon of synchronization which is discussed in a later section.

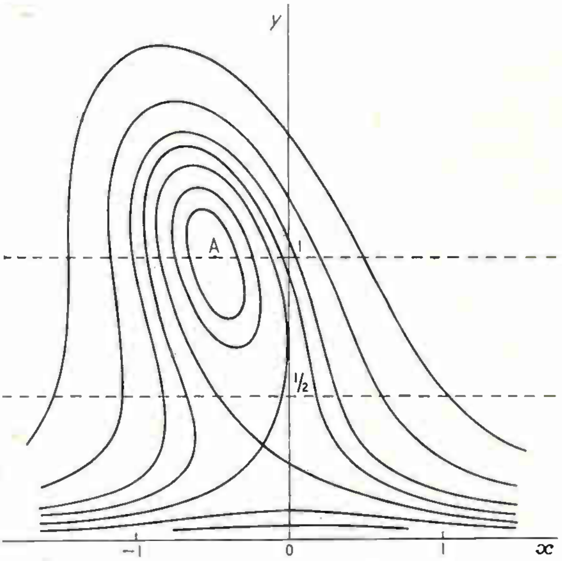

Fig. 5. *Unsymmetrical resonance curves; the point A represents the free oscillation with zero-applied e.m.f.*

11. Form of Resonance Curves and Synchronization

The general form of the resonance curves was shown in Figs. 2 and 5. For a large applied e.m.f. the amplitude has a maximum value for a frequency near to ω_0 due to the fact that near to this frequency the flow of current in the linear circuit produces a potential which reinforces the generator e.m.f. at the NLR terminals. If the effect of harmonic voltages is ignored, the maximum amplitude, for a given amplitude of applied e.m.f., occurs exactly at frequency ω_0, the voltage developed across the linear circuit being then in phase with the applied e.m.f. As the frequency varies from this value the amplitude falls, the phase relation being upset owing to the phase shift of the linear circuit reactance, until when the amplitude falls below the critical value $V_0/\sqrt{2}$ the free oscillation begins to develop.

This may be regarded as an ordinary resonance phenomenon. In the case of the free oscillation it was shown earlier that when harmonic voltages were taken into account, the intermodulation of these with the fundamental gave rise to out-of-phase currents which necessitated a deviation of frequency from ω_0. That is, the effective resonance frequency of the complete non-linear circuit was slightly less than ω_0. The same effect occurs in the case of the forced oscillation and causes a displacement of the maximum of the resonance curve to a frequency less than ω_0. The amount of this displacement increases with the amplitude, being always greater than the frequency deviation of the free oscillation, approaching that value as the amplitude E of the applied e.m.f. tends to zero, the amplitude V_1 at the same time approaching V_0.

When the amplitude E of the applied e.m.f. is small there is always a frequency interval within which there is no separate free oscillation, no matter how small the value of E may be. This is the phenomenon of synchronization, the oscillator being said to be synchronized to the frequency of the small applied e.m.f.

As the frequency is varied with a fixed small amplitude E, the breakaway from synchronization, or 'pull out' as it is called (that is, the frequency at which a separate free oscillation begins to develop), occurs before the amplitude has fallen to the critical value $V_0\sqrt{2}$ applicable for a large applied e.m.f.

The width of the frequency interval in which synchronization occurs decreases and tends to zero with E, its centre being displaced in the manner described above, the frequency and amplitude approaching those of the free oscillation when no external e.m.f. is applied, as $E \rightarrow 0$.

It has been remarked that at the maximum of the amplitude of the forced oscillation for a given amplitude of e.m.f., the conditions are those of resonance, the current, potential at the NLR terminals, and e.m.f. being all in phase with each other. We may therefore regard the effect of the applied e.m.f. at this point as the production of a frequency deviation of the free oscillation, which brings the frequency of the free oscillation into coincidence with that of the forced oscillation, thereby enabling a large amplitude to be developed with only a small applied e.m.f. without any separate free oscillation appearing.

It appears, as one might anticipate, that a sufficient deviation is only possible when the frequency of the applied e.m.f. is near to the normal frequency of free oscillation. With a large e.m.f. the interval of synchronization is extended by the intermodulation effect already described preventing the development of a free oscillation until the amplitude V_1 falls to the critical value $V_0/\sqrt{2}$. The free oscillation then develops at a frequency near to its normal value, still of course modified slightly by the intermodulation of harmonic and sum and difference voltages with the fundamental. A normal beat is obtained between the free and forced oscillation.

With a small e.m.f. synchronization ceases when V_1 is of the same order as V_0, and a very slow beat is obtained indicating a free oscillation frequency very close to that of the forced oscillation. The form of the beat is not that obtained from two approximately simple periodic functions but has been described as appearing as though the oscillation slipped at regular intervals, synchronization being maintained between successive 'slips.'[2]

As the frequency of applied e.m.f. is varied further, the slips occur at shorter intervals and later the form of the beat changes progressively until a normal beat is obtained.

The last paragraph describes what is actually observed without attempting an explanation. It supports the view put forward earlier, that the effect of a small applied e.m.f. of frequency near to the normal frequency of free oscillation is to bring the frequency of the free oscillation into coincidence with the frequency of the applied e.m.f. The conditions during pull-out with a small e.m.f. are clearly much more complicated than any which have been considered so far.

12. The Pull-out from Synchronization

The discussion of the phenomenon of pull-out from synchronization with a small applied e.m.f. involves the consideration of potentials of two different frequencies both near to ω_0 existing at the NLR terminals, namely the potentials of the free and forced oscillations. To appreciate the significance of this it is necessary to consider more closely the process which has been used in the previous discussion.

In discussing the free oscillation without any external e.m.f., for example, we considered a potential of fundamental frequency at the NLR terminals. This produces currents in NLR of the fundamental and harmonic frequencies. The harmonic currents flowing in the linear circuit produce small harmonic potentials at the NLR terminals. These in turn intermodulate with each other and the fundamental to produce further modulation currents, which in turn flow through the linear circuit and produce potentials at the NLR terminals, and so on. The process is thus, in reality, one of successive approximation, taking into account modulation currents of successively higher orders. So long as the potentials developed by the currents of harmonic and combination frequencies flowing through the linear circuit are small, the process converges; the previous discussion takes into account the first approximations using modulation products up to the third order.

Now when we have two fundamental e.m.fs at the NLR terminals, the forced oscillation of frequency ω_1, and a free oscillation of frequency ω, and both these frequencies are near to ω_0, then currents of these frequencies will develop appreciable potentials in the linear circuit which will be very close to the correct phase relation for the oscillation to build up until limited by the effect of the higher modulation products. Furthermore modulation products will be formed of frequencies $2\omega_1 - \omega$, $2\omega - \omega_1$, $3\omega_1 - 2\omega$, $3\omega - 2\omega_1$, and so on, which will likewise be near to ω_0, and so when potentials of these frequencies are produced they will also tend to build up. If we put $\omega_1 = \omega + \delta\omega$ then $2\omega_1 - \omega = \omega_1 + \delta\omega$, $2\omega - \omega_1 = \omega_1 - 2\delta\omega$, $3\omega_1 - 2\omega = \omega_1 + 2\delta\omega$, $3\omega - 2\omega_1 = \omega_1 - 3\delta\omega$, and so on. Thus as soon as a free oscillation of frequency $\omega_1 - \delta\omega$ begins to develop, we shall have potentials at the NLR terminals of frequencies $\omega_1 \pm \delta\omega$, $\omega_1 \pm 2\delta\omega$, $\omega_1 \pm 3\,\delta\omega$, and so on, all of which tend to build up; that is, currents of these frequencies will develop appreciable potentials when flowing in the linear circuit, all approximately in the same phase relationship. In such circumstances the successive approximation process will no longer converge.

If we consider the situation at the frequency at which the oscillator is just on the point of pulling out, there is a considerable oscillation of frequency ω_1 and a free oscillation we suppose of very slightly different frequency $\omega_1 - \delta\omega$. Then from what was said above the whole set of frequencies $\omega_1 \pm \delta\omega$, $\omega_1 \pm 2\delta\omega$, $\omega_1 \pm 3\delta\omega$, ... will appear simultaneously, any one of which might be regarded as a free oscillation, since they are all of frequencies near to ω_0 and owing to the frequency deviation, no one of them is likely to have a frequency exactly equal to ω_0. We may therefore anticipate a whole set of components of frequencies $\omega_1 \pm r\delta\omega$ developing with small amplitudes approximately equal for small values of r, and ultimately falling away to zero as r increases so that the frequencies are further removed from ω_0.

The effect of such a train of frequency components may be estimated by considering frequencies up to $\omega_1 \pm n\delta\omega$ of equal amplitudes v. The sum will be

$$\sum_{r=-n}^{+n} \mathrm{v}\cos(\omega_1 + r\delta\omega)t = \mathrm{v}.\frac{\sin(n+\frac{1}{2})\delta\omega t}{\sin\frac{1}{2}\delta\omega t}.\cos\omega_1 t$$

Thus we obtain an oscillation of forced oscillation frequency ω_1 and amplitude v. $\left|\dfrac{\sin(n+\frac{1}{2})\delta\omega t}{\sin\frac{1}{2}\delta\omega t}\right|$ which oscillates between small positive and zero values except when $\frac{1}{2}\delta\omega t$ is an integral multiple of π when it rises to a sharp maximum value $(2n + 1)$v. The graph of this amplitude for $n = 7$ is shown in Fig. 6. This clearly gives the sort of effect we require. For most of the time the amplitude is negligible the different components mutually cancelling each other, to leave the undisturbed forced oscillation, but at regular intervals $2\pi/\delta\omega$, all the components come into phase to give, as it were, a pulse which combines with the forced oscillation to produce the apparent slip from synchronization.

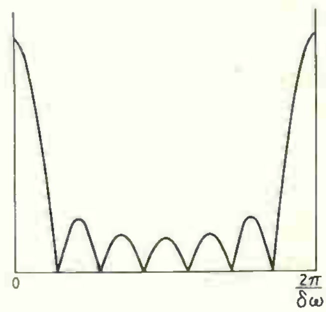

Fig. 6. *Amplitude of sum of equally-spaced frequency components.*

As the frequency of applied e.m.f. is varied we may assume that $\delta\omega$ increases so that the number of these components within what one might term the resonance region of the frequency scale will decrease; i.e., n decreases while v may increase so that $(2n + 1)$v retains roughly the same value. The dip in amplitude of the forced oscillation would then occur at more frequent intervals and at the same time become less sharp until it ultimately merged into a normal beat note between the free and forced oscillations.

13. Impedance of Circuit when Amplitude and Phase of Current Vary Slowly

It has been remarked that during the growth of an oscillation there will be a variation of amplitude and frequency until the voltage returned to the NLR terminals is just sufficient both in amplitude and phase to maintain the oscillation. We may formulate the mathematical equations of this transient stage on the assumption that the amplitude and phase of the oscillation vary so slowly that the rates of change are sensibly constant over a number of cycles of the oscillation.

Suppose we have a current $i = I\cos(\omega_1 t + \phi)$ in which I and ϕ vary slowly, having values I_0 and ϕ_0 at a particular instant t_0. Then for several cycles we will have approximately

$$\phi = \phi_0 + \left(\frac{d\phi}{dt}\right)_0 (t - t_0) \qquad \text{and}$$

$$\log I = \log I_0 + \frac{1}{I_0}\left(\frac{dI}{dt}\right)_0 (t - t_0)$$

or

$$I = I_0 \exp\left\{\frac{1}{I_0}\left(\frac{dI}{dt}\right)_0 (t - t_0)\right\}$$

Therefore

$$i = I_0 \exp\left\{\frac{1}{I_0}\left(\frac{dI}{dt}\right)_0 (t - t_0)\right\}$$

$$\cos\left\{\left[\omega_1 + \left(\frac{d\phi}{dt}\right)_0\right](t - t_0) + \omega_1 t_0 + \phi_0\right\}$$

$$= \mathrm{Re}\,\mathbf{I} \exp\left\{\frac{1}{I_0}\left(\frac{dI}{dt}\right)_0 + j\left[\omega_1 + \left(\frac{d\phi}{dt}\right)_0\right]\right\}(t - t_0)$$

with $\mathbf{I} = I_0 \exp j(\omega_1 t_0 + \phi_0)$ as the instantaneous complex amplitude.

Thus

$$i = \mathrm{Re}\,\mathbb{I} \exp p(t - t_0)$$

with

$$p = \frac{1}{I_0}\left(\frac{dI}{dt}\right)_0 + j\left[\omega_1 + \left(\frac{d\phi}{dt}\right)_0\right]$$

The current oscillation at any instant t is thus represented in the standard exponential form with frequency $\omega_1 + \frac{d\phi}{dt}$ and amplitude varying at the fractional rate $\frac{1}{I}\frac{dI}{dt}$ represented by the real term in the index of the exponential.

When this current flows through an impedance Z the potential drop across the impedance, instead of being given by $Z(j\omega_1).\mathbb{I}$ will now be given by $Z(p).\mathbb{I}$.

14. The Equations determining Transient Behaviour

Consider now the circuit of Fig. 1, assuming for simplicity a cubic characteristic

$$i = a_1 v + a_3 v^3$$

or alternatively assuming that the effect of modulation products of frequencies not near to ω_0 is negligible. Let an e.m.f. $E \cos \omega_1 t$ be introduced into the circuit and let a potential $V_1 \cos (\omega_1 t + \phi)$ exist at the NLR terminals. A fundamental current

$$(a_1 V_1 + \tfrac{3}{4} a_3 V_1^3) \cos (\omega_1 t + \phi) = I \cos (\omega_1 t + \phi)$$

will flow round the circuit.

Regarding amplitudes and phases as slowly varying as already indicated, the impedance of the circuit is

$$Z(p) = Z\left[\frac{1}{I}\frac{dI}{dt} + j\left(\omega_0 + \Delta\omega + \frac{d\phi}{dt}\right)\right]$$

where $\omega_1 = \omega_0 + \Delta\omega$, and if it is assumed that $\frac{1}{I}\frac{dI}{dt} + j\left(\Delta\,\omega + \frac{d\phi}{dt}\right)$ is small then

$$Z(p) = Z(j\omega_0) + Z'(j\omega_0)\left\{\frac{1}{I}\frac{dI}{dt} + j\left(\Delta\,\omega + \frac{d\phi}{dt}\right)\right\}.$$

For the circuit of Fig. 1 we have $Z(j\omega_0) = R$ and $Z'(j\omega_0) = -2CR^2$ so that

$$Z(p) = R - 2CR^2 \left\{\frac{1}{I}\frac{dI}{dt} + j\left(\Delta\omega + \frac{d\phi}{dt}\right)\right\}.$$

The potential developed across the linear circuit is therefore

$$I\left[\left(R - 2CR^2.\frac{1}{I}\frac{dI}{dt}\right)\cos (\omega_1 t + \phi) + 2CR^2\left(\Delta\omega + \frac{d\phi}{dt}\right)\sin (\omega_1 t + \phi)\right]$$

For the complete circuit we now have instead of equation (5)

$$-\left(IR - 2CR^2\frac{dI}{dt}\right)\cos (\omega_1 t + \phi) - 2CR^2 I\left(\Delta\omega + \frac{d\phi}{dt}\right)\sin (\omega_1 t + \phi) + E \cos \omega_1 t = V_1 \cos (\omega_1 t + \phi) \quad .. \quad (10)$$

Separating the 'in phase' and quadrature components as before we obtain

$$\begin{cases} -\left(IR - 2CR^2\frac{dI}{dt}\right) + E\cos\phi = V \\ -2CR^2\left(\Delta\omega + \frac{d\phi}{dt}\right)I + E\sin\phi = 0 \end{cases}$$

Putting $I = a_1 V_1 + \frac{3}{4}a_3 V_1^3$ in the first equation but neglecting the $\frac{3}{4}a_3 V_1^3$ in the second equation, as was done before, and for the same reason neglecting the term $\frac{3}{4}a_3 3V_1^2\frac{dV_1}{dt}$ in

$$\frac{dI}{dt} = a_1\frac{dV_1}{dt} + \tfrac{3}{4}a_3.3V_1^2\frac{dV_1}{dt}$$

we have

$$\begin{cases} -R(a_1 V_1 + \frac{3}{4}a_3 V_1^3) + 2CR^2 a_1\frac{dV_1}{dt} + E\cos\phi = V_1 \\ -2CR^2\left(\Delta\omega + \frac{d\phi}{dt}\right)a_1 V_1 + E\sin\phi = 0 \end{cases}$$

or

$$E\cos\phi = (1 + a_1 R)V_1 + \tfrac{3}{4}a_3 R V_1^3 - 2CR^2 a_1\frac{dV_1}{dt} \quad .. \quad .. \quad .. \quad .. \quad (11)$$

$$E\sin\phi = 2CR^2 a_1 V_1\left(\omega + \Delta\frac{d\phi}{dt}\right) \quad .. \quad (12)$$

Comparing (10) with (5), or (11) and (12) with

(6) and (7), it is seen that when the voltage returned to the NLR terminals differs from that required to maintain the current, the difference is corrected by additional voltages due to the varying amplitude and phase of the current in the linear circuit. When a stationary periodic oscillation exists dV_1/dt and $d\phi/dt$ are both zero and the present equations reduce to the earlier ones.

Equations (11) and (12) determine the non-periodic solutions of the circuit equations to a first approximation, provided the external e.m.f. is of small amplitude, and the frequency near to ω_0. Both the transient building up of a stationary oscillation, and combination oscillations, in which both free and forced oscillations are present, are included.

Unfortunately an exact analytical solution of these equations is not known. They were first derived using a different procedure by Van der Pol, and the general form of their solutions has been extensively discussed, the most thorough treatment being due to Dr. Cartwright.[4]

If we divide equations (11) and (12) by $V_0(|a_1|R - 1)$ we have

$$\begin{cases} \dfrac{E}{V_0(|a_1|R-1)}\cos\phi = -\dfrac{V_1}{V_0} + \dfrac{V_1^3}{V_0^3} + \dfrac{2CR^2|a_1|}{|a_1|R-1}\dfrac{d}{dt}\left(\dfrac{V_1}{V_0}\right) \\ \dfrac{E}{V_0(|a_1|R-1)}\sin\phi = -\dfrac{2CR^2|a_1|}{|a_1|R-1}\dfrac{V_1}{V_0}\left(\Delta\omega + \dfrac{d\phi}{dt}\right) \end{cases}$$

and introducing F, x and also $b = \dfrac{V_1}{V_0}$ and $\dfrac{a_1|R-1}{2CR^2|a_1|}\ t = \tau$ we obtain the standard form

$$\begin{cases} F\cos\phi = -b(1-b^2) + \dfrac{db}{d\tau} \quad \ldots \quad \ldots \quad (13) \\ F\sin\phi = -bx - b\dfrac{d\phi}{d\tau} \quad \ldots \quad \ldots \quad \ldots \quad (14) \end{cases}$$

These changes amount only to changes of scale, which allow the discussion of these equations independently of particular circuit parameters.

F, b, x and τ are proportional respectively to the amplitude of applied e.m.f., to the amplitude of the potential at the NLR terminals, to the frequency difference from ω_0 and to the time t.

15. Vector Representation of Transient Behaviour and Combination Oscillations

Taking account of these changes of scale the vector representing the voltage at the NLR terminals will be of length b at an angle ϕ to the vector representing the applied e.m.f., to which the phase angle ϕ is referred. Its end point is the point (b, ϕ) on a plane with b and ϕ as polar co-ordinates. This plane is called the Van der Pol plane. Equations (13) and (14) may be regarded as determining the radial and transverse components of velocity, $\dfrac{db}{d\tau}$ and $b\dfrac{d\phi}{d\tau}$, of the end point of this vector. A definite direction and speed of motion of this end point is defined at each point of the (b, ϕ) plane, with the exception of those points for which $\dfrac{db}{d\tau}$ and $b\dfrac{d\phi}{d\tau}$ both vanish. These points are the singular points of the differential equations and give the positions of the vector corresponding to the stationary synchronized oscillations.

If the vector is initially at any other position its end point will move along a trajectory which is an integral curve of the differential equations. For certain values of the parameters the end point of the vector may return to its initial position after a certain interval, the trajectory being a closed curve which will be described repeatedly in the same period until the system is again disturbed. Such a closed integral curve is called a 'limit cycle' and represents a combination oscillation in which free and forced oscillations are present simultaneously. For example, if there were only the two frequency components the vector would be the resultant of a fixed vector representing the forced oscillation and a second vector of constant length rotating with the difference frequency, giving the free oscillation. Its end point would describe a circle with constant speed. Actually other combination frequencies are always present so that the situation is never as simple as this.

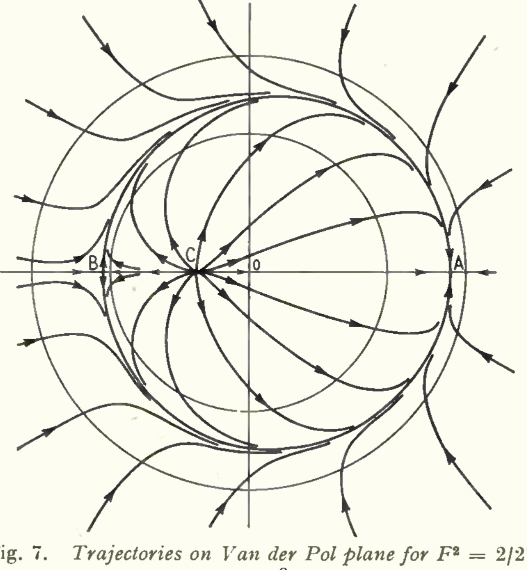

Fig. 7. *Trajectories on Van der Pol plane for $F^2 = 2/27$, $x = 0$.*

The first of the two equations shows that $db/d\tau$ is always negative if b is sufficiently large, so that a trajectory starting at a sufficient distance from the origin always moves inwards; i.e., an oscillation of large amplitude always decays.

The theory of such differential equations shows that other trajectories either approach a limit cycle or a stable singular point as $\tau \rightarrow \infty$, and when traced backwards through negative values of τ, arose at $\tau \rightarrow -\infty$ at an unstable singular point, from an unstable limit cycle, or moved in from infinity. Thus from any initial condition the system settles down ultimately either to a stable synchronized oscillation, or to a combination oscillation. The trajectories may be constructed by a graphical procedure for any particular values of the parameters. Some examples are shown in Figs. 7, 8 and 9. Fig. 7 shows the situation for a small value of F and $x = 0$ (i.e., E small and $\omega_1 = \omega_0$) when three stationary oscillations are possible. These correspond to the singular points A, B, C, and it will be seen that A is stable, C unstable, while B appears to be stable for variations of amplitude only, but when the phase is disturbed at all the vector at once commences to rotate away from the position OB, approaching finally the stable oscillation OA, in agreement with the analysis given earlier. It will be noticed that from arbitrary initial conditions there is first of all a rapid adjustment of amplitude to a value near to the stationary value, followed by a slower rotation of phase. Fig. 8 for the same F shows the stable synchronized oscillation about to give way to a combination oscillation. As the frequency moves further away from $x = 0$, A and B will move into coincidence at a phase angle very slightly beyond $-90°$, leaving a closed trajectory representing the combination oscillation. This limit cycle approximates to a circle of unit radius (i.e., amplitude equal to that of the free oscillation) but the cycle is not described uniformly with respect to the time. Near to the position at which A and B disappeared the velocity is small and the vector very nearly stationary, giving the appearance of a synchronized oscillation. When the vector has moved through this region it then moves more rapidly round the greater part of the cycle until it comes into this region again from the other side, giving the apparent slip from synchronization, followed again by an interval in which the oscillation appears to be synchronized. Fig. 9 for the same F shows the limit cycle just after the two singular points have disappeared.

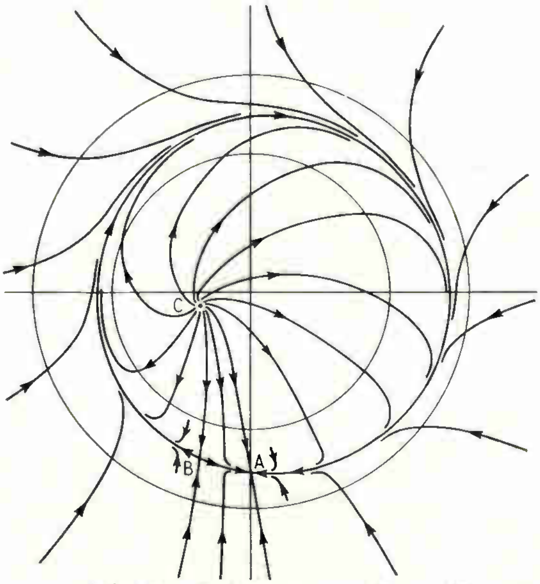

Fig. 8. *Trajectories on Van der Pol plane for $F^2 = 2/27$, $x = 0{\cdot}2722 = F$.*

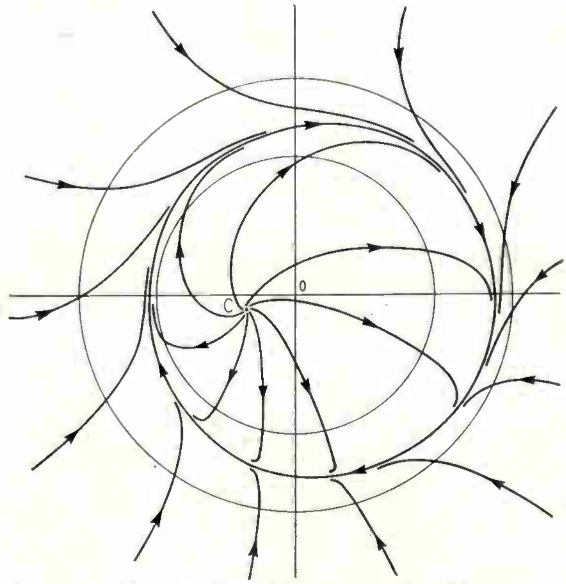

Fig. 9. *Trajectories on Van der Pol plane for $F^2 = 2/27$ $x = 0{\cdot}28$.*

16. The Frequency Expansion of the Combination Oscillation for Small F

The system of differential equations (13) and (14) is non-linear and an analytical solution is not obtainable by present methods. There is one case, however, in which an approximate integration may be carried out. If F is very small it is possible to define two circles, with centre at the origin, one with radius less than unity, the other with radius greater than unity, such that the smaller circle is always crossed by the solution curves in an outward direction, and the larger is always crossed inwards. Thus, once the end point has moved into the annulus between these circles, it remains there. Two such circles are shown in the Figs. 7, 8 and 9. This value of F is, however, large for the present argument. For a very much smaller F the radii of the two circles may be taken very near to unity, so that if the end point of the vector moves into the annulus, the length of the vector remains thereafter very near to $b = 1$. We may then assume $b \approx 1$. The first equation then determines a small fluctuation in the value of b which is now assumed to be negligible, and in the second equation we may put $b = 1$. Equation (14) then becomes

$$\frac{d\phi}{d\tau} = -F \sin \phi - x$$

This equation may be integrated[3] to give

$$\tan \frac{\phi}{2} = -\frac{F}{x} + \frac{\sqrt{(F^2 - x^2)}}{x} \tanh \tfrac{1}{2} \sqrt{(F^2 - x^2)}(\tau - \tau_0) \qquad (15)$$

if $|x| < F$; while if $|x| > F$

$$\tan \frac{\phi}{2} = -\frac{F}{x} - \frac{\sqrt{(x^2 - F^2)}}{x} \tan \tfrac{1}{2}\sqrt{(x^2 - F^2)}(\tau - \tau_0) \qquad (16)$$

In either case τ_0 represents a constant of integration which may be zero if the origin of time is suitably chosen.

If $|x| < F$ equation (15) applies, and as $\tau \to \infty$

$$\tanh \tfrac{1}{2}\sqrt{(F^2 - x^2)}(\tau - \tau_0) \to 1$$

and

$$\tan \frac{\phi}{2} \to -\frac{F}{x} + \frac{\sqrt{(F^2 - x^2)}}{x}$$

Thus ϕ approaches a definite limiting value and the vector a limiting position representing the synchronized oscillation. As x increases from zero this solution ceases to be valid when $x = F$ with $\tan (\phi/2) = -1$ or $\phi = -\pi/2$ and at this value of x the pull out from synchronization takes place.

Beyond this point the solution given by equation (16) becomes valid. When $\frac{1}{2}\sqrt{(x^2 - F^2)}\ (\tau - \tau_0)$ increases by π,

$$\left(\text{i.e., in time } \frac{2\pi}{\sqrt{(x^2 - F^2)}}\right)$$

the tangent on the right runs through all values between $-\infty$ and $+\infty$, the tangent on the left therefore does the same and, if x is positive, $\phi/2$ decreases by π (i.e., the vector makes a complete revolution in the negative direction) showing a combination oscillation with a free oscillation of frequency lower than that of the applied e.m.f. by the difference angular frequency $\sqrt{(x^2 - F^2)}$. When x is large compared to F this approximates to x, showing the free oscillation of frequency ω_0 on the original time scale. As $x \to F$, however, the difference frequency tends to zero, the period of rotation of the vector tending to infinity, showing that at synchronization the free oscillation frequency is brought into coincidence with that of the applied e.m.f.

The angular speed of the vector is not uniform, showing that other frequencies are present in the oscillation. When the factor $\sqrt{(x^2 - F^2)}/x$ is small we have $\tan (\phi/2) \approx -F/x$ except when the tangent on the right is passing through the infinite value at $\frac{1}{2}\sqrt{(x^2 - F^2)}(\tau - \tau_0) = \pi/2 + n\pi$. Thus the vector is nearly stationary at $\tan (\phi/2) = -F/x$ for most of the period, making a rapid rotation round the origin at the values of τ determined by the previous equation. On the other hand, when x is large compared to F the coefficient $\sqrt{(x^2 - F^2)}/x$ approximates to unity and the rotation of the vector approximates to a uniform angular speed. This would correspond to a free oscillation of amplitude near to $b = 1$ with the forced oscillation as well as other combination frequencies of negligible amplitude.

The vector of unit length at the angle ϕ may be represented by the complex number $1.e^{j\phi}$. This may be written

$$e^{j\phi} = \frac{e^{j\phi/2}}{e^{-j\phi/2}} = \frac{\cos (\phi/2) + j \sin (\phi/2)}{\cos (\phi/2) - j \sin (\phi/2)}$$

$$= \frac{1 + j \tan (\phi/2)}{1 - j \tan (\phi/2)}$$

On substituting for $\tan (\phi/2)$ from equation (16) and setting $\frac{1}{2}\sqrt{(x^2 - F^2)}(\tau - \tau_0) = \theta/2$ as an abbreviation, we have

$$e^{j\phi} = \frac{x - jF - j\sqrt{(x^2 - F^2)} \tan (\theta/2)}{x + jF + j\sqrt{(x^2 - F^2)} \tan (\theta/2)}$$

$$= \frac{(x - jF)\ (e^{j\theta/2} + e^{-j\theta/2}) - \sqrt{(x^2 - F^2)}\ (e^{j\theta/2} - e^{-j\theta/2})}{(x + jF)\ (e^{j\theta/2} + e^{-j\theta/2}) + \sqrt{(x^2 - F^2)}\ (e^{j\theta/2} - e^{-j\theta/2})}$$

$$= \frac{[x - \sqrt{(x^2 - F^2)} - jF] + [x + \sqrt{(x^2 - F^2)} - jF]e^{-j\theta}}{[x + \sqrt{(x^2 - F^2)} + jF] + [x - \sqrt{(x^2 - F^2)} + jF]e^{-j\theta}}$$

Let $$\frac{x - \sqrt{(x^2 - F^2)} + jF}{x + \sqrt{(x^2 - F^2)} + jF} = Ae^{-j\alpha}$$

Then $$\left|\frac{x - \sqrt{(x^2 - F^2)} + jF}{x + \sqrt{(x^2 - F^2)} + jF}\right| = A < 1$$

We now have

$$e^{j\phi} = \frac{x + \sqrt{(x^2 - F^2)} - jF}{x + \sqrt{(x^2 - F^2)} + jF} \times (Ae^{j\alpha} + e^{-j\theta})\ (1 + Ae^{-j(\theta+\alpha)})^{-1}$$

If we expand the last bracket by the binomial, multiply by the second factor and re-arrange the terms we obtain

$$e^{j\phi} = \frac{x + \sqrt{(x^2 - F^2)} - jF}{x + \sqrt{(x^2 - F^2)} + jF} A e^{j\alpha} \times$$
$$\left[1 + \frac{1 - A^2}{A} e^{-j(\theta + \alpha)} - (1 - A^2) e^{-2j(\theta + \alpha)} + A(1 - A^2) e^{-2j(\theta + \alpha)} - + \ldots\right]$$

The constant of integration may be chosen so that

$$\theta + \alpha = \sqrt{(x^2 - F^2)}\tau = \delta\omega . t$$

where $\delta\omega$ is the difference frequency on the original time scale. The factor $\dfrac{x + \sqrt{(x^2 - F^2)} - jF}{x + \sqrt{(x^2 - F^2)} + jF}$ of modulus unity may be written as $e^{-j\beta}$ so that we have

$$e^{j\phi} = A e^{j(\alpha - \beta)} \left[1 + \frac{1 - A^2}{A} e^{-j\delta\omega t} - (1 - A^2) e^{-2j\delta\omega t} + A(1 - A^2) e^{-3j\delta\omega t} - + \ldots\right]$$

Now

$$\begin{aligned} v &= V_1 \cos(\omega_1 t + \phi) \\ &= V_0 b \cos(\omega_1 t + \phi) \\ &= V_0 \cos(\omega_1 t + \phi) \end{aligned}$$

with $b \approx 1$.

thus

$$\begin{aligned} v &= \mathrm{Re}\, V_0 e^{j(\omega_1 t + \phi)} \\ &= \mathrm{Re}\, V_0 e^{j(\omega_1 t + \alpha - \beta)} [A + (1 - A^2) e^{-j\delta\omega_t} \\ &\quad - A(1 - A^2) e^{-2j\delta\omega t} + A^2(1 - A^2)^{-3j\delta\omega t} + \ldots] \\ &= V_0 A \cos\{\omega_1 t + \alpha - \beta\} + V_0(1 - A^2) \cos \\ &\qquad \{(\omega_1 - \delta\omega)t + \alpha - \beta\} \\ &\quad - V_0 A(1 - A^2) \cos\{(\omega_1 - 2\delta\omega)t + \alpha - \beta\} \\ &\quad + V_0 A^2(1 - A^2) \cos\{(\omega_1 - 3\delta\omega)t + \alpha - \beta\} \\ &\qquad - + \ldots \end{aligned}$$

which gives the resolution of v into its frequency components.

The first term is the forced oscillation; the second of frequency $\omega_1 - \delta\omega = \omega$ is the free oscillation; the third of frequency $\omega_1 - 2\delta\omega = 2\omega - \omega_1$, the fourth of frequency $\omega_1 - 3\delta\omega = 3\omega - 2\omega_1$ and the subsequent terms are combination terms with frequencies of the form $p\omega - (p - 1)\omega_1$. Moreover as $x \to F$ (i.e., as the frequency approaches the pull-in to synchronization) $A \to 1$ and the terms after the first do in fact approach equality of amplitude at a value which will be small by virtue of the factor $(1 - A^2)$. This is in agreement with the description given previously, with the difference that only those frequencies on the resonance side of ω_1 are involved. Had the small variation of amplitude been taken into account combination frequencies of the form $\omega_1 + q\delta\omega = (q + 1)\omega_1 - q\omega$ would have been introduced but with smaller amplitudes of the order F times those given above.

If x is large compared to F, the latter still being assumed small we shall have $A \approx \dfrac{F}{2x}$, $\alpha = \dfrac{\pi}{2}$ and $\beta = 0$ as limiting values, or neglecting terms involving F^2 we obtain

$$\begin{aligned} v &= \frac{V_0 F}{2x} \cos\left\{\omega_1 t + \frac{\pi}{2}\right\} + V_0 \cos\left\{(\omega_1 - \delta\omega)t + \frac{\pi}{2}\right\} \\ &\qquad - \frac{V_0 F}{2x} \cos\left\{(\omega_1 - 2\delta\omega)t + \frac{\pi}{2}\right\} \\ &= -\frac{V_0 F}{2x} \sin\omega_1 t - V_0 \sin\omega t + \frac{V_0 F}{2x} \sin(2\omega - \omega_1)t. \end{aligned}$$

The free oscillation is now large, approximating to the amplitude and frequency it would have with no applied e.m.f. Superposed on this is a small forced oscillation and a small combination term of frequency $2\omega - \omega_1$ on the other side of the free oscillation frequency, these being of approximately equal amplitudes.

17. Synchronization to Harmonics and Sub-Harmonics

The general discussion already given of the forced oscillation can easily be modified to account for the phenomena associated with the synchronization of the oscillator to harmonics or sub-harmonics of an applied e.m.f. These cases are, however, omitted from the present discussion.

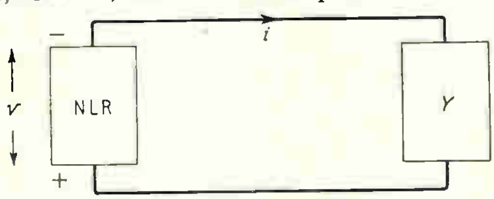

Fig. 10. *Circuit with general linear impedance and non-linear resistance.*

18. Oscillator Circuits of Other Types

The previous analysis has been set out with special reference to the circuit of Fig. 1, but it is applicable to other types of oscillator.

If the admittance of the linear circuit is denoted by Y the differential equation (1) may be written

$$i = f(v) = -Y(D)v \quad \ldots \quad \ldots \quad \ldots \quad (17)$$

No use has been made of the special form of the linear circuit in Fig. 1 other than the value of its admittance $1/R$ at resonance, so the analysis is equally applicable to the circuit of Fig. 10 in which Y is a general admittance.

In the circuit of Fig. 11 the linear network may be regarded as a four-terminal network with two terminals connected to the grid and filament and the other pair connected to anode and filament.

Assuming no grid current and using the short-circuit driving point and transfer impedances, the equations are

$$v_1 = z_{12}\, i_2;\ v_2 = z_{22}\, i_2$$

$$i_2 = f\left(v_1 + \frac{v_2}{\mu}\right)$$

in which v_1 and v_2 are the grid and anode potentials, i_2 the anode current, μ the amplification factor of the valve, and the third equation the characteristic.

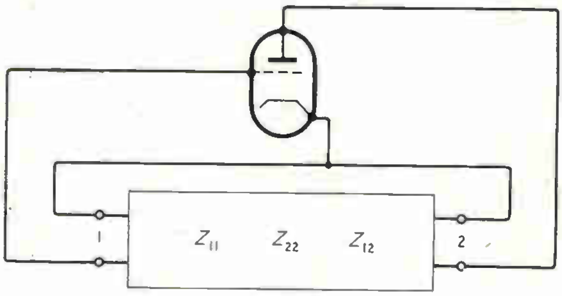

Fig. 11. *Triode circuit.*

$$\text{Putting } v = v_1 + \frac{v_2}{\mu} = \left\{z_{12} + \frac{z_{22}}{\mu}\right\} i_2$$

$$\text{we have } i = f(v) = -\ Y(D)v$$

$$\text{where } Y = -\frac{1}{z_{12} + z_{22}/\mu}$$

giving equations identical with equation (17) above.

If the valve is a pentode, for which we may assume the anode current dependent on v_1 only, then putting μ infinite we obtain

$$i_2 = f(v_1) = -\ Y(D)v_1$$

where now $Y = -1/z_{12}$

The analysis is therefore applicable with only a small re-interpretation of the symbols to triode or pentode oscillators provided no grid current flows.

19. Conclusion

In concluding it must be emphasized that this article does not pretend to give a detailed mathematical discussion. That has been given elsewhere.[5] The present purpose is merely to give a physical explanation of the phenomena of electrical oscillators which may contribute to the understanding of the physical processes involved and which gives a uniform approach to a variety of different aspects which have not previously been related. The physical ideas involved are also suggestive of the mathematical procedure for obtaining the solutions of some of the problems discussed.

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CORRESPONDENCE

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

Three-Probe Method of Impedance Measurement

SIR,—I read with interest the paper "Three-Probe Method of Impedance Measurement" by W. J. Duffin in the December 1952 issue of *Wireless Engineer*. Such a method of fixed probes, instead of a single sliding probe in a slotted waveguide in the microwave region (commonly known as standing-wave detector), is an improvement in the right direction because of its ability to measure impedance variations with sufficient accuracy and in much less time. Duffin has used the method in coaxial waveguide but his method can, of course, be used in rectangular waveguides as well. Practical difficulties will limit its use at wavelengths below 3 cm in both cases.

My suggestion is that, instead of having a probe as described by Duffin, we have a 'probe' consisting merely of a small hole coupling to a branch guide that is soldered to the main guide and terminated in a detector. The admittance of the probe of this type is only adjustable by tuning the branch guide.

In such a case the distortion of the field due to reflections from the three probes inside the waveguide is completely eliminated. The hole has the advantage of simplicity and of a predictable admittance.

PREM SWARUP.
Physics Department,
Allahabad University,
India.

2nd April 1953.

Networks with Maximally-Flat Delay

SIR—A recent paper concerning "Networks with Maximally-Flat Delay" (W. E. Thomson: *Wireless Engr*, Vol. 29, October 1952) describes an interesting relation between maximally-flat delay network and Gaussian impulse response. Dr. Thomson's work has an intimate relationship to our past research. The outline of our work, which is divided into two parts, is as follows.

First, Kiyasu worked out a general theory on the "Design of Delay Networks" (*The Journal of the Institute of Electrical Communication Engineers of Japan*, No.

245, August 1943) from the consideration of group propagation time. By applying this theory, he got a delay network of all-pass type having the Wagner characteristic (we use this terminology instead of maximally-flat characteristic), the network being exactly identical with that described in Thomson's previous paper (*Proc. Instn elect. Engrs*, Vol. 90, Part III, 1949).

Secondly, for the application of impulse technique, a need for applying Gaussian impulses also arose. For this purpose we designed a network whose attenuation characteristic is approximated to that of the Gaussian form, without regard to its phase property, that is, the attenuation factor $T(\omega)\ T(-\omega)$ is expressed in terms of a polynomial up to and including the $2n$-th term in the power series expansion of $\exp\{-(\omega/\omega_o)^2\}$. (The Joint Convention of the Institutes of Electrical Engineering of Japan and other two Institutes, May 1951). Thomson's network, on the other hand, is designed merely from the point of its phase characteristic. The exploitation of a proper basis of evaluation for the two cases is believed to be a problem yet to be solved.

KIYASU-ZEN'ITI,
NOBUITI IKENO,
SIGEHARU YAMADA.

Electrical Communication Laboratory,
Tokyo, Japan.
4th April 1953.

NEW BOOKS

Television Receiver Design: Monograph 1—I.F. Stages

By A. G. W. UITJENS. Pp. 177 + x. Cleaver-Hume Press, Ltd., 42a South Audley Street, London, W.1. Price 21s.

This is Book VIIIA in the series "Electronic Valves" in the Philips Technical Library. Its chapter headings are: 1. Gain and bandwidth with two-terminal coupling networks; 2. Response curve of the complete amplifier; 3. Distortion; 4. Gain, bandwidth and distortion with four-terminal coupling networks; 5 Noise; 6. Feedback; 7. Practical considerations following on the theory. There are five appendixes, four tables and a list of symbols, but there is no index.

In spite of its title, the book is not confined to intermediate-frequency amplification, but covers signal-frequency amplification as well. In fact, as pointed out in the preface, "This monograph deals particularly with pentode amplifiers operating in a frequency range lying roughly between 10 Mc/s and over 100 Mc/s."

Chapter 3 is probably the most important one, because in it there is given an approximate method of computing the transient response from the frequency response, and ways of doing this are not at all widely known. In the rest of the book the material is, as is only natural, of a more familiar nature. Taken as a whole, the book contains a great deal of information which is necessary to the designer, but it does not contain everything.

In Chapter 4, the four-terminal networks are limited to pairs of coupled circuits and there is no mention of the three- and four-circuit couplings which are sometimes used. Of more importance is the inadequate way in which matters pertaining to the sound channel are treated. The use of a rejector circuit coupled to a two-terminal coupling circuit is actually dealt with much more fully than appears at first sight. This is because it is discussed in terms of a coupled pair of circuits and most of the information appears in an appendix. There is no mention of sound pick-out circuits, cathode rejectors or of bridged-T rejectors.

The mathematical part of the book is, in the main, simple and straightforward. The text is made unnecessarily difficult because of the author's habit of using in it mathematical symbols as a kind of shorthand. While this is sometimes legitimate and is sometimes an aid to clarity, it is only so when the reader can easily remember the meaning of the symbols. When, as in this book, the reader has continually to turn back, or refer to the list of symbols in order to elucidate their meaning, this kind of shorthand is nothing but a hindrance. It means that one cannot obtain proper value from the book until one has become thoroughly soaked in the author's symbolism. Matters are in no way helped by this being partly non-standard; thus, S is used for mutual conductance instead of g_m, and g_c is not conversion conductance, as one would expect it to be, but the conductance of a coupling circuit.

W. T. C.

Einführung in die Mikrowellen-Elektronik.—Part 1.

By DR. WERNER KLEEN. Pp. 300 + xix with 129 illustrations. S. Hirzel Verlag, Zürich. Price 30 Fr. (Swiss)

The author is now attached to the valve research laboratories of Siemens and Halske at Munich, but this book is the outcome of lectures given by him in 1950 and 1951 in Madrid and Stockholm, and is largely based on his previous experience in the research laboratories of the Compagnie Générale de Télégraphie sans Fil, Paris. The book is published in Switzerland but printed in Holland; it is therefore quite an international production. It is to be in three parts of which the volume under review is Part 1, Fundamentals; Parts II and III will deal with grid-controlled valves, klystrons and travelling-wave tubes. The reader is expected to have a knowledge of mathematics of a final year standard, also of the fundamentals of valves and high-frequency transmission; it also says that the reader is expected to work in co-operation with the author. This soon becomes necessary, for on page 2 it is stated that $dW/dt = -\delta U/\delta t$ where W is the energy of the electron and U the potential; the reader can co-operate with the author by inserting the charge of the electron e before $\delta U/\delta t$ and thus making sense of the equation.

The book is divided into 19 chapters, and most chapters conclude with a very full list of references, sometimes as many as 30 or 40; in the preface the author apologises for the small number of German references. The book deals primarily with wavelengths between 1 m and 1 cm. The system of units employed is the Mie system (i.e., the V, A, cm, s, system) in which the unit of mass is 10 tons; he departs from this system in the case of magnetic induction B which is given in gauss. Another somewhat disturbing departure from standard practice is that the direction of electric field and current is taken as the direction of the force on an electron so that $\vec{E} = dU/\vec{ds}$ without a minus sign. In order that the fundamental equations of electromagnetism such as the Maxwell equations may retain their usual signs, the author reverses the direction of the magnetic field H and assumes that inside a magnet the flux goes from the north to the south pole. This is the penalty that we have to pay for the negative charge of the electron.

The first two chapters deal with the movement of electrons in static fields, chapter 3 with alternating fields and travelling waves, and chapter 4 with the resulting external current. The next four chapters deal with velocity modulation, phase focusing, the klystron, and the complex conductance of space-charge tubes. The subjects are developed in the following chapters on the lines of the article "Electron Beams and Electromagnetic Waves" contributed by the author and two of his colleagues to *Wireless Engineer* of June 1951. Then follow chapters on the generation and amplification of waves under crossed fields, the klystron, magnetron, and other types of tubes. These are all described with photographs and their applications discussed. The tubes are then discussed as circuit elements, including noise, bandwidth, available power, etc. A whole chapter is then devoted to noise and 43 references given to papers on the subject; this is followed by a chapter on microwave resonators. A chapter entitled Retarding Lines deals with waveguides of different types, corrugated, spiral, disc-loaded, etc., and the final chapter on electron guns discusses various methods of focusing the beam.

After the table of contents at the beginning seven pages are devoted to classified lists of the symbols and units employed. The treatment throughout is very thorough and all the mathematical results are clearly illustrated by means of curves. To those with the necessary knowledge of German this book will form a very useful post-graduate introduction to microwave electronics.

G. W. O. H.

Radio Designers' Handbook (4th Edition)

By F. Langford-Smith, B.Sc., B.E., Sen.M.I.R.E., A.M.I.E.(Aust.). Pp. 1,474. Iliffe & Sons, Ltd., Dorset House, Stamford Street, London, S.E.1. Price 42s. (postage 1s. 6d.).

This new edition is the work of 10 authors with 23 collaborating engineers under the editorship of F. Langford-Smith. It covers the receiving side of wireless equipment with special emphasis on broadcast receivers and audio-frequency apparatus. Summaries of both British and American component specifications are included.

Principles of Radar (Third Edition)

By Members of the Staff of the Radar School, Massachusetts Institute of Technology. 3rd Edition by J. Francis Reintjes and Godfrey T. Coate. Pp. 985 + xv. McGraw-Hill Publishing Co., Ltd., 95 Farringdon St., London, E.C.4. Price 55s. 6d.

This is a greatly-revised edition of a book which was originally produced offset from typescript; it is now in normal letterpress.

Essentials of Microwaves

By Robert B. Muchmore. Pp. 236 + vi. Chapman & Hall, Ltd., 37 Essex Street, London, W.C.2.

STANDARD-FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)

From 26th May 1953 the schedule of British Standard-Frequency Transmissions has been changed. The transmissions from MSF at Rugby now cover 24 hours a day and take place on 5 Mc/s and 10 Mc/s, with a power of 0·5 kW. The transmissions will be interrupted from 15 to 20 minutes past each hour so that one station alone can be measured under conditions where, for example, WWV is received at similar strength. Initially, there is also a transmission on 2·5 Mc/s and later, 15 Mc/s and 20 Mc/s may also be used, but no more than three frequencies simultaneously.

The modulation is in accordance with the schedule of Table 1. This applies also to the 60-kc/s transmissions with a power of 10 kW which are still being made to serve this country. However, from this station, transmission is carried out from 1429 to 1530 G.M.T. only.

Details of the modulation and general characteristics of the signal were given in the March 1952 issue of *Wireless Engineer*, p. 82. The results of measurements on the transmissions during April 1953 are given in Table 2.

TABLE 1

Minutes past each hour	Modulation
0– 5, 15–20, 30–35, 45–50	1000 c/s
5–10, 20–25, 35–40, 50–55	1-c/s pulses, 59th of each minute omitted
10–14, 25–29, 40–44, 55–59	Unmodulated
14–15, 29–30, 44–45, 59–60	Speech

TABLE 2

Date 1953 April	Frequency deviation from nominal: parts in 10^8 — MSF 60 kc/s 1029-1130 G.M.T.	Droitwich 200 kc/s 1030 G.M.T.	Lead of MSF impulses on GBR 1000 G.M.T. time signal in milliseconds
1	+ 0·5	0	+ 49·8
2	+ 0·4	0	+ 50·3
3*	+ 0·4	− 3	+ 49·0
4	+ 0·4	− 3	+ 48·5
5	+ 0·4	0	+ 48·4
6	+ 0·5	− 3	+ 46·7
7	+ 0·5	+ 1	+ 48·0
8	+ 0·5	− 3	+ 45·9
9**	—	− 3	—
10	+ 0·4	− 1	+ 45·9
11	+ 0·4	− 1	+ 45·0
12	+ 0·5	0	+ 45·6
13	+ 0·5	0	+ 43·9
14	+ 0·5	0	+ 43·8
15	+ 0·5	− 1	+ 43·4
16	+ 0·4	− 1	+ 41·6
17	+ 0·4	− 1	+ 41·1
18	+ 0·5	0	+ 40·1
19	+ 0·5	0	+ 39·4
20	+ 0·5	0	+ 38·2
21*	+ 0·5	0	+ 37·7
22*	+ 0·5	+ 1	+ 36·0
23*	+ 0·6	0	+ 35·3
24	+ 0·5	0	+ 34·5
25	+ 0·6	+ 1	+ 34.2
26	+ 0·6	+ 2	+ 34·0
27	+ 0·7	+ 2	+ 33·8
28	+ 0·6	+ 2	+ 33·4
29	+ 0·6	+ 2	+ 33·3
30	+ 0·7	+ 2	+ 32·8

The values are based on astronomical data available on 1st May 1953.

The transmitter employed for the 60-kc/s signal is sometimes required for another service.

* = No MSF Transmission at 1029 G.M.T. Results for 1429-1530 G.M.T.

** = No MSF Transmission at 1029 or 1429 G.M.T.

ABSTRACTS and REFERENCES

Compiled by the Radio Research Organization of the Department of Scientific and Industrial Research and published by arrangement with that Department.

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. U.D.C. numbers marked with a dagger (†) must be regarded as provisional. The abbreviations of journal titles conform generally with the style of the World List of Scientific Periodicals. An Author and Subject Index to the abstracts is published annually; it includes a list of journals abstracted, the abbreviations of their titles and their publishers' addresses.

ACOUSTICS AND AUDIO FREQUENCIES

016 : 534 1549

References to Contemporary Papers on Acoustics.—R. T. Beyer. (*J. acoust. Soc. Amer.*, Jan. 1953, Vol. 25, No. 1, pp. 164–168.) Continuation of 910 of April.

534.1 + 621.396.615] : 621.3.016.37 1550

Electrical and Acoustic Oscillation Build-Up Phenomena.—Dänzer. (See 1626.)

534.231 : 532.517 1551

Acoustic Streaming due to Attenuated Plane Waves.—W. L. Nyborg. (*J. acoust. Soc. Amer.*, Jan. 1953, Vol. 25, No. 1, pp. 68–75.)

534.231 : 532.527 1552

The Theory of Steady Rotational Flow Generated by a Sound Field.—P. J. Westervelt. (*J. acoust. Soc. Amer.*, Jan. 1953, Vol. 25, No. 1, pp. 60–67.)

534.231.1 : 621.3.018.7 1553

Nonlinear Dissipative Distortion of Progressive Sound Waves at Moderate Amplitudes.—J. S. Mendousse. (*J. acoust. Soc. Amer.*, Jan. 1953, Vol. 25, No. 1, pp. 51–54.)

534.231.3 1554

The Physical Interpretation of the Expression for an Outgoing Wave in Cylindrical Coordinates.—M. C. Junger. (*J. acoust. Soc. Amer.*, Jan. 1953, Vol. 25, No. 1, pp. 40–47.) The sound field generated by a vibrating cylinder of infinite length is expressed in terms of acoustic impedance ratios. The effect of variation of wavelength on the propagation of the various possible modes is examined. Graphs of the impedance ratios corresponding to certain modes are presented.

534.232 1555

Effect of a Finite Circular Baffle Board on Acoustic Radiation.—T. Nimura & Y. Watanabe. (*J. acoust. Soc. Amer.*, Jan. 1953, Vol. 25, No. 1, pp. 76–80.) The oblate-spheroidal wave function developed by Kotani and by Bouwkamp is used to calculate the effect of a circular baffle on the radiation field of a circular disk. Results are shown graphically.

534.232 : 534.321.9 : 538.652 1556

Critical Consideration of Ultrasonic-Wave Generators using Linear-Magnetostrictive Dumb-Bell-Type Oscillators.—H. H. Rust & E. Bailitis. (*Akust. Beihefte*, 1952, No. 2, pp. 89–90.) In this type of generator the amplitude of the longitudinal oscillations suffers reduction as a result of the transverse effect. For Ni oscillators of usual dimensions, off-resonance amplitudes are less than half the values to be expected from the magnetostriction curve, taking account of the reduction of amplitude due to remanence. An arrangement is indicated in which the effect is compensated by using Ni laminations for the exciting bar and Fe-Ni laminations for the radiating pistons.

534.26 1557

Diffraction of Sound Waves by a Circular Aperture in a Rigid Screen.—H. Severin & C. Starke. (*Akust. Beihefte*, 1952, No. 2, pp. 59–66.) Measurements were made of the sound pressure along the axis behind the aperture and in the plane of the screen, using apertures of radius $0{\cdot}5\lambda$, λ, $1{\cdot}5\lambda$, 2λ, 3λ and 4λ. The results are compared with two approximate solutions calculated from Kirchhoff's theory. The observed distribution agrees best with theoretical values obtained by using a velocity potential derived on the assumption that the aperture area is covered with radiating dipoles. Corresponding work on e.m. waves is noted in 2412 of 1951 (Severin).

534.321.9 1558

Interference of Ultrasonic Beams [incident on and] **reflected by a Plane Wall. Conduction of Waves.**—F. Canac. (*C. R. Acad. Sci., Paris*, 26th Jan. 1953, Vol. 236, No. 4, pp. 360–362.) Shadow patterns obtained by a stroboscopic method show the wave formation for different angles of incidence. An analogy with waveguide propagation is illustrated using two parallel reflectors.

534.321.9 1559

Velocity of Ultrasonic Waves in Argon at Pressures up to 950 Atm.—A. Lacam & J. Noury. (*C. R. Acad. Sci., Paris*, 26th Jan. 1953, Vol. 236, No. 4, pp. 362–364.)

Results of measurements by a light-diffraction method at about 900 kc/s show that the velocity increases nearly linearly with pressure over the range 250–650 atm.

534.321.9 : 532.528 **1560**

Methods for investigating Oscillation Cavitation in Liquids by means of Ultrasonics.—T. Lange. (*Akust. Beihefte*, 1952, No. 2, pp. 75–82.)

534.321.9 : 534.614 **1561**

Ultrasonic Measurements.—A. Giacomini. (*Ricerca sci.*, Jan. 1953, Vol. 23, No. 1, pp. 75–80.) A pulse method for measuring the velocity of ultrasonic waves in solids is based on a comparison between the unknown velocity and the velocity in water.

534.321.9 : 534.833.4 **1562**

Investigations of Ultrasonic Absorption in Animal Tissues and in Plastics.—R. Esche. (*Akust. Beihefte*, 1952, No. 2, pp. 71–74.) A method is described for measurements in the range 300–600 kc/s. The absorption coefficient is found to increase linearly with frequency, as at higher frequencies.

534.321.9-14 **1563**

The Behaviour of Ultrasonics in Liquids.—J. Lamb. (*Research, Lond.*, Dec. 1952, Vol. 5, No. 12, pp. 553–560.) The results of recent research are reviewed, with particular reference to the measurement and physical interpretation of relaxation effects in pure liquids, electrolytes and polymer liquids. About 50 references.

534.414 **1564**

The Effects of Viscous Dissipation in the Spherical Acoustic Resonator.—H. G. Ferris. (*J. acoust. Soc. Amer.*, Jan. 1953, Vol. 25, No. 1, pp. 47–50.) Analysis shows that the effect of viscosity is to decrease slightly the natural frequencies and to cause decay of the free vibrations. The attenuation of the purely radial modes of vibration is much less than that for modes with tangential velocity components.

534.613 : 534.37 **1565**

Radiation Pressure and Dispersion.—E. J. Post. (*J. acoust. Soc. Amer.*, Jan. 1953, Vol. 25, No. 1, pp. 55–60.) Discussion of basic theories relative to propagation in a dispersive medium, with particular reference to boundary conditions and the essential difference between acoustic and electromagnetic radiation pressure. An extensive annotated bibliography, dating back to 1870, is given.

534.614 **1566**

Precision Measurement of the Velocity of Sound in Air.—P. W. Smith, Jr. (*J. acoust. Soc. Amer.*, Jan. 1953, Vol. 25, No. 1, pp. 81–86.) A method is described in which the driving-point impedance of a loudspeaker connected to a closed tube of adjustable length serves as the frequency-controlling element of a bridge-stabilized oscillator. Results of measurements at 1 kc/s in air, corrected to standard conditions, give a velocity of $331.45 \pm 0{\cdot}05$ m/s.

534.62 **1567**

Investigations for improving the Linings of Acoustically Damped Rooms.—G. Kurtze. (*Akust. Beihefte*, 1952, No. 2, pp. 104–107.) By arranging Helmholtz-type resonators behind absorbing wedges, the same amount of absorption is achieved with a reduction of one-third in the thickness of the lining and without raising the lower frequency limit.

534.78 : 681.613 **1568**

The Phonetic Typo-sonograph or Phonetograph.—J. Dreyfus-Graf. (*Tech. Mitt. schweiz. Telegr.-Teleph-Verw.*, 1st Dec. 1952, Vol. 30, No. 12, pp. 363–379. In French.) Developments of the steno-sonograph (271 and 1330 of 1950) are described. The present models give a phonetic transcript of speech in the form of known alphabetic symbols by means of an electric typewriter. The speech is transformed into groups of nine pulses which are compared two-by-two in an electrical computer operating on the trinary system (0 + –). A system of lamps associated with the computer enables conversation to be followed by the deaf. The width of the frequency channel necessary for transmitting phonetograms is 270 c/s according to Fourier theory, i.e. 135 c/s in practice.

534.833 **1569**

Wide-Band Absorbers for Liquid-Borne Sound.—E. Meyer & K. Tamm. (*Akust. Beihefte*, 1952, No. 2, pp. 91–104.) Description, with theory and performance figures, of absorbers for the frequency range 5–50 kc/s, suitable for lining measuring tanks. Porous elastic-rubber materials are used and the devices have ribs arranged to form parallel ducts whose walls are sloped for impedance matching.

534.844/.845 **1570**

Comparative Absorption-Coefficient Measurements 1950.—A. Eisenberg. (*Akust. Beihefte*, 1952, No. 2, pp. 108–114.) Measurements previously reported by Meyer & Schoch (2412 of 1939) did not agree sufficiently well to serve as a basis for a test specification. Further measurements are now reported and a tentative test specification is presented for the determination of absorption coefficients by the reverberation-chamber method.

534.844.1 **1571**

The Accuracy of Reverberation-Time Measurements with Warble Tones and White Noise.—W. Furrer & A. Lauber. (*J. acoust. Soc. Amer.*, Jan. 1953, Vol. 25, No. 1, pp. 90–91.) Investigations in three rooms of very different volumes show that the probable error of reverberation-time measurements is less with warble tones than with white noise at 200 c/s, the reverse being the case at 800 and 3 200 c/s.

534.846.4 **1572**

Improvement of Audibility in a Theatre by means of a Sound-Delaying 'Quiet-Speaker' System.—G. R. Schodder, F. K. Schröder & R. Thiele. (*Akust. Beihefte*, 1952, No. 2, pp. 115–116.) Account of experiments made inside a hall at Recklinghausen; the technique was similar to that used by Parkin & Scholes (1294 of 1951) in their investigations of an open-air theatre.

534.862 : 534.874.1 **1573**

Acoustic Problems at the "Waldbühne" Open-Air Sound Theater in Berlin.—H. Simon. (*J. Soc. Mot. Pict. Telev. Engrs*, Dec. 1952, Vol. 59, No. 6, pp. 512–516.) General arrangements for vision-sound synchronization for film reproduction are discussed. Uniform sound volume for an audience of 25 000 is achieved by careful adjustment of the beam direction of loudspeaker units at each side of the screen, each unit comprising two groups of three loudspeakers with different frequency characteristics. The calculated amplifier power required to give the required audibility over a wide frequency range is about 120 W, the acoustic power being 8–10 W.

534.874.1 : 621.395.61 **1574**

Methods of Directional Sound Reception.—H. Grosskopf. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, Nov./Dec. 1952, Vol. 4, Nos. 11/12, pp. 209–218.) Various methods for the directional reception of speech and music are reviewed, including those using (*a*) gradient microphones of the first, second or higher order, either singly or in groups, (*b*) microphones utilizing the interference of sound waves, some using reflectors, others groups of receivers, as in the tubular and line microphones of Mason & Marshall (1518 of 1939) and Olson (3600 of 1939). Figure 2 in the article should be rotated through 180°.

621.395.625.3 **1575**

The Field of Harmonically Magnetized Recording Tape. Playback under Open-Circuit Conditions with Ideal and with Very-Wide-Gap Head.—O. Schmidbauer. (*Frequenz*, Oct. & Nov. 1952, Vol. 6, Nos. 10 & 11, pp. 281–290 & 319–324.)

621.395.625.3 **1576**

Performance of High-Output Magnetic Tape.—L. B. Lueck & W. W. Wetzel. (*Electronics*, March 1953, Vol. 26, No. 3, pp. 131–133.) Performance figures are given for a new recording tape, to be known as Scotch Brand No. 120, which will give a 6-db greater output than the standard Scotch Brand No. 111 without rise of noise level. Remanence is increased by a factor of 2·2 compared with No. 111 tape, and the new tape has considerably lower distortion, so that greater signal voltages can be used without the distortion exceeding permissible limits.

621.395.625.3 + 534.852] : 061.3 **1577**

U.E.R. [Union Européenne de Radiodiffusion] **Conference on Magnetic Sound Recording, Hamburg, 18th–22nd November 1952.**—K. E. Gondesen. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, Nov./Dec. 1952, Vol. 4, Nos. 11/12, p. 230.)

AERIALS AND TRANSMISSION LINES

621.315.212 **1578**

Investigations on a Broadcast-Programme Line using Star-Quad Cable without Coil Loading.—E. A. Pavel & H. v. Schau. (*Fernmeldetech. Z.*, Dec. 1952, Vol. 5, No. 12, pp. 551–562.) Extensive measurements at frequencies up to 15 kc/s are reported on the 'high-quality' cable running from Hamburg to Frankfort and back, a total distance of 1 072 km, and incorporating 33 line amplifiers. A difference between input and output quality could be detected only after a repetition process corresponding to transmission over a cable length of about 5 000 km.

621.392 **1579**

The Input Impedance of a Line, and the Location of Nonuniformities.—R. Monelli. (*Alta Frequenza*, Dec. 1952, Vol. 21, No. 6, pp. 260–287.) Formulae are derived for the input resistance and reactance of a very long line with purely resistive characteristic and terminating impedances. A special function is introduced expressing the relation between attenuation and terminating-impedance mismatch. The information obtainable from r.f.-bridge impedance measurements is discussed.

621.392.09 **1580**

Theory of the Single-Wire Transmission Line.—T. E. Roberts, Jr. (*J. appl. Phys.*, Jan. 1953, Vol. 24, No. 1, pp. 57–67.) Equations are derived for the current induced in an infinitely long straight wire of finite surface impedance, connected to a flanged coaxial line. Radiation patterns are computed and the input conductance is determined. The current comprises two components, one of which is a modal current. Graphs are given for computing the efficiency of excitation of this component. A second system comprising a single wire between perfectly conducting parallel plates is investigated as a boundary-value problem; the solution is used in the discussion of the physical properties of a finite-length single-wire transmission line. Usual transmission-line concepts are applicable under certain restricting conditions.

621.392.22 : 517.9 **1581**

Applications of the Theory of Systems of Differential Equations to Multiple Nonuniform Transmission Lines.—R. L. Sternberg & H. Kaufman. (*J. Math. Phys.*, Jan. 1953, Vol. 31, No. 4, pp. 244–252.)

621.392.26 **1582**

Theory of Propagation in Waveguides.—E. Ledinegg & P. Urban. (*Acta phys. austriaca*, Nov. 1951, Vol. 5, No. 1, pp. 1–11.) See 2433 of 1952.

621.392.26 **1583**

An Exact Formulation of the Problem of Diffraction by Diaphragms with Slits in Waveguides with Rectangular Cross-Section.—R. Müller. (*Z. angew. Phys.*, Nov. 1952, Vol. 4, No. 11, pp. 424–433.) From results previously reported (625 of March) for waveguides of arbitrary cross-section, formulae are derived for the case of a diaphragm, with a symmetrically placed slit, in a rectangular waveguide. The original two-dimensional problem turns into a series of one-dimensional problems. In the absence of an incident wave a field can exist in the waveguide only for some particular values of wave number. Conditions for the existence of various modes are investigated and the singular behaviour of H_{0n} waves is discussed. Analysis is given for the case of an incident H_{10} or H_{01} wave at a frequency such that it is the only homogeneous wave of its type of symmetry. For the H_{10} wave the solution is given by a symmetrical Fredholm integral equation of the first kind, for the H_{01} wave by a pair of such equations.

621.392.26 **1584**

Waveguide Attenuation and its Correlation with Surface Roughness.—F. A. Benson. (*Proc. Instn elect. Engrs*, Part III, March 1953, Vol. 100, No. 64, pp. 85–90.) The attenuation of 9·375-kMc/s waves in samples of standard commercial waveguide was determined by a method involving examination of the standing-wave pattern near a voltage minimum. Detailed studies were made of the roughness of the internal surfaces of the waveguides. Discrepancies between measured and calculated values of attenuation are attributed to surface roughness, though in some cases very high values of measured attenuation are attributable to high resistivity of the waveguide material. One particular sample of 'copper' waveguide, with a d.c. resistivity nearly three times that of pure Cu, was analysed and found to be made from deoxidized arsenical copper. Effects of small changes in waveguide dimensions on the attenuation are estimated and a modification of Kuhn's attenuation formula (1328 of 1947) is given which takes account of surface roughness.

621.392.26 **1585**

Low-Loss Waveguide Transmission.—S. E. Miller & A. C. Beck. (*Proc. Inst. Radio Engrs*, March 1953, Vol. 41, No. 3, pp. 348–358.) Report of experiments on long-distance transmission, using frequencies around 9 kMc/s with circular waveguide of diameter about 5 in. and propagating the TE_{01} mode, the theoretical loss being 2 db/mile. 0·1-μs pulses were transmitted over a distance of 40 miles. Observed losses averaged about 50% more than the theoretical values, as a result of surface roughness and mode conversion. Compared with dominant-mode transmission, this multimode system has the advantages of increased bandwidth, reduced delay distortion and reduced size, and the disadvantage of reduced mechanical tolerances. Mode filtering devices are discussed. See also 2992 of 1952 (King).

621.392.26 : 538.221 **1586**
Modes in Waveguides containing Ferrites.—M. L. Kales. (*Trans. Inst. Radio Engrs*, Dec. 1952, No. PGAP-4, pp. 104–105.) Analysis shows that in a cylindrical waveguide filled with a homogeneous ferrite material subjected to a constant axial magnetic field no TE, TM or TEM modes are possible.

621.396.67 **1587**
Radiation from a Horizontal Dipole in a Semi-infinite Dissipative Medium.—R. H. Lien. (*J. appl. Phys.*, Jan. 1953, Vol. 24, No. 1, pp. 1–4.) Formulae are derived for the radiation field. Solutions are obtained in the form of complex integrals which can be evaluated in closed form for low frequencies.

621.396.67 **1588**
The Radiation Resistance of a Straight Wire in Free Space.—R. Müller. (*Arch. elekt. Übertragung*, Jan. 1953, Vol. 7, No. 1, pp. 56–57.) The function expressing the dependence of the radiation resistance on the ratio of wire length to wavelength is shown to have a periodic component superimposed on the monotonically increasing term.

621.396.67 : 621.397.6 **1589**
Multiple Television and Frequency-Modulation Transmitting Antenna Installation on the Empire State Building.—J. B. Dearing, H. E. Gihring, R. F. Guy & F. G. Kear. (*Proc. Inst. Radio Engrs*, March 1953, Vol. 41, No. 3, pp. 324–337.) Account of the development and performance of the installation, which comprises a steel tower supporting six aerial systems, radiating altogether five picture carriers, five sound carriers and three f.m. carriers. See also 1568 of 1951 (Kear & Hanson).

621.396.67 : 621.397.62 **1590**
Television Aerials.—G. W. Luscombe. (*Wireless Engr*, April 1953, Vol. 30, No. 4, pp. 82–90.) An investigation of receiving aerials. The most important characteristics determining performance are the gain, the radiation pattern and the impedance. Methods of measuring these properties are described, the precautions necessary to ensure good accuracy being indicated. Results obtained with nine outdoor aerials of British manufacture are presented; measurements were made at vision-carrier and sound-carrier frequencies.

621.396.67.013.24 **1591**
Impulse Electromagnetic Fields.—J. A. Feyer. (*Trans. S. Afr. Inst. elect. Engrs*, Oct. 1952, Vol. 43, Part 10, p. 291.) An error in Kitai's paper (38 of January) is pointed out. At distances $>\lambda/2\pi$ the radiation field predominates over the induction field in respect of both intensity and the spectrum function at a given frequency.

621.396.677 **1592**
Further Effects of Manufacturing Tolerances on the Performance of Linear Shunt Slot Arrays.—H. F. O'Neill & L. L. Bailin. (*Trans. Inst. Radio Engrs*, Dec. 1952, No. PGAP-4, pp. 93–102.) Calculations are made of the increase of side-lobe level due to specified deviations from the optimum array parameters, thus providing data of service in fixing manufacturing tolerances. Curves are given to facilitate the design of Dolph-Tchebycheff-type arrays with 12, 24 or 48 elements. See also 1241 of May (Bailin & Ehrlich).

621.396.677 **1593**
A Helical Beam [aerial] **for Citizens' Radio.**—E. F. Harris. (*Electronics*, March 1953, Vol. 26, No. 3, pp. 134–135.) Description, with radiation diagrams, of a 6-turn helical aerial of ⅜-in. Cu braid embedded in a fiberglas moulding of length 29 in. and diameter 6·5 in. With a 16 × 16-in. Al ground plate, good directivity is obtained from 390 to 600 Mc/s. With four such elements on a common ground plate at each end of a radio link, the gain is increased by 12 db compared with the use of single units. See also 3511 of 1948 (Lurie) and back references.

621.396.677 **1594**
Optimum Design of Linear Arrays in the Presence of Random Errors.—D. Ashmead. (*Trans. Inst. Radio Engrs*, Dec. 1952, No. PGAP-4, pp. 81–92.) Assuming the maximum error in the current at any radiator element is a constant fraction of the designed current, a method is given for finding the mean side-lobe suppression that can be obtained for any Tchebycheff distribution. Over-correction in the design may result in less suppression and lower gain. A method is also given for finding the aperture efficiency of a Tchebycheff distribution without calculating the individual currents. See also 1241 of May (Bailin & Ehrlich).

621.396.677 **1595**
Lattice Lenses for Centimetre Waves.—J. Moussiegt. (*Onde élect.*, Dec. 1952, Vol. 32, No. 309, pp. 515–518.) See 3004 and 3355 of 1952.

621.396.677.029.6 **1596**
On Spherically Symmetric Lenses.—J. E. Eaton. (*Trans. Inst. Radio Engrs*, Dec. 1952, No. PGAP-4, pp. 66–71.) Generalization of Luneberg's spherical lens, with a solution of the integral equation for the ray paths.

621.396.677.029.6 **1597**
Astigmatic Diffraction Effects in Microwave Lenses.—A. S. Dunbar. (*Trans. Inst. Radio Engrs*, Dec. 1952, No. PGAP-4, pp. 72–80.) Discussion of the effects of astigmatism on the diffraction pattern and gain of a lens aerial.

CIRCUITS AND CIRCUIT ELEMENTS

621.314.22.015.7 **1598**
Design of Series Peaking Transformers.—A. C. Sim. (*Wireless Engr*, April 1953, Vol. 30, No. 4, pp. 90–93.) Essential features of this type of transformer are (*a*) the load current is arranged to be negligible compared with the magnetizing current, and (*b*) the magnetizing current is allowed to reach a value much greater than that required to saturate the core. An easy and accurate design method is presented which does not require an analytical representation of the B/H characteristic. The example of a transformer to produce a voltage pulse of about 100 V for triggering a thyratron is worked out numerically. The voltage and duration of the pulse can be predicted to within about 20% of the measured values.

621.314.222 **1599**
Investigations on Voltage Transformers.—H. Kafka. (*Elektrotech. u. Maschinenb.*, 1st Dec. 1952, Vol. 69, No. 23, pp. 523–532.) A conductance diagram is developed to show directly the variation of secondary voltage and phase angle as a function of the load. Its application in transformer design is illustrated.

621.314.5 : 537.525.92 **1600**
D.C./A.C. Converter utilizing the Space-Charge Capacitance of a Valve.—H. Wechsung. (*Frequenz*, Nov. 1952, Vol. 6, No. 11, pp. 336–337.) Two circuits are described for converting feeble d.c. to a.c. prior to

amplification for measurement purposes. Both make use of the variation of grid-cathode capacitance with the voltage applied to the grid, the first circuit using a bridge balanced for a certain value of the grid-cathode capacitance, so that any variation produces a proportionate deflection of a meter. The second circuit uses a balanced arrangement of two valves with a differential transformer. Zero stability is good in both cases.

621.314.7(083.3) **1601**

Transistor Equations.—F. R. Stansel. (*Electronics*, March 1953, Vol. 26, No. 3, pp. 156, 158.) Formulae giving circuit voltage and current gain, and impedance characteristics, in terms of transistor parameters are tabulated for grounded-base, grounded-emitter, and grounded-collector transistor amplifier circuits.

621.318.57 **1602**

Static-Dynamic Design of Flip-Flop Circuits.—C. L. Wanlass. (*Trans. Inst. Radio Engrs*, Dec. 1952, No. PGEC-1, pp. 6-18.) Discussion of the design of flip-flop circuits for operation from a relatively low supply voltage, a desirable feature for computer elements. The circuit adopted uses negative-pulse triggering and a diode input circuit, the minimum trigger voltage being adjustable by means of two variable resistors. Stability is determined by use of Routh's criteria. General directions for the design of flip-flop circuits for low-voltage operation are summarized.

621.392.5 **1603**

Design of Low-Frequency Constant-Time-Delay Lines.—C. M. Wallis. (*Elect. Engng, N.Y.*, Dec. 1952, Vol. 71, No. 12, p. 1123.) Digest only. Analysis is presented for ladder-type low-frequency delay lines consisting of Pierce asymmetrical T sections, each section having two unequal coils in series, wound on a common core to give a coupling coefficient near unity, a capacitor being connected from the junction of the coils to earth. When properly terminated, such delay lines give delays constant to within $\pm 1\%$ over 60% of the pass band. Where many sections must be used, they have the decided advantage of requiring only two components per section, a double-winding inductor and a capacitor.

621.392.5 **1604**

Transfer Matrix of a Four-Terminal Passive Network in terms of its Image Parameters.—H. P. Biggar. (*Electronic Engng*, April 1953, Vol. 25, No. 302, pp. 152–153.) The constants in the fundamental simultaneous equations relating the input and output voltages and currents for a linear passive quadripole have a particular significance and can be used to derive the transfer matrix of any passive network in terms of the image impedances. This is demonstrated for the L type of network.

621.392.5 **1605**

The Measurement of 'A' Matrix Elements of Passive Networks.—W. R. Hinton. (*Electronic Engng*, April 1953, Vol. 25, No. 302, pp. 151–152.) The general method of using the 'A' matrix has previously been described (840 of 1950). A method is here described for determining the matrix elements from measurements of the impedance looking in at the sending end of a network, the receiving end being first open-circuited and then short-circuited. The elements so found apply only for the measurement frequency. The analysis shows that networks are electrically equivalent at a given frequency when their 'A' matrices are equal at that frequency.

621.392.5 **1606**

The Maximum Gain of an RC Network.—A. D. Fialkow & I. Gerst. (*Proc. Inst. Radio Engrs*, March 1953, Vol. 41, No. 3, pp. 392–395.) Analysis is presented indicating that any desired voltage gain can be achieved at a given real frequency by using networks of sufficient complexity, as represented by the degree of the corresponding formulae. Transfer-function theory developed previously (3369 of 1952) is used. The relation between the maximum possible gain and the degree of the transfer function is determined. The results are illustrated by examples.

621.392.5 **1607**

Note on a Network Theorem.—L. Storch. (*Wireless Engr*, April 1953, Vol. 30, No. 4, pp. 77–81.) The formulation adopted by Zepler (1535 of 1952) is criticized, and it is shown that his results can be obtained more simply by using the conventional impedance concept rather than his unconventional power concept.

621.392.5.018.7 **1608**

Waveform Computations by the Time-Series Method.—N. W. Lewis. (*Proc. Instn elect. Engrs*, Part III, March 1953, Vol. 100, No. 64, pp. 109–110.) Discussion on paper abstracted in 55 of January.

621.392.5.018.75 **1609**

The Synthesis of a Network to have a Sine-Squared Impulse Response.—W. E. Thomson. (*Proc. Instn elect. Engrs*, Part III, March 1953, Vol. 100, No. 64, p. 110.) Discussion on paper abstracted in 348 of February.

621.392.5.029.64 : 538.221 **1610**

Reduction of the Loss in Ferrite Materials in the Microwave Region.—H. N. Chait, N. G. Sakiotis & R. E. Martin. (*J. appl. Phys.*, Jan. 1953, Vol. 24, No. 1, pp. 109–110.) The loss in microwave gyrators, of the type comprising a ferrite rod inside a circular waveguide surrounded by a solenoid producing an axial magnetic field, has in certain cases been reduced by application of an auxiliary transverse static magnetic field.

621.392.52 **1611**

Filter Transfer Function Synthesis.—G. L. Matthaei. (*Proc. Inst. Radio Engrs*, March 1953, Vol. 41, No. 3, pp. 377–382.) Theory is developed for circuits in which approximately constant attenuation is achieved over one or more frequency bands. An electrostatic-potential-analogue method is used [see 2750 of 1949 (Klinkhamer)]. The method is applied to some examples of low-pass filters.

621.392.52 **1612**

A Systematic Procedure for the Design of Filters.—K. H. Haase. (*Frequenz*, Dec. 1952, Vol. 6, No. 12, pp. 363–368 & Jan. 1953, Vol. 7, No. 1, pp. 8–14.) Continuing previous work on the formulation of the characteristic functions of filters (3377 of 1952), tables are developed which facilitate the determination of the circuit elements of coupling-free filters of various types.

621.392.52 **1613**

Multistage Band-Pass Filter with Optimum Approximation to the Ideal Rectangular Shape of the Response Curve.—G. Drexler & H. H. Voss. (*Frequenz*, Nov. 1952, Vol. 6, No. 11, pp. 337–342.)

621.392.52 **1614**

The Design of Linear-Phase Low-Pass Filters.—C. F. Floyd, R. L. Corke & H. Lewis. (*Proc. Instn elect. Engrs*, Part IIIA, 1952, Vol. 99, No. 20, pp. 777–787. Discussion, pp. 860–866.) General design principles are outlined and a method applicable to certain low-pass filters required in television equipment is described. Adequate phase linearity of response can be obtained over the whole pass band only by limiting the discrimination required in the attenuation region of the insertion-loss characteristic. Practical examples include (*a*) a set of low-pass filters with linear phase characteristics from zero nearly to their cut-off frequencies, which range from 1·5 to 4 Mc/s, (*b*) an asymmetric-sideband filter with a cut-off frequency of 6·12 Mc/s and phase linearity from 3 to 7 Mc/s, with 30-db discrimination above 7 Mc/s.

621.392.52 : 538.652 **1615**

How to Use Mechanical I.F. Filters.—M. L. Doelz & J. C. Hathaway. (*Electronics*, March 1953, Vol. 26, No. 3, pp. 138–142.) Discussion of the construction and performance of electromechanical filters using magnetostrictive Ni wires as coupling units between metal-disk resonators. A typical 6-disk filter has a 6-db bandwidth of 3·1 kc/s centred on 455 kc/s, with a transmission loss $<$ 26 db and an operating temperature from $-$ 30° to $+$ 80°C. The size of the complete assembly is only $2\frac{13}{16} \times 1 \times \frac{15}{16}$ in.

621.392.6 **1616**

Fundamental Results and Outstanding Problems of Network Synthesis.—V. Belevitch. (*Tijdschr. ned. Radiogenoot.*, Jan. 1953, Vol. 18, No. 1, pp. 33–51. In English.) See also 1547 of 1952.

621.392.6 **1617**

Synthesis of *n*-Terminal-Pair Networks.—M. Bayard. (*Ann. Télécommun.*, Dec. 1952, Vol. 7, No. 12, pp. 517–524.) Dissipative networks only are considered. Generalization of the Gewertz reduced-matrix method and three methods involving auxiliary reactive networks are outlined and critically reviewed. The first auxiliary reactive network method considered was developed independently by Oono (65 of 1951) and Leroy (1616 of 1950 and 64 of 1951), the second was due to Bayard (593 of 1951) and the third to Belevitch (1547 of 1952).

621.396.6 : 061.4 **1618**

R.E.C.M.F. Exhibition Preview.—(*Electronic Engng*, April 1953, Vol. 25, No. 302, pp. 168–171.) Short descriptions of selected exhibits at the exhibition arranged by the Radio and Electronics Component Manufacturers' Federation, London, April 1953.

621.396.611.21 **1619**

Piezoelectric Oscillations of Quartz Plates at Even and Half-Odd Harmonics.—S. Parthasarathy, M. Pancholy & A. F. Chhapgar. (*Nature, Lond.*, 31st Jan. 1953, Vol. 171, No. 4344, pp. 216–217.) A number of quartz plates of different dimensions were plated with Ag, Al or Cu by various methods and tested in the same circuit. Results show that a quartz plate has preferred modes of oscillation at the odd and even harmonics and at frequencies midway between these, though the amplitude of the odd harmonics is very much the greater.

621.396.611.33 **1620**

The Double Resonance in the Secondary of a System of Two Circuits with Inductive Coupling.—R. Cazenave. (*Rev. gén. Élect.*, Dec. 1952, Vol. 61, No. 12, pp. 586–595.) The resonance condition is defined in terms of the primary and secondary impedances and the mutual inductance. The variation of secondary current with the coupling, frequency, and series capacitance is determined. In the case of a variable frequency with fixed coupling a graphical method is used to derive 'equal-current' curves. By means of these a region can be defined within which two resonance peaks occur.

621.396.611.4 **1621**

Some Exactly Integrable Cases of Electromagnetic Oscillations in Two Coupled Cavity Resonators.—L. Breitenhuber. (*Acta phys. austriaca*, Nov. 1951, Vol. 5, No. 1, pp. 45–68.) Calculations of resonance frequencies are made for cavity configurations whose boundaries coincide with co-ordinate surfaces of an elliptical-cylindrical co-ordinate system. The expressions obtained are developed explicitly for the case of weak coupling, and are compared with known approximations.

621.396.611.4 **1622**

Theory of Forced Oscillations in Electrodynamic Systems.—E. Ledinegg & P. Urban. (*Acta phys. austriaca*, June 1952, Vol. 5, No. 4, pp. 510–528.) The method of analysis using vector series progressing according to the eigenfunctions of the wave equation is applied to a system comprising cylindrical resonator and coaxial line. The theoretical assumption underlying the method are examined and the Green tensor appropriate to linear electrodynamic processes is formulated.

621.396.611.4 **1623**

The Complete System of Natural Electromagnetic Oscillations of Uniaxal Anisotropic Cavity Resonators.—E. Hafner. (*Arch. elekt. Übertragung*, Jan. 1953, Vol. 7, No. 1, pp. 47–56.) Analysis of the field inside an arbitrary cylinder completely or partially filled by a uniaxal crystal. The special cases of the circular-section cylinder and the parallelepiped are treated explicitly. See also 1315 of May.

621.396.615 **1624**

Perturbations of Filtered Oscillators: Part 2.—G. Cahen. (*C. R. Acad. Sci., Paris*, 26th Jan. 1953, Vol. 236, No. 4, pp. 356–358.) Results previously obtained (1279 of May) regarding the stability, synchronization and frequency pulling of a nonlinear oscillator subjected to excitation by an external source are generalized by considering the dependence of amplifier gain on the rate of phase change.

621.396.615 : 621.3.016.35 **1625**

Nonlinear Oscillators and the Nyquist Diagram.—A. Blaquière. (*J. Phys. Radium*, Nov. 1952, Vol. 13, No. 11, pp. 527–540.) Two types of nonlinear oscillators having amplitude limiting are considered. Their equations become linear when a particular value is chosen for the parameter involving the amplitude. A single curve can then be drawn for this particular value on the same principle as the ordinary Nyquist curve. Other values of the parameter yield a family of curves showing the stages in the evolution of oscillations. Conditions for stable and unstable oscillations are derived, and the results extended to cover response to a small perturbation, frequency stability and synchronization limits. A brief account of part of the work was given in 335, 1867 and 2735 of 1952.

621.396.615 + 534.1] : 621.3.016.37 **1626**

Electrical and Acoustic Oscillation Build-Up Phenomena.—H. Dänzer. (*Ann. Phys., Lpz.*, 15th June 1952, Vol. 10, Nos. 6/7, pp. 395–412.) Mathematical treatment of build-up phenomena such as those investigated experimentally by Trendelenburg et al. (1939 of 1938) for organ-pipe sounds.

621.396.615 : 621.396.611.34/.35 **1627**

Multimesh *RC* Networks for Phase-Shift Oscillators.—R. Krastel. (*Funk u. Ton*, Dec. 1952, Vol. 6, No. 12, pp. 649–653.) The required phase-shift for each mesh is calculated for a terminated 3-mesh network. The values, which total 180°, differ considerably from 60°.

621.396.615.029.53/.55 : 621.384.612 **1628**

A Wide-Range Oscillator of High Stability.—D. E. Caro & L. U. Hibbard. (*J. sci. Instrum.*, Dec. 1952, Vol. 29, No. 12, pp. 403–408.) Description of a beat-frequency oscillator for the r.f. system of the Birmingham University proton synchrotron [700 of 1951 (Hibbard)]. The two similar oscillators used are tuned by $\lambda/4$ coaxial resonators on rubber shock mounts in a constant-temperature enclosure, the fixed-frequency oscillator operating at about 33 Mc/s. A capacitor using a single-plate rotor with a stepped profile serves for varying the tuning of the other oscillator via a servomechanism. The beat frequency as a function of rotor setting is reproducible to within 0·03% throughout the range 0·25–10 Mc/s.

621.396.615.17 : 621.314.25.015.7 **1629**

A Phase-Shifting Pulse Generator for Thyratron Control.—J. C. West & D. K. Partington. (*Electronic Engng*, March 1953, Vol. 25, No. 301, pp. 120–121.) Description, with detailed circuit diagram, of equipment producing positive pulses with repetition frequency that of the mains (50 c/s), but with phase variable with respect to that of the supply voltage. The phase lag is controlled by application of a small direct voltage to the grid of a high-vacuum valve.

621.396.615.17 : 621.317.755 **1630**

Pulse-Operated Time-Bases.—H. A. Dell. (*Electronic Engng*, March 1953, Vol. 25, No. 301, pp. 94–97.) Discussion of the basic principles and description of a practical timebase circuit of the single-sweep type, triggered by a delayed pulse developed from the transient or transmitter pulse to be examined.

621.396.615.17 : 621.397.621.2 **1631**

Timebase Circuits.—H. Bähring. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, Nov./Dec. 1952, Vol. 4, Nos. 11/12, pp. 219–227.)Discussion of the principles of operation of various types of timebase, including an anastigmatic system, and of h.v. generation by the fly-back method.

621.396.645 **1632**

The Three Possible Types of Valve Amplifier Circuit and Applications in Television Receivers.—W. Taeger. (*Frequenz*, Nov. 1952, Vol. 6, No. 11, pp. 330–335.) Analysis and comparison of grounded-cathode, grounded-grid and grounded-anode amplifier circuits.

621.396.645 : 621.317.3 **1633**

A Selective Detector Amplifier for 10–10 000 c/s.—G. H. Rayner. (*J. sci. Instrum.*, Jan. 1953, Vol. 30, No. 1, pp. 17–20.) An amplifier is described for use as a detector in bridge or other measurements where a null indication is required. Tuning is accomplished by means of parallel-T feedback networks including only resistors and capacitors. Noise is equivalent to an input signal of 0·2 μV across $10^5\Omega$.

621.396.645.015.7 : 621.396.822 **1634**

Background Noise in Ionization-Chamber Pulse Amplifiers.—G. Valladas & A. Leveque. (*J. Phys. Radium*, Nov. 1952, Vol. 13, No. 11, pp. 521–526.) Two methods of particle-energy measurement are considered. For the charge-collection method Milatz & Keller (1750 of 1945) have calculated the absolute limit to measurement accuracy for the amplifier circuit used. For the pulse-slope method initiated by Sherr & Peterson (*Rev. sci. Instrum.*, 1947, Vol. 18, p. 567) a slightly different limit is found. Circuits giving optimum signal/noise ratios are proposed in each case.

621.396.645.029.3 **1635**

Constant-Current Audio Power Amplifiers.—H. T. Sterling & A. Sobel. (*Electronics*, March 1953, Vol. 26, No. 3, pp. 122–125.) The design features are discussed of a push-pull amplifier using a pair of Type-5881 valves in the output stage to feed a General Radio Type-942A toroidal output transformer. Constant-current operation of the output stage, with 1-kΩ resistors in the cathode leads for d.c. degeneration, minimizes variations due to valve aging or replacement. Full circuit details of the amplifier and its simple power-supply unit are given.

621.396.645.029.4 **1636**

Comments on the Performance of Pentode Output Stages.—B. G. Dammers. (*Philips tech. Commun., Aust.*, 1952, No. 6, pp. 3–13.) Excerpt from Book V of the Philips series of books on electronic valves (2765 of 1952).

621.396.645.36 **1637**

The Differential Amplifier.—V. S. E. Lewis: B. F. Davies. (*Electronic Engng*, Feb. 1953, Vol. 25,No. 300, pp. 82–83.) Comment on 3392 of 1952 and author's reply.

621.396.645.37 + 621.396.615.11 **1638**

Electrically Tuned RC Oscillator or Amplifier.—O. G. Villard, Jr, & F. S. Holman. (*Proc. Inst. Radio Engrs*, March 1953, Vol. 41, No. 3, pp. 368–372.) Two variants are described of a feedback circuit including two all-pass phase-shifting networks. Tuning is performed by adding a variable amount of the output of the one phase shifter to the output of the other, the variation being controlled electronically by means of a modulator circuit. Theoretically the tuning range is from zero to infinite frequency; in practice the frequency ratio obtainable without appreciable change in gain is about 4 : 1.

621.397.645.029.62 **1639**

Fundamental Problems of H.F. and I.F. Amplifiers for TV Reception: Part 1 — Gain and Bandwidth.—A. G. W. Uitjens. (*Electronic Applic. Bull.*, Nov. 1950, Vol. 11, No. 11, pp. 205–228. Errata, *ibid.*, Jan. 1951, Vol. 12, No. 1, p. 17.) The gain-bandwidth product GB is shown to depend exclusively on the ratio of valve mutual conductance to the sum of input and output capacitances. Factors affecting G and B are discussed. Amplifier response curves are considered and the total bandwidth is deduced for amplifiers with synchronous circuits. Staggered tuning is discussed in detail. The unit-function response curves for double- and vestigial-sideband systems are treated and the distortion in the latter system is analysed. The gain and bandwidth of quadripole coupling networks, particularly the Π network and the double-tuned band-pass filter, are discussed and characteristic data, including GB values, are tabulated for nine commonly used pentode circuits. Data for unit-function response curves when several identical groups are connected in cascade, and bandwidth factors for double-tuned coupling circuits, are also tabulated.

621.397.645.029.62 : 621.397.822 **1640**

Fundamental Problems of H.F. and I.F. Amplifiers for TV Reception: Part 2 — Noise.—A. G. W. Uitjens. (*Electronic Applic. Bull.*, Jan. 1951, Vol. 12, No. 1, pp. 2–17. Correction, *ibid.*, March 1951, Vol. 12, No. 3, p. 64.) Noise factors of amplifiers operating under various specified conditions are calculated and valve noise is discussed. Formulae are derived for the noise factor at metre wavelengths for four special simplified cases, a general formula being given in an appendix. The concept of noise reference frequency f_n, the frequency at which the equivalent noise resistance of a valve is equal to the input resistance, is discussed and formulae are tabulated for both triode and pentode circuits with predominating partition noise. Graphs of noise factor for various cases show that above the frequency f_n the noise factor increases rapidly. The signal/noise ratio is calculated from the noise factor. It appears that the weakest signal that can be received reasonably well depends on the frequency, on the aerial noise, and particularly on the noise reference frequency of the first valve of the receiver.

GENERAL PHYSICS

535.13 : 517.941.9 **1641**

The First Boundary-Value Problem of Maxwell's Equations.—E. Ledinegg & P. Urban. (*Ann. Phys., Lpz.*, 15th June 1952, Vol. 10, Nos. 6/7, pp. 349–360.) Mathematical analysis of the generation of oscillations in an e.m. system by an external source, such as excitation of a cavity resonator by means of a waveguide feeder.

535.325-1 **1642**

Refractive Index of Dipolar Gas Mixtures at Decimetre Wavelengths.—R. Schachenmeier & F. Weller. (*Naturwissenschaften*, Jan. 1953, Vol. 40, No. 1, pp. 17–18.) The refractive index of mixtures of air and water vapour was measured at frequencies from 600 to 1 700 Mc/s over a range of pressures, using a cavity-resonator method. Results are shown in curves. Discrepancies between the observations and values predicted from theory are attributed to the fact that the relaxation time of the water molecules is not negligible at low pressures.

535.37 **1643**

Capture and Recapture of Electrons in Phosphorescence.—D. Curie. (*Ann. Phys., Paris*, Nov./Dec. 1952, Vol. 7, pp. 749–801.) The relative importance of these phenomena is discussed and the mechanisms with which they are associated in phosphorescence decay are analysed.

535.37 : 537.52 **1644**

The Luminescence of Phosphors in Strong Electric Fields.—A. Herwelly. (*Acta phys. austriaca*, Nov. 1951, Vol. 5, No. 1, pp. 30–44.) The account of the phenomenon of electroluminescence given by Destriau (see e.g. 110 of 1949) is critically discussed. Cells used for demonstrating the phenomenon always take the form of a stratified capacitor including a layer of gas and a layer of solid dielectric of much greater breakdown strength, the gas being occluded either in the electrodes or in or between the phosphor grains. With such an arrangement a glow discharge occurs in the gas layer for voltages above a certain threshold and below sparking voltage. It is concluded that this is the true explanation of the observed phenomenon.

537.1 : 530.12 **1645**

Circulation in the Flow of Electricity: Dirac's New Variables.—O. Buneman. (*Proc. roy. Soc. A*, 5th Dec. 1952, Vol. 215, No. 1122, pp. 346–352.) "Kelvin's circulation theorems are shown to be applicable to the generalized momentum vector in the relativistic flow of electricity. Vortex filaments become 'vortex ribbons' in space-time, and Dirac's variables ξ and η are identified as the parameters which characterize the family of vortex ribbons."

537.213 **1646**

The Electric Potential in an Electrically Shielded Space.—G. Piccardi & R. Cini. (*Ricerca sci.*, Jan. 1953, Vol. 23, No. 1, pp. 113–114.) Experiments confirm the previously observed variation of potential inside a metallic enclosure; the phenomenon is attributed to the production of space charge by ionization due to spontaneous radiation from the enclosure walls.

537.226 **1647**

The Frequency Dependence of the Dielectric Properties of Dipolar Substances.—J. J. O'Dwyer & R. A. Sack. (*Aust. J. sci. Res., Ser. A*, Dec. 1952, Vol. 5, No. 4, pp. 647–660.) Inconsistencies in previous theories of dipolar interaction are pointed out. A new theory presented is based on Debye's theory of molecular relaxation and on an application of Onsager's method for treating e.s. interaction.

537.291 + 538.691 **1648**

The Solution of the Schrödinger Equation for Finite Systems, with special reference to the Motion of Electrons in Coulomb Electric Fields and Uniform Magnetic Fields.—R. B. Dingle. (*Proc. Camb. phil. Soc.*, Jan. 1953, Vol. 49, Part 1, pp. 103–114.)

537.291 **1649**

Motion of Gaseous Ions in Strong Electric Fields.—G. H. Wannier. (*Bell Syst. tech. J.*, Jan. 1953, Vol. 32, No. 1, pp. 170–254.) Amplified and extended version of papers noted in 97 of 1952 and 376 of February.

537.3 **1650**

Current Flow in Cylinders.—W. R. Smythe. (*J. appl. Phys.*, Jan. 1953, Vol. 24, No. 1, pp. 70–73.) The mixed-boundary-value problem of the flow of current into a right circular cylinder through a perfectly conducting coaxial disk electrode at one end is solved approximately.

537.315.6 **1651**

Numerical Calculation of the Potential and Field due to a Uniformly Charged Disk.—M. Laudet. (*J. Phys. Radium*, Nov. 1952, Vol. 13, No. 11, pp. 549–553.) Various methods of calculating the potential at any point are discussed, the particular advantages of the relaxation method in the most general case being pointed out.

537.52 : 538.639 **1652**

New Types of Discharge in Magnetic Fields.—K. Siebertz. (*Acta phys. austriaca*, Nov. 1951, Vol. 5, No. 1, pp. 134–142.) Discharge phenomena described by Ehrenhaft et al. (*Acta phys. austriaca*, 1950, Vol. 4, No. 1, p. 129) are explained in terms of known concepts and laws of gas-discharge physics.

537.52 : 538.639 **1653**

Investigations of Discharges in Magnetic Fields. A Note on Righi's Magnetic Rays.—W. H. Bergmann. (*Acta phys. austriaca*, June 1952, Vol. 5, No. 4, pp. 425–428.)

537.525 : 538.566 **1654**

The Microwave Admittance of a Mercury-Vapour Discharge.—H. A. Prime. (*Aust. J. sci. Res., Ser. A*, Dec. 1952, Vol. 5, No. 4, pp. 607–617.)

537.525 : 538.63 **1655**

Theory of the Discharge Potential for Coaxial Cylindrical Electrodes in a Transverse Magnetic Field.—G. Valle. (*Nuovo Cim.*, 1st Feb. 1952, Vol. 9, No. 2, pp. 145–168.)

537.527 : 538.566 **1656**

The Absorption and Reflection of Microwave Radiation by a Mercury-Vapour Discharge.—H. A. Prime. (*Aust. J. sci. Res., Ser. A*, Dec. 1952, Vol. 5, No. 4, pp. 592–606.) Results of measurements at 3-cm λ show that with increasing discharge current the real component of the complex conductivity of the discharge increases almost linearly, the imaginary component, which is negative, increasing nonlinearly and less rapidly.

537.533 : [537.29 + 538.63 **1657**

The Imaging Properties of Electron Beams in Arbitrary Static Electromagnetic Fields.—P. A. Sturrock. (*Phil. Trans. A*, 19th Aug. 1952, Vol. 245, No. 894, pp. 155–187.) Detailed theory of the aberrations of electron-optical systems with curved axes, with application to calculation of the imaging properties of a helical beam moving in the field of a pair of coaxial cylindrical electrodes.

537.533.8 : [546.56-1 + 546.57-1] : 539.234 **1658**

The Secondary Emission from Copper and Silver Films obtained with Primary-Electron Energies below 10 eV.—H. P. Myers. (*Proc. roy. Soc. A*, 5th Dec. 1952, Vol. 215, No. 1122, pp. 329–345.) Values obtained for the secondary-emission ratio are lower than those hitherto recorded, both Ag and Cu having a ratio $< 0{\cdot}1$ in the primary-energy range 1–5 eV. The difference is attributed to the improved vacuum conditions used. Measurements of the energy distribution of the secondary electrons for primary energies < 5 eV show that the secondary electrons are in fact elastically reflected primary electrons.

537.562 **1659**

Electromagnetic Properties of Plasmas.—M. Bayet. (*J. Phys. Radium*, Nov. 1952, Vol. 13, No. 11, pp. 579–586.) Conditions for the propagation of longitudinal and transverse e.m. waves are considered, taking account of the effect of collisions on electron velocities and assuming no applied magnetic field. The mobility coefficient relating the electric field to the mean velocity imparted by it to the electrons, and dispersion formulae linking wavelength and frequency of the propagated disturbance, are derived.

537.562 : 538.56 **1660**

Studies in Magneto-hydrodynamics.—S. Lundquist. (*Ark. Fys.*, 26th Aug. 1952, Vol. 5, Part 4, pp. 297–347.) A systematic treatment of the subject, designed to bring out the underlying physical principles. Basic equations for e.m. phenomena in a conducting liquid and their conditions of validity are discussed. These are applied in specific static, kinematic and dynamic problems. Experimental work is briefly summarized.

537.562 : 538.566 : 532.51 **1661**

A New Type of Propagation.—C. H. Papas & W. W. Salisbury. (*Trans. Inst. Radio Engrs*, Dec. 1952, No. PGAP-4, pp. 47–52.) Discussion of the properties of magneto-hydrodynamic waves in various types of conducting fluids. See also 3469 of 1945 (Alfvén).

538.221 **1662**

Overlapping Energy Bands and the Theory of Collective Electron Ferromagnetism.—A. B. Lidiard. (*Proc. Camb. phil. Soc.*, Jan. 1953, Vol. 49, Part 1, pp. 115–129.)

538.221 : 538.65 **1663**

Motions of Material Particles in the Field of an Electromagnet.—F. Stockinger. (*Acta phys. austriaca*, June 1952, Vol. 5, No. 4, pp. 440–448.) Description and discussion of two different effects observed when ferromagnetic particles of order of magnitude 10^{-4}–10^{-2} cm are exposed to fields of strength up to 30 000 oersted in the air-gap of a magnet.

538.221 : 538.65 **1664**

The Behaviour of Ferromagnetic Particles in a Rotating Magnetic Field.—K. V. Desoyer. (*Acta phys. austriaca*, June 1952, Vol. 5, No. 4, pp. 429–434.) Account of an experimental investigation of the motion of ferromagnetic particles of order of magnitude 10^{-4}–10^{-2} cm, suspended in gas at pressures between 2 and 20 mm Hg, and subjected to illumination and to a magnetic field rotating at 50 c/s.

538.566 : 535.42 **1665**

Theory of Diffraction by a Cylinder, taking account of the Surface Wave.—W. Franz & K. Deppermann. (*Ann. Phys., Lpz.*, 15th June 1952, Vol. 10, Nos. 6/7, pp. 361–373.) Maxima and minima observed by A. Limbach in diffraction of centimetre e.m. waves by metal cylinders are found to be due to interference between the waves geometrically reflected from the cylinder surface and the surface wave creeping round to the back of the cylinder. Quantitative calculations based on the integral equations of the diffraction theory of Maue (*Z. Phys.*, 1949, Vol. 126, p. 606) give results in good agreement with experiment.

538.566 : 535.42 **1666**

Diffraction Patterns in Circular Apertures less than One Wavelength in Diameter.—H. L. Robinson. (*J. appl. Phys.*, Jan. 1953, Vol. 24, No. 1, pp. 35–38.) Measurements were made of the field intensity at points along the aperture diameters parallel respectively to the electric and magnetic vectors, using normally incident radiation of wavelengths 16 cm and 32 cm. The results disagree with values calculated from Young's theory in two respects, viz., (*a*) the observations indicate a sharp increase at the ends of the diameter parallel to the electric vector; (*b*) along the diameter parallel to the magnetic vector the observed central peak for apertures of diameter near $\lambda/2$ was 50% greater than the calculated value, the high intensity being attributed to multiple reflections from the edges.

538.566 : 535.42 **1667**

The Electromagnetic Field in the Plane of a Circular Aperture due to Incident Spherical Waves.—D. C. Hogg. (*J. appl. Phys.*, Jan. 1953, Vol. 24, No. 1, pp. 110–111.) Measurements were made of the field in an aperture of diameter 50 cm illuminated with radiation of wavelength 1·25 cm from small radiators at various distances. Results are shown graphically; the curves obtained can be represented approximately by a modified form of Andrews' equation (3141 of 1950).

538.566.2 : 512.831 **1668**

Application of Matrices to the Problem of Transmission through a Multi-Layered Dielectric Wall.—R. A. Henschke. (*Trans. Inst. Radio Engrs*, Dec. 1952, No. PGAP-4, pp. 117–134.) The matrix methods described by Watson (1329 of 1947) are applied to the determination of design formulae for dielectric walls of the symmetrical sandwich type.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.746 : [523.77 + 53 **1669**

The Physical Conditions in Sunspots, deduced from their Spectra.—R. Michard. (*C. R. Acad. Sci., Paris*, 12th Jan. 1953, Vol. 236, No. 2, pp. 182–184.) Discussion of the theoretical consequences resulting from the sunspot model recently proposed (1324 of May). The vertical gradient of the magnetic field is found to be about 1·5 gauss/km.

551.510.52 **1670**

Nomograms for the Computation of Tropospheric Refractive Index.—D. M. Swingle. (*Proc. Inst. Radio Engrs*, March 1953, Vol. 41, No. 3, pp. 385–391.) Charts for calculating refractive index, refractive-index discontinuities, and vertical gradient are presented in a form convenient for use with standard radiosonde data. They are valid for wavelengths $> 1{\cdot}5$ cm.

551.510.52/.53 : 551.571.7(422) **1671**

Humidity of the Upper Troposphere and Lower Stratosphere over Southern England.—J. K. Bannon, R. Frith & H. C. Shellard. (*Geophys. Mem., Lond.*, 1952, Vol. 11, No. 88, 36 pp.) Detailed report of observations made on 130 flights during 1943–1946 at altitudes up to about 40 000 ft. For a summarized account, see *Nature, Lond.*, 28th Feb. 1953, Vol. 171, No. 4348, pp. 381–382.

551.510.535 **1672**

Abnormalities in the Ionosphere at High Latitudes.—W. R. Piggott. (*Nature, Lond.*, 17th Jan. 1953, Vol. 171, No. 4342, pp. 124–125.) Study of the variations, with respect to time and location, of the frequency of occurrence of storm types E and D shows that the polar region may be divided into three zones, the polar zone, the storm belt and the polar quiet zone. Within the storm belt, the most dense forms of storm-E ionization are practically confined to an area centred near Canada, while the existence can be demonstrated of two centres of activity in the D region, spaced 180° apart, which rotate once around the earth every 48 hours.

551.510.535 **1673**

Geo-morphology of F_2-Region Ionospheric Storms.—D. F. Martyn. (*Nature, Lond.*, 3rd Jan. 1953, Vol. 171, No. 4340, pp. 14–16.) F_2-region ionospheric storms may be regarded as manifestations of a single world-wide phenomenon, viz., the e.s. field of the solar diurnal current system, whose local effects depend markedly on local time and season. The theoretical (Chapman) characteristics of the F_2 region are normally perturbed by solar tidal electric fields (1053 of 1949). During magnetic storms the phase of the additional electric field present is such as to increase these perturbations at high latitudes and to reduce them at low latitudes, though the effect may be masked by local time variations of ionization density. In contradistinction to Appleton & Piggott (891 of 1950), no evidence is found that any magnetic storm affects the northern and southern hemispheres differently, when due allowance is made for the seasonal effect.

551.510.535 : [523 : 621.396.822 **1674**

Study of the Ionosphere by Extraterrestrial Radio Waves.—A. P. Mitra. (*Indian J. Phys.*, Oct. 1952, Vol. 26, No. 10, pp. 495–511.) A paper surveying experimental work on the use of galactic r.f. radiation for investigating ionospheric refraction, absorption, irregularities and the effects of sudden ionospheric disturbances. Results are compared with theory, and further lines of investigation are suggested.

551.510.535 : 523.78 **1675**

E Region during the Solar Eclipse of February 25 1952.—C. M. Minnis. (*Nature, Lond.*, 13th Sept. 1952, Vol. 170, No. 4324, p. 453.) Analysis of observations of E-layer critical frequencies at Khartoum during the eclipse indicates that the ionizing radiation from the solar disk is not uniformly distributed, as is also the case for the optical radiation. Part of the nonuniform component of the ionizing radiation appears to be divided about equally between the east and west limbs of the sun. This is consistent with observations made at Meudon of the distribution over the sun's disk of the intensity of the green coronal line. The main discrepancy between the values of the parameter J (proportional to the radiation flux) calculated from the coronal data and from the E-region measurements appears abruptly just after totality and suggests the existence of an intense source of ionizing radiation near the west limb of the sun which cannot be explained in terms of solar data available at present.

551.510.535 : 551.55 **1676**

A Study of Ionospheric Winds and Turbulence utilizing Long Radio Waves.—G. H. Millman. (*Ann. Géophys.*, Oct./Dec. 1952, Vol. 8, No. 4, pp. 365–384. In English.) An experimental investigation of winds in the E-layer was made during the period July 1951–March 1952, using vertical-incidence pulse technique, with waves of frequency 150 kc/s. The time variations of the reflected amplitudes are analysed statistically and compared with results to be expected from theory. Fades of duration 0.5–6.5 min are most prevalent. Diurnal and seasonal effects are observed. See also 720 of March.

551.510.535 : 621.3.087.4 **1677**

A High-Precision Ionospheric Sounding Equipment.—B. M. Banerjee & R. Roy. (*Indian J. Phys.*, Oct. 1952, Vol. 26, No. 10, pp. 473–494.) Description, with circuit diagrams and oscillograms, of high-resolution equipment which provides pulses variable from 6 to 30 μs in duration and from 10 to 50 kW in peak power. The delta aerials are of the type considered by Cones (1567 of 1951), alternative terminations giving improved signal strength at frequencies $<2 \cdot 5$ Mc/s. During conditions of severe interference a horizontal receiving dipole is used. Interchangeable receivers covering a bandwidth of 50–200 kc/s are provided. The timebase circuit previously described (2477 of 1952) is modified to give a 12-line raster, each line corresponding to 50-km height. The evaluation of resolution is discussed in an appendix.

551.510.535 : 621.396.11 **1678**

Effect of the Geomagnetic Field on the Absorption of Short Waves in the Ionosphere (Vertical Incidence).—É. Argence, K. Rawer & K. Suchy. (*C. R. Acad. Sci., Paris*, 12th Jan. 1953, Vol. 236, No. 2, pp. 190–192.) An approximate formula for the index of refraction is used to calculate the coefficient of absorption for five frequencies around 3 Mc/s. The results are sensibly the same as those obtained by use of the rigorous Appleton-Hartree formula; those previously given by Argence, Mayot & Rawer (1300 of 1952) are about 3% higher.

551.510.535 : 621.396.11 **1679**

The Reflection and Absorption of Radio Waves in the Ionosphere.—W. R. Piggott. (*Proc. Instn elect. Engrs*, Part III, March 1953, Vol. 100, No. 64, pp. 61–72.) Investigations of the absorption of r.f. waves made in Britain since 1935 are reviewed and the theory of ionospheric absorption is outlined. The effects of double refraction, polarization, spatial attenuation, dispersion, ionospheric inhomogeneities and partial reflection, which modify the apparent attenuation of radio waves reflected in the ionosphere, are discussed and methods used to minimize absorption-measurement errors due to these factors are given, the experimental techniques used in routine absorption measurements being described fully. Detailed results of measurements of ionospheric absorption made in South-East England in the period 1935–1948 are being published separately.

551.510.535 : 621.396.812 **1680**

The Investigation of Ionospheric Absorption by a New Automatic Method: Part 1—Measurements on Vertical-Incidence Pulse Signals.—J. B. Jenkins & G. Ratcliff. (*Electronic Engng*, April 1953, Vol. 25, No. 302, pp. 140–145.) Equipment is described which automatically measures echo pulse amplitude and produces records, at minute intervals, of the integral of all echo pulse amplitudes received during each minute. The output of a modified communications receiver, blanked so as not to respond to the transmitter pulse, is fed to two gated amplifiers, one of which selects the echo signal and associated noise, while the other deals with noise only. After integration for 1 min, the difference of the output voltages of the integrators is applied to a pen recorder, the integrators at the same time being reset to zero. Under good conditions, with little rapid fading, the results obtained are in good agreement with results noted by a skilled observer from a c.r.o. display of the transmitted and echo pulses. Wide differences are found under poor conditions, with rapid fading, and the results furnished by the automatic equipment are considered the more reliable.

551.594 **1681**

Measurement of the Electric Field of the Atmosphere by means of Radioactive Probes.—H. Wichmann. (*Arch. Met. A, Wien*, 28th April 1952, Vol. 5, No. 1, pp. 86–99.)

551.594.21 **1682**

Boys-Camera Study of the 'Air Discharge' and the 'Flash to Ground with Horizontal Top'.—M. Sourdillon. (*Ann. Géophys.*, Oct./Dec. 1952, Vol. 8, No. 4, pp. 349–364.)

551.594.21 **1683**

Air Discharges and the Positive Charge at the Bottom of a Thundercloud.—D. J. Malan. (*Ann. Géophys.*, Oct./Dec. 1952, Vol. 8, No. 4, pp. 385–401.)

LOCATION AND AIDS TO NAVIGATION

621.396.9 **1684**

Omnidirectional Reflectors for Microwaves.—L. Grifone. (*Ricerca sci.*, Dec. 1952, Vol. 22, No. 12, pp. 2307–2313.) Measurements at 3·2-cm wavelength of the reflecting power of a corner reflector relative to that of a plane reflector were made by rotating the reflector about a vertical axis through a total angle of about 90°. The results are shown graphically for angles of tilt of the reflector up to 40°. A reflector system with satisfactory omnidirectional characteristics, consisting of two staggered groups of four, one above the other, should be detectable at a distance of over 3 sea miles under adverse conditions, using radar equipment with peak pulse power of 35 kW.

621.396.9 : 627.3 **1685**

Shore-Based Radar for Harbor Surveillance.—E. J. Isbister & W. R. Griswold. (*Elect. Engng, N.Y.*, Dec. 1952, Vol. 71, No. 12, pp. 1072–1077.) General descriptions are given of the installations at the ports of Boston, Long Beach, New York and Liverpool.

MATERIALS AND SUBSIDIARY TECHNIQUES

533.5 **1686**

An Assessment of some Working Fluids for Diffusion Pumps.—D. Latham, B. D. Power & N. T. M. Dennis. (*Vacuum*, Jan. 1952, Vol. 2, No. 1, pp. 33–49.) Choice of working fluid is largely determined by the particular pump to be used and conditions of use. Silicones are outstanding for general purposes and valveless pumping to really low pressures, Apiezon-C is considered the best for lowest possible pressures in a fractionating pump, while Aroclor-1245 is good for applications where very low pressures are not required.

535.215 : 546.482.21 **1687**

Photoconductivity of Cadmium-Sulphide Crystals at Relatively High Temperatures.—R. Caspary & H. Müser. (*Z. Phys.*, 17th Dec. 1952, Vol. 134, No. 1, pp. 101–105.) A variation of the method described by Frerichs (450 of 1948) resulted in the production of crystals up to 2 in. long. Curves are given showing the response to light of wavelength in the range 425–610 mμ at temperatures from 18° to 148°C. Maximum sensitivity occurs at wavelengths a little above 500 mμ; it decreases with increasing temperature and is displaced slightly towards longer wavelengths.

535.215 : 546.482.21 : 548.55 **1688**

Growth and Decay of the Photoconductivity of CdS Single Crystals.—J. Fassbender & B. Seraphin. (*Ann. Phys., Lpz.*, 15th June 1952, Vol. 10, Nos. 6/7, pp. 374–394.)

535.215.1 : [546.22.36 + 546.22.35 **1689**

Photoemission from Cesium and Rubidium Tellurides.—E. Taft & L. Apker. (*J. opt. Soc. Amer.*, Feb. 1953, Vol. 43, No. 2, pp. 81–83.) These materials are similar to Cs_3Sb (see 1690 below) except that their forbidden energy bands are almost twice as wide. Consistent with this, the energy distributions of the emitted photoelectrons show no evidence for electron-electron scattering up to $h\nu = 6{\cdot}7$ eV. These materials are of interest for ultraviolet photometry.

535.215.1 : 546.86.36 **1690**

Electron Scattering and the Photoemission from Cesium Antimonide.—L. Apker, E. Taft & J. Dickey. (*J. opt. Soc. Amer.*, Feb. 1953, Vol. 43, No. 2, pp. 78–80.) Measurements of the energy distribution of photoelectrons emitted from Cs_3Sb are reported. At values of $h\nu > 4$ eV only a small proportion of the photoelectrons have energies > 2 eV; this is attributed to scattering of excited electrons by valence electrons. The results give support to the energy-band structure proposed by Burton (*Phys. Rev.*, 1947, Vol. 72, p. 531A).

535.37 + 535.215] : 546.482.21 : 548.55 **1691**

Luminescence and Photoconductivity in Cadmium Sulfide at the Absorption Edge.—C. C. Klick. (*Phys. Rev.*, 1st Jan. 1953, Vol. 89, No. 1, pp. 274–277.) Measurements on single crystals of CdS at 77°K and 4°K suggest that emission at the edge of the absorption band may occur at special centres and not be characteristic of the pure lattice.

537.226.2/.3 **1692**

Study of Dielectrics; Dielectric Constant, Losses and Breakdown.—M. Bouix. (*Ann. Télécommun.*, Nov. & Dec. 1952, Vol. 7, Nos. 11 & 12, pp. 466–479 & 497–512.) A review of modern theories relating to calculation of the dielectric constant of and losses in nonpolar, polar and dipolar materials, and concerning dielectric breakdown.

537.226.2 : 539.22 **1693**

The Dielectric Constants of Heterogeneous Mixtures of Isotropic and Anisotropic Materials.—W. Niesel. (*Ann. Phys., Lpz.*, 15th June 1952, Vol. 10, Nos. 6/7, pp. 336–348.) A general theory applicable to the calculation of the dielectric constants of mixtures is developed. The individual particles of the mixtures are assumed to be ellipsoidal. For the particular cases of aggregates of spherical or laminar particles the general formulae reduce to those given by Bruggemann (1284 and 2424 of 1936 and 1087 of 1937). Measurements on multicrystal $BaTiO_3$ aggregates are discussed in relation to the theory.

537.228.1 : 546.431.824-31 **1694**

Measurements on Barium Titanate.—H. Schmidt. (*Akust. Beihefte*, 1952, No. 2, pp. 83–88.) Formulae are given for calculating the mechanical constants of piezoelectric disks from electrical measurements. Results of observations of the temperature dependence of electric and elastic parameters of $BaTiO_3$ and the mixed ceramic $BaTiO_3$-$PbTiO_3$ are shown in graphs, and values of the parameters for $BaTiO_3$ are tabulated.

537.311.33 **1695**

Modulation of the Conductance of a Semiconductor by an Electric Field.—P. Aigrain, J. Lagrenaudie & G. Liandrat. (*J. Phys. Radium*, Nov. 1952, Vol. 13, No. 11, pp. 587–588.) An a.c. field is applied normally to the surface of an evaporated Te film through which a.c. of the same frequency is passed. The modulation of the conductance by the normally applied field gives rise to a d.c. component in the current through the specimen; from measurements of this component the conductance variations are deduced. Very small changes can be estimated by this method.

537.311.33 : 546.27.03 1696

Electrical and Optical Properties of Boron.—J. Lagrenaudie. (*J. Phys. Radium*, Nov. 1952, Vol. 13, No. 11, pp. 554–557.) Measurements on three differently prepared samples are reported. The shape of the conductivity/temperature curves over a low-temperature range indicates the existence of several impurity energy levels. The material is a *p*-type semiconductor. Photoconductivity measurements indicate an intrinsic energy of about 1·2 eV.

537.311.33 : 546.289 1697

The Question of Thermally Produced Lattice Defects in Germanium.—K. Seiler, D. Geist, K. Keller & K. Blank. (*Naturwissenschaften*, Jan. 1953, Vol. 40, No. 2, p. 56.) Experiments are described which confirm that the specific resistance of pure Ge is not altered by heat treatment followed by rapid cooling; a change of resistance in these conditions indicates the presence of impurities, possibly only on the surface. Ca, Mg, Fe and Cu are likely impurities, Si is improbable. These results leave it undecided whether lattice defects can be produced in Ge by heat treatment.

537.311.33 : 546.289 1698

Surface Properties of Germanium.—W. H. Brattain & J. Bardeen. (*Bell Syst. tech. J.*, Jan. 1953, Vol. 32, No. 1, pp. 1–41.) Measurements are reported of contact potential (c.p.) and change of c.p. with illumination, using a Pt reference electrode. The influence of the ambient medium is investigated; the c.p. varies over a range of about 0·5 V between the value corresponding to the largest surface dipole, obtained in ozone or peroxide vapours, and that corresponding to the smallest surface dipole, obtained in vapours with OH radicals. For chemically prepared surfaces, the value of the surface recombination velocity is 50–70 cm/s for *n*-type and 180–200 cm/s for *p*-type Ge and is independent of c.p. A theoretical explanation is given in terms of surface traps. The experimental results afford direct evidence for the existence of a space-charge layer at the free surface of a semiconductor.

537.311.33 : 546.289 1699

Electronic Structure of the Germanium Crystal.—F. Herman & J. Callaway. (*Phys. Rev.*, 15th Jan. 1953, Vol. 89, No. 2, pp. 518–519.) Study of the energy-band structure of the Ge crystal by the method of orthogonalized plane waves. Preliminary note only.

537.311.33 : 546.47-31 1700

New Method of Studying Lattice Defects in the Semiconductor ZnO by Radio-Frequency Spectroscopy.—M. Freymann & R. Freymann. (*J. Phys. Radium*, Nov. 1952, Vol. 13, No. 11, pp. 589–590.) ZnO specimens were examined at frequencies from 1 kc/s to 1 Mc/s and temperatures from 100° to 293°K. Two components of Debye absorption were found, the respective intensity and position of which vary according to specimen preparation, temperature and frequency. Energy levels for diffusion of the lattice defects are deduced, and the results compared with those of other workers.

537.311.33 : 546.561-31 1701

The Semiconducting Properties of Cuprous Oxide.—J. S. Anderson & N. N. Greenwood. (*Proc. roy. Soc. A*, 5th Dec. 1952, Vol. 215, No. 1122, pp. 353–370.) A procedure is described for preparing reproducible specimens of Cu_2O at the oxygen-poor limit of its composition range. Measurements of the electrical conductivity and thermoelectric power of such specimens are reported for the temperature range 20–1 030°C. The curves for both properties consist of two sections, the discontinuity occurring at 355°C in both cases. Below this temperature the thermoelectric power has a constant value; at higher temperatures it decreases and remains characteristic of a positive-hole conductor up to the highest temperatures used. The effect of added oxygen was investigated; the theoretical implications of the results obtained are discussed.

537.311.33(083.72) 1702

Technical Vocabulary relating to Semiconductors.—(*Onde élect.*, Jan. 1953, Vol. 33, No. 310, pp. 62–63.) Definitions (in French) are given of terms included in the Vocabulaire Électrotechnique International Électronique. The definitions were adopted by a sub-committee at Brussels, August 1952, but have still to be approved by national committees.

537.311.33.001.8 1703

Physical Properties of Semiconductors and Application in Radio Technique.—P. Aigrain & C. Dugas. (*Onde élect.*, Jan. 1953, Vol. 33, No. 310, pp. 5–14.)

538.221 1704

Effects on the Magnetization Characteristic of Ferromagnetic Materials carrying Current.—S. Krapf. (*Elektrotech. Z.*, *Edn A*, 1st Dec. 1952, Vol. 73, No. 23, pp. 745–746.) A short account of changes produced in the magnetization curves of ferromagnetic wires by passing auxiliary current through them or subjecting them to mechanical stress. These effects find application particularly in measurement technique.

538.221 1705

The Complex Permeability of a High-Permeability Ferrite Core.—R. Feldtkeller & O. Kolb. (*Z. angew. Phys.*, Dec. 1952, Vol. 4, No. 12, pp. 448–451.) Analysis of the locus-diagram representation of the variation of complex permeability with frequency in Ni-Zn ferrite. The shape of the curve is decided primarily by the main relaxation effect of the ferrite, and to a lesser extent by after-effect, electrical and mechanical resonances, and hysteresis.

538.221 : 538.632 1706

Hall Effect and Temperature in Ferromagnetic Conductors.—J. P. Jan. (*Helv. phys. Acta*, 15th Dec. 1952, Vol. 25, No. 7, pp. 677–700. In French.) Measurements of the variation of the Hall effect in Fe and Ni are reported for the temperature range −190° to +600°C. The magnetization Hall constant was found to depend mainly on the purity of the material. Graphical comparison is made between the temperature variation curves obtained and results published by various workers.

538.221 : 621.318.4 1707

Analysis of Measurements on Magnetic Ferrites.—C. D. Owens. (*Proc. Inst. Radio Engrs*, March 1953, Vol. 41, No. 3, pp. 359–365.) The relations between the magnetic quality factor μQ and the characteristics of coils and transformers are analysed, and the advantages of using the μQ value of the ferrite as a design parameter are indicated.

538.221 : 669.14.018.8-131.2 1708

Magnetic Properties of Stainless Steel Wire. Effect of Cold Work.—W. Sucksmith. (*Metal Treatm.*, Dec. 1952, Vol. 19, No. 87, pp. 545–546, 549.) An account of experiments conducted on wire for magnetic-recording purposes. The increase in coercivity due to heat treatment at 450°–550°C is ascribed primarily to a decrease in size of ferromagnetic inclusions. In cold-worked specimens the coercivity is reduced and longer heat-treatment is required for maximum coercivity to develop.

538.614 : 538.221 **1709**

Theory of the Microwave Permeability Tensor and Faraday Effect in Nonsaturated Ferromagnetic Materials.—G. T. Rado. (*Phys. Rev.*, 15th Jan. 1953, Vol. 89, No. 2, p. 529.) A calculation is made of the microwave Faraday rotation in a ferromagnetic material without introducing the restriction that the material be magnetized to saturation. See also 1233 of 1952 (Hogan).

538.632/.633 **1710**

Corbino Effect and Magnetic Variation of Resistance in Bismuth.—L. Halpern & K. M. Koch. (*Acta phys. austriaca*, Nov. 1951, Vol. 5, No. 1, pp. 129–133 & June 1952, Vol. 5, No. 4, p. 567.) The accuracy of Hall-constant measurements can be increased and information can be obtained about the mechanism of resistance variation in a magnetic field by using specimens of shapes different from the usual strip.

538.652 **1711**

Reversal of Sign of Magnetostriction by Expansion: Part 2.—A. Elsas & E. Vogt. (*Z. Naturf.*, May 1951, Vol. 6a, No. 5, pp. 233–238.)

546.431-31 : [535.37 + 537.58 **1712**

Luminescence and Thermionic Emission of Barium Oxide.—V. L. Stout. (*Phys. Rev.*, 1st Jan. 1953, Vol. 89, No. 1, pp. 310–314.) Two of the six luminescence bands identified varied in intensity with increase of thermionic emission, one decreasing, the other increasing. Peak intensity in both bands varied inversely with temperature.

620.19 : 621.396.6 **1713**

Tests relating to the Resistance of Electrical Equipment to Tropical Conditions.—A. Delrieu. (*Rev. gén. Élect.*, Dec. 1952, Vol. 61, No. 12, pp. 551–559.) Report of laboratory tests on materials and circuit components.

621.314.632 : 546.28 **1714**

Thermal Treatment of Silicon Rectifiers.—L. Esaki. (*Phys. Rev.*, 15th Jan. 1953, Vol. 89, No. 2, pp. 398–399.) Measurements were made on Si point-contact rectifiers heated at temperatures between 900° and 1400°C for various periods. With both *p*- and *n*-type Si, on heating at over 1250°C in vacuo, the rectifier property disappears, but is restored by subsequent heating at 1000°C in an oxygen atmosphere at low pressure. Possible mechanisms responsible for the changes are discussed briefly.

621.314.632 : 546.289 **1715**

Reverse Characteristics of High-Inverse-Voltage Point-Contact Germanium Rectifiers.—J. H. Simpson & H. L. Armstrong. (*J. appl. Phys.*, Jan. 1953, Vol. 24, No. 1, pp. 25–34.) A theoretical investigation is made of the effect of the geometry of the contact and of the high concentration of holes in the 'inversion region' on the field at the contact. The latter factor results in a lowering of the effective barrier height for rectifiers made of very pure material. The former factor gives rise to increases in reverse current resulting from image force and tunnel effect at high voltages. Experimentally determined current/voltage characteristics were of the form predicted by the theory.

621.315.61.029.5 **1716**

Behavior of Insulating Materials at Radio Frequencies.—J. J. Chapman, J. W. Dzimianski, C. F. Miller & R. K. Witt. (*Elect. Mfg*, July 1951, Vol. 48, No. 1, pp. 107–109 . . 238.) Data on the breakdown stress, dielectric constant and dissipation factor at frequencies from 60 c/s to 18 Mc/s are tabulated for 16 modern materials.

621.315.612.4 **1717**

Rare Earth Titanates with a Perovskite Structure.—J. Brous, I. Fankuchen & E. Banks. (*Acta cryst., Camb.*, 10th Jan. 1953, Vol. 6, Part 1, pp. 67–70.) Report of experiments on the production of new materials having interesting dielectric properties. The compounds $EuTiO_3$, $(La,Li)TiO_3$, $(La,Na)TiO_3$, $(La,K)TiO_3$ and $(La,Rb)TiO_3$ having the same crystal structure as $SrTiO_3$ were prepared and their lattice constants determined.

621.315.612.4.011.5 **1718**

Lattice Constants and Dielectric Properties of Barium-Titanate/Barium-Stannate/Strontium-Titanate Bodies.—R. H. Dungan, D. F. Kane & L. R. Bickford, Jr. (*J. Amer. ceram. Soc.*, Dec. 1952, Vol. 35, No. 12, pp. 318–321.) An account, with tabulated results, of investigations of the properties of $BaTiO_3$ ceramics including varying proportions of $BaSnO_3$ and $SrTiO_3$. Substitution of Sn for Ti increases the lattice constant, while substitution of Sr for Ba decreases it. Both substitutions decrease the Curie temperature, which has no direct relation to the lattice constant at room temperature.

621.315.614.6 **1719**

Electrical Properties of Glass-Fiber Paper.—T. D. Callinan, R. T. Lucas & R. C. Bowers. (*Elect. Mfg*, Aug. 1951, Vol. 48, No. 2, pp. 94–97 . . 252.) Experimental values of dielectric strength, dielectric constant, power and loss factors, and volume resistivity are tabulated for papers made from AAA-superfine glass fibres and for glass-fibre papers impregnated with 10 different varnishes.

621.315.616 **1720**

Polyvinyl Chloride.—A. G. Thomson. (*Elect. Times*, 25th Dec. 1952, Vol. 122, No. 3190, pp. 1157–1160.) A review of the properties and applications of PVC, particularly as an insulating material.

621.318.424.042.4.017.3 **1721**

Alternating-Field Losses in Magnetic Cores with Air Gaps.—M. Kornetzki. (*Frequenz*, Nov. 1952, Vol. 6, No. 11, pp. 313–318.) Formulae previously given (3073 of 1950) for the inductance and losses of coils with pot cores are extended so as to apply to coils with an air-gap in the core.

MATHEMATICS

681.142 **1722**

The Use of a 'Floating Address' System for Orders in an Automatic Digital Computer.—M. V. Wilkes. (*Proc. Camb. phil. Soc.*, Jan. 1953, Vol. 49, Part 1, pp. 84–89.)

681.142 **1723**

Modern Mathematical Machines.—L. Biermann & H. Billing. (*Naturwissenschaften*, Jan. 1953, Vol. 40, No. 1, pp. 7–13.) Description of the G1 and G2 digital computers at Göttingen.

681.142 **1724**

Rapid-Access Magnetostatic Store for Electronic Computers.—H. Billing. (*Naturwissenschaften*, Jan. 1953, Vol. 40, No. 2, pp. 49–50.) A system using a matrix arrangement of ferrite ring-cores is described.

681.142 **1725**

Present Development of Programme-Controlled Computers in Germany.—G. Overhoff. (*Phys. Blätter*, Jan. 1953, Vol. 9, No. 1, pp. 31–36.) Details are included of electronic machines under development at the Darmstadt Technische Hochschule and in the Max-Planck-Gesellschaft Institute at Göttingen, and of commercial e.m.-relay machines.

681.142 1726

The Numerograph.—P. Vauthieu. (*Onde élect.*, Dec. 1952, Vol. 32, No. 309, pp. 496–499.) Description of the general principles of a pulse-operated system of displaying on a c.r.o. successive 5-digit numbers provided by a digital computer, the numbers being then photographed on moving film. Recording speed is about 1 000 times that of a teletype machine.

681.142 1727

Numerical Determination of Biharmonic Functions by an Analogue Method using Superposed [resistor] **Networks.**—J. Boscher. (*C. R. Acad. Sci., Paris*, 5th Jan. 1953, Vol. 236, No. 1, pp. 44–46.) Description of the network used and of its application to the determination of Airy's function.

681.142 : [537.222 + 538.242 1728

Matrix Storage Systems.—J. A. Rajchman. (*Onde élect.*, Dec. 1952, Vol. 32, No. 309, pp. 479–491.) Various e.s. and e.m. methods of digit storage on structures of the matrix type are described, with details of their operation and control systems. See also 2064 of 1951 and 3156 of 1952.

681.142 : 621.314.7 1729

Typical Block Diagrams for a Transistor Digital Computer.—J. H. Felker. (*Elect. Engng, N.Y.*, Dec. 1952, Vol. 71, No. 12, pp. 1103–1108.) Detailed discussion of digital-computer circuits in which transistors can be used with advantage.

MEASUREMENTS AND TEST GEAR

538.71 1730

New Rotating-Iron Magnetometer.—H. Gondet. (*J. Rech. Centre nat. Rech. sci.*, Dec. 1952, No. 21, pp. 286–291.) The instrument comprises three ferrite rods, of which two are fixed and aligned while the third rotates in the space between them, round an axis perpendicular to them. The fixed rods carry windings in which is induced an e.m.f. whose value depends on the speed of rotation and on the local magnetic field. The possibility of using the instrument in aircraft is discussed.

621.3.018.41(083.74) + 621.396.91 1731

Improvements in the N.B.S. Primary Standard of Frequency.—(*Tech. News Bull. nat. Bur. Stand.*, Jan. 1953, Vol. 37, No. 1, pp. 8–12.) The N.B.S. primary frequency standard comprises eight 100-kc/s quartz-crystal resonators and nine 100-kc/s crystal-controlled oscillators, three of which are located at the WWV station. The newer oscillators are driven by a current $<100\,\mu$A and show increased short-time stability and overall reliability. The quartz-crystal resonators are used only once a day as part of a balanced-bridge network for comparison with one of the standard oscillators. Driving current for the resonators is $10\,\mu$A. Once a day the frequency of each standard oscillator is compared with that of each crystal resonator and then with one of the standard oscillators as reference. A frequency-difference method is used for the purpose with an electronic frequency counter with a sensitivity of ± 1 part in 10^{10}. Elaborate oven arrangements keep the temperature variations to $<0{\cdot}001$°C. Details of a new automatic announcing system for WWV are also given. See also *Radio & Telev. News, Radio-Electronic Engng Section*, Jan. 1953, Vol. 49, No. 1, pp. 10–11, 30.

621.3.018.41(083.74) : 529.786 1732

"Geared to Greenwich".—R. S. J. Spilsbury. (*Wireless World*, April 1953, Vol. 59, No. 4, p. 167.) Authoritative comment on the Rugby MSF standard-frequency service, with particular reference to the transmission of 1-c/s impulses. See also 1421 and 1423 of May.

621.317.3.029.64 : 621.392.5 1733

The Experimental Determination of Linear Quadripoles for Centimetre Wavelengths.—H. Oppitz. (*Acta phys. austriaca*, Dec. 1951, Vol. 5, No. 2, pp. 214–236.) Weissfloch's transformation law (711, 2083 and 2084 of 1943) for linear reversible loss-free quadripoles is discussed, and a proof is given using quadripole matrices. An experimental method for determining quadripole characteristics, based on this law, requires the use of a terminal piston providing a perfect short-circuit for a coaxial line. An investigation is made of the effect of piston losses on the accuracy of the determinations. A method is described for measuring the apparent impedance of the piston; results are compared with those obtained by the usual method.

621.317.335.3.029.64 1734

A Simple Method for determining the Dielectric Constant of Fluids at Centimetre Wavelengths.—F. Reder & E. Hafner. (*Acta phys. austriaca*, Dec. 1951, Vol. 5, No. 2, pp. 189–201.) An improved coaxial-cylinder method is described. Only a relatively small quantity of the test material is required; the accuracy is to within 0·5%. Results are tabulated for benzene, toluene, carbon tetrachloride and paraffin oil. See also 1664 of 1952 (Ledinegg et al.).

621.317.34 : 621.396.645 1735

An Instrument for Measuring Complex Voltage Ratios in the Frequency Range 1–100 Mc/s.—G. Thirup. (*Philips tech. Rev.*, Sept./Oct. 1952, Vol. 14, Nos. 3/4, pp. 102–114.) Full description of an instrument designed for research on the stability of wide-band feedback amplifiers. The input and output h.f. voltages are converted to a fixed i.f. of 0·3 Mc/s and one of the i.f. voltages is adjusted in amplitude and phase to cancel the other; the phase-shifting device used for this purpose comprises a metal tube containing crossed loops providing a rotating magnetic field, and also a pickup loop. Balance is indicated by the disappearance of a beat note from a loudspeaker. Accuracy is to within 0·2 db in respect of amplitude and 2° in respect of phase angle.

621.317.35 : 621.397.5 1736

A Television-Waveform Monitor.—J. E. Attew. (*Electronic Engng*, March 1953, Vol. 25, No. 301, pp. 106–107.) Complete circuit details are given of a monitor for examination of television line, frame and synchronizing-signal waveforms. Construction details are included for the delay line used in the pulse-duration calibrator.

621.317.353.3 : 621.3.018.78 1737

Intermodulation Distortion; its Significance and Measurement.—E. W. Berth-Jones. (*J. Brit. Instn Radio Engrs*, Jan. 1953, Vol. 13, No. 1, pp. 57–63.) Discussion indicates that in the majority of cases intermodulation testing has little advantage over the simpler total-harmonic testing methods. For the special cases in which intermodulation test methods are justified, great care must be taken in the design of equipment to ensure reasonably accurate results. In particular, phase shifts in filters may affect peak values considerably, and results mean little on an absolute basis unless full details of test conditions are quoted. Typical intermodulation test equipment is described.

621.317.361.029.3 1738

An Accurate Method of Measuring Frequency in the Audio Range.—A. F. B. Nickson. (*J. sci. Instrum.*, Dec. 1952, Vol. 29, No. 12, pp. 391–393.) The unknown frequency is compared directly with a neighbouring frequency given by a v.f.o. locked to a suitable harmonic of a crystal-controlled oscillator. The comparison is effected by means of Lissajous figures on the screen of a c.r.o.

621.317.373 : 621.317.755 1739

A Note on Phase-Angle Measurements using a Cathode-Ray Tube.—F. A. Benson & M. S. Seaman. (*Electronic Engng*, March 1953, Vol. 25, No. 301, p. 100.) A method of eliminating measurement errors due to incorrect location of the horizontal reference line on the c.r.o. screen is described. A double-beam oscillograph is used to generate two ellipses, the intersections of which determine accurately the position of the reference line. See also 1965 of 1950 (Benson & Carter).

621.317.373 : 621.317.755 1740

Phase-Angle Measurements.—F. A. Benson. (*Wireless World*, April 1953, Vol. 59, No. 4, p. 157.) Comparison of errors due to trace width in three methods of measurement. See also 1965 of 1950 (Benson & Carter).

621.317.373.029.6 1741

Application of the Microwave Homodyne.—F. L. Vernon, Jr. (*Trans. Inst. Radio Engrs*, Dec. 1952, No. PGAP-4, pp. 110–116.) The principle of homodyne detection is explained and its application in the measurement of the phase difference between two microwave signals is described. The method is effective over a wide range of signal amplitudes.

621.317.41 1742

Measurement of the Constants of Ferromagnetic Materials by means of Transmission Lines (80–4 000 Mc/s).—A. Hersping. (*Frequenz*, Dec. 1952, Vol. 6, No. 12, pp. 345–356.) A short length of a straight coaxial measurement line is completely filled with the material to be tested, such as a nonmetallic ferrite. The front portion of the line is slotted to permit probe measurements along the line. The ferrite-filled section acts as an attenuator with a complex impedance and propagation constant, formulae for which involve the line parameters and the magnetic and dielectric loss angles. The line is fed through a shielded cable and the permeability, dielectric constant and loss angles are determined from measurements of the no-load and short-circuited impedance of the line. Detailed theory of the method is given. The range 80–600 Mc/s is covered using the Rohde & Schwarz Type-BN3916 measurement line, smaller equipment, details of which are given, being used for the range 600–4 000 Mc/s. Measurement results for Ni-Zn and Cu-Zn ferrites are shown graphically.

621.317.42 : 621.385.833 1743

Pendular System for Electrodynamic Measurement of the Axial Field of a Magnetic Electron Lens.—P. Durandeau. (*C. R. Acad. Sci., Paris*, 26th Jan. 1953, Vol. 236, No. 4, pp. 366–368.) Description of an accurate method based on the displacement of a compensated system of two solenoids.

621.317.421 1744

Null Method for the Precision Measurement of Magnetic Induction.—R. Tenzer. (*Arch. Elektrotech.*, 1952, Vol. 40, No. 7, pp. 406–421.) A method developed for investigating the constancy of permanent magnets is described. The voltage pulse generated in a small coil on withdrawing it from the magnet is opposed by another pulse of the same magnitude derived by mutual induction on opening or closing a primary circuit. A fluxmeter rather than a ballistic galvanometer is used as the null instrument. Relevant theory is given and system errors are discussed.

621.317.71 1745

A Direct Current Microampere Integrator.—R. N. Schweiger. (*Rev. sci. Instrum.*, Dec. 1952, Vol. 23, No. 12, pp. 735–738.) A feedback-stabilized current amplifier feeds the current coil of a modified Type C-6 Thompson watt-hour meter. The no-load speed of this meter is directly proportional to the current supplied, so that the value of the integrated current is given on the scale of a revolution counter operating at nearly zero torque. The nett accuracy is to within about 1% of full scale. Seven ranges are provided, with full-scale readings ranging from 10 μA to 1 mA.

621.317.725 1746

Nonlinearity in a Voltmeter using Cathode-Follower and Thermocouple.—D. G. Tucker. (*J. sci. Instrum.*, Jan. 1953, Vol. 30, No. 1, pp. 11–13.) The arrangement considered comprises a galvanometer connected to a thermocouple whose heater is used as the cathode load of a cathode follower; it is useful for comparing an alternating voltage with a direct-voltage standard. Analysis shows that the effect of valve nonlinearity is usually so small that the calibration error is $<0{\cdot}1\%$ for a voltage of about 0·8 V.

621.317.73.029.6 : 621.392.43 1747

New Equipment for Impedance Matching and Measurement at Very High Frequencies.—A. Bloch, F. J. Fisher & G. J. Hunt. (*Proc. Instn elect. Engrs*, Part III, March 1953, Vol. 100, No. 64, pp. 93–99.) A section of low-loss coaxial line, which may be only a third of the length of the equivalent slotted line, is fitted with three fixed probes to measure relative voltage amplitudes. From these ratios an unknown impedance loading the line can be found, either analytically or graphically, each ratio defining a circular locus on the Smith chart. The equipment is particularly suitable for field tests; it has the decided advantage that adjustment of a load to a prescribed value is shown by the simultaneous zero reading of two meters. The frequency range covered is about 3 : 1. This can be more than doubled if the two outer probes are movable; such an arrangement has the advantage that a single chart can be used for all the frequencies covered.

621.317.733 : [621.316.86 : 537.312.6 1748

Use of Thermistors in a R.F. Bridge for Measurement of Low Admittances.—M. Soldi. (*Alta Frequenza*, Dec. 1952, Vol. 21, No. 6, pp. 243–259.) A bridged-T circuit is used, with thermistors as variable resistance standards (see 2134 of 1952). The thermistors are simultaneously regulated and calibrated by means of a special automatic d.c. bridge. Results obtained with an experimental model confirm the theoretically predicted high sensitivity and good accuracy of the arrangement.

621.317.733 : [621.316.86 : 537.312.6 1749

The Design of a Direct-Reading Thermistor Bridge with Temperature Compensation.—R. M. Pearson & F. A. Benson. (*Electronic Engng*, Feb. 1953, Vol. 25, No. 300, pp. 51–57.) Characteristic data are given for thermistors of British manufacture, of both bead and disk types. The data are used to design a bridge in which power at a wavelength of 3 cm is measured in terms of the change of resistance of a thermistor on to which the power is directed. Effects due to variations of ambient temperature are balanced out by means of compensating thermistor networks arranged in parallel with the bridge supply source and in series with the meter.

621.317.74 : 621.397.2 1750

Apparatus for the Measurement of Phase Delay in Television Transmission Circuits and in Associated Equipment.—C. W. Goodchild & R. C. Looser. (*Proc. Instn elect. Engrs*, Part IIIA, 1952, Vol. 99, No. 20, pp. 788–795. Discussion, pp. 860–866.) A waveform rich in harmonics of a convenient fundamental frequency is applied to the circuit under test, the output from which is compared with a similar reference waveform, of opposite polarity, by means of a calibrated delay unit. The frequency range covered is 100 kc/s–10 Mc/s. For delays $<1\ \mu s$ the error is $<\pm 0{\cdot}01\ \mu s$.

621.317.76.089.6 1751

Wide-Range Frequency Calibrator.—J. F. Sterner. (*Radio & Telev. News, Radio-Electronic Engng Section*, Jan. 1953, Vol. 49, No. 1, pp. 12–13, 31; *Proc. nat. Electronics Conf., Chicago*, 1952, Vol. 8, pp. 831–835.) Description of simple equipment providing subharmonics and harmonics of a 2·5-Mc/s crystal throughout the range 0·25–250 Mc/s. An *LC* multivibrator, locked to the crystal-controlled oscillator, provides harmonics of its 0·25-Mc/s fundamental frequency whose amplitudes are increased by means of a 1-Mc/s 'bumper' circuit in the cathode lead.

621.317.784.029.51 1752

High-Frequency Power Meter with Dry Rectifier.—H. Wilde. (*Arch. tech. Messen*, Jan. 1953, No. 204, pp. 5–6.) Description of an instrument using a dynamometer type of indicator, with phase control of the rectified voltages applied to the coils. The frequency limit is about 200 kc/s; power consumption is $<0{\cdot}2$ W on the lowest power range of 0–8 W.

621.317.79 : 621.397.8 1753

A Variable-Definition Camera Channel for the Appraisal of Television Standards.—A. V. Lord & C. B. B. Wood. (*Proc. Instn elect. Engrs*, Part IIIA, 1952, Vol. 99, No. 20, pp. 811–820. Discussion, pp. 860–866.) Description of equipment which can be switched rapidly between six preselected line-scanning standards in the range of line frequency 10–32 kc/s, the frame frequency in all cases being that of the mains. The more interesting units are described in detail and results obtained with the equipment are reported.

621.397.2.001.4 1754

The Testing of Television Transmission Systems by means of Square Waves.—J. Müller. (*Funk u. Ton*, Dec. 1952, Vol. 6, No. 12, pp. 617–631.) Description of the method, with two practical examples illustrating the connection between transient response and transfer function. See also 209 and 1879 of 1952.

621.397.24.001.4 1755

The Specification and Testing of Television Wire-Broadcasting Systems.—F. Hollinghurst & D. S. Tod. (*Proc. Instn elect. Engrs*, Part IIIA, 1952, Vol. 99, No. 20, pp. 680–695. Discussion, pp. 757–760.) Technical conditions and recommended standards of performance specified for such systems are discussed and test methods described. The main features of eight experimental or working systems are tabulated and results of performance tests on four systems are presented.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

534.321.9 : 620.179.1 1756

Efficient Small Apparatus for Ultrasonic Testing of Materials with Barium-Titanate Oscillator.—V. Met & E. Skudrzyk. (*Elektrotech. u. Maschinenb.*, 1st Dec. 1952, Vol. 69, No. 23, pp. 519–523.) Description and circuit details of equipment including a specially damped $BaTiO_3$ disk operated below its natural frequency, pulse recurrence frequencies ranging from 25 to 200/sec.

534.321.9 : 639.245.1 1757

Ultrasonic Equipment for Locating Whales.—(*Engineering, Lond.*, 16th Jan. 1953, Vol. 175, No. 4538, pp. 90–91.) Pulse-reflection apparatus is described using a narrow horizontal beam with a maximum range of about 2 000 yards. The generator is of magnetostriction type, operating at 14 or 25 kc/s.

535.33.071 : 621.383.4 1758

A Simple High-Speed Spectrometer for the Infrared Region.—D. A. H. Brown & V. Roberts. (*J. sci. Instrum.*, Jan. 1953, Vol. 30, No. 1, pp. 5–8.) A single-prism spectrometer is described which uses a rapid-response photoconductive detector together with c.r.o. presentation at a repetition rate of 150 per sec. Using a PbTe cell, the long-wave limit is about $5{\cdot}5\ \mu$.

535.822.9 : 621.397.611.2 1759

The Flying-Spot Microscope.—F. Roberts & J. Z. Young. (*Proc. Instn elect. Engrs*, Part IIIA, 1952, Vol. 99, No. 20, pp. 747–757. Discussion, pp. 757–760.) For another account see 1733 of 1951.

621.316.7 1760

Calculation of Control Systems having Given Limits of Control Error.—D. Ströle. (*Arch. elekt. Übertragung*, Jan. & Feb. 1953, Vol. 7, Nos. 1 & 2, pp. 37–46 & 107–116.) The criteria for assessing a control system are that the damping and the maximum and final deviations of the system under control shall not exceed given limits. The problem is treated in terms of frequency response. A practical method of designing the control system is developed and is illustrated by means of two examples.

621.316.7 1761

Stabilization of Nonlinear Feedback Control Systems.—R. L. Cosgriff. (*Proc. Inst. Radio Engrs*, March 1953, Vol. 41, No. 3, pp. 382–385.)

621.317.083.7 1762

Radio Telemetering.—E. D. Whitehead & J. Walsh. (*Proc. Instn elect. Engrs*, Part III, March 1953, Vol. 100, No. 64, pp. 45–56. Discussion, pp. 57–59.) Basic principles of present-day telemetry systems are discussed, together with the characteristics of various types of modulation and methods of multiplexing. Particular systems are outlined and a 6-channel frequency-multiplex f.m./a.m. system developed for aerodynamic research is described in some detail.

621.383.001.8 : 778.37 1763

An Industrial Instrument for the Observation of Very-High-Speed Phenomena.—M. S. Richards. (*Proc. Instn elect. Engrs*, Part IIIA, 1952, Vol. 99, No. 20, pp. 729–746. Discussion, pp. 757–760.) Detailed description of equipment designed in conjunction with an experimental image-converter tube. Repetitive exposures down to $1\ \mu s$ are available whose repetition rate can be varied between 12/sec and 10^5/sec for stroboscopic viewing. Single exposures down to $0{\cdot}05\ \mu s$, for recording transient phenomena, can also be obtained.

621.384.611/.612 1764

Synchrocyclotron for 450-MeV Protons.—H. L. Anderson, J. Marshall, L. Kornblith, Jr, L. Schwarcz & R. Miller. (*Rev. sci. Instrum.*, Dec. 1952, Vol. 23, No. 12, pp. 707–728.) Detailed description of the particle accelerator recently completed at the Institute for Nuclear Studies, Chicago University.

621.384.611 1765

Electric Fields within Cyclotron Dees.—R. L. Murray & L. T. Ratner. (*J. appl. Phys.*, Jan. 1953, Vol. 24, No. 1, pp. 67–69.)

621.384.622.1 1766

A New Focusing Principle applied to the Proton Linear Accelerator.—J. S. Bell. (*Nature, Lond.*, 24th Jan. 1953, Vol. 171, No. 4343, pp. 167–168.) A modification of the drift tube is described by which suitable electric or magnetic fields can be applied transverse to the particle motion so that particles converge strongly in one direction and diverge in the direction at right angles. Overall focusing results if the field directions are interchanged periodically along the accelerator. See also 1455 of May (Blewett).

621.385.83 : 538.691 1767

Determination of an Electron Trajectory by Successive Integrations.—É. Durand. (*C. R. Acad. Sci., Paris*, 26th Jan. 1953, Vol. 236, No. 4, pp. 364–366.) A method of integration of second-order linear equations is adapted for systematic calculations in electron optics.

621.385.833 1768

The Scanning Electron Microscope and the Electron-Optical Examination of Surfaces.—D. McMullan. (*Electronic Engng*, Feb. 1953, Vol. 25, No. 300, pp. 46–50.) A conventional two-stage e.s. electron microscope has been modified to operate as a scanning microscope by removal of the projector lens and replacement of the fluorescent screen by a unit comprising deflection coils, e.s. lens, specimen stage and electron multiplier; the picture is reconstituted in a separate c.r. tube. For direct viewing, the system uses 405 lines (interlaced) per picture and about 0·9 pictures/sec; for recording, with beam current reduced to give maximum resolving power, the system uses 550 lines per picture and 1 picture in 300 sec. It is possible to obtain pictures of surfaces directly without making replicas.

621.385.833 1769

Electrostatic Lenses for Focusing Very-High-Energy Particles.—M. Y. Bernard. (*C. R. Acad. Sci., Paris*, 12th Jan. 1953, Vol. 236, No. 2, pp. 185–187.) The type of lens described, which is similar to that proposed recently by Courant, Livingston & Snyder (1454 of May), uses two sets of electrodes in line. Each set comprises four nearly quadrantal sections of a cylinder, two opposite sections being maintained at a positive potential and the other two at an equal negative potential. One set is rotated 90° with respect to the other. With such systems, ions with energies of several megavolts can be focused, using very much lower voltage sources.

621.385.833 1770

Experimental Study of Electrostatic Immersion Objective with Plane Electrodes: Extractor Field E_0 under Conditions of Focusing. Geometrical Aberrations off the Axis.—A. Septier. (*C. R. Acad. Sci., Paris*, 5th Jan. 1953, Vol. 236, No. 1, pp. 58–60.)

621.385.833 : 061.3 1771

Summarized Proceedings of a Conference on Electron Microscopy — Bristol, September 1952.—V. E. Cosslett, J. Nutting & R. Reed. (*Brit. J. appl. Phys.*, Jan. 1953, Vol. 4, No. 1, pp. 1–5.) Summaries are given of papers and discussion on design, operating technique and applications.

PROPAGATION OF WAVES

538.566.029.45/.51 : 551.594.6 1772

The Propagation of Very Long Electromagnetic Waves and of [waves due to] Lightning Discharges round the Earth.—W. O. Schumann. (*Z. angew. Phys.*, Dec. 1952, Vol. 4, No. 12, pp. 474–480.) See 802 of March.

621.396.11 1773

Ionospheric Propagation.—R. Gea Sacasa. (*Rev. Telecomunicación, Madrid*, Dec. 1952, Vol. 8, No. 30, pp. 2–10. In Spanish, French and English.) Documents discussed at Stockholm in May 1952 by the C.C.I.R. committee on ionospheric propagation are analysed in detail to show that the accuracy of predictions made by the 'Spanish method' (3536 of 1952) compares favourably with that of predictions made by British and U.S. methods.

621.396.11 1774

Investigations of the Propagation of Space Waves: Part 2.—W. Budde. (*Z. Naturf.*, May 1951, Vol. 6a, No. 5, pp. 238–242.) Measurements made during the period October 1944 to January 1945 over the path Ismaning (Munich)–Kölby (Denmark) are discussed; the frequency used was 9·83 Mc/s. Short-term fluctuations of azimuth angle were observed. The fluctuations decreased from a fairly high early-morning value to a minimum around midday and then increased to a maximum during the afternoon, decreasing again in the early evening. No seasonal variation of the mean fluctuations was observed.

621.396.11 1775

Determination of the Sense of Polarization of the Magneto-ionic z-Component.—B. Landmark. (*Tellus*, Nov. 1952, Vol. 4, No. 4, pp. 319–323.) See 2867 of 1952.

621.396.11 1776

The Diffraction of Radio Waves by the Curvature of the Earth.—M. H. L. Pryce. (*Advances Phys.*, Jan. 1953, Vol. 2, No. 5, pp. 67–95.) A simplified approximate derivation is given of the field at moderate heights and distances due to an oscillating dipole at moderate height, taking into account the curvature and electrical constants of the earth. The treatment is approximately equivalent to assuming the earth to be flat and the wave path to be curved upwards, and is mathematically convenient because Fourier integrals can be used instead of expansions in Legendre functions. The method applies to both vertically and horizontally polarized waves.

621.396.11 1777

Tables of Functions occurring in the Diffraction of Electromagnetic Waves by the Earth.—C. Domb. (*Advances Phys.*, Jan. 1953, Vol. 2, No. 5, pp. 96–102.) In a previous paper [800 of 1948 (Domb & Pryce)] curves and formulae were given which enabled field strength to be calculated with fair accuracy; the approximation of the first term of the differential series was used for the region well beyond the optical range. Details are now given of relevant tables of functions to enable more accurate calculations to be made and the region of validity of the one-term approximation to be assessed. See also 1776 above.

621.396.11 : 551.510.535 1778

Effect of the Geomagnetic Field on the Absorption of Short Waves in the Ionosphere (Vertical Incidence).—Argence, Rawer & Suchy. (See 1678.)

621.396.11 : 551.510.535 1779

The Reflection and Absorption of Radio Waves in the Ionosphere.—Piggott. (See 1679.)

621.396.11 : 551.510.535 1780

Gradient Reflections from the Atmosphere.—J. Feinstein. (*Trans. Inst. Radio Engrs*, Dec. 1952, No. PGAP-4, pp. 2–13.) Analysis is presented showing the difference between (*a*) the effect on the field far beyond the horizon of a refractive-index gradient with no discontinuities in the derivatives of any order, and (*b*) the effect of a gradient having a discontinuity in a particular derivative. A theory of scattering from regions of random gradient of refractive index is developed, under the assumption of complete horizontal stratification of the atmosphere.

621.396.11.029.51/.53 **1781**

A Review of Present Knowledge of the Ionospheric Propagation of Very-Low, Low- and Medium-Frequency Waves.—F. A. Kitchen, B. G. Pressey & K. W. Tremellen. (*Proc. Instn elect. Engrs*, Part III, March 1953, Vol. 100, No. 64, pp. 100–108.) Observations at frequencies <3 Mc/s and for ranges mainly >500 km are reviewed. The results for frequencies <300 kc/s are summarized under the headings of field strength, phase and polarization measurement, the effects of ionospheric disturbances, the effects of local ionospheric characteristics, and theoretical studies. Propagation investigations on medium frequencies prior to 1937 were adequately summarized in an I.R.E. report (27 of 1939), so that only work done since that date is considered here. Present knowledge of the subject is summarized and suggestions are made for further study. 70 references.

621.396.11.029.51 **1782**

The Measurement of the Phase Velocity of Ground-Wave Propagation at Low Frequencies over a Land Path. —B. G. Pressey, G. E. Ashwell & C. S. Fowler. (*Proc. Instn elect. Engrs*, Part III, March 1953, Vol. 100, No. 64, pp. 73–84.) An account of measurements of the change of phase of a 127·5-kc/s wave with distance along the irregular 177-km land path between the Decca stations at Lewes (Sussex) and Warwick. Mobile equipment was used at 25 points along the path to measure the phase difference of the signals from the two transmitters. The results (accurate to within 0·25°) were used to plot a curve showing the deviation of the measured phase difference from that calculated on the assumption of a propagation velocity equal to that in free space. The mean velocity deduced for the path was 299 230 ± 12 km/s. No definite evidence was found of any ground-contour effect on the phase, but the effect of the nature of the ground was very marked, a high velocity being obtained over ground of good conductivity and a low velocity over ground of poor conductivity.

621.396.11.029.55(98) **1783**

Ionospheric Propagation of Decametre Waves in the Arctic Polar Regions.—J. Bouchard. (*C. R. Acad. Sci., Paris*, 12th Jan. 1953, Vol. 236, No. 2, pp. 220–222.) Discussion of reception in France of signals mainly from stations on the Pacific coast of North America, the frequencies concerned ranging from 12 to 17 Mc/s. Distortion on these frequencies is at times so great that the signals are unreadable. The distortion appears to be due to ionospheric turbulence in the arctic polar regions, and related to geomagnetic disturbance. The distortion of signals whose transmission paths involve reflection at points in the arctic regions can be attributed to fluctuations of the critical frequency at the reflection points. In some cases, when the signals follow a great-circle path with a considerable length in the arctic regions, similar distortion of signals from the antipodes occurs, but when the signals travel by other paths, no great distortion is observed. Correlation has been established between the received field-strength and distortion magnitude and the data on geomagnetic variations published by the French Ionospheric Bureau, and it is known that the corresponding variations in the zone of maximum auroral activity are 4 or 5 times greater than in France.

621.396.11.029.6 : 551.510.52 **1784**

Meteorological Effects on V.H.F. Propagation.—K. Toman, W. G. Albright & E. C. Jordan. (*Trans. Inst. Radio Engrs*, Dec. 1952, No. PGAP-4, pp. 20–30.) Continuous records of the signal strength of several f.m. and TV stations in the Chicago area have been obtained during a period of 18 months at Urbana, Illinois, about 127 miles away. Correlation is discussed between the observed signal strengths and meteorological data obtained at two radiosonde stations along the transmission path. Low values of signal strength are obtained when the distribution of the effective dielectric constant ϵ of the troposphere is nearly linear, with a gradient corresponding to a well-mixed atmosphere. Medium and high signal levels result from larger surface gradients of ϵ and from larger gradients at certain heights above ground. Very high signal levels are obtained when the trapping conditions of duct propagation are satisfied.

621.396.11.029.63 **1785**

Propagation Characteristics of Microwave Optical Links.—L. G. Trolese, J. P. Day & R. U. F. Hopkins. (*Trans. Inst. Radio Engrs*, Dec. 1952, No. PGAP-4, pp. 31–36.) A report of tests, at frequencies in the band 1·7–1·85 kMc/s, on two paths near San Francisco, the transmitter being on Mt. Diablo at a height of 3 800 ft and the receivers respectively at heights of 7–65 ft and 165 ft above sea level. Owing to the height of the transmitter, the path difference between the direct and the ground-reflected component varies rapidly with the height of the receiving aerial, so that closely spaced maxima and minima occur in the signal-strength/aerial-height curve. The measure of the path difference in wavelengths also varies rapidly with frequency, so that the maxima and minima for frequencies of 1·7 and 1·85 kMc/s do not correspond, an aerial at a height of 55 ft giving maximum signals on 1·7 kMc/s and minimum signals on 1·85 kMc/s, essentially the same response on the two frequencies being obtained for an aerial height of 7 ft, where the first maximum occurs for both frequencies and for which less fading is observed.

621.396.11.029.63 **1786**

Variation of Field Intensity over Irregular Terrain within Line of Sight for the U.H.F. Band.—H. Fine. (*Trans. Inst. Radio Engrs*, Dec. 1952, No. PGAP-4, pp. 53–65.) Statistical analysis of available data from field-strength surveys enabled empirical propagation formulae to be deduced for the whole u.h.f. television band from 300 to 900 Mc/s, to assist the Federal Communications Commission in the problem of frequency allocation for u.h.f. television broadcasting stations.

621.396.11.029.64/.65 **1787**

Microwave Radio Reflection from Ground and Water Surfaces.—A. W. Straiton. (*Trans. Inst. Radio Engrs*, Dec. 1952, No. PGAP-4, pp. 37–45.) Reflection measurements for wavelengths from 0·86 to 26·5 cm and angles within 5° of grazing incidence showed that the reflection coefficient increased with wavelength and decreased with increase of the angle between the beam direction and the ground or water surface. The 0·86-cm signals were very sensitive to roughness of the reflecting surface. The reflection coefficient was greater for horizontal than for vertical polarization.

RECEPTION

621.396.4 : 621.396.619.16 : 621.396.621 **1788**

Channel Separation in Pulse-Phase Modulation (P.P.M.).—K. Steinbuch. (*Fernmeldetech. Z.*, Dec. 1952, Vol. 5, No. 12, pp. 535–538.) An outline description of separators for multichannel p.p.m. telephony systems, and an examination of their relative cost.

621.396.621.029.62 : 621.396.82 **1789**

Noise Performance of V.H.F. Receivers.—E. G. Hamer. (*Electronic Engng*, Feb. 1953, Vol. 25, No. 300, pp. 68–71.) The frequency distribution of different types of noise is shown graphically, and receiver design is discussed in relation to the reduction particularly of man-made interference and of noise inherent in the receiving system itself, including the aerial. A convenient method of assessing receiver noise performance makes use of a diode as a calibrated noise generator; a miniature diode with associated disk resistor for measurements at frequencies up to 500 Mc/s is illustrated. The additional cost of a special low-noise-factor design may be justified for services operating at frequencies >200 Mc/s.

621.396.621.54.029.64 : 621.396.822 **1790**

The Noise Factor of Centimetric Superheterodyne Receivers.—J. H. Evans. (*Electronic Engng*, March 1953, Vol. 25, No. 301, pp. 98–100.) The noise contributions from the various units of a receiver comprising a crystal frequency changer, klystron local oscillator, and i.f. amplifier, are discussed and curves are given which show the effect on the receiver noise factor of the noise in the local oscillator and the i.f. amplifier. Methods of reducing local-oscillator noise are noted and their attendant disadvantages are considered briefly.

621.396.622.63 : 621.396.621.54 **1791**

Harmonic Mixing and Distortion with Crystal Diodes.—H. F. Mataré. (*Arch. elekt. Übertragung*, Jan. 1953, Vol. 7, No. 1, pp. 1–15.) Basic theory is outlined for heterodyning with oscillator harmonics and for the distortion resulting from curvature of the diode characteristic. The principal properties of mixers, viz. slope of rectifier characteristic, conversion slope and optimum conversion efficiency, are shown in families of curves for two basic forms of presentation of characteristics. Distortion may be caused by pure impedance variations. From the curves it is possible to estimate the losses with harmonic mixing, or the efficiency in the presence of distortion. In this respect Ge diodes are better than Si diodes. See also 2824 of 1951.

STATIONS AND COMMUNICATION SYSTEMS

621.39.001.11 **1792**

Technical Characteristics related to Information. Application to Telecommunications and [radar] **Detection.**—J. Maillard. (*Onde élect.*, Dec. 1952, Vol. 32, No. 309, pp. 500–514.) The terms 'quantity of information' and 'rate of information' are defined and basic relations between them are discussed. These relations enable comparison to be made between the efficiency of the different methods used for the transmission of information, in particular, between the different methods of modulation, whose advantages and disadvantages are here discussed. The use of e.m. or acoustic waves for the detection of objects is also considered in the light of information theory. The analysis indicates that though f.m. equipment appears to have theoretical advantages, pulse equipment has in most cases technical and practical advantages.

621.396.41 : 621.396.619.16 **1793**

Time-Division Multiplex Systems.—J. E. Flood. (*Electronic Engng*, Jan.–April 1953, Vol. 25, Nos. 299–302, pp. 2–5, 58–63, 101–106 & 146–150.) A series of articles forming an introduction to the subject. Part 1 describes different forms of pulse modulation which can be used and discusses features common to different time-division multiplex systems. Part 2 deals in detail with p.a.m., part 3 with p.w.m. and p.p.m., and part 4 with p.c.m. 94 references.

621.396.44 : 621.315.052.63 : 621.396.822 **1794**

The Reduction of Radiation from Carrier Communication Circuits on Overhead Power Lines.—E. P. L. Westell. (*J. Instn Engrs Aust.*, Dec. 1952, Vol. 24, No. 12, pp. 213–218. Discussion, pp. 218–219.) An account of measurements of noise voltages on (*a*) a single-channel a.m. carrier system on a 132-kV power line, (*b*) a multichannel s.s.b. suppressed-carrier system, with suggestions as to practical means of reducing the r.f. interference from such systems.

621.396.5 : 621.396.8 **1795**

150–3 700 Mc/s Performance Tests.—W. R. Young, Jr. (*Commun. Engng*, Jan./Feb. 1953, Vol. 13, No. 1, pp. 15–18, 34.) See 823 of March.

621.396.65 : 621.396.8 **1796**

Intermodulation Interference in Radio Systems.—W. C. Babcock. (*Bell Syst. tech. J.*, Jan. 1953, Vol. 32, No. 1, pp. 63–73.) Formulae are derived for the number of third- and fifth-order intermodulation products liable to be formed within a continuous frequency band comprising a number of channels. Calculations are made of the probability of encountering interference in channels selected at random from this band, and of the number of interference-free channels that can be obtained by careful selection.

621.396.65 : 621.396.93 **1797**

Frequency Economy in Mobile-Radio Bands.—K. Bullington. (*Bell Syst. tech. J.*, Jan. 1953, Vol. 32, No. 1, pp. 42–62.) Factors affecting the suitability of different channels for mobile radio services are discussed, and estimates are made of the number of usable channels per Mc/s for some present and proposed methods of operation, taking into account the possibilities of geographical and operational coordination as well as circuit selectivity.

621.396.933 : 621.396.82 **1798**

Ground-to-Air Cochannel Interference at 2·9 kMc/s.—P. L. Rice, W. V. Mansfield & J. W. Herbstreit. (*Trans. Inst. Radio Engrs*, Dec. 1952, No. PGAE-6, pp. 1–10.) Estimates are made of the average interference-limited coverage to be expected for a ground station using a Bendix Type-ASR-1 aerial with vertical polarization and operating at 2·9 kMc/s. Coverage is defined in terms of protection ratios of 0, 6, 12 and 18 db with respect to interfering transmissions from a similar station 10, 50, 100 or 150 miles distant.

621.396.97 + 621.397.61].029.62 : 061.3 **1799**

The Technical Principles of the Allocation of Frequencies at the European Broadcasting Conference of 1952 in Stockholm.—F. Kirschstein. (*Fernmeldetech. Z.*, Dec. 1952, Vol. 5, No. 12, pp. 563–567.) For other accounts see 3560 of 1952.

SUBSIDIARY APPARATUS

621-526 **1800**

The Principle of a Servo-Type Mechanism requiring Variable Elements.—R. Drenick. (*Proc. Inst. Radio Engrs*, March 1953, Vol. 41, No. 3, pp. 373–377.)

621-526 : 621.316.7 **1801**

The Analysis of Sampled-Data Systems.—J. R. Ragazzini & L. A. Zadeh. (*Elect. Engng, N.Y.*, Dec. 1952, Vol. 71, No. 12, p. 1102.) Digest only. The Laplace and Fourier transform methods of analysis for sampled-data systems are unified and extended. Formulae representing the input-output relations of such systems in the frequency domain are derived.

621.314.634.011.4 **1802**

Capacitance of Selenium Rectifiers.—W. Oldekop. (*Z. Phys.*, 17th Dec. 1952, Vol. 134, No. 1, pp. 66–77.) In a.c. measurements on Se rectifiers, a temperature-dependent phase shift between current and voltage is produced. Representation of the rectifier by an equivalent circuit consisting of a resistor and parallel capacitor affords a satisfactory explanation of this. Discussion shows that a negative component of capacitance is concerned which decreases with decreasing current and with increasing frequency. The abnormal decrease of capacitance with high reverse voltages observed by Hoffmann (941 of 1951) is in agreement with this. See also 3566 of 1952 (Schottky).

621.316.722 : 621.311.6 : 621.387.422 **1803**

A Corona Stabilizer E.H.T. Supply for Proportional Counters.—P. Holton & J. Sharpe. (*Electronic Engng*, Feb. 1953, Vol. 25, No. 300, pp. 63–65.)

621.316.722.1 **1804**

An Electronic Control Device.—W. Hartel. (*Elektrotech. Z., Edn A*, 11th Dec. 1952, Vol. 73, No. 24, pp. 769–771.) Description of a high-precision voltage regulator suitable for a rotary generator.

621.316.722.1 **1805**

Cold-Cathode Voltage Stabilizer.—G. O. Crowther. (*Electronic Engng*, March 1953, Vol. 25, No. 301, p. 127.) Comment on 554 of February (Goulding).

TELEVISION AND PHOTOTELEGRAPHY

621.397.2 : 621.317.74 **1806**

Apparatus for the Measurement of Phase Delay in Television Transmission Circuits and in Associated Equipment.—Goodchild & Looser. (See 1750.)

621.397.2.001.4 **1807**

The Testing of Television Transmission Systems by means of Square Waves.—Müller. (See 1754.)

621.397.24 **1808**

Television Distribution by Wire.—P. Adorian. (*Proc. Instn elect. Engrs*, Part IIIA, 1952, Vol. 99, No. 20, pp. 665–672. Discussion, pp. 757–760.) Discussion of systems particularly suitable for densely populated urban areas. Distribution on screened quad cable and coaxial cable is mainly considered. Typical transmitting and receiving systems are described.

621.397.24.001.4 **1809**

The Specification and Testing of Television Wire-Broadcasting Systems.—Hollinghurst & Tod. (See 1755.)

621.397.26 : 621.396.65 **1810**

Directional Radio Links on Ultra-short Waves for Television Programme Exchange with Berlin.—W. Scholz. (*Fernmeldetech. Z.*, Dec. 1952, Vol. 5, No. 12, pp. 539–544.) An account of the planning and design of the Berlin-Höhbeck-Hamburg relay system, including test measurements of field strengths in the 174–216-Mc/s band. For the Berlin-Höhbeck 136-km link, aerial systems comprise 30 units mounted on a 150-m mast, each unit consisting of 8 horizontal dipoles. The 10-kW Berlin and 1-kW Höhbeck transmitters use negative a.m. The attenuation of the aerial feeders, which are 360-m coaxial lines of brass with polystyrene-disk insulation, is 1·4 db at 175 Mc/s and 1·8 db at 240 Mc/s.

621.397.26 : 621.396.72 **1811**

Television 'Booster' Stations.—P. J. Harvey. (*Wireless World*, April 1953, Vol. 59, No. 4, pp. 148–152.) Discussion, based on experimental investigations, of the operation of satellite stations in fringe areas, receiving the normal broadcast transmission and reradiating locally on suitable frequencies. Circuit arrangements for 10–250-W vision and 5–50-W sound transmitters are outlined. In the case of retransmission at u.h.f. the expense of converters in domestic receivers would be largely offset by the saving on preamplifiers and special aerials. Initial and maintenance costs of a typical unattended station are estimated and administrative aspects of the system are considered.

621.397.335 : 535.88 **1812**

A Precision Synchronizing System for Large-Screen Television Equipment.—A. W. Keen. (*Proc. Instn elect. Engrs*, Part IIIA, 1952, Vol. 99, No. 20, pp. 696–707. Discussion, pp. 757–760.) A description is given of a complete system developed for experimental cinema equipment. It provides, under local control, line- and frame-trigger and black-level-clamp switching-pulse waveforms, all of which are derived from a type of automatic phase follower and are therefore substantially unaffected by interference or all but the most severe degradation of the signal. Automatic interlacing is introduced, with fine manual control of interframe spacing to eliminate the mean interlace error. Schematic diagrams are given of all the main sections of the equipment, which was designed for a 625/50 European standard, with negative modulation. Since the method adopted is practically equivalent to regeneration of the synchronizing signal, it would appear to be adaptable to other problems, particularly in network repeater stations.

621.397.5 : 535.62/.65 **1813**

A Survey of the Methods and Colorimetric Principles of Colour Television.—J. E. Benson. (*J. Brit. Instn Radio Engrs*, Jan. 1953, Vol. 13, No. 1, pp. 9–49. Reprinted from *Proc. Instn Radio Engrs, Aust.*, July & Aug. 1951, Vol. 12, Nos. 7 & 8, pp. 201–205 & 237–258.) The historical development of colour television is reviewed and the basic requirements of a satisfactory system are discussed. A detailed account is given of the principles of colour specification, measurement and calculation, using the concept of colour space as a means of explaining the algebraic development of the trichromatic theory of colour representation, and leading to an explanation of the origin and principal characteristics of the chromaticity diagram of the International Commission on Illumination. An outline is given of the principles of three-colour reproduction in television. About 140 references.

621.397.5 : 535.623/.624 **1814**

Some Fundamental Aspects of Colour Television.—W. N. Sproson, M. Gilbert & W. West. (*Proc. Instn elect. Engrs*, Part IIIA, 1952, Vol. 99, No. 20, pp. 842–853. Discussion, pp. 860–866.) A review of the physical and physiological factors relevant to the analysis and synthesis of television pictures in colour.

621.397.5 : 535.623/.624 **1815**

Compatible Systems of Colour Television.—G. Valensi. (*Ann. Télécommun.*, Nov. & Dec. 1952, Vol. 7, Nos. 11 & 12, pp. 439–458 & 482–496.) The concept of compatibility, as applied to a system of television in colour, is defined, and the specification of colours by defining the trichromatic coefficients, the dominant wavelength, and colorimetric purity factor, is described. Some physiological characteristics concerned in colour vision are discussed briefly. Two only of the numerous methods of realizing television in colour are described: (*a*) a system using the electrical double-refraction properties of a Kerr cell in the receiving system; (*b*) an entirely electronic system. Discussion indicates a possible reduction of the total bandwidth required for the transmission of colour-television signals. Two of the compatible systems developed in the U.S.A. are also described: the R.C.A. system with dot interlacing of the image, and the N.T.S.C. system using line interlacing.

621.397.5 : 535.623 **1816**

Colorimetric Analysis of Gamma-Corrected Shunted-Monochrome Simultaneous Colour Television Systems.—D. C. Livingston. (*Sylvania Technologist*, Jan. 1953, Vol. 6, No. 1, pp. 13–17.) A formula is developed for use in analysing the colorimetric properties of colour-television systems; in a subsequent paper the method is to be applied to an analysis of the colour system used by the U.S. N.T.S.C. for field testing.

621.397.5 : 535.623(083.71) **1817**

Standards on Television: Definitions of Color Terms, Part 1, 1953.—(*Proc. Inst. Radio Engrs*, March 1953, Vol. 41, No. 3, pp. 344–347.) Standard 53 IRE 22.S1.

621.397.5 : 535.65 **1818**

Colour Television: Some Subjective and Objective Aspects of Colour Rendering.—G. T. Winch. (*Proc. Instn elect. Engrs*, Part IIIA, 1952, Vol. 99, No. 20, pp. 854–860. Discussion, pp. 860–866.)

621.397.5 : 621.317.35 **1819**

A Television-Waveform Monitor.—Attew. (See 1736.)

621.397.5 : 621.396 **1820**

Television as a Communication Problem.—L. C. Jesty. (*Proc. Instn elect. Engrs*, Part IIIA, 1952, Vol. 99, No. 20, pp. 761–770. Discussion, pp. 860–866.) A survey of the art of television communication and discussion of possible lines of development. 26 references.

621.397.6 **1821**

Signal Corps Mobile Television System.—J. S. Auld. (*J. Soc. Mot. Pict. Telev. Engrs*, Dec. 1952, Vol. 59, No. 6, pp. 462–469.) Brief description of a complete microwave system operating over distances up to 20 miles. All equipment for transmission, reception, recording and power supply is housed in five vehicles.

621.397.6 : 535.623 **1822**

Generation of N.T.S.C. Color Signals.—J. F. Fisher. (*Proc. Inst. Radio Engrs*, March 1953, Vol. 41, No. 3, pp. 338–343.) Equipment is described for generating a composite compatible colour signal in accordance with the N.T.S.C. specifications discussed previously [1750 of 1952 (Hirsch et al.)]. A flying-spot system is used which yields voltages proportional to the tristimulus values of the C.I.E. system; these are corrected to suit the red, green and blue channels by a process of electronic addition and subtraction termed 'matrixing'. An indication is given of modifications made to the signal specification in the light of the results of the field tests carried out during 1951 and 1952.

621.397.6 : 621.317.74 **1823**

Test Equipment for Television Transmission Circuits.—S. H. Padel, A. R. A. Rendall & S. N. Watson. (*Proc. Instn elect. Engrs*, Part IIIA, 1952, Vol. 99, No. 20, pp. 821–833. Discussion, pp. 860–866.) The standard of performance of a complete television transmission system is considered, and definite limits are suggested for distortion of amplitude and phase, with respect to frequency, noise and linearity. Descriptions are given of instruments designed to measure amplitude and phase distortions; these are (*a*) video-frequency oscillator, (*b*) video-frequency amplifier detector, (*c*) test-pulse generator, (*d*) waveform monitor, (*e*) group-delay measurement equipment. Applications of these instruments are discussed.

621.397.61/.62 **1824**

Compressed Television.—Radionyme. (*Télévision*, Nov. 1952, No. 28, pp. 259–260, 274.) The principles of various techniques developed by Toulon for reducing the bandwidth required for a vision channel are outlined. Combinations of vertical and horizontal interlacing can be used in conjunction with storage devices at the receiver. By using picture-difference transmission, the bandwidth required for a radio link can be reduced to 1 Mc/s for a system normally occupying 14 Mc/s. Magnetic-drum and phosphorescent-film devices for storing the picture signals are described.

621.397.61/.62 **1825**

Saving Television Bandwidth.—(*Wireless World*, April 1953, Vol. 59, No. 4, pp. 158–162.) Methods proposed for reducing the bandwidth required for the vision channel are reviewed. These include prediction techniques for reducing redundancy in successive images or elements, variable-velocity scanning, decreasing the picture frequency and other methods of exploiting limitations in visual acuity.

621.397.61 **1826**

A Gamma-Control Circuit using Crystal Diodes.—L. Lax & D. Weighton. (*Proc. Instn elect. Engrs*, Part IIIA, 1952, Vol. 99, No. 20, pp. 804–810. Discussion, pp. 860–866.) A basic bridge-type circuit is described which meets the requirements of gamma-control circuits more nearly than those using thermionic valves as nonlinear elements. A practical circuit with a high degree of d.c. stabilization is also described.

621.397.61.029.62 : 061.3 **1827**

The Stockholm Plan.—(*Télévision*, Dec. 1952, No. 29, pp. 296–300.) Details of French television transmissions based on the frequency allocations of the 1952 Stockholm Conference.

621.397.611.2 **1828**

The Mechanism of Signal Generation in Storage-Type Television Camera Tubes.—R. Theile. (*J. Telev. Soc.*, Oct./Dec. 1952, Vol. 6, No. 12, pp. 457–477.) The principles of operation of the iconoscope, image iconoscope, orthicon and image orthicon are described in detail. Charge-storage action and target-stabilization processes are discussed with reference to storage efficiency and transfer characteristic. Two methods of deriving the signal current are described and the signal/noise ratios in the two cases are compared; at low light intensities a better ratio is maintained with preamplification by secondary-emission multiplication. Operational features of the four tubes and improvements in design are discussed, particularly the photoelectron-stabilization and pulsed-biasing techniques applied to the image iconoscope, and performance characteristics directly dependent on the mechanism of signal generation in the four types of tube are compared.

621.397.611.2 **1829**

Generation of Signals in Television Camera Tubes: Part 1 — Fundamentals of the Transformation of the Optical Image into a Charge Image used for Signal Generation.—R. Theile. (*Arch. elekt. Übertragung*, Jan. 1953, Vol. 7, No. 1, pp. 15–27.) See 1828 above.

621.397.611.2 : 778.53 **1830**

The Electronic Camera in Film-Making.—N. Collins & T. C. Macnamara. (*Proc. Instn elect. Engrs*, Part IIIA, 1952, Vol. 99, No. 20, pp. 673–679. Discussion, pp. 757–760. *J. Soc. Mot. Pict. Telev. Engrs*, Dec. 1952, Vol. 59, No. 6, pp. 445–457. Discussion, pp. 458–461.) Limitations of the cinematograph camera are reviewed and the advantages of the use of several electronic cameras, with the attendant possibilities of image distribution for control purposes, are discussed. See also *Wireless World*, April 1953, Vol. 59, No. 4, pp. 153–156.

621.397.62 **1831**

Television Sound Reception. The Critical Capacitance-Coupling System.—S. L. Fife. (*Electronic Engng*, March 1953, Vol. 25, No. 301, pp. 114–117.) In receivers with common r.f. and (sometimes) i.f. stages for the vision and sound signals, it has been found advantageous to use critical coupling, by means of capacitors of the order of 1 pF, in the succeeding sound-channel i.f. amplifier. Suitable i.f. circuits are shown. The use of an inter-carrier f.m. sound circuit in television receivers for the American and European line standards is also considered briefly.

621.397.62 : 621.396.665 **1832**

Television-Receiver A. G. C. Systems.—E. S. White. (*Electronics*, March 1953, Vol. 26, No. 3, pp. 146–149.) Discussion of the operation of seven a.g.c. circuits commonly used in television receivers, and of design factors important for obtaining an efficient low-noise circuit.

621.397.62 : 621.396.665 **1833**

Vision A.G.C.—(*Wireless World*, April 1953, Vol. 59, No. 4, pp. 173–174.) Description of the 'automatic picture control' system which operates on the black-level signal following each line-synchronizing pulse.

621.397.621.2 **1834**

Some Factors in the Design of Deflecting Coils for Cathode-Ray Beam Systems.—E. W. Bull. (*Proc. Instn elect. Engrs*, Part IIIA, 1952, Vol. 99, No. 20, pp. 771–776. Discussion, pp. 860–866.) Conditions for the optimum field distribution are determined for a deflecting-coil system and practical arrangements are briefly described.

621.397.621.2 : 535.88 **1835**

Some Aspects of a Cathode-Ray-Tube Projector for Large-Screen Television in Cinemas.—E. D. McConnell. (*Proc. Instn elect. Engrs*, Part IIIA, 1952, Vol. 99, No. 20, pp. 708–720. Discussion, pp. 757–760.) An account of the development of a large-screen projector system using a Schmidt optical system with a 9-in. c.r. tube which has a back-aluminized screen. Details of the design of the tube and optical system are discussed.

621.397.621.2 : 621.396.615.17 **1836**

Timebase Circuits.—Bähring. (See 1631.)

621.397.645 : 621.397.822 **1837**

Fluctuation Noise in Television-Camera Head Amplifiers.—I. J. P. James. (*Proc. Instn elect. Engrs*, Part IIIA, 1952, Vol. 99, No. 20, pp. 796–803. Discussion, pp. 860–866.) An improvement in signal/noise ratio for the higher video frequencies is obtained with a modified input circuit for an amplifier used with a pickup tube of the C.P.S. emitron type. A method of equalizing the frequency/amplitude response of the modified circuit is described. An improvement of picture quality is obtained, particularly when aperture correction is applied.

621.397.645.018.424 **1838**

Some Factors in the Design of Wide-Band Amplifiers for Television.—W. S. Percival. (*Proc. Instn elect. Engrs*, Part IIIA, 1952, Vol. 99, No. 20, pp. 834–841. Discussion, pp. 860–866.)

621.397.645.029.62 **1839**

Fundamental Problems of H.F. and I.F. Amplifiers for TV Reception: Part 1 — Gain and Bandwidth.—Uitjens. (See 1639.)

621.397.645.029.62 : 621.397.822 **1840**

Fundamental Problems of H.F. and I.F. Amplifiers for TV Reception: Part 2 — Noise.—Uitjens. (See 1640.)

621.397.8 : 535.735 **1841**

An Investigation of the Phenomenon of Flicker and a Possible Explanation of an Observed Resonance Effect.—H. de Lange Dzn. (*Tijdschr. ned. Radiogenoot.*, Jan. 1953, Vol. 18, No. 1, pp. 1–31. Discussion, pp. 31–32.) The dependence on frequency and mean brightness of the ability of the eye to perceive rapid variations of brightness can be explained by assuming that the physiological system transmitting the visual impressions from the retina to the brain includes a low-pass filter. The problem can then be examined in the frequency domain, using Fourier analysis. An important parameter is the ripple ratio r, defined as the ratio of the amplitude of the first Fourier component to the mean brightness. At high brightness levels a resonance effect occurs at about 9 c/s; under these conditions the value of r at which flicker disappears is $< 1{\cdot}35\%$. The results are consistent with the presence in the physiological system of a feedback mechanism having a time delay of about 1/80 sec.

621.397.8 : 621.317.79 **1842**

A Variable-Definition Camera Channel for the Appraisal of Television Standards.—Lord & Wood. (See 1753.)

621.397.828 **1843**

Reduction of Pulse Interference in Television Receiving Systems.—A. W. Keen. (*J. Brit. Instn Radio Engrs*, Jan. 1953, Vol. 13, No. 1, pp. 51–55.) A method of interference reduction, known as 'black spotting', which has been used in receivers designed for reception of British positive-modulation transmissions (405/50 standard) is discussed. Existing circuit techniques are reviewed and a more detailed description is given of examples chosen to illustrate the particular suitability of the cathode-coupled-pair type of amplifier as a basis for improved circuit arrangements.

621.397.9 **1844**

Industrial and Professional Applications of Television Technique.—R. C. G. Williams. (*Proc. Instn elect. Engrs*, Part IIIA, 1952, Vol. 99, No. 20, pp. 651–664. Discussion, pp. 757–760.) A survey of British developments.

621.397.9 : 617(07) **1845**

An Application of Television to the Demonstration of Operative Surgery.—G. C. Newton & H. J. B. Atkins. (*Proc. Instn elect. Engrs*, Part IIIA, 1952, Vol. 99, No. 20, pp. 721–728. Discussion, pp. 757–760.) Description of the equipment used at Guy's Hospital, London, and discussion of the value of television as a teaching aid and of the possibilities for its future development, based on extensive experience with the equipment. See also 3276 of 1949.

TRANSMISSION

621.396.619.23 **1846**

Serrasoid Modulation.—E. Paulsen. (*Frequenz*, Jan. 1953, Vol. 7, No. 1, pp. 14–18.) An explanation is given of the operation of the serrasoid modulator [342 of 1949 (Day)], with particular reference to the capacitor charge and discharge phenomena in the modulation section, which have a great influence on the attainable frequency swing and the incidental a.m. See also 3267 (Gundlach) and 3589 (Bünemann & Pethke) of 1952.

621.396.3 : 621.396.61 **1847**

Rounding of Signals by a Filter in a Radiotelegraphy Transmitter.—A. Tchernicheff. (*Ann. Télécommun.*, Dec. 1952, Vol. 7, No. 12, pp. 513–516.) Results are given of measurements carried out on a transmitter in which a low-pass filter is used to reduce the transmission bandwidth. The frequency spectra of rectangular signals and of signals filtered by means of filters with cut-off frequencies of 75 c/s and 100 c/s are compared. Measurements of the telegraphy distortion were made by a stroboscopic method using a c.r.o.

VALVES AND THERMIONICS

537.533 : 621.396.822 **1848**

Noise Analysis of a Single-Velocity Electron Gun of Finite Cross Section in an Infinite Magnetic Field.—H. E. Rowe. (*Trans. Inst. Radio Engrs*, Jan. 1953, No. PGED-2, pp. 36–46.) The various sinusoidal modes of propagation of a finite accelerated electron beam, converging or parallel, are investigated and boundary conditions for matching at the cathode and at the anode are considered. An idealized model is analysed in which an infinite magnetic field parallel to the axis of the beam prevents all transverse motion of the electrons both in the drift space and in the electron gun. Only a summary of the analysis is presented. The results are applied to numerical calculations for seven different guns, for which noise-standing-wave curves are given for different anode voltages and, in the case of converging beams, for different angles of convergence. These curves show that for beams of the same total current and radius in the drift space, a converging-beam gun has a lower s.w.r. than a parallel-beam gun. Reduction of the beam radius, keeping beam voltage and current constant, reduces the level of the maxima while increasing the s.w.r., thus further reducing the minima of the noise standing wave.

621.314.632 **1849**

Germanium Diodes Sealed in Glass.—J. W. Dawson. (*Sylvania Technologist*, Jan. 1953, Vol. 6, No. 1, pp. 1–4.) To prevent the entry of moisture into the diodes they may be sealed either in glass or in ceramic cartridges filled with wax. Humidity-exposure tests indicate that the glass-cartridge assembly is the more effective in maintaining the diode characteristics stable.

621.314.7 + 621.314.632 **1850**

Germanium in Telecommunications Technique.—M. C. Weill. (*Onde élect.*, Jan. 1953, Vol. 33, No. 310, pp. 15–26.) A review of the many uses of Ge rectifiers and transistors.

621.314.7.001.8 **1851**

Properties and Utilization of Transistors.—J. M. Moulon. (*Onde élect.*, Jan. 1953, Vol. 33, No. 310, pp. 27–35.) A review of the characteristics of point-contact and *p-n* junction transistors and of their applications in amplifiers, negative-impedance repeaters, gating circuits, etc.

621.383 + 621.314.63 **1852**

Height of the Potential Barrier in Barrier-Layer Cells.—M. S. Ridout. (*Nature, Lond.*, 31st Jan. 1953, Vol. 171, No. 4344, p. 219.) If the barrier height is assumed to have a normal distribution about a mean value and to have a standard deviation, the equation relating zero-voltage resistance to temperature can be recast in a form which in general fits well the experimental results of Billig & Ridout (3153 of 1951) for Se, Ge and Si rectifiers and for Se photocells.

621.383.27 **1853**

Development and Use of Electron-Multiplier Photocells.—A. Lallemand. (*Onde élect.*, Dec. 1952, Vol. 32, No. 309, pp. 492–495.) The general principles of construction of electron-multiplier photocells are discussed, with illustrations of the actual construction of cells with 7 and 19 stages respectively.

621.385 : 621.3.018.78 **1854**

Parasitic Effects and Distortion due to Curvature of Valve Characteristics.—B. G. Dammers, J. Haantjes, J. Otte & H. van Suchtelen. (*Philips tech. Commun., Aust.*, 1952, No. 7, pp. 3–26.) Reprint from 'Applications of the Electronic Valve in Radio Receivers and Amplifiers' (991 of 1950).

621.385.029.63/.64 **1855**

Phase Distortion in Amplification by means of Travelling-Wave Valves.—L. Brück. (*Arch. elekt. Übertragung*, Jan. 1953, Vol. 7, No. 1, pp. 28–36.) Reflections and dispersion introduced by the delay line give rise to a variation of the group transit time over the pass band which is of particular importance in wide-band amplifiers for multichannel radio links. An approximate determination is made of the distortion factor for f.m. signals. For a valve with a helix, distortion due to dispersion is negligible compared with that due to mismatching. Methods of reducing the latter are discussed.

621.385.029.64 **1856**

Space-Charge-Wave Amplifier Tubes, Basic Principles of Operation.—R. G. E. Hutter. (*Sylvania Technologist*, Oct. 1952, Vol. 5, No. 4, pp. 94–99, & Jan. 1953, Vol. 6, No. 1, pp. 6–12.) The performance of space-charge-wave valves (i.e., travelling-wave valves, electron-wave valves, single-stream electron-wave valves, etc.) is analysed by considering them as composed of basic regions, viz., gaps between apertured electrodes, drift regions and slow-waveguide regions. A general electronic theory applicable to all the regions is developed. The circuit problem is discussed, first in general and then for each region. The combined electronic and circuit relations are derived.

621.385.032.216 **1857**

Thermionic Emission from Oxide-Coated Tungsten Filaments.—C. P. Hadley. (*J. appl. Phys.*, Jan. 1953, Vol. 24, No. 1, pp. 49–52.) An experimental investigation including (*a*) observations with a phosphor-coated tube of emission distribution, (*b*) retarding-field measurements, (*c*) accelerating-field measurements, and (*d*) X-ray diffraction study of the coating.

621.385.032.216 **1858**

The Effect of Impurity Migrations on Thermionic Emission from Oxide Cathodes.—I. E. Levy. (*Proc. Inst. Radio Engrs*, March 1953, Vol. 41, No. 3, pp. 365–368.) Emission measurements were made using diodes (*a*) similar to the A.S.T.M. standard, and (*b*) with specially purified materials. Results indicate the dependence of the work function on impurities migrating from parts of the valve other than the cathode.

621.385.032.216 : 537.311.33 **1859**

Semiconductors and Oxide Cathodes.—R. Bourion. (*Onde élect.*, Jan. 1953, Vol. 33, No. 310, pp. 36–39.) Electron emission from semiconductors in vacuo is discussed in relation to Fermi energy levels. Reasons are given for considering that the active material in an oxide cathode is an excess semiconductor. Experimental results support this view, but the complex effects observed in oxide cathodes cannot be adequately explained by this simple theory.

621.385.032.8 **1860**

Vacuum-Tube Sockets for Industrial Equipment.—W. Clarkin & L. F. Biosca. (*Elect. Mfg*, July 1951, Vol. 48, No. 1, pp. 72–77 . . 218.) A review of available types of wafer and moulded sockets suitable for receivers, and also for heavy-duty industrial service and for applications requiring low leakage, low loss, and reliable operation under vibration and humid conditions.

621.385.2 **1861**

Schottky's Equation for Thermal Electron Emission.—H. Bonifas. (*Rev. gén. Élect.*, Nov. 1952, Vol. 61, No. 11, pp. 539–540.) An error in the development of this equation is pointed out, and a corrected formula is given for the anode current of a diode operating at constant temperature.

621.385.2 **1862**

Extension of the Planar Diode Transit-Time Solution.—W. E. Benham. (*Proc. Inst. Radio Engrs*, March 1953, Vol. 41, No. 3, pp. 395–396.) Comment on 497 of 1950 (Begovich).

621.385.2 : 537.533 **1863**

Electron-Gas Equations for Electron Flow in Diodes.—H. Poritsky. (*Trans. Inst. Radio Engrs*, Jan. 1953, No. PGED-2, pp. 60–84.) Analysis is presented which shows that Langmuir's results relative to the electron current in a diode can be recast in a form involving the concepts of electron gas. Examination of the discrepancies between Langmuir's results and those of Hahn (566 of 1949) indicates that they are due to somewhat arbitrary assumptions made by Hahn, and that the discrepancies disappear when these assumptions are properly modified. Langmuir's planar-diode equations, when put into a form involving gas motion, thus yield results identical with those obtained from analysis of the electron motion.

621.385.2 : 621.396.822 **1864**

Noise in Thermionic Valves.—J. H. Fremlin. (*Proc. Instn elect. Engrs*, Part III, March 1953, Vol. 100, No. 64, pp. 91–92.) The relation between Nyquist's formula for the mean square noise voltage generated by a resistive circuit and the formula for the fluctuation current in a temperature-limited diode is discussed. In the practical space-charge-limited case the thermodynamical approach is of little assistance; an explanation of this is given.

621.385.3 **1865**

Inherent Feedback in Triodes.—H. Stockman. (*Wireless Engr*, April 1953, Vol. 30, No. 4, pp. 94–96.) The triode is considered as an infinite-impedance pentode, with the effect of the anode on the field at the cathode represented by negative feedback. By means of this transformation practical formulae are obtained for the triode circuit from conventional feedback theory.

621.385.3/.5].012 **1866**

The Effect of Filament Voltage upon Vacuum Tube Characteristics.—A. J. Winter. (*Trans. Inst. Radio Engrs*, Jan. 1953, No. PGED-2, pp. 47–59.) Operation of valves with reduced filament voltages is discussed. By using the concept of an image cathode, located behind the real cathode at the position from which the electrons appear to be emitted, the principles of electrostatics can be applied to determine the effect of lowered cathode temperature on the amplification factor. Curves for a Type-6C4 triode and a Type-6AG5 pentode illustrate the fact that a higher gain can be obtained with a lower filament voltage. A 50% increase of gain has been obtained by reducing the filament voltage from 6·3 V to 3·0 V.

621.385.5.032.212 **1867**

A Ten-Stage Cold-Cathode Stepping Tube.—D. S. Peck. (*Elect. Engng, N.Y.*, Dec. 1952, Vol. 71, No. 12, pp. 1136–1139.) Description of the construction and operating characteristics of the Western Electric Type-6167 decade counter tube, which is capable of operating at pulse rates up to about 2 000/sec.

621.385.832 **1868**

Single-Gun Storage Tube Writes, Reads and Erases.—R. C. Hergenrother & A. S. Luftman. (*Electronics*, March 1953, Vol. 26, No. 3, pp. 126–130.) An article based on a 1952 National Electronics Conference paper (*Proc. nat. Electronics Conf., Chicago*, 1952, Vol. 8, pp. 543–552). Improvements effected in an earlier model [2942 of 1950 (Hergenrother & Gardner)] are described which permit storage of charges for periods as long as a week with little loss, and up to 27 000 playbacks with little effect on the storage pattern. Typical applications are noted.

621.385.832 : 621.397.62 : 535.37 **1869**

Phosphors for Television Cathode-Ray Tubes.—H. G. Jenkins, A. H. McKeag & E. E. Welch. (*Proc. Instn elect. Engrs*, Part IIIA, 1952, Vol. 99, No. 19, pp. 542–550. Discussion, pp. 571–576.) The development of television c.r.-tube screens and phosphors is reviewed, the theory of luminescence outlined, the preparation of the principal television-tube phosphors described and their characteristics noted. 44 references.

621.385.832.002.2 : 621.397.62 **1870**

Some Aspects of Modern Cathode-Ray-Tube Manufacture.—L. J. Bayford. (*Proc. Instn elect. Engrs*, Part IIIA, 1952, Vol. 99, No. 19, pp. 514–523. Discussion, pp. 571–576.) An account of the more important processes in the manufacture of aluminized television c.r. tubes, including descriptions of (*a*) a semi-automatic method of screen deposition, with film forming before screen drying, (*b*) the pipette specially designed to facilitate the continuous production of nitrocellulose films on fluorescent screens prior to aluminizing.

621.387 : 621.316.722.1 **1871**

Voltage-Regulator Tubes.—W. Kiryluk. (*Electronic Engng*, Feb. 1953, Vol. 25, No. 300, p. 83.) A method is outlined for determining the dynamic characteristics, from which the best operating point for each tube can be rapidly found.

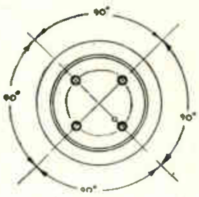
621.396.615.14 : 621.385.029.6 **1872**

Traveling-Wave-Tube Oscillators.—H. R. Johnson & J. R. Whinnery. (*Trans. Inst. Radio Engrs*, Jan. 1953, No. PGED-2, pp. 11–35.) Analysis is presented for a type of oscillator comprising a travelling-wave valve with feedback via an external path including a cavity resonator of the transmission-line type. The analysis indicates that a tuning range of 4–8% should be obtainable by variation of the helix voltage. A range of 4·5% has been obtained experimentally, the centre frequency being 2·72 kMc/s and power output 0·3 W. Theory shows that the valve used should be short, with a high gain per unit length.

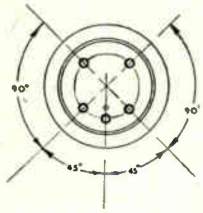
MISCELLANEOUS

061.3 : 621.39 **1873**

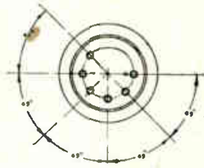
1953 I.R.E. National Convention, 23rd–26th March.—(*Proc. Inst. Radio Engrs*, March 1953, Vol. 41, No. 3, pp. 401–427.) Summaries are given of the papers presented.



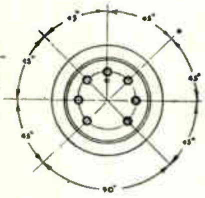
TYPE 3244
Rating 1KV. 5A.



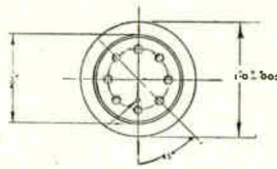
TYPE 3245
Rating 1KV. 5A.



TYPE 3246
Rating 1KV. 5A.

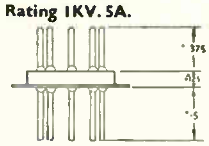


TYPE 3247
Rating 1KV. 5A.



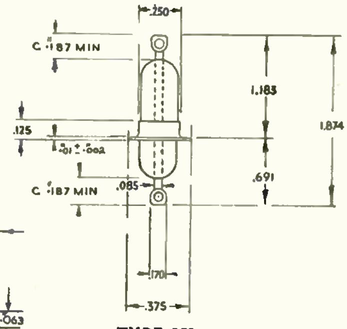
TYPE 3248
Rating 1KV. 5A.

EDISWAN *metal glass seals*



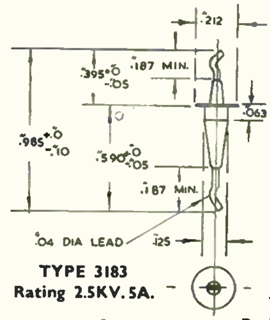
Side view 3244-8
NOTE. All pins are ".04/25 Top Ends Rounded Spaced on ".5 PDC.

These outline diagrams illustrate a range of Ediswan metal-glass seals delivery of which can be effected promptly. Many additional types can be supplied by arrangement, and our Engineers will welcome the opportunity of collaborating with you in the design of special seals for your particular requirements.

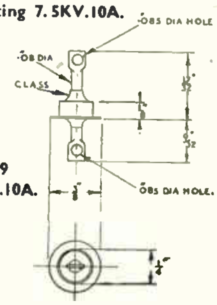


TYPE 252
Rating 7.5KV.10A.

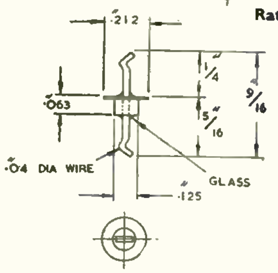
TYPE No.	VOLTAGE RATING (Volts)	CURRENT PER TERMINAL (amps)	REMARKS
252	7,500	10.0	
3279	2,000	10.0	
3167	10,000	10.0	
3183	2,500	5.0	
3244	1,000	5.0	Voltage between pins and flange.
3245	1,000	5.0	
3246	1,000	5.0	
3247	1,000	5.0	
3248	1,000	5.0	
3280	750	5.0	



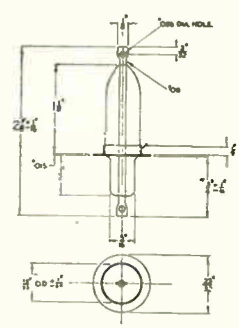
TYPE 3183
Rating 2.5KV. 5A.



TYPE 3279
Rating 2.0KV.10A.



TYPE 3280
Rating 750V. 5A.



TYPE 3167
Rating 10KV. 10A.

EDISWAN

For full information please write for publication R.1602

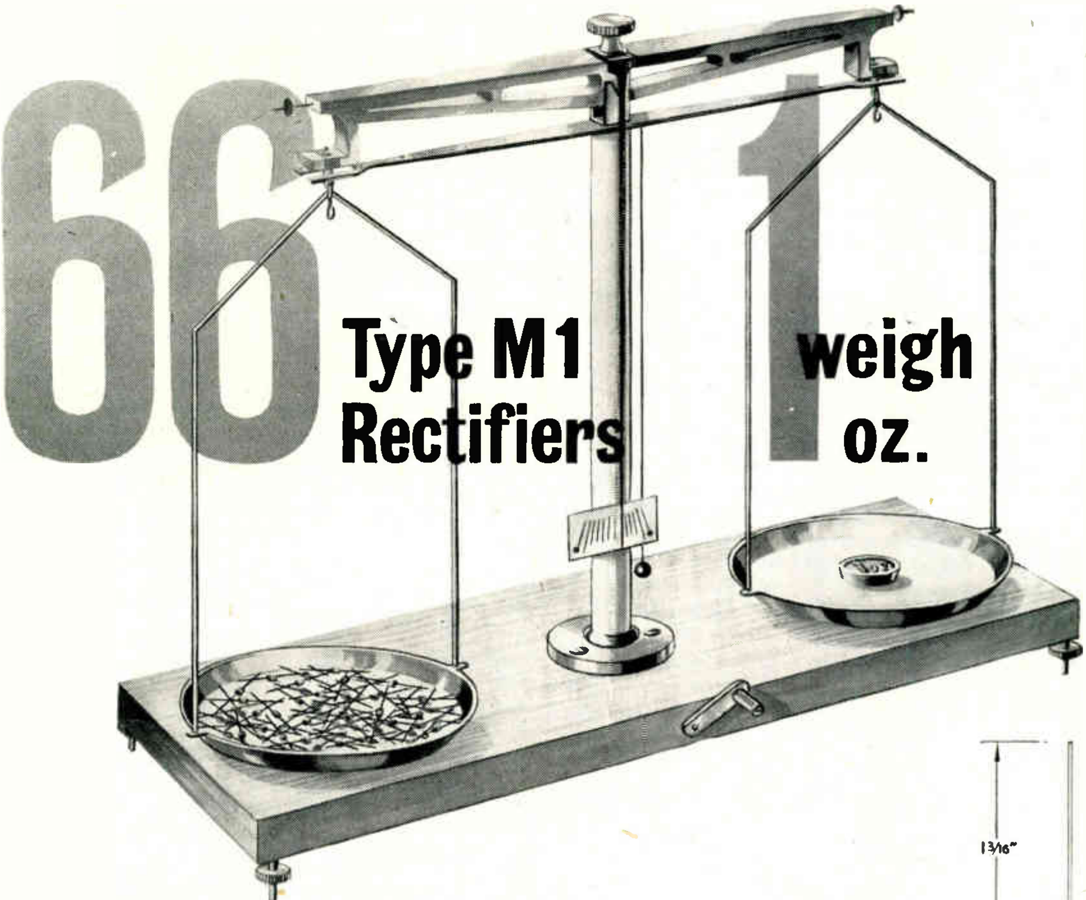
RADIO DIVISION

THE EDISON SWAN ELECTRIC CO. LTD., 155 CHARING CROSS RD., LONDON, W.C.2

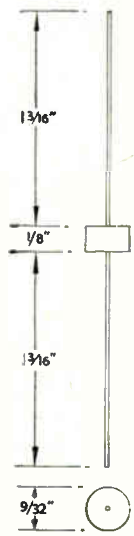
Telephone: Gerrard 8660 *Member of the A.E.I. Group of Companies* *Telegrams: Ediswan, Westcent, London*

MG104

66 Type M1 Rectifiers weigh 1 oz.



Inexpensive, featherweight SenTerCel Rectifiers types M-1 and M-3 can be suspended even in light wiring, and in most cases will replace a thermionic rectifier. They measure 9/32″ diam. x ⅛″ thick . . . and will operate at input levels of 0·5 volt upwards at frequencies up to 5 Mc/s. Less wiring! No valve-holder ! No heat !



TYPE M-1

For A.F. 2nd detectors. AGC rectifiers. Muting circuits. Contrast expansion and compression. Level indicators. Modulation depth indicators. Limiters. Automatic frequency control.

Average characteristics :

Self Capacitance	22 pF
Forward resistance at 5 V D.C.	10 kΩ
Reverse resistance at 5 V D.C.	1000 MΩ
Maximum peak inverse voltage	68 V
Minimum A.C. input	0·5 V

TYPE M-3

Has similar characteristics and application with a maximum frequency of 100 kc/s.

Average characteristics :

Self Capacitance	65pF
Forward resistance at 5 V D.C.	1·2 kΩ
Reverse resistance at 5 V D.C.	45 MΩ
Maximum peak inverse voltage	68 V
Minimum A.C. input	0·5 V

SELENIUM
SenTerCel
RECTIFIERS

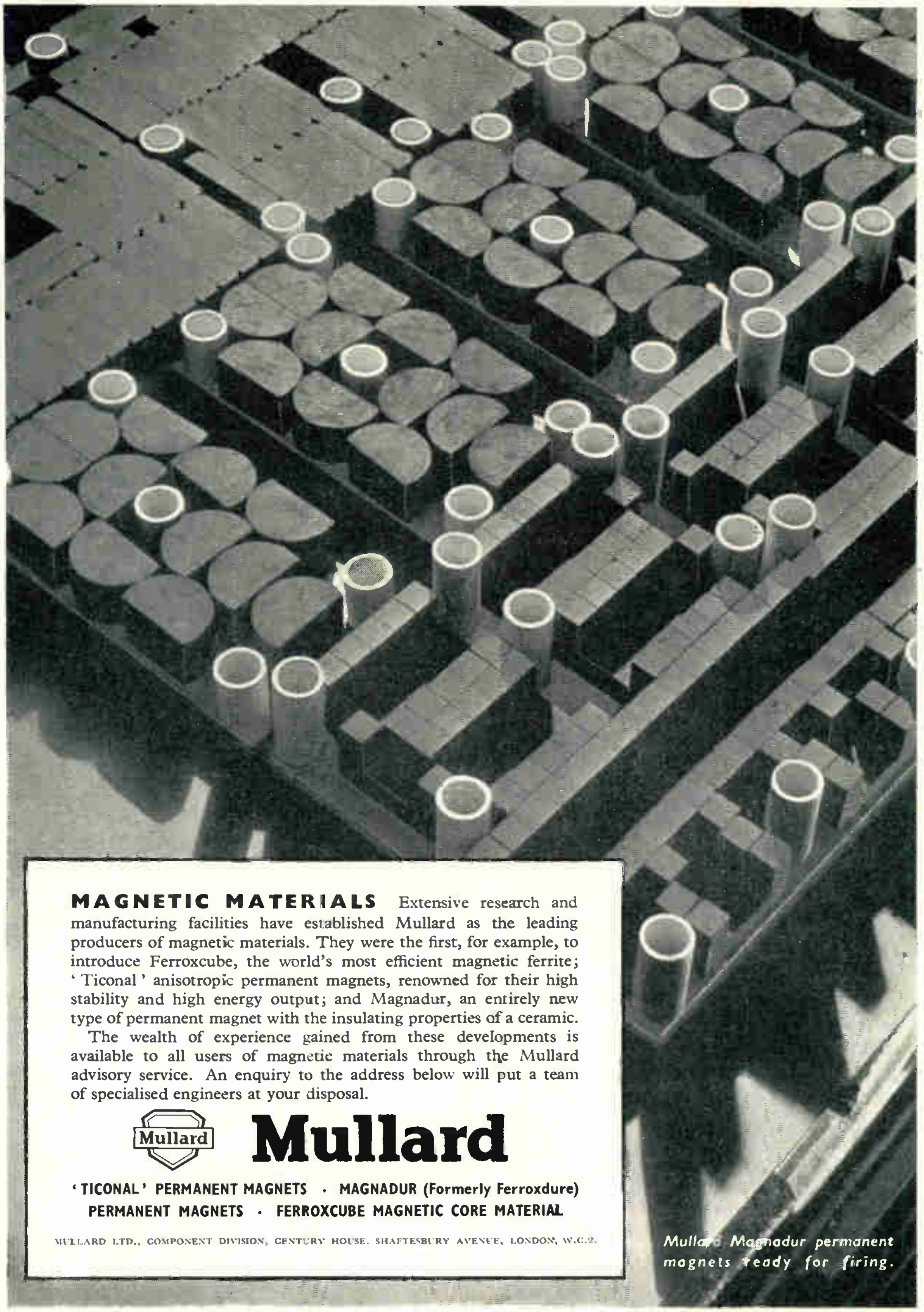
REGISTERED TRADE MARK

Standard Telephones and Cables Limited

Registered Office : Connaught House, Aldwych, London, W.C.2.

RECTIFIER DIVISION : Warwick Road, Boreham Wood, Hertfordshire

Telephone : Elstree 2401 Telegrams : SenTerCel, Borehamwood



MAGNETIC MATERIALS Extensive research and manufacturing facilities have established Mullard as the leading producers of magnetic materials. They were the first, for example, to introduce Ferroxcube, the world's most efficient magnetic ferrite; 'Ticonal' anisotropic permanent magnets, renowned for their high stability and high energy output; and Magnadur, an entirely new type of permanent magnet with the insulating properties of a ceramic.

The wealth of experience gained from these developments is available to all users of magnetic materials through the Mullard advisory service. An enquiry to the address below will put a team of specialised engineers at your disposal.

Mullard

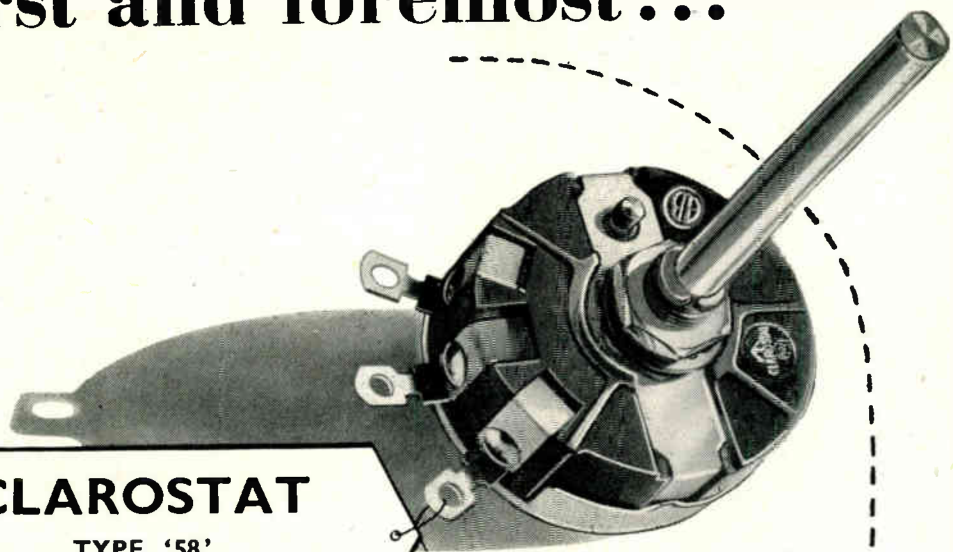
Mullard

'TICONAL' PERMANENT MAGNETS · MAGNADUR (Formerly Ferroxdure) PERMANENT MAGNETS · FERROXCUBE MAGNETIC CORE MATERIAL

MULLARD LTD., COMPONENT DIVISION, CENTURY HOUSE, SHAFTESBURY AVENUE, LONDON, W.C.2.

Mullard Magnadur permanent magnets ready for firing.

First and foremost...



CLAROSTAT

TYPE '58'

WIRE-WOUND POTENTIOMETER

We are proud to introduce the CLAROSTAT Type 58 Wire-Wound Potentiometer, designed by the world's foremost Potentiometer manufacturers and produced for the first time in this country. The Type 58 has many features which will immediately appeal to the discriminating user.

FEATURES

★ Rated at 3 watts and available in all values up to **100,000** ohms linear.

★ Special windings are available to requirements.

★ High Grade Bakelite Casing, of rugged construction. Solder Tags heavily silver-plated and of special design making them completely immovable under all conditions.

★ Metal cover firmly keyed into Bakelite Casing, cannot loosen or turn. Connected to fixing bush providing automatic "grounding" of cover.

★ Obtainable with single or double pole mains switch.

★ Dimensions: Diameter, $1\frac{21}{32}$ in. Depth of case without switch, $\frac{25}{32}$ in.

★ Samples and full specifications available on application.

★

WHOLESALERS

CLAROSTAT TYPE 58 Wire-Wound Potentiometers are supplied individually packed in attractive printed cartons. Delivery is prompt. Write for price list and details of very attractive trade terms.

★

CLAROSTAT

Controls and Resistors

METAL PRODUCTS LIMITED.

16, Berkeley Street, London, W.1.

Phone: GROsvenor 5206/7

Farnborough, Hants., Royal Aircraft Establishment Technical College requires:—

Assistant "B" in Electrical Engineering to teach subject to H.N.C. and possibly final B.Sc.(Eng.) standard. Facilities for research may be available. Quals.: Degree or equivalent in electrical engineering, preferably with industrial experience.

Demonstrator in Electrical Engineering and Physics to assist in organization, preparation and control of work in electrical laboratory (to B.Sc. standard) and in physics (to Inter B.Sc. standard). A little teaching may be involved. Facilities for research or further study may be available. Quals.: Electrical Engineering Degree or equivalent, preferably with industrial experience.

Salaries on Burnham Scales for Technical Colleges with appropriate allowances for training, experience, etc., and subject to usual deductions for superannuation.

Details and forms from Principal, to whom completed forms must be returned within 14 days of this notice.

APPOINTMENTS—The engagement of persons answering these advertisements must be made through the local office of the Ministry of Labour and National Service, etc., if the applicant is a man aged 18-64 or a woman 18-59 inclusive, unless he or she or the employer is excepted from the provisions of The Notification of Vacancies Order, 1952.

CITY OF COVENTRY EDUCATION COMMITTEE

Coventry Technical College

Principal: H. V. Field, B.Sc., Wh.Sch., M.I.E.E.

Electrical Engineering and Physics Department

Required September 1st, 1953, full-time Lecturer in Electronics and Telecommunications. Candidates should be graduates or hold equivalent qualifications, have had appropriate industrial or service experience and preferably some teaching experience.

Salary in accordance with the Burnham Technical Scale (£940 × £25—£1,040).

Application forms and further particulars of the above appointment are obtainable from the Director of Education, Council House, Coventry, to whom completed forms should be returned as soon as possible.

London Electrical Manufacturing Co., Ltd., require Senior Production Engineers with theoretical and/or practical knowledge. Applicants should have had experience in the industry, preferably with components. Preference and considerably higher salaries would be offered to men who have been engaged in the production of ceramic condensers. Please write, giving full particulars, in confidence to: L.E.M., Beavor Lane, W.6, or telephone Riverside 4824, asking for Mr. Lyons.

Draughtsmen, Senior and Junior Required. Design and layout experience of electronic and electro-mechanical equipment and/or mechanical computors essential. Facilities exist for further technical training under the supervision of E.M.I. Institutes, Ltd. Apply by letter stating age, nationality and experience to: E.M.I. Engineering Development, Ltd. (Wells Division), Penleigh Works, Wells, Somerset.

Belling & Lee, Ltd., Cambridge Arterial Road, Enfield, Middlesex, require research assistants in connection with work on electronic components, fuses, interference suppressors and television aerials. Applicants must be graduates of the I.E.E. or possess equivalent qualifications together with similar laboratory experience. Salary will be commensurate with previous experience: five-day week, contributory pension scheme. Applications must be detailed and concise, and will be treated as confidential.

EXPERIENCED RADIO TESTERS AND INSPECTORS

Required for production of communication and radio apparatus. Also instrument makers, wirers and assemblers for factory test apparatus. Apply Personnel Manager, E. K. Cole, Ltd., Ekco Works, Malmesbury, Wilts.

Electronic Engineer required for development work in a laboratory (London S.E. Area) engaged in the application of electronic techniques in new fields. Pension Scheme. Give full details of qualifications, experience, age and commencing salary required.—Box 7282, c/o *Wireless Engineer*.

Required spare-time Television Aerial Development Engineer. Field Testing and Experimental Work. Write full details, age, experience, etc., in confidence.—Box 7290, c/o *Wireless Engineer*.

THE COLLEGE OF AERONAUTICS

Applications are invited for the following posts in the expanding Electrical Section:—

(1) **Lecturer in Electrical Engineering (Electronics)**, who will be required to cover fundamental electronic circuits, including pulse circuits and their application to the measurement of non-electrical quantities. He will also assist in the development of the specialized electronic instruments required by all Departments of the College in their research work. Honours Degree in Physics or Electrical Engineering required with experience in at least one of the following fields:—airborne radio and radar systems; servomechanisms; electronic techniques in measurement and control. Salary within range £600-£1,000 p.a. with superannuation under F.S.S.U. and family allowance. Further particulars available.

(2) **Design Draughtsman**, who will be responsible for the layout and detail design of the great variety of electronic and electrical equipment developed by the Section. Applicants must have had experience with electronic and/or light electro-mechanical equipments. Preference will be given to a person capable of assisting and supervising students in their project-design work. Salary according to qualifications and experience.

Applications, giving full particulars and containing the names and addresses of not more than three referees, to the Recorder, The College of Aeronautics, Cranfield, Bletchley, Bucks.

Television Engineer, conversant with T/V theory, testing methods and test equipment. Able to organize and handle all servicing needs of large rental organization in London. Post offers permanency and prospects, attractive remuneration to man of integrity and drive. Application (fully confidential) to: Telefusion, Ltd., 69 Southwark Park Road, Bermondsey, S.E.16.

Admiralty invites applications for Experimental Officer or Assistant Experimental Officer vacancy at Royal Greenwich Observatory, Herstmonceux, Sussex, for recording of Cosmic Ray intensity and development of photoelectric instruments. Appointment unestablished, but opportunities occur for establishment through open competition. Starting salary depending upon age and experience within ranges E.O. £597-£755 (Men); £502-£623 (Women); A.E.O. £264-£555 (£467 Women). Candidates must be British, at least 17½ years of age and should possess at least H.N.C. or equivalent with some experience of modern electronics development work and circuit design.

Application forms from M.L.N.S., Technical and Scientific Register (K), Almack House, 26 King Street, London, S.W.1, quoting A.101/53/A. Closing date June 20, 1953.

Several experienced graduate electronic and electro-mechanical engineers are required for long term development in centimetric radar equipment, servo mechanisms and instrumentation. Candidates must be British born and should be capable of working as members of a team with little supervision. The posts are permanent and pensionable. Applications, giving full particulars of experience to date will be treated in confidence and should be made in writing, quoting reference SP, to the Personnel Manager, The Plessey Co., Ltd., Vicarage Lane, Ilford.

THE BRITISH IRON & STEEL RESEARCH ASSOCIATION

Young Electronics Engineer required by the Plant Engineering Division of the above Association to assist in instrumentation work involving use of strain gauges and multichannel metring and recording circuits in connection with new developments in steelworks plant.

Applicants should have an Ordinary or Higher National Certificate in electrical engineering. Age range 20-24. Starting salary according to age, qualifications and experience within the range £430-£480 p.a. Location of work London, with visits to steelworks.

Written applications only quoting "Instruments," to Personnel Officer, B.I.S.R.A., 11 Park Lane, London, W.1.

Electronic and Television Engineer required, with sound technical qualifications, suitable design and practical experience, to take charge of small development unit (due for expansion) in London, producing specialized equipment for H.F. wide-band line and/or space transmission systems. The post offers opportunities for initiative and progress with generous remuneration. Replies (fully confidential) Telefusion, Ltd., 69 Southwark Park Road, Bermondsey, S.E.16.

The De Havilland Engine Co., Ltd. Electronic Engineer required, interested in physical measurement rather than construction of apparatus, required for vibration measurement and analysis on gas turbine and piston engines. Please write in confidence stating age and full details of previous experience to the Personnel Officer, The de Havilland Engine Co., Ltd., Stag Lane, Edgware, Middlesex.

The English Electric Co., Ltd., Stafford, requires the following personnel for its new factory in the North Staffs. area:—

1. **Engineer** with experience of magnetic amplifiers; good academic and practical background is essential.

2. **Engineer** with experience on electronic servo control gear and servo analysis.

Housing facilities may be available in 12 months. Salary up to £1,000 per annum.

Apply to Department Central Personnel Services, 336 Strand, W.C.2, quoting ref. 1095.

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