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VOL. 31 No. 7 THREE SHILLINGS AND SIXPENCE

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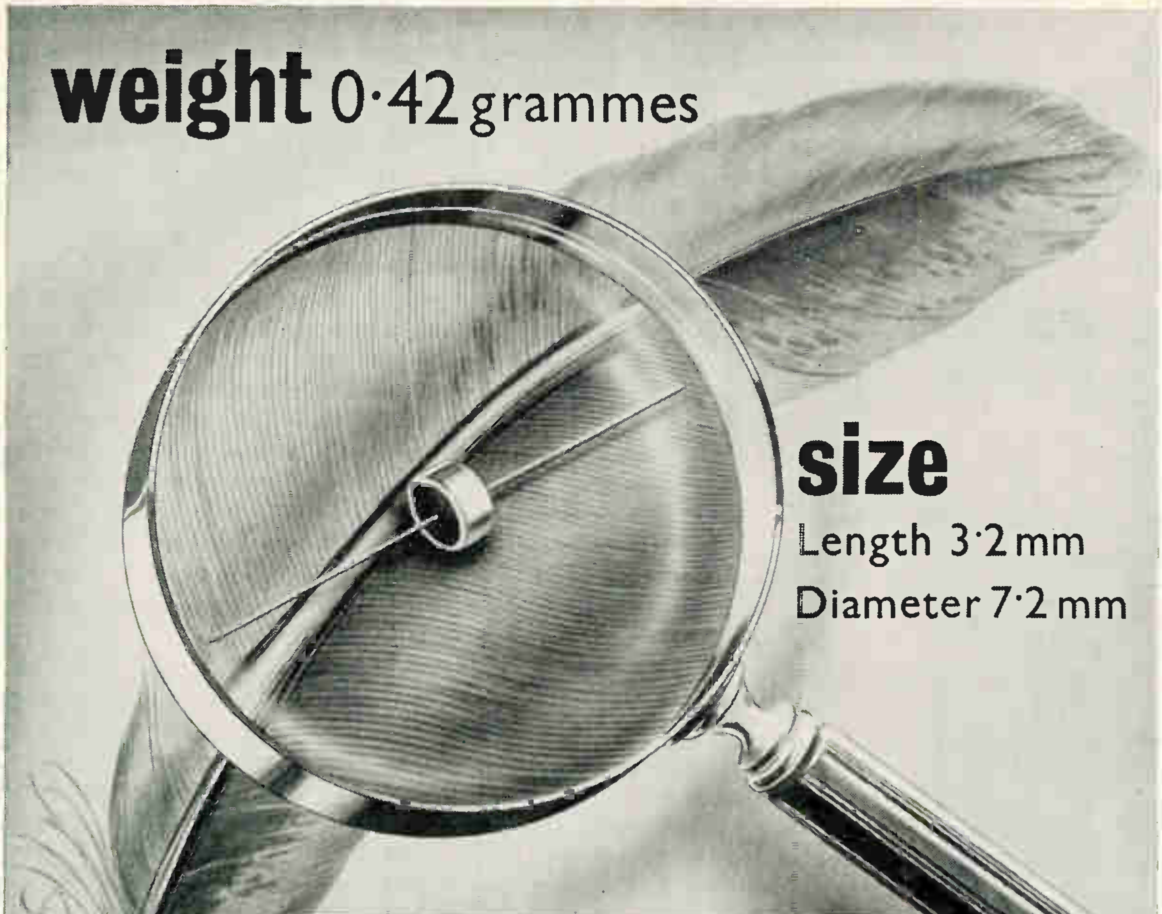

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Maximum Frequency	5 Mc/s.

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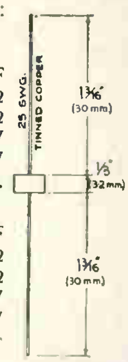
Average Characteristics

Self Capacitance	65 pF
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
Maximum Frequency	100 kc/s



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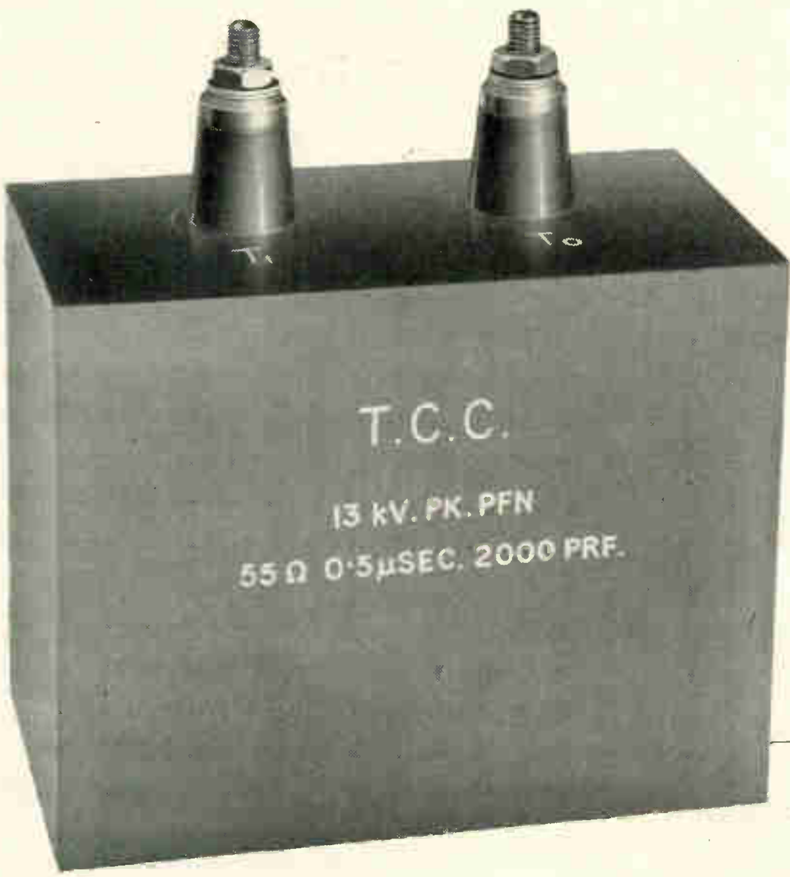

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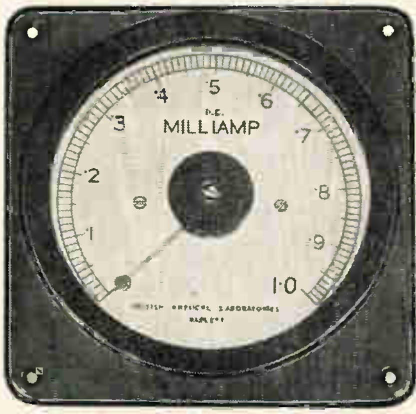

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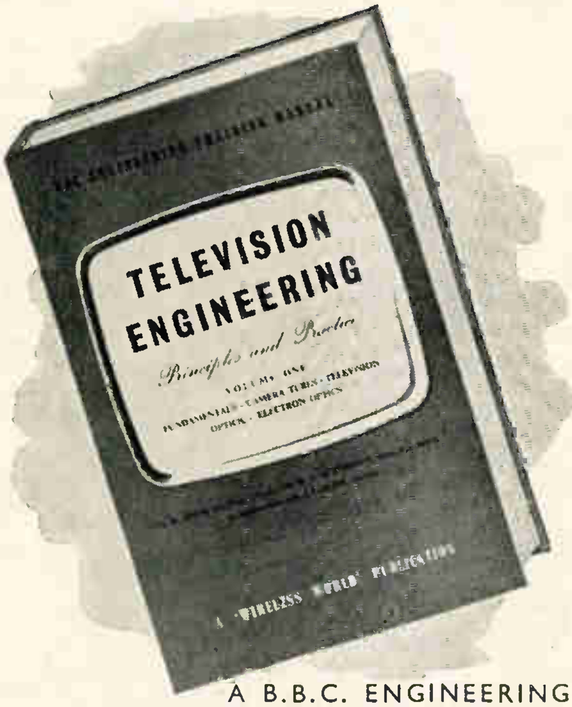

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WIRELESS ENGINEER

The Journal of Radio Research and Progress

ESTABLISHED 1923

Managing Editor: HUGH S. POCOCK, M.I.E.E.
Editor: W. T. COCKING, M.I.E.E. Technical Editor: Professor G. W. O. HOWE, D.Sc., LL.D., M.I.E.E.

Volume 31 · Number 7

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JULY 1954

Published on the fifth of each month

Annual Subscription: Home and overseas £2 7s. 0d; Canada and U.S.A. $7.50
(*Including Annual Index to Abstracts and References*)

Editorial, Advertising and Publishing Offices: Dorset House, Stamford Street, London, S.E.1

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WIRELESS ENGINEER

Vol. 31 JULY 1954 No. 7

Improved Morphology of Electromagnetics

THIS is the title of an article of 83 pages in the *Bulletin of the Technical University of Istanbul* (1952) by Paul de Belatini of the Telecommunication Department. The dictionary tells us that morphology is the science of organic form. There has been a tendency in recent years to merge the two systems, one based on the unit charge and the other on the unit pole, into a common system, but the author suggests eliminating the anomalies of the usual duality of electricity and magnetism by "replacing the duality by triplicity, where the third sector is that of 'dielectricity' in duality to magnetism, with electricity midway between them".

The meaning of this is made clear by the following:—

TABLE II

Triple Analogies

$\mu = B/H$	$\kappa = J/E_c$	$\epsilon = D/E_c$
$B = \Phi/F$	$J = i/F$	$D = \Psi/F$
$W_k = \frac{1}{2}M\Phi$	$P = ei$	$W = \frac{1}{2}eq$
$D_k = \frac{1}{2}HB$	$N = E_cJ$	$D_p = \frac{1}{2}E_cD$

Double Analogies

$H = M/l$	$E_c = e/l$
$M = \Phi R_M$	$e = iR_E$
$R_M = l/\mu F$	$R_E = l/\kappa F$

Without Analogies

$C = \epsilon F/l$, $L = n^2/R_M$, $M = ni$, $q = eC$, etc.

In these formulae F = area, E_c = electric field strength, J = current density, W = energy, D = energy density, also displacement density P = power, N = power density, e = e.m.f., M = m.m.f., R_e = resistance, R_m = reluctance, κ = conductivity. We have copied the author's symbols except where they were obviously wrong.

In the triple analogies the first column is magnetic, the third electrostatic or dielectric and the second column is what the author regards as their electric analogues. In the double analogies one of the three is missing from ordinary usage, and there are many formulae without any analogy. The author's object is to fill these gaps as far as possible by introducing new conceptions and reshaping equations. He claims that this results in a perfectly symmetrical classification of electromagnetic quantities and laws. A tabulation is proposed, which he calls the Pe iodic System, "in which all known quantities are included and the laws are shown by the geometry of the tabulation". This is applied to other branches of physics, and he develops three periodic systems, viz., electromagnetic, mechanical and thermal, which he then unifies in a single general periodic system, with three absolute basic dimensions, length, time, and energy (or one of its time derivatives) and one relative dimension, viz., "the impedance of the domain in question". The name 'periodic' is adopted from Mendeleev's chemical system.

An interesting example of the author's procedure is the introduction of the multi-plate-pair capacitor. The reciprocal of the capacitance is called the dielectric reluctance R_d and in the formula $R_d = l/(\epsilon F)$, l and F are regarded as the length and cross-section of the dielectric core, just as in the formula $L = 1/R_m$ for the inductance of a single turn inductor, R_m is the reluctance of the core. If the inductor has n turns $L = n^2/R_m$, and to obtain the dielectric analogy, the author uses a number of capacitors in series but all using the same slab of dielectric as shown in Fig. 1. If the reluctance R_d of the whole slab is thus divided between p pairs of plates, the reluctance of each section will be pR_d and the capacitance of the p pairs in series will be

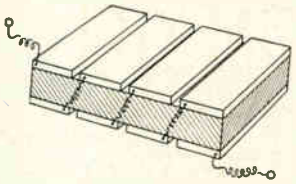

Fig. 1

$$C = (1/p)^2/R_D$$

which is analogous to

$$L = n^2/R_M$$

The dual of n is $1/p$ because ωL and ωC are of an inverse character. This example shows how, by departing somewhat from ordinary usage, analogies can be devised and the gaps in the table filled up.

His first example of the lack of analogy between magnetic and electric formulae is rather surprising; he says that although $\int_o Hdl = M$ where M is the m.m.f., the line integral $\int_o F_c dl = 0$ where E_c is the electric field strength. This is surely wrong, since the line integral of E_c in a closed path is the e.m.f. which he designates e. He returns to the subject later on and it is seen that the anomaly is of his own invention. In a section entitled "Reshaping of the Maxwell vector equations" he uses three different symbols for electric field strength, viz., E, E_c and U; he says, "Up to now we have used a common space derivation, $E_c = de/dl$. Obviously we must split this into dielectric field strength $U = du/dl$ and the rate of voltage drop $E = de/dl$." The word 'obviously' is ominous. Four lines after stating that $\int_o E_c dl = 0$, he has curl $E_c = -dB/dt$ which seems contradictory. He then says, "U is restricted to media characterized by ϵ or dielectrics; for conducting media characterized by κ it is the rate of voltage drop which is to be used." He does not say what is to be done when current flows in a conductor without any voltage drop. Consider two closed rings, one of copper and the other of a dielectric, on an iron core through which the magnetic flux is changing. The e.m.f. e induced will be the same in each ring, viz., $-d\Phi/dt$ and the electric field strength E in each case will be the e.m.f. divided by the length of the path around the ring. In the copper ring the current will be e/R and the current density $J = E/\rho$; in the dielectric ring the displacement density $D = E\epsilon$ where ϵ is the permittivity of the material. In neither case does potential enter into the case since all points of the ring are at the same potential; in both cases the effect is due to the induced electric field and not to fall of potential. In a recent programme for a meeting of the Education Discussion Circle of the Institution of Electrical Engineers we were surprised to see that one of the questions to be discussed was "Is there any electric field strength in a short-circuited ring?" One of the free electrons might be consulted as to why it was moving. Perhaps further light is thrown on his point of view by considering a dynamo and its external circuit. In the external circuit the electric field strength at any point is equal to the fall of potential per unit length, but in the armature winding there are three electric forces, viz., the induced electric field due to the movement in the magnetic field, the opposing electric field due to the fall of potential and the resultant electric field equal to the difference between them. The only one of these for which the line integral around the circuit is zero is that due to the fall of potential, and the author must have had this in mind when he wrote $\int_o E_c dl = 0$, although this does not agree with curl $E_c = -dB/dt$ which he writes four lines further on. Since the line integral around the circuit of the forces due to the potential difference must be zero, the integral of the actual resultant force must be the same as that of the induced electric force; i.e., $-n\,d\Phi/dt$.

On page **42** he states that E = electric rate of voltage-drop and U = dielectric field-strength, and on page **43** we are told that $U = -E$ and curl $E = dB/dt$ and then on page **44** that curl $E = -dB/dt$. Apart from this final contradiction, surely the field-strength or electric force in a dielectric is in the same direction as the voltage drop.

We have only touched on a fraction of this lengthy paper, but we must confess that, after struggling with these apparent contradictions for many hours, we gave it up. In view of the great amount of work that has evidently been put into the paper, it is to be hoped that the author will make an effort to set out more clearly the initial basic sections.

G. W. O. H.

A Heaviside-Lorentz Network Theorem

IN both the April and May Editorials we referred to a theorem of Heaviside which was developed by Lorentz and recently discussed in a paper by van der Pol and also in a paper by Tellegen. In both cases the reference given was to Heaviside's Electrical Papers, Vol. II, p. 412. Our attention has been drawn to the fact that Heaviside is here merely quoting a special application of a general theorem which he had derived in Vol. I, p. 462, where a section is entitled "Work done by impressed forces during transient states".

G. W. O. H.

SELF-HEATING TRIODE FOR VOLTAGE STABILIZATION

By E. G. Hopkins, B.E., A.M.I.E.(Aust.).

(School of Electrical Engineering, N.S.W. University of Technology, Sydney.)

SUMMARY.—A self-heating triode for use in voltage-stabilizing circuits is described. It contains two planar oxide cathodes separated by a planar grid and acts as a variable resistance directly in series with the a.c. supply to small fixed loads. The valve has two auxiliary anodes which connect the grid circuit to the appropriate cathode during each half-cycle. A large cathode area of 8·4 cm² relative to the total power dissipation of 16 W renders the triode more suitable than conventional valves for operation in low-voltage circuits.

1. Introduction

THE alternating voltage supplied to a piece of small electrical equipment is sometimes stabilized or controlled by a pair of high-vacuum triodes or pentodes acting as a variable resistance.[1] The valves are usually connected across the centre-tapped secondary of a transformer whose primary is placed in series with the supply line. The transformer is necessary to increase the low equivalent conductance and the low rated current of standard receiving valves to more appropriate values.

The present paper describes a self-heating triode[2, 3] designed especially to replace the two valves and the transformer by acting as a variable a.c. resistance directly in the supply line to a constant or nearly-constant load. The triode has two parallel planar cathodes equally spaced about a planar grid. Each cathode acts alternately as anode and cathode. Heat generated by electron bombardment maintains the cathode temperature within its working range. The cathodes are provided with heaters which are used only for starting. Two auxiliary anodes clamp the grid circuit to the appropriate cathode during each half-cycle of operation.

The basic principles of self-heating valves are discussed in other papers.[2, 4]

2. Description of the Valve

The working parts of the triode are illustrated in Fig. 1. The cathodes are 42 mm in length and have a hollow rectangular cross-section measuring 10 mm × 1·5 mm outside, with a wall thickness of 0·25 mm. They are made of polished Grade A nickel and are supported at each end in the mica spacers by two nickel rods 1 mm in diameter. The cathodes are induction-heated during exhaust. One face of each cathode is completely coated with barium-strontium oxide. A length of 1 cm at the centre of each reverse side is also coated to form the emitting surface of a clamping diode.

Each cathode is provided with a well-insulated 6·3-V, 0·6-ampere heater which is retained within the cathode between 0·85-mm diameter nickel rods. It is proposed to replace the present heaters in future valves by heaters rated at 0·2 ampere and 33 V which will be connected in place of the electron stream during the starting period, as is shown in the circuit of Fig. 2. Any value of heater power between the limits of about 6 W and 15 W is satisfactory for starting the triode.

The grid consists of molybdenum wires 0·175 mm in diameter welded to 1-mm nickel side rods. Each side rod carries four carbonized nickel cooling fins 10 mm long and 5 mm wide. The plain nickel clamping anodes measuring 1 cm × 1 cm are attached to two supporting members which are eyeletted to the mica spacers.

The spacing between cathodes is 0·8 mm. This relatively large spacing is used only to avoid the necessity for close mechanical tolerances.

The envelope is 29 mm in diameter and 85 mm in length. The envelope, stem, base and heaters are standard "GT" receiving-valve parts.

3. Performance of the Valve

Fig. 2 shows a circuit in which the self-heating triode is connected, together with the shunt

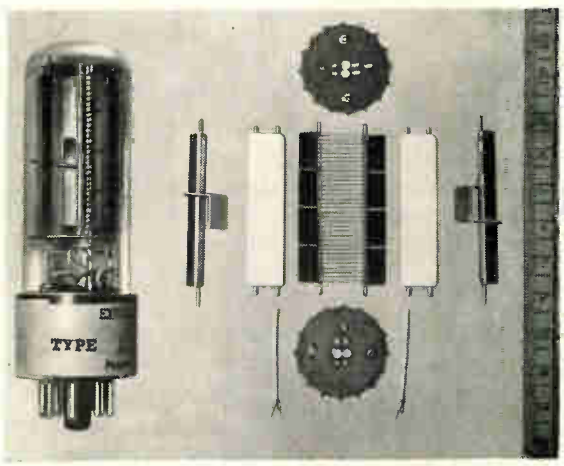

Fig. 1. *The parts of the self-heating triode.*

MS accepted by the Editor, September 1953

resistor R_s, between a constant load and an a.c. supply which varies about a nominal voltage of 240. Because the resistances R_g are large compared with those of the clamping diodes, each cathode is effectively connected to one end of the grid transformer winding during the half-cycle in which it is negative with respect to the other cathode. A control signal in phase with the anode voltage V_a is fed into the transformer so that the grid is driven negative with respect to the appropriate cathode by V_g, a voltage of sinusoidal form. The control signal, which is derived from a measuring circuit, maintains the load current, I_L constant as the supply voltage changes.

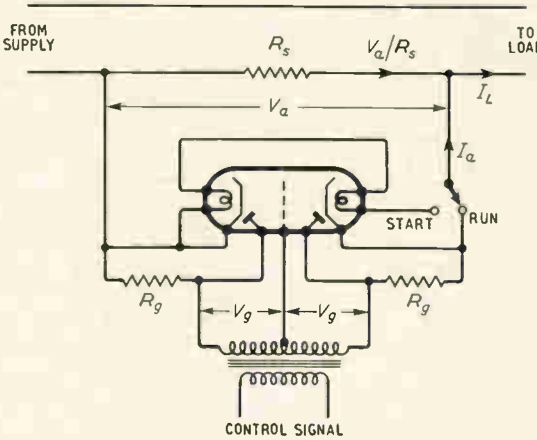

Fig. 2. *A circuit in which the self-heating triode acts as a variable resistance to maintain a constant load current, I_L when controlled by a mains-frequency signal derived from a measuring circuit.*

The characteristic curves of r.m.s. anode current, I_a, versus r.m.s. anode voltage, V_a, measured with sinusoidal anode and grid voltages are shown in Fig. 3. Since $I_L = I_a + V_a/R_s$, the operating point is defined by the intersection of the straight line $I_a = I_L - V_a/R_s$ and the V_a/I_a characteristic for the particular grid voltage being considered. Harmonic currents may be ignored.[2] The region between the curves of maximum and minimum dissipation (16 W and 10 W) defines the safe working range of the valve. The power delivered to a load of unity power factor is I_L $(240 - V_a')$, where V_a' is the centre value of anode voltage. R_s is chosen to give the best centre value and range of anode voltage with the specified load current. For example, when R_s is 400 ohms a control of ± 22 V is possible for a load current of 0·4 ampere (56 W load power), and when R_s is 600 ohms ± 40 V variation is possible for a load current of 0·32 ampere (40 W load power). The voltage-gain of these circuits is about unity and the total harmonic distortion introduced into the load current is about 2%.

4. Comments

The emitting areas of the cathodes are much greater than those of two conventional valves of similar total bulb dissipation. For example, if a double triode were designed with cathodes of similar area to those of the present self-heating triode it would require a filament power alone of at least 16 W, which equals the total bulb dissipation of the self-heating triode. This power would have to be dissipated by the anodes and bulb of such a valve in addition to the power lost at the anodes themselves.

The present triode will operate stably between the limits of 8-W and 25-W dissipation. The range 10 W to 16 W quoted above was chosen arbitrarily pending an investigation of life performance with relatively large numbers of valves.

With negative grid voltage, the convergence of the electron streams between grid wires causes a large increase in current density on the bombarded regions of the cathode surfaces. In the present experimental valves, which use sprayed coating and highly polished base metal, considerable localized erosion of cathode coating takes place. Experiments are continuing to ascertain whether the base metal surfaces[5, 6] and coatings, which have been developed to improve thermal and electrical conduction between cathode surface

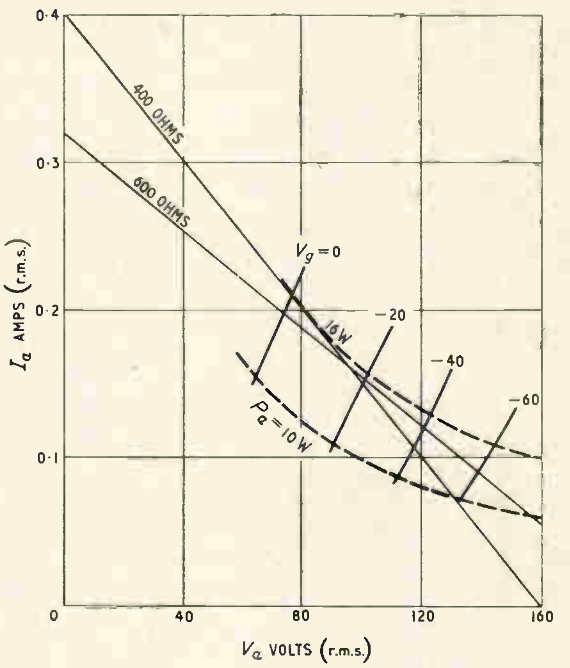

Fig. 3. *Characteristic curves of r.m.s. anode current versus r.m.s. anode voltage with sinusoidal anode and grid voltages. The load lines intersect the current axis at the load current I_L and have a slope equal to R_s, the shunt resistance of Fig. 2.*

and base metal, will reduce this effect to negligible proportions.

It is thought that the performance of the present valve may be improved by a small increase in cathode area and a decrease in inter-electrode spacing to accommodate a variation in V_a of ± 40 V about a centre value of 80 V for a load current of 0·5 ampere. This performance would be adequate for stabilizing the supply voltage to pieces of electronic equipment with power consumptions of up to 80 W or the heater voltage of a group of about 40 receiving valves.

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SPECTRUM OR WAVEFORM EQUALIZATION?

By D. A. Bell, M.A., Ph.D., M.I.E.E.

(University of Birmingham)

SUMMARY.—The relationship between the wave-filter approach to equalization and the approach via differentiating and integrating circuits is discussed in the light of their historical development and of present-day needs in both communication and servo systems.

1. Introduction

THE supremacy of the electric wave filter in circuit engineering is a phenomenon of the past quarter-century; and this, together with the general use of steady-state methods of circuit calculation, has tended to blind us to the possibility of carrying out engineering design in the time–amplitude plane instead of in the frequency plane. It was, therefore, very refreshing to read G. G. Gouriet's paper[1] (strangely entitled "*Spectrum* Equalization" when it approaches the problem from the angle of time functions rather than frequency functions) on the use of differentiating and integrating networks to restore the waveform of a signal which has been distorted at some point in the communication channel.* It is of course agreed that no system of correction can restore a component of the signal which has been *entirely* eliminated in transmission, or has been reduced to an intensity well below noise level: any process of equalization must, in principle, operate by levelling down to the weakest of the surviving components, accompanied usually by a uniform scalar amplification of the whole signal either before or after equalization. Moreover, if the specification for equalization requires that it should introduce no time delay, one can only work on the components as they exist at the first instant of the received signal.

An example may clarify this last point. If delay is permitted, the response of a typical physically-occurring low-pass filter can be phase-equalized by means of all-pass filter sections;[2] this will improve the symmetry and steepness of rise of the output response to a step-function input without introducing additional attenuation, but it will introduce delay. Alternatively, Bode has stated[3] that the frequency-discriminating effect of any physically-realizable minimum-phase network can be cancelled by adding in series with it an inverse network, the pair together then having a transfer characteristic of zero phase and constant magnitude. The apparent time delay (Gouriet's 'virtual delay') of the first circuit must then have been cancelled by the second circuit, but only at the cost of an overall attenuation not less than that at the critical point on the characteristic of the first network. (The 'critical point' may be defined either in terms of spectral response as the maximum frequency which is required to be transmitted, or in terms of waveform response as the earliest time at which full response to a step function is required.)

* The author has also had the pleasure and benefit of personal discussion of this topic with Mr. Gouriet.

MS accepted by the Editor, November 1953

2. History

It is interesting now to note that an early proposal to use differentiating circuits for equalization related to the telegraph signals received through a submarine cable; i.e., it was concerned with counteracting the frequency-dispersion of a distributed circuit. This was described in U.S. Patent No. 1,315,539, issued to J. R. Carson on 9th September, 1919, the first claim being as follows:

"In a wave form correcting device, means for producing from the arrival current separate currents having wave forms corresponding substantially to the arrival current itself and one or more of its derivatives respectively, and means for superimposing the separate currents in a single output circuit."

Carson proposed in the first place to use inductive differentiating circuits (the low-inductance primary of a transformer was to be fed from a comparatively high-impedance source so as to carry a current proportional to the signal amplitude, and the secondary e.m.f. proportional to the derivative of the current was to be applied to a valve grid or other high-impedance circuit) with valves as buffer amplifiers between successive differentiations. Partial outputs could be tapped off along a chain of such differentiators to give a sum of the form

$$V_o = (1 + ap + bp^2 + cp^3 + \ldots)V_i \quad .. \quad (1)$$

where V_o is the output of the equalizing system, V_i its input, and $p = d/dt$. So long as the differentiating circuits are nearly perfect and the various derivatives are combined by simple addition, such a system has the advantage that the several coefficients $a, b, c \ldots$ in equation (1) can be adjusted separately. But Carson also suggested taking the output simply from the end of a cascade of imperfect differentiators. The output would then be of the form

$$V_o = (1 + a_1 p)(1 + a_2 p)(1 + a_3 p) \ldots (1 + a_n p) V_i \quad .. \quad (2)$$

and the coefficients of the powers of p in the polynomial form of the right-hand side of (2) would not be mutually independent.

Formula (2) is in fact a typical circuit *frequency response*, in which p is interchangeable with $j\omega$. The transition from the 'derivative' interpretation to the 'filter' interpretation is carried a stage further by considering American patent No. 1,586,821 issued to R. C. Mathes on 1st June, 1926.* Mathes placed far less stress on the use of buffer amplifiers and the desirability of mutual independence of the several differentiating circuits. He also proposed quite complex circuits, ranging from Fig. 1(a) to (b), as the 'differentiating' units, though these clearly would not produce pure single derivatives. When one has circuits as complex as this with, in some cases, mutual interaction between successive stages, one would nowadays design them in terms of frequency response as filters; the mathematical theory of electric wave filters barely existed in 1919, but has since become highly developed.

* Carson's patent, dated 9th September, 1919, was based on his application filed on 3rd July, 1918. Mathes's application was originally filed on 19th June, 1918 (i.e., a few days *before* Carson's), but was left in abeyance until 27th March, 1925, so that Mathes's patent was not granted until 1926, seven years after Carson's.

3. Application to Communication Systems

Now Gouriet proposes[1] to abandon the idea of equalization of the spectrum by wave filters, and to design instead in terms of equalization of the system response by means of differentiating circuits. Two questions are inevitable: "Is this really different?" and "Does it give any substantial advantage?" So far as the ultimate theory is concerned, there is no difference. The loose statement above that p and $j\omega$ are interchangeable can for this purpose be made sufficiently rigorous by pointing out that the various forms of operational calculus allow the time-function response to be completely derived from a frequency (p or $j\omega$) specification of the circuit. Hence correction of the $j\omega$ response necessarily corrects the waveform response and vice versa.*

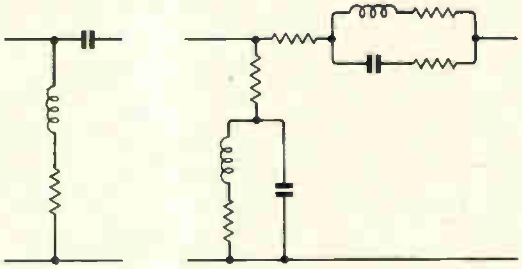

Fig. 1. *Complex differentiating circuits proposed by Mathes.*

The practical advantage of derivative equalization, however, is that it greatly simplifies the construction of an approximate correcting circuit for the overall distortion which may have been contributed to by many separate circuits of unknown form. The theoretical method of equalizing by networks might be to factorize the total transfer function so that its inverse could be constructed from a cascade of networks, using in principle one section of correcting network per factor in the transmission function; but this is cumbersome, and one would prefer to construct empirically a correcting network of polynomial form having the least number of terms consistent with a fair fit to the required characteristic. The great advantage of using pure derivatives is that each term in the polynomial can be adjusted *independently*, whereas a cascaded-network response of the type of formula (2) is such that change of any one parameter will affect the values of most of the terms in the polynomial and vice versa.

On these grounds it is difficult to see why the use of derivative equalization has been neglected from the time of Carson (1919) to Gouriet (1953).

* Only if the frequency response is expressed as a function of ω (instead of $j\omega$) will there be a possibility of failure through lack of phase agreement.

A possible answer is that whereas the wave filter is a passive network, and therefore has stable characteristics, the derivative equalizer commonly has a number of separate amplifiers, the relative gains of which affect the proportion of the different derivatives in the output; i.e., the equalization characteristic. This objection can now be largely overcome by the use of negative feedback to stabilize gains; e.g., by using cathode-followers as buffer stages between successive differentiating networks.

4. Application to Feedback and Servo Systems

For practical reasons, however, it is desirable to distinguish between communication systems and servo systems. In the former it can usually be assumed that although the communication channel is of limited bandwidth, the terminal equipment can be made with virtually infinite bandwidth; but in the servo system there is no such distinction between 'channel' and 'terminal equipment'. It is true that there may be a distinction between valve amplifying stages, of virtually infinite bandwidth, and rotary-machine amplifying stages of very limited bandwidth; but in this case the bandwidth limitation comes second in the chain, and one cannot put much pre-correction in the valve stages without reaching the power limitation which is the sole reason for the existence of the machine amplifying stages.

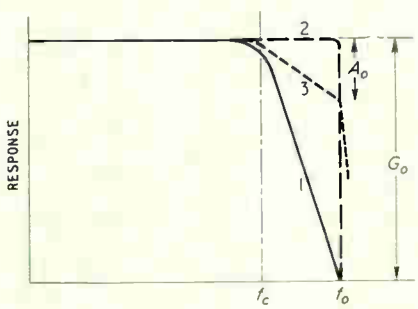

Fig. 2. *Gouriet's method of stabilization. Curve 1 is for the original characteristic and curve 2 for the equalized system. Curve 3 shows the final characteristic including a quadratic filter.*

In connection with feedback systems, Gouriet proposes an apparently new principle of stabilization: make the response perfect over a frequency range extending far beyond the required working band, then insert a quadratic low-pass filter to provide the cut-off. The quadratic filter cannot produce a phase-shift in excess of 180°, so the complete system is necessarily stable. In principle, however, this in no way differs from the Bode technique of adjusting the rate of cut-off of the system. According to Gouriet, stabilization is to be achieved as shown in Fig. 2: the original characteristic with nominal cut-off at f_c (curve 1) is equalized out to some higher frequency f_o (curve 2) and then a new and safe cut-off characteristic (curve 3) is introduced by means of a quadratic filter; i.e., the slope of curve 3 is less than 12 db per octave for a sufficient range.

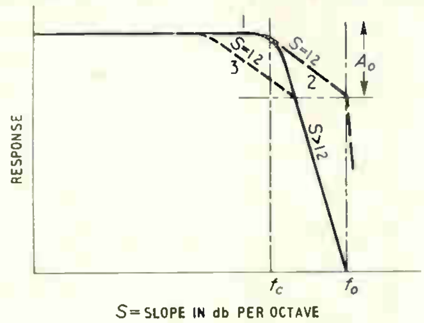

Fig. 3. *Bode's method of stabilization. S is the slope of the curve in db/octave.*

This clearly requires additional gain at high frequencies, the amount of which is arrived at as follows: (i) The additional gain G_o at f_o must equal the loss of the original circuit at f_o. (ii) The amount of attenuation A_o introduced by the quadratic filter at f_o must be at least equal to the loop gain, (to ensure stability when the overall characteristic leaves the quadratic characteristic 3 and begins to take on the large phase-shift which must be associated with the sharp cut-off at f_o) and this controls the distance of f_o from f_c. Now stabilization by the Bode type of technique is illustrated in Fig. 3. By increasing the bandwidth of a sufficient number of the stages, the open-loop cut-off is eased to 12 db/octave until the attenuation A_o exceeds the loop gain, after which the cut-off characteristic is allowed to take its own course (curve 2). Alternatively, if greater overall bandwidth is not obtainable one must ease the slope back as shown in curve 3, at the cost of curtailment of the working frequency band. The final result is then the same in Fig. 2 and Fig. 3—a cut-off slope reduced to a value not greater than 12 db/octave down to the point of zero loop gain. (Both methods ignore the question of shaping the characteristic within the working frequency band, so the final result may not after all be ideal for a servo system, but that is another matter.) However, the claim which may well be made for the system of Fig. 2 is that the best approximation to the equalization of curve 2 by means of derivatives can easily be established and verified empirically, and the quadratic filter is then simple to design and certain in operation. The design of the modifying filter for Fig. 3 is far more tricky, especially as calculated values of system characteristics can be unreliable in the presence of non-linearities.

Any equalization technique requires additional amplification, whether this be included in series with the main signal chain and a network type of equalizer, or in parallel where the derivatives of the signal are amplified to the required extent. If there is no practical limitation and little economic limitation on increasing the amplification of the system at a fixed bandwidth, this presents no difficulty; but in a servo system real difficulties can arise.

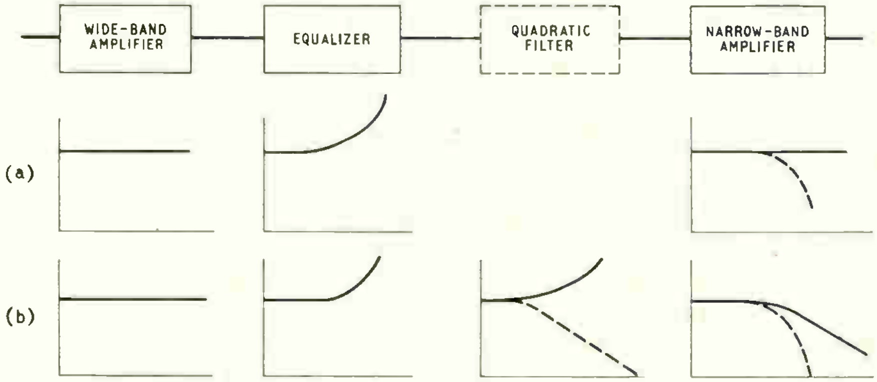

Fig. 4. *Transfer characteristic (a) for equalized system; (b) for equalized system plus quadratic filter. Cumulative characteristic shown broken line, unit characteristic shown solid line.*

In theory the gain-bandwidth product of a given type* of amplifier is independent of power rating, and in practice the gain-bandwidth may decrease only slowly as the power rating is increased. But as already remarked, there is for various reasons a change of type of amplifier from thermionic valve to rotary machine at a certain power level in most servo systems; and the bandwidth of the whole is effectively controlled by the machine stages. The achievement of curve 2 in Fig. 2 therefore implies a very large pre-distortion of the signal in the thermionic stages before it is applied to the machine stages, as shown diagrammatically at (a) in Fig. 4 for the exactly equalized system and at (b) with the addition of a quadratic cut-off. Even case (b) indicates a considerable rise in the peak power input to the narrow-band amplifier; and if the output power of the wide-band amplifier is fixed, the mean input to the narrow-band amplifier must necessarily be reduced, and its gain correspondingly increased. This will require more stages, and the question is whether one might not just as well have used the extra stages to shape the cut-off characteristic of the narrow-band amplifier, thus saving the equipment involved in the equalizer. As a first step one would in any case surely use part of the narrow-band amplifier as the quadratic filter, and this complicates the experimental setting-up of the equalizer.

There is certainly a case for adjusting the characteristics of a rotary amplifier by means other than a cascaded network, (e.g., by some form of feedback) because of the physical difficulties of constructing a network for extremely low frequencies; but this is not comparable with *equalization* by the addition of derivatives, since the latter requires additional gain with negligible time-constant (or in other words, additional gain at high frequencies) for the derivative components, and this additional gain is not available in a narrow-band amplifier. It is sometimes thought to be available when the amplifier output circuit has a much shorter time-constant than its input circuit.* But if the design of the amplifier or the load-matching conditions have been so adjusted that the input and output time-constants are approximately equal, there is in fact no such reserve.

5. Conclusions

The addition to the signal of derivatives amplified in parallel paths, instead of modification of the signal by frequency-selective networks connected in series with the main amplifying chain, greatly facilitates the empirical design of the correction for a given system; but the results which can, in principle, be obtained from the two systems (disregarding economic or constructional limitations) are identical, and the use of derivative circuits does not avoid the need for additional amplification at high frequencies (or with short time-constant, if one prefers to express it in that way). When using the derivative method, it must be remembered that the result depends on the constancy of the relative gains of any parallel paths through which different components of the output signal may be amplified.

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[2] A. E. Brain, "The Compensation for Phase Errors in Wide-band Video Amplifiers", *Proc. Instn elect. Engrs*, 1950, Vol. 97, Pt. III, p. 243.
[3] H. W. Bode, "Feedback Amplifier Design and Network Analysis", Van Nostrand, New York, 1945.

* E.g., thermionic valve, rotary machine, magnetic amplifier.

* Bode (ref. 3, p. 284) gives an example of the converse case, a cathode-feedback circuit in which the cathode–earth capacitance is assumed to provide the only time-constant.

L.F. COMPENSATION FOR VIDEO AMPLIFIERS

Part 1. Amplifiers without Feedback

By J. E. Flood, Ph.D., A.M.I.E.E.

SUMMARY.—The effect of distortion at low frequencies on the indicial response (i.e., the response to a unit-step input voltage) of amplifier circuits is considered. Low-frequency distortion causes the output waveform to depart from the ideal flat top of the unit step, but it can be made as flat as possible by making the maximum number of its differential coefficients to be zero at $t = 0$. If r derivatives are zero, the amplifier is said to have rth-order compensation. A general method for doing this is described and is applied to some typical resistance-coupled amplifier circuits. A single resistance-capacitance coupled stage can have second-order compensation and multi-stage amplifiers can have still higher orders of compensation. Experimental results confirm the analysis.

1. Introduction

VIDEO-FREQUENCY and pulse amplifiers must introduce as little change as possible in the waveforms of the signals they amplify. A useful criterion for judging the fidelity of an amplifier is its indicial response; that is, its response to a unit-step input voltage (i.e., a voltage which is zero when $t < 0$ and is unity when $t > 0$). The bandwidth of the amplifier is large, so the effects of gain and phase distortion at low frequencies and at high frequencies can be considered independently. Analyses of high-frequency distortion can be made from equivalent circuits which omit the coupling and decoupling networks. Analyses of low-frequency distortion can be made from equivalent circuits which omit stray capacitances and components included to compensate for their effect.

Distortion at high frequencies affects the initial build-up period of the indicial response and may cause overshoot. Distortion at low frequencies affects the subsequent waveform, causing it to depart from the ideal flat top of the unit step. The output waveform can be perfectly flat only if the frequency response is uniform down to zero frequency. If the amplifier cannot pass d.c. the output voltage ultimately must tend to zero with increasing time, but it is possible to design an amplifier so that the response may approximate to the ideal within given limits for a specified time.

The effects of high-frequency distortion have been fully dealt with elsewhere. In this paper the effect of distortion at low frequencies is analysed for some typical amplifier circuits. The indicial response is made as flat as possible by causing as many as possible of its differential coefficients to be zero at $t = 0$, as suggested by W. E. Thomson.[1] Some well-known results are included as particular cases.

MS accepted by the Editor, October 1953

2. General Theory

The indicial response of the low-frequency equivalent circuit is given by its Heaviside operational expression $h(p)1$ where

$$h(p) = \frac{p^n + a_1p^{n-1} + a_2p^{n-2} + \ldots + a_rp^{n-r} + \ldots + a_n}{p^n + b_1p^{n-1} + b_2p^{n-2} + \ldots + b_rp^{n-r} + \ldots + b_n} \quad \ldots (1)$$

Expanding this in inverse powers of p gives

$$h(p) = 1 + \frac{c_1}{p} + \frac{c_2}{p^2} + \ldots + \frac{c_r}{p^r} + \ldots \quad \ldots (2)$$

The values of the coefficients in this expansion are derived in Appendix 1 and are as follows:

$$c_1 = a_1 - b_1$$
$$c_2 = a_2 - b_2 - a_1b_1 + b_1^2$$
$$c_3 = a_3 - b_3 - a_2b_1 + 2b_1b_2 - a_1b_2 + a_1b_1^2 - b_1^3$$
$$\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots$$
$$c_r = \sum_{s=0}^{r-1} d_s(a_{r-s} - b_{r-s}) \quad \ldots \quad \ldots \quad \ldots (3)$$

$$\text{where } d_s = \sum \frac{(-1)^m m!}{\alpha!\,\beta!\ldots\epsilon!} b^{\alpha}{}_{\lambda} \cdot b^{\beta}{}_{\mu} \ldots b^{\epsilon}{}_{\rho} \quad (4)$$

where the summation extends over all possible values of $\alpha, \beta \ldots \epsilon, \lambda, \mu, \ldots \rho$ and where

$$m = \alpha + \beta + \ldots + \rho; \quad \alpha\lambda + \beta\mu + \ldots + \epsilon\rho = s;$$
$$1 \leqslant \lambda < \mu \ldots < \rho \leqslant s; \quad a_0 = b_0 = d_0 = 1$$

Now $\quad \frac{1}{p^r} 1 = \frac{t^r}{r!} 1$

Therefore from equation (2):

$$h(t) = 1 + c_1t + c_2\frac{t^2}{2!} + c_3\frac{t^3}{3!} + \ldots$$
$$+ c_r\frac{t^r}{r!} + \ldots (t > 0)$$

But by Maclaurin's theorem

$$h(t) = h(0) + h_1(0)t + h_2(0)\frac{t^2}{2!} + \dots$$
$$+ h_r(0)\frac{t^r}{r!} + \dots$$

where $h_r(0)$ is the rth differential coefficient of $h(t)$ at $t = 0$.

$$\therefore \quad h(0) = 1$$
$$h_1(0) = c_1 \ , \ h_2(0) = c_2$$
$$\dots\dots\dots\dots\dots\dots\dots$$
$$h_r(0) = c_r \text{, etc.}$$

The indicial response may be made to approach more and more nearly to the ideal $[h(t) = 1$ for all $t > 0]$ by making more and more of the differential coefficients vanish at $t = 0$. Equation (3) shows that if $a_1 = b_1$, $a_2 = b_2$, . . . $a_r = b_r$, then $c_1 = c_2 = c_3 = \ . \ . \ . = c_r = 0$ and the first r differential coefficients of $h(t)$ vanish at $t = 0$. The amplifier will then be said to have rth-order compensation of its indicial response. If the indicial response has $(n - 1)$th-order compensation, it will be called maximally flat. If $a_1 = b_1$, $a_2 = b_2$, . . . $a_n = b_n$, then equation (1) clearly gives a completely distortionless indicial response. It is shown in Section 5 that, except when $n = 2$, a maximally-flat indicial response corresponds to neither a maximally-flat gain–frequency response nor a maximally-flat phase–frequency response. This confirms that the gain–frequency response and phase–frequency response are not entirely satisfactory for judging the waveform distortion caused by gain and phase distortion at low frequencies.

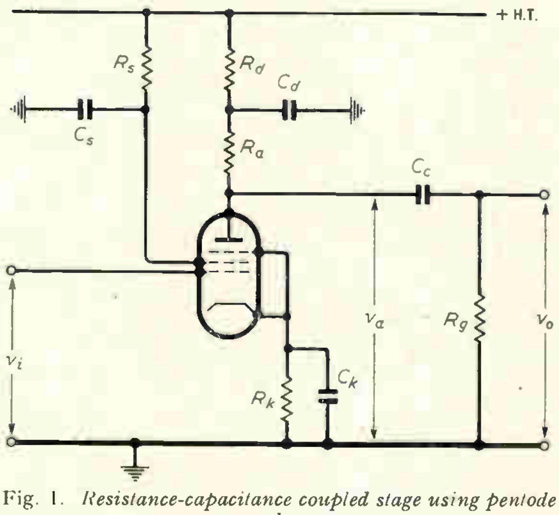

Fig. 1. *Resistance-capacitance coupled stage using pentode valve.*

3. RC Coupled Stage

Fig. 1 shows a typical resistance-capacitance coupled amplifier using a pentode valve. The usual assumptions will be made, namely:

(i) the anode and screen currents are independent of anode voltage, and the ratio betweeen them is independent of grid voltage, screen voltage and anode voltage;
(ii) the impedance of the d.c. power supply is negligible;
(iii) the grid resistance R_g is large compared with the anode load resistance R_a and decoupling resistance R_d.

Edwards and Cherry[2] have shown that the gain between grid and anode is then given by

$$\frac{v_a}{v_i} = \frac{g_m (R_a + Z_d)}{1 + g_k Z_k + \sigma_s Z_s} \qquad \dots \qquad \dots \qquad (5)$$

where g_m is the grid–anode mutual conductance.

g_k is the sum of the screen conductance and the grid–anode, grid–screen and screen–anode mutual conductances.

σ_s is the screen conductance.

Z_d, Z_k and Z_s are the impedances of the anode, cathode and screen decoupling networks respectively.

The operational expression for v_a when v_i is a unit step is therefore

$$v_a = g_m R_a \frac{1 + \dfrac{R_d}{R_a} \cdot \dfrac{1}{1 + pC_d R_d}}{1 + g_k \dfrac{R_k}{1 + pC_k R_k} + \sigma_s \dfrac{R_s}{1 + pC_s R_s}} 1$$

$$= G \frac{1 + \dfrac{\gamma}{p + \lambda}}{1 + g_k R_k \dfrac{\eta}{p + \eta} + \sigma_s R_s \dfrac{\theta}{p + \theta}} \qquad \dots \qquad (6)$$

where $\gamma = 1/R_a C_d$, $\lambda = 1/R_d C_d$

$\eta = 1/R_k C_k$, $\theta = 1/R_s C_s$

and $G = g_m R_a$

$$\text{Now} \quad \frac{v_0}{v_a} 1 = \frac{pC_c R_g}{1 + pC_c R_g} 1 = \frac{p}{p + d} 1 \qquad (7)$$

where $d = 1/R_g C_c$

It follows from equations (6) and (7) that the overall indicial response is given by

$$v_0 = G \frac{p^4 + a_1 p^3 + a_2 p^2 + a_3 p}{p^4 + b_1 p^3 + b_2 p^2 + b_3 p + b_4} 1 \qquad (8)$$

where $a_1 = \gamma + \lambda + \eta + \theta$

$a_2 = \eta\theta + (\gamma + \lambda)(\eta + \theta)$

$a_3 = \eta\theta(\gamma + \lambda)$

$b_1 = \alpha + \lambda + \eta(1 + g_k R_k) + \theta(1 + \sigma_s R_s)$

$b_2 = \alpha\lambda + \eta(\alpha + \lambda)(1 + g_k R_k) + \theta(\alpha + \lambda)(1 + \sigma_s R_s) + \eta\theta(1 + g_k R_k + \sigma_s R_s)$

$b_3 = \alpha\lambda\eta(1 + g_k R_k) + \alpha\lambda\theta(1 + \sigma_s R_s) + \eta\theta(1 + g_k R_k + \sigma_s R_s)(\alpha + \lambda)$

$b_4 = \alpha\lambda\eta\theta(1 + g_k R_k + \sigma_s R_s)$

3.1. *Compensation for Anode Decoupling by means of Cathode and Screen Decoupling*

At low frequencies, the gain tends to rise because of the increase in the anode load impedance due to the finite reactance of C_d, but there is also a fall in gain due to the negative feedback introduced by the finite reactance of C_k and C_s. Edwards and Cherry[2] have shown that these effects can be made to cancel completely so that the gain between grid and anode is independent of frequency. Other circuits in which an increase in anode load impedance offsets screen and cathode feedback at low frequencies have been described by Zeidler and Noe.[3]

From equation (6) it follows that

$$v_a = G\,1$$

if $$\frac{\gamma}{p+\lambda} = g_k R_k \frac{\eta}{p+\eta} + \sigma_s R_s \frac{\theta}{p+\theta}$$

$\therefore \quad \gamma = \eta\, g_k R_k + \theta \sigma_s R_s$

and $\quad \lambda = \eta = \theta$

i.e., $$C_d R_d = C_k R_k = C_s R_s \qquad (9)$$

and $$R_d = R_a\,(g_k R_k + \sigma_s R_s) \qquad (10)$$

If equations (9) are satisfied, the anode waveform is distortionless. However, unless the distortionless stage has d.c. coupling to the next stage, distortion will be introduced by the coupling capacitance. The overall indicial response is then of zero order, it being that of the coupling circuit alone, namely a simple exponential curve with time constant $1/\alpha$. This time constant must therefore be adequately large to keep the sag on the top of the waveform within the permitted limit.

3.2. *Compensation for the Coupling Circuit by means of the Anode Decoupling Circuit*

At low frequencies the gain is reduced by the increasing reactance of the coupling capacitor (corresponding to a sagging indicial response), but the increasing reactance of the anode decoupling capacitor tends to increase the gain (corresponding to a rising indicial response). The effects of the coupling and anode decoupling circuits can thus be made to offset each other. Perfect compensation cannot be obtained because the coupling circuit reduces the gain to zero at zero frequency but the increase in gain caused by the decoupling circuit is only finite. This compensating effect has been analysed by several authors.[1,4,5]

The effect of the screen and cathode decoupling can be avoided either by removing these circuits or by making their time constants long compared with those of the coupling circuit and the anode decoupling circuit. In either case (i.e., letting $R_k = R_s = 0$ or letting η and $\theta \to 0$), equation (6) becomes

$$v_a = G\left[1 + \frac{\gamma}{p+\lambda}\right] 1 \qquad (11)$$

Alternatively, if the screen and cathode coupling capacitors are omitted but R_s and R_k are retained, η and θ become infinite and equation (6) again reduces to the form of equation (11), but with a modified gain given by

$$G' = \frac{G}{1 + g_k R_k + \sigma_s R_s}$$

From equations (7) and (11) the overall indicial response is

$$v_0(p)\,1 = G\frac{p^2 + p\,(\gamma+\lambda)}{p^2 + p\,(\alpha+\lambda) + \alpha\lambda}\,1 \qquad (12)$$

First-order compensation is obtained when the coefficients of p are equal in the numerator and denominator

i.e., $\quad \alpha = \gamma$

$$\therefore \quad R_g C_c = R_a C_d \qquad (13)$$

Other circuits in which an increase in anode impedance offsets the falling response of the coupling circuit at low frequencies have been described by Schlesinger.[6]

3.3. *Second-Order Compensation*

One method of obtaining second-order compensation which has been proposed by Thomson[1] involves adding further components to the coupling circuit to make it more nearly compensate for the effect of the anode decoupling circuit: the effects of the screen and cathode decoupling circuits are assumed to be negligible. It is possible, however, to obtain second-order compensation without adding any components to the circuit of Fig. 1 by making use of the screen and cathode decoupling circuits as well as the anode decoupling circuit.

From equation (8), first-order compensation is obtained when $a_1 = b_1$

$$\therefore \gamma = \alpha + \eta\, g_k R_k + \theta \sigma_s R_s \qquad (14)$$

For second-order compensation it is required in addition that $a_2 = b_2$

$$\therefore \gamma\,(\eta+\theta) = \alpha\,(\lambda+\eta+\theta) + \eta\, g_k R_k\,(\alpha+\lambda+\theta) + \theta\sigma_s R_s\,(\alpha+\lambda+\eta) \qquad (15)$$

Eliminating γ from equation (15) by means of equation (14) gives

$$\alpha = \frac{\eta\, g_k R_k\,(\eta-\lambda) + \theta\sigma_s R_s\,(\theta-\lambda)}{\lambda + \eta\, g_k R_k + \theta\sigma_s R_s} \qquad (16)$$

If we attempt to obtain third-order compensation we also require $a_3 = b_3$ and, from equation (8) this is obtained when

$$\gamma\eta\theta = \alpha\,(\lambda\eta + \eta\theta + \theta\lambda) + \eta\, g_k R_k\,(\lambda\theta + \theta\alpha + \alpha\lambda) + \theta\sigma_s R_s\,(\lambda\eta + \eta\alpha + \alpha\lambda)$$

Eliminating γ by means of equation (14) gives

$$\eta\theta\,(\eta\,g_k R_k + \theta\sigma_s R_s)$$
$$= \alpha\lambda\,[\eta\,(1 + g_k R_k) + \theta\,(1 + \sigma_s R_s)] + (\alpha + \lambda)\,\eta\theta\,(g_k R_k + \sigma_s R_s)$$

But, from equation (16),

$$\alpha + \lambda = \frac{\eta^2\,g_k R_k + \theta^2\sigma_s R_s - \alpha\lambda}{\eta\,g_k R_k + \theta\sigma_s R_s}$$

$$\therefore\ \alpha\lambda = -\frac{\eta\theta\,(\eta - \theta)^2\,g_k R_k\,\sigma_s R_s}{\eta^2 g_k R_k\,(1 + g_k R_k) + \theta^2\sigma_s R_s(1 + \sigma_s R_s) + 2\eta\,\theta g_k R_k\,\sigma_s R_s}$$

Now α, λ, η and θ are all positive, so it is impossible to obtain third-order compensation.

The highest order of compensation which can therefore be obtained for the indicial response of the circuit shown in Fig. 1 is the second order. This is obtained when α satisfies equation (16) and γ satisfies equation (14).

3.4. *Numerical Example*

Consider, as an example, a stage using an EF91 valve. Typical parameters obtained from the published characteristics of this valve are as follows:

grid–anode mutual conductance	7·6 mA/V
grid–screen mutual conductance	2·0 mA/V
screen–anode mutual conductance	0·1 mA/V
screen conductance	0·029 mA/V

(These figures relate to the following conditions: $V_a = V_s = 250$ V, $V_g = -2$ V, $I_a = 10$ mA, $I_s = 2{\cdot}5$ mA.)

The value of g_k is therefore 9·7 mA/V. If $R_k = 160\ \Omega$ and $R_s = 18$ kΩ, then $g_k R_k = 1{\cdot}5$ and $\sigma_s R_s = 0{\cdot}5$.

In order to compensate for the effect of anode decoupling by means of the cathode and screen decoupling circuits, equations (9) and (10) show that the anode, cathode and screen time constants should be equal and

$$R_d = 2R_a$$

The indicial response is then of zero order, it being a simple exponential curve with the time constant of the coupling circuit.

If the effects of the screen and cathode circuits can be made negligible and the anode decoupling circuit is used to compensate for the effect of the coupling circuit ($\alpha = \gamma$), the indicial response is given by equation (12). If we keep $R_d = 2R_a$ ($\gamma = 2\lambda$) equation (12) becomes

$$v_0(p)\,1 = \frac{p^2 + 1{\cdot}5\,p\gamma}{p^2 + 1{\cdot}5\,\gamma p + 0{\cdot}5\,\gamma^2}\,1$$

whence

$$v_0(t) = 1 - 0{\cdot}25\,(\gamma t)^2 + 0{\cdot}125\,(\gamma t)^3 - 0{\cdot}0365\,(\gamma t)^4 + 0{\cdot}0078\,(\gamma t)^5 - \ldots\ldots \quad (t > 0)$$

The response calculated from this series is shown as curve B in Fig. 2. This figure also shows as curve A the response due to the coupling circuit alone (when $\alpha = \gamma$), thus showing the improvement brought about by the first-order compensation.

If all the time constants are made finite and are adjusted to give second-order compensation the indicial response is given by equation (8). If we keep $\gamma = 2\lambda$ and put $\eta = 2\theta$ then, from equations (14) and (16),

$$\alpha = 0{\cdot}016\,\gamma, \qquad \lambda = 0{\cdot}5\,\gamma$$
$$\eta = 0{\cdot}56\,\gamma, \qquad \theta = 0{\cdot}28\,\gamma$$

Equation (8) then becomes

$$v_0(p)\,1 = \frac{p^4 + 2{\cdot}34\gamma p^2 + 1{\cdot}42\gamma^2 p^2 + 0{\cdot}235\gamma^3 p}{p^4 + 2{\cdot}34\gamma p^3 + 1{\cdot}42\gamma^2 p^2 + 0{\cdot}257\gamma^3 p + 0{\cdot}00376\gamma^4}\,1$$

whence

$$v_0(t) = 1 - 0{\cdot}00037\,(\gamma t)^3 + 0{\cdot}0020\,(\gamma t)^4 - 0{\cdot}00067\,(\gamma t)^5 + \ldots\ldots$$

The response calculated from this series is shown as curve D in Fig. 2. The use of second-order compensation thus gives a great improvement over the response for first-order compensation, assuming that the value of anode time constant ($1/\gamma$) is the same in each case. However, the value of coupling time constant ($1/\alpha$) required to obtain second-order compensation is 62 times that required for first-order compensation. Fig. 2 shows as curve C the response due to the coupling circuit alone. This response is still inferior to that of curve D and it can only be obtained if the decoupling circuits are omitted, or have much larger time constants than the coupling circuit, or are arranged to compensate for each other as described in Section 3.1. If the decoupling time

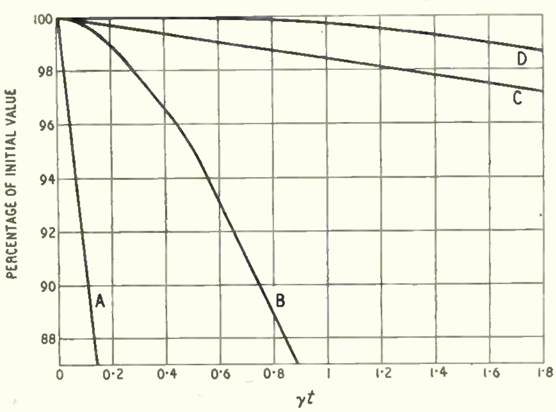

Fig. 2. *Indicial response of amplifier stage; A, due to coupling circuit alone ($\alpha = \gamma$); B, with 1st-order compensation; C, due to coupling circuit alone ($\alpha = 0{\cdot}016\gamma$); D, with 2nd-order compensation.*

constants can be chosen to compensate for each other they can be arranged to give second-order compensation as shown in curve D. The time taken by curve D to drop by 1% is over $2\frac{1}{2}$ times that taken by curve C.

4. Stages in Tandem

Let the indicial responses of two stages be

$$h_1(p)\,1 = \frac{p^n + a_1 p^{n-1} + a_2 p^{n-2} + \ldots + a_r p^{n-r} + \ldots + a_{n-1} p + a_n}{p^n + b_1 p^{n-1} + b_2 p^{n-2} + \ldots + b_r p^{n-r} + \ldots + b_{n-1} p + b_n}\,1$$

and

$$h_2(p)\,1 = \frac{p^m + a_1' p^{m-1} + a_2' p^{m-2} + \ldots + a_r' p^{m-r} + \ldots + a'_{m-1} p + a_m}{p^m + b_1' p^{m-1} + b_2' p^{m-2} + \ldots + b_r' p^{m-r} + \ldots + b'_{m-1} p + b_m}\,1$$

The overall response of the two stages in tandem is therefore

$$h(p)\,1 = h_1(p) . \; h_2(p)\,1$$

$$= \frac{p^{n+m} + c_1 p^{n+m-1} + \ldots + c_r p^{n+m-r} + \ldots + c_{n+m-1} p + c_{n+m}}{p^{n+m} + d_1 p^{n+m-1} + \ldots + d_r p^{n+m-r} + \ldots + d_{n+m-1} p + d_{n+m}}\,1$$

where

$$c_1 = a_1 + a_1', \qquad d_1 = b_1 + b_1'$$

$$c_2 = a_2 + a_1 a_1' + a_2', \qquad d_2 = b_2 + b_1 b_1' + b_2'$$

$$c_r = a_r + a_r' + \sum_{q=1}^{r-1} a_q a'_{r-q}, \qquad d_r = b_r + b_r' + \sum_{q=1}^{r-1} b_q b'_{r-q}$$

$$c_{n+m-1} = a_{n-1} a_m' + a_n a'_{m-1}, \qquad d_{n+m-1} = b_{n-1} b_m' + b_n b'_{m-1}$$

$$c_{n+m} = a_n a_m' \qquad d_{n+m} = b_n b_m'$$

If one stage has rth-order compensation and the other has qth order, then

$$a_1 = b_1,\ a_2 = b_2, \ldots a_r = b_r,\ a_{r+1} \neq b_{r+1}$$

and

$$a_1' = b_1',\ a_2' = b_2', \ldots a_q' = b_q',\ a'_{q+1} \neq b'_{q+1}$$

where

$$q \geqslant r \text{ and } a_{r+1} \neq b'_{r+1},\ b_{r+1} \neq a'_{r+1}$$

therefore

$$c_1 = d_1,\ c_2 = d_2, \ldots .\ c_r = d_r,\ c_{r+1} \neq d_{r+1}$$

The order of compensation of the complete amplifier is therefore the same as that of the stage with the lower-order compensation. It follows that if an amplifier has n stages all with rth-order compensation then the complete amplifier will also have rth-order compensation. If each stage can be given rth-order compensation, however, it is usually possible for the complete amplifier to have a higher order of compensation.

If two stages are connected in tandem and one can be made to have rth-order compensation and the other can be made to have qth-order compensation it should be possible to adjust the values of the coefficients of at least the first $(q + r)$ terms in the numerator and denominator of $h(p)$ so as to produce at least $(q + r)$th compensation. Thus it should be possible to design a multi-stage amplifier whose order of compensation is at least the sum of those which can be obtained from the individual stages. A very simple example is the tandem connection of a coupling circuit and an anode decoupling circuit. Each, considered separately, has a zero-order indicial response but when the two are connected in tandem the coefficients can be adjusted to give a first-order response as shown in Section 3.2. Some more complicated circuits are considered in Sections 4.1 to 4.4.

Because a multi-stage amplifier has a higher order of compensation than a simpler circuit it does not follow that its indicial response will approximate to the ideal more closely at all times. The multi-stage circuit with the high order of compensation will have an indicial response nearer to the ideal at times close to $t = 0$ but its response may subsequently fall more rapidly than that of the simple circuit so as to cross it and ultimately lie well below it. As an example, curve C in Fig. 2 relates to a zero-order response ($n = 1$, $r = 0$), but curve B relates to a first-order response ($n = 2$, $r = 1$). However, curve B crosses curve C quite close to $t = 0$ and, subsequently, departs from the ideal very much more rapidly.

4.1. *One Coupling Network and Two Stages with Anode Decoupling*

The operational form of the overall indicial response, from equations (7) and (11), is

$$h(p)\,1 = \frac{p}{p+\alpha}\left[1 + \frac{\gamma_1}{p+\gamma_1}\right] . \left[1 + \frac{\gamma_2}{p+\gamma_2}\right] 1$$

$$= \frac{p^3 + a_1 p^2 + a_2 p}{p^3 + b_1 p^2 + b_2 p + b_3}$$

where

$$a_1 = \gamma_1 + \gamma_2 + \lambda_1 + \lambda_2$$

$$a_2 = \gamma_1\gamma_2 + \gamma_1\lambda_2 + \gamma_2\lambda_1 + \lambda_1\lambda_2$$

$$b_1 = \alpha + \lambda_1 + \lambda_2$$

$$b_2 = \alpha\lambda_1 + \alpha\lambda_2 + \lambda_1\lambda_2$$

$$b_3 = \alpha\gamma_1\gamma_2$$

For second-order compensation we require $a_1 = b_1$ and $a_2 = b_2$

whence $\alpha = \gamma_1 + \gamma_2$ .. .. .. (17)

and $\gamma_1\gamma_2 = \gamma_1\lambda_1 + \gamma_2\lambda_2$ .. .. .. (18)

In particular, if $\gamma_1 = \gamma_2$ and $\lambda_1 = \lambda_2$ then equations (17) and (18) give

$$\alpha = 2\gamma_1 = 2\gamma_2 = 4\lambda_1 = 4\lambda_2 \qquad (19)$$

Thus, when the two anode circuits are identical, second-order compensation requires the anode decoupling resistance to be equal to twice the anode load resistance.

4.2. *Two Coupling Networks and Two Stages with Anode Decoupling*

The operational form of the overall indicial response, from equation (12) is

$$h(p)\,1 = \frac{p^2 + p(\gamma_1 + \lambda_1)}{p^2 + p(\alpha_1 + \lambda_1) + \alpha_1\lambda_1}\cdot\frac{p^2 + p(\gamma_2 + \lambda_2)}{p^2 + p(\alpha_2 + \lambda_2) + \alpha_2\lambda_2}\,1$$

$$= \frac{p^4 + a_1p^3 + a_2p^2}{p^4 + b_1p^3 + b_2p^2 + b_3p + b_4}\,1$$

where

$a_1 = \gamma_1 + \gamma_2 + \lambda_1 + \lambda_2,$ $\quad b_1 = \alpha_1 + \alpha_2 + \lambda_1 + \lambda_2$

$a_2 = \gamma_1\gamma_2 + \gamma_1\lambda_2 + \gamma_2\lambda_1 + \lambda_1\lambda_2,$ $\quad b_2 = \alpha_1\lambda_1 + \alpha_2\lambda_2 + (\alpha_1 + \lambda_1)(\alpha_2 + \lambda_2)$

$b_3 = \alpha_1\lambda_1(\alpha_2 + \lambda_2) + \alpha_2\lambda_2(\alpha_1 + \lambda_1),$ $\quad b_4 = \alpha_1\alpha_2\lambda_1\lambda_2$

Although the circuit considered in this section has more elements than that in Section 4.1, resulting in polynomials one degree higher for the numerator and denominator of $h(p)$, third-order compensation cannot be obtained because $a_3 = 0$.

For second-order compensation we require $a_1 = b_1$ and $a_2 = b_2$, which gives

$$\alpha_1 + \alpha_2 = \gamma_1 + \gamma_2$$

and $\quad \alpha_1\alpha_2 = \gamma_1\gamma_2 - \gamma_1\lambda_1 - \gamma_2\lambda_2$

$$\therefore \alpha_1, \alpha_2 = \tfrac{1}{2}\{(\gamma_1 + \gamma_2) \pm \sqrt{[(\gamma_1 - \gamma_2)^2 + 4(\gamma_1\lambda_1 + \gamma_2\lambda_2)]}\} \qquad (20)$$

Equation (20) cannot have equal roots, so second-order compensation cannot be obtained when the coupling circuits have equal time constants. The anode circuits can be identical however. In this particular case $\gamma_1 = \gamma_2 = \gamma$ and $\lambda_1 = \lambda_2 = \lambda$. Equation (20) then gives

$$\alpha_1, \alpha_2 = \gamma \pm \sqrt{2\gamma\lambda} \qquad (21)$$

Second-order compensation can therefore be obtained provided that $\gamma > 2\lambda$. The anode decoupling resistance can thus have any value greater than $2R_a$.

4.3. *m Coupling Networks and n Stages with Anode Decoupling*

The operational form of the indicial response from equations (7) and (11) is

$$h(p)\,1 = \prod_{r=1}^{m}\frac{p}{p + \alpha_r}\cdot\prod_{s=1}^{n}\frac{p + \gamma_s + \lambda_s}{p + \lambda_s}\,1 = \frac{p^{m+n} + a_1p^{m+n-1} + \ldots + a_np^m}{p^{m+n} + b_1p^{m+n-1} + \ldots + b_np^m + \ldots + b_{m+n}}\,1$$

This shows that the maximum order of compensation that can be achieved is n, the number of decoupling circuits. Coupling circuits do not increase the order of compensation, although they may assist in providing a sufficient number of variables to make a desired order of compensation realizable.

In order to obtain a response with a higher order than zero, there must be at least one coupling circuit. If there is not (i.e., if $m = 0$), then

$$a_1 = \sum_{s=1}^{n}(\gamma_s + \lambda_s), \qquad b_1 = \sum_{s=1}^{n}\lambda_s$$

The condition $a_1 = b_1$ gives $\sum_{s=1}^{n}\gamma_s = 0$, which is impossible because all γ are positive.

It can be shown that one coupling circuit is also sufficient to obtain nth-order compensation for at least $n = 1, 2, 3, 4, 5$. The cases $n = 1$ and $n = 2$ have been dealt with in Sections 2.2 and 4.1. A further example, $n = 3$, is given in Section 4.4.

4.4. *One Coupling Network and Three Stages with Anode Decoupling*

The operational form of the indicial response, from equations (7) and (11) is

$$h(p)\,1 = \frac{p}{p + \alpha}\cdot\frac{p + \gamma_1 + \lambda_1}{p + \lambda_1}\cdot\frac{p + \gamma_2 + \lambda_2}{p + \lambda_2}\times\frac{p + \gamma_3 + \lambda_3}{p + \lambda_3}$$

$$= \frac{p^4 + a_1p^3 + a_2p^2 + a_3p}{p^4 + b_1p^3 + b_2p^2 + b_3p + b_4}\,1$$

where

$a_1 = \gamma_1 + \gamma_2 + \gamma_3 + \lambda_1 + \lambda_2 + \lambda_3$

$a_2 = (\gamma_1 + \lambda_1)(\gamma_2 + \lambda_2) + (\gamma_2 + \lambda_2)(\gamma_3 + \lambda_3) + (\gamma_3 + \lambda_3)(\gamma_1 + \lambda_1)$

$a_3 = (\gamma_1 + \lambda_1)(\gamma_2 + \lambda_2)(\gamma_3 + \lambda_3)$

$b_1 = \alpha + \lambda_1 + \lambda_2 + \lambda_3$

$$b_2 = \lambda_1\lambda_2 + \lambda_2\lambda_3 + \lambda_3\lambda_1 + \alpha(\lambda_1 + \lambda_2 + \lambda_3)$$
$$b_3 = \lambda_1\lambda_2\lambda_3 + \alpha(\lambda_1\lambda_2 + \lambda_2\lambda_3 + \lambda_3\lambda_1)$$
$$b_4 = \alpha\lambda_1\lambda_2\lambda_3$$

For first-order compensation $a_1 = b_1$

$$\therefore \quad \alpha = \gamma_1 + \gamma_2 + \gamma_3 \quad \ldots \quad \ldots \quad \ldots \quad (22)$$

For second-order compensation, putting $a_2 = b_2$ and eliminating α by means of equation (22) gives

$$\gamma_1\gamma_2 + \gamma_2\gamma_3 + \gamma_3\gamma_1 = \gamma_1\lambda_1 + \gamma_2\lambda_2 + \gamma_3\lambda_3 \quad (23)$$

Third-order compensation requires in addition $a_3 = b_3$ which, after eliminating α by means of equation (22) gives

$$\gamma_1\lambda_1(\lambda_2 + \lambda_3) + \gamma_2\lambda_2(\lambda_1 + \lambda_3) + \gamma_3\lambda_3(\lambda_1 + \lambda_2)$$
$$= \lambda_1\gamma_2\gamma_3 + \lambda_2\gamma_1\gamma_3 + \lambda_3\gamma_1\gamma_2 + \gamma_1\gamma_2\gamma_3 \quad \ldots \quad (24)$$

Third-order compensation is, therefore, obtained when equations (23) and (24) are satisfied. It can be shown that these equations cannot be satisfied when $\lambda_1 = \lambda_2 = \lambda_3$ but they can be satisfied when $\gamma_1 = \gamma_2 = \gamma_3$. In the particular case when $\gamma_1 = \gamma_2 = \gamma_3 = \gamma$ and $\lambda_1 = \lambda_2 = \lambda$, equations (23) and (24) become

$$3\gamma = 2\lambda + \lambda_3$$

and $\quad \gamma^2 + \gamma(2\lambda + \lambda_3) = 2\lambda^2 + 4\lambda\lambda_3$

which give

$$\gamma_1 = \gamma_2 = \gamma_3 = \gamma = \lambda_3(2\sqrt{3} - 3) \quad \ldots \quad (25)$$
$$\lambda_1 = \lambda_2 = \lambda = \lambda_3(3\sqrt{3} - 5) \quad \ldots \quad (26)$$
$$\alpha = \lambda_3(6\sqrt{3} - 9) \quad \ldots \quad (27)$$

The values of the coefficients are given in the appendix.

The gain–frequency response will be said to have rth-order compensation when the first r derivatives of $|g(\omega)|$ with respect to ω vanish at infinite frequency. Similarly, the phase–frequency response will be said to have rth-order compensation when the first r derivatives of $\phi(\omega)$ vanish at infinite frequency. The gain–frequency response will be called maximally-flat when it has $(2n-1)$th-order compensation and the phase–frequency response will be called maximally flat when it has $(2n-2)$th-order compensation.

It is shown in the appendix that when the indicial response has compensation of order $2r$ or $(2r + 1)$ the gain–frequency response has compensation of order $(2r + 1)$. The maximally-flat indicial response, therefore, always corresponds to a gain–frequency response with a lower order of compensation than that required for maximal flatness.

When the indicial response has compensation of order $2r$ or $(2r - 1)$ the phase response has compensation of order $2r$. Thus only when $n = 2$ does the maximally-flat indicial response correspond to a maximally-flat phase response. For all higher values of n the maximally-flat indicial response corresponds to a phase response of a lower order of compensation than that required for maximal flatness. This is illustrated by the examples shown in Table 1.

TABLE 1

Order of max. flat indicial response	Order of corresponding gain-freq. response	Order of corresponding phase-freq. response	Order of max. flat gain-freq. response	Order of max. flat phase-freq. response
1	1	2	3	2
2	3	2	5	4
3	3	4	7	6
4	5	4	9	8

5. Comparison with Gain and Phase Responses

It is shown in Appendix 2 that the gain–frequency response corresponding to the indicial response given by equation (1) is

$$|g(\omega)| = \left[\frac{\omega^{2n} + c_2\omega^{2n-2} + \ldots + c_{2r}\omega^{2n-2r} + \ldots + c_{2n}}{\omega^{2n} + d_2\omega^{2n-2} + \ldots + d_{2r}\omega^{2n-2r} + \ldots + d_{2n}}\right]^{\frac{1}{2}} \quad \ldots \quad (28)$$

and the phase–frequency response is given by

$$\phi(\omega) = \tan^{-1} \left[\frac{c_1\omega^{2n-1} + c_3\omega^{2n-3} + \ldots + c_{2n-1}\omega}{\omega^{2n} + d_2\omega^{2n-2} + \ldots + d_{2r}\omega^{2n-2r} + \ldots + d_{2n}}\right] \quad \ldots \quad (29)$$

Consider, as an example, the indicial response given by

$$h(p)\,1 = \frac{p^3 + a_1p^2 + a_2p}{p^3 + b_1p^2 + b_2p + b_3}\,1$$

This has a maximally-flat indicial response, which is second-order, when $a_1 = b_1$ and $a_2 = b_2$.

The corresponding gain–frequency response, from equation (28), is given by

$$|g(\omega)|^2 = \frac{\omega^6 + c_2\omega^4 + c_4\omega^2}{\omega^6 + d_2\omega^4 + d_4\omega^2 + d_6}$$

where the coefficients, given by Appendix 2.1 are

$$c_2 = a_1^2 - 2a_2, \qquad d_2 = b_1^2 - 2b_2$$
$$c_4 = a_2^2, \qquad d_4 = b_2^2 - 2b_3b_1$$
$$d_6 = b_3^2$$

This has a maximally-flat gain-frequency response, which is fifth-order, when $c_2 = d_2$ and $c_4 = d_4$.

$\therefore \quad b_1 = \sqrt{[a_1^2 + 2(a_2 - b_2)]}$

and $\quad b_3 = [a_2^2 - b_2^2]/2\sqrt{[a_1^2 + 2(a_2 - b_2)]}$

The phase-frequency response, from equation (29), is given by

$$\tan \phi(\omega) = \frac{c_1\omega^5 + c_3\omega^3 + c_5\omega}{\omega^6 + d_2\omega^4 + d_4\omega^2}$$

where the coefficients, given by Appendix 2.2, are

$$c_1 = b_1 - a_1 \qquad d_2 = a_1b_1 - a_2 - b_2$$
$$c_3 = a_1b_2 - a_2b_1 - b_3 \qquad d_4 = a_2b_2 - a_1b_3$$
$$c_5 = b_3a_2$$

This has a maximally-flat phase-frequency response, which is fourth-order, when

$$c_1 = c_3 = 0$$
$$\therefore b_1 = a_1$$

and $\quad b_2 = a_2 + b_3/a_1$

Thus the condition for a maximally-flat indicial response corresponds neither to the maximally-flat gain-frequency response nor to the maximally-flat phase-frequency response. The latter, however, does give a first-order compensated indicial response whereas the maximally-flat gain-frequency response corresponds to an uncompensated (i.e., zero-order) indicial response.

6. Experimental Work

Fig. 3 shows an amplifier which was built in order to obtain some experimental verification of the theory. The values of the components determining the time-constants of the coupling circuits and decoupling circuits are accurate to $\pm$ 1%. The time-constants of the screen decoupling circuits are of the order of 100 times those of the coupling circuits and anode decoupling circuits, which were made much smaller than normal in order to avoid distortion being introduced by the oscilloscope amplifier. The effects of time-constants other than those of the two coupling circuits and two anode decoupling circuits are, therefore, negligible and the analysis given in Section 4.2 is applicable. The voltage divider before each stage reduces its gain to about unity in order to facilitate the taking of oscillograms.

Oscillograms were taken of the response of the amplifier to a 50-c/s square-wave input voltage. Because the period of this wave is long compared with the amplifier time-constants, the response to each edge of the waveform is substantially independent of its predecessors and approximates closely to the indicial response. A double-beam oscilloscope was used with the output voltage deflecting one beam and a 500-c/s sine-wave calibrating signal deflecting the other beam.

When each coupling capacitor is 0·0125 μF the time-constants of the coupling and decoupling circuits are equal. Each stage should then have first-order compensation of its indicial response, and so should the complete amplifier. Figs. 4(a) and (b) show the response of each stage separately and Fig. 4(c) shows the response of the two stages in tandem. The initial flatness of the response indicates that first-order compensation was obtained in each case.

When the coupling capacitances are 0·00688 μF and 0·0682 μF respectively, the time-constants have the values given by equation (21). The complete amplifier should then have second-order compensation. Figs. 4 (d) and (e) show the response of each stage and (f) shows the response of the two stages in tandem. The response in Fig. 4(f) maintains its initial flatness longer than that in Fig. 4(c), as is expected with second-order

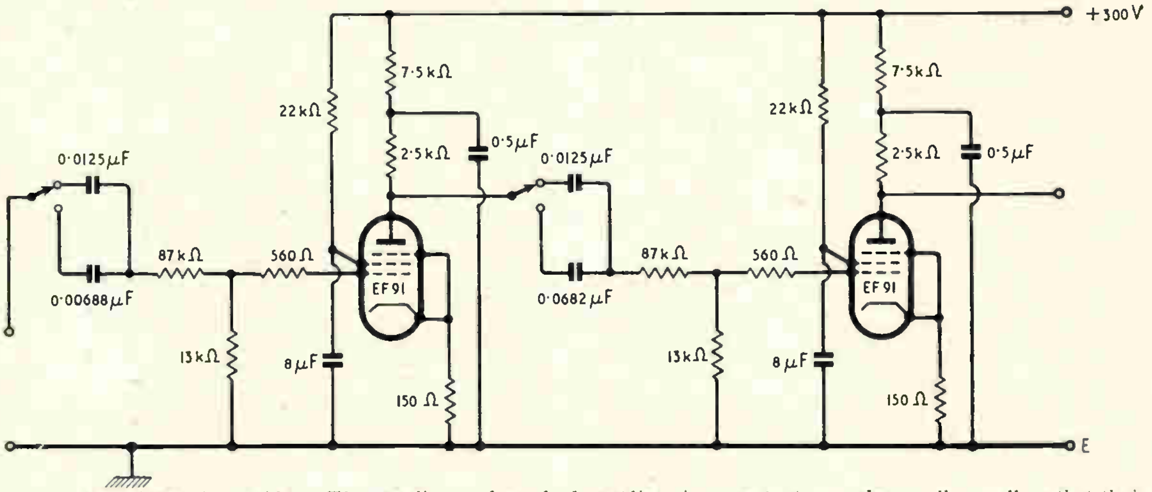

Fig. 3. *Experimental amplifier. The coupling and anode decoupling time constants are abnormally small so that their effect can be observed with an input of a convenient frequency.*

compensation. The total amount by which the top of the response sags in 2 ms in the second-order case is about one-third of that with first-order compensation.

The time-constants obtained in practice will depart from the calculated values because of the tolerances on components. Fig. 5 shows the effect on the overall response of a $\pm 10\%$ variation of the coupling capacitance of the first stage. The response of the amplifier with incorrectly-adjusted second-order compensation compares favourably with that obtained with correctly-adjusted first-order compensation.

A single resistance-capacitance coupled stage can have second-order compensation. As the number of stages in an amplifier is increased, the order of compensation which can be obtained increases, but so does the complexity of the calculations required to obtain the conditions for compensation. In many practical cases it is sufficient to obtain first- or second-order compensation. A complicated circuit can then be dealt with by considering it as a number of simpler stages connected in tandem, each of which can be made to have the required order of compensation.

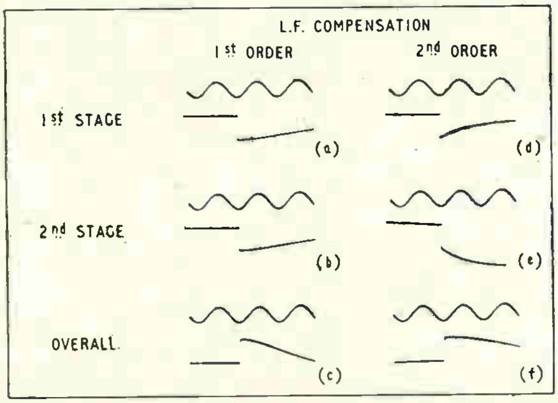

Fig. 4. *Oscillograms of response of model amplifier to a 50-c/s square wave. (The timing wave is 500 c/s.)*

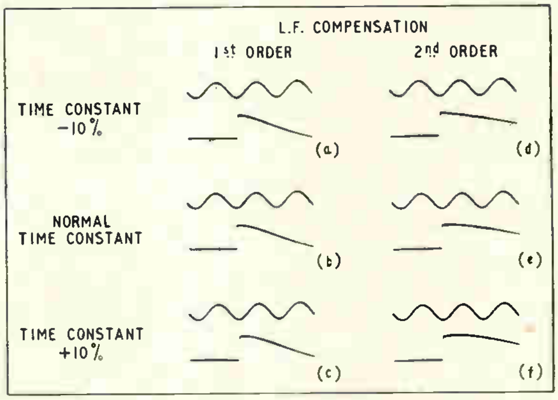

Fig. 5. *Effect of varying the time constant of one coupling circuit on the response of model amplifier to a 50-c/s square wave. (The timing wave is 500 c/s.)*

The gain–frequency response of the amplifier was measured with the component values corresponding to first- and second-order indicial responses. The results obtained are shown in Fig. 6. In each case the response is far from being maximally flat, thus confirming that a flat gain–frequency response is not the criterion for a flat indicial response. The frequency at which the second-order response has fallen 3 db is nearly half of that for the first-order case but the peak in the second-order curve is higher.

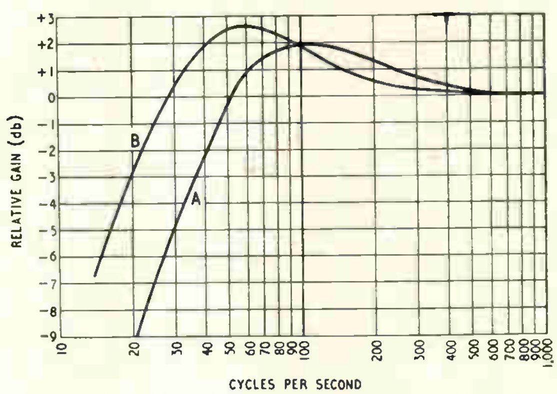

Fig. 6. *Gain frequency response of model amplifier; A, 1st-order compensation; B, 2nd-order compensation of indicial response.*

7. Conclusion

The indicial response of an amplifier can be made maximally flat by causing as many as possible of its differential coefficients to be zero at $t = 0$. A general method for doing this has been developed and applied to some typical circuits.

Some experimental verification has also been obtained. In order to obtain as high an order of compensation as possible, in many cases, it is necessary for the anode decoupling resistance to be several times the anode load resistance. This is quite practicable in wideband amplifiers without the use of an excessive supply voltage because the anode load resistance must be small in order to obtain adequate high-frequency performance.

APPENDIX 1

The substitution $x = 1/p$ in equation (1) gives

$$h(x) = \frac{1 + a_1 x + a_2 x^2 + \ldots + a_r x^r + \ldots + a_n x^n}{1 + b_1 x + b_2 x^2 + \ldots + b_r x^r + \ldots + b_n x^n}$$

$$= \frac{1 + A}{1 + B}$$

where $A = \sum_{r=1}^{n} a_r x^r$ and $B = \sum_{r=1}^{n} b_r x^r$

$$\therefore h(x) = (1 + A)[1 - B + B^2 - \ldots\ldots + (-1)^r B^r + \ldots\ldots] \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (30)$$

$$= 1 + c_1 x + c_2 x^2 + \ldots + c_r x^r + \ldots\ldots$$

Let $[1 - B + \ldots\ldots + (-1)^r B^r \ldots] = 1 + d_1 x + d_2 x^2 + \ldots + d_r x^r + \ldots$ then by the multinomial theorem

$$d_r = \sum \frac{(-1)^m m!}{\alpha!\, \beta! \ldots \epsilon!} b_\lambda{}^\alpha \,.\, b_\mu{}^\beta \ldots\ldots b_\rho{}^\epsilon \quad \ldots \quad (4)$$

where the summation extends over all possible values of $\alpha, \beta, \ldots \epsilon$ and $\lambda, \mu \ldots \rho$ and where

$m = \alpha + \beta + \ldots + \epsilon;\ \alpha\lambda + \beta\mu + \ldots + \epsilon\rho = r;$
$1 \leqslant \lambda < \mu < \ldots < \rho \leqslant r$

Then from equation (30)

$$c_r = d_r + \sum_{s=0}^{r-1} d_s a_{r-s}$$

It is proved below that

$$d_r = -\sum_{s=0}^{r-1} d_s b_{r-s} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (31)$$

it therefore follows that

$$c_r = \sum_{s=0}^{r-1} d_s a_{r-s} - \sum_{s=0}^{r-1} d_s b_{r-s}$$

$$= \sum_{s=0}^{r-1} d_s (a_{r-s} - b_{r-s}) \quad \ldots \quad \ldots \quad \ldots \quad (3)$$

where $a_0 = b_0 = d_0 = 1$

To prove equation (31), assume it to be true for a particular value of r and derive from it the value of d_{r+1}. In each factor $b_\lambda{}^\alpha \,.\, b_\mu{}^\beta \ldots b_\rho{}^\epsilon$ (where $\alpha\lambda + \beta\mu + \ldots + \epsilon\rho = r$ and $\alpha + \beta + \ldots + \epsilon = m$) one of the m factors b_q is replaced by b_{q+1}, so that a term like $d_s b_{r-s}$ is replaced by $d_s b_{r+1-s}$; in addition, each term in d_r occurs multiplied by b_1. Also, when like terms are collected, the numerical factor of $b_{\lambda'}{}^{\alpha'} \,.\, b_{\mu'}{}^{\beta'} \ldots b_{\rho'}{}^{\epsilon'}$ (where $\alpha' + \beta' + \ldots + \epsilon' = m + 1$ and $\alpha'\lambda' + \beta'\mu' + \ldots + \epsilon'\rho' = r + 1$) is

$$\frac{m!}{(\alpha' - 1)!\beta'! \ldots \epsilon'!} + \frac{m!}{\alpha'!(\beta' - 1)! \ldots \epsilon'!} + \ldots + \frac{m!}{\alpha'!\, \beta'! \ldots (\epsilon' - 1)!}$$

$$= \frac{m!}{\alpha'!\, \beta'! \ldots \epsilon'!}[\alpha' + \beta' + \ldots + \epsilon']$$

$$= \frac{m!(m + 1)}{\alpha'!\, \beta'! \ldots \epsilon'!} = \frac{(m + 1)!}{\alpha'!\, \beta'! \ldots \epsilon'!}$$

which is the correct multinomial coefficient, so that

$$d_{r+1} = -\left\{\sum_{s=0}^{r-1} d_s b_{r+1-s} + d_r b_1\right\} = -\sum_{s=0}^{r} d_s b_{r+1-s}$$

which is of the same form as d_r [given by equation (31)] with r replaced by $r + 1$ throughout. The values of d_r for $r = 1, 2, 3$ give the values of c_1, c_2 and c_3 shown in Section 2. These values are easily obtained from equation (1) by direct division, thus verifying equations (3) and (4) for $r = 1$, 2 and 3. But it has been shown that if equation (31) holds for r it holds for $r + 1$. Therefore, by induction, equations (3) and (4) are valid for all positive integral values of r.

APPENDIX 2

The steady-state complex gain, from equation (1) is

$$g(\omega) \underline{/\phi(\omega)}$$

$$= G \frac{a_0(j\omega)^n + a_1(j\omega)^{n-1} + \ldots + a_r(j\omega)^{n-r} + \ldots + a_n}{b_0(j\omega)^n + b_1(j\omega)^{n-1} + \ldots + b_r(j\omega)^{n-r} + \ldots + b_n}$$

where $a_0 = b_0 = 1$

$$= G \frac{A + jB}{C + jD} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (32)$$

If n is even ($n = 2m$) then

$$A = \sum_{r=0}^{m} a_{2r}(-\omega^2)^{m-r} \qquad B = \omega \sum_{r=1}^{m} a_{2r-1}(-\omega^2)^{m-r}$$

$$C = \sum_{r=0}^{m} b_{2r}(-\omega^2)^{m-r} \qquad D = \omega \sum_{r=1}^{m} b_{2r-1}(-\omega^2)^{m-r}$$

If n is odd ($n = 2m + 1$) then

$$A = \sum_{r=0}^{m} a_{2r+1}(-\omega^2)^{m-r} \qquad B = \omega \sum_{r=0}^{m} a_{2r}(-\omega^2)^{m-r}$$

$$C = \sum_{r=0}^{m} b_{2r+1}(-\omega^2)^{m-r} \qquad D = \omega \sum_{r=0}^{m} b_{2r}(-\omega^2)^{m-r}$$

2.1. *Gain–Frequency Response*

From equation (32)

$$|g(\omega)|^2 = \frac{A^2 + B^2}{C^2 + D^2}$$

$$= \frac{\omega^{2n} + c_2\, \omega^{2n-2} + \ldots + c_{2r}\omega^{2n-2r} + \ldots + c_{2n}}{\omega^{2n} + d_2\omega^{2n-2} + \ldots + d_{2r}\omega^{2n-2r} + \ldots + d_{2n}} \quad (28)$$

If n is even ($n = 2m$)

$$A^2 = \sum_{q=0}^{m} a_{2q}(-\omega^2)^{m-q} \,.\, \sum_{s=0}^{m} a_{2s}(-\omega^2)^{m-s}$$

The coefficient of $\omega^{2(n-r)}$ (where $r = q + s$) is

$$(-1)^r \sum_{q=0}^{r} a_{2q} a_{2(r-q)}$$

$$B^2 = \omega^2 \sum_{q=1}^{m} a_{2q-1}(-\omega^2)^{m-q} \,.\, \sum_{s=1}^{m} a_{2s-1}(-\omega^2)^{m-s}$$

The coefficient of $\omega^{2(n-r)}$ is

$$(-1)^{r+1} \sum_{q=1}^{r} a_{2q-1} a_{2(r-q)+1}$$

$$\therefore A^2 + B^2 = \omega^{2n} + \sum_{r=1}^{n} c_{2r}\omega^{2(n-r)}$$

where $c_{2r} = (-1)^r \sum_{q=0}^{r} (a_{2q} a_{2(r-q)} - a_{2q-1} a_{2(r-q)+1})$

$$= (-1)^r \sum_{q=0}^{2r} (-1)^q a_q a_{2r-q}$$

$$= 2 \sum_{s=0}^{r} (-1)^s a_{r+s} a_{r-s} \quad \ldots \quad \ldots \quad \ldots \quad (34)$$

Similarly, $C^2 + D^2 = \omega^{2n} + \sum_{r=1}^{n} d_{2r}\omega^{2(n-r)}$

where $$d_{2r} = 2\sum_{s=0}^{r} (-1)^s b_{r+s}b_{r-s} \quad \ldots \quad (35)$$

If n is odd ($n = 2m + 1$)

$$A^2 = \sum_{q=0}^{m} a_{2q+1}(-\omega^2)^{m-q} \cdot \sum_{s=0}^{m} a_{2s+1}(-\omega^2)^{m-s}$$

The coefficient of $\omega^{2(n-r)}$ is

$$(-1)^{r+1}\sum_{q=0}^{r-1} a_{2q+1}a_{2(r-q)-1}$$

$$B^2 = \omega^2 \sum_{q=0}^{m} a_{2q}(-\omega^2)^{m-q} \cdot \sum_{s=0}^{m} a_{2s}(-\omega^2)^{m-s}$$

The coefficient of $\omega^{2(n-r)}$ is

$$(-1)^r \sum_{q=0}^{r} a_{2q}a_{2(r-q)}$$

$$\therefore\ c_{2r} = (-1)^r \sum_{q=0}^{r} (a_{2q}a_{2(r-q)} - a_{2q-1}a_{2(r-q)+1})$$

$$= 2\sum_{s=0}^{r} (-1)^s a_{r+s}a_{r-s}$$

and similarly for d_{2q}.

The coefficients c_{2q} and d_{2q} therefore have the same values whether n is even or odd.

Putting $1/\omega = x$ in equation (28) and expanding it ascending powers of x gives

$$|g(\omega)|^2 = 1 + e_2x^2 + e_4x^4 + \ldots + e_{2r}x_2^r + \ldots$$

where $$e_{2r} = \sum_{s=0}^{r-1} g_s(c_{2(r-s)} - d_{2(r-s)}) \quad \ldots \quad (36)$$

and $$g_s = \sum (-1)^m \frac{m!}{\alpha!\ \beta! \ldots \epsilon!} d_{2\lambda}{}^{\alpha} d_{2\mu}{}^{\beta} \ldots d_{2\rho}{}^{\epsilon} \quad (37)$$

where the sum extends over all possible values of α, $\beta \ldots \epsilon$ and $\lambda, \mu \ldots \rho$ and

$$\alpha + \beta + \ldots + \epsilon = m\ ,\ \alpha\lambda + \beta\mu + \ldots + \epsilon\rho = s$$

Now by Maclaurin's theorem $e_{2r} = \frac{1}{(2r)!} f_{2r}(0)$

and $f_{2r+1}(0) = 0$

where $f_r(0)$ is the rth differential coefficient of $|g(\omega)^2|$ at $x = 0$ (i.e., $\omega = \infty$). The gain–frequency response therefore has $(2r + 1)$th-order compensation when $e_2 = e_4 \ldots e_{2r} = 0$. Equation (36) shows that this occurs when $c_1 = d_1,\ c_2 = d_2 \ldots c_{2r} = d_{2r}$.

If the indicial response has compensation of order $2r$ or $(2r + 1)$ then $a_q = b_q$ for $q = 1, 2 \ldots 2r$ or $q = 1, 2 \ldots (2_r + 1)$ respectively. It follows from equations (34) and (35) that $c_1 = d_1, c_2 = d_2, \ldots c_{2r} = d_{2r}$. Therefore, from equation (36), $e_2 = e_4 = \ldots = e_{2r} = 0$. Thus, when the indicial response has compensation of order $2r$ or $(2r + 1)$ the gain–frequency response has compensation of order $(2r + 1)$. The converse is not true, because the condition $e_2 = e_4 = \ldots = e_{2r}$ is insufficient to determine the coefficients $a_1 \ldots a_r$ and $b_1 \ldots b_r$ uniquely.

2.2. *Phase–Frequency Response*

From equation (32)

$$\tan\phi(\omega) = \frac{BC - AD}{AC + BD} \quad \ldots \quad (38)$$

$$= \frac{c_1\omega^{2n-1} + c_3\omega^{2n-3} + \ldots + c_{2r-1}\omega^{2(n-r)+1} + \ldots + c_{2n-1}\omega}{\omega^{2n} + d_2\omega^{2n-2} + \ldots + d_{2r}\omega^{2(n-r)} + \ldots + d_{2n}} \quad \ldots \quad (29)$$

If n is even ($n = 2m$)

$$BC = \omega \sum_{q=0}^{m} b_{2q}(-\omega^2)^{m-q} \cdot \sum_{s=1}^{m} a_{2s-1}(-\omega^2)^{m-s}$$

The coefficient of $\omega^{2(n-r)+1}$ (where $r = q + s$) is

$$(-1)^r \sum_{q=1}^{r} a_{2q-1}b_{2(r-q)}$$

$$AD = \omega \sum_{q=0}^{m} a_{2q}(-\omega^2)^{m-q} \cdot \sum_{s=1}^{m} b_{2s-1}(-\omega^2)^{m-s}$$

The coefficient of $\omega^{2(n-r)+1}$ is

$$(-1)^r \sum_{q=1}^{r} a_{2(r-q)}b_{2q-1}$$

$$\therefore\ BC - AD = \sum_{r=1}^{n} c_{2r-1}\omega^{2(n-r)+1}$$

where

$$c_{2r-1} = (-1)^r \sum_{q=1}^{r} (a_{2q-1}b_{2(r-q)} - a_{2(r-q)}b_{2q-1}) \quad \ldots \quad (39)$$

$$AC = \sum_{q=0}^{m} a_{2q}(-\omega^2)^{m-q} \cdot \sum_{s=0}^{m} b_{2s}(-\omega^2)^{m-s}$$

The coefficient of $\omega^{2(n-r)}$ in AC is

$$(-1)^r \sum_{q=0}^{r} a_{2q}b_{2(r-q)}$$

$$BD = \omega^2 \sum_{q=1}^{m} a_{2q-1}(-\omega^2)^{m-q} \cdot \sum_{s=1}^{m} b_{2s-1}(-\omega^2)^{m-s}$$

The coefficient of $\omega^{2(n-r)}$ in BD is

$$(-1)^{r+1} \sum_{q=1}^{r} a_{2q-1}b_{2(r-q)+1}$$

$$\therefore\ AC + BD = \sum_{r=0}^{n} d_{2r}\omega^{2(n-r)}$$

where $d_0 = 1$.

and

$$d_{2r} = (-1)^r \left\{ a_0b_{2r} + \sum_{q=1}^{r} (a_{2q}b_{2(r-q)} - a_{2q-1}b_{2(r-q)+1}) \right\} \quad \ldots \quad (40)$$

If n is odd ($n = 2m + 1$)

$$BC = \omega \sum_{q=0}^{m} a_{2q}(-\omega^2)^{m-q} \cdot \sum_{q=0}^{m} b_{2s+1}(-\omega^2)^{m-s}$$

The coefficient of $\omega^{2(n-r)+1}$ in BC is

$$(-1)^{r-1} \sum_{q=0}^{r-1} b_{2(r-q)-1}a_{2q}$$

$$AD = \omega \sum_{q=0}^{m} a_{2q+1}(-\omega^2)^{m-q} \cdot \sum_{s=0}^{m} b_{2s}(-\omega^2)^{m-s}$$

The coefficient of $\omega^{2(n-r)+1}$ in AD is

$$(-1)^{r-1}\sum_{q=0}^{r-1} a_{2(r-q)-1}b_{2q}$$

$$\therefore\ c_{2r-1} = (-1)^{r-1}\sum_{q=0}^{r-1}(a_{2q}b_{2(r-q)-1} - a_{2(r-q)-1}b_{2q})$$

$$= (-1)^{r}\sum_{q=1}^{r}(a_{2q-1}b_{2(r-q)} - a_{2(r-q)}b_{2q-1})$$

$$AC = \sum_{q=0}^{m} a_{2q+1}(-\omega^2)^{m-q}\sum_{s=0}^{m} b_{2s+1}(-\omega^2)^{m-s}$$

The coefficient of $\omega^{2(n-r)}$ in AD is

$$(-1)^{r-1}\sum_{q=0}^{r-1} a_{2q+1}b_{2(r-q)-1}$$

$$BD = \omega^2\sum_{q=0}^{m} a_{2q}(-\omega^2)^{m-q}\ .\ \sum_{s=0}^{m} b_{2s}(-\omega^2)^{m-s}$$

The coefficient of $\omega^{2(n-r)}$ in BD is

$$(-1)^{r}\sum_{q=0}^{r} a_{2q}b_{2(r-q)}$$

$$\therefore\ d_{2r} = (-1)^{r}\left\{a_0b_{2r} + \sum_{q=1}^{r}(a_{2q}b_{2(r-q)} - a_{2q-1}b_{2(r-q)+1}\right\}$$

The coefficients c_{2r-1} and d_{2r} therefore have the same values whether n is even or odd.

Putting $y = \omega^2$ in equation (29)

$$\tan\phi = \frac{c_1}{\omega}\cdot\frac{y^n + \frac{c_3}{c_1}y^{n-1} + \frac{c_5}{c_1}y^{n-2} + \ldots + \frac{c_{2r+1}}{c_1}y^{n-r} + \ldots + \frac{c_{2n-1}}{c_1}y}{y^n + d_2y^{n-1} + d_4y^{n-1} + \ldots + d_{2r}y^{n-r} + \ldots + d_{2n}}$$

Expanding this in ascending powers of $1/y$ gives

$$\tan\phi = \frac{c_1}{\omega}\left(1 + \frac{e_1}{y} + \frac{e_2}{y^2} + \ldots + \frac{e_r}{y^r} + \ldots\right)$$

$$= k_1x + k_3x^3 + \ldots + k_{2r+1}x^{2n+1} + \ldots$$

where $x = 1/\omega$

and $$k_{2r+1} = c_1e_r = \sum_{s=0}^{r-1} g_s(c_{2r+1-s} - c_1d_{2r-s}) \quad \ldots \quad (41)$$

and $$g_s = \sum(-1)^m\frac{m!}{\alpha!\,\beta!\ldots\epsilon!}d_{2\lambda}{}^{\alpha}d_{2\mu}{}^{\beta}\ldots d_{2\rho}{}^{\epsilon} \quad (42)$$

where the sum extends over all possible values of α, $\beta\ldots\epsilon$ and $\alpha\lambda + \beta\mu + \ldots + \epsilon\rho = s$

and $\alpha + \beta + \ldots + \epsilon = m$; $\alpha\lambda + \beta\mu + \ldots + \epsilon\rho = s$

Now by Maclaurin's theorem

$$k_{2r+1} = \frac{1}{(2r+1)!}f_{2r+1}(0)$$

and $f_{2r}(0) = 0$

where f_r is the rth differential coefficient of $\tan\phi$ at $x = 0(\omega = \infty)$

The phase frequency response therefore has $2r$th-order compensation when $k_1 = k_3 = \ldots = k_{2r-1} = 0$. Equation (41) shows that this occurs when $c_1 = c_3 = \ldots = c_{2r-1} = 0$.

If the indicial response has compensation of order $(2r-1)$ or $2r$ then $a_q = b_q$ for $q = 1, 2\ldots(2r-1)$ or $q = 1, 2, \ldots 2r$ respectively. It follows from equation (39) that $c_1 = c_3 = \ldots = c_{2r-1} = 0$. Thus, when the indicial response has compensation of order $(2r-1)$ or $2r$ the phase–frequency response has compensation of order $(2r-1)$. The converse is not true since the condition $c_1 = c_3 = \ldots c_{2r-1} = 0$ is insufficient to determine the coefficients $a_1\ldots a_r$ and $b_1\ldots b_r$ uniquely.

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AUTOMATIC TUNING FOR PRIMARY RADAR

By S. Ratcliffe, B.Sc.

(*Concluded from p. 165, June issue*)

Choice of A.F.C. System

In drawing up the requirements for the a.f.c. system, it is important to avoid placing an unnecessary load on the a.f.c. by calling on it to compensate for a defect elsewhere in the equipment which may be more readily dealt with by attention to the design of the component in question. For example, a thermostat in series with the blower motor may be used to reduce temperature drifts in the magnetron. The a.f.c. designer should, therefore, avoid being tempted to display his ingenuity in solving problems of this type, since the increased complexity of the a.f.c. usually brings with it an increased risk of component failures. Alternatively, it is necessary to avoid lowering the reliability of the a.f.c. by attempting to perform a complex task with too few valves and too many preset adjustments.

It is necessary, in particular, to decide whether or not to make any serious attempt to deal with scanner pulling. In general, the wider the pulse and the higher the transmitter power, the more likely it is that the improvement in performance will justify the increased complexity. It is then also possible to decide whether to use a narrow (i.e., electronic) or wide-band a.f.c. system. The objections to a narrow-range system have been given earlier, and in an airborne equipment the arguments in favour of a wide-range system are very strong indeed. On a ground or shipborne equipment where a mechanic is in attendance, there is more difference of opinion. The argument may be summarized by saying that a human being usually shows more signs of intelligence but often also of fatigue.

The choice of local oscillator will be influenced by many factors which have little connection with a.f.c., but care should be taken to avoid the trap of seeking an oscillator with the widest possible tuning range, regardless of its other demerits. The most important thing, from the point of view of the a.f.c. design, is the difference between the available electronic-tuning range and the possible drift in local-oscillator frequency due to changes in temperature and pressure. If a wide-range reflector-cum-mechanical a.f.c. is to be used, it is advisable to avoid flexible mechanical couplings between the tuning mechanism and the control motor and gear train. This is best achieved by making the resonant cavity an integral part of the radar plumbing, with a plug-in klystron. Valves designed for use in this way are now becoming available.

The typical circuits given previously, together with the theoretical analysis given in the Appendix, should assist in the choice between the 'instantaneous' and the 'extrapolating' types of reflector a.f.c. If a pure reflector a.f.c. is being designed, it may be necessary to provide a search valve, either to get an adequate pull-in range, or to enable the circuit to find a suitable reflector voltage to obtain oscillations. Two main types of search circuit have been described, and the ingenious reader may readily devise many variants on either of these themes. The essential point is to arrive at a designable circuit in which the bare minimum of preset controls is required.

It must again be pointed out that it is essential to take every possible step to keep down tolerances on the gain round the reflector a.f.c. loop. Even if the transient response of the system is unimportant, the problems set by the harmonic output of the mixer will usually necessitate a close control of the gain, up to the search-stopping mechanism at least. This point must be clearly understood. It is completely erroneous to argue from normal negative-feedback principles, but it is a common misconception that gain is not critical above a useful level.

Performance Tests on an A.F.C. Loop

Assuming that a design has been completed and an experimental model is available, it is necessary to check that the performance is satisfactory.

Consider first the problem of determining the locking accuracy. A very convenient means of checking this involves the use of a spectrum analyser of the type consisting of a narrow-band superheterodyne receiver having a local oscillator which is frequency modulated by means of a low-frequency sawtooth, and a c.r.t. display on which the receiver response is shown as a function of frequency. The intermediate frequency of this receiver is made exactly half the i.f. of the a.f.c. under test, and the spectrum analyser is supplied with a signal input both from the magnetron of the radar, and from the local oscillator to which the a.f.c. is being applied. Such a spectrum analyser will, at any given point in its sweep, respond to signals on two frequencies one above and the other below that of the analyser oscillator, and the difference of the two frequencies will be twice the i.f. of the spectrum analyser; i.e., equal to the radar receiver i.f. If, then, the a.f.c. under test is holding a local oscillator at the correct frequency separation from a pulse source, it is possible, with an analyser such as that described, to display the spectra of the pulse and of the local-oscillator output superimposed on each other. Fig. 24 should help to make this clear. Any failure of the a.f.c. to maintain the desired radar i.f. will result in a deviation of the observed position of the local oscillator 'pip' from the centre of the pulse spectrum.

With the aid of such a device it is possible to check the locking accuracy of the a.f.c. under various conditions. Fig. 25 shows a typical picture obtained when an a.f.c. was locking correctly to a one-microsecond pulse, although some undesired ripple on the power supplies to the klystron was causing a small amount of frequency modulation of the local oscillator.

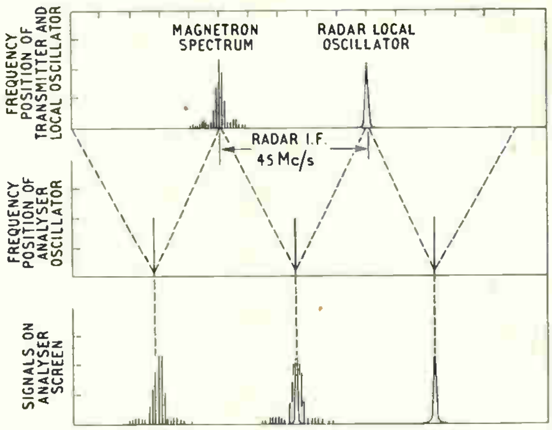

Fig. 24. *Action of analyser in viewing signals.*

It will be appreciated that the i.f. of the spectrum analyser is being adopted as the standard of frequency, for the purpose of this test. It is, therefore, necessary to ensure high stability of this frequency. A double superheterodyne receiver was used in the spectrum analyser from which Fig. 25 was obtained. The first i.f. was 22·5 Mc/s and the final i.f. 2·5 Mc/s, a crystal-controlled oscillator being used to determine the remaining 20 Mc/s of the first i.f. so that the stability of the 22·5 Mc/s response could be relied on to 10 kc/s or so. Present experience suggests that such a device in skilled hands is a very valuable tool, but since it is as likely to go wrong as is the a.f.c. it is testing, it is emphatically not suitable for routine maintenance work.

The ideal and universally-applicable method of testing the transient response of an a.f.c. is to produce a step-function change in the transmitter frequency and to observe the rate at which the reflector voltage or tuning shaft responds. It is not easy to produce such a change in the frequency of a transmitter, however, and this test is only practicable if the transmitter sample to the a.f.c. mixer is replaced by the output of a suitable pulsed-signal generator. For example, Fig. 25 is

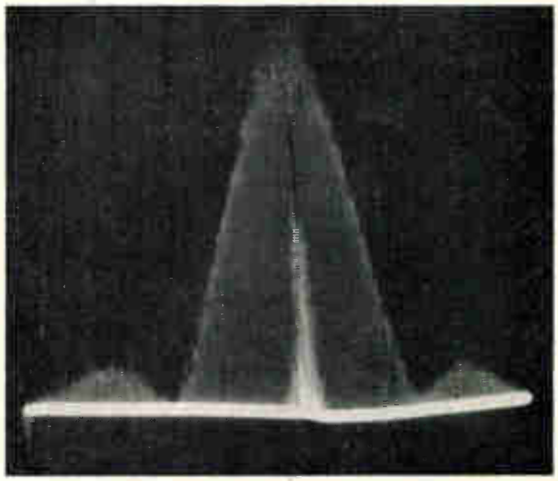
Fig. 25. *Typical oscillogram of the pulse spectrum with an a.f.c. system locking correctly on a 1-μsec pulse.*

a photograph of part of the spectrum of a one-microsecond X-band pulse produced by pulse-modulating the grid of a CV129 klystron. It is easy to apply a suitable square-wave to the reflector of this klystron and thereby frequency modulate the output pulse.

Despite the usefulness of such pulse signal generators as laboratory tools, this method of testing the a.f.c. performance is somewhat artificial, since it is difficult to ensure that the signal level at the a.f.c. mixer is that which obtains during normal operation. It is, therefore, desirable to have some way of testing the transient response of an a.f.c. system using as signal source the magnetron with which the a.f.c. is normally required to work.

One such method, which can often be applied, is to introduce a step-function disturbance in the reflector potential of the local oscillator. It is convenient to make this disturbance a periodic one, so that an oscilloscope can be used to observe the response. For example, if a resistance of a few thousand ohms is connected in series with the lead to the reflector, a high-speed relay may be arranged periodically to connect a 1½-V battery, say, across this resistance. An oscilloscope can then be connected to display the reflector waveform, from which the transient response of the a.f.c. can readily be deduced. A useful refinement is to brighten up the c.r.t. picture momentarily each time the transmitter fires, and to synchronize the switching frequency to some sub-multiple of the p.r.f.

Conclusion

A considerable amount of research has been devoted to improving the noise factor of centimetre-wave receivers. Much of the available improvement in performance can be lost by an unfortunate choice of receiver bandwidth and magnetron-pulling figure or by the use of an unsuitable a.f.c. system. This article has attempted to show the extent of the contribution which a good a.f.c. can make towards improving the overall performance of an equipment; to give enough circuit details to indicate the price that must be paid in increased numbers of valves and components; and to show that, despite the increased complexity, a good a.f.c. may simplify the maintenance problem by eliminating a number of preset adjustments, and thereby give both a better and more reliable performance.

It may reasonably be objected that it has been assumed that the equipment designer has complete control over the factors, such as the receiver bandwidth and magnetron-pulling figure, whereas in practice he is often compelled to use a particular i.f. amplifier or magnetron because of its availability. It is felt, however, that many systems are designed to use some particular components, not so much because better ones are not available, or cannot be developed in time, but because the designer is not aware of the extent of the loss in performance which he is tolerating for the sake of the convenience.

Experience has shown that by adopting the 'designable' a.f.c. circuit a better system can be obtained with less development work.

It is of some interest to compare the present-day problems of scanner pulling, frequency splitting, etc., with the problems of transmitter design in the early days of wireless telegraphy, when self-oscillatory output stages were in use. The latter problems included frequency shifts due to aerials swaying in the wind and varying in capacitance, and the 'Ziehen effect',[19] which showed itself as a spasmodic jump in transmitter frequency.

Automatic frequency control was invented to deal with these transmitter difficulties, but the solution finally adopted was to use a power amplifier at the output stage and to drive it from a separate master oscillator. It may be that eventually the same solution will be adopted in cm-wave radar systems, but much more progress in the technique of power amplification at these frequencies will be needed before this step becomes economical.

In the meantime, existing techniques are capable of dealing with the problem, at a price, and the main field in which improvement is to be expected is in the engineering details of the wide-range a.f.c. systems which are superseding the earlier systems.

Acknowledgments

The author is indebted to the Chief Scientist, Ministry of Supply, for permission to publish the results of this work, most of which was conducted under his auspices. The material has been drawn from the unpublished results of a large number of workers, but especial thanks are due to Mr. C. Baron of R.R.E., Malvern, for the circuit of Fig. 20, and for part of the discussion given in the Appendix.

APPENDIX

Transient Response of an A.F.C. Loop

The behaviour of any servomechanism can be summed up in an 'Equation of Motion'. For a continuous feedback servo, this relates the position of the output shaft (or its equivalent) and, usually, derivatives of this quantity with respect to time, to the position of the input shaft (or equivalent), and possibly time derivatives of this quantity also.

The resulting differential equation can be solved, for a linear system, by straightforward, but possibly laborious, methods to give the response of the system to any applied disturbance, or to establish the stability conditions. For a linear sampling servo the same situation arises, except that the infinitesimal calculus must be replaced by the less familiar calculus of finite differences. Thus the equation of motion takes the form:

$$q_1 u_n + q_2 u_{n-1} + \dots\dots + q_n u_1 = f(n) \quad \dots \quad \dots \quad (1)$$

where $u_1, u_2 \dots\dots u_n$ denote the values of the output quantity at the successive instants $t_1, t_2 \dots\dots t_n$ when samples are taken, and $f(n)$ is the applied disturbance.

It can be shown by substitution that equation (1) is satisfied by:

$$u_n = Aa^n + Bb^n + \dots\dots Nn^n + F(n) \quad \dots \quad (2)$$

where $a, b, \dots . n$, are the roots of the characteristic equation:

$$q_1 x^n + q_2 x^{n-1} + \dots\dots + q_n = 0 \quad \dots \quad \dots \quad (3)$$

and $A, B, \dots . N$, are determined by the initial conditions and $F(n)$ gives the forced response to the input $f(n)$.

It follows from equation (2) that the stability condition for the sampling servo is that the modulus of each of the roots of equation (3), real or complex, should be less than unity. Just as the Laplace transform is commonly used by servo designers to simplify the solution of ordinary linear differential equations, integral transform techniques[21,22] are available for the solution of finite difference equations. For the reader who does not wish to master these techniques, we will present a simpler though more laborious treatment of the only two equations of serious interest to the a.f.c. designer, those relating to systems having respectively one and two significant time-lags.

Single Time-Lag

Consider the local oscillator and transmitter frequencies at the equally-spaced instants of time $t_0, t_1, \dots . t_n$, etc., when the transmitter fires, and let the local-oscillator frequency at these instants be $f_0, f_1 \ldots f_n$, while the oscillator frequency required for correct tuning at these instants is $g_0, g_1 \ldots g_n$.

Then the tuning error e_n at time t_n is $(f_n - g_n)$, and the characteristic equation of the system is

$$f_n = f_{n-1} - Ae_{n-1} \quad \ldots \quad (4)$$

where A is a constant determined by the circuit design.

Consider the response of the system to a step-function. Put $e_0 = E$ and $g_0 = g_1 = \ldots = g_n$. Then equation (4) becomes

$$e_n = E\,(1 - A)^n \quad \ldots \quad (5)$$

Thus the system is stable if $0 < A < 2$, and the response will be monotonic, 'dead-beat' or oscillatory, according as A is less than, equal to, or greater than unity.

Two Time-Lags

Using the above notation, suppose the output of the discriminator to be zero except for the time interval $t_0 < t < t_1$, when it then has a value which we may, without loss of generality, assume to be a constant e_0. Then during the interval $t_0 < t < t_1$:

$$\frac{df}{dt} = -Ke_0\left[1 - \exp - \left(\frac{t - t_0}{T}\right)\right]$$

where T is the speed build-up time-constant of the system.

$$f_0 - f_1 = Ke_0 \int_{t_0}^{t_1}\left[1 - \exp - \left(\frac{t - t_0}{T}\right)\right] dt$$

Putting $\dfrac{t_1 - t_0}{T} = r$,

$$f_0 - f_1 = e_0\,KT\,[r + \exp(-r) - 1] \quad \ldots \quad (6)$$

At any time $t_1 < t$

$$\frac{df}{dt} = -Ke_0\,[1 - \exp(-r)]\exp - \left(\frac{t - t_1}{T}\right)$$

whence $\quad f_1 - f_2 = Ke_0 \displaystyle\int_{t_1}^{t_2} [1 - \exp(-r)] \exp - \left(\frac{t - t_1}{T}\right) dt$

Therefore $\quad f_1 - f_2 = e_0\,KT\,[1 - \exp(-r)]^2$

and generally, for $n \geqslant 2$

$$f_{n-1} - f_n = e_0\,KT\,[1 - \exp(-r)]^2 \exp[-(n-2)\,r] \quad \ldots \quad (7)$$

Now if the output from the discriminator consists not of one signal e_0 but a sequence $\{e_n\}$, then the response of the oscillator frequency will be given by the sum of its responses to a series of inputs of the form just discussed. Thus equations (6) and (7), when applied successively to each term of $\{e_n\}$, give:

$$\begin{aligned} f_{n-1} - f_n = {} & e_{n-1}\,KT\,[r + \exp(-r) - 1] \\ & + e_{n-2}\,KT\,[1 - \exp(-r)]^2 \\ & + KT\,[1 - \exp(-r)]^2\,[\exp(-r)\,e_{n-3} \\ & + \exp(-2r)\,e_{n-4} + \text{etc.}] \end{aligned}$$

With the aid of a similar expression for $(f_{n-2} - f_{n-1})$, the infinite series can be eliminated, giving:

$$\begin{aligned} f_n - f_{n-1}\,[1 + \exp(-r)] + f_{n-2}\exp(-r) \\ = -e_{n-1}\,KT\,[r + \exp(-r) - 1] \\ - e_{n-2}\,KT\,[1 - \exp(-r)]^2 \\ + e_{n-2}\,KT\,[r + \exp(-r) - 1]\exp(-r) \end{aligned} \quad \ldots \quad (8)$$

To find the response of the system to a step function, as before, put $g_0 = g_1 = \ldots = g_n = G$, say. Then equation (8) leads to:

$$\begin{aligned} e_n - \{1 + \exp(-r) - KT\,[r - 1 + \exp(-r)]\}\,e_{n-1} \\ + \{\exp(-r) + KT\,[1 - (1 + r)\exp(-r)]\}\,e_{n-2} \\ = 0 \end{aligned} \quad \ldots \quad (9)$$

It is then possible to compute the response to a step E by using:

$$e_1 = E\,\{1 - KT\,[r + \exp(-r) - 1]\}$$

$$e_2 = e_1 - KT\,[1 - \exp(-r)]^2 E$$

and equation (9) for the later terms.

Alternatively, writing equation (9) as:

$$e_n + Ae_{n-1} + Be_{n-2} = 0 \quad \ldots \quad (10)$$

it is possible to use equation (2) for terms later than the second. Since, in deriving equation (9), it was assumed that the transmitter frequency had remained constant for at least two pulses, this equation is not valid for e_1 and e_2. The stability conditions for the system are that the roots of the characteristic equation should have a modulus less than unity. These are satisfied if:

$$f(1) = 1 + A + B > 0$$

$$f(-1) = 1 - A + B > 0$$

$$f(0) = B < 0$$

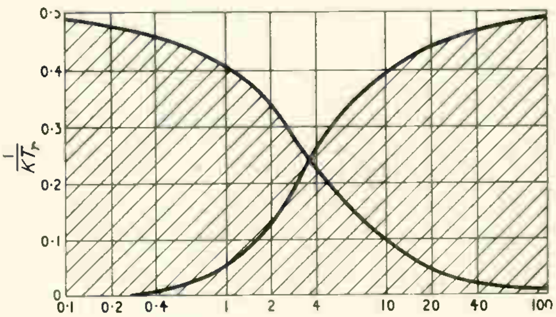

Fig. 26. *Stability conditions for system with two time-lags.*

These conditions have been used to plot the curves of Fig. 26. The shaded area shows the unstable region.

The interested reader should compare this analysis with a different approach due to Hurewicz.[22]

Effect of Component Tolerances on the Performance of an Extrapolating A.F.C. Loop

Consider the single time-lag system discussed above. Equation (5) gives the response to a step-function change in transmitter frequency as

$$e_n = E\,(1 - A)^n$$

where e_n denotes the tuning error n pulses after the application of the step of height E, and A denotes the ratio of the amount the oscillator frequency moves between one pulse and the next, to the tuning error during the corresponding period. The best performance will be attained when $A = 1$, but some deviation from this value must obviously occur in practice, and it was shown previously that A may well vary over a range of 10 : 1. It is now proposed to derive an expression showing the effect on performance of given tolerances on A.

As a measure of the efficiency of the system, consider the number of pulses necessary to reduce e_n/E to some amount $1/K$.

From equation (5)

$$1/K = (1 - A)^n$$

$$\text{or} \quad n = \frac{\log 1/K}{\log(1-A)} = \frac{-\log K}{\log(1-A)} \quad \cdot\cdot \quad (11)$$

It is necessary to choose a, the geometric-mean value of A. This will be chosen to keep the maximum value of n as low as possible, by making the performance fall off equally at both limits of A.

Let these limits be ar, a/r where $r < 1$.

Since we are only concerned with the modulus of the error, the desired condition is given by putting

$$|1 - ar| = \left|1 - \frac{a}{r}\right|$$

$$\text{or } a = \frac{2r}{r^2 + 1} \quad \cdot\cdot \quad \cdot\cdot \quad \cdot\cdot \quad (12)$$

Substituting ar for A in equation (11):

$$n = \frac{-\log K}{\log \dfrac{1 - r^2}{1 + r^2}}$$

Since $r^2 < 1$

$$= \frac{\log_e K}{-2\left(r^2 - \dfrac{r^6}{3} - \dfrac{r^{10}}{5}\right)}$$

If $r^2 < 0{\cdot}5$, we may neglect the terms in r^6, r^{10}, etc., without introducing a serious error. Thus we may write

$$n = \frac{\log_e S}{2r^2}$$

$$\text{or, writing} \quad R = \frac{1}{r^2},\ n = \frac{R \log_e K}{2} \quad \cdot\cdot \quad \cdot\cdot \quad (13)$$

where R is the ratio of the greatest possible value of A to the smallest. In other words, the system has a time-constant equal to, or less than, $\frac{1}{2}RT$, where:

$$T = (t_n - t_{n-1}) = \frac{1}{\text{p.r.f.}}$$

It is clear from equation (13) that a good transient response can only be guaranteed if R can be kept low.

Similar results hold for systems having more than one time-lag, since, although it is possible to devise circuits which to some extent compensate for variations in A when following a smoothly-varying transmitter frequency, it has already been shown that this problem is not that with which we are concerned.

'Instantaneous' A.F.C. System

This system, discussed previously, attempts to avoid the fundamental limitation discussed above by arranging that the a.f.c. behaves, during the pulse, as a normal negative-feedback system.

Baron's circuit, shown in Fig. 20 has a peak separation sufficiently wide to make it possible to neglect the phase shift in the i.f. amplifier. It may be assumed that there are, in all, three significant time-lags: in the diode loads, in the 'clamp' circuit, and in the anode load of the subsequent d.c. amplifier. It may be shown that the most economical arrangement makes the time-constant of two of these circuits equal to each other, while the third must be much larger in order to ensure stability. Let the long and short time-constants be t_1 and t_2 respectively, and let the gain round the feedback loop be μ. Then if F_t is the frequency error at time t, and if F' is the frequency change produced at the local oscillator, the 'equation of motion' is given by $(1 + t_1 D)(1 + t_2 D)^2 F' = \mu F_t$.

Since $t_1 \gg t_2$, the auxiliary equation may be simplified approximately to:

$$p^3 + \frac{2}{t_2}p^2 + \frac{1}{t_2{}^2}p + \frac{\mu}{t_1 t_2{}^2} = 0$$

Putting $p = 1/t_2\,P$ (i.e., changing the time scale so that $t = t_2 T$) we obtain:

$$P^3 + 2P^2 + P + \mu\, t_2/t_1 = 0 \quad \cdot\cdot \quad \cdot\cdot \quad \cdot\cdot \quad (14)$$

This equation may be solved graphically by plotting $f(P) = p^3 + 2p^2 + p$ in the range $0 > P > -2$ (this being the range within which the system is stable). The solution for any value of $\mu t_2/t_1$ is then obtained direct from the curve, since $\mu t_2/t_1 = -f(P)$. The system is critically damped for $\mu t_2/t_1 = 0{\cdot}14$, and the response to a unit step function for this value and other values of $\mu t_2/t_1$ are shown in Fig. 27, which is due to C. Baron, as is an analysis from which the above treatment is derived.

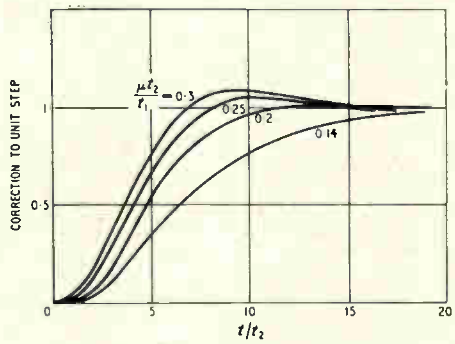

Fig. 27. *Response of 'Baron' a.f.c.*

Effect of Component Tolerances on the Performance of an 'Instantaneous' A.F.C. Loop

Since the main objection to 'extrapolating' a.f.c. systems is the considerable departures from the ideal performance which are occasioned by the tolerances on the gain round the feedback loop, it is of some interest to consider the effect of these tolerances on the system discussed above.

We first require an approximation to the response of the system for values of $\mu t_2/t_1$ appreciably higher or lower than the optimum value. For low values the response consists of three exponential terms, and it will be assumed here that only the slowest of these is important. Thus it will be assumed that the response is determined by the smallest root of equation (14). For values of $\mu t_2/t_1$ greater than $0{\cdot}14$, the equation (14) has only one real root and the response is oscillatory. It is proposed to consider the response as determined by the envelope of this oscillation; i.e., by the real part of the complex roots.

Equation (14) can readily be solved graphically, as indicated above. The value of the real part of the complex roots, when these occur, can be found when the real root, $-r$ say, is known, by using the relation that the sum of the three roots is minus one times the coefficient of x^2 in equation (14). It follows that the real part of the complex roots is given by:

$$a = \frac{r - 2}{2} \quad \cdot\cdot \quad \cdot\cdot \quad \cdot\cdot \quad \cdot\cdot \quad \cdot\cdot \quad \cdot\cdot \quad \cdot\cdot \quad (15)$$

Fig. 28 then gives a value of p giving a suitable approximation to the response of the system to a step function of unit height in the form

$$C_t = 1 - \exp\left(-p\frac{t}{t_2}\right) \quad \cdot\cdot \quad \cdot\cdot \quad \cdot\cdot \quad \cdot\cdot \quad (16)$$

where C_t is the error at time t after the application of the step. It will be remembered that, for values of $\mu t_2/t_1$ corresponding to the upper part of the curve, equation (16) gives the envelope only of the oscillation.

Suppose now that the tolerances on the gain round the loop can be estimated, and that the ratio of the maximum possible gain round the loop to the minimum is R say, where a probable value for R is about 10. It is now necessary to choose the mean value for $\mu t_2/t_1$ which gives the best performance over the range R. Since we are only concerned with the modulus of the error, it will be

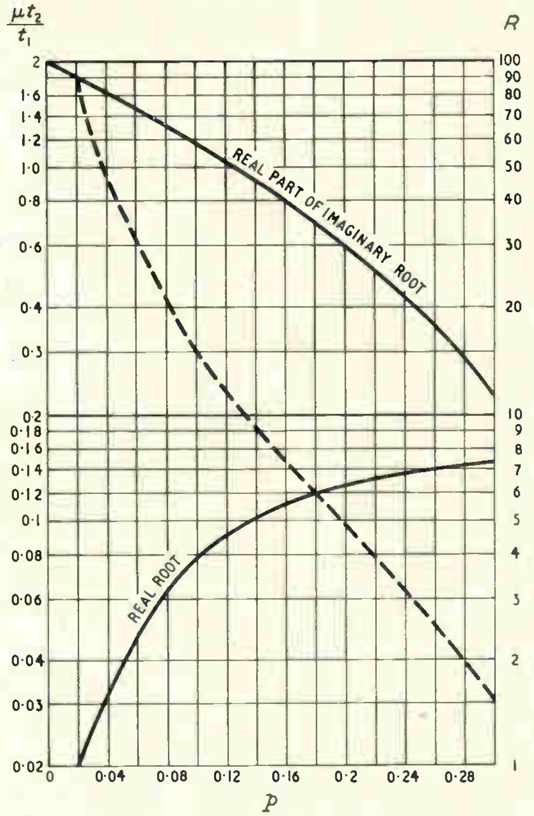

Fig. 28. *Smallest real root or real part of complex root of* $p^3 + 2p^2 + p + \mu t_2/t_1 = 0$.

sufficient to choose the mean value such that the two extreme tolerances give the same value of p in Fig. 28, for clearly if any other mean is chosen, the 'guaranteeable' performance will deteriorate. The dotted curve gives the ratio R, of the two values of $\mu t_2/t_1$ for each value of p.

The response time-constant p must be chosen to suit the pulse width and the extent to which it is proposed to attempt to correct the tuning error during a given pulse.

It will be seen from Fig. 27 that, although the Baron a.f.c. makes possible a more rapid response for given tolerances, these tolerances have a similar effect on the response speed to that which they have on the response of an 'extrapolating' servo.

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STANDARD-FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)

Values for May 1954

Date 1954 May	Frequency deviation from nominal: parts in 10^8		Lead of MSF impulses on GBR 1000 G.M.T. time signal in milliseconds
	MSF 60 kc/s 1429-1530 G.M.T.	Droitwich 200 kc/s 1030 G.M.T.	
1	−1·3	+4	NM
2	−1·3	+4	NM
3	−1·2	+5	NM
4	−1·2	+4	+6·5
5	−1·2	+4	+5·0
6	−1·3	+4	+4·2
7	−1·2	+4	+2·5
8	−1·2	+3	NM
9	−1·1	+4	NM
10	−1·2	+4	−1·6
11	−1·1	+3	−2·6
12	−1·0	+4	−3·6
13	−1·1	+3	−4·8
14	−1·3	+2	−5·9
15	−1·3	+4	NM
16	−1·3	+5	NM
17	−1·3	+4	−8·5
18	−1·3	+4	−9·2
19	−1·2	+4	−10·6
20	−1·2	+4	−11·0
21	NM	+4	−11·4
22	−1·2	+4	NM
23	−1·3	+4	NM
24	−1·2	+5	−12·8
25	−1·2	+5	−12·6
26	−1·2	+4	−13·8
27	−1·1	+6	−14·4
28	−1·1	−1	−14·2
29	−1·1	−2	NM
30	NM	−2	NM
31	−1·1	−2	−13·4

The values are based on astronomical data available on 1st June 1954.
NM = Not Measured.

NEW BOOKS

Statistical Methods in Electrical Engineering

By D. A. Bell. Pp. 175. Chapman & Hall, 37 Essex Street, London, W.C.2. Price 25s.

Some knowledge of statistics is almost indispensable for scientists, research workers and 'technologists'. Unfortunately, from the point of view of the majority of these workers, statistics is an essentially mathematical subject. From the point of view of a mathematician, like the reviewer, statistics is best treated as a branch of mathematics. There is a certain amount of fundamental technique and theory which must be mastered whatever the application; this includes the practical estimation of the mean, variance, and other moments, of correlation coefficients, and the theory of various standard statistical distributions and the relations between them. But there are some who, perhaps rightly from their own point of view, regard mathematics as a necessary evil which they wish to avoid as far as possible. If they are experienced engineers with little previous knowledge of statistics, they will find Dr. Bell's book helpful. For though Dr. Bell's style is somewhat involved, and there are too many confusing parenthetic insertions and asterisked explanatory footnotes, he has written in a language readily understood by experienced engineers. The fundamental technique and theory mentioned above is covered, but the distinctive part of Dr. Bell's book is the relation of statistics to such subjects as waveform analysis, noise and information theory.

The book arose from a series of lectures to post-graduate students of Birmingham University: it is not intended for the undergraduate student, for whom the language familiar to electrical engineers may require as much learning as the language of mathematics. There should not be the present unnatural separation between the subjects of radio-electricity and statistics, which are in fact closely related. Dr. Bell's book is welcome because it helps to reduce this separation. J. W. H.

Prediction of Optimum Traffic Frequencies

By Ing. Prof. Rufino Gea Sacasa. Pp. 20 with 2 diagrams and 5 nomograms. Obtainable from the author, El Encimar 10, Madrid, Spain. Price 38s.

For some years the production of ionospheric forecasts has been an important ancillary service to organizations engaged in radio communication. The laboratories responsible for preparing the forecasts usually issue six charts to represent conditions in the F_2 layer during a given month at a particular epoch in the solar cycle. These charts already contain approximations and it would be difficult to reduce their number without a serious loss of accuracy. In spite of the fact that the use of such charts is complicated, the tendency on the part of the operational users is to demand higher accuracy rather than a simpler means of presenting the data.

Prof. Gea Sacasa's method represents an attempt to compress into five nomograms all the information required to make forecasts of the optimum traffic frequency (f.o.t.) in all parts of the world at any part of the solar cycle and for any desired month. To achieve this, a very drastic simplification of what does in fact occur is inevitable and it is worth pointing out the implications which are inherent in the use of the nomograms.

The basic assumption which is made, referred to as the 'probable law', is that the f.o.t. for a 1300-km NS path is a linear function of time of day and that it increases until 1400 L.M.T. and falls thereafter. The functions used are: Morning: f.o.t. $= 10 + 2n$ Mc/s; Evening: f.o.t. $= 18 - 2n$ Mc/s where n = number of hours after sunrise or sunset as appropriate. From the way in which these relations are used, it is possible to derive from them the implied relations for the critical frequencies (f_0E, f_0F_2) of E and F_2 layers:

Morning: $f_0E = 2{\cdot}2 + 0{\cdot}5n$ Mc/s; $f_0F_2 = 3{\cdot}9 + 0{\cdot}8n$ Mc/s
Evening: $f_0E = 4{\cdot}0 - 0{\cdot}5n$ Mc/s; $f_0F_2 = 7{\cdot}0 - 0{\cdot}8n$ Mc/s

A representative sample of actual values of f_0F_2 and its slope at sunrise and sunset has been taken for 1952; the mean values of these parameters and their standard deviations lead to the relations:

Morning: $f_0F_2 = 3{\cdot}5 \pm 0{\cdot}7 + n(1{\cdot}2 \pm 0{\cdot}7)$ Mc/s
Evening: $f_0F_2 = 5{\cdot}9 \pm 1{\cdot}8 - n(1{\cdot}0 \pm 0{\cdot}9)$ Mc/s

It seems likely that the differences between the observed relations and those implied by the 'probable law' could be explained in terms of different levels of solar activity. For this reason it is difficult to see how any single 'probable law' can be satisfactory throughout the solar cycle. Quite apart from this criticism of the method, potential users of the nomograms must satisfy themselves that their needs can be met by a triangular wave approximation for the diurnal changes in f_0E and f_0F_2; and by the substitution of mean values for parameters whose standard deviations vary from 20% to 90% of the mean value.

Although a compact nomographic technique for calculating approximate values of f.o.t. would be useful, it seems fair to suggest that Prof. Gea Sacasa may have simplified his approximations to such an extent as to limit rather severely possible applications of his method.

C. M. M.

Thermionic Valves

By A. H. W. Beck, B.Sc.(Eng.), A.M.I.E.E. Pp. 570 + xvi. Cambridge University Press, 200 Euston Road, London, N.W.1. Price 60s.

This book is in three parts. The first covers the physical theory of electronics in three chapters on thermionic emission, secondary, field and photoelectric emission, and fluorescence and phosphorescence. The second provides the mathematical theory of electronics and covers electrostatic fields, electrostatic and magnetic electron optics of fields without space charge, space-charge flow and the diode, transit-time effects and fluctuation noise in valves. In the third part, types of valve are discussed and there is one chapter on triodes for low and medium frequencies, another on multigrid receiving valves and a third on transmitter valves. The interest then shifts to the higher frequencies in chapters covering velocity-modulated valves, triodes at ultra-high frequencies, travelling-wave tubes and beam interaction tubes, magnetrons and picture convertors and storage tubes.

The book is intended mainly for "graduates with a first degree in physics or electrical engineering and those who are starting independent work in this field, either in industry or in the various post-graduate courses on electronics".

Mathematics are freely used throughout, but not for their own sake. There are many consecutive pages of reading matter with no sign of an equation. It is, however, true to say that a good mathematical knowledge is necessary to follow a great deal of the discussion.

Ordinary types of valve are treated somewhat briefly, but not cathode-ray tubes nor gas-filled types. The whole emphasis is on valves for the higher frequencies.

Radio Research 1953

Report of the Radio Research Board for 1953. Pp. 40 + iv. Published for D.S.I.R. by H.M. Stationery Office, York House, Kingsway, London, W.C.2. Price 1s. 9d.

Laplace Transforms for Electrical Engineers

By B. J. STARKEY, Dipl.Ing., A.M.I.E.E. Pp. 279. Published for *Wireless Engineer* by Iliffe & Sons, Ltd., Dorset House, Stamford Street, London, S.E.1. Price 30s.

In the first two chapters, this book deals with the symbolic method, Fourier transformations, generalized impedances and cisoidal oscillations and, in Chapter 3, introduces the Laplace transform. Its application to electrical circuits comes in Chapter 4 and the next four chapters cover functions of a complex variable, integration in the complex plane, classification of functions of a complex variable and further analysis of contour integration. Chapters on Laplace transformation, its applications and theorems, the use of the inverse Laplace transformation and the generalization of the theory follow.

The book is intended for engineers rather than for the mathematician and the author claims to have used a physical rather than a mathematical vocabulary.

Applied Electronics Annual 1953-1954

Edited by R. E. BLAISE, A.M.Brit.I.R.E. Pp. 257. British-Continental Trade Press, Ltd., 222 Strand, London, W.C.2. Price 20s.

Discrete Sources of Extra-Terrestrial Radio Noise

Special Report No. 3 of the International Scientific Radio Union. Pp. 56. Available from the General Secretariat of the U.R.S.I., 42 Rue des Minimes, Brussels, Belgium. Price 75 fr.

Index to "Proc.I.E.E.", 1942-51

A 10-year index to the *Proceedings of the Institution of Electrical Engineers*, covering Parts I, IA, II, IIA, III, IIIA and Monographs (i.e., Part IV). It includes author and subject indexes. Pp. 500. The Secretary, The Institution of Electrical Engineers, Savoy Place, London, W.C.2. Price 25s.

Handbook of Industrial Electroplating (2nd Edition)

By E. A. OLLARD, A.R.C.S., F.R.I.C., F.I.M., and E. B. SMITH. Pp. 364. Published for *Metal Industry*, by Iliffe & Sons, Ltd., Dorset House, Stamford Street, London, S.E.1. Price 30s.

The Oscilloscope at Work

By A. HAAS and R. W. HALLOWS, M.A.(Cantab.), M.I.E.E. Pp. 171 with 102 diagrams and 217 oscillograms. Published for *Wireless World* by Iliffe & Sons, Ltd., Dorset House, Stamford Street, London, S.E.1. Price 15s.

This book was originally published in France. It has been translated and enlarged, and covers the form and applications of the general-purpose oscilloscope.

OBITUARY

It is with deep regret that we have to announce the death on 21st June of Francis Morley Colebrook after a short illness. Born in 1893, Mr. Colebrook was educated at Birkbeck College, London, and the City and Guilds College. He was with the Radio Division of the National Physical Laboratory from 1926-1949 when he became Officer-in-Charge of the Electronics Section.

He has contributed to *Wireless Engineer* and *Wireless World* and was the author of "Basic Mathematics for Radio Students". Since 1946 he has been a member of the *Wireless Engineer* Editorial Advisory Board.

As announced elsewhere, he was appointed an Officer of the Order of the British Empire in this year's Birthday Honours.

NATIONAL BUREAU OF STANDARDS:

Statistical Theory of Extreme Values and Some Practical Applications

By E. J. GUMBEL. National Bureau of Standards Applied Mathematics Series 33. Pp. 51. Price 40 cents.

Optical Image Evaluation

Proceedings of Symposium held October 1951. National Bureau of Standards Circular 526. Pp. 289. Price $2.25.

Tables of Lagrangian Coefficients for Sexagesimal Interpolation

National Bureau of Standards Applied Mathematics Series 35. Pp. 157. Price $2.

Table of Secants and Cosecants to Nine Significant Figures at Hundredths of a Degree

National Bureau of Standards Applied Mathematics Series 40. Pp. 46. Price 35 cents.

Effective Radio Ground-Conductivity Measurements in the United States

By R. S. KIRBY, J. C. HARMAN, F. M. CAPPS and R. N. JONES. National Bureau of Standards Circular 546. Pp. 87. Price 65 cents.

The National Bureau of Standards' publications can be obtained from the Government Printing Office, Washington 25, D.C., U.S.A. One-third of the publication price should be added to cover postage.

BIRTHDAY HONOURS

In the Birthday Honours List, Major-General L. B. Nicholls, C.B., C.B.E., M.I.E.E. (Chairman of Cable & Wireless) is appointed K.C.M.G.

T. E. Goldup, M.I.E.E. (Director, Mullard, Ltd., and Chairman of the Board of Governors of the Ministry of Supply School of Electronics at Malvern), has been appointed a Commander of the Order of the British Empire.

F. M. Colebrook, B.Sc., A.C.G.I., D.I.C. (Officer in charge of the Electronics Division, National Physical Laboratory), becomes an Officer of the Order of the British Empire.

A. I. Forbes Simpson of the radio equipment design department in the Coventry works of the General Electric Co., Ltd. becomes a Member of the Order of the British Empire.

INSTITUTE OF PHYSICS

Sir John Cockcroft (Director of the Atomic Energy Research Establishment, Harwell) was recently elected President of the Institute of Physics. G. R. Noakes (Assistant Master, Uppingham School) was elected Vice-President. Dr. S. Whitehead and Dr. B. P. Dudding were re-elected Honorary Treasurer and Honorary Secretary respectively.

INSTITUTION OF ELECTRONICS EXHIBITION

The ninth annual Electronics Exhibition is being held at the Manchester College of Technology, Sackville Street, Manchester, 1, from 14th-20th July. It will be open on Wednesday, 14th July, from noon to 10 p.m. and on other days (except Sunday) from 10 a.m. to 10 p.m. except Saturday, 17th July, when it will close at 7 p.m. The exhibition will be in two sections, one covering scientific and industrial research items and the other commercial apparatus.

Tickets are obtainable from W. Birtwistle, 78 Shaw Road, Thornham, Rochdale, Lancs.

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 selected list of journals abstracted, the abbreviations of their titles and their publishers' addresses.

ACOUSTICS AND AUDIO FREQUENCIES

534 + 621.395.61/.62] : 061.3 **1977**

Proceedings of the First I.C.A. [International Commission on Acoustics] **Congress on Electroacoustics.**—(*Acustica*, 1954, Vol. 4, No. 1, pp. 1–306.) Text of papers presented at the congress held in the Netherlands, June 1953. See also 2855 of 1953.

534.6 : 621.375 **1978**

Use of Variable-Gain Amplifiers in Acoustic Measurements.—A. Moles. (*Rev. gén. Élect.*, Jan. 1954, Vol. 63, No. 1, pp. 35–52.) Two methods of use are distinguished, namely (*a*) for maintaining a constant intensity of sound from the test source, and (*b*) for obtaining response curves of a sound source or detector or of an intervening transmission medium. Applications to the investigation of the acoustics of halls and of the transmission properties of partitions, etc., are described. See also 3767 of 1947.

534.75 : 621.3.018.78 **1979**

The Audibility of Linear Distortion of Natural [musical] **Sounds.**—N. Mayer. (*Funk u. Ton*, Jan. 1954, Vol. 8, No. 1, pp. 1–6.) Subjective tests were made using music reproduced by a loudspeaker associated with a variable-frequency-response amplifier. The least perceptible decrease of the mean amplitude over an octave is 4 db in the medium and high a.f. range and 10 db at the low-frequency end of the range.

534.84 **1980**

Behaviour of Sound in a Room with Absorbent Walls.—S. Eutizi. (*Alta Frequenza*, Feb. 1954, Vol. 23, No. 1, pp. 3–15.) The equation of propagation for a parallelepipedal room with absorbent walls is obtained in a form involving a series expansion, and an explicit third-approximation solution is given for a cubic room. Resonance frequencies and reverberation times are calculated for some examples. An improved formula is derived for calculating Sabine's absorption coefficient.

534.862.4 **1981**

Measurement and Evaluation of the Interference Effect of Noises.—E. Belger. (*Fernmeldetech. Z.*, Jan. 1954, Vol. 7, No. 1, pp. 25–32.) Report of subjective tests of the effect of warbled or interrupted tones and of thermal noise with a bandwidth of a musical third superimposed on reproduced music.

621.395.61 **1982**

Experiments on the Construction of a 'One-Dimensional' Directional Microphone.—G. Kurtze. (*Tech. Mitt. schweiz. Telegr.-TelephVerw.*, 1st Jan. 1954, Vol. 32, No. 1, pp. 27–31.) An arrangement in which sound arriving from undesired directions is suppressed by interference is achieved by mounting a slotted tube on the front of an ordinary microphone. A modification which is frequency independent within certain limits is also described. Calculated and measured polar diagrams are shown.

621.395.61 : 534.321.9 **1983**

Modulation of Ultrasonic Standing Waves in Air.—L. Pimonow. (*Ann. Télécommun.*, Jan. 1954, Vol. 9, No. 1, pp. 24–28.) Standing waves from a single ultrasonic source may be modulated by a l.f. source, e.g. the human voice. A 'gas-microphone' arrangement based on this principle is described, having 3% modulation depth at 30 kc/s. The device will not operate at frequencies below 15 kc/s. Theory is developed to explain the phenomena.

621.395.623.7 : 534.321.9 **1984**

Marked Demodulation in Air of Two Ultrasonic Waves of Different Frequencies.—S. Klein. (*Ann. Télécommun.*, Jan. 1954, Vol. 9, No. 1, pp. 21–23.) See 958 of April (Maulois).

621.395.625.3 : 621.385.832 **1985**

Investigation of Core Structures for the Electron-Beam Reproducing Head in Magnetic Recording.—J. W. Gratian. (*Trans. Inst. Radio Engrs*, Jan./Feb. 1954, Vol. AU-2, No. 1, pp. 27–38.) An investigation has been made of factors contributing to losses at the upper end of the frequency range in the electron-tube magnetic pickup described by Skellett et al. (11 of January). Strip-type structures have been designed for the external cores; an output of the order of 0·2 V and a frequency range of about 1 c/s–15 kc/s is obtained with a tape speed of 10 in./s.

AERIALS AND TRANSMISSION LINES

621.315.212 1986

The Evaluation of Cable Irregularities at Very High Frequencies.—W. W. H. Clarke & J. D. S. Hinchliffe. (*Proc. Instn elect. Engrs*, Part IV, Feb. 1954, Vol. 101, No. 6, pp. 55–60. Digest, *ibid.*, Part III, Jan. 1954, Vol. 101, No. 69, pp. 44–46.) "The magnitude and distribution of cable irregularities are related statistically to the end-to-end input-impedance difference under conditions involving a large number of wavelengths and appreciable attenuation. The manner of variation of the measured quantities with frequency and irregularity magnitude and distribution is revealed. These relationships are shown in the form of master curves for a particular cable on the assumption of exponential fault correlation."

621.372 1987

Coupling of Modes of Propagation.—J. R. Pierce. (*J. appl. Phys.*, Feb. 1954, Vol. 25, No. 2, pp. 179–183.) "When two lossless modes of propagation are coupled, waves which increase or decrease with distance may arise when the power flow of the two modes is in opposite directions or when power is generated in the coupling device. This behavior is characteristic of wave filters, traveling-wave tubes, double-stream amplifiers, and space-charge-wave amplifiers. Such behavior is analysed assuming linearity and conservation of energy only."

621.372.2 1988

Designing Surface-Wave Transmission Lines.—G. Goubau. (*Electronics*, April 1954, Vol. 27, No. 4, pp. 180–184.) The theory of the surface-wave line is outlined and the effects of precipitation are discussed. Measurements have been made at about 250 Mc/s on a 2-mile line with a loss of 6 db per mile, and at about 2 kMc/s on a 130-ft aerial feeder line with a total loss of about 2 db. Design graphs are given for lines using polyethylene-coated Cu wire. Launching loss is also considered; best results were obtained using a horn with an inner conductor tapering down to the line diameter.

621.372.2 : 621.396.67 1989

Foundations of an Exact Theory of the Wave Field of a Transmission Line.—G. A. Grinberg & B. E. Bonshtedt. (*Zh. tekh. Fiz.*, Jan. 1954, Vol. 24, No. 1, pp. 67–95.) The propagation of e.m. waves along a straight conductor of circular cross-section, located above and parallel to the surface of a plane homogeneous earth is considered theoretically. The exact solution is derived first, then approximate solutions are found and the effect of the earth on the field above it is calculated. The complex velocity of propagation along the conductor, the effective parameters of the line and wave field near the earth are also calculated. The solutions by Sommerfeld, Mie, Pollaczek (*Elekt. NachrTech.*, 1927, Vol. 4, pp. 295–304 & 515–525) and others are briefly discussed.

621.372.2.029.64 1990

Microstrip — A New Transmission Technique for the Kilomegacycle Range.—D. D. Grieg & H. F. Engelmann. (*Proc. Instn Radio Engrs, Aust.*, Jan. 1954, Vol. 15, No. 1, pp. 13–19.) Reprint. See 621 of 1953.

621.372.8 1991

Transmission and Matching Theory of Homogeneously Guided Waves.—F. J. Tischer. (*Arch. elekt. Übertragung*, Jan. & Feb. 1954, Vol. 8, Nos. 1 & 2, pp. 8–14 & 75–84.) Mathematical expressions for propagation in waveguide systems are derived directly from Maxwell's equations. The orthogonal curvilinear coordinate system is used. Particular cases considered include propagation in cylindrical systems and the effects of terminations, inhomogeneities, discontinuities and coupling.

621.372.8 1992

Propagation of Microwaves through a Cylindrical Metallic Guide filled Coaxially with Two Different Dielectrics: Part 4.—S. K. Chatterjee. (*J. Indian Inst. Sci.*, Section B, Jan. 1954, Vol. 36, No. 1, pp. 1–13.) The field components and propagation constants for hybrid modes are derived theoretically. Part 3: 1671 of June.

621.372.8 1993

Representation of the Electromagnetic Field in a Waveguide with Absorbent Walls.—M. De Socio. (*R. C. Accad. naz. Lincei*, Jan. 1954, Vol. 16, No. 1, pp. 63–68.) Analysis is given for a system comprising plane parallel conductors with a homogeneous dielectric filling the space between them. The field can be represented by two superimposed evanescent plane waves, of which the one can be considered as the reflection of the other at the walls of the interspace.

621.396.67 1994

Radiation Resistance and Gain of Homogeneous Ring Quasi-Array.—H. L. Knudsen. (*Proc. Inst. Radio Engrs*, April 1954, Vol. 42, No. 4, pp. 686–695.) Continuation of analysis presented previously (2570 of 1953). Ring arrays of tangential or radial dipoles are considered. To simplify the calculation of gain and radiation resistance the number of dipoles is assumed to be infinite. The method of calculation is similar to that of Foster (46 of 1945), who investigated the particular case of constant phase round the ring, and is also related to that of Page (1862 of 1948) for axial dipoles.

621.396.676 1995

Designing Flush Antennas for High-Speed Aircraft.—J. V. N. Granger. (*Electronics*, March 1954, Vol. 27, No. 3, pp. 136–140.) A general discussion of the design problems encountered.

621.396.677.3.091.22 1996

A New Method of Measuring the Gain of Linear-Array Antennas.—S. Uda & Y. Mushiake. (*Sci. Rep. Res. Inst. Tohoku Univ., Ser. B*, Dec. 1952, Vol. 4, No. 1, pp. 51–65.) A method of calculating the gain of linear aerials from the measured radiation patterns in the equatorial plane is briefly described. The necessary formulae, which involve sine power-series and use of Simpson's summation rule, are derived. An alternative graphical method is indicated. The methods are particularly applicable to the Yagi-Uda type aerial.

621.396.677.4 1997

Radiation from a Ground Antenna.—J. R. Wait. (*Canad. J. Technol.*, Jan. 1954, Vol. 32, No. 1, pp. 1–9.) Theory of the wave aerial is developed and possible alternative modes of operation are discussed. Curves are presented for determining radiation characteristics.

621.396.677.8 : 538.52 : 537.311.5 1998

On the Current induced in a Conducting Ribbon by a Current Filament Parallel to it.—E. B. Moullin. (*Proc. Instn elect. Engrs*, Part IV, Feb. 1954, Vol. 101, No. 6, pp. 7–17.) Full paper. See 704 of March.

621.396.677.832 1999

Measurements on Corner-Reflector Aerials.—G. Burkhardt. (*Fernmeldetech. Z.*, Feb. 1954, Vol. 7, No. 2, pp. 55–56.) Curves are presented showing variation of gain with distance *a* between dipole and vertex, at 50–70 cm λ, for 60° and 72° corner-reflector aerials, for values of *a* up to 2λ and $1{\cdot}6\lambda$ respectively. For wide-band aerials a reflector angle of 72° is better than one of 60° or 90° because the gain variation is not so great. See also 2224 of 1953 (Harris).

621.396.677.85 **2000**

Tolerances in Parameters of Microwave Lenses.—D. H. Shinn & T. C. Cheston. (*Marconi Rev.*, 1st Quarter 1954, Vol. 17, No. 112, pp. 1–9.) Lenses comprising metal plates or square metal tubes are considered. Formulae for the constructional and electrical tolerances derived by Risser (*Microwave Antenna Theory and Design*, chapter 11) are generalized and extended. Some consistent errors can be corrected by refocusing. Tolerances in phase-corrected reflectors are treated in an appendix.

621.396.677.85 **2001**

Secondary Beams from Metal Lenses.—T. C. Cheston. (*Marconi Rev.*, 1st Quarter 1954, Vol. 17, No. 112, pp. 10–15.) Microwave lenses of parallel-plate or square-tube type are considered, and the conditions for the occurrence of secondary beams are derived using diffraction-grating theory. A sphero-ellipsoidal lens is less liable to generate secondary beams than is a plano-ellipsoidal or a plano-hyperboloidal lens. For scanning over wide angles it is impractical to design lenses free from secondary beams, because the required low value of refractive index is undesirable from the point of view of matching.

AUTOMATIC COMPUTERS

681.142 **2002**

Partial Drift Compensation in Electronic D.C. Analog Computers for Differential Equations.—L. E. Löfgren. (*Appl. sci. Res.*, 1954, Vol. B4, Nos. 1/2, pp. 109–123.) A method particularly suitable for computers with time-shared elements is described. The number of compensation points can be much smaller than the number of drift sources. The method is illustrated by describing the compensation of a machine for computing direction cosines.

681.142 **2003**

Analog Computing by Heat Transfer.—P. H. Savet. (*Tele-Tech*, Feb. 1954, Vol. 13, No. 2, pp. 101, 122.) Multiplier, divider, a.c. integrating and differentiating circuits are described, based on the use of a thermal transducer. This transducer consists basically of a pair of heater elements each one of which is in thermal contact with one of a pair of temperature sensing elements mounted in a Wheatstone bridge circuit. Appropriate application to the heater elements of two electrical input signals creates a temperature difference proportional to their product between the sensing elements.

681.142 **2004**

FOSDIC — A Film Optical Sensing Device for Input to Computers.—(*Tech. News Bull. nat. Bur. Stand.*, Feb. 1954, Vol. 38, No. 2, pp. 24–27.) An instrument is described for processing written records, such as answers to questionnaires. Marks made with ordinary pen or pencil at special locations on microfilm are converted into electrical pulses, by means of a flying-spot scanning system. The pulses are recorded on magnetic tape for input to the computer. See also *Tele-Tech*, Feb. 1954, Vol. 13, No. 2, pp. 78–79..140.

681.142 **2005**

Fast-Acting Digital Memory Systems.—I. L. Auerbach. (*Elect. Mfg*, Oct. & Nov. 1953, Vol. 52, Nos. 4 & 5, pp. 100–107 & 136–143.) A survey of moderate- and high-speed storage techniques for digital computers.

681.142 : 538.221 **2006**

Penetration of an Electromagnetic Wave into a Ferromagnetic Material.—A. Papoulis. (*J. appl. Phys.*, Feb. 1954, Vol. 25, No. 2, pp. 169–176.) An investigation of the magnetization of metallic-ribbon toroidal cores with rectangular hysteresis loop, used as computer storage elements. Approximate solutions are obtained for the field inside the core under different conditions of loading. Theoretical and experimental output waveforms show close agreement for high magnetizing force and thickness of core material $> 0{\cdot}5$ mil.

681.142 : 538.221 : 621.318.134 **2007**

Ferromagnetic Spinels with Rectangular Hysteresis Loops.—I. J. Hegyi. (*J. appl. Phys.*, Feb. 1954, Vol. 25, No. 2, pp. 176–178.) Tests made on materials of different composition show $MgO.3MnFe_2O_4$ to be most suitable for magnetic storage elements. For rings of outside diameter 55 mil, inside diameter 35 mil, the coercive force is about 3 oersteds and switch-over time $0{\cdot}3\ \mu s$. Rings of 0·5-cm outside diameter have a lower coercive force.

681.142 : 621-526 **2008**

Servo Control of the Position and Size of an Optical Scanning System.—T. Kilburn & E. R. Laithwaite. (*Proc. Instn elect. Engrs*, Part IV, Feb. 1954, Vol. 101, No. 6, pp. 129–134.) "A method of controlling the position and size of a television-type raster relative to arbitrary references by means of four servo mechanisms is described. One of the applications of such a system is that of a reading device for punched cards. A system for reading any one of a stack of cards without removing the selected card has been developed using the servo-controlled raster. The design of the servo mechanisms for this application is described in detail."

681.142 : 621.3 (084.2) **2009**

Electronic-Circuit Technique for a High-Speed Computer.—G. Piel. (*Onde élect.*, Jan. 1954, Vol. 34, No. 322, pp. 38–46.) Commonly used computer-circuit units based on a triode or Ge diode are represented by a symbol indicating the unit and its mode of operation. The functions of different combinations of the basic units are listed and their application in a serial binary-scale computer is illustrated using the notation described.

681.142 : 621.318.572 **2010**

New Flip-flop Chain Circuits used in Computers for counting to Base 10 and Base 12.—J. J. Bruzac. (*Onde élect.*, Jan. 1954, Vol. 34, No. 322, pp. 59–62.) Modifications to the Potter decade arrangement are described.

681.142 : 621.372 **2011**

An Improved Experimental Iteration Method for Use with Resistance-Network Analogues.—G. Liebmann & R. Bailey. (*Brit. J. appl. Phys.*, Jan. 1954, Vol. 5, No. 1, pp. 32–35.) A development of the method previously described by Liebmann (2839 of 1952). The error signals for a number of network nodes are displayed simultaneously on a c.r.-tube screen, so that it is always possible to work on the worst error instead of making adjustments cyclically.

681.142 : 621.374.5 **2012**

The Mercury-Delay-Line Storage System of the Ace Pilot Model Electronic Computer.—E. A. Newman, D. O. Clayden & M. A. Wright. (*Proc. Instn elect. Engrs*, Part II, Feb. 1954, Vol. 101, No. 79, p. 65.) Discussion on 61 of January.

681.142 : 621.375.2.024 **2013**

Time-Shared Amplifier stabilizes Computers.—D. W. Slaughter. (*Electronics*, April 1954, Vol. 27, No. 4, pp. 188–190.) Drift in the d.c. operational amplifiers of analogue computers is reduced by connecting them periodically to the output of an auxiliary stabilizing amplifier. The advantages of using a direct-coupled rather than a chopper type of amplifier for this purpose are indicated. Circuit details are described, including the switching and filtering arrangements.

CIRCUITS AND CIRCUIT ELEMENTS

534.321.9 : 534.213 : 621.396.6 **2014**

Metal Ultrasonic Delay Lines.—R. W. Mebs, J. H. Darr & J. D. Grimsley. (*J. Res. nat. Bur. Stand.*, Nov. 1953, Vol. 51, No. 5, pp. 209–220.) An experimental investigation was made with the object of finding a metal or alloy suitable for the transmission of 10-Mc/s pulses at temperatures between −50°C and + 200°C with a delay < 50 μs independent of temperature. The effect of various constructions, crystal attachment methods and materials, and treatment of the delay-line material was investigated; the results are tabulated and shown graphically. An isoelastic iron alloy containing ~ 36% Ni, 7–8% Cr and other minor constituents, used with overcured epoxy-resin quartz-crystal attachments, gave the best transmission characteristics. A shorter account of the work is given in *Tech. News Bull. nat. Bur. Stand.*, March 1954, Vol. 38, No. 3, pp. 38–39.

621.3.015.3 : 517.942.82 **2015**

The Calculation of Transients in Dynamical Systems.—Ward. (See 2163.)

621.3.018.75 : 621.387.4 **2016**

A Stable High-Speed Multichannel Pulse Analyzer.—E. Gatti. (*Nuovo Cim.*, 1st Feb. 1954, Vol. 11, No. 2, pp. 153–162. In English.) Pulse selection in each channel is accomplished by a single discriminator whose threshold is set at a height corresponding to the upper boundary of the channel, the channel width being determined by suitably shaping the incoming pulses.

621.316.726.029.6 : 621.376.3 **2017**

Frequency Discrimination and Stabilization of Square-Wave Modulated Microwave Transmissions.—C. H. M. Turner. (*Proc. Instn elect. Engrs*, Part IV, Feb. 1954, Vol. 101, No. 6, pp. 61–72.) Full paper. See 652 of March.

621.316.729 **2018**

Pulling Effect in Synchronized Systems.—Z. Jelonek, O. Celinski & R. Syski. (*Proc. Instn elect. Engrs*, Part IV, Vol. 101, No. 6, pp. 108–117. Digest, *ibid.*, Part III, Vol. 101, No. 69, pp. 50–52.) Synchronization of a valve oscillator with a low-pass filter in the feedback loop is considered, and the phase-equation method of analysis applied, with parameters M and T, proportional to the detuning and to the filter time-constant respectively. For $|M| < 1$, the operating point lies within the synchronization range, but the system remains in the asynchronous state (i.e., pulling effect exists) for values of $|M|$ greater than a critical value M_0. For $|M| < M_0$ the system is synchronized, and, if this is an initial condition of the system, synchronization will be maintained for values of $|M|$ up to unity. A graph of the pulling function, i.e., the relation between M_0 and T, obtained in part by analysis, in part from experiment, is shown. The case of a limiter in the feedback loop is also considered.

621.316.8.029.5 **2019**

The Design of a Radio-Frequency Coaxial Resistor.—C. T. Kohn. (*Proc. Instn elect. Engrs*, Part IV, Feb. 1954, Vol. 101, No. 6, pp. 146–153. Digest, *ibid.*, Part III, Jan. 1954, Vol. 101, No. 69, pp. 48–50.) Characteristics of the cylindrical coaxial resistor, intended for use as a reflection-free termination, are discussed and design curves given. The jacket diameter is chosen to maintain the resistive component of the resistor at the required value. The reactive component may be compensated in two ways. The first, suitable for resistor lengths < 0·08λ, consists in undercutting the line. The second requires the insertion of a short-circuited transmission line in series with the resistor, and will give either perfect compensation over a restricted frequency range, or less accurate compensation over a wider range. The effect of the holder is taken into account.

621.318.4 **2020**

Iron-free Cylindrical Coils with Reduced External Field.—G. Kirschstein. (*Arch. Elektrotech.*, 1954, Vol. 41, No. 4, pp. 222–230.) The design of coils with zero magnetic moment is described and the necessary formulae are given. The arrangement consists of three coaxial coils connected so that the outer coil produces a field opposing that of the two inner coils and of such magnitude as to reduce the external field to nearly zero. The internal field is approximately equal to that of the innermost coil alone. Coils of this type are useful, e.g., for focusing electron beams, for magnetic measurements and for inductance standards. See also *Phys. Z.*, 15th Feb. 1940, Vol. 41, Nos. 3/4, pp. 53–55 (Steinhaus & Kussmann).

621.318.423 **2021**

Design of High-Frequency Coils for High Currents.—E. Karger. (*Funk u. Ton*, Jan. 1954, Vol. 8, No. 1, pp. 7–18.) A practical guide to the design of single-layer air-cored cylindrical coils and a short-circuit variometer. The dissipation of heat, electrical breakdown potential, and the various losses are considered. Formulae, tables and an inductance/coil-dimensions nomogram are given.

621.319.4 **2022**

The Performance of Dried and Sealed Mica Capacitors.—G. H. Rayner & L. H. Ford. (*J. sci. Instrum.*, Jan. 1954, Vol. 31, No. 1, pp. 3–6.) Good stability has been achieved in an 0·01-μF capacitor by drying it for a year and then mounting it in a sealed container. The interdependence of the variations of capacitance and power factor over the frequency range 10 c/s–10 kc/s is discussed.

621.319.45 **2023**

Tantalum Electrolytic Capacitors.—Nguyen Thien-Chi & J. Vergnolle. (*Ann. Radioélect.*, Jan. 1954, Vol. 9, No. 35, pp. 83–97.) A general review of types of electrolytic capacitors and problems of design, manufacture and testing of Ta types. Data are presented for three C.S.F. capacitors, having volumes of 1·7, 1·1 and 0·5 cm³, with average capacitances of 50, 25 and 12 μF respectively at 70-V operation.

621.372 **2024**

Normalization of the Frequency Dependence of Impedance and Amplifier Circuits: Outline of Generalized Circuit Theory.—K. H. R. Weber. (*NachrTech.*, Jan. 1954, Vol. 4, No. 1, pp. 13–19.)

621.372 : 517.63 **2025**

Some Observations on Time considered as a Complex Variable.—E. C. Cherry. (*Onde élect.*, Jan. 1954, Vol. 34, No. 322, pp. 7–13.) Results of applying this concept in Fourier and Laplace transformations are examined, and its application to the analysis of transients is discussed. A practical interpretation of reversing the real-time/complex-frequency relation is given with reference to echo phenomena.

621.372 : 621.314.7 **2026**

The Transistor as a Network Element.—J. T. Bangert. (*Bell Syst. tech. J.*, March 1954, Vol. 33, No. 2, pp. 329–352.) The use of transistors for (*a*) reduction of dissipation, (*b*) elimination of inductance, (*c*) production of delay, (*d*) inversion of impedance is discussed. Improvements in performance can be achieved which would otherwise be unobtainable or uneconomic.

621.372 : 621.396.822 : 530.162 2027

The Brownian Movement of Linear and Nonlinear Systems.—D. K. C. MacDonald. (*Phil. Mag.*, Jan. 1954, Vol. 45, No. 360, pp. 63–68.) The application of Brownian-movement analysis to systems with a nonlinear relaxation mechanism (e.g., electrical conductivity) is considered. A statistical hypothesis is proposed which enables a calculation to be made of the Brownian movement and corresponding frequency spectrum of such systems. The results are relevant to the study of electrical noise.

621.372 : 621.396.822 : 530.162 2028

Note on the Theory of Brownian Motion in Nonlinear Systems.—D. Polder. (*Phil. Mag.*, Jan. 1954, Vol. 45, No. 360, pp. 69–72.) The hypothesis advanced by MacDonald (2027 above) is shown to be not self-consistent.

621.372.5 2029

A New Method of Synthesis of Reactance Networks.—A. Talbot. (*Proc. Instn elect. Engrs*, Part IV, Feb. 1954, Vol. 101, No. 6, pp. 73–90. Digest, *ibid.*, Part III, Jan. 1954, Vol. 101, No. 69, pp. 46–48.) Since the chain matrix of a cascade combination of quadripoles is the product of the individual chain matrices, and since synthesis of a complicated network by the insertion-loss method involves obtaining a number of simple sections to be connected in cascade, the chain matrix is chosen for this purpose. A method of factorizing any realizable chain matrix into two such matrices of lower order is presented. Repeated factorizations ultimately yield sections simple enough to be synthesized. The method involves only simple algebra and a new theorem concerning reactance and impedance functions.

621.372.5 2030

Lumped-Parameter Delay Lines.—H. Feissel. (*Onde élect.*, Jan. 1954, Vol. 34, No. 322, pp. 53–58.) The application of rigorous filter theory to delay lines comprising a series of coaxial coils is difficult owing to the coupling between nonadjacent coils. An experimental method of obtaining frequency characteristics is outlined. The latter can be corrected by connecting a capacitor across adjacent coils. Filter theory is applied to give an indication of suitable capacitance values.

621.372.5 : 621.3.015.3 2031

Monotonic Transient Response.—O. P. D. Cutteridge. (*Proc. Instn elect. Engrs*, Part IV, Feb. 1954, Vol. 101, No. 6, pp. 46–54. Digest, *ibid.*, Part III, Jan. 1954, Vol. 101, No. 69, pp. 43–44.) Restrictions are derived which must be placed on the poles and zeros of the system function of a linear system with lumped constants in order that the response of the system to a step-function drive shall be monotonic. System functions having up to three zeros are considered. A method of dealing with systems containing more than three zeros is explained.

621.372.5 : 621.3.018.78 2032

Distortion of Arbitrarily Shaped Curves by *RC* Sections.—E. William. (*Funk u. Ton*, Jan. & Feb. 1954, Vol. 8, Nos. 1 & 2, pp. 30–46 & 87–99.) A method is given of calculating the distortion by an *RC* network of any signal which is expressible as the sum of a power series and terms containing exponential functions of the circuit time constants. Tables and nomograms are given for calculating the output form given the input curve, and vice versa, for rectangular, trapezoidal, triangular, parabolic and exponential-type pulses.

621.372.51 : 621.396.67 2033

Matching Circuits for Asymmetrical Wire Aerials.—A. Simon. (*Frequenz*, Feb. 1954, Vol. 8, No. 2, pp. 48–56.) A comparison of various types of aerial coupling circuits used particularly at medium and long wavelengths, and of the ease of adjustment for use at different frequencies. The discussion shows that the most advantageous circuits consist of two variable reactances used in conjunction with an impedance meter. Examples of this type for use at 1·5–12 Mc/s or at medium waves are shown.

621.372.512 2034

Action of Periodic Telegraphy Signals on Cascaded *RLC* Resonant Circuits.—J. Marique. (*Rev. HF, Brussels*, 1954, Vol. 2, No. 9, pp. 233–244.) The response to rectangular, trapezoidal and sine-squared pulses of a cascaded arrangement of up to three similar tuned circuits is investigated theoretically and the results compared with the response of a single tuned circuit (3241 of 1953). The effect of bandwidth and signal repetition frequency is studied from the point of view of using the circuit for Fourier analysis and for reception of telegraphy.

621.372.54 2035

Explicit Formulae for the Calculation of Filter Circuits with Generalized Parameters.—V. Fetzer. (*Arch. elekt. Übertragung*, Jan. 1954, Vol. 8, No. 1, pp. 31–46.) An extension of previous work (1545 of 1952) to the antimetrical low-pass, the symmetrical and the antimetrical band-pass characteristic functions and the determination of the transmission factor.

621.372.54 2036

The Smoothing of Loss Attenuation in Basic Types of Ladder Filter.—H. Matthes. (*Frequenz*, Dec. 1953, Vol. 7, No. 12, pp. 360–368 & Jan. 1954, Vol. 8, No. 1, pp. 17–28.) The characteristics of basic and terminating half-sections with losses are derived from image-parameter theory, and the results applied to show how the attenuation can be smoothed by introducing mismatch at the end of the filter or by inserting series and/or shunt resistors at properly chosen points within the network.

621.372.54 : 621.3.015.3 2037

The Transient Response of R.F. and I.F. Filters to a Wave Packet.—A. W. Gent. (*Proc. Instn elect. Engrs*, Part IV, Feb. 1954, Vol. 101, No. 6, p. 164.) Discussion on 963 of 1953.

621.372.56 2038

A Stable Voltage-Controlled Logarithmic Attenuator.—G. E. Boggs. (*Proc. Inst. Radio Engrs*, April 1954, Vol. 42, No. 4, pp. 696–700.) The variable-impedance element is a triode operated so that the cathode impedance varies approximately inversely as the transconductance. Stabilization is achieved by the use of d.c. feedback with a suitable increase in loop gain. Design procedure is given for both high- and low-input types. Experimental results are discussed.

621.372.6 2039

Design of *RC* Wide-Band 90-Degree Phase-Difference Network.—D. K. Weaver, Jr. (*Proc. Inst. Radio Engrs*, April 1954, Vol. 42, No. 4, pp. 671–676.) The arrangement comprises two all-pass networks, as described by Orchard (1356 of 1950). The pole-zero pairs are first determined so as to give the 90° difference of phase shift over the frequency band, and the networks are synthesized from the response functions thus found. Design procedure is detailed step by step and is illustrated by a numerical example. Notes on construction and alignment are included.

621.372.8 : 538.614 **2040**

A Nonreciprocal Microwave Component.—M. L. Kales, H. N. Chait & N. G. Sakiotis. (*J. appl. Phys.*, June 1953, Vol. 24, No. 6, pp. 816–817.) In the presence of a steady magnetic field the permeability of a ferrite is an asymmetrical tensor. This property is applied in the construction of a nonreciprocal device comprising a waveguide in which a ferrite block is arranged asymmetrically and parallel to the axis, a static magnetic field being applied transversely. Experimental results are given.

621.373.4 **2041**

Valve Oscillators with Voltage-Controlled Frequency Dependence.—H. Wilde. (*Frequenz*, Jan. 1954, Vol. 8, No. 1, pp. 1–7.) A discussion of various circuits in which reactance is varied by application of control voltage, and a comparison of their performance in respect of bandwidth and output at medium and high frequencies.

621.373.42 **2042**

Ultra-Low-Frequency Three-Phase Oscillator.—G. Smiley. (*Proc. Inst. Radio Engrs*, April 1954, Vol. 42, No. 4, pp. 677–680.) An oscillator for the frequency range 0·01 c/s–1 kc/s uses three identical networks each comprising a polystyrene capacitor and a stable resistor, in a star arrangement radiating from the power supply. The circuit is in effect a three-stage amplifier, oscillations occurring at the frequency for which the phase shift is 60° in each *RC* network. Use is made of the Miller effect to keep down the size and cost of the frequency-determining elements.

621.373.42.029.55/.62 : 621.318.572 **2043**

Electronically-Tuned Wide-Range Oscillator.—D. D. King & R. L. Konigsberg. (*Electronics*, March 1954, Vol. 27, No. 3, pp. 184–186.) A helical transmission line with Ge diodes mounted at intervals of $\lambda/4$ is used as tuning element in a Colpitts circuit. Tuning is accomplished by applying a switching pulse to a selected diode to short the line. The five tuning positions give frequencies of about 8·6, 14, 22, 34 and 46 Mc/s respectively. The length of helical line required is very much less than that of a parallel-conductor line for the same frequency. Satisfactory operation of an experimental model was achieved at switching rates up to 1 000 pulses/sec.

621.373.421 **2044**

Perturbations of a Nonlinear Filtered Oscillator.—G. Cahen. (*Onde élect.*, Jan. 1954, Vol. 34, No. 322, pp. 80–89.) Analysis, summarized earlier (1279 and 1624 of 1953), of the effect of a disturbance on the behaviour of an oscillator comprising a nonlinear amplifier, feedback loop and low-pass filter. See also 972 of 1953 (Cahen & Loeb).

621.373.421.029.4/.51 **2045**

Audio Oscillator uses New *RC* Design.—J. H. Owens. (*Electronics*, March 1954, Vol. 27, No. 3, pp. 176–177.) A signal generator covering the range 11 c/s–100 kc/s in four bands has both low-pass and high-pass *RC* filters incorporated in the feedback loop to give good frequency stability and good waveform. Amplified a.g.c. is included in a commercial adaptation of the circuit.

621.374 : 621.314.632 : 546.289 **2046**

"Positive-Gap" Germanium Diode.—Reeves. (See 2256.)

621.374 : 621.396.82 **2047**

Noise Discriminator for Periodic Signals.—R. L. Conger & L. E. Schilberg. (*Rev. sci. Instrum.*, Jan. 1954, Vol. 25, No. 1, pp. 52–54.) The periodic signal, of any waveform, is applied through a phase inverter to one pair of terminals of a resistance bridge including two diode rectifiers. Periodic pulses are applied to the other pair of bridge terminals. During each pulse the diodes are unblocked and the signal passes to an integrating circuit with an adjustable time constant. If the pulse frequency is slightly greater than the signal frequency the output obtained has the waveform of the noise-free input signal and a much lower frequency; random noise, mains interference and any other disturbance of a frequency which is not a multiple of the signal frequency will not be reproduced. The unit was designed for use with nuclear magnetic resonance apparatus.

621.374.4 : 621.373.421.13 **2048**

Subharmonic Crystal Oscillators.—M. O. Thompson, Jr, C. E. Tschiegg & M. Greenspan. (*Rev. sci. Instrum.*, Jan. 1954, Vol. 25, No. 1, pp. 8–12.) The operating frequency of a relaxation oscillator can be stabilized at a submultiple of a crystal frequency. In a conventional blocking oscillator with a crystal connected across the third winding of the pulse transformer, frequency division by factors up to several thousand can be effected. A multivibrator with crystal between anode and grid of one valve gave good frequency and pulse-width stability when operating with a division factor of 100. Details of these two circuits are given; three others have been examined, namely a screen-coupled-phantastron, a thyratron and a transistor circuit.

621.375 : 534.6 **2049**

Use of Variable-Gain Amplifiers in Acoustic Measurements.—Moles. (See 1978.)

621.375.132.018.78 **2050**

Harmonic Distortion and Negative Feedback.—E. E. Zepler. (*Wireless Engr*, May 1954, Vol. 31, No. 5, pp. 118–121.) Different methods of defining gain in relation to the output-voltage/input-voltage characteristic of an amplifier are discussed; for a nonlinear amplifier it is reasonable to base the definition on the average slope. A method is indicated for constructing the characteristic of the amplifier with feedback from that of the amplifier without. Distortion can be reduced by means of feedback even for flat regions of the characteristic; this result is in disagreement with that of Rowlands (2278 of 1953). Detailed analysis is given for several types of amplifier.

621.375.2.018.75 : 621.373.431.1 **2051**

Application of the Multivibrator Principle in Counter-Tube Amplifiers.—W. Kroebel & G. Stutzer. (*Z. angew. Phys.*, Jan. 1954, Vol. 6, No. 1, pp. 14–19.) A description is given of a pulse amplifier based on a multivibrator circuit, which gives an output pulse of amplitude 180 V with a leading-edge slope of $1·1 \times 10^{-9}$ sec/V for a 0·3-mV input pulse.

621.375.222 **2052**

Cathode-Coupled Valves.—T. W. Brady. (*Wireless Engr*, May 1954, Vol. 31, No. 5, pp. 111–114.) A graphical method is described for investigating cathode-coupled amplifiers. The method is demonstrated in relation to a circuit using a Type-ECC33 twin triode, and involves the construction of composite load lines on the anode characteristics. An initial trial solution must be found to determine the voltage across the cathode resistor in the quiescent case.

621 375.232.3 **2053**

An A.C. Cathode-Follower Circuit of Very High Input Impedance.—J. R. Macdonald. (*Rev. sci. Instrum.*, Feb. 1954, Vol. 25, No. 2, pp. 144–147.) A push-pull amplifier for the range 10^2–10^4 c/s is described. Each half of the input consists of a cathode follower with gain approximately unity with a constant-current cathode load constituted by a further valve. The effect of grid-anode capacitance in the input valve is reduced by driving the anode by another cathode follower in series with it. Input capacitance is $< 0{\cdot}3$ pF and input resistance $> 4 \times 10^9\,\Omega$ up to $3{\cdot}2$ kc/s. Circuit modifications to obtain zero or negative input capacitance are explained.

621.375.3 **2054**

Flux Preset High-Speed Magnetic Amplifiers.—C. B. House. (*Elect. Engng, N.Y.*, Jan. 1954, Vol. 73, No. 1, p. 51.) Digest of paper to be published in *Trans. Amer. Inst. elect. Engrs*, 1953, Vol. 72.

621.375.3 **2055**

Parallel-Connected Magnetic Amplifiers.—S. H. Chow. (*J. appl. Phys.*, Feb. 1954, Vol. 25, No. 2, pp. 216–221.) A complete analysis, based on a single-valued B/H relation, is made for three types of load: resistive, inductive and capacitive. Transient response is considered. Analysis of feedback amplifiers shows clearly the 'jump' phenomenon in the output characteristic.

621.375.4 **2056**

High-Frequency Transistor Amplifiers.—W. F. Chow. (*Electronics*, April 1954, Vol. 27, No. 4, pp. 142–145.) Transistor equivalent circuits appropriate for h.f. wide-band operation are presented. Input and output impedance and power gain for grounded-base and grounded-emitter arrangements are discussed. The design of an i.f. amplifier using four *n-p-n* transistors is described as an example; the bandwidth is 14 kc/s centred at 455 kc/s, and the gain is 18 db per stage.

621.375.4 **2057**

Transistor Equations using h-Parameters.—C. C. Cheng. (*Electronics*, April 1954, Vol. 27, No. 4, pp. 191–192, 194.) Transistor circuit calculations are simplified by making use of fundamental parameters for the four-terminal network corresponding to the base-input common-emitter circuit. Circuit equations in terms of these parameters are tabulated for the three basic configurations.

621.375.4 **2058**

Practical Two-Stage Transistor Amplifiers.—R. L. Riddle. (*Electronics*, April 1954, Vol. 27, No. 4, pp. 169–171.) The effects of feedback and choice of coupling circuits are discussed. Experimentally determined operating characteristics are presented for the three most useful arrangements, (*a*) grounded-emitter to grounded-emitter, (*b*) grounded-base to grounded-emitter, and (*c*) grounded-collector to grounded-emitter, using commercially available junction transistors.

621.375.4 **2059**

Design of Transistor Power Amplifiers.—S. K. Ghandhi. (*Electronics*, March 1954, Vol. 27, No. 3, pp. 146–149.) Methods are outlined for obtaining maximum power output consistent with acceptable distortion. Factors considered include stage gain and power supply. Using values obtainable with the GE2N34 *p-n-p* transistor, two numerical examples are worked out, (*a*) a class-A push-pull grounded-base arrangement for a 100-mW low distortion a.f. amplifier, and (*b*) a class-B grounded-emitter arrangement for a 300-mW output stage.

621.375.4.024 **2060**

Temperature-Compensated D.C. Transistor Amplifier.—E. Keonjian. (*Proc. Inst. Radio Engrs*, April 1954, Vol. 42, No. 4, pp. 661–671.) The variation of transistor parameters with temperature is discussed. Compensation can be provided by using temperature-sensitive resistors. A description is given of an experimental d.c. amplifier compensated in this way. See also 1367 of May.

621.376.5 : 621.385.029.6 **2061**

Pulse Formation and Supply for Magnetrons.—H. G. Bruijning. (*Tijdschr. ned. Radiogenoot.*, Jan. 1954, Vol. 19, No. 1, pp. 11–23.) Discussion of the design of a modulator capable of handling pulses of the order of 30 A at 30 kV with a duration of 1 μs and a repetition rate of 1 000 per sec. The method of pulse formation used is based on the slow charging and rapid discharging of a capacitor. The use of artificial lines for pulse shaping is described. If a pulse transformer is used, the charging voltage need not be unduly high.

621.396.6 **2062**

Trends of Development in the Field of Electrical Components for Telecommunication.—P. Henninger. (*Frequenz*, Dec. 1953, Vol. 7, No. 12, pp. 345–359, & Jan. 1954, Vol. 8, No. 1, pp. 7–17.) Component design is considered from the point of view of the physics and chemistry involved in meeting the requirements of (*a*) operation over a particular frequency band, (*b*) operation under conditions of prescribed current waveform and intensity, (*c*) operation under peak loads, (*d*) operation at low loss, (*e*) operation under given ambient conditions, (*f*) operation under restrictions of tolerance, (*g*) stability and durability.

621.396.6 : 061.4 **2063**

Components Exhibition. Trends in Developments Portrayed at the R.E.C.M.F. Show.—(*Wireless World*, May 1954, Vol. 60, No. 5, pp. 206–210.) A review of the exhibition held in London, April 1954.

621.396.6.002.2 **2064**

Printed-Circuit Design Sources.—(*Elect. Mfg*, Dec. 1953, Vol. 52, No. 6, pp. 129–132..316.) A detailed guide to commercial design and manufacturing sources available in the U.S.A.

621.396.6.002.2.001.4 **2065**

Standardization of Printed Circuit Materials.—W. Hannahs, J. Caffiauz & N. Stein. (*Tele-Tech*, Feb. 1954, Vol. 13, No. 2, pp. 68–70..155.) Results of tests on metal-clad plastic laminates for (*a*) thermal endurance under simulated manufacturing conditions, (*b*) adhesion variation between the centre and the edge of the sheet, (*c*) bond strength during solder dipping and hand soldering, (*d*) effects of dust, are shown graphically. These and other types of bond-strength tests are discussed from the point of view of establishing quality standards and standard conditions of test. The positioning of components to avoid bond stress, and the proper choice of operating temperatures are also considered.

GENERAL PHYSICS

537.122 **2066**

A Theory of the Electron.—H. T. Flint & E. M. Williamson. (*Nuovo Cim.*, 1st Feb. 1954, Vol. 11, No. 2, pp. 188–199. In English.) An essentially classical theory is proposed by analogy with the gravitational theory of matter; it is based on equations of the form suggested by Mie in 1912. The mass of the electron is accounted for by means of the energy of the field.

537.311.5 : 621.3.015.3 **2067**

Diffusion of Pulsed Currents in Conductors.—L. M. Vallese. (*J. appl. Phys.*, Feb. 1954, Vol. 25, No. 2, pp. 225–228.) Theoretical investigation of the density distribution and the equivalent depth of penetration of a transient current, for the case of a plane TEM wave incident normally upon a plane conductor. Particular pulse waveforms are considered. See also 1900 of 1950.

537.52 **2068**

The Decay of the Space Charges in Intermittent Discharges in Neon and Argon.—D. Brini & P. Veronesi. (*Nuovo Cim.*, 1st Dec. 1953, Vol. 10, No. 12, pp. 1662–1672. In English.) The influence of the gas pressure and electrode separation on the decay of the space charges is investigated experimentally. The decay times for both neon and argon appear to be of the order of 10^{-3} sec, and are nearly proportional to the pressure over the range of measurement (about 1–40 Torr). Possible explanations of these values are discussed, and the various regions of the static characteristic are related to the functioning of counters and periodic discharges.

537.523 **2069**

The High-Pressure Glow Discharge in Air.—W. A. Gambling & H. Edels. (*Brit. J. appl. Phys.*, Jan. 1954, Vol. 5, No. 1, pp. 36–39.) Report of observations on the glow discharge in air at a pressure of about 760 mm Hg, between Cu and W electrodes; characteristics are given for discharge lengths of 0–8 mm and currents of 0·01–0·5 A.

537.525.5 : 621.396.822 **2070**

Spectrum Intensities and Radio Frequency Noise in a D.C. Hydrogen Arc.—S. E. Williams & V. Maslen. (*Nature, Lond.*, 20th Feb. 1954, Vol. 173, No. 4399, pp. 361–362.) R.f. noise generated by the arc was reduced by shunting the arc with a capacitor; this caused an incidental modification of the ultraviolet spectrum. The results suggest a connection between the noise intensity and the distribution of energy among the electrons.

537.533 : 546.74 **2071**

Investigation of the Electron Emission from Nickel.—G. V. Spivak & A. Gel'berg. (*C. R. Acad. Sci. U.R.S.S.*, 21st Jan. 1954, Vol. 94, No. 3, pp. 455–458. In Russian.) Experiments show that the work function of a single crystal of Ni, reduced in hydrogen, is higher above the Curie point than below it. This result is shown graphically. Field strengths of about 6×10^{7} V/cm were used in measurements of electron emission from a point source. Photographs of the emission patterns are shown.

537.56 **2072**

A Note on the Formula for the Mobility of Electrons with Mean Free Path varying with Velocity.—P. M. Davidson. (*Proc. phys. Soc.*, 1st Feb. 1954, Vol. 67, No. 410B, pp. 159–161. Correction, *ibid.*, 1st March 1954, Vol. 67, No. 411B, p. 279.) A correct generalized formula is derived for the electronic drift velocity in an ionized gas, taking the Townsend-Ramsauer effect into account.

537.582 **2073**

Theory of the Work Function of Metals.—W. Oldekop & F. Sauter. (*Z. Phys.*, 25th Jan. 1954, Vol. 136, No. 5, pp. 534–546.) The polarizing action of individual electrons on the distribution of the remaining electrons in the conduction band is investigated by means of extended Thomas-Fermi statistics and the effect on the magnitude of the work function is calculated. In alkali metals the work function is determined largely by this polarization effect (image force) and only slightly by the electrostatic double-layer at the metal boundary.

538.248 **2074**

Fluctuation Magnetic After-Effect.—J. C. Barbier. (*Ann. Phys., Paris*, Jan./Feb. 1954, Vol. 9, pp. 84–140.) Thesis, based on Néel's theory, on the irreversible after-effect in ferromagnetic materials which is identified with the residual loss. Earlier reports of the work were noted in 2237 of 1952 and 1661 of 1950, in which it should have been stated that the square root of the remanent magnetization is proportional to the logarithm of time elapsed from the suppression of the field and to the logarithm of the duration of application.

538.3 **2075**

Self-Consistent Electrodynamics.—O. Buneman. (*Proc. Camb. phil. Soc.*, Jan. 1954, Vol. 50, Part 1, pp. 77–97.) "The idea of direct action between streams is applied to a continuous charged fluid and combined with the new formulation of the electrodynamical laws of motion in terms of conservation of circulation. A simple and rigorous integrated formulation is thus obtained from the Maxwell-Lorentz differential equations, applicable to co-existing positive and negative fluids, as well as vacuum. Exact solutions are obtained, among them one which represents self-consistent, self-maintained flow in a hollow tubular region of infinite axial extent. It is hoped this tube might be bent into a torus and that an electron model will result from merely quantizing the one or two vortices around which this flow-pattern circulates."

538.52 : 537.311.5 : 621.396.677.8 **2076**

On the Current induced in a Conducting Ribbon by a Current Filament Parallel to it.—E. B. Moullin. (*Proc. Instn elect. Engrs*, Part IV, Feb. 1954, Vol. 101, No. 6, pp. 7–17.) Full paper. See 704 of March.

538.56 : 535.422 **2077**

On the Complete Theory of Diffraction of Electric Waves by a Perfect Conducting Wedge.—Y. Nomura. (*Sci. Rep. Res. Inst. Tohoku Univ., Ser. B*, Dec. 1952, Vol. 4, No. 1, pp. 29–50.) Continuation of work abstracted in 374 of 1952. The analysis is revised and completed, and errata are corrected.

538.566 : 535.42 **2078**

Diffraction of Electromagnetic Waves by an Aperture in a Large Screen.—J. H. Crysdale. (*J. appl. Phys.*, Feb. 1954, Vol. 25, No. 2, pp. 269–270.) Bekefi's approximate formula (709 of March) is derived simply by another method.

538.569.4.029.65 : 535.33 **2079**

One-to-Two Millimeter Wave Spectroscopy: Part 4 — Experimental Methods and Results for OCS, CH_3F, and H_2O.—W. C. King & W. Gordy. (*Phys. Rev.*, 1st Feb. 1954, Vol. 93, No. 3, pp. 407–412.) Continuation of work reported previously (2623 of 1953). Multiplier and detector performance were greatly improved by using crystals small compared to the wavelength to be detected. A new tuning technique based on known absorption lines was developed. Measurements made included that of a new water-vapour line at 183·31130 ± 0·00030 kMc/s.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.2/.8 : 621.396.822] : 621.396.621 **2080**

A D.C. Comparison Radiometer.—Selove. (See 2205.)

523.746 **2081**

Revised Data for the Mean Sunspot Curve.—W. Gleissberg. (*Naturwissenschaften*, Feb. 1954, Vol. 41, No. 4, p. 82.) When the data for the present nearly concluded 11-year cycle are included with those for the seventeen preceding cycles, the mean curve exhibits a slightly steeper rise and a slightly shallower fall, thus increasing its asymmetry.

523.746 : 550.386 **2082**

Controls of Geomagnetic Activity by Sunspots.—U. Becker & J. F. Denisse. (*J. atmos. terr. Phys.*, March 1954, Vol. 5, No. 1, pp. 70–72.) Becker (*Z. Astrophys.*, 1953, Vol. 32, p. 195) showed that a decrease of magnetic activity occurs when two sunspots or groups of sunspots are symmetrically located on the solar disk with respect to the solar equator. On the basis of this work and that of Denisse (2642 of 1953), it should be possible to forecast each year, on an average, about 80 perturbed and 75 quiet days.

523.854 : 621.396.822 **2083**

Identification of the Most Powerful Discrete Sources of Radio Emission of the Galaxy with the Residua of Supernovae Exploded in the Last 2000 Years.—I. S. Shklovski. (*C. R. Acad. Sci. U.R.S.S.*, 21st Jan. 1954, Vol. 94, No. 3, pp. 417–420. In Russian.)

550.38 : 551.510.535 : 621.396.11 **2084**

Experimental Determination of the Total Intensity of the Terrestrial Magnetic Field in the Lower Region of the Upper Atmosphere (E Layer).—M. Cutolo. (*Nuovo Cim.*, 1st July 1953, Vol. 10, No. 7, pp. 915–925.) Fuller account of work described previously (720 of March).

551.510.535 **2085**

Plasma Theory of the Ionosphere.—I. Lucas & A. Schlüter. (*Arch. elekt. Übertragung*, Jan. 1954, Vol. 8, No. 1, pp. 27–30.) The basic equations for the dynamic and electrical behaviour of the ionosphere are derived from the theory of a neutral plasma, and are used to investigate the problems of diffusion of the plasma at layer boundaries, e.m. damping of air movements and tidal oscillations of the ionosphere.

551.510.535 **2086**

Electron Density in the Upper Atmosphere and Interpretation of the *h′f* Curves of Ionosphere Virtual Height: Part 2.—F. Mariani. (*Ann. Geofis.*, Oct. 1953, Vol. 6, No. 4, pp. 533–553.) A calculation is made of the optical path of a wave reflected by a Chapman layer with a semithickness 2·5 H (where H is the scale height). Comparison of the results with those obtained previously (3598 of 1953) indicates that the lower half of such a layer is closely approximated by a parabolic layer of semithickness 1·5 H. The electron density distribution is investigated for regions of superposition of the F_1 and true F_2 layers, taking recombination into account. The linear superposition formula used previously is found to be inadequate and is replaced by the formula $N = (N_1^2 + N_2^2)^{\frac{1}{2}}$, where N_1 and N_2 are the independent densities of the F_1 and the true F_2 layers.

551.510.535 **2087**

A Tentative Model of the Equilibrium Height Distribution of Nitric Oxide in the High Atmosphere and the Resulting D Layer.—A. P. Mitra. (*J. atmos. terr. Phys.*, March 1954, Vol. 5, No. 1, pp. 28–43.) Two possible processes for the production of NO are considered, (*a*) three-body recombination with O, and (*b*) photodissociation of N_2O into N and NO. The distribution of known constituents of the atmosphere in the 50–100-km range is discussed, and the NO height distribution derived for the two possible production processes. D-layer ionization characteristics are satisfactorily explained, assuming photoionization of NO at $\lambda < 1\,300$ Å, and the electron distribution is derived, both for the case of photochemical equilibrium of NO and for a nonequilibrium distribution under the condition of complete mixing.

551.510.535 **2088**

The Determination of the Electron Density Distribution of an Ionosphere Layer in the Presence of an External Magnetic Field.—J. M. Kelso. (*J. atmos. terr. Phys.*, March 1954, Vol. 5, No. 1, pp. 11–27.) "The electron density distribution is determined by finding the actual heights at which waves of various frequencies are reflected. These true heights are obtained from experimental *h′-f* curves by obtaining the exact solution, as a convergent series of integrals, of the integral equation giving the group height as a function of frequency. The errors arising in the numerical work are shown by applying the procedure to *h′-f* curves obtained theoretically from layers of known shape, and such errors are seen to be small compared with those made through use of other procedures."

551.510.535 **2089**

Regularities in the F Region of the Ionosphere.—B. Chatterjee. (*Nature, Lond.*, 6th Feb. 1954, Vol. 173, No. 4397, pp. 263–264.) Justification of the assumptions made in 407 of February is presented.

551.510.535 : 551.543 **2090**

Correlation between Variations of Surface Pressure and Ionospheric Parameters.—M. R. Kundu. (*Indian J. Phys.*, May 1953, Vol. 27, No. 5, pp. 235–243.) A statistical analysis was made of meteorological and ionospheric data obtained at Calcutta during the period 1948–1951. Correlation between variations of ionosphere parameters and of surface pressure was found at least for some months of the year. The results are in agreement with those obtained in Australia by Martyn & Pulley (*Proc. roy. Soc. A*, 1936, Vol. 154, p. 455). Previously proposed tentative explanations based on variations of ozone content or the effects of circulation in the troposphere are examined; further data are required in order to assess their correctness.

551.510.535 : 551.55 **2091**

Recent Advances in the Study of Ionospheric Winds.—L. A. Manning. (*Bull. Amer. met. Soc.*, Nov. 1953, Vol. 34, No. 9, pp. 401–405.) The radio-fading method developed by Mitra (96 of 1950) and the meteor-trail-echo method developed by Manning et al. (3052 of 1950) for investigating ionospheric winds are briefly reviewed, and results so far obtained are discussed.

551.510.535 : 621.396.11 **2092**

The Effect of Sunrise on the Reflection Height of Low Frequency Waves.—S. B. Brown & W. Petrie. (*Canad. J. Phys.*, Jan. 1954, Vol. 32, No. 1, pp. 90–98.) The observed sudden changes of phase and amplitude of 16-kc/s waves transmitted over distances of 540 km are discussed in relation to the geometry of the system at the change-over from night-time to daytime reflection height (95 km and 70 km respectively). The effect cannot be due to photoionization of atmospheric molecules, but may be due to the removal of electrons from negative oxygen ions by visible and near-infrared radiation; calculations are presented giving support to this view. The rate of fall of the reflection height is discussed.

551.594.13 : 551.508.11 **2093**

Measurement of the Electrical Conductivity in the Upper Air by Radiosonde.—S. P. Venkiteshwaran, B. K. Gupta & B. B. Huddar. (*Proc. Indian Acad. Sci. A*, Aug. 1953, Vol. 38, No. 2, pp. 109–115.) Description of a measurement technique using the Gerdien conductivity apparatus in conjunction with a valve electrometer and an a.f.-modulated American type radiosonde.

551.594.21 **2094**

Generation of Electricity in Thunderstorms.—B. J. Mason. (*Elect. Times*, 4th Feb. 1954, Vol. 125, No. 3248, pp. 159–163.) A general account of the origin and nature of lightning.

551.594.6 : 538.566 **2095**

The Higher-Order Modes in the Propagation of Long Electric Waves in the Earth-Air-Ionosphere System and Two Applications (Horizontal and Vertical Dipole).—Schumann. (See 2195.)

LOCATION AND AIDS TO NAVIGATION

621.396.93 : 621.396.663 **2096**

Investigation of the Interference Field of Electromagnetic Waves with a C.R. Direction Finder.—J. Pietzner. (*Fernmeldetech. Z.*, Feb. 1954, Vol. 7, No. 2, pp. 80–84.) Description of a system for determining the directions of two interfering transmitters operating on the same frequency. The Adcock- and vertical-aerial systems are used in conjunction with a goniometer so that the signals applied to the Y and X plates of the c.r.o. are the signals received effectively by aerials with cardioid and figure-of-eight polar diagrams respectively. The two goniometer settings at which the c.r.o. trace reduces to a straight line are ϕ_1 and ϕ_2–ϕ_1–180°, respectively, where ϕ_1 and ϕ_2 are the bearings for the two transmitters.

621.396.962.3 : 621.396.621 **2097**

Use of Superregenerative Receivers as Intermediate-Frequency Amplifiers for Pulsed Radar Reception.—S. Marmor. (*Onde élect.*, Jan. 1954, Vol. 34, No. 322, pp. 73–79.) Operating principles of a superregenerative receiver for pulse reception (481 of 1952) are outlined. Its performance at a wavelength of 3·2 cm is compared with that of a conventional superheterodyne receiver. Typical p.p.i. and Type-A displays in the two cases are compared.

621.396.962.3 : 621.396.822 **2098**

The Minimum Usable Signal in Radar Reception and its Improvement by Certain Correlation Techniques.—L. Gérardin. (*Onde élect.*, Jan. 1954, Vol. 34, No. 322, pp. 67–72.) The definition of minimum usable signal in terms of signal/noise ratio and probability of detection is discussed. Experimental results obtained at a wavelength of 10 cm, using a 12-in. c.r. tube with p.p.i. display, are in agreement with calculations reported by Ross (1653 of 1951). For a p.p.i. display, under normal conditions the minimum usable signal is the tangential signal, viz. a signal 8 db above r.m.s. noise power. Principles of design of linear and nonlinear filters effectively increasing signal/noise ratio are outlined.

621.396.962.38 **2099**

Secondary Surveillance Radar.—D. A. Levell. (*Wireless World*, May 1954, Vol. 60, No. 5, pp. 227–230.) An experimental system tried at London Airport uses an interrogation frequency of 1·215 kMc/s and a response frequency of 1·375 kMc/s. The interrogating signal comprises three 1-μs pulses, (*a*) a reference pulse actuating gating circuits in the airborne transponder, (*b*) a control pulse, and (*c*) the interrogator pulse proper. The separation between the leading edges of (*a*) and (*b*) is 5 μs and that between the leading edges of (*a*) and (*c*) is 16 μs. (*a*) and (*c*) are produced by the same oscillator and radiated from the same directional aerial, (*b*) is produced separately and radiated omnidirectionally. When the received control pulse exceeds the following interrogator pulse by more than 3 db the interrogator pulse is prevented from passing through the gate. Side-lobe responses are thus prevented.

621.396.963.325 **2100**

Circular Radar cuts Rain Clutter.—W. D. White. (*Electronics*, March 1954, Vol. 27, No. 3, pp. 158–180.) A 1·3-kMc/s radar unit was modified for use with circularly polarized radiation by covering the dish-shaped reflector with wire mesh. Improvements in target/precipitation ratio varying from 8 to 25 db were obtained. The adverse effect of ground reflections is noted. Details are given of the statistical method used in measurements on aircraft targets.

621.396.969.33 + 621.396.969.36 **2101**

A Survey of Five Years' Progress in Marine Radar.—F. J. Wylie. (*J. Inst. Nav.*, Jan. 1954, Vol. 7, No. 1, pp. 59–77.)

MATERIALS AND SUBSIDIARY TECHNIQUES

531.788.7 **2102**

Ultra-High Vacuum: Part 2 — Limiting Factors on the Attainment of Very Low Pressures.—D. Alpert & R. S. Buritz. (*J. appl. Phys.*, Feb. 1954, Vol. 25, No. 2, pp. 202–209.) Continuation of work reported in 3603 of 1953 (Alpert). Calibration of the Bayard-Alpert ionization gauge shows that its characteristic is linear over most of its useful range, 5×10^{-10}–10^{-3} mm Hg. The achievement of very low pressures in glass systems is limited ultimately by diffusion of atmospheric He through the walls. A simplified omegatron [*Phys. Rev.*, 1st June 1951, Vol. 82, No. 5, pp. 697–702 (Sommer et al.)] has been developed for the measurement of partial pressures in a highly evacuated system.

535.215 : 546.817.221 **2103**

Bulk Photoconductivity in Lead Sulfide.—D. E. Soule & R. J. Cashman. (*Phys. Rev.*, 1st Feb. 1954, Vol. 93, No. 3, pp. 635–636.) Preliminary results of experiments at 77°K on three cleaved synthetic PbS crystals are reported. The crystal surface was scanned by a light spot and the resulting photocurrent measured. Current maxima were found at points on the crystal at which internal potential barriers were known to exist.

535.215.3 : 538.639 : 546.289 **2104**

Contribution to the Study of the Photomagnetoelectric Effect in Germanium.—H. Bulliard. (*Ann. Phys., Paris*, Jan./Feb. 1954, Vol. 9, pp. 52–83.) Detailed report of a theoretical and experimental investigation. See also 2015 and 2016 of 1953 (Aigrain & Bulliard). Values for the coefficient of volume recombination and the surface recombination velocity are calculated.

535.3 : [546.817.221 + 546.817.231 + 546.817.241 **2105**

Further Measurements on the Optical Properties of Lead Sulphide, Selenide and Telluride.—D. G. Avery. (*Proc. phys. Soc.*, 1st Jan. 1954, Vol. 67, No. 409B, pp. 2–8.) Continuation of earlier work (2018 of 1953). Measurements have been made at wavelengths up to 6 μ, using reflection techniques. Temperature effects on PbS and the effects of baking in an O_2 atmosphere are discussed.

535.34 : [537.311.31 + 537.311.33 **2106**

On the Theory of Optical Absorption in Metals and Semiconductors.—R. Wolfe. (*Proc. phys. Soc.*, 1st Jan. 1954, Vol. 67, No. 409A, pp. 74–84.) A quantum-mechanics method is described for calculating the effect on the optical absorption of any of the factors responsible for electrical resistance. The electrons scattered by imperfections in the crystal lattice are considered to absorb light by a photoelectric process. The results calculated, using a first-order approximation, for the case where the conduction electrons are scattered by dissolved impurities are similar to those obtained using semi-classical theory in which the current set up by the light is assumed to be damped by the impurities. For semiconductors, the new method gives much lower absorption values than the semiclassical method. The use of exact wave functions for the very slow electrons in semiconductors would give greatly increased absorption coefficients.

535.37 **2107**

Emission Spectra of Multiply Activated Electroluminescent Materials.—G. Destriau. (*J. Phys. Radium*, Jan. 1954, Vol. 15, No. 1, pp. 13–15.) The spectra are composed of bands associated respectively with the different activators. The electroluminescence spectra may be very different from the fluorescence spectra, and are greatly influenced by the frequency and intensity of the exciting field.

535.37 **2108**

Periodic Variations of Brightness in Electroluminescence and in Luminous Surface Effects.—G. Curie & D. Curie. (*J. Phys. Radium*, Jan. 1954, Vol. 15, No. 1, pp. 61–62.) The observed phase shift of electroluminescence brightness variations with respect to the exciting field is interpreted as supporting the view that electroluminescence is essentially a volume rather than a surface effect.

535.37 **2109**

Temperature Dependence of the Electroluminescence of ZnS and ZnO Phosphors.—H. Gobrecht, D. Hahn & H. E. Gumlich. (*Z. Phys.*, 25th Jan. 1954, Vol. 136, No. 5, pp. 623–630.) Further investigations (see 2110 below) revealed an electrothermoluminescence effect in variously activated ZnS and ZnO phosphors, the luminescence/temperature curve exhibiting maxima at a number of points when the phosphor is slowly heated from $-$ 100°C to room temperature. The positions of the maxima are independent of field strength and frequency. On cooling, luminescence increases monotonically over a large part or the whole of the temperature range.

535.37 : 537.228 **2110**

Electroluminescence of Various Phosphors and its Dependence on the Strength and Frequency of the Alternating Electric Field.—H. Gobrecht, D. Hahn & H. E. Gumlich. (*Z. Phys.*, 25th Jan. 1954, Vol. 136, No. 5, pp. 612–622.) Electroluminescence was produced in phosphors at $-$ 100°C by using alternating fields of strengths between 10^4 and 10^5 V/cm and frequencies between 100 c/s and 10 kc/s. Results show that Zn and Cd sulphides and ZnO are excited more easily than other materials, that oxidizing pre-treatment of the sulphides reduces their electroluminescence field-strength threshold, and that luminescence can be excited in other materials by a glow-discharge caused by the field. Several criteria for distinguishing true electroluminescence from this latter case are given. The intensity of electroluminescence increases with field strength and frequency. Possible causes of these phenomena are discussed.

535.37 : 537.533.8 **2111**

The Effect of Organic Vapor on the Secondary Emission of Phosphors.—P. H. Dowling & J. R. Sewell. (*J. appl. Phys.*, Feb. 1954, Vol. 25, No. 2, pp. 228–230.) Experiments show that surface contamination causes a rapid reduction of the secondary emission ratio. If the bombardment voltage is sufficiently high this results in a progressive decrease in the surface potential and a corresponding decrease in apparent fluorescence efficiency. The relation between this effect and true 'phosphor burn' is discussed. Organic-vapour contamination may accelerate the development of 'cross burn' in c.r. tubes.

535.376 **2112**

Characteristics of Radioluminescence in Crystals.—G. T. Wright & G. F. J. Garlick. (*Brit. J. appl. Phys.*, Jan. 1954, Vol. 5, No. 1, pp. 13–18.) Report of an experimental investigation of the variation of light output with particle energy for single crystals of organic and inorganic phosphors excited by α particles.

535.376 **2113**

A Luminescent Screen for Use with Very-Low-Velocity Electrons.—S. F. Kaisel & C. B. Clark. (*J. opt. Soc. Amer.*, Feb. 1954, Vol. 44, No. 2, pp. 134–135.) Screens responding to excitation by electrons with energy as low as 3 eV are prepared by settling a phosphor designated as 'hex ZnO: (Zn)' on to a glass plate previously coated with a transparent conducting film.

535.376 **2114**

Excitation of Zinc Oxide Phosphors by Low-Energy Electrons.—R. E. Shrader & S. F. Kaisel. (*J. opt. Soc. Amer.*, Feb. 1954, Vol. 44, No. 2, pp. 135–139.) An experimental investigation was made for cases where the energy of the exciting electrons was not greater than that of the observed photons. The results suggest that any electron accepted by the crystal lattice is capable of producing luminescence in ZnO phosphors, no matter how low the bombarding voltage.

537.227 **2115**

Phase Transitions in Ferroelectric $KNbO_3$.—G. Shirane, H. Danner, A. Pavlovic & R. Pepinsky. (*Phys. Rev.*, 15th Feb. 1954, Vol. 93, No. 4, pp. 672–673.) A third phase transition has been observed at $-$ 10°C on heating and $-$ 55°C on cooling, a change from orthorhombic to rhombohedral structure being accompanied by an abrupt change of dielectric constant.

537.227 : 546.431.824-31 **2116**

Effect of a Two-Dimensional Pressure on the Curie Point of Barium Titanate.—P. W. Forsbergh, Jr. (*Phys. Rev.*, 15th Feb. 1954, Vol. 93, No. 4, pp. 686–692.) A disk-shaped single-crystal specimen of $BaTiO_3$ was subjected to pressure on its edges and not on its faces. The transition temperature increased with the two-dimensional pressure. This result is discussed in relation to previous experiments in which hydrostatic pressure was applied.

537.311.31 : 548.0 **2117**

Lattice Defects and the Electrical Resistivity of Metals.—T. Broom. (*Advances Phys.*, Jan. 1954, Vol. 3, No. 9, pp. 26–83.) Calculations of the effect of lattice defects on resistivity are summarized and the results are discussed in relation to observations made by quenching, irradiation and deformation experiments.

537.311.31 : 669-124.2 **2118**

The Effect of Cold-Work on the Electrical Resistivity of Alloys and the Law of Recovery.—J. O. Linde. (*Appl. sci. Res.*, 1954, Vol. B4, Nos. 1/2, pp. 73–86.) Measurements are reported on alloys with Cu, Ag or Au as matrix component. The resistivity changes resulting from cold working differ greatly for the various alloys. Au-Cr and Au-Fe alloys exhibit a decrease of resistivity. Recovery is studied for various annealing temperatures. A law is established expressing the resistivity change as a function of temperature and time. The results are discussed in relation to theory.

537.311.33 **2119**

Theory of the Differential Thermoelectric Power of Semiconductors.—G. Lautz. (*Z. Naturf.*, June 1953, Vol. 8a, No. 6, pp. 361–371.) Two formulae are derived for determining the Fermi level, the one applicable at the low temperatures associated with impurity semiconduction and the other applicable at the medium and high temperatures associated with the transition from impurity to intrinsic conditions and with purely intrinsic semiconduction. Expressions for the thermoelectric power are derived which are valid over a wide temperature range. These can be used to determine activation energy, width of energy gap and apparent mass of charge carriers. The limits of validity of the approximate formulae are demonstrated by comparison with exact solutions for two examples. Some experimental results on the temperature dependence of the thermoelectric power are discussed, showing qualitative agreement with the theoretical results.

537.311.33 **2120**

On the Existence of Hertzian Absorption Bands in Monatomic Semiconductors (Boron, Selenium).—J. Meinnel. (*J. Phys. Radium*, Feb. 1954, Vol. 15, No. 2, pp. 124–125.) Preliminary results of measurements at temperatures in the range 88°–300°K and at frequencies from 100 c/s to 400 kc/s show that one or more absorption bands exist.

537.311.33 : 546.24 : 535.323 **2121**

Infrared Index of Refraction of Tellurium Crystals.—P. A. Hartig & J. J. Loferski. (*J. opt. Soc. Amer.*, Jan. 1954, Vol. 44, No. 1, pp. 17–18.)

537.311.33 : 546.24-1 : 535.343 **2122**

Infrared Optical Properties of Single Crystals of Tellurium.—J. J. Loferski. (*Phys. Rev.*, 15th Feb. 1954, Vol. 93, No. 4, pp. 707–716.) The absorption and photoconductivity were investigated experimentally, the effects of temperature variation, crystal anisotropy and Se additions being studied. The variation of energy gap with temperature is -2×10^{-5} eV/°C. The photoconductivity is barely detectable at room temperature but is considerably enhanced at 90°K. Results obtained with two SeTe alloys indicate that the reduction of the lattice parameter is accompanied by an increase of the energy gap.

537.311.33 : 546.289 **2123**

Orientation Relationships in Cast Germanium.—W. C. Ellis & J. Fageant. (*J. Metals, N.Y.*, Feb. 1954, Vol. 6, No. 2, pp. 291–294.)

537.311.33 : 546.289 **2124**

Dislocations in Plastically Deformed Germanium.—G. L. Pearson. W. T. Read, Jr, & F. J. Morin. (*Phys. Rev.*, 15th Feb. 1954, Vol. 93, No. 4, pp. 666–667.) Hall-effect, conductivity and lifetime measurements were made on rods of *n*-type and of *p*-type Ge which had been bent while heated to about 650°C, and on control specimens. The results are consistent with the hypothesis that edge dislocations are associated with acceptor levels in the middle or upper half of the energy gap.

537.311.33 : 546.289 **2125**

Redistribution of Solutes by Formation and Solidification of a Molten Zone.—W. G. Pfann. (*J. Metals, N.Y.*, Feb. 1954, Vol. 6, No. 2, pp. 294–297.) Description of the production of step or graded *p-n* junctions in semiconductors by melting and re-solidifying a zone of a homogeneous ingot containing suitable concentrations of a donor and an acceptor in solid solution.

537.311.33 : 546.289 **2126**

Some Electrical Properties of Germanium Crystals containing Compensated Impurities.—V. Ozarow. (*Phys. Rev.*, 1st Feb. 1954, Vol. 93, No. 3, pp. 371–372.) Three compensated *n*-type Ge crystals containing both radioantimony and radioindium in varying amounts, and one uncompensated crystal doped with Sb only were prepared. Measurements of Hall constant and resistivity were made in the temperature range 78°–393°K. Marked differences in the characteristics of compensated and uncompensated crystals of comparable resistivities were noted.

537.311.33 : 546.289 **2127**

Electrical Properties of N-Type Germanium.—P. P. Debye & E. M. Conwell. (*Phys. Rev.*, 15th Feb. 1954, Vol. 93, No. 4, pp. 693–706.) Measurements of the conductivity and Hall effect over the temperature range 11°–300°K were made on specimens with various controlled amounts of added arsenic, the room-temperature resistivities ranging from 43 to 0·005Ω.cm. The results are used as the basis of a comprehensive review of semiconductor theory. The evidence supports the view that the constant-energy surfaces are not spherical, though results based on this simplifying assumption agree in many respects with the experimental results.

537.311.33 : 546.289 **2128**

Self-Diffusion in Germanium.—H. Letaw, Jr, L. M. Slifkin & W. M. Portnoy. (*Phys. Rev.*, 15th Feb. 1954, Vol. 93, No. 4, pp. 892–893.) Diffusion in thin slabs of Ge with a plating of Ge^{71} was determined by measuring the concentration of Ge^{71} in cuts of known mass and thickness. The results are discussed in relation to the vacancy-motion theory of diffusion.

537.311.33 : 546.289 **2129**

Measurement of Minority Carrier Lifetime and Contact Injection Ratio on Transistor Materials.—A. Many. (*Proc. phys. Soc.*, 1st Jan. 1954, Vol. 67, No. 409B, pp. 9–17.) Lifetime is determined by observing the decay of resistance of a filament of the material during an injecting pulse. By using a bridge circuit incorporating a *RC* network electrically analogous to the filament, the lifetime can be read directly on a calibrated dial. The measurement range extends down to 1μs, with an accuracy usually to within about 5%. Measurements on *n*-type Ge with soldered contacts are reported; the injection ratio is proportional to the current through the contact.

537.311.33 : 546.289 **2130**

The Temperature Dependence of the Drift Mobility of Injected Holes in Germanium.—R. Lawrance. (*Proc. phys. Soc.*, 1st Jan. 1954, Vol. 67, No. 409B, pp. 18–27.) Report of an experimental investigation of the effect of trapping in *n*-type Ge, and of the nature of the traps. The results have been reported previously (2332 of 1953).

537.311.33 : 546.289 : 535.323 **2131**

The Index of Refraction of Germanium measured by an Interference Method.—D. H. Rank, H. E. Bennett & D. C. Cronemeyer. (*J. opt. Soc. Amer.*, Jan. 1954, Vol. 44, No. 1, pp. 13–16.) Measurements in the wavelength range 2·0–2·4 μ are reported.

537.311.33 : 546.289 : 537.533.8 **2132**

Secondary Electron Emission from Germanium.—J. R. Johnson & K. G. McKay. (*Phys. Rev.*, 15th Feb. 1954, Vol. 93, No. 4, pp. 668–672.) Measurements were made on single crystals with *p-n* junctions. The secondary-emission yield δ exhibited a maximum value of about 1·15 at room temperature at a primary voltage of about 500 V. δ has a small negative temperature coefficient but is independent of donor or acceptor concentrations up to 10^{19}/cm³. No effects due to space-charge fields under the surface were observed in the case of Ge. The secondary-emission process in semiconductors generally is compared with that in metals and insulators.

537.311.33 : 546.682.86 **2133**

Radiation Effects in Indium Antimonide.—J. W. Cleland & J. H. Crawford, Jr. (*Phys. Rev.*, 15th Feb. 1954, Vol. 93, No. 4, pp. 894–895.) Polycrystalline specimens of both *n*-type and *p*-type InSb were subjected to neutron irradiation. The results of conductivity and Hall-coefficient measurements indicate that (*a*) donor impurities are introduced by transmutations in the expected manner, and (*b*) lattice defects produced by fast neutrons act as electron traps in *n*-type material.

The evidence is insufficient to indicate whether these defects behave predominantly as acceptors or hole traps in p-type material.

537.311.33 : 546.682.86 **2134**

Anomalous Optical Absorption Limit in InSb.—E. Burstein. (*Phys. Rev.*, 1st Feb. 1954, Vol. 93, No. 3, pp. 632–633.) Experimental results in agreement with those of Tanenbaum & Briggs (1098 of April) were obtained. An alternative theoretical explanation is offered, based on the very small effective mass of the electrons in InSb. This is associated with small effective density of states and with a small degeneracy concentration. InSb therefore becomes degenerate at relatively low electron densities.

537.311.33 : 546.811-17 : 538.632 **2135**

Electronic Conduction in Grey Tin.—J. T. Kendall. (*Phil. Mag.*, Feb. 1954, Vol. 45, No. 361, pp. 141–157.) The conductivity and Hall constant of grey tin containing antimony and/or gallium were measured over the temperature range 77°–286°K, and the number of charge carriers and their mobility were calculated from the results. A comparison is made with the corresponding properties of Si and Ge. Results are tabulated and shown graphically.

537.311.33 : 546.817.221 **2136**

Antimony Content and Semiconductor Properties of Synthetic Lead Sulphide Photoconductive Elements.—H. G. Smolczyk. (*Naturwissenschaften*, Feb. 1954, Vol. 41, No. 4, p. 84.) A preliminary report of experimental work. Measurements were made of the internal photoelectric effect and of the thermoelectric effect for specimens containing differents amounts of Sb and prepared under controlled pressure of sulphur vapour. The possibility of producing PbS p-n junctions is indicated.

537.311.33 : 548.0 : 546.817.221 **2137**

A Variation Principle for Electronic Wave Functions in Crystals.—D. P. Jenkins & L. Pincherle. (*Phil. Mag.*, Jan. 1954, Vol. 45, No. 360, pp. 93–99.) A variation principle giving the energy levels of electrons in polyatomic lattices is presented. The method is used to redetermine some of the energy levels of PbS; the results are in agreement with those obtained by other methods [2037 of 1953 (Bell et al.)].

537.311.33 : 548.55 **2138**

Preparation of Single Crystals of Semiconductor Compounds of Type A^{III} B^{V}.—R. Gremmelmaier & O. Madelung. (*Z. Naturf.*, May 1953, Vol. 8a, No. 5, pp. 333, 304A.) A method of pulling from the melt is used similar to that described by Teal et al. (1682 of 1951). Photographs of a single-crystal specimen of InSb and a polycrystalline specimen of AlSb produced by the method are shown.

537.311.33 : 621.396.822 **2139**

Some Notes on Gisolf's Theory of Electron Fluctuation Phenomena in Semiconductors.—K. W. Böer. (*Ann. Phys., Lpz.*, 5th Jan. 1954, Vol. 14, Nos. 1/2, pp. 87–96.) Gisolf's formulae (667 of 1950) are evaluated by introducing a statistical distribution of lifetimes for the conduction electrons. The effect of field distortion in the barrier layer is taken into account. The magnitude of the effect is compared with that of the Nyquist noise.

538.22 **2140**

Effects of Band Shape on the Magnetic and Thermal Properties of Metals and Alloys.—E. W. Elcock, P. Rhodes & A. Teviotdale. (*Proc. roy. Soc. A*, 7th Jan. 1954, Vol. 221, No. 1144, pp. 53–77.)

538.221 **2141**

Notes on the Theory of the Magnetic Properties of Hard Materials.—L. Néel. (*Appl. sci. Res.*, 1954, Vol. B4, Nos. 1/2, pp. 13–24. In French.) The theory of magnetic hysteresis for single-domain small grains is considered. By taking into account (*a*) the dispersion of the values of the coercive force for the individual grains and (*b*) the interaction force between the grains and the dispersion of the values of this force, numerical results are obtained which agree well with experimental results for good permanent magnets.

538.221 : 532.111 **2142**

The Change of Ferromagnetic Curie Points with Hydrostatic Pressure.—L. Patrick. (*Phys. Rev.*, 1st Feb. 1954, Vol. 93, No. 3, pp. 384–392.) Consistent results for eight materials measured in a liquid compression system were obtained. Results for five other materials measured in a gas compression system were less satisfactory. Neither the Bozorth nor the Néel theoretical interaction curve agrees with the experimental observations; this may be due to neglect of the part played by conduction electrons in the interaction.

538.221 : 532.111 **2143**

The Influence of Pressure on the Curie Temperature of Iron and Nickel.—R. Smoluchowski. (*Phys. Rev.*, 1st Feb. 1954, Vol. 93, No. 3, pp. 392–393.) A comparison is made between Patrick's measurements (2142 above) and theory based on a Brillouin function.

538.221 : [621.318.124 + 621.318.134 **2144**

Magnetic Resonance Phenomena in Ferrites.—F. Brown & D. Park. (*Phys. Rev.*, 1st Feb. 1954, Vol. 93, No. 3, pp. 381–384.) The resonance frequencies in the microwave and infrared regions are calculated, taking into account the effects of anisotropy, external field and differences in magnetization and g values between the two sublattices. The relation between the anisotropy fields in the expressions derived and the measured equivalent anisotropy field is discussed. Theoretical and experimental results for single-crystal Ni ferrite are in good agreement. See also 1828 of June (Wangsness).

538.221 : 621.318.134 **2145**

Microwave Resonance Absorption in Nickel Ferrite-Aluminate.—T. R. McGuire. (*Phys. Rev.*, 15th Feb. 1954, Vol. 93, No. 4, pp. 682–686.) Results are reported of experiments on materials having the composition $NiOAl_tFe_{2-t}O_3$, with particular attention to the composition for which the value of t is 0·7, at which value the magnetic moment is zero.

538.221 : 621.318.134 **2146**

Ferrites for Microwave Circuits and Digital Computers.—E. Albers-Shoenberg. (*J. appl. Phys.*, Feb. 1954, Vol. 25, No. 2, pp. 152–154.) Properties of two classes of commercially available Mg-Mn ferrites are discussed. One class is characterized by low losses and a Faraday rotation effect at microwave frequencies; the other by high resistivity and a rectangular hysteresis loop useful in magnetic storage systems operating at high speed.

538.221 : 621.318.134 : 681.142 **2147**

Ferromagnetic Spinels with Rectangular Hysteresis Loops.—Hegyi. (See 2007.)

538.221 : 681.142 **2148**

Penetration of an Electromagnetic Wave into a Ferromagnetic Material.—Papoulis. (See 2006.)

548.0 : 546.39.185-841 : 539.374 **2149**

The Plastic Deformation of Ammonium Dihydrogen Phosphate.—P. L. Smith & E. I. Salkovitz. (*J. appl.*

Phys., Feb. 1954, Vol. 25, No. 2, pp. 237–239.) Plastic deformation was produced in ADP single crystals by loading specimens as a simple beam at about 100°C. This deformation does not appreciably alter the piezoelectric or elastic properties.

549.514.51 : 621.372.412.002.2 **2150**

V.H.F. Crystal Grinding.—E. A. Gerber. (*Electronics*, March 1954, Vol. 27, No. 3, pp. 161–163.) Round crystals for frequency control in the range 20–180 Mc/s are finished by fastening the blanks to a work-holder when about 0·7 mm thick and reducing them to the required thickness on optical lapping machines. Crystals with low series resistance are obtained. Unwanted modes of vibration are reduced.

621.315.61 **2151**

New Nonrigid Materials for the Functional Design of Electrical Insulating Systems.—A. E. Javitz. (*Elect. Mfg*, Sept. 1953, Vol. 52, No. 3, pp. 123–138.) A survey of the properties and performance of recently developed types of insulation in the form of film, tape, sheet and paper.

621.315.612.029.6 **2152**

Materials and Problems of High-Frequency Ceramics.—J. Kainz. (*Elektrotech. u. Maschinenb.*, 1st Nov. & 1st Dec. 1953, Vol. 70, Nos. 21 & 23, pp. 473–478 & 525–530.) A survey paper. Composition and properties of various ceramic insulating materials are tabulated and shown in graphs. Investigations of the colloid systems by means of the electron microscope are described.

621.315.616 **2153**

Conductivity induced in Insulating Materials by X-rays.—J. F. Fowler & F. T. Farmer. (*Nature, Lond.*, 13th Feb. 1954, Vol. 173, No. 4398, pp. 317–318.) Measurements on polythene and perspex are reported; the results are of interest in connection with the determination of the energy-level distribution of the material.

621.791.342.6 : 546.682 **2154**

A Technique of Soldering to Thin Metal Films.—R. B. Belser. (*Rev. sci. Instrum.*, Feb. 1954, Vol. 25, No. 2, pp. 180–183.) By using In and certain of its alloys as a solder, without a flux, adherence of thin metal films to glass or quartz substrates may be obtained without damage to the film. The technique has been successfully applied in the mechanical suspension of mirrors, for making electrical contact with thin metal films and for mounting piezoelectric crystals. Soldered connections have been made to films of 18 metals including Al, Ti and Zr.

MATHEMATICS

514.1 **2155**

Some Formulae of P. Stein and Others concerning Trigonometrical Sums.—N. B. Slater. (*Proc. Camb. phil. Soc.*, Jan. 1954, Vol. 50, Part 1, pp. 33–39.) Formulae relevant to problems of alternating currents in cables are discussed.

517.43 **2156**

Approximations in Operational Methods.—J. Brodin. (*Ann. Télécommun.*, Jan. 1954, Vol. 9, No. 1, pp. 1–8.)

517.5 **2157**

Recurrence Relations for Prolate Spheroidal Wave Functions.—I. Marx. (*J. Math. Phys.*, Jan. 1954, Vol. 32, No. 4, pp. 269–275.)

517.6 **2158**

An Approximate Method of Evaluating Integral Transforms.—A. H. Zemanian. (*J. appl. Phys.*, Feb. 1954, Vol. 25, No. 2, pp. 262–266.) The method developed is quite general and has been applied to Fourier, Laplace, Mellin and Hankel transforms. It may also be effective in evaluating numerically an integral with a highly oscillatory term in its integrand.

517.63 **2159**

On Inverting Laplace Transforms of the Form $h(s)/p(s) + q(s)e^{-\tau s}$.—T. E. Hull & W. A. Wolfe. (*Canad. J. Phys.*, Jan. 1954, Vol. 32, No. 1, pp. 72–80.)

517.9 **2160**

A Sufficient Condition for an Infinite Discrete Spectrum.—C. R. Putnam. (*Quart. appl. Math.*, Jan. 1954, Vol. 11, No. 4, pp. 484–487.) A study of the problem of obtaining a sufficient criterion in order that the equation $x'' + f(t)x = 0$ be oscillatory, for the particular case that $f(t)$ satisfies the limit relation $f(t) \rightarrow 0$ as $t \rightarrow \infty$.

517.9 **2161**

On the Gaps in the Spectrum of the Hill Equation.—C. R. Putnam. (*Quart. appl. Math.*, Jan. 1954, Vol. 11, No. 4, pp. 496–498.)

517.93 **2162**

Geometrical Integration of Nonlinear Second-Order Differential Equations with Second Member.—G. Cahen. (*Bull. Soc. franç. Élect.*, Jan. 1954, Vol. 4, No. 37, pp. 44–50.) A study is made of equations of the type

$$x + b(x)\dot{x} + r(x) = g(t),$$

and is extended to equations containing a further term $a(x)\dot{x}^2$ or $c(\dot{x})$.

517.942.82 : 621.3.015.3 **2163**

The Calculation of Transients in Dynamical Systems.—E. E. Ward. (*Proc. Camb. phil. Soc.*, Jan. 1954, Vol. 50, Part 1, pp. 49–59.) The calculation of transients by Tricomi's method, using Laguerre functions, is a practical alternative to the use of partial fractions. This is shown by considering numerical examples of Laplace transforms ranging from quadratic to sixth-power polynomials. The composition of the coefficients of the Laguerre series is analysed, and the conditions for rapid convergence are indicated.

519 : 53 **2164**

A Method of Solving Very Large Physical Systems in Easy Stages.—G. Kron. (*Proc. Inst. Radio Engrs*, April 1954, Vol. 42, No. 4, pp. 680–686.) By using the subdivision method of dealing with large systems (202 of January), the amount of calculation required to obtain the solution can be reduced approximately in the ratio $2/n^2$, where n is the number of subdivisions. The method is illustrated by solving the two-dimensional Maxwell field equations by subdividing their electric-circuit models.

MEASUREMENTS AND TEST GEAR

529.7 **2165**

The Determination of Time and Frequency.—H. M. Smith. (*Proc. Instn elect. Engrs*, Part II, Feb. 1954, Vol. 101, No. 79, pp. 64–65.) Discussion on 2223 of 1951.

621.317.3 : 621.314.632 : 546.289 **2166**

Use of the Germanium Rectifier for the Measurement of Current, Voltage and Power at High Frequency: Part 2 — Power Measurement.—J. Schiele. (*Arch. tech. Messen*, Jan. 1954, No. 216, pp. 21–22.) Two examples are illustrated. Part 1: 1498 of May.

621.317.3 : 621.372.56.029.63/.64 **2167**

Mismatch Errors in the Measurement of Ultrahigh-Frequency and Microwave Variable Attenuators.—R. W. Beatty. (*J. Res. nat. Bur. Stand.*, Jan. 1954, Vol. 52, No. 1, pp. 7–9.) Expressions for the mismatch error are derived by analysis and an example is given to show that the mismatch error in measuring the difference in attenuation between two attenuators is less than the sum of the mismatch errors obtained when measuring each attenuator individually.

621.317.3 : 621.396.722 **2168**

The B.B.C. Measurement and Technical Receiving Station at Tatsfield.—Griffiths. (See 2210.)

621.317.3.089.6 : 621.372.8 **2169**

The Calibration of the Slotted Section for Precision Microwave Measurements.—A. A. Oliner. (*Rev. sci. Instrum.*, Jan. 1954, Vol. 25, No. 1, pp. 13–20.) A calibration procedure is described in which compensation is made for the slight change in guide wavelength and characteristic impedance due to the slot, and for discontinuities at the end of the slot and at coupling elements, bead supports, etc. Practical instructions are given for constructing a calibration curve valid for purely reactive terminations, from which correction factors are derived for the position of the voltage node and the value of s.w.r. in dissipative structures.

621.317.326 : 621.314.626 **2170**

Pulse Measurement with Peak-Voltage Rectifier Circuit.—E. de Gruyter. (*Bull. schweiz. elektrotech. Ver.*, 6th Feb. 1954, Vol. 45, No. 3, pp. 61–70.) The suitability of direct-indicating peak voltmeters for measuring pulse trains is examined. For instruments using contact rectifiers, general error curves are given taking account of the instrument constants, the pulse width and the voltage reading, for different pulse shapes. Calibration of individual instruments is thus rendered unnecessary. An unknown pulse width can be determined from comparative measurements.

621.317.335.3.029.63 : 621.315.615 **2171**

Two New Methods of determining the Electrical Constants of Liquids in the Decimetre Waveband.—O. Huber. (*Z. angew. Phys.*, Jan. 1954, Vol. 6, No. 1, pp. 9–14.) The dielectric constant and the loss tangent are determined from measurements using a vertical coaxial line. In the first method, the voltage variations at a fixed probe are determined as a function of the depth of the liquid, the input termination being reflection-free. The second method depends on the change of resonance frequency of the short-circuited line when the liquid is introduced. See also 3046 of 1951.

621.317.336 : 621.372.8 **2172**

Measurement of Waveguide Impedance.—A. Cunliffe & D. P. Saville. (*Wireless Engr*, May 1954, Vol. 31, No. 5, pp. 115–118.) The two horizontal arms of a T junction are connected respectively to a waveguide section with a sliding short-circuiting termination and to the unknown impedance, while the vertical arm is connected to the source. Voltage is measured by means of a fixed probe in the arm connected to the unknown impedance, and the value of the latter is found by adjusting the short-circuit to produce resonance. Theory of the method is given. Results of experiments using a wavelength of 3·2 cm indicate that the accuracy compares favourably with that of the usual methods. The apparatus can be used over a wide frequency band.

621.317.34 : 621.315.212 **2173**

A Pulse Method for the Quantitative Determination of Nonuniformities in Wide-Band Cables.—L. Krügel. (*Fernmeldetech. Z.*, Jan. 1954, Vol. 7, No. 1, pp. 3–9.) A simple method suitable for investigating very small irregularities is described.

621.317.42 : 538.221 **2174**

The Effect of the Förster Probe on Measurements in the Vicinity of a Ferromagnetic Material.—F. Brandstaetter. (*Elektrotech. u. Maschinenb.*, 1st Nov. 1953, Vol. 70, No. 21, pp. 484–487.) An investigation of the influence of the macroscopic structure of the ferromagnetic material on the divergence between readings obtained with this instrument (1506 of May) and theoretically calculated values.

621.317.7 : 621.372.412 **2175**

Quartz Crystal Testing.—R. Rollin. (*Wireless World*, May 1954, Vol. 60, No. 5, pp. 220–223.) The principles described by Biggs & Wells (969 of 1946) are applied in equipment for evaluating the quality of crystals over the frequency range 50 kc/s–2 Mc/s. A two-valve oscillator with band-switching and input-capacitance switching is used in conjunction with a calibrated variable-impedance element which can be substituted for the crystal. The complete circuit is shown.

621.317.7.087 **2176**

Direct-Indicating Recording Instruments.—S. R. Gilford. (*Elect. Mfg.* Nov. & Dec. 1953, Vol. 52, Nos. 5 & 6, pp. 114–121 & 120–128.) A survey covering applications, functional requirements, design and operating principles.

621.317.723 **2177**

A Vibrating Needle Electrometer.—Y. L. Yousef & R. Kamel. (*J. sci. Instrum.*, Jan. 1954, Vol. 31, No. 1, pp. 13–15.) The e.s. force due to the charge under test is modulated so that instead of simply deflecting the suspended needle it produces a resonant vibration whose amplitude is proportional to the charge.

621.317.73/.74 : 621.315.212 **2178**

The Development of a Precision Termination for 0·375-inch Polythene-Disc-Insulated Coaxial Cable.—R. J. Cheetham, E. L. Mather & W. W. H. Clarke. (*Proc. Instn elect. Engrs*, Part IV, Feb. 1954, Vol. 101, No. 6, pp. 135–145.) The adjustment facilities necessary in a precision termination are determined from theoretical considerations and test requirements. Details are given of the development of a circuit suitable for use in pulse testing and in the determination of end impedances to within about 0·05 Ω. In bridge testing, similar accuracies in both real and imaginary parts may be obtained between 50 kc/s and 8 Mc/s.

621.317.755 **2179**

Gated Time Markers for C.R.O. Display.—P. Steinberg. (*Electronics*, March 1954, Vol. 27, No. 3, pp. 150–151.) The waveform under examination and a series of marker dots are presented in alternate horizontal sweeps of the c.r.o.; the position and brightness of the two traces can thus be adjusted separately. A brief description is given of the circuit. Reading accuracy is increased by providing a vernier scale of marker dots.

621.317.755 : 621.314.7.012.6 **2180**

Characteristic-Curve Tracer for Transistors.—H. Lennartz.)*Funk u. Ton*, Jan. 1954, Vol. 8, No. 1, pp. 25–29.) Description of equipment used in conjunction with a c.r.o.

621.317.755 : 621.385.832 **2181**

An Interesting Four-Beam Cathode-Ray Tube.—Wendling. (See 2267.)

621.317.791 **2182**

Universal Meter for Measuring Voltages at High Impedances, Micromicroamperes, and Insulation Resistance.—W. R. Clark, R. E. Watson & G. C. Mergner. (*Elect. Engng, N.Y.*, Jan. 1954, Vol. 73, No. 1, pp. 41–45.) Direct voltages up to 500 V, direct currents down to 10^{-12}A and insulation resistance up to 10^8 MΩ are measured by a null method with maximum full-scale-deflection errors of 1·5, 3·5 and 5% respectively. The meter circuit includes a chopper-type stabilized feed-back amplifier. To be published also in *Trans. Amer. Inst. elect. Engrs*, 1953, Vol. 72.

621.373.44 **2183**

Current-Step Waveform Generator.—V. A. Babits, S. R. Spengler & R. V. Morris. (*Electronics*, March 1954, Vol. 27, No. 3, pp. 164–167.) A circuit is described for producing stepped currents capable of providing stepped values of magnetic field such as are required, e.g. for rotating the plane of polarization in microwave experiments. The arrangement comprises voltage-step generator, pulse generator and output stage driving a coil of inductance 1·5–3 H.

621.387.001.4 **2184**

A Method of Testing Cold-Cathode Tubes.—D. L. Benson & D. H. Vogan. (*P.O. elect. Engrs' J.*, Jan. 1954, Vol. 46, Part 4, pp. 196–197.) In the usual methods of testing d.c. breakdown potential, using manual control, it is difficult to vary the potential across the gap at the optimum speed. An automatically operating circuit for this purpose is described.

621.396.6.002.2.001.4 **2185**

Standardization of Printed Circuit Materials.—Hannahs, Caffiauz & Stein. (See 2065.)

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

621.317.755 : 621.3.018.75 : 615.471 **2186**

A Time-Marker for Electrocardiography.—M. A. Bullen & L. A. Daynes. (*J. sci. Instrum.*, Jan. 1954, Vol. 31, No. 1, pp. 6–7.) A simple circuit using gas-filled counting tubes is described by means of which time-marking signals at four repetition rates between 5 and 100 per sec can be derived from a.c. mains or a suitable oscillator; the signals are superposed on the electrocardiograph displayed on the c.r.o. screen, no separate channel being required.

621.37 : 794 **2187**

Electronic Air-War Game simulates Missile Strikes.—L. I. Davis. (*Electronics*, April 1954, Vol. 27, No. 4, pp. 146–152.) Circuit arrangements are described for illustrating the main principles involved in air warfare. The two participants each have two electric potentials at their disposal, one of which controls the attack and the other the defence. The operating time scale is such as to enable logical decisions to be made regarding choice of targets, etc.

621.383.2 **2188**

Development of [electron-optical] **Image Converters and Image Intensifiers.**—F. Eckart. (*Ann. Phys., Lpz.*, 5th Jan. 1954, Vol. 14, Nos. 1/2, pp. 1–13.) Resolving power is estimated, particular designs, including two-stage types, are described, and applications are indicated.

621.384.611/.612 **2189**

A Synchrocyclotron with Fixed Operating Frequency.—F. Ollendorff. (*Elektrotech. u. Maschinenb.*, 1st Jan. 1954, Vol. 71, No. 1, pp. 10–15.)

621.384.611 **2190**

Cyclotron Oscillators and the Shifting Inter-Dee Ground Surface.—F. H. Schmidt & M. J. Jakobson. (*Rev. sci. Instrum.*, Feb. 1954, Vol. 25, No. 2, pp. 136–139.) The push-pull mode of oscillation of a two-dee cyclotron can be described in terms of a ground potential surface between the dees. Unbalance in dee voltages can be interpreted as a shift of the position of the ground surface. The electrical implications of a particular adjustment can by application of this concept be quickly evaluated by the cyclotron operator.

621.385.833 **2191**

Calculation of some Magnetic and Electric Fields with Cylindrical Symmetry.—Y. Axner. (*Appl. sci. Res.*, 1954, Vol. B4, Nos. 1/2, pp. 124–136.) Fields having cylindrical symmetry and also mirror symmetry with respect to a median plane are discussed; the field strength in the median plane is given. Solutions are presented in the form of power expansions; analytical solutions are obtained when the field strength in the median plane as a function of the radius is given by a polynomial.

621.385.833 **2192**

Investigation of the Mechanism of the Formation of the Image in an Electron Microscope.—I. G. Stoyanova & A. I. Frimer. (*C. R. Acad. Sci. U.R.S.S.*, 21st Jan. 1954, Vol. 94, No. 3, pp. 459–462. In Russian.) Experimental investigation of the dependence of the image contrast on the thickness of the specimen and on the electron energy, in dark-field and bright-field image presentations.

621.387.4 **2193**

The New Counters.—S. C. Curran. (*Sci. Progr.*, Jan. 1954, Vol. 42, No. 165, pp. 32–45.) A review of developments in particle counters, with 47 references.

621.387.424 **2194**

Velocity of Discharge Propagation in Self-Quenching Geiger-Müller Counters.—P. A. C. Mortier & J. F. Roose. (*Proc. phys. Soc.*, 1st Feb. 1954, Vol. 67, No. 410B, pp. 161–163.)

PROPAGATION OF WAVES

538.566 : 551.594.6 **2195**

The Higher-Order Modes in the Propagation of Long Electric Waves in the Earth-Air-Ionosphere System and Two Applications (Horizontal and Vertical Dipole.)—W. O. Schumann. (*Z. angew. Phys.* Jan. 1954, Vol. 6, No. 1, pp. 35–43.) Expressions for the propagation and the radiation intensities of higher-order modes of very long (>3 km) waves are derived and applied. For very long waves the Bessel functions can be approximated by exponential terms. The numerical illustrations include the case of lightning signals. See also 1544 of May.

538.566.3 **2196**

Propagation of Plane Electromagnetic Waves in a Homogeneous Plasma (Ionosphere).—R. Jancel & T. Kahan. (*J. Phys. Radium*, Jan. 1954, Vol. 15, No. 1, pp. 26–33.) Mathematical analysis based on magneto-ionic theory developed previously (1030 of April). Refractive index, double refraction, phase and group velocities, attenuation, polarization and critical frequencies are studied. The range of validity of classical formulae relating to ionospheric propagation is discussed.

621.396.11 **2197**

Diffraction of Plane Radio Waves by a Parabolic Cylinder.—S. O. Rice. (*Bell Syst. tech. J.*, March 1954, Vol. 33, No. 2, pp. 417–504.) Expressions are given for the diffraction field far behind, and the surface currents on, a parabolic cylinder. Approximate values for the field strength and current density are given for the case when the radius of curvature of the cylinder is large compared with λ. The method of analysis makes use of parabolic cylinder functions of large complex order. The results indicate that the knife-edge representation is valid even for gently rounded hills when the angle of diffraction is small, but that the formulae developed are applicable to calculation of the shadows cast by hills in microwave propagation when the angle of diffraction is so large that the knife-edge representation is invalid.

621.396.11 : 535.15 2198

Birefringence in Crystals and in the Ionosphere.—C. H. M. Turner. (*Canad. J. Phys.*, Jan. 1954, Vol. 32, No. 1, pp. 16–34.) Propagation of plane e.m. waves in the ionosphere is compared with that in an optically inactive crystal, as in the work of Lange-Hesse (2868 of 1952); damping due to collisions of electrons with other particles is taken into account. In the ionosphere the plane wave is always characterized by three components of the field vectors, one of which is linearly polarized along the direction of the uniform magnetic field while the other two are circularly polarized in opposite senses in the plane perpendicular to the magnetic field.

621.396.11 : 551.510.535 2199

Oblique Propagation of Radio Waves over a Curved Earth.—B. Chatterjee. (*Indian J. Phys.*, May 1953, Vol. 27, No. 5, pp. 257–268.) Booker's analysis of oblique propagation (*Phil. Trans. A*, 1938, Vol. 237, pp. 411–451) is extended to take account of the earth's curvature by introducing appropriate correction factors. The values obtained for group retardation, attenuation and lateral deviation in the refracting region are higher than those for the flat earth and are in better agreement with experimental results.

621.396.11.029.51 2200

The Ionospheric Propagation of Radio Waves of Frequency 30-65 kc/s over Short Distances.—R. N. Bracewell, J. Harwood & T. W. Straker. (*Proc. Instn elect. Engrs*, Part IV, Feb. 1954, Vol. 101, No. 6, pp. 154–162. Digest, *ibid.*, Part III, March 1954, Vol. 101, No. 70, pp. 108–110.) Experiments carried out at Cambridge on waves reflected from the ionosphere at steep incidence are described. The phases and amplitudes of two linearly polarized components of the downcoming wave were measured with reference to the ground wave. Results are compared with those obtained by Straker for 16-kc/s waves, and show that (*a*) the day-to-day variations of the downcoming wave were greater at the higher frequencies, (*b*) the change in height of reflection in passing from day to night was about the same for all the frequencies, (*c*) the amplitudes by day and night in summer were very different, except at 16 kc/s, and (*d*) the polarization at all the frequencies was approximately circular, left-handed and constant.

621.396.11.029.62 : 621.396.812 2201

U.S.W. Extended-Range [propagation] **and Inversions.**—H. Wisbar. (*Funk-Technik, Berlin*, Jan. 1954, Vol. 9, No. 1, pp. 8–10.) Reception and transmission over abnormally long distances by a North-West German amateur station operating in the 2-m band, in the period from March to October 1953, is correlated with meteorological conditions, particularly barometric pressure systems. Propagation distances up to 1 100 km were recorded. The effect of widespread fog is also noted.

621.396.11.029.62 : 621.397.5 2202

Wave Propagation and Television Broadcasting at Very High Frequencies.—Smith-Rose. (See 2234.)

621.396.81 + 621.396.65 2203

Microwave as Applied to Railroad Operation in the Gulf Coast Area.—Thomas. (See 2219.)

621.396.82.029.62 2204

Prediction of the Likelihood of Interference at Frequencies of 30 to 42 Megacycles in Alaska.—T. N. Gautier, Jr, & C. J. Sargent. (*J. Res. nat. Bur. Stand.*, Jan. 1954, Vol. 52, No. 1, pp. 21–31.) A full report is presented of an investigation of the likelihood of interference with the operation of a proposed v.h.f. line-of-sight network, either from ionospheric propagation of signals between stations of the network, or from stations outside the network. Diurnal, seasonal and sunspot-cycle variations are studied. The likelihood of interference is inferred from the likelihood of occurrence of m.u.f.s equal to or greater than the operation frequencies. Calculations were made for paths involving reflections from (*a*) the F_2 layer and (*b*) the sporadic-E layer.

RECEPTION

621.396.621 : [523.2/.8 : 621.396.822 2205

A D.C. Comparison Radiometer.—W. Selove. (*Rev. sci. Instrum.*, Feb. 1954, Vol. 25, No. 2, pp. 120–122.) Apparatus for measuring r.f. radiation from astronomical sources is described in which a comparison signal is transmitted through the receiving system along with the incoming signal, as a means of measuring the gain of the system and of cancelling the effect of gain fluctuations. The two signals are separated at the receiver output by a pair of filters. The rectified outputs of the two filters are compared, and the difference constitutes the final output signal. For a signal with bandwidth narrow compared with the receiving system bandwidth, the maximum improvement attainable over the switching-type comparison radiometer is 3·2 db.

621.396.621 : 621.314.7 2206

An Experimental Transistor Personal Broadcast Receiver.—L. E. Barton. (*Trans. Inst. Radio Engrs*, Jan. 1954, No. PGBTR-5, pp. 6–13.) Details are given of an a.m. receiver using 6 r.f. junction transistors of the type described in 1955 of June (Mueller & Pankove), three conventional junction transistors for class-B audio driver and output stages, and two diodes. Battery drain is <12 mA, and maximum output is 150 mW. Sensitivity and signal/noise ratio are comparable to those of conventional receivers.

621.396.621 : 621.372.2 2207

The History of the Homodyne and Synchrodyne.—D. G. Tucker. (*J. Brit. Instn Radio Engrs*, April 1954, Vol. 14, No. 4, pp. 143–154.) An account is given of the independent development of the synchronous demodulation system from Colebrook's homodyne and from Tucker's synchrodyne. 62 references.

621.396.621 : 621.396.822 2208

The Minimum Detectable Change in the Mean Noise-Input Power to a Radio Receiver.—D. G. Lampard. (*Proc. Instn elect. Engrs*, Part IV, Feb. 1954, Vol. 101, No. 6, pp. 118–128. Digest, *ibid.*, Part III, March 1954, Vol. 101, No. 70, pp. 111–113.) Expressions for the minimum detectable change involving only the pre-detector and the post-detector filter responses have been derived for receivers using either a power-law instantaneous detector (full- or half-wave type) or an ideal power-law envelope detector. Using these expressions and making appropriate approximations, an example, in which the pre-detector filter is a single tuned circuit and the post-detector filter is a simple *RC* integrator, is worked out in detail. The results, presented in simple form, show clearly the effect of the filter bandwidths and the detector law on the sensitivity of the receiver to changes in noise input power.

621.396.662.078 2209

A Wideband Searching Automatic Frequency Control Circuit of New Type.—H. Wallman. (*Chalmers tek. Högsk. Handl.*, 1953, No. 132, 21 pp.) A system in which the local oscillator automatically 'searches' a wide frequency range to locate its correct frequency before the a.f.c. operates, cannot be based on a simple i.f. discriminator. In the system described the frequency-sensing element is an i.f. resonant circuit with a detector and a circuit differentiating the detector output with respect to time. Frequency search is effected by a motor-driven coarse-tuning capacitor; an appropriate output from the differentiator actuates a multivibrator which changes the phase of the motor supply, causing the motor to reverse. The coarse-tuning capacitor stops and a fine-tuning capacitor is engaged. The continued presence of a signal gives a periodic output from the differentiator so that the latter capacitor oscillates back and forth every second through about $\pm$ 10°. Over this range a free-play gear operates so that the coarse-tuning capacitor is at rest. The system has important application at u.h.f. Details are given of a model operated at 2·4 Mc/s with an automatic search range of 500 kc/s.

621.396.722 : 621.317.3 2210

The B.B.C. Measurement and Technical Receiving Station at Tatsfield.—H. V. Griffiths. (*B.B.C. Quart.*, Spring 1954, Vol. 9, No. 1, pp. 43–56.) The development, functions and equipment of the station are described. Work undertaken is classified under eight main headings. It involves measurement of frequency (mainly carrier frequencies between 150 kc/s and 150 Mc/s), field strength, and atmospheric noise; relaying programmes for broadcast or transcription; interference identification and programme-schedule checking; d.f. and wave angle measurements; measurement of modulation depth and sideband dispersion. Frequency-checking procedure and apparatus in particular are noted. Four frequency standards are maintained.

621.396.828 2211

Funk-Entstörung. [Book Review]—F. Seelemann. Publishers: O. Elsner, Darmstadt, 1954, 832 pp., DM56. (*Frequenz*, Feb. 1954, Vol. 8, No. 2, pp. 61–62.) Published on behalf of the Federal German Post Office. Sources of radio interference are discussed and methods of measurement and suppression described. A chapter is devoted to the construction of low-interference electrical apparatus and machines.

STATIONS AND COMMUNICATION SYSTEMS

621.376.2 2212

A General Solution of the Two-Frequency Modulation Product Problem: Part 1.—R. L. Sternberg & H. Kaufman. (*J. Math. Phys.*, Jan. 1954, Vol. 32, No. 4, pp. 233–242.) A method is presented for readily obtaining approximate numerical values of the amplitudes of the modulation products occurring in the output of an arbitrary modulator with continuous output/input characteristic, to which a two-frequency input is applied. A partial analytical solution of the problem is also given. Exact values are obtainable in cases such as that of the biased ideal rectifier considered by Bennett (3506 of 1947).

621.376.3.015.7 2213

Frequency Spectra of Individual H.F. Pulses with Varying Carrier Frequency.—S. I. Bytschkow. (*NachrTech.*, Jan. 1954, Vol. 4, No. 1, pp. 7–13.) Translated from *Radiotekhnika, Moscow*, 1950, Vol. 5, No. 1. Analytica expressions are derived for the spectral density for different types of variation of the carrier frequency during the pulse; a graphical method for finding the spectrum is given. The choice of receiver bandwidth for such systems is discussed.

621.39 + 621.317] : 519.213 2214

Probability of Causal Events in Telecommunications and Measurement Technique.—J. Loeb. (*Ann. Télécommun.*, Jan. 1954, Vol. 9, No. 1, pp. 15–19.) Application of Bayes' Law to the calculation of Shannon's equivocation $H_y(x)$ and to the matching of a source to a channel with noise.

621.39.001.11 2215

Two Types of Error due to Noise.—J. Loeb. (*Ann. Télécommun.*, Feb. 1954, Vol. 9, No. 2, pp. 29–34.) Probability theory (2214 above) is discussed in relation to binary-code telegraphy, when the probability of mistaking a signal for noise is quite independent of the probability of accepting a noise voltage as a signal. The effect of additive and multiplicative iteration techniques on the probability of error is examined. According to a 'principle of complements', for a given signal/noise ratio no receiver modification can give a simultaneous reduction of the two types of error.

621.39.001.11 2216

Geometric Aspects of Least Squares Smoothing.—A. A. Hauser, Jr. (*Proc. Inst. Radio Engrs*, April 1954, Vol. 42, No. 4, pp. 701–704.) Function space techniques used by Shannon and Zadeh to clarify concepts in communication theory are considered. "An understanding of the operation of a least squares smoother is enhanced by establishing an m-dimensional space in which inputs and outputs are vectors and the smoother is a transformation. The concept of transmission and rejection manifold as established by Zadeh is introduced and the manner in which signal and noise are separated is illustrated geometrically. An explicit pictorial representation of the process is given for the three dimensional case."

621.395.44 : 621.315.212 2217

S900-101 Coaxial-Pair Equipment.—(*Câbles & Transm.*, Jan. 1954, Vol. 8, No. 1, pp. 4–125.) A series of articles gives details of the planning of the system used in France, of the design of certain individual items of equipment for repeater and terminal stations and of gear for servicing, testing and maintenance work.

621.396.4 : 621.396.65 2218

Experimental Radio Bearer Equipment for Carrier Telephone Systems.—W. S. McGuire & A. G. Bird. (*J. Brit. Instn Radio Engrs*, April 1954, Vol. 14, No. 4, pp. 171–188.) Reprint. See 3412 of 1953.

621.396.65 + 621.396.81 2219

Microwave as Applied to Railroad Operation in the Gulf Coast Area.—L. R. Thomas. (*Elect. Engng, N.Y.*, Jan. 1954, Vol. 73, No. 1, pp. 63–67.) A radio link for a path of approximately 70 miles and comprising two terminal and three repeater stations is described. Eight p.a.m. channels are provided; the equipment operates in the 6·7-kMc/s range. Propagation tests over a period of nearly two years show that, even when the aerial heights have been suitably chosen, fading may be experienced due to (*a*) temperature inversions or changes in humidity, and (*b*) fog at one of the station sites. To be published also in *Trans. Amer. Inst. elect. Engrs*, 1953, Vol. 72.

621.396.65 2220

Carrier-Current Radio Links in the Light of C.C.I.F. Recommendations.—J. P. Vasseur. (*Ann. Radioélect.*, Jan. 1954, Vol. 9, No. 35, pp. 47–82.) C.C.I.F. recommendations in respect of signal/noise and signal/crosstalk ratios are considered in relation to f.m. links. The noise contribution of each element in the transmission chain is assessed, and the minimum performance required of each element specified. A 2 500-km f.m. link will require careful planning if it is to satisfy C.C.I.F. requirements. Sine-wave signals are suitable for testing individual elements, but noise-modulated sine-wave signals are preferable for overall tests of the complete link.

621.396.97 : 621.397.24/.26 **2221**
Technical Arrangements for the Sound and Television Broadcasts of the Coronation Ceremonies on 2nd June, 1953.—Procter, Pulling & Williams. (See 2231.)

SUBSIDIARY APPARATUS

621-526 : 621.3.015.7 **2222**
The Pulse Transfer Function and its Application to Sampling Servo Systems.—R. H. Barker. (*Proc. Instn elect. Engrs*, Part IV, Feb. 1954, Vol. 101, No. 6, pp. 162–163.) Discussion on 1131 of 1953.

621.3.013.783.001.4 **2223**
Measuring the Effectiveness of Shielding Materials.—C. DeVore. (*Elect. Mfg*, Aug. 1953, Vol. 52, No. 2, pp. 122–125.) A method developed at the U.S. Naval Research Laboratory is described. Only a small sample of the material under test, e.g. wire mesh or conducting coating, is needed. The sample is inserted between two shield cans containing transmitting and receiving aerials respectively, and the attenuation increase produced by the sample is observed. The frequency range investigated is 15 kc/s–100 Mc/s; it may be possible to extend the method to higher frequencies by using waveguide technique.

621.314.57 : 621.387 **2224**
Thyratron Inverter.—J. D. Howells. (*Wireless World*, May 1954, Vol. 60, No. 5, pp. 237–241.) Description of a unit giving 100 W at 240 V, 50 c/s from d.c. mains.

621.314.6 **2225**
Rectifier with Smoothing Capacitor and High-Vacuum Valves or Selenium Rectifiers.—K. Müller-Lübeck. (*Arch. Elektrotech.*, 1954, Vol. 41, No. 4, pp. 181–195.) Exact equations are derived for a p-phase rectifier circuit with a capacitance connected across the load. For practical use approximate formulae are derived, and their application is illustrated by the design of a two-phase rectifier for an output of 2·25 kV, 0·7 A.

621.314.63 + 621.316.93] : 546.28 **2226**
Silicon *P-N* Junction Power Rectifiers and Lightning Protectors.—G. L. Pearson & C. S. Fuller. (*Proc. Inst. Radio Engrs*, April 1954, Vol. 42, No. 4, p. 760.) A method of preparing large-area-junction diodes is indicated, in which donor (e.g. P) or acceptor (e.g. B) impurities are diffused into high-purity single crystals of Si at temperatures above 1 000°C. Low-resistivity surface layers are produced, facilitating the application of contacts. Electrical characteristics are presented of a power rectifier and a lightning protector prepared by this method.

621.316.722.1 **2227**
A Simple Mains Unit for Highly Constant Alternating Voltages and Currents.—H. Helke. (*Elektrotech. Z.*, *Edn A*, 1st Jan. 1954, Vol. 75, No. 1, pp. 11–13.) A unit giving a voltage continuously variable up to 600 V and currents up to 6 A is described, the voltage varying not more than ±0·1% with mains voltage variations up to ±10% and mains frequency variations up to 0·5%.

621.316.722.1 **2228**
A New D.C. Electronic Voltage-Stabilizing Circuit.—R. B. Mackenzie. (*Proc. Instn elect. Engrs*, Part II, Feb. 1954, Vol. 101, No. 79, pp. 59–63.) The detecting element across the supply consists of a negative-resistance and a positive-resistance element of equal numerical values in series. By controlling the negative resistance complete compensation can be achieved. A circuit built on a similar principle but using a different type of negative-resistance element was described by Patchett (227 of 1951). A stabilizer with an output of 400 V is described in detail; the output can be maintained constant to within ±0·001% for an input fluctuation of ±10% and a load variation of 1–20 mA. The circuit is suitable for voltages from 70 V upwards.

621.318.5 **2229**
Relay Standardization.—(*Elect. Mfg*, Sept. 1953, Vol. 52, No. 3, pp. 146–154..380.) Summarized report of proceedings of symposium on e.m. relays, held at Oklahoma in June 1953. Military applications were the main consideration.

621.355 **2230**
Recent Patents on Electrical Accumulators.—L. Jumau. (*Rev. gén. Élect.*, Feb. & March 1954, Vol. 63, Nos. 2 & 3, pp. 59–70 & 129–142.) Review of developments. For previous review see 3177 of 1950.

TELEVISION AND PHOTOTELEGRAPHY

621.397.24/.26 : 621.396.97 **2231**
Technical Arrangements for the Sound and Television Broadcasts of the Coronation Ceremonies on 2nd June, 1953.—W. S. Procter, M. J. L. Pulling & F. Williams. (*Proc. Instn elect. Engrs*, Part I, March 1954, Vol. 101, No. 128, pp. 57–72. Discussion, pp. 73–78.) A comprehensive description of the arrangements and equipment used for the simultaneous broadcasts to the United Kingdom and parts of Western Europe. An account is included of the sound and television recording arrangements of the B.B.C. and the North American organizations, which enabled telerecordings to be screened in America within 11 hours of the event. See also 2804 of 1953 (Bridgewater).

621.397.26 **2232**
U.H.F.-TV Satellite Operation.—J. S. Allen. (*Sylvania Technologist*, Jan. 1954, Vol. 7, No. 1, pp. 3–7.) An unattended system is described consisting of two transmitters, one on a hilltop and the other about 1½ miles away in the small town to be served. Programmes picked up at the hilltop station are re-broadcast on channel 22 (518–524 Mc/s) and relayed by low-power microwave link to the town transmitter, broadcasting on channel 82 (878–884 Mc/s).

621.397.335 **2233**
Portable Sync Generator for TV Broadcasting.—H. E. Ennes. (*Electronics*, April 1954, Vol. 27, No. 4, pp. 138–141.) Description, including detailed circuit diagram, of equipment using 22 miniature valves and providing a standard R.E.T.M.A. synchronizing signal at 4 V negative peak to peak across 75Ω.

621.397.5 : 621.396.11.029.62 **2234**
Wave Propagation and Television Broadcasting at Very High Frequencies.—R. L. Smith-Rose. (*Proc. Instn Radio Engrs, Aust.*, Jan. 1954, Vol. 15, No. 1, pp. 7–12.) "The fundamental factors determining the bandwidth of a television system are reviewed and the present arrangement of the v.h.f. channels in the B.B.C. service is outlined. Experiments on the propagation of radio

waves over various distances in the frequency range 30 to 100 Mc/s are then described and the results related to the problem of extending the service by means of shared channels."

621.397.5(083.74)(931) **2235**
The Choice of a Television Standard.—N. R. Palmer. (*Radio & Electronics, Wellington, N.Z.*, 1st Feb. 1954, Vol. 8, No. 12, pp. 17–19.) Discussion of various considerations affecting the choice of appropriate standards for operation in New Zealand. The only difference between the New Zealand recommendations and standard British practice was in the choice of horizontal polarization. The frequency channels recommended were 48–53 Mc/s for Wellington, 53–58 Mc/s for Dunedin, 58–63 Mc/s for Christchurch and 63–68 Mc/s for Auckland.

621.397.61 : 535.316/.319].001.4 **2236**
The Measurement of the Performance of Lenses.—W. N. Sproson. (*B.B.C. Quart.*, Spring 1953, Vol. 8, No. 1, pp. 55–64.) Lenses for television cameras and other television purposes are discussed.

621.397.611 : 778.5 **2237**
New Method for Television Film Scanning.—T. Stutz. (*Tech. Mitt. schweiz. Telegr.-TelephVerw.*, 1st Jan. 1954, Vol. 32, No. 1, pp. 1–27.) A comprehensive analysis is made of known methods, and a description is given of a continuous-motion picture projector using two optical systems in alternation. The film motion is compensated by moving the axes of the optical systems alternately forward and backward with respect to the film travel. The arrangement is used in conjunction with a flying-spot scanner.

621.397.611 : 778.5 **2238**
New 35-mm Television Film Scanner.—E. H. Traub. (*J. Soc. Mot. Pict. Telev. Engrs*, Jan. 1954, Vol. 62, No. 1, pp. 45–54.) A flying-spot system with continuous exposure and continuous film motion combined with optical compensation by means of a new type of rotating-prism device is described.

621.397.611.2 **2239**
The Supericonoscope IS 9 mm/10.—R. Just. (*Radio Tech., Vienna*, Jan. 1954, Vol. 30, No. 1, pp. 13–21.) Detailed description of a recently developed German camera tube with magnetic deflection and focusing systems. Associated supply, scanning and protection circuits are shown and operating characteristics are discussed, particularly sensitivity and performance at different levels of illumination.

621.397.611.2 : 537.531 **2240**
X-Ray Noise Observation using a Photoconductive Pickup Tube.—A. D. Cope & A. Rose. (*J. appl. Phys.*, Feb. 1954, Vol. 25, No. 2, pp. 240–242.) X-ray images have been recorded directly using a vidicon tube with Se target 0·001-in. thick. Noise effects due to the absorbed photons are clearly visible. Clear transmitted pictures at a rate of 30/sec were obtained when irradiating the vidicon with a 5-mA 100-kV beam at a distance of 2 ft.

621.397.62 **2241**
An Outline of the British Television System: Part 3—Receiving the Picture Waveform.—D. Wray. (*P.O. elect. Engrs' J.*, Jan. 1954, Vol. 46, Part 4, pp. 166–170.) Typical aerials and circuits for domestic receivers are described; the effects of receiver faults and external interference on reception are discussed. Part 2: 873 of March.

621.397.62 **2242**
Band III Convertor.—G. H. Russell. (*Wireless World*, May 1954, Vol. 60, No. 5, pp. 211–213.) A simple circuit for adapting Band-I television receivers to accommodate either of the two proposed British Band-III television channels uses a self-oscillating type of mixer whose output is fed to the receiver via a step-down i.f. transformer. Details are given of components and alignment procedure.

621.397.62 **2243**
Signal Overload Relay for Television Receivers.—C. Masucci, J. R. Peltz & W. B. Whalley. (*Electronics*, April 1954, Vol. 27, No. 4, pp. 153–155.) When a high-sensitivity receiver is operated in an area where alternative strong and weak signals are available, it is desirable to make provision for automatically reducing the gain for the strong signal and restoring it for the weak one; the same arrangement can be used simply to prevent overloading. The device described consists of a relay in the r.f. and i.f. anode circuits which is arranged to cut out the r.f. amplifier valve when the a.g.c. voltage reaches a set level.

621.397.62 : 535.623 **2244**
Color Decoder Simplifications based on a Beam Deflection Tube.—R. Adler & C. Heuer. (*Trans. Inst. Radio Engrs*, Jan. 1954, No. PGBTR-5, pp. 64–70.) In a synchronous demodulator for the N.T.S.C. two-component colour signal, the usual multigrid valve is replaced by a beam valve; one signal component is applied to a conventional control grid, and the other to a deflection control system. Circuit arrangements and experimental results are presented.

621.397.62 : 535.623 **2245**
TV Color Detectors use Pulsed-Envelope Method.—K. Schlesinger. (*Electronics*, March 1954, Vol. 27, No. 3, pp. 142–145.) Synchronous detection of the two quadrature components of the N.T.S.C. colour signal is accomplished by means of rectifiers, which may be conventional diodes or triodes, keyed in and out by a local oscillator. Balanced and unbalanced arrangements are described. See also *Trans. Inst. Radio Engrs*, Jan. 1954, No. PGBTR-5, pp. 53–63.

621.397.62 : 621.385.832 : 537.531 **2246**
The Production of X Rays by Television C.R. Tubes.—K. H. J. Rottgardt. (*Fernmeldetech. Z.*, Feb. 1954, Vol. 7, No. 2, pp. 74–75.) X-ray film measurements on the Valvo MW 36-44, the Lorenz Bs 42 R-3 and Bs 42 R-6 c.r. tubes, at anode voltages $\geqslant$ 16 kV, indicate that the amount of X-ray radiation penetrating the glass envelope is negligible. An editorial note adds that the radiation intensity from a 15-kV tube was equivalent to that from 1 μg Ra, i.e. of the same order as that from a luminous watch dial.

621.397.621.018.75 **2247**
Development of Television Pulse-Regeneration Equipment.—K. H. Vogt. (*Fernmeldetech. Z.*, Feb. 1954, Vol. 7, No. 2, pp. 53–55.) Description of equipment used at Höhbeck relay station. For a description of similar equipment see 558 of February (Dröscher).

621.397.621.2 : 535.623: 535.37 **2248**
The Preparation of Phosphor Screens for Color Television Tubes.—S. Levy & A. K. Levine. (*J. electrochem. Soc.*, Feb. 1954, Vol. 101, No. 2, pp. 99–103.) See 3440 of 1953.

621.397.813 **2249**
'Sensation-Correct' Gamma Correction of Television Pictures.—P. R. Arendt. (*Arch. elekt. Übertragung*,

Jan. 1954, Vol. 8, No. 1, pp. 1–4.) Results of subjective tests indicate that equalization to unity gamma causes a distortion of the half-tone scale. The construction of subjectively correct equidistance brightness scales for given adaptation levels is discussed.

621.397.828 **2250**

Beat between Sound Carrier and Color Signal Components in a Television Receiver.—J. E. Allen. (*Trans. Inst. Radio Engrs*, Jan. 1954, No. PGBTR-5, pp. 71–86.) Reception of signals conforming to the N.T.S.C. standards is discussed. Analysis of the beat produced in the detector between the sound carrier and the colour sub-carrier, and visible as streaks in the picture, shows that the sound carrier attenuation must be at least 14 db greater than the desired reduction in beat level below a full black-to-white transition. Tests show that 30-db attenuation is required for beat suppression. The sound level must therefore be at least 44 db below mid-band response. To allow for transmission variations, a further margin should be provided. The requirement is more easily realizable in a receiver with a response curve sloping so as to reduce the response to the colour sub-carrier. Effects of the beat in the chrominance channel are simultaneously suppressed.

TRANSMISSION

621.376.32 **2251**

Frequency Modulation of Microwaves.—S. Freedman. (*Radio & Telev. News, Radio-Electronic Engineering Section*, Jan. 1954, Vol. 51, No. 1, pp. 16–18.) Direct frequency modulation of the carrier frequency in a system comprising microwave generator and resonant cavity is achieved by altering the dimensions of the resonant cavity. One wall of the cavity consists of a light framework supporting a series of concentric wire rings. This framework replaces the cone in a loudspeaker, and vibrates when a.f. signals are applied to the loudspeaker. The vibrating wall is not in electrical connection with the main body of the cavity. A TE_{0n} mode must be used. Modifications may be made so that the system operates as an a.f.c. unit

621.396.61 **2252**

High Power H.F. Broadcast Transmitters.—D. F. Bowers & J. F. Ennos. (*Marconi Rev.*, 1st Quarter 1954, Vol. 17, No. 112, pp. 16–36.) A detailed account is given of the design of a series of 100-kW transmitters Type BD.253 for the frequency bands 160–285 kc/s, 525–1 605 kc/s and 5·9–26·1 Mc/s respectively. Good overall efficiency is obtained by using valves with thoriated tungsten filaments; air cooling is used. Valve filaments are heated by a.c., and hum is reduced by feedback. Hot-cathode Hg-vapour rectifiers provide the h.v. supply.

621.396.61 **2253**

Design of Transmitter Power Stages from Valve Data (with Formulae).—W. Lacmann. (*Frequenz*, Dec. 1953, Vol. 7, No. 12, pp. 369–375.) Formulae for calculating operating conditions for these circuits are presented, based mainly on applied anode voltage, maximum anode dissipation and maximum emission current as parameters. It is assumed that the valve is operated so that the voltage drop across it is a minimum consistent with maximum anode current, but modifications to the formulae are indicated for other cases. Diagrams are given from which efficiency can be estimated under various operating conditions. The practical application of the formulae is illustrated for power amplifier and frequency-doubler stages.

621.396.61 : 621.314.7 **2254**

160-Metre Transistor Transmitter.—A. Cockle. (*Wireless World*, May 1954, Vol. 60, No. 5, p. 217.) A Type-OC51 transistor (alpha cutoff at 1·5 Mc/s) is used in an amateur transmitter employing the negative-resistance base-oscillator principle and locked over about 1 kc/s by a 1·8 Mc/s crystal. With power up to 100 mW, signals have been received at distances up to 30 miles.

621.396.61.029.58 : 621.396.65 **2255**

Single-Sideband Multi-Channel Operation of Short-Wave Point-to-Point Radio Links: Part 4(b) — An Independent-Sideband High-Power Short-Wave Transmitter — Design and Performance.—H. E. Sturgess & F. W. Newson. (*P.O. elect. Engrs' J.*, Jan. 1954, Vol. 46, Part 4, pp. 191–195.) Part 4(a): 882 of March.

621.396.61.029.62 **2256**

V.H.F./U.H.F. Transmitters for Experimental Work.—(*Elect. J.*, 26th March 1954, Vol. 152, No. 13, pp. 1031–1032.) A band-III transmitter for use by the B.B.C. is described. Square-waveform modulation at 1 kc/s is provided for purposes of field strength measurements, and pulse modulation for investigation of multipath transmission and echo effects. Only one r.f. circuit is used for both conditions of modulation. The transmitter is intended for operation in a van, and the power-supply unit is accordingly designed to deal with large variations of mains supplies. The output is 150 W c.w. The development of another band-III transmitter and of band-IV and band-V transmitters is mentioned.

VALVES AND THERMIONICS

621.314.632 : 546.289 : 621.374 **2257**

"Positive-Gap" Germanium Diode.—A. H. Reeves. (*Onde élect.*, Jan. 1954, Vol. 34, No. 322, pp. 32–37.) A suitable point contact, such as Ag with As impurity, and a suitable electrical forming technique produce a discontinuity in the diode forward characteristic useful in pulse techniques. Its application in a 20-Mc/s pulse generator and in counting and trigger circuits operating at higher frequencies is illustrated.

621.314.7 **2258**

The Theory of Physical Principles involved in the A-Type Transistor Action.—Y. Watanabe & N. Honda. (*Sci. Rep. Res. Inst. Tohoku Univ., Ser.B*, Dec. 1952, Vol. 4, No. 1, pp. 117–163.) Detailed consideration of the nature of the barrier layer of the collector contact and the mechanism of rectification of high-back-voltage Ge rectifiers, and of the mechanism of transistor action principally in *n*-type Ge. A comparison of the theoretical results with the experimental results of Bardeen & Brattain (2979 of 1949) shows satisfactory agreement. The current feedback factor is also considered.

621.314.7.001.4 **2259**

Measuring Transistor Temperature Rise.—J. Tellerman. (*Electronics*, April 1954, Vol. 27, No. 4, pp. 185–187.) The temperature rise as a function of the power dissipation is determined by noting the change in the reverse collector current, for zero emitter current. Circuits are illustrated for providing an oscilloscope or a meter indication of collector current. Experiments with Type 2520 and 2521 transistors using various cooling arrangements showed that the temperature rise could be reduced from about 26° to about 20° above an ambient temperature of 27°C, thus permitting an increase of 25–30% in the power dissipation.

621.314.7.012.6 : 621.317.755 **2260**

Characteristic-Curve Tracer for Transistors.—H. Lennartz. (*Funk u. Ton*, Jan. 1954, Vol. 8, No. 1, pp.

25–29.) Description of equipment used in conjunction with a c.r.o.

621.383.2 **2261**

Alkali Photocells: Part 2.—M. Ploke. (*Arch. tech. Messen*, Jan. 1954, No. 216, pp. 15–18.) 70 references. Part 1: 1608 of May.

621.385.029.63/.64 **2262**

On the Focusing of High-Current Electron Beams.—G. R. Brewer. (*J. appl. Phys.*, Feb. 1954, Vol. 25, No. 2, pp. 243–251.) The trajectories of electrons under the influence of magnetic and space-charge forces in a system of periodically spaced magnetic lenses are determined. Graphs are presented for use in designing a focusing system for given values of current, voltage, and beam diameter. Use of multiple magnetic-lens systems of the type investigated offers considerable economies over the use of long focusing coils for travelling-wave valves.

621.385.032.2 : 537.533.9 **2263**

Some Effects of Slow Electron Bombardment in Thermionic Valves.—D. A. Wright. (*Brit. J. appl. Phys.*, March 1954, Vol. 5, No. 3, pp. 108–111.) A review of recent work. See also 1748 of June (Wright & Woods).

621.385.032.216 **2264**

Decay of Emission from an Oxide-Coated Cathode due to Adsorption of Matter liberated from the Anode.—S. Deb. (*J. Brit. Instn Radio Engrs*, April 1954, Vol. 14, No. 4, pp. 157–167.) Two types of adsorption procedure are distinguished, (*a*) when the adsorbed particles remain mobile on the surface, (*b*) when the adsorbed particles are immobile. Equations for the decay processes in the two cases are given in integral form and are solved numerically using estimated values of the constants involved. An approximate analytical solution is also given for case (*a*). The decay in this case is generally of short duration and the emission current is likely to recover spontaneously, whereas in case (*b*) the decay is of long duration and is likely to be permanent. Differences between the pulsed and d.c. behaviour are explained on the hypothesis that the poisoning of the cathode is only temporary in the pulsed case.

621.385.032.216 **2265**

On the Crystallizations of Alkaline Earth Carbonates and Effects of Sizes and Shapes of these Crystals on the Oxide-Coated Cathodes.—H. Nukiyama, E. Takagi & A. Sato. (*Sci. Rep. Res. Inst. Tohoku Univ., Ser. B*, Dec. 1952, Vol. 4, No. 1, pp. 87–103.) The principal conclusions of this experimental investigation are: (*a*) in the crystallization of carbonates by the precipitation method, temperature, concentration of reactants and pH of precipitant are important, (*b*) the velocity of decomposition into oxides on heating in vacuum depends mainly on the shape rather than the size of the crystals, and (*c*) thermionic emission is not strongly influenced by crystal shape or size.

621.385.2 **2266**

Test of Langmuir's Three-Halves Power Law, Deviations due to β^2-Factor and Secondary Emission.—P. L. Walraven. (*Appl. sci. Res.*, 1954, Vol. B3, No. 6, pp. 393–399.) Results of measurements on a cylindrical diode, in which the filament current and the anode voltage were interrupted alternately at a rate of 3 500/sec, indicate that there is a large deviation from Langmuir's law due to an additional space charge which is produced mainly by reflected electrons at anode voltages up to 30 V; the correction factor should then be $\beta^2(1 + 2\alpha)$ instead of β^2, where α is the reflection coefficient. For a Ta anode α is 14%. The correction for the space charge in a planar diode is also derived.

621.385.3.012 **2267**

Calculation of Characteristic Curves for Planar Triodes.—O. Heymann. (*Frequenz*, Feb. 1954, Vol. 8, No. 2, pp. 33–40.) Equations for the slope and valve current are derived from the valve geometry, taking into consideration the formation of 'islands' by the grid wires. The slope equation relates the cut-off potential to the grid-cathode distance and can conveniently be used as the starting point for valve design. The numerical values of the elliptic integrals occurring in the equations are shown graphically and tabulated. See also 3609 of 1952 (Dahlke).

621.385.8 **2268**

Approximate Electrode Shapes for a Cylindrical Electron Beam.—E. R. Harrison. (*Brit. J. appl. Phys.*, Jan. 1954, Vol. 5, No. 1, pp. 40–41.) With the aid of some approximations, a simple equation is derived for the shapes of electrodes to produce a parallel beam of rotational symmetry in which the current is space-charge limited. The method has been used for designing a proton-beam accelerator.

621.385.832 : 537.533 **2269**

The Geometrical-Optical Distribution of Intensity over the Focused Spot of the Deflected Electron Beam.—H. Grümm. (*Optik, Stuttgart*, 1954, Vol. 11, No. 1, pp. 32–43.) Analysis is presented which gives the approximate distribution with far less calculation than by the method based on wave mechanics.

621.385.832 : 621.317.755 **2270**

An Interesting Four-Beam Cathode-Ray Tube.—H. F. Wendling. (*Funk-Technik, Berlin*, Feb. 1954, Vol. 9, No. 4, pp. 102–103.) The single-gun oscillograph tube, Type V113, has a first cylindrical e.s. lens producing a ribbon beam; this is deflected as a whole by the e.s. horizontal-deflection system and is split into four in passing through the multiple e.s. vertical-deflection system, the individual ribbon beams being restored to pencil form by the multiple cylindrical lens constituted by the apertures of the vertical-deflection system in combination with the post-deflection accelerator.

621.396.822 : 537.533 **2271**

Cathode Boundary Conditions and Noise Minima in Electron Beams.—R. Wiesner & H. W. König. (*Arch. elekt. Übertragung*, Jan. 1954, Vol. 8, No. 1, pp. 5–7.) The assumption of the existence of a convection-current fluctuation at the cathode in addition to the velocity fluctuation leads to a theoretical noise distribution in good agreement with that found experimentally by Cutler & Quate (1274 of 1951). In particular the noise minima have values differing from zero.

MISCELLANEOUS

061.4 : [621.317.7 + 621.35 **2272**

Physical Society's Exhibition [1954].—(*Wireless Engr*, May 1954, Vol. 31, No. 5, pp. 132–136.) Brief descriptions are given of selected exhibits. See also *Elect. Times*, 8th April 1954, Vol. 125, No. 3257, pp. 506–509, and *Instrum. Practice*, March 1954, Vol. 8, No. 3, pp. 234–244.

621.37/.39 **2273**

Radio Progress during 1953.—(*Proc. Inst. Radio Engrs*, April 1954, Vol. 42, No. 4, pp. 705–759.) A review presented with the main purpose of giving specialist workers an idea of developments in fields other than their own. Over 1 100 references are given. The headings under which the material is arranged are almost identical with those used for the 1952 review.

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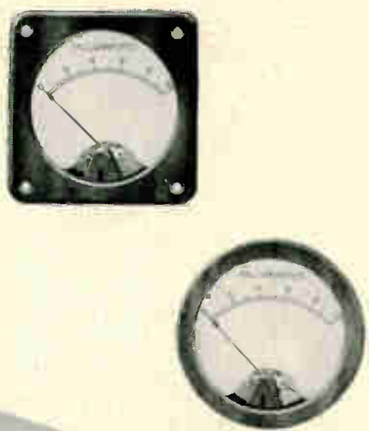

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