WIRELESS ENGINEER

Vol. XXVIII OCTOBER 1951 No. 337

Electrostatic D.C. Transformer

IN a paper read at a meeting of the American Physical Society[*] in June 1950, by J. M. Malpica, there is described certain research work being carried out at the National Polytechnic Institute of Mexico. A well-known method of obtaining a very high voltage discharge is to charge a number of capacitors in parallel from a relatively low-voltage source and then to connect them in series. This is, of course, an intermittent process, but the apparatus described in the paper is such that the process becomes continuous and gives a steady output current at a multiple of the input voltage. The object of the research is to obtain a source of d.c. high voltage for the acceleration of electrified particles.

A glass disc (Fig. 1) 1 metre diameter and 7 mm thick is rotated at a speed of 200 to 350 revs. per min. between two brass brushes extending radially from $r = 25$ cm to $r = 39$ cm; that is, the brushes are 14 cm long and about 5 cm wide. The primary source of supply, consisting of a mechanical rectifier with a bank of capacitors in parallel, is connected to the brushes, the potential difference being 40 to 50 kilovolts. The glass between the brushes is thus subjected to a steady electric field strength of about 6 kV per mm. As the glass moves away from the brushes some interesting questions arise. If the brushes were not in contact with the glass but separated from it by an air-gap, then, neglecting sparking and corona effects, one would not expect the glass to retain its charged condition to any appreciable extent as it moved out of the field, but if the brushes are actually in contact with the glass, the question is not so simple.

Benjamin Franklin devised an experiment that became a classic and was described in every text-book on electricity. He made a slightly conical Leyden jar in which the inner and outer metallic coatings could be removed from the glass. When the coatings were carefully removed from the charged jar it was found that they had little if any charge, but, on putting the jar together again, it was found to be highly charged, thus proving—or so he thought—that the charges remained on the inner and outer surfaces of the glass dielectric.

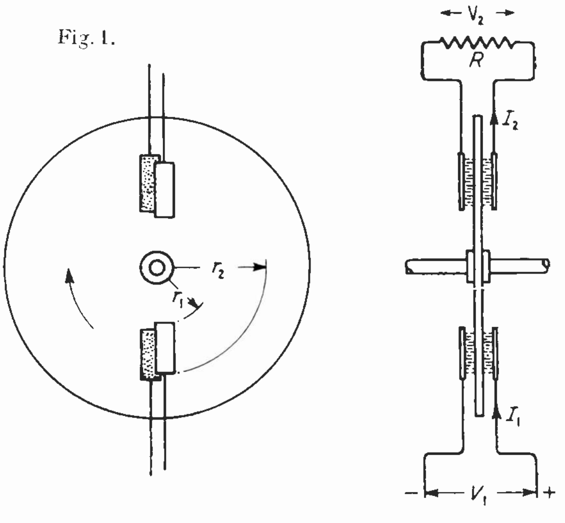


Fig. 1.

Nobody seemed to bother about the nature of these charges on the glass until it occurred to G. L. Addenbrooke to do the experiment with perfectly clean and dry glass in a desiccator. Under these circumstances the results were entirely contrary to those of Benjamin Franklin; the charges remained on the metal coatings and,

---

[*] "Electrostatic Direct-Current Transformer of 300 Kilovolts," *Journal of Scientific Instruments*, June 1951, p. 364

if these were brought into contact and then replaced on the glass jar, it was found to be discharged. Hence, under ordinary circumstances there is sufficient moisture on the glass to serve as a conducting coating and maintain the charged condition. In the case of the rotating glass disc the brass brush may modify the nature of the surface layer.

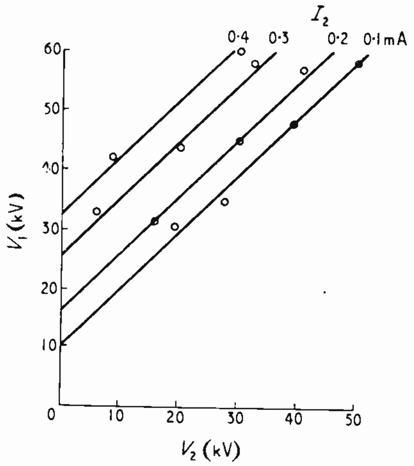


Fig. 2.

In the apparatus described, if the disc were at rest there would be a certain steady current over the surface and round the edge from the positive to the negative side, but this would be small because of the very high resistance of the surface layer, especially if the moisture were in the form of discrete particles. The p.d. between opposite faces would fall off rapidly with increasing distance from the brushes, reaching a small value near the edge. On rotating the disc, there will be an increased flow of current from the supply to recharge to the full p.d. the incoming partly-discharged glass. Tests made at a speed of 278 r.p.m. at various voltages gave the following results:

p.d. =	27·5	31	36	45	54 kV
current =	0·01	0·015	0·02	0·035	0·08 mA

It will be seen that the current increases very rapidly when the p.d. exceeds about 40 kV.

If now a similar pair of brushes are placed on another part of the disc at the same radial distance and connected by means of a high resistance, the charged glass passing between them will partly discharge through the resistance and maintain a steady current through it. If the brushes extend radially from r_1 to r_2 the area passing through them per second is $\pi(r_2^2 - r_1^2)n$ where n = revs. per sec. and if the capacitance per unit area is C_a, and the primary and secondary voltages V_1 and V_2 respectively, then, neglecting losses, the quantity discharged at the secondary brushes per sec. will be $\pi(r_2^2 - r_1^2)nC_a(V_1 - V_2)$. This then must be the secondary current I_2 flowing through the load resistance R and $V_2 = I_2R$. Theoretically the primary and secondary currents should be equal but, as we saw above, there was a primary current even before the secondary brushes were fitted, and the primary current always exceeds the secondary current. The primary voltage must always exceed the secondary voltage since the action depends on the difference $V_1 - V_2$. Hence, one cannot expect a very high efficiency. The author introduces a ratio S of the observed secondary current I_{ob} to the value calculated by the above formula, C_a having been determined by a.c. measurements; from the above formula, we then have

$$V_1 = V_2 + \frac{I_{ob}}{SC_a n\pi(r_2^2 - r_1^2)}$$

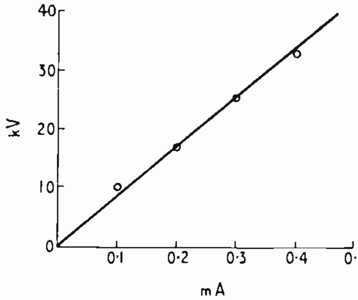


Fig. 3.

This agrees with the experimental results shown in Fig. 2 for which the speed was 278 r.p.m. The current was adjusted by varying the load resistance R, which had a maximum value of 1,000 megohms; for example, the point 0·1 mA with $V_2 = 50$ kV corresponds to a load of 500 megohms. These results show that for a given current, $V_1 = V_2 +$ a constant, and that the constant is roughly proportional to the current, as indicated by the above formula. The results of Fig. 2 are plotted in a different way in Fig. 3; here the values of V_1 when $V_2 = 0$ (i.e., the intercepts on the vertical in Fig. 2) are plotted against the current, and are seen to be approximately on a straight line through the origin. This gives the value of V_1 when the secondary is short-circuited; under these circumstances a current of 0·2 mA requires a primary voltage of 17 kV, and since Fig. 2 shows that the difference between V_1 and V_2 is constant for a given current, a secondary voltage of 30 kV would require a primary voltage of 30 + 17 = 47 kV. As one would expect, the ratio S varies considerably, depending on the conditions; for $dA/dt = 16490$ cm²/sec and $I_2 = 0{\cdot}1$ mA, $S = 0{\cdot}654$, whereas for $dA/dt = 9880$ cm²/sec and $I_2 = 0{\cdot}2$ mA, $S = 0{\cdot}939$.

Instead of two sets of brushes placed 180° apart, four sets may be used at 90°, alternately primary and secondary, the primary brushes being connected in parallel and the secondary either in parallel or series. In the actual transformer a number of glass discs are mounted on the shaft; to obtain a step-up ratio of 1 to 3 six active discs are employed as shown in Fig. 4, an extra disc being fitted at each end to act as a shield. A similar set with a ratio of 1 to 6 has 12 active discs and two shields. Using the two sets with their secondaries in series an output of 1 mA was obtained at about 300 kV. Each disc has four pairs of brushes, the top and bottom pair, connected in parallel, being the primary, and the two side pairs, also connected in parallel, acting as the secondary. Fig. 4 is a plan; the top primaries can be seen; the bottom primaries are exactly similar; the three terminals marked + are connected together and to the three similar bottom terminals; similarly the eight – terminals are connected together; thus all the primaries are in parallel. The sides of the discs facing one another are of the same polarity. As connected in Fig. 4, it is seen that, not only are the two sets of secondary brushes in parallel, but also the secondary brushes of discs 1 and 2; these are then connected in series with those of discs 3 and 4, and those of discs 5 and 6. The secondary voltage is thus that of three pairs of brushes in series, with four parallel paths through the transformer. Insulating rims are fitted to the discs, both to strengthen them and to increase the length of leakage paths. The shaft is covered with insulating material. To reduce the corona loss the connections are made with closely-wound coils of wire about 14 cm dia.

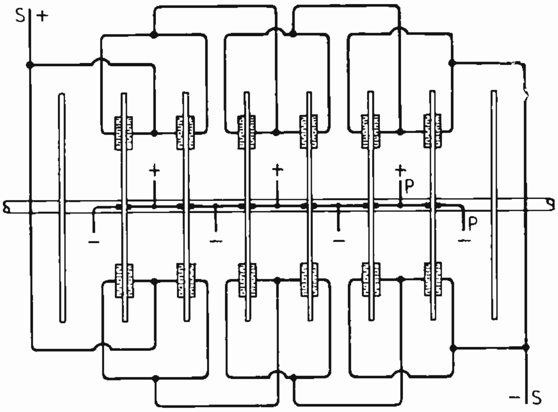


Fig. 4.

Such a machine can be used for insulation measurements with a sensitive d.c. instrument, because the short-circuit current is very limited. It is suitable for X-ray work because of the greater efficiency with the steady d.c. source instead of a rectified a.c. voltage of the same peak value.

This apparatus is stated to be an intermediate step in the development of a one-million volt transformer, and one will await with interest its further developments.

G. W. O. H.

FLYWHEEL SYNCHRONIZING CIRCUIT

For Television-Receiver Time Bases

By A. B. Starks-Field, B.Sc.

(*Marconi's Wireless Telegraph Co., Ltd.*)

The triggering method of synchronizing line time bases in television receivers is satisfactory so long as the triggering pulses are very decisive and free from irregularities. Normally no synchronizing waveform can be ideally square in form and in general it is made up of exponential forms as shown in Fig. 1(a). In 'fringe' areas where the signal-to-noise ratio may be low the edges of these pulses may be seriously modified in form by the superposition of noise, as is shown in Fig. 1(b). Normally such pulses are differentiated and the spike corresponding to the leading edge used to trigger the time base. Referring to Fig. 1(c), which represents the differentiation of 1(b), the dotted line indicates the level at which the time base might be considered to fire, and it will be seen that differences of the order of a microsecond from the correct timing occur. This results in the displacement of lines from each other by anything of the order of 3 or 4 picture elements with the consequent jagged appearance of the verticals in a picture. The effect is entirely random and at normal viewing distance results in an apparent loss in focus of the vertical characteristics.

If, however, the time base could be made to run consistently at an almost constant rate with a continuous automatic adjustment taking place to keep it in synchronism with the picture signal, the irregularities would average out and each line would be correctly placed with respect to its neighbour, resulting in a very much clearer picture. Fig. 4 shows the results of operating an English Electric Model 1550 in a noisy location with and

---

MS accepted by the Editor, May 1951

without flywheel synchronizing.

Probably the most effective way of dealing with the problem is to take a suitable waveform from the time base, mix it with the incoming synchronizing signal and derive a unidirectional voltage which is a function of the phase difference between these two waveforms. This voltage is then used as a control voltage to govern the recurrence frequency of the time base. Such a system must inevitably have a long time constant so that it will be insensitive to the irregularities of the synchronizing pulses.

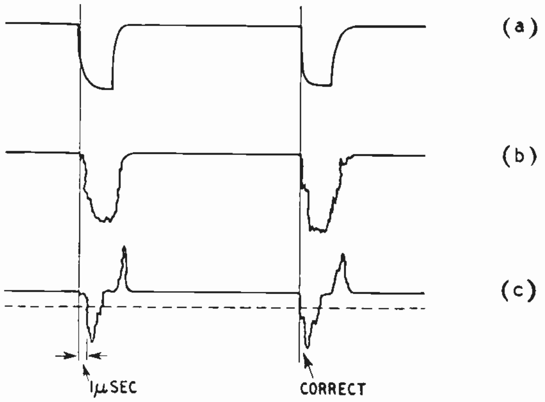


Fig. 1. *Typical sync pulse waveform (a) and with added noise (b). After differentiation the latter takes the form (c).*

The circuit about to be described is one method of achieving this, though there are many variations. The waveform required from the time base is a saw-tooth, shown in Fig. 2(d), which may be derived from any suitable point in the circuit. Its form is not critical except that the slope of the leading edge should be fairly steep, of the order of 2 V/μsec and about 10 V peak-to-peak in amplitude. In the particular arrangement to be described this waveform is derived from a blocking oscillator which is supplying artificial synchronizing pulses to a triggered time base, but there is no reason why the waveform should not be derived from the time base itself.

Referring to Fig. 2, (a) shows the incoming synchronizing waveform including part of the frame synchronizing pulses, since the behaviour of the circuit under *their* influence must be considered. This waveform is semi-differentiated by an RC circuit to that shown in (b). The negative pulses of this waveform correspond to the correct firing times of the line time base except during the frame period when alternate ones correspond to the middle of the scan. This waveform is then phase-inverted and clipped so that the positive-going pulses in the output are all of the same height and approximately the same width, as shown in (c). Waveforms (c) and (d) are then added together resulting in a waveform between the two extreme conditions shown in (e) and (f); (e) represents the result when the time base is firing late while (f) that when the time base is firing early. The peak voltage is greater when the time base is firing late. This then may be fed to a peak-voltage detector and the resulting output used as a control voltage for an oscillator, which is arranged so that an increase of control voltage increases the frequency. A reactance valve may be used to control the frequency of a sinusoidal oscillator which in turn controls the time base, but a blocking oscillator itself has the required characteristic and is much simpler, the only complication being that it requires a low-impedance bias source.

In common with various a.f.c. circuits this system depends for its action upon a voltage being derived which is a function of phase error. In this case the permissible phase error is extremely small. The standard B.B.C. waveform allows $\frac{1}{2}$ μsec of front porch and 5 μsec of back porch. It is therefore essential that the circuit should synchronize within the limits $+\frac{1}{2}$ and -5 μsec

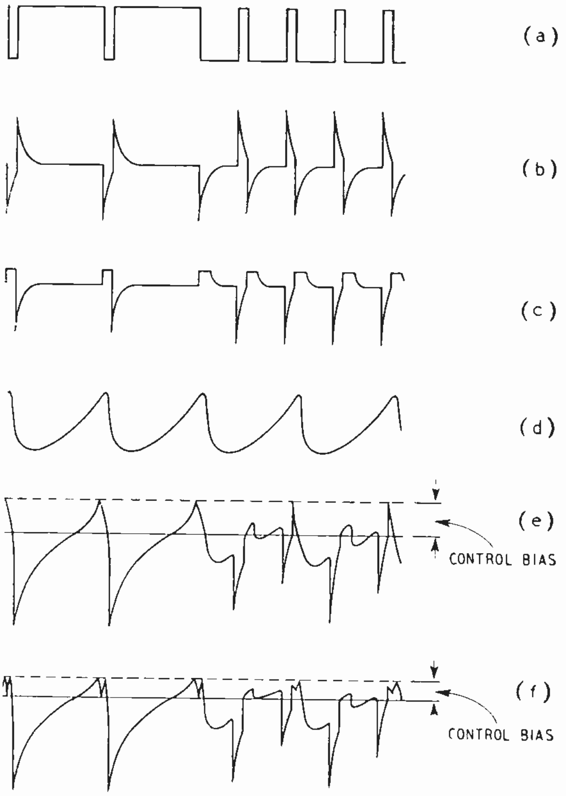


Fig. 2. *General sync pulse waveform (a), including two frame pulses. After differentiating, it takes the form (b) and subsequent clipping brings it to (c). A saw-tooth wave derived from the line time base is shown at (d) and its sum with (c) appears in (e) and (f) for two different phase conditions.*

so that the picture may be centred on the raster. This assumes that the fly-back period of the time base does not exceed 10μsec. Referring again to Fig. 2, the waveform shown at (d) has been arranged to have a very steep leading edge. The peak voltage of waveforms (e) and (f) will therefore change very rapidly with a change in phase of 1 μsec or so and if the leading edge of waveform (d) has a slope 2 V/μsec and a peak-to-peak voltage of 10 V then the maximum variation in control voltage of 10 V would occur within a phase variation of 5μsec. There is a theoretical tendency in this circuit for the time base to fire early at one extreme setting, resulting in a loss of the extreme right-hand edge of the picture, but in practice the slight delay of the synchronizing pulses due to strays in the mixing circuits is enough to offset the effect. A correctly set up circuit allows a shift of 2% in picture position on the raster.

One important requirement that must not be overlooked in the operation of these circuits is that the response during the frame-synchronizing pulses shall be the same as that during the line pulses; if not, the phasing of the line time base will be upset. It will be seen later that such a circuit will have a long time constant and may take several lines to settle down. Normally, isolated interference pulses will be unable to change conditions sufficiently to produce any disturbance to the running of the time base, but the frame pulses occupy a period of 400μsec and if they produce a change in control voltage several lines at the top of the picture may be displaced, before the conditions settle back to normal.

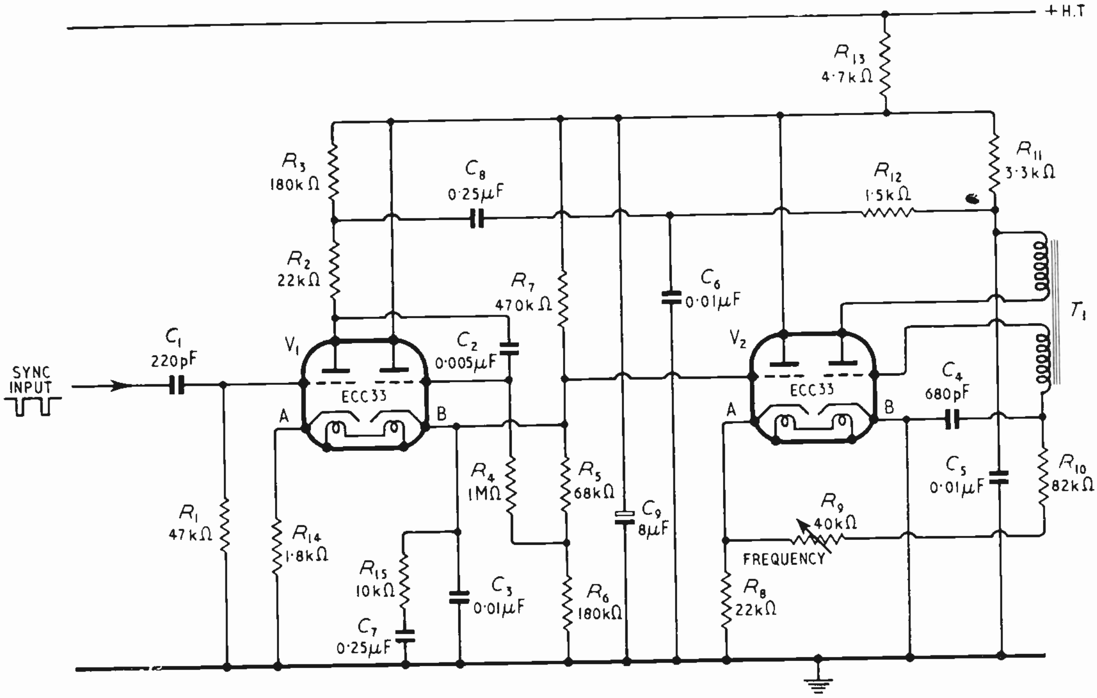


Fig. 3. *Flywheel synchronizing circuit. The blocking-oscillator transformer T_1 is a Wearite Type 232.*

This is the reason for the semi-differentiation of the synchronizing waveform. Referring again to Fig. 2, the degree of partial differentiation must be such that the positive-going pulses shown in waveform (c) must be of approximately the same duration during line pulses as they are during the frame pulses. If not they may be able to influence the output voltage of the peak-voltage detector since such a device cannot have a zero source impedance during its conduction period. If differentiation is too complete the pulses will all be too short and if incomplete the pulse length will be longer during the frame period. The frame pulses which occur at mid-scan lie in the troughs of waveform (d) and therefore do not influence the peak voltage of (e) or (f). In practice, it is impossible to achieve perfect equality though a good approximation can be effected.

Fig. 3 shows a suitable circuit. V_{2B} is a blocking oscillator giving, at the junction between R_{11} and C_5, a saw-tooth voltage waveform which, when differentiated, produces artificial negative-going synchronizing pulses suitable for synchronizing a line time base. The frequency of this blocking oscillator can either be controlled by the total value of R_9 and R_{10} or the voltage at the cathode of V_{2A}. R_{12} and C_6 lengthen the leading edge of the saw-tooth voltage, which is then fed via C_8 to the junction between R_2 and R_3. The synchronizing waveform is fed via the semi-differentiating circuit C_1 and R_1 to the grid of V_{1A} which amplifies and clips the waveform to that shown in Fig. 2(c). The waveform at the anode of V_{1A} is therefore a combination of saw-tooth voltage, which is fed through R_2, and that shown in Fig. 2(c). The practical waveform here need not be maintained so carefully as that shown in the diagrams, and the rounding of the edges due to stray capacitances has no disastrous effects on the performance of the circuit. Furthermore, it permits the use of higher anode-load resistors and a greater resulting valve gain than would be possible if faithful reproduction were necessary.

The mixed waveform is then fed to V_{1B}, which is an infinite-impedance detector, the cathode load consisting of R_5, R_6 and R_7; C_3 is the reservoir capacitor. The three resistors are used to bias this stage to an operating voltage of about 50. V_{2A} is a cathode-follower which reproduces the voltage developed by V_{1B} at low impedance to control the blocking oscillator.

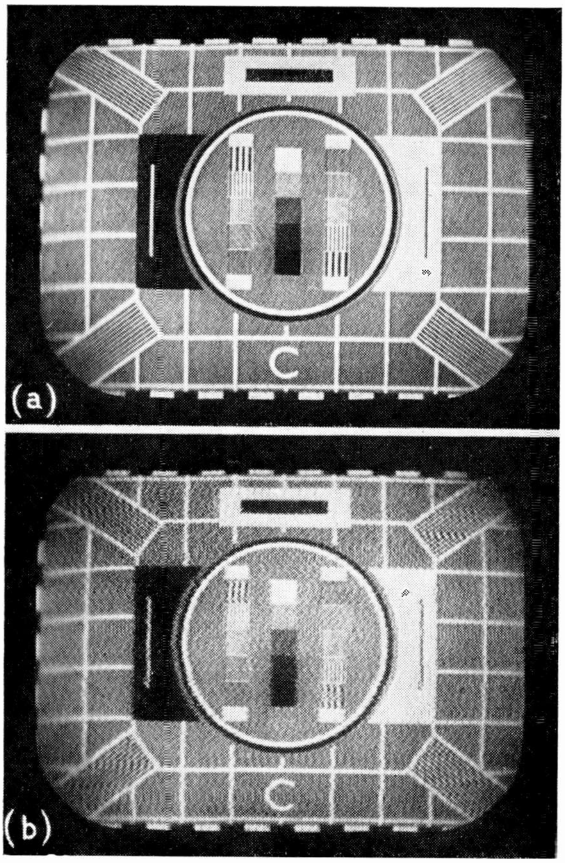


Fig. 4. *Pictures received under noisy conditions with (a) and without (b) flywheel synchronizing.*

The voltage at the cathode of V_{2A} will, in the absence of synchronizing pulses, be of the order of 50 V. In the condition shown in Fig. 2(f) it will also be of the same order, while in the condition shown in Fig. 2(e) it will be of the order of 58–60 V. The circuit will therefore hold as long as R_9 is adjusted so that the recurrence frequency of the blocking oscillator is correct at some voltage between these two limits. For example, if R_9 is adjusted so that the blocking oscillator is correct when the cathode of V_{2A} is at 54 V, then it will be necessary for the pulses of the mixed waveform to stand 4 V above the peaks of the saw-tooth waveform. Any tendency for the blocking oscillator to slow (i.e., for the saw-tooth waveform to run later) would cause the peaks to rise further up the slope, thereby increasing the control voltage, and tending to increase the frequency of the blocking oscillator.

The range of hold on R_9 is about 0 to + 15% of the total resistance value necessary to attain the correct frequency under no-sync conditions.

It can be shown (see Appendix) that this type of control circuit can have a natural frequency, usually quite low, and unless care is taken a disturbance of any sort can cause the circuit to 'hunt' around the stable condition before finally settling down. In extreme cases instability occurs and it never settles down, but runs alternately fast and slow; giving a wavy edge to the picture. R_{15} and C_7 are introduced to damp out any tendency for this to happen and the recovery from a disturbance becomes exponential; that is, it approaches the stable condition without overshooting. Usually the disturbance of the frame period is enough to excite any such instability. It might be mentioned that poor synchronizing separation is likely to cause apparent instability as the hold position then becomes a function of picture content. Furthermore, the synchronizing amplitude must not be influenced by mains ripple, as this usually causes curvature to the sides of the picture.

APPENDIX

Symbols

P	Pulse repetition frequency of blocking oscillator.
P_1	Pulse repetition frequency of synchronizing pulses.
ϕ	Phase difference in seconds between P and P_1.
V	Effective peak voltage of waveform 2(e) or (f).
V_c	Voltage at cathode of V_{2A}, assumed equal to that at cathode of V_{1B}.
C_3	Capacitance of reservoir capacitor at cathode of V_{1B}, R_{15} and C_7 being disconnected.
R	Effective source impedance of V_{1B}.

I_c Charging current to C_3.

D Differential operator. $\left(D \equiv \frac{d}{dt}; \frac{1}{D} \equiv \int dt\right)$

Since the frequency of the blocking oscillator is proportional to V_c, then:

$P = KV_c$, where K is some constant dependent on the design of the blocking oscillator.

The phase advance of the blocking oscillator, compared with the sync pulse waveform:

$$\phi = \left(\frac{1}{P_1} - \frac{1}{P}\right)t \text{ seconds per second.}$$

$$= \left(\frac{P - P_1}{PP_1}\right)t$$

If P and P_1 are very nearly the same:

$$\phi \approx \frac{P - P_1}{P_1^2}t$$

$$\therefore \frac{d\phi}{dt} \approx \frac{P - P_1}{P_1^2},$$

but $P = KV_c$,

$$\therefore \frac{d\phi}{dt} = \frac{KV_c}{P_1^2} - \frac{1}{P_1},$$

and this may be written:

$$\frac{d\phi}{dt} = K_1V_c + K_2 \qquad \ldots \quad \ldots \quad \ldots \quad \ldots \qquad (1)$$

where $K_1 = \frac{K}{P_1^2}$ and $K_2 = \frac{1}{P_2}$.

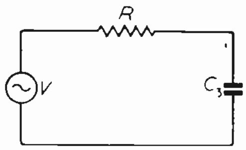


Fig. 5.

Since V, the peak voltage of waveform 2(e) or (f), decreases with a phase advance of the blocking oscillator $V = -K_3\phi$, where K_3 is a constant dependent on the slope of the steep portion of the saw-tooth waveform.

Fig. 5 shows the equivalent circuit of V_{IB}.

$$V = RI_c + \frac{1}{C_3}\int I_c dt,$$

where I_c is the current to C_3.

Using the differential operator form $\frac{1}{D} \equiv \int dt$;

$$V = \left(R + \frac{1}{C_3}\frac{1}{D}\right)I_c, \quad \therefore I_c = \frac{V}{R + \frac{1}{C_3}\frac{1}{D}}$$

$$\text{Now } V_c = \frac{1}{C_3}\frac{1}{D}I_c = \frac{V}{1 + DC_3R}$$

but $V = -K_3\phi$,

$$\therefore V_c = -\frac{K_3\phi}{1 + DC_3R} \qquad \ldots \quad \ldots \quad \ldots \quad \ldots \qquad (2)$$

Substituting (2) in (1), and continuing to use the operational form, where $D \equiv \frac{d}{dt}$;

$$D\phi = K_2 - \frac{K_1K_3\phi}{1 + DC_3R}, \quad \therefore D^2\phi = -\frac{K_1K_3D\phi}{1 + DC_3R}$$

whence

$$\left(D^2 + \frac{D}{RC_3} + \frac{K_1K_3}{C_3R}\right)\phi = 0 \qquad \ldots \quad \ldots \qquad (3)$$

The solution of (3) is:

$$\phi = A \exp \tfrac{1}{2}\left[-\frac{1}{C_3R} + \sqrt{\frac{1}{C_3^2R^2} - 4K_1K_3}\right] + B$$
$$\exp \tfrac{1}{2}\left[-\frac{1}{C_3R} - \sqrt{\frac{1}{C_3^2R^2} - 4K_1K_3}\right]$$

If $\frac{1}{RC_3} < 4K_1K_3$ the expression under the surd sign becomes negative and the solution of ϕ takes on a damped harmonic form, indicating that any disturbance to the circuit would cause ϕ to advance and retard a few times (i.e., to hunt) before settling down.

If $\frac{1}{RC_3} = 4K_1K_3$ the circuit is critically damped and settles exponentially to its steady state in the shortest possible time.

Unfortunately R is not a constant, but depends on whether V is increasing or decreasing. When V is increasing V_{IB} is supplying the charging current, R being apparently low. When V is decreasing C_3 is being discharged by R_5, R_6 and R_7, R appearing large. Careful choice of R_{15} and C_7 is, however, quite effective in making the apparent source-resistance sufficiently low to maintain critical damping.

I.E.E. TELEVISION CONVENTION

The Institution of Electrical Engineers is holding a convention on "The British Contribution to Television" during the week 28th April to 3rd May 1952. It is to cover the whole field of television in nine two-hour sessions at each of which a survey paper and a number of supporting papers will be discussed. In addition, visits to industrial and other organizations are being arranged.

The titles of the various sessions are:—

Programme Origination.
Point-to-point Transmission.
Broadcasting Stations.
Propagation.
Receiving Equipment (two sessions).
Non-broadcasting Applications.
System Aspects.

Attendance at the convention is not confined to members of the Institution, but non-members will be required to pay a registration fee.

INSTITUTE OF PHYSICS

An examination has been established by which those who do not hold a recognized university degree may satisfy the academic requirements for election to Graduateship of the Institute of Physics. Subsequently when the necessary approved experience has been acquired, transference to the corporate grade of Associate of the Institute may be effected.

Candidates for the examination must normally have followed a suitable course of study at a college or other recognized institution. The subjects are physics (three papers and a practical examination), mathematics and, either applied physics, more advanced physics, mathematical physics or statistics.

The first examination will be held during 1952 and copies of a booklet containing the full regulations and syllabus are obtainable from the Institute, 47 Belgrave Square, London, S.W.1.

FORECASTING SUNSPOT VARIATIONS TO 1957

By A. F. Wilkins, O.B.E., M.Sc., M.I.E.E.

(Communication from the National Physical Laboratory)

SUMMARY.—Published forecasts of the variation of 12-month running averages of relative sunspot numbers for the period succeeding the 1947 maximum and up to the next maximum are presented. Although the trend of sunspot number in the immediate future has been obscured by an unexpected variation in the observed values following the 1947 maximum, it is considered that the forecasts should be suitable for the purposes of planning radio-communication services for most of the remainder of the present sunspot cycle.

IN order to facilitate the long-term planning of high-frequency radio-communication services it is necessary to be able to estimate the likely annual trend of world-wide vertical-incidence critical frequency (foF2). This, in turn, necessitates an estimate of the future trend of sunspot number, for there is an almost linear relationship between the critical frequency of ionospheric regions and the sunspot number.

The lack of detailed knowledge of solar processes is such that no forecast of sunspot variations can be completely reliable. This is borne out by the large divergencies occurring in the forecasts which have been made for previous cycles and for the early part of the present cycle.

The object of this note is to present in a convenient form the forecasts which have been published for the present cycle and for the beginning of the next. No independent forecast is attempted.

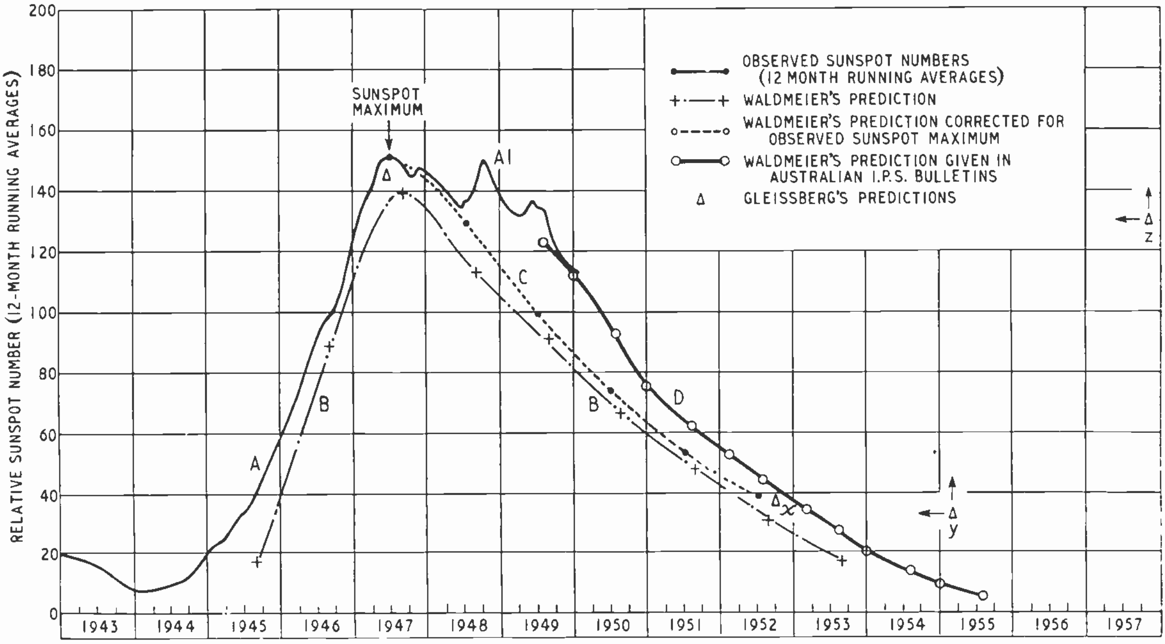
In the diagram the 12-month running averages of the Zürich relative sunspot numbers have been plotted from 1943 up to January 1950 (Curve A).

Curve B is plotted from information published by Waldmeier[1] in 1946. Substitution of the observed maximum sunspot number (151·8 in 1947·4) in Waldmeier's formulae gives rise to Curve C, the apparent accuracy of which has been greatly impaired by the appearance of the subsidiary peaks on the curves of observed sunspot number. A more recent forecast of Waldmeier's[2] is shown in Curve D.

Considerable success in forecasting the course of the present cycle has been achieved by Gleissberg[3] who has recently published a forecast

---

MS accepted by the Editor, September 1950



Variation with time of observed and predicted relative sunspot numbers.

for the next cycle.[4] His expectations are that, with a probability of 0·95, (*a*) the smoothed relative sunspot numbers will rise from a quarter of their maximum value to the maximum within less than 32 months; (*b*) the highest smoothed relative number will be greater than 130; and (*c*) that the period of low activity which will precede the next cycle will be shorter than 29 months. He also concludes that the sunspot minimum will not be a deep one.

By using Gleissberg's methods and extrapolating it is found that the time taken by the sunspot number to fall from its maximum in the present cycle (151·8) to one quarter of this value may be about 63 months. This gives us point (x) on the diagram. Gleissberg defines the period of low activity as the interval between the time at which the sunspot number drops to one quarter of its maximum in one cycle and attains one quarter of its maximum in the succeeding cycle. On the basis of this definition and on Gleissberg's expectations given above, points (y) and (z) may be plotted, bearing in mind that their positions may have to be moved in one or both of the directions shown by the arrows.

The trend of sunspot numbers has been obscured by the appearance of the bulge (A1) shown in Curve A but, from the information given, it should be possible to give estimates accurate enough for ionospheric purposes for most of the remainder of the present cycle. At the sunspot minimum period it will be seen that the forecasts are at variance and any estimate of sunspot numbers then obtaining cannot, at present, be much more than a guess.

Acknowledgment

The work described above was carried out as part of the programme of the Radio Research Board. This note is published by permission of the Director of the National Physical Laboratory and the Director of Radio Research of the Department of Scientific and Industrial Research.

REFERENCES

[1] Waldmeier, M. "A Prediction of the Next Maximum of Solar Activity." *Terr. Mag. and Atmos. Elect.*, 1946, Vol. 51, p. 270.

[2] Radio Propagation Bulletin, Ionospheric Prediction Service of the Commonwealth Observatory, Australia, August 1950.

[3] Gleissberg, W. "Probability Laws of Sunspot Variations," *Astrophys. J.*, 1942, Vol. 96, p. 234.

[4] Gleissberg, W. "The Probable Behaviour of the Next Sunspot Cycle," *Astrophys. J.*, 1949, Vol. 110, p. 90.

DIELECTRIC-LENS AERIAL

For Marine Navigational Radar

By D. G. Kiely, M.Sc., A.Inst. P.

(Royal Naval Scientific Service)

SUMMARY.—The design and performance of a dielectric-lens aerial for marine-navigational radar are described. The aerial has a fan-beam radiation pattern and is designed for horizontal polarization; its aperture and focal length are 4 ft and the maximum sidelobes are some 30 db below the main-beam level over the frequency band 9320-9500 Mc/s. This low sidelobe performance makes the aerial particularly suitable for marine-navigational radar application where the suppression of 'ghost' echoes due to sidelobes is important.

Introduction

It is particularly important that the sidelobe level in the radiation pattern of a navigational-radar aerial should be as low as possible to minimize the appearance of spurious 'ghost' echoes on the display screen. To achieve this the sidelobe level should be more than 20 db and, if possible, at least 30 db below the main-beam level.

The sidelobe amplitude in the radiation pattern is governed by the field amplitude and phase distributions across the aperture of the aerial and is related to these distributions by a Fourier transformation.[1] With a uniform phase distribution and an amplitude distribution of field given by

$$f(x) = m + \cos^2\alpha$$

where x is the distance across the aperture measured from the centre, α is of period $2B$ where B is the total aperture width and m is a constant. Maximum sidelobes 36·5 db below the main beam are predicted by theory when $m = 0{\cdot}111$ due to the existence of a state of balanced cancellation. This, however, assumes a distribution which follows the above expression exactly and if either the value of m or the shape of the distribution curve departs from the theoretical requirement the sidelobe level will increase rapidly and, allowing for an experimental approximation to the required squared-cosine distribution tapering to 10%, a more realistic figure for the attainable sidelobe level is 30 db.

In the dielectric-lens aerial the function of the lens is to create a uniform phase distribution across the aperture while the feed horn is designed to

---

MS accepted by the Editor, July 1950

have a radiation pattern which produces the required field distribution across the aerial aperture. The detailed design of these two principal parts of the aerial system will be described below.

As shipborne navigational-radar aerials are not stabilized the vertical beam width must be approximately 20° or more to prevent the target from being lost due to pitch and roll of the ship. For this reason a cylindrical lens with a small vertical aperture dimension is used and the energy flow between the feed horn and the lens is constrained between two triangular metal plates. The component parts of the aerial system may be seen in Fig. 1.

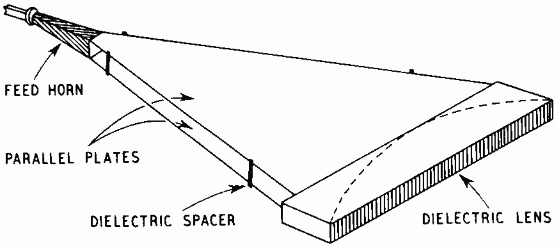


Fig. 1. *Sketch of aerial system.*

The Dielectric Lens

An unstepped cylindrical plano-convex polystyrene lens having a refractive index of 1·59 was designed. The lens height was 3 in., and its aperture and focal length were 4 ft, that is its f number (ratio of focal length to aperture) was unity. The lens was unstepped to obtain the best possible sidelobe performance as diffraction at the shadow areas of a stepped lens increases the overall sidelobe level. The lens was thus much thicker and considerably heavier than an equivalent stepped lens which would have a smaller volume of material. This was a disadvantage in its mechanical design off-set by better electrical performance. The lens height was made 3 in. to obtain large tolerances on the metal parallel-plate spacing. If the height of the aerial system is reduced to, say, 1 in. then, as horizontal polarization is used, the wave phase velocity between the metal parallel-plate system becomes critically dependent on the plate spacing and the mechanical construction of the aerial presents considerable difficulty.

Polystyrene was available in sheets of 1 in. and 2 in. thickness and several of these were cemented together to give the required lens thickness of 3 in. The cemented junctions of these sheets and the degree of homogeneity of the material had no apparent adverse effect on the radiation patterns.

The lens contour was obtained from simple ray theory. Assuming a point source O at the focus of the lens, Fig. 2, the function of the lens is to correct the phase of the radiation from O and to produce a uniform phase surface at ABD.

The lens-surface contour is obtained by considering the electric path length of any ray OPQ where P is on the contour and is designated by x, y, the co-ordinates of a rectangular system with origin at B.

Thus

$$\mathrm{OP} + \mu\mathrm{PQ} = (\mathrm{OB} - \mathrm{BC}) + \mu\mathrm{BC} = \mathrm{OA} = \text{constant}.$$

where μ is the Refractive Index;

$$\text{i.e.,}\quad \sqrt{y^2 + (\mathrm{OB} - x)^2} + \mu x = \text{constant} \quad \ldots (1)$$

This is the equation of the lens contour and it is hyperbolic. Snell's Law is automatically satisfied at the plane surface and the ray path through the lens can be immediately drawn parallel to the axis, thus simplifying the contour-determining equation.

The lens was contained in an open-fronted metal box with its top and bottom walls continuous with the parallel plates.

Primary Feed Requirements

The required amplitude distribution at the aperture of the aerial is obtained by designing the primary feed to have a certain radiation pattern, but this alone is not sufficient as the refraction at the lens will affect this radiation pattern and so also will affect the amplitude distribution at the aperture. Therefore, in order to obtain a particular aperture distribution it is necessary to know what effect the refraction at the lens has on the primary-feed radiation pattern.

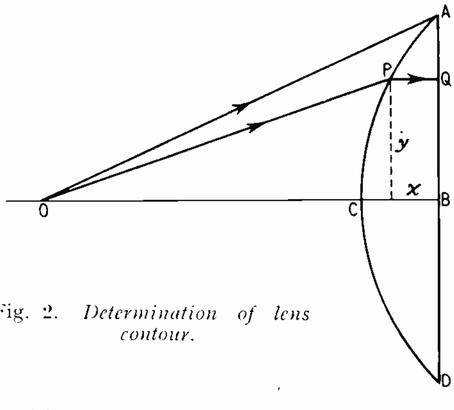


Fig. 2. *Determination of lens contour.*

Consider a point source O at the focus of a lens whose first surface is AB and a pencil of rays of small constant angle passing through the lens from O, as in Fig. 3. Assuming that all the energy is transmitted through the lens it is at once evident that the energy flow per unit area across the surface between two near points P and P′ is different from the energy flow per unit area across

QQ′ the corresponding section of the out-going beam since $PP' \neq QQ'$.

Moreover, since the distance PP′ varies along the surface AB for different values of θ and a small constant pencil, there will be a continuous variation in the energy flow per unit area across the surface.

Thus an isotropic radiator at O will not produce a uniform amplitude distribution at the lens aperture plane CD even allowing for the attenuation due to path-length differences.

Consider a cylindrical plano-convex lens (with its outer surface plane) designed to be fed by a horizontally-polarized vertical line source, the radiation from which is constrained to move between two parallel horizontal plates. As the lens is cylindrical (i.e., has no curvature in the vertical plane) we can consider a horizontal slice of unit thickness.

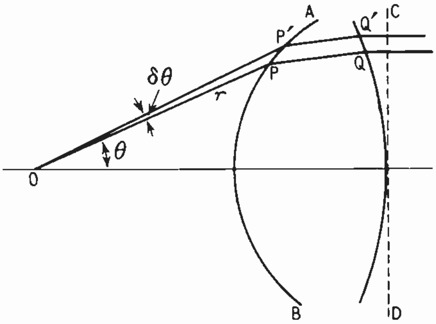


Fig. 3. *Illustrating the variation of energy flow per unit area.*

In Fig. 4, let O be a point source, $P(\tau, \theta)$ any point on the curved surface AB and P′ a point near P on AB such that PP′ subtends a small angle $\delta\theta$ at O. If δE is the total energy emitted from O in any sector of angle $\delta\theta$ the energy flow per unit area across P′N is approximately $\delta E/\tau\delta\theta$ where P′N is the perpendicular from P′ to OP.

The energy flow F_p per unit area across PP′ is, therefore, given by $F_p = \delta E \cos\alpha/\tau\delta\theta$ where α is the angle of incidence of OP upon the surface, provided P′ is taken close enough to P for the chord PP′ to approximate to the target at P.

After refraction the energy flow F_Q per unit area across the corresponding region of the aperture is given by

$$F_Q = E \cos\alpha/\tau\delta\theta\cos\beta$$

where β is the angle of refraction at P.

For a sector of angle $\delta\theta$ symmetrically placed about OC the energy flow per unit area at D

$$F_o = \delta E/R\delta\theta \text{ where } R = OC$$

The ratio $F_Q/F_o = R\cos\alpha/\tau\cos\beta$ is thus a measure of the variation in energy flow across the aperture due to refraction at the lens surface. From the geometry of the lens it will be seen that $\alpha = \beta + \theta$ and that, from Snell's Law

$$\sin\alpha = \mu\sin\beta$$

Thus α and β can both be found in terms of θ and μ

$$\alpha = \cos^{-1}\{(\mu\cos\theta - 1)/\sqrt{1 - 2\mu\cos\theta + \mu^2}\}$$

$$\beta = \cos^{-1}\{(\mu - \cos\theta)\sqrt{1 - 2\mu\cos\theta + \mu^2}\}$$

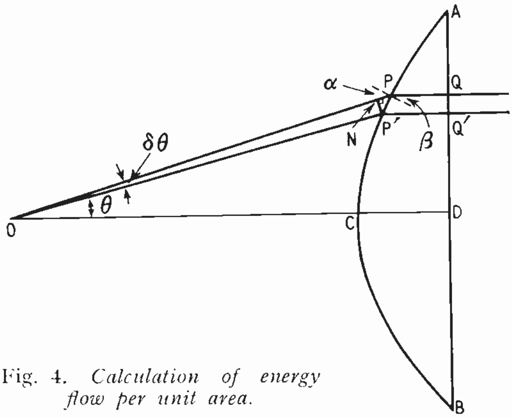


Fig. 4. *Calculation of energy flow per unit area.*

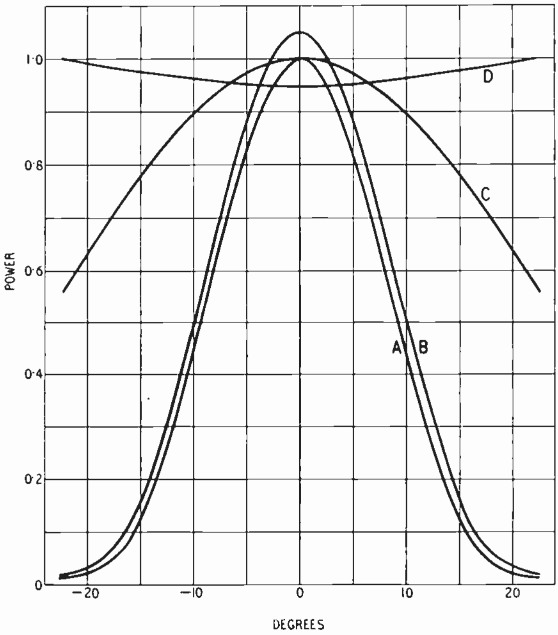


Fig. 5. *Curves: A, desired aperture distribution; B, required radiation pattern; C, correction curve; D, transmission coefficient, for dielectric lens.*

Also, from geometry, we have as the condition for an equiphase wavefront at the plane surface ADB that

$$R + \mu(D - R) = \tau + \mu(D - \tau\cos\theta)$$

$$\text{where } D = OD \quad R = OC \text{ (Fig. 4)}$$

$$\text{i.e., } R = \tau + \mu(R - \tau\cos\theta)$$

$$= \tau(\mu\cos\theta - 1)/(\mu - 1)$$

been calculated, they can be substituted in the formulae $F = \psi(L, P, S)$ or $F = \psi'(L, T, S)$, thus giving three equations with three unknowns, L, S and P or T. When the equations are solved, the fourth parameter (i.e., T or P) is assumed to be zero. But the real oscillator circuit always contains both P and T, because of the unavoidable coil, wiring and stray capacitances. Therefore, conversion formulae are used to calculate the components of the real oscillator circuit, after L, S and either P or T have been found. The real oscillator circuit, having four independent variables, has an infinite number of possible combinations of parameters. Such a domain of combinations, limited by the wiring capacitances, can be represented by curves or by scales.

Numerous articles have been published on algebraical and graphic solutions of the three equations mentioned above. The methods of solution seem to have been more or less unsatisfactory; a critical analysis of those articles is found in a paper of Kj. Prytz.[3] Moreover, in those articles only three components were calculated, the fourth being neglected.

Choice of Tracking Frequencies

The claim that the oscillator circuit and the signal circuit(s) shall be tuned with a frequency difference $F - f$ that has to be 'as constant as possible', requires a criterion in order to determine whether this claim is fulfilled. In the existing literature on the subject quite different criteria have been proposed.

M. Wald[1] pointed out that the absolute maximum value of the frequency error had to be as low as possible. In order to convert this claim into a more practical formulation, imagine that the four maximum errors ΔF_1 at f_1, ΔF_3 at f_3, ΔF_5 at f_5 and ΔF_7 at f_7 have been made numerically equal, irrespective of positive or negative sign. If now one tracking frequency is varied, the four extreme errors will change in value; some will decrease, others will increase. The latter means that the absolute maximum error in the tuning range increases. Therefore, the criterion mentioned by Wald can be formulated by the claim that the four maximum errors ΔF must be numerically equal.

A different criterion is the one, proposed by K. Fränz[4] who claims that the four maximum relative errors $\Delta F/f$ shall be numerically equal.

The third criterion has only been proposed in some articles,[3] but has not been applied to actual calculations. It claims that one should allow increasing tracking errors at the higher frequencies within the tuning range for the benefit of smaller errors at the lower frequencies, because the signal circuits are more selective with the lower frequencies. Evidently this means that the maximum attenuation (by detuning) within the tuning range, has to be as low as possible. By considerations similar to those above it is found that the four extreme values of attenuation (which is a function of ΔF/h.f. resistance) must be equal to each other.

First Criterion

M. Wald[1] has given a method for approximate calculation of tracking frequencies f_2, f_4, f_6 of a tracking-error curve with four equal maximum errors ΔF. The method was based on the approximate function

$$\Delta F = \lambda \frac{(f-f_2)(f-f_4)(f-f_6)}{f+\phi},$$

$$\text{where } \lambda = -\tfrac{1}{2}\frac{f_{\text{mean}} + f_2 + f_4 + f_6 + 2\phi}{f_{\text{mean}}^2 + \text{constant}} \quad \ldots (1)$$

and on the assumption that

$$\left.\begin{aligned} f_3 &= f_1 + 0{\cdot}25(f_7 - f_1) \\ f_5 &= f_7 - 0{\cdot}25(f_7 - f_1) \end{aligned}\right\} \quad \ldots \quad \ldots (2)$$

In many cases, however, this assumption does not hold, since f_3 and f_5 are found to deviate considerably from these values; e.g., in the example in the next paragraph, where

$$\left.\begin{aligned} f_3 &= f_1 + 0{\cdot}2(f_7 - f_1) \\ f_5 &= f_7 - 0{\cdot}3(f_7 - f_1) \end{aligned}\right\} \quad \ldots \quad \ldots (2a)$$

By substituting (1) and (2) in

$$\left|\frac{\Delta F_3}{\Delta F_1}\right| = 1\,;\ \left|\frac{\Delta F_5}{\Delta F_7}\right| = 1\,;\ \left|\frac{\Delta F_5}{\Delta F_3}\right| = 1 \quad \ldots (3)$$

Wald obtained three complicated equations, that were arranged in such a manner that $f_2 - f_1$, $f_{mid} - f_4$ and $f_7 - f_6$ were the unknowns. In order to convert these equations into linear ones that can easily be solved, all powers and products of the (relatively large) unknowns were neglected and some other approximations were made.

Consequently Wald's method results in tracking frequencies that deviate about 10 to 16 kc/s from those required and maximum errors that have 12–18% mutual difference. Moreover, the condition (2) limits the application of the method to the first criterion only.

Exact Calculation; First Criterion

The method of exact calculation to be described in this paper, starts with an arbitrary tracking-error curve that approximates to the criterion. Then three corrections (expressed in frequency) are calculated; these have to be added to the former tracking frequencies, in order to obtain the exact tracking frequencies for the ideal error curve. Since the corrections are going to be calculated as functions of the approximation properties of the arbitrary curve, it is necessary to compute its maximum errors ΔF quite exactly.

This is done by using the following formula:

$$\left.\begin{aligned} \Delta F &= \frac{(f-f_2)(f-f_4)(f-f_6)(f-f_8)}{-2\left(f+\phi+\dfrac{\Delta F}{2}\right)\left(f^2+\dfrac{\theta}{2\phi}\right)} \\ \text{where } f_8 &= -f_2-f_4-f_6-2\phi \\ \theta &= -f_2f_4f_6f_8\left(\frac{1}{f_2}+\frac{1}{f_4}+\frac{1}{f_6}+\frac{1}{f_8}\right) \end{aligned}\right\} \quad (4)$$

The factors that mainly determine the shape of the tracking error curve are $(f-f_2)(f-f_4)(f-f_6)$. Therefore, the ratio of the errors of two curves that have nearly the same tracking frequencies (viz, f_2, f_4, f_6 and f_2+c_2, f_4+c_4, f_6+c_6, where c_2, c_4, c_6 are small), can almost exactly be expressed by

$$\frac{\Delta F'}{\Delta F} = \frac{(f-f_2-c_2)(f-f_4-c_4)(f-f_6-c_6)}{(f-f_2)(f-f_4)(f-f_6)} \quad (5)$$

The ratios $\Delta F_1'/\Delta F_1$, $\Delta F_3'/\Delta F_3$, etc., can be written by putting f equal to f_1, f_3, etc. The expressions thus obtained can be considerably simplified by using the following abbreviations

$$\left.\begin{array}{lll} f_2-f_1=A & f_4-f_1=D & f_5-f_1=G \\ f_3-f_1=B & f_4-f_2=E & \ldots\ldots\text{etc} \\ f_3-f_2=C & f_4-f_3=F & f_7-f_6=W \end{array}\right\} \quad (6)$$

With these abbreviations it is found that

$$\left.\begin{aligned} \Delta F_1'/\Delta F_1 &= (A+c_2)(D+c_4)(L+c_6)/ADL \\ \Delta F_3'/\Delta F_3 &= (C-c_2)(F+c_4)(N+c_6)/CFN \\ \Delta F_5'/\Delta F_5 &= (H-c_2)(K-c_4)(Q+c_6)/HKQ \\ \Delta F_7'/\Delta F_7 &= (S-c_2)(U-c_4)(W-c_6)/SUW \end{aligned}\right\} \quad (7)$$

which can be transformed to

$$\Delta F_1' = \left(1+\frac{c_2}{A}\right)\left(1+\frac{c_4}{D}\right)\left(1+\frac{c_6}{L}\right)\Delta F_1, \ldots \text{etc.} \quad (8)$$

Substituting the four expressions (8) in three equations (9) results in fractions that can be simplified like this one:

$$\frac{1+c_2/A}{1-c_2/C} \approx \left(1+\frac{c_2}{A}\right)\left(1+\frac{c_2}{C}\right) \approx 1+\frac{c_2}{A}+\frac{c_2}{C}$$

$$= 1+c_2\frac{A+C}{AC} = 1+c_2\frac{B}{AC} \quad \ldots \quad \ldots \quad (10)$$

Thus one obtains

$$\left.\begin{aligned} \left(1+c_2\frac{B}{AC}\right)\left(1-c_4\frac{B}{DF}\right)\left(1-c_6\frac{B}{LN}\right) &= \left|\frac{\Delta F_3}{\Delta F_1}\right| \\ \left(1-c_2\frac{J}{CH}\right)\left(1+c_4\frac{J}{FK}\right)\left(1-c_6\frac{J}{NQ}\right) &= \left|\frac{\Delta F_5}{\Delta F_3}\right| \\ \left(1-c_2\frac{V}{HS}\right)\left(1-c_4\frac{V}{KU}\right)\left(1+c_6\frac{V}{QW}\right) &= \left|\frac{\Delta F_7}{\Delta F_5}\right| \end{aligned}\right\} \quad (11)$$

In these equations the products of the relatively small c_2, c_4, c_6 can really be neglected; therefore, three linear equations are obtained

$$1+B\left(\frac{c_2}{AC}-\frac{c_4}{DF}-\frac{c_6}{LN}\right) = \left|\frac{\Delta F_3}{\Delta F_1}\right|, \ldots\text{etc} \quad (12)$$

which by writing X for $\left|\dfrac{\Delta F_3}{\Delta F_1}\right|$, Y for $\left|\dfrac{\Delta F_5}{\Delta F_3}\right|$ and Z for $\left|\dfrac{\Delta F_7}{\Delta F_5}\right|$ can be converted to

$$\left.\begin{aligned} \frac{c_2}{AC}-\frac{c_4}{DF}-\frac{c_6}{LN} &= \frac{X-1}{B} \\ -\frac{c_2}{CH}+\frac{c_4}{FK}-\frac{c_6}{NQ} &= \frac{Y-1}{J} \\ -\frac{c_2}{HS}-\frac{c_4}{KU}+\frac{c_6}{QW} &= \frac{Z-1}{V} \end{aligned}\right\} \quad \ldots \quad (13)$$

By solving these linear equations the following formulae for c_2, c_4, c_6 are found:

$$\left.\begin{aligned} c_2 &= \frac{ACHS}{EMR}\left[\frac{DL}{BG}\left(X-1\right)+\frac{FW/T+KL/G}{J}\left(Y-1\right)+\frac{UW}{TV}\left(Z-1\right)\right] \\ c_4 &= \frac{FKU}{P}\left[c_2\frac{M}{CHS}+\frac{N(Y-1)/J+W(Z-1)/V}{T}\right] \\ c_6 &= QW\left[\frac{c_2}{HS}+\frac{c_4}{KU}+\frac{Z-1}{V}\right] \end{aligned}\right\} \cdot (14)$$

The tracking error curve with f_2+c_2,...etc., and $\Delta F_1'$, $\Delta F_3'$,...etc., is supposed to fulfil the claim of the criterion, viz.:

$$|\Delta F_1'| = |\Delta F_3'| = |\Delta F_5'| = |\Delta F_7'|$$

$$\text{or } \left|\frac{\Delta F_1'}{\Delta F_3'}\right| = 1\,;\ \left|\frac{\Delta F_3'}{\Delta F_5'}\right| = 1\,;\ \left|\frac{\Delta F_5'}{\Delta F_7'}\right| = 1 \quad (9)$$

A tracking-error curve with tracking frequencies f_2+c_2, f_4+c_4, f_6+c_6 has four maximum errors that are exactly equal. This is illustrated by the following example. As said in the introduction, the examples will be computed with the utmost accuracy, in order to serve as a proof of the correctness. For practical use, however, it will be found that calculation with fewer significant figures is sufficient. Let the signal circuit(s) cover the range from $f_1 = 506$ kc/s to $f_7 = 1882$ kc/s, and let the intermediate frequency be $\phi = 475$ kc/s. If the approximate calculation of M. Wald[1] had been used, the following frequencies would have been found:

$$\left.\begin{array}{l} f_2 - f_1 = 67{\cdot}5 \text{ kc/s} \\ f_{mid} - f_4 = 108{\cdot}5 \text{ kc/s} \\ f_7 - f_6 = 125 \text{ kc/s} \end{array}\right\} \text{unknowns of equations}$$

$$\left.\begin{array}{l} f_2 = 573{\cdot}5 \text{ kc/s} \\ f_4 = 1085{\cdot}5 \text{ kc/s} \\ f_6 = 1757 \text{ kc/s} \end{array}\right\} \text{tracking frequencies}$$

This tracking-error curve can be taken as the arbitrary curve in the method of exact calculation. The errors are

$$\Delta F = \frac{(f - 573{\cdot}5)(f - 1085{\cdot}5)(f - 1757)(f + 4366)}{-2\left(f + 475 + \frac{\Delta F}{2}\right)\left(f^2 + 15\,106\,100\right)} \text{ kc/s}$$

with the following extreme values

$\Delta F_1 = +\,7{\cdot}878$ kc/s at $f_1 = 506$ kc/s

$\Delta F_3 = -\,8{\cdot}070$ kc/s at $f_3 = 776$ kc/s

$\Delta F_5 = +\,8{\cdot}594$ kc/s at $f_5 = 1455$ kc/s

$\Delta F_7 = -\,9{\cdot}278$ kc/s at $f_7 = 1882$ kc/s

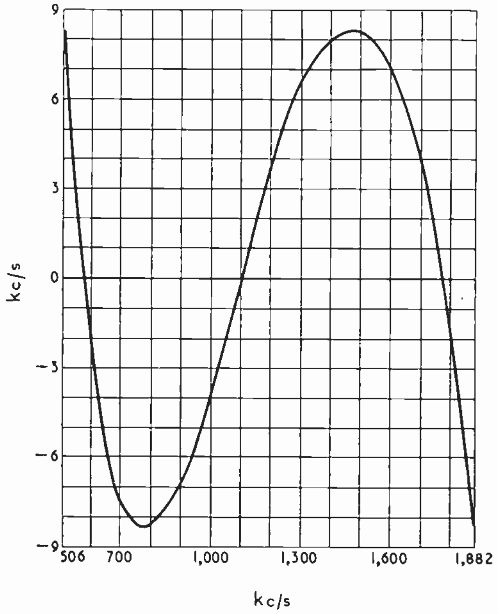


Fig. 3. *Tracking-error curve for equal maximum errors.*

From these figures the following frequency differences and error ratios are found [see (6) and the significance of symbols X, Y and Z]:

$$\left.\begin{array}{llll} A = 67{\cdot}5 \text{ kc/s} & D = 579{\cdot}5 \text{ kc/s} & & U = 796{\cdot}5 \text{ kc/s} \\ B = 270 \text{ kc/s} & E = 512 \text{ kc/s} & \ldots\text{etc}\ldots & V = 427 \text{ kc/s} \\ C = 202{\cdot}5 \text{ kc/s} & F = 309{\cdot}5 \text{ kc/s} & & W = 125 \text{ kc/s} \end{array}\right\}$$

and: $X = 1{\cdot}024\,36$ $Y = 1{\cdot}064\,87$ $Z = 1{\cdot}079\,62$

Substitution in (14) results in:

$$c_2 = \frac{ACHS}{EMR}(68{\cdot}930 + 49{\cdot}876 + 16{\cdot}785)10^{-3} \approx +\,2{\cdot}5 \text{ kc/s}$$

$$c_4 = \frac{FKU}{P}(12{\cdot}991 + 105{\cdot}811)10^{-6} \approx +\,16 \text{ kc/s}$$

$$c_6 = QW(2{\cdot}223 + 54{\cdot}757 + 186{\cdot}460)10^{-6} \approx +\,9 \text{ kc/s}$$

The tracking frequencies are thus changed to:

$f_2 + c_2 = 576$ kc/s; $f_4 + c_4 = 1101{\cdot}5$ kc/s; $f_6 + c_6 = 1766$ kc/s

The error curve with these tracking frequencies has four maximum errors that are exactly equal; viz,

$\Delta F_1' = +\,8{\cdot}31$ kc/s at $f_1 = 506$ kc/s

$\Delta F_3' = -\,8{\cdot}31$ kc/s at $f_3' = 782$ kc/s

$\Delta F_5' = +\,8{\cdot}31$ kc/s at $f_5' = 1467$ kc/s

$\Delta F_7' = -\,8{\cdot}31$ kc/s at $f_7 = 1882$ kc/s

where f_1, f_3', f_5', f_7 denote resonant frequencies of the signal circuit(s). The curve is shown in Fig. 3. ΔF is expressed here to an unusual degree of accuracy solely in order to show the exact equality of the four tracking errors.

Second Criterion

This is the criterion of K. Fränz,[4] according to which the four maximum relative errors $\Delta F/f$ should be numerically equal. The method of calculation developed in the present paper is much simpler than that of Fränz who wrote that his method "would take about half a day after one had become familiar with it."

Since a simple approximation does not exist, the tracking frequencies of the arbitrary tracking-error curve should be estimated. The proper correction values c_2, c_4, c_6 are computed in exactly the same way as described in the previous paragraph, with the following exceptions only:

1. f_3 and f_5 are those frequencies where $\left|\frac{\Delta F}{f}\right|$ is a maximum;
2. X, Y and Z denote:

$$X = \left|\frac{\Delta F_3}{\Delta F_1}\right| \cdot \frac{f_1}{f_3}; \; Y = \left|\frac{\Delta F_5}{\Delta F_3}\right| \cdot \frac{f_3}{f_5}; \; Z = \left|\frac{\Delta F_7}{\Delta F_5}\right| \cdot \frac{f_5}{f_7} \quad \ldots \quad (15)$$

The calculation is illustrated by the following example, set up for the same f_1, f_7 and ϕ. Regarding the extreme errors that shall be proportional to f, the tracking frequencies have to be changed towards the lower frequencies. Estimating $f_2 = 550$ kc/s, $f_4 = 960$ kc/s and $f_6 = 1680$ kc/s, an error curve with the following maximum values of relative error is found:

$$\Delta F_1/f_1 = + \; 0{\cdot}863\,7\% \text{ at } f_1 = 506 \text{ kc/s}$$
$$\Delta F_3/f_3 = - \; 0{\cdot}874\,2\% \text{ at } f_3 = 694 \text{ kc/s}$$
$$\Delta F_5/f_5 = + \; 0{\cdot}809\,3\% \text{ at } f_5 = 1318 \text{ kc/s}$$
$$\Delta F_7/f_7 = - \; 1{\cdot}059\,3\% \text{ at } f_7 = 1882 \text{ kc/s}$$

From these figures the following frequency differences and ratios of relative error [see (6) and (15)] are found:

$A = 44$ kc/s $\quad D = 454$ kc/s $\quad U = 922$ kc/s
$B = 188$ kc/s $\quad E = 410$ kc/s ...etc... $V = 564$ kc/s
$C = 144$ kc/s $\quad F = 266$ kc/s $\quad W = 202$ kc/s
and: $X = 1{\cdot}012\,10 \quad Y = 0{\cdot}925\,74 \quad Z = 1{\cdot}308\,86$

Substitution in (14) gives:

$c_2 \approx +1$ kc/s; $c_4 \approx +0{\cdot}5$ kc/s; $c_6 \approx +40$ kc/s

and thus

$$f_2 + c_2 = 551 \text{ kc/s}; \; f_4 + c_4 = 960{\cdot}5 \text{ kc/s}; \; f_6 + c_6 = 1720 \text{ kc/s}.$$

The correction c_6 has a rather high value; the method of calculation is, however, based on small corrections. Therefore, it may be expected that the error curve, obtained with the changed tracking frequencies, still deviates slightly from the ideal one. Truly enough it is found:

$$\Delta F_1'/f_1 = +0{\cdot}894\,9\% \text{ at } f_1 = 506 \text{ kc/s}$$
$$\Delta F_3'/f_3' = -0{\cdot}886\,0\% \text{ at } f_3' = 695 \text{ kc/s}$$
$$\Delta F_5'/f_5' = +0{\cdot}881\,2\% \text{ at } f_5' = 1338 \text{ kc/s}$$
$$\Delta F_7'/f_7 = -0{\cdot}832\,9\% \text{ at } f_7 = 1882 \text{ kc/s}$$

By repeating the correction, with the auxiliary magnitudes

$A' = 45$ kc/s, $B' = 189$ kc/s, $C' = 144$ kc/s...etc.

and

$X' = 0{\cdot}990\,04$, $Y' = 0{\cdot}994\,59$, $Z' = 0{\cdot}945\,36$

the following values are obtained:

$c_2' \approx -0{\cdot}5$ kc/s, $c_4 \approx -3$ kc/s, $c_6 \approx -7$ kc/s

and thus:

$$f_2' + c_2' = 550{\cdot}5 \text{ kc/s}, f_4' + c_4' = 957{\cdot}5 \text{ kc/s}, f_6' + c_6' = 1713 \text{ kc/s}.$$

A tracking-error curve with these tracking frequencies has four equal maximum relative errors:

$$\Delta F_1''/f_1 = +0{\cdot}878\% \text{ at } f_1 = 506 \text{ kc/s}$$
$$\Delta F_3''/f_3'' = -0{\cdot}878\% \text{ at } f_3'' = 694 \text{ kc/s}$$
$$\Delta F_5''/f_5'' = +0{\cdot}878\% \text{ at } f_5'' = 1333 \text{ kc/s}$$
$$\Delta F_7''/f_7 = -0{\cdot}878\% \text{ at } f_7 = 1882 \text{ kc/s}$$

The tracking error curve $\frac{\Delta F}{f} = \psi(f)$ is shown in Fig. 4.

Third Criterion

According to the third criterion the four maximum values of attenuation shall be made equal. The attenuation factor of each signal circuit is

$$\alpha = \sqrt{1 + \left(2\frac{\Delta f}{f}Q\right)^2} \quad \ldots \quad (16)$$

where $Q = \frac{2\pi f l}{r}$ = quality factor of coil. If p denotes $\frac{f}{2Q} = \frac{r}{4\pi l}$ = detuning that causes $\sqrt{2}$ times attenuation (about 3 db), then (16) can also be written as

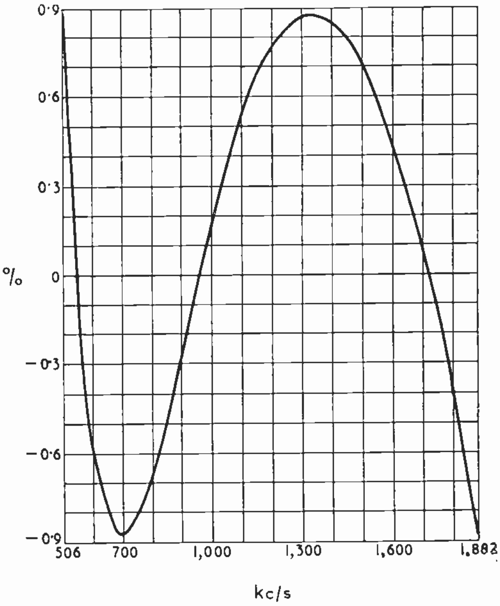


Fig. 4. *Tracking-error curve for equal maximum percentage error.*

$$\alpha = \sqrt{1 + \left(\frac{\Delta f}{p}\right)^2} \quad \ldots \quad (17)$$

The values of p as a function of f can be measured directly on the preselector circuit. Let p_1, p_3, p_5, p_7 denote the values at f_1, f_3, f_5, f_7 (i.e., at the frequencies of maximum attenuation) then the condition for the ideal tracking-error curve is

$$\frac{|\Delta F_1|}{p_1} = \frac{|\Delta F_3|}{p_3} = \frac{|\Delta F_5|}{p_5} = \frac{|\Delta F_7|}{p_7} \quad \ldots \quad (18)$$

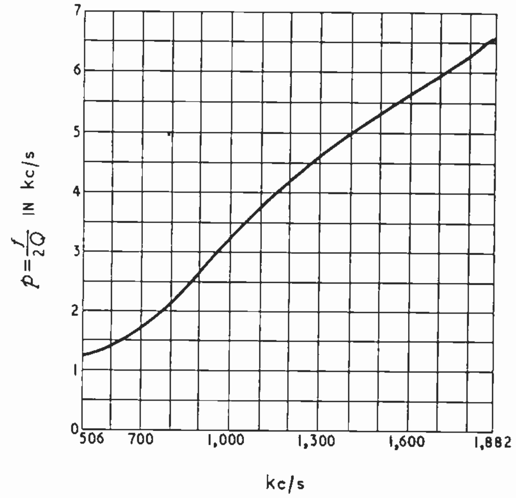


Fig. 5. *Bandwidth of signal-frequency circuit.*

The calculation of tracking frequencies can be carried out in the same manner as described in the previous cases, with the following exceptions only:

1. f_3, f_5 denote those frequencies at which $\frac{|\Delta F|}{p}$ is a maximum;
2. X, Y and Z denote:

$$X = \left|\frac{\Delta F_3}{\Delta F_1}\right| \cdot \frac{p_1}{p_3}; \; Y = \left|\frac{\Delta F_5}{\Delta F_3}\right| \cdot \frac{p_3}{p_5}; \; Z = \left|\frac{\Delta F_7}{\Delta F_5}\right| \cdot \frac{p_5}{p_7} \quad \ldots \quad (19)$$

The calculation is illustrated with an example with the same f_1, f_7 and ϕ as in the previous ones. Fig. 5 shows the results of the bandwidth measurements on the preselector circuit. p is the detuning that causes an attenuation factor $\alpha = 2^{n/2}$ in n tuned circuits (or $\alpha = \sqrt{2}$ for one circuit).

Since in practice the third criterion approximates to the second one, the tracking frequencies may, with respect to the previous examples, be estimated to be $f_2 = 550$ kc/s, $f_4 < 957$ kc/s say 900 kc/s and $f_6 < 1713$ kc/s, say 1700 kc/s. By applying the same method of calculation as before, the following corrections are found:

$$c_2 \approx -2 \text{ kc/s}; \; c_4 \approx -5 \text{ kc/s}; \; c_6 \approx +2 \text{ kc/s}.$$

Thus the tracking frequencies become

$$f_2 + c_2 = 548 \text{ kc/s}; \; f_4 + c_4 = 895 \text{ kc/s};$$
$$f_6 + c_6 = 1702 \text{ kc/s}.$$

The tracking-error curve $\Delta F = \psi(f)$ is shown in Fig. 6; the attenuation curve $\alpha = \psi'(f)$ in Fig. 7. The four maximum values of attenuation, caused by the tracking error $\Delta F'$, are equal and amount to

$$\alpha = 3{\cdot}16 \text{ at } \begin{cases} f_1 = 506 \text{ kc/s} \\ f'_3 = 670 \text{ kc/s} \\ f'_5 = 1300 \text{ kc/s} \\ f_7 = 1882 \text{ kc/s} \end{cases}$$

Calculation of Circuit Parameters

In the introduction of this paper it was pointed out that the real oscillator circuit always contains both capacitances P and T, as indicated in Fig. 1. Thus the circuit has four unknown components L, S, P and T. However, substitution of the calculated tracking frequencies f_2, f_4, f_6 in the formulae for F results in three equations only. Therefore, it is necessary to fix one of the unknowns before solving for the others. The simplest equations are found by fixing P or T, since these may become zero. In the following the circuit parameters L, S, T will be solved assuming $P = 0$.

Imagine that the inductance l of the signal circuit(s) has been computed from the given magnitudes: desired waveband, available tuning capacitor, trimmer, coil and wiring capacitances, etc. Then the values of total capacitance C_2, C_4,

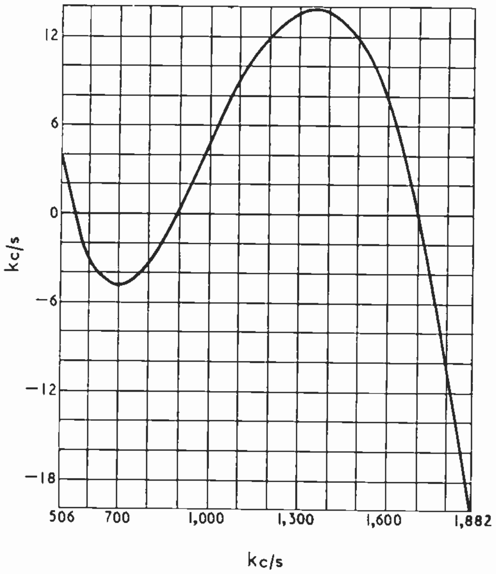


Fig. 6. *Tracking-error curve for equal attenuation at the maximum error points.*

C_6 of the signal circuit at frequencies f_2, f_4, f_6 can be computed from

$$C_2 = \frac{1}{4\pi^2 f_2^2 l}\ ;\ C_4 = \ldots\text{etc.} \qquad \ldots \qquad (20)$$

Furthermore the oscillator frequencies F_2, F_4, F_6 are known since at these frequencies $\Delta F = 0$ and, therefore, $F_2 = f_2 + \phi$, $F_4 = \ldots$etc. Thus three equations are obtained:

$$F_2 = \frac{1}{2\pi}\sqrt{\frac{C_2 + T + S}{LS(C_2 + T)}}\ ;$$
$$F_4 = \frac{1}{2\pi}\sqrt{\frac{C_4 + T + S}{LS(C_4 + T)}}\ ;$$
$$F_6 = \frac{1}{2\pi}\sqrt{\frac{C_6 + T + S}{LS(C_6 + T)}} \qquad \ldots \qquad (21)$$

By solving these equations the following equations are found:

$$\left.\begin{aligned} S &= \frac{C_4 + T}{4\pi^2 F_4^2 L(C_4 + T) - 1} \\ L &= \frac{C_2 - C_4}{4\pi^2 (F_4^2 - F_2^2)(C_2 + T)(C_4 + T)} \\ T &= \frac{C_2(C_4 - C_6)(F_4^2 - F_2^2) - C_6(C_2 - C_4)(F_6^2 - F_4^2)}{(C_2 - C_4)(F_6^2 - F_4^2) - (C_4 - C_6)(F_4^2 - F_2^2)} \end{aligned}\right\} \quad (22)$$

These formulae can be simplified considerably by introducing the magnitudes

$$\alpha = \left(\frac{f_4}{f_2}\right)^2 ;\ \beta = \left(\frac{f_4}{f_6}\right)^2 ;\ A = \left(\frac{F_2}{F_4}\right)^2 ;\ B = \left(\frac{F_6}{F_4}\right)^2 \qquad \ldots \qquad \ldots \qquad (23)$$

Thus the whole calculation becomes independent of the use of units as pF, cm, μH, mH, kc/s, Mc/s, etc. The capacitances T and S are expressed by dimensionless numbers k_1 and k_3 in relation to the tuning capacitance at the tracking frequency f_4. (Instead of C_2, C_4, C_6 according to (20), only C_4 has to be calculated). The inductance L is expressed by the dimensionless number k_2 in relation to the inductance l of the signal circuit(s). In this way the calculation formulae are:

$$\left.\begin{aligned} T &= k_1\, C_4\,,\ \text{where } k_1 \\ &= \frac{\alpha(1 - \beta)(1 - A) - \beta(\alpha - 1)(B - 1)}{(\alpha - 1)(B - 1) - (1 - \beta)(1 - A)} \\ L &= k_2 \left(\frac{f_4}{F_4}\right)^2 l\,,\ \text{where } k_2 \\ &= \frac{\alpha - 1}{(1 - A)(\alpha + k_1)(1 + k_1)} \\ S &= k_3 C_4\,,\ \text{where } k_3 = \frac{1 + k_1}{k_2(1 + k_1) - 1} \end{aligned}\right\} \quad (24)$$

As an example of this method of calculation the circuit parameters (preliminary values for $P = 0$) will be calculated for the case: $f_2 = 576$ kc/s, $f_4 = 1101{\cdot}5$ kc/s, $f_6 = 1766$ kc/s, $\phi = 475$ kc/s (this was the first example of choice of tracking points in the first part of this paper). Let the tuning capacitor have a capacitance range 11–511 pF for each section. The necessary total parallel capacitance C_o that reduces the ratio $C_{\max}/C_{\min}$ of the signal circuit(s) from 511/11 to a far lower value $\dfrac{C_{\max} + C_o}{C_{\min} + C_o}$, can be computed by means of the formula

$$C_o = \frac{C_{\max} - C_{\min}\left(\dfrac{f_7}{f_1}\right)^2}{\left(\dfrac{f_7}{f_1}\right)^2 - 1} \qquad \ldots \qquad \ldots \qquad (25)$$

With $f_1 = 506$ kc/s and $f_7 = 1882$ kc/s this will be: $C_o = 27{\cdot}96$ pF ≈ 28 pF which value includes coil winding capacitance, wiring capacitance and the adjusting (trimmer) capacitor(s) of the signal circuit(s). (Thus the total tuning capacitance, denoted by C in this paper, varies from $C_7 = 38{\cdot}96$ pF to $C_1 = 538{\cdot}96$ pF; i.e., in the desired ratio $\dfrac{f_7^2}{f_1^2}$.) The total tuning capacitance is:

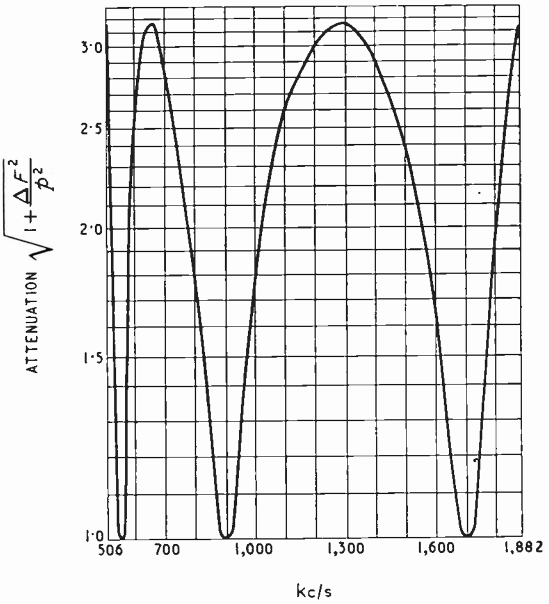


Fig. 7. *Error curve of Fig. 6 expressed in the form of attenuation through tracking error.*

$C_4 = \left(\frac{f_1}{f_4}\right)^2 C_1 = 113{\cdot}73$ pF. The inductance in the signal circuit(s) is: $l = 1/4\pi^2 f_1^2 C_1 = 183{\cdot}56\ \mu$H. The auxiliary magnitudes according to (23) are: $\alpha = 3{\cdot}656\,98$; $\beta = 0{\cdot}389\,03$; $A = 0{\cdot}444\,44$; $B = 2{\cdot}020\,67$.

These magnitudes inserted in the formulae (24) give the following results:

$$\left.\begin{array}{ll} k_1 = 0{\cdot}078\,51 & T = 8{\cdot}9 \text{ pF} \\ k_2 = 1{\cdot}187 & L = 106{\cdot}4\ \mu\text{H} \\ k_3 = 3{\cdot}848 & S = 437{\cdot}6 \text{ pF} \end{array}\right\} \text{ when } P = 0$$

As in the first part of the paper all the calculations have been carried out to more significant figures than are required in practical design work.

Parameters of the Real Circuit

As indicated in the preceding paragraph, the symbol C (with its special values $C_1, C_2, \ldots$ etc.) denotes the total capacitance in the signal circuit(s), including the parallel capacitances of coil windings, amplifier or modulator valve, wiring, trimmer capacitor, etc. In the oscillator circuit, however, these parallel capacitances are split in two portions, denoted as P and T in Fig. 1. Either P or T includes the oscillator trimmer; P includes furthermore the parallel capacitances of coil windings and oscillator valve; T includes wiring capacitance only. According to Fig. 1 and equations (21) the symbol T denotes the difference between the capacitances that shunt the oscillator section and the signal section(s) of the tuning capacitor. It is evident that this difference T can be negative, especially when the oscillator trimmer shunts L and thus is a part of P.

The calculations with formulae (22) and (24) were based on the assumption that $P = 0$. Actually, however, the oscillator circuit always contains both P and T, where P has a positive and T a negative minimum value, as explained above. It is necessary to investigate, whether an oscillator circuit with both P and T oscillates at the same frequencies at corresponding values of C; i.e., whether the real oscillator circuit can be composed of such values of components, that it is equivalent to the calculated circuit with $P = 0$.

When $P = 0$ the oscillator frequency F as a function of C is expressed by

$$F = \frac{1}{2\pi}\sqrt{\frac{C+S+T}{LS(C+T)}} \text{ or } \frac{1}{4\pi^2F^2} = \frac{CLS + LST}{C + (S+T)} \quad \ldots \quad \ldots \quad (26)$$

When $P \neq 0$, then

$$\left.\begin{array}{l} F = \dfrac{1}{2\pi}\sqrt{\dfrac{C + S' + T'}{L'[C(S'+P') + S'P' + S'T' + P'T']}} \\ \text{or} \quad \dfrac{1}{4\pi^2F^2} = \dfrac{CL'(S'+P') + L'(S'P' + S'T' + P'T')}{C + (S'+T')} \end{array}\right\} \quad \ldots \quad \ldots \quad (27)$$

where L', S', T' indicate modified values. These functions (26) and (27) are both of the general form:

$$\frac{1}{4\pi^2F^2} = \frac{xC + y}{C + z} \quad \ldots \quad \ldots \quad \ldots \quad (28)$$

The equivalence of the two cases is a fact, if the parameters L', S', T' (with P') can be chosen in such a manner that they give the same values of x, y and z as L, S, T did (with $P = 0$). Putting x, y and z in the two cases equal, the following equations are found:

$$\left.\begin{array}{l} LS = L'(S' + P') \\ LST = L'(S'P' + S'T' + P'T') \\ S + T = S' + T' \end{array}\right\} \quad \ldots \quad (29)$$

that have the following solutions:

1. if T' is given:

$$\left.\begin{array}{l} P' = (T - T')\left(1 + \dfrac{T - T'}{S}\right) \\ S' = S + T - T' \\ L' = \dfrac{L}{1 + \left(\dfrac{T - T'}{S}\right)^2} \end{array}\right\} \quad \ldots \quad (30)$$

2. if P' is given:

$$\left.\begin{array}{l} T' = T + S\left(\dfrac{1}{2} - \sqrt{\dfrac{1}{4} + \dfrac{P'}{S}}\right) \\ S' = S\left(\dfrac{1}{2} + \sqrt{\dfrac{1}{4} + \dfrac{P'}{S}}\right) \\ L' = \dfrac{L}{\left(\dfrac{1}{2} + \sqrt{\dfrac{1}{4} + \dfrac{P'}{S}}\right)^2} \end{array}\right\} \quad \ldots \quad (31)$$

(In the case of formula (31) (i.e., P' being given) it is easier to calculate S' with the aid of (30), after T' has been found.)

For the case $T' = 0$ which is theoretically possible, the formulae (30) have the simpler form:

$$\left.\begin{array}{l} P' = T\left(1 + \dfrac{T}{S}\right) \\ S' = S + T \\ L' = \dfrac{L}{\left(1 + \dfrac{T}{S}\right)^2} \end{array}\right\} \quad \ldots \quad \ldots \quad (32)$$

As an example the formulae obtained will be applied to the parameters found in the preceding paragraph. For some arbitrary values of T the other parameters are computed by means of formula (30); for $T = 0$ they are calculated by means of (32):

T	P	L	S	
8·9 pF	0 pF	106·4 μH	437·6 pF	(preceding §)
0 pF	9·1 pF	102·2 μH	446·5 pF	[formula (32)]
− 12 pF	21·9 pF	96·9 μH	458·5 pF	[formula (30)]
− 24 pF	35·4 pF	92·0 μH	470·5 pF	

As shown in Figs. 8 and 9 the result of the calculation is not one single value for each circuit element, but a domain of possible combinations. This domain is limited by the minimum value of T on one side and by that of P on the other side. If, in the present example, the wiring capacitance across the oscillator section be 5 pF, then the minimum value of T is $- 28 + 5 = - 23$ pF. The limit of P is the sum of the coil winding capacitance, oscillator valve and wiring capacitances; suppose that this capacitance amounts to 18 pF. The two limits ($T = - 23$ pF and $P = 18$ pF) are shown in Fig. 9. The capacitances indicated in Figs. 8 and 9, are total values: T is written as − 23 pF plus added trimmer-capacitance; P as 18 pF plus added capacitance.

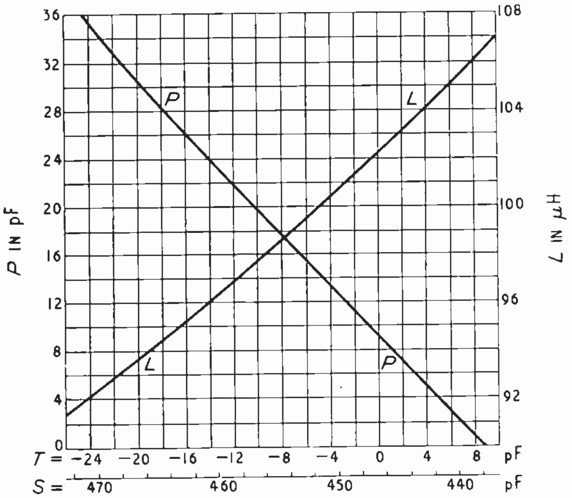


Fig. 8. *Variation of component values with distribution of capacitance between P and T.*

In order to adjust the oscillator circuit at the three tracking (trimming) frequencies F_2, F_4, F_6, it is necessary that three of the four elements be adjustable, the fourth being fixed. The usual way is to adjust L, S and P while T is at minimum value or to vary L, S and T while P has minimum value. In principle, trimming at three tracking frequencies is also possible with L, P and T adjustable and S fixed; or S P and T adjustable and L fixed. In the first cases (T or P at minimum) the fixed element is known fairly exactly and varies but little from one receiver to another. In the latter cases (S or L fixed), however, it is interesting to know with what percentage tolerance the fixed component has to be manufactured and pre-set. From Fig. 9 it is seen that in the present example S should have a capacitance within about 3% tolerance; or L should (in the

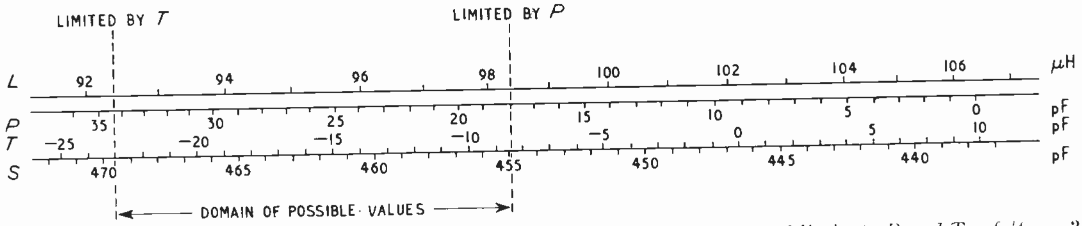


Fig. 9. *Illustrating that the domain of possible circuit values is bounded by the practical limits to P and T ;* $f_7/f_1 = 3·72$.

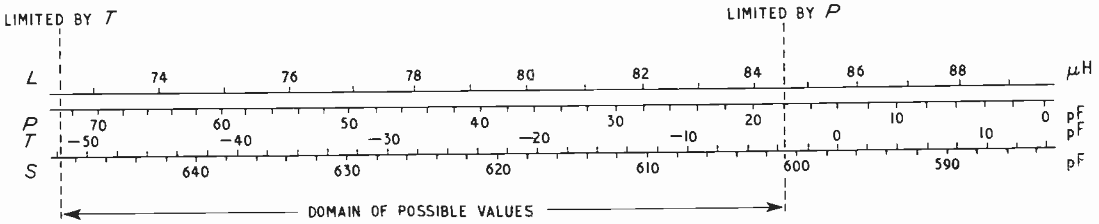


Fig. 10. *With a lower value of* f_7/f_1 (= 2·86) *the domain of possible values is much greater than in Fig. 6 and wider tolerance components can be used.*

case of fixed L) have an inductance within about 6%. These rather low tolerances are a consequence of the high ratio $f_7/f_1 = 3{\cdot}72$.

A different example with a lower ratio $f_7/f_1 = 2{\cdot}86$ may illustrate that the tolerances for S or L can approach much higher values, as 8% for S and 16% for L. Let the frequency range be from $f_1 = 542$ kc/s to $f_7 = 1548$ kc/s, and the intermediate frequency be $\phi = 445$ kc/s. A tuning-capacitor covering 11–496 pF per section requires an extra capacitance of 56·7 pF in the signal circuit(s). The total capacitance at frequency f_4 amounts to $C_4 = 166{\cdot}6$ pF; the inductance is $l = 156{\cdot}2\ \mu$H. Formulae (24) give in the hypothetical case $P = 0$:

$k_1 = 0{\cdot}082\,39 \qquad T = 13{\cdot}7$ pF

$k_2 = 1{\cdot}209 \qquad L = 89{\cdot}7\ \mu$H

$k_3 = 3{\cdot}503 \qquad S = 583{\cdot}5$ pF

For calculation of the scales diagram formulae (30) and (32) give the following results.

T	P	L	S	
13·7 pF	0 pF	89·7 μH	583·5 pF	(see above)
0 pF	14·1 pF	85·6 μH	597·3 pF	[form. (32)]
− 13 pF	27·9 pF	82·0 μH	610·3 pF	[form. (30)]
− 26 pF	42·4 pF	78·6 μH	623·3 pF	
− 39 pF	57·5 pF	75·5 μH	636·3 pF	
− 52 pF	73·2 pF	72·5 μH	649·3 pF	

The results are shown in Fig. 10. Reckoning with the same capacitances of wiring, etc., as in the first example, the limits $T = -51{\cdot}7$ pF and $P = 18$ pF are found. This permits, as stated before, a 16% tolerance of L or (in the case of fixed S) an 8% tolerance of S.

These large tolerances allow the construction of an oscillator circuit, built with a cheap fixed oscillator coil and three capacitive trimmers, which nevertheless permits adjustment at three tracking frequencies in order to obtain satisfactory results over the whole tuning range.

It is seen from Figs. 8, 9 and 10 that S and T are linear functions of each other and both L and P are nearly linear functions of S and T. Therefore formulae (30) can be simplified to:

$$\left.\begin{aligned} P' &\approx T - T' \\ S' &\approx S + T - T' \\ \frac{\Delta L'}{L} &\approx -2\frac{P'}{S} \end{aligned}\right\} \quad \cdots \quad \cdots \quad (33)$$

From the examples and from formulae (30), (31), (33) it is evident that the relative variation domain of L is about twice that of S.

APPENDIX

Derivation of formula (4); the curve of $\Delta F = \psi(f)$. From Fig. 1 it is seen that the oscillator frequency is

$$F = \frac{1}{2\pi\sqrt{L\left[P + \dfrac{S(C+T)}{S+C+T}\right]}}$$

$$= \frac{1}{2\pi}\sqrt{\frac{(S+T)+C}{L(SP+ST+PT)+L(S+P).C}} \quad (27)$$

By substitution of $C = 1/4\pi^2 f^2 l$ this can be rearranged to:

$$f + \phi + \Delta F = \sqrt{\frac{af^2+b}{f^2+c}} \quad \cdots \quad \cdots \quad (34a)$$

where

$$\left.\begin{aligned} a &= \frac{S+T}{4\pi^2 L(SP+ST+PT)} \\ b &= \frac{1}{16\pi^4 lL(SP+ST+PT)} \\ c &= \frac{S+P}{4\pi^2 l(SP+ST+PT)} \end{aligned}\right\} \quad \cdots \quad (34b)$$

Taking the square of both sides of (34a) and arranging in order of powers of ΔF:

$$\Delta F^2(f^2+c) + 2\Delta F(f^3+\phi f^2+cf+c\phi) + \\ + \{f^4+2\phi f^3+(\phi^2+c-a)f^2+2\phi cf+(c\phi^2-b)\} \\ = 0 \quad \cdots \quad \cdots \quad \cdots \quad \cdots \quad (35)$$

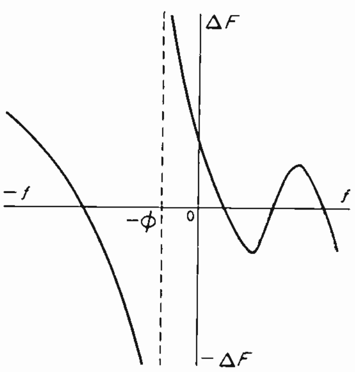


Fig. 11.

i.e., a quadratic equation in ΔF. The solution of any quadratic may be written in the implicit form without any radical and with the unknown on both sides. The solution of (35) reads in this way:

$$\Delta F = \frac{f^4+2\phi f^3+(\phi^2+c-a)f^2+2\phi cf+(c\phi^2-b)}{-2(f+\phi)(f^2+c)-\Delta F(f^2+c)} \quad \cdots \quad \cdots \quad \cdots \quad \cdots \quad (36)$$

This expression shows that the tracking-error curve can have at the most four real frequencies of zero error, since in the case $\Delta F = 0$, f is determined by

$$f^4+2\phi f^3+(\phi^2+c-a)f^2+2\phi cf+(c\phi^2-b) = 0 \quad \cdots \quad \cdots \quad \cdots \quad \cdots \quad (37)$$

or: $$(f-f_2)(f-f_4)(f-f_6)(f-f_8) = 0 \quad \cdots \quad (37a)$$

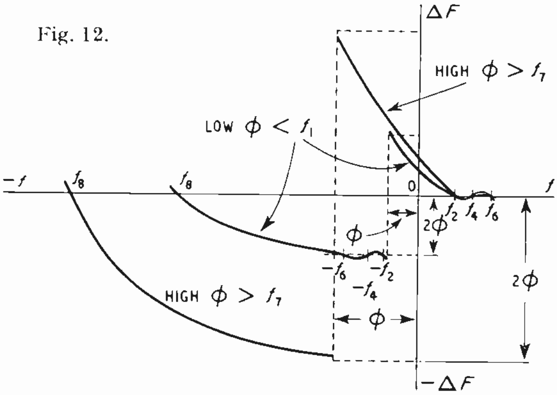
where f_2, f_4, f_6, f_8 are the roots of equation (37).

The following mathematical connections exist between the roots of (37, 37a):

$$\left.\begin{aligned} f_2+f_4+f_6+f_8 &= -2\phi \\ f_2f_4f_6+f_2f_4f_8+f_2f_6f_8+f_4f_6f_8 &= -2c\phi \end{aligned}\right\} \cdots (37b)$$

Formula (4) is obtained by writing (36) with the numerator arranged as (37a); the formulas of f_8 and $\theta = 2c\phi$ are obtained from (37b).

Fig. 12.



M. Wald[2] has published a curve of the complete function $\Delta F = \psi(f)$, including negative signal frequencies down to the point f_8 where $\Delta F = 0$. Wald's curve which is reproduced in Fig. 11, has a negative value of $d\Delta F^2/d^2f$ in the left part. It can be proved, however, that the second derivative is positive in that domain. Fig. 12 shows the correct curvature; it is drawn for two cases; viz, $\phi < f_1$(low i.f.) and $\phi > f_7$ (high i.f.).

Negative frequencies occur in Figs. 11 and 12; it should be noted that the negative domain has to be regarded only as the mathematical completion of the function $\Delta F = \psi(f)$ and without any physical interpretation.

REFERENCES

[1] M. Wald, "Ganging Superheterodyne Receivers," *Wireless Engineer* March 1940, Vol. 17, pp. 105–109.

[2] M. Wald, "Further notes on Ganging Superheterodyne Receivers," *Wireless Engineer*, April 1941, Vol. 18, p. 148.

[3] Kj. Prytz, "The Padding Condenser," *Ingeniörvidenskabelige Skrifter*, 1941, No. 3, pp. 11–19, Copenhagen, Academy of Technical Sciences.

[4] K. Fränz, "Dimensionierung der Dreipunktabgleichsschaltung beim Gleichlauf von Uberlagerungsempfängern." *Hochfrequenztechnik und Elektroakustik*, 1942, Vol. 59, pp. 144–150.

NEW BOOKS

Integral Transforms in Mathematical Physics

By C. J. Tranter. Pp. 118. Methuen, London, and John Wiley & Sons, New York, 1951. Price 6s.

This book, by the Associate Professor of Mathematics at the Military College of Science, Shrivenham, belongs to the series of 'Methuen's Monographs on Physical Subjects,' intended to supply science students at university level, and research workers in related sciences with a compact statement of the modern position in each subject. The subject here discussed is the use of integral transforms in obtaining solutions to problems governed by *partial* differential equations with assigned boundary and initial conditions.

Rigorous derivation of solutions has not been attempted; the object has been rather to show the reader what techniques are worth trying for the solution of various types of problem. Several specific examples are solved in more than one way. An extensive bibliography is included.

The integral transform $\bar{f}(p)$ of a function $f(x)$ is defined by the integral equation

$$\bar{f}(p) = \int_a^b f(x)K(p, x)dx$$

where $K(p, x)$ is a known function of p and x called the *kernel* of the transformation. In order that such a transform may be useful, it must be possible to 'invert' the transform; that is, to find $f(x)$ given $\bar{f}(p)$.

When inversion is possible, the integral transform replaces the original partial differential equation involving n independent variables by one involving $(n - 1)$ independent variables and the parameter p.

The five different kernels discussed in this book are those mainly used hitherto, namely: (a) e^{-px}, (b) sin px and cos px, (c) e^{jpx}, (d) $xJ_n(px)$ and (e) x^{p-1}, the limits being usually 0 to ∞ for (a), (b), (d) and (e), and $-\infty$ to $+\infty$ for (c), though the case when both limits are finite is discussed in Chapter 6. There is no reason why other types of kernel could not also be used. Chapter 1 deals with inversion formulae for the five kernels already mentioned, while Chapters 2–4 deal with the kernels separately. Chapter 5 deals with the numerical evaluation of integrals in solutions, and includes a method due to Willis for obtaining asymptotic series for large x for integrating functions of the type $f(p)F(x, p)$ with respect to p where F is oscillatory, and a formula due to Filon for integrating functions of the type $f(p)$ cos xp with respect to p which reduces to Simpson's Rule when x tends to zero.

In Chapter 7, the possibility is considered of using an integral transform to reduce an equation involving three independent variables to one involving two variables, and then using relaxation methods to solve the latter equation approximately.

The author claims that the technique of integral transforms can be reduced almost to a drill, whereas classical methods of solution often demand great ingenuity in assuming at the outset the correct form of solution. The book makes it quite clear that the possibility of using integral transforms to solve partial differential equations must never be forgotten, and it can be recommended mainly to those who are concerned with problems leading to such equations. The book should also be in the libraries of firms and institutions whose outlook is mainly practical as a reminder that there is really no clear dividing line between mathematics—even so-called 'pure' mathematics—and practical engineering. For the function $\bar{f}(p)$ mentioned above is initially a mathematical abstraction, but this book shows that $\bar{f}(p)$ has great practical importance.

J. W. H.

Les Hyperfréquences

By J. Voge. Pp. 317 + xii. Les Éditions Eyrolles, 61 boulevard Saint-Germain, Paris Ve, France. Price 1,980 francs.

A.C./D.C. Test Meters

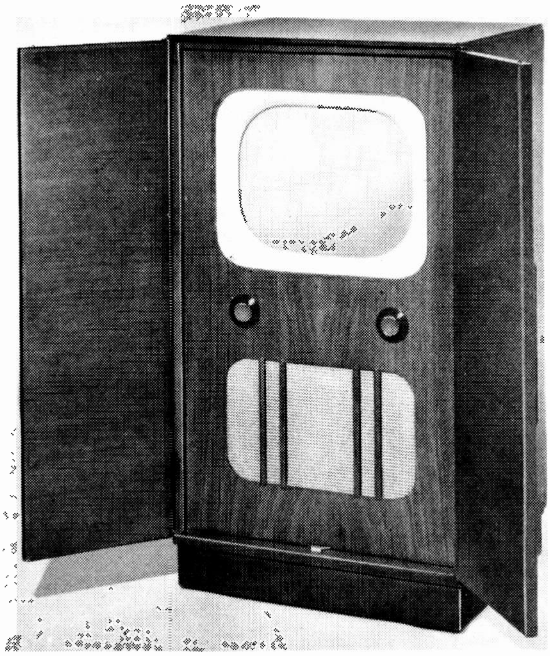
By W. H. Cazaly and Thomas Roddam. Pp. 180 + viii with 112 illustrations. Sir Isaac Pitman & Sons Ltd., Parker Street, Kingsway, London, W.C.2. Price 18s.

18TH NATIONAL RADIO EXHIBITION

Trends in Receiver Design

EVEN the most cursory glance around this year's Radio Exhibition showed that television receivers formed its most prominent section, and closer investigation revealed that they also formed a section of major technical interest. This comes about because, although the final aim of most television designers is much the same, there is still no unanimity among them about the best way of achieving it. The variations between the sets are somewhat less than in previous years, however, and there are signs that the same kind of uniformity that has been reached in broadcast receivers will appear in the television sets of the future, although probably not of the immediate future.

On the signal side there is a good deal of likeness between the different sets. The superheterodyne is preferred by most designers and probably the majority of sets have one signal-frequency amplifier and a frequency-changer which are common to both sound and vision signals. They are followed by separate sound- and vision-channel i.f. amplifiers of which the latter has two, or sometimes three, stages with a diode detector and one video stage with a diode noise limiter. Two or three sound-channel rejector circuits are fitted. On the sound side there are usually two i.f. stages with a diode detector, a diode noise limiter and a pentode output stage. Uniformity in the choice of frequency for the i.f. amplifier has not yet been achieved, but there is a tendency for it to shift upwards and a figure around 34·5 Mc/s is quite common.



Bush TUG26 with 16-in metal tube.

Television Station Selection

Although the general form of the signal channels may be very similar in different sets, this similarity does not extend to the methods used for station selection. It is the need to cater for more than a single station that is spelling the end of the straight set; the superheterodyne is usually preferred because station selection can be accomplished by changes to three circuits only instead of upwards of twelve.

Quite a few sets are provided with signal- and oscillator-frequency coils having sufficient trimming range to cover the television band; the coils are of small diameter wound on thin-walled and relatively-long formers, so that a sufficient change of inductance can be secured with a dust-iron core. In many cases the adjustment is intended to be carried out by the dealer, since it is usually needed only when an owner moves from one district to another. In one case, however (Bush), they are a user adjustment, for the coils are so arranged that the trimming controls are knobs at the rear of the set.

As an alternative, Philips use plug-in coils. They are pre-set in the factory for the various television-channel frequencies and the appropriate coils are plugged-in. Another method is to build all the coils into the set and to select the appropriate ones by altering connections or by switching. Replaceable r.f. and frequency-changer units are adopted by some firms (Murphy), while others still change the whole receiver unit; among these are naturally those who still adhere to the straight set; for example, the G.E.C. B.T.5145.

It is on the time-base side that the major differences arise, but even here uniformity of practice is on its way. There is a very general tendency to use 'economy' line-scan circuits with h.t. boost, and to derive the e.h.t. supply from the line flyback. The forms of the circuits differ very considerably, however. The present tendency is towards the use of larger and shorter c.r. tubes so that bigger pictures can be obtained and the equipment still housed in a cabinet of moderate depth. The angle through which the c.r. beam must be deflected has thus increased and now tends to be about 70° instead of only 50°. The increase of picture size—12-in. tubes are common, while 15-in. and 16-in. are becoming so, and there was even one set (H.M.V.) with a 21-in. tube—has necessitated an increase of operating voltage to maintain picture brightness. The tendency to use tinted 'safety glass,' which permits greater contrast in daylight viewing, has also resulted in a need for basically brighter pictures in order to offset the loss of light in the tinted 'glass.' In fact, the material is usually a plastic and is often moulded to the contour of the tube face.

Because of this, the present practice is to operate at 12–14 kV for 15- or 16-in. tubes and at about 10 kV for 12-in. types. The old heavily-damped scanning circuits, which sufficed a few years ago, are now quite inadequate.

Line-Scan Circuits

The so-called economy circuits[1,2,3] are much more efficient because, instead of supplying the energy for deflection afresh for every scanning line and dissipating it during the flyback, a large proportion of the energy is recovered and utilized again. In fact, the energy supplied need only be sufficient to provide for the unavoidable copper and iron losses in the coils and the anode dissipations of the valves. The circuits are used to control large amounts of circulating energy and do it with quite a small loss. Basically, a pentode V$_1$ (Fig. 1) is used to control the supply of energy to the circuit, and it does this by driving current through the deflector coil via an auto-transformer. During the flyback V$_1$ and V$_2$ are cut-off. The inductance of the circuit resonates with the self-capacitance and the energy stored in the magnetic field is converted first to electric energy in the capacitance and then back to magnetic energy in one-half cycle of free oscillation at around 35–50 kc/s.

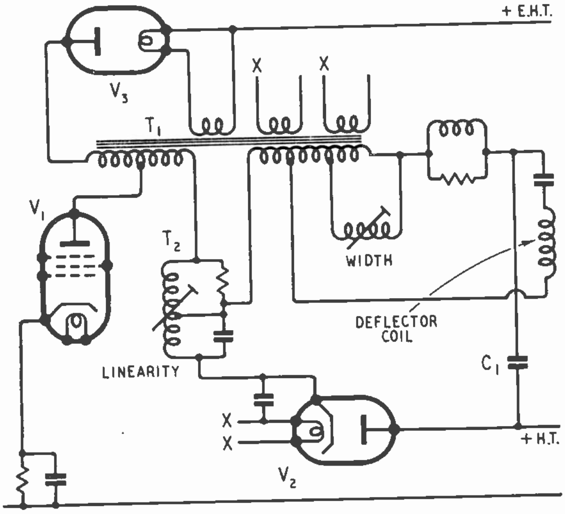


Fig. 1. *Simplified circuit of the line time-base output stage of the Pye FV1 receiver. The heater of the diode V$_2$ is connected in series with those of other valves but through bifilar windings on T$_1$ in order to prevent the appearance of a high pulse voltage between heater and cathode.*

When in electric form the energy produces a high voltage[4] which is stepped up by auto-transformer action to provide the e.h.t. supply of 9–14 kV through the rectifier V$_3$. The reservoir capacitor is not shown in Fig. 1 since it is provided by the capacitance between internal and external conducting coatings on the cathode-ray tube itself. Flyback e.h.t. is obviously extremely economical of material, since nothing but a rectifier and a transformer winding is needed.

At the end of flyback when the energy is back in magnetic form, V$_2$ comes into operation to inhibit further oscillation and to control the further decay of current. This now takes place linearly and V$_1$ need only start to drive again when the circuit has fallen almost to zero. The current through V$_2$ is used to charge a capacitor C$_1$ which is connected in series with the h.t. supply to V$_1$. The voltage developed across it thus augments the h.t. supply voltage. This h.t. boost may amount to several hundred volts, although it is not always as much as this. It is an extremely important development and one which has made possible the a.c./d.c. television set; in such sets the h.t. supply is limited to about 190 V and scanning at such a low voltage would be an expensive matter without a boost voltage.

Circuits of this nature demand low-loss components and Ferroxcube is finding great application for the transformer core and even for the 'iron circuit' of the deflector coil. It is necessary that the mean currents of V$_1$ and V$_2$ should be the same and the transformer ratio is chosen to make them so. The ideal condition is a unity ratio, for then the boost voltage is a maximum, but this demands that there should be no loss of energy during flyback—an impossible condition.

In order to obtain both linearity of scan and high efficiency the diode must be controlled. Two ways of doing this are found in present-day sets. In Fig. 1 the transformer T$_2$ is energized by the current of V$_1$ and, being a resonant transformer, it develops a control voltage for V$_2$ throughout the scan. The second method which is just coming into use, utilizes an inductance in series with the deflector coil; this inductance has a Ferroxcube core which is operated under saturated conditions. The result is to make the voltage across it fall as the scanning current increases and this offsets the increasing voltage drop across the winding resistances. In some cases an open core is used and saturated by a permanent magnet; in others a closed core is employed and saturated by a winding carrying direct current.

A further development of these circuits, which does not yet seem to have found its way into production sets, was shown by Mullard and is indicated in Fig. 2. No line-scan transformer is needed. An extra coil L$_1$ is included in the anode circuit of V$_1$ and some of the energy stored in this is communicated to the deflector coil during flyback to offset its resistance losses. As a result, the pentode and diode mean currents can be equalized without a transformer. In effect, there are two resonant circuits coupled by C$_1$. A saturated inductance L$_4$ is used for a linearity control. The coil L$_1$ is provided with an extra winding to give a step-up of voltage and enable e.h.t. to be obtained. In view of its simplicity the circuit is one which may well find its place in next year's receivers.

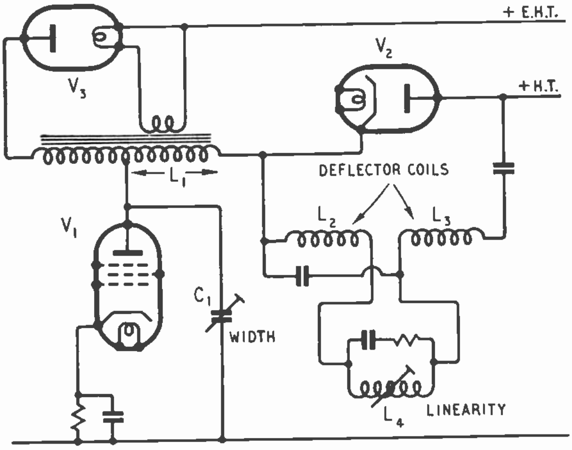
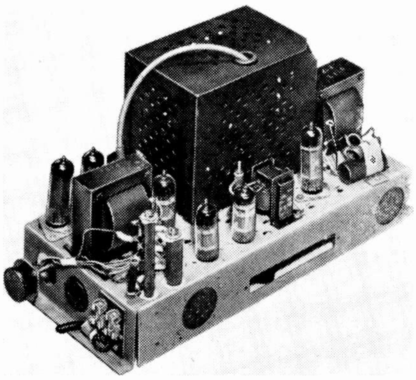


Fig. 2. *Mullard circuit in which no line-scan transformer is needed. Equality of mean pentode and diode currents is achieved by coupled circuit action.*

The output valve of the time base has, of course, to be driven and the general practice is to do so with a saw-tooth voltage derived from a blocking oscillator. There are, however, many alternatives. The output valve can be made to generate its own grid voltage in a so-called self-oscillating circuit and such arrangements are more common this year than before. There are several varieties of circuit. In one used by G.E.C., Invicta and some Pye models there is a transformer between the control and screen-grid circuits to provide the self-oscillating action. In others (Baird, for instance) an extra winding on the output transformer is connected in the control-grid circuit to provide positive feedback.

Frame-scan circuits exhibit much less variety and consist commonly of a triode blocking oscillator with a pentode output valve feeding the deflector coil through a transformer. The differences arise chiefly in the methods used to obtain linearity of scan. There are two main methods—pre-distortion of the input waveform to the pentode by adding an integrated component to the saw-tooth wave and negative feedback through an RC network having the form of a differentiator and an integrator in series.



Time-base and power-supply chassis of Philips television receiver.

One development which may again be a foretaste of the future appears in the Ekco T165. It is spot-wobble.[5] In order to reduce the visibility of the scanning lines a small vertical deflection at high frequency is given to the spot. A small extra pair of deflector coils is fitted between the normal deflector coils and the focus coil and is fed from a 11·5-Mc/s oscillator. In effect, the scanning spot is made taller than it is wide and it certainly removes the line structure with little or no effect on the definition.

Mention has already been made of the a.c./d.c. receiver. Largely because of the low h.t. supply obtainable this form is only favoured by a few manufacturers and many more adopt a similar technique which is suited to a.c. operation only. The h.t. supply is taken almost directly from the mains through a half-wave rectifier and smoothing circuit, and the valve heaters are series-connected. An auto-transformer is included, however, and both valve heaters and h.t. supply are fed from fixed tapping points on it. The mains-voltage adjustment is provided by means of tappings. In this way a 250-V supply can be obtained from all mains and the component is much smaller and cheaper and has less stray field than a double-wound transformer. An unusual feature found in the Murphy V202C is the use of a full-wave bridge rectifier.

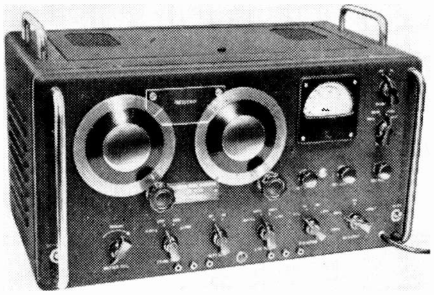
Projection television equipment has not greatly changed during the year. All sets are based on the Philips' scheme using a $2\frac{1}{2}$-in. tube operating at 25 kV with a folded optical system of the Schmidt type. The e.h.t. supply is developed from a pulsed choke through a voltage-tripler rectifier and with a new stabilizer is claimed to have a constant output voltage for currents up to 300 μA.

Test Equipment

Television has brought in its train the need for more elaborate test equipment. A simple method of measuring voltages up to 30 kV is based on the separation required between two spheres for sparking to commence and a 'voltmeter' operating on this principle has been produced by Waveforms, Ltd. The spheres are mounted inside a tube. This same firm has developed quite an elaborate signal generator in which eight different modulation waveforms are available to give various patterns of horizontal and/or vertical bars. On the r.f. side it provides outputs simultaneously at the vision and sound carrier frequencies, and both are tunable to any channel in the television band.

The Murphy TP911 generator is now available in models designed for 525- and 625-line systems as well as the original 405 lines. It provides an r.f. output modulated by a bar pattern and sync pulses which conform to the specification for television transmitters.

Frequency-modulated oscillators (wobbulators) for visual alignment of r.f. and i.f. circuits are widely used in factory production. In many cases they are used for initial alignment and the final adjustments are done with a pulse-modulated generator. Here the pulse output of the receiver is viewed on a c.r. tube and adjustments are made so that the rise time, overshoot and shape of the top are all within prescribed limits. So far as is known such pulse methods have not yet made their way into the field of the serviceman, but the wobbulator has, although in simpler form than the factory type. As an example, the Taylor model 260A is self-contained with its own c.r. tube and gives a sweep adjustable from 0–5 Mc/s at a carrier frequency from 5 Mc/s to 70 Mc/s. The sweep frequency is 50 c/s and an output up to 10 mV is available. A frequency marker is included.



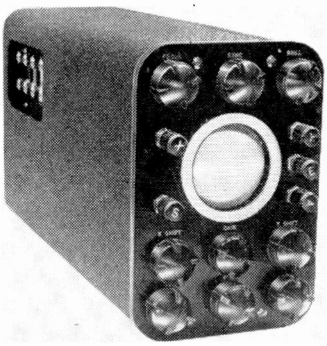
Waveforms television signal generator, type W.90.

The oscilloscope is, of course, an indispensable tool for all work on time bases and the double-beam type is particularly useful in enabling the waveforms in two different parts of a circuit to be observed simultaneously. The Cossor Model 339B is a fairly simple type, and in addition to Y amplifiers for electric deflection, magnetic deflection coils are included.

The size and weight of most oscilloscopes makes them inconvenient for the serviceman and a miniature model has been introduced by John Bell & Croyden. Its size is approximately 3½ in. by 5 in. by 9½ in. and it has a 1¾-in. tube. The frequency response is stated to be 10 c/s to 3 Mc/s and the sensitivity 50–100 mV(r.m.s.)/cm. The time-base frequency range is 10 c/s to 50 kc/s.

Sound Broadcast Receivers

After television, broadcast receivers form the main part of the exhibition but, technically speaking, there is no major trend in development to report. There are, of course, individual receivers which are of especial technical interest because of their deviations from normal practice. There is a general tendency towards a reduction in size of receivers, which applies in all sections and arises largely through the widespread adoption of miniature valves. There is, too, evidence of the growing popularity of forms of portable receivers. The mains/battery set was well in evidence and the 'all-dry' portable is still popular. Here there is some tendency towards an increase of size in order to accommodate larger batteries. The so-called 'personal portable,' for instance, is tending to increase more towards a slim attaché-case form.



Belclere miniature c.r. oscilloscope.

An unusual feature of the Ferguson 300 radio-gramophone is a cathode-coupled push-pull output stage. The preceding pentode amplifier is directly coupled to the stage and has its screen fed from a tapping on the cathode-coupling resistor. Some 30-db negative feedback is applied and an output of 3 W for 0·05% distortion is claimed and 6 W for 1%.

Quite a few sets exhibited were export models. An interesting development here is the use of an earthed-grid triode on short waves, as in the H.M.V. models 5312 and 5412 and Marconiphone T28 series. Another export-model trend is towards operation from a 6-V accumulator; Bush, Invicta and Marconiphone all have such models. One very unusual feature to be found in the Ace Radio 'Selector' is the use of a tuning capacitor of only 187 pF. This has necessitated splitting the medium-wave band into two ranges of 130–275 m and 270–570 m. The set covers long waves and one short-wave band of 16–33 m. The export version omits the long-wave band, but has five short-wave bands.

All apparatus depends in one way or another on the valve. For some years the trend has been towards miniaturization and it still persists. The characteristics, so far from suffering, have improved, and there are quite a few special types designed primarily for television. These generally are intended for use with series-connected heaters to meet the demands of a.c./d.c. technique and include such types as the Mullard PL81 with a peak anode-current rating of 350 mA and a peak anode pulse rating of 7 kV, and the Brimar 6U4GT damping diode with a peak *heater-cathode* voltage rating of 3·85 kV. In addition, there are triode-pentodes suitable for frame time-base use among other things.

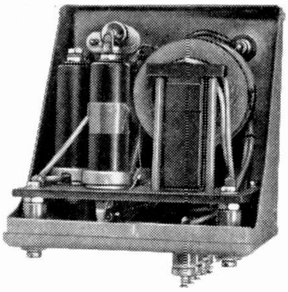
Twinvicta Model 55 portable receiver.

In cathode-ray tubes aluminizing is now common, but many firms as well or instead fit ion traps. As already mentioned the tendency is to larger screens and deflection angles and, to reduce weight, several firms have produced so-called metal tubes in which the flare portion is of metal, among them are English Electric and Mullard. In order to reduce the space occupied by the tube Brimar have produced models with rectangular screens.

A foretaste of the future appeared in the G.E.C. broadcast receiver using germanium crystal triodes. This is purely an experimental model to illustrate the capabilities of the crystal valve; this valve is stated to operate at frequencies up to 10 Mc/s and a pair in push-pull will give an audio-frequency output of 100 mW operating from a 70-V h.t. supply with a current consumption for the complete receiver of 10 mA. It is not suggested that the crystal triode is likely to displace the thermionic valve in broadcast equipment for many years, but when the production problems are solved it has obvious advantages in hearing aids and in computers.

Components

In addition to valves, all receivers naturally depend upon components. In most cases development proceeds steadily and the changes which occur are often far from obvious ones. This is especially so with such things as capacitors and resistors. The general trend is towards a reduction of size and improved sealing so that the initial insulation is maintained over long periods in spite of adverse climatic conditions. T.C.C. have a method by



Interior of Haynes Radio oil-filled line-scan transformer and e.h.t. supply unit.

which capacitors of more-or-less normal construction can be sealed by a plastic immersion moulding (Plimoseal). Such capacitors are claimed to withstand 100% humidity at 100°C. The problem of insulation becomes particularly acute in the high-voltage parts of television sets and there appear to be two schools of thought. One favours oil immersion. The whole line-scan transformer is immersed in an oil-filled can, sometimes alone, but sometimes with the e.h.t. rectifier and even reservoir capacitors. The other school adopts wax impregnation for the windings; in addition to impregnation, however, the wax is built up thickly around the high-voltage parts so that it forms, as it were, a tyre around the coil. The aim is to reduce corona.

In resistors the problem of heat radiation is an important one, and in the Dubilier BT insulated types the end wires penetrate a considerable distance inside the element in order to bring out the heat by thermal conduction.

Television aerials are in the main unchanged but, as an alternative to the Yagi form for fringe-area reception, simplified forms of broadcast array are appearing. Antiference, for instance, have one in which two X aerials are mounted side by side, $\lambda/2$ apart and fed in phase. Generally speaking, however, the H and X forms predominate, but the Yagi types with one or two directors are great favourites when increased gain is wanted. Such aerials, of course, require to be strongly made if they are to resist wind pressure, and Belling-Lee claim that their 4-element Multirod will withstand gusts up to 80 m.p.h.

REFERENCES

[1] "Magnetic-Deflection Circuits for Cathode-Ray Tubes," by Otto H. Schade, *R.C.A. Review*, September 1947, p. 506.

[2] "Characteristics of High-Efficiency Deflection and High-Voltage Supply Systems for Kinescopes," by Otto M. Schade, *R.C.A. Review*, March 1950, p. 5.

[3] "Efficiency Line-Scan Circuits," by W. T. Cocking, *Wireless World*, August, September and October 1951, pp. 302, 347 and 425.

[4] "Flyback E.H.T.," by W. T. Cocking, *Wireless World*, August and September 1950, pp. 279 and 313.

[5] "Spot Wobble," by R. W. Hallows, *Wireless World*, March 1950, p. 84, "More About Spot Wobble," by T. C. Nuttall, *Wireless World*, May 1950, p. 189.

CORRESPONDENCE

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

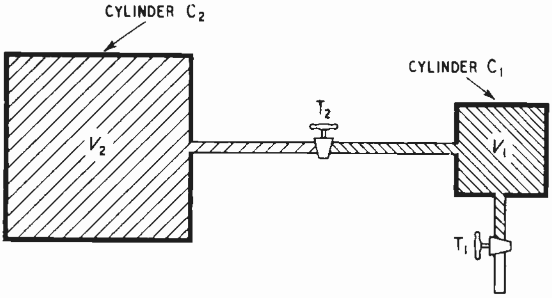
Stepping-Counter Analogy

SIR,—An interesting analogy exists between the simple stepping counter, as, for instance, described in Mr. Wintle's paper ("Precision Calibrator for L.F. Phase-meters," *Wireless Engineer*, July 1951), and the following 'gas-law' problem.

Two cylinders, C_1, C_2, of volumes V_1, V_2, are joined as shown in the figure. Each cylinder contains, initially, a certain mass of air, such that the pressure in C_1 is less than P_1, the atmospheric pressure, and the pressure in C_2 is P_2, where $P_2 \ll P_1$. The cylinder system is now subjected to a series of cycles, each consisting of:—

(a) Closing T_2, opening T_1 ('charging' C_1);

(b) Closing T_1, opening T_2 (sharing the charge between C_1 and C_2).



The first cycle starts when, with T_2 closed, T_1 is opened. It is assumed that the operations are performed at constant temperature and that sufficient time is allowed for a stable condition to be reached after each operation.

(a) T_2 closed, T_1 is opened. Air enters C_1 until the pressure is P_1. Then,

$$P_1V_1 = m_1RT \quad \ldots \quad (1)$$

where m_1 is the mass of air in C_1, R the gas constant for unit mass of air and T the absolute temperature. Similarly, for the isolated cylinder C_2,

$$P_2V_2 = m_2RT \quad \ldots \quad (2)$$

where m_2 is the initial mass of air.

(b) T_1 is closed, T_2 is opened. Air flows into C_2 to equalize the pressure. If the system pressure at the end of the 1st cycle is denoted by $(P)_1$, then,

$$(P)_1(V_1 + V_2) = (m_1 + m_2)RT \quad \ldots \quad (3)$$

$$\text{Therefore, } (P)_1 = \frac{P_1V_1 + P_2V_2}{V_1 + V_2} \quad \ldots \quad (4)$$

Subsequent cycles raise the system pressure by ever decreasing amounts, so that after p cycles:—

$$(P)_p = \frac{1}{V_2}\{P_2V_2n^p + P_1V_1(n + n^2 + n^3 + \ldots + n^p)\} \quad (5)$$

$$\text{where} \quad n = \frac{V_2}{V_1 + V_2} \quad \ldots \quad (6)$$

Equation (5) may be compared with the step-function voltage e_p of the simple stepping counter, given by Mr. Wintle as:—

$$e_p = \frac{1}{C_{14}}\{qn^p + Q(n + n^2 + n^3 + \ldots + n^p)\} \quad (7)$$

$$\text{where} \quad n = \frac{C_{14}}{C_{13} + C_{14}} \quad \ldots \quad (8)$$

The analogy between the corresponding symbols of equations (5) and (7) and the modes of operation of the electrical and mechanical circuits is clearly evident.

It is interesting to calculate the 'time-constant' of the mechanical system for the simple case $P_2 = 0$. Then,

$$(P)_p = \frac{P_1 \; V_1}{V_2}\frac{n(1 - n^p)}{(1 - n)} \qquad \text{since } n < 1$$

$$\text{Or, } (P)_p = P_1(1 - n^p) = P_1(1 - \epsilon^{-ap}) \quad \ldots \quad (9)$$

$$\text{where} \quad a = -\log_\epsilon n \quad \ldots \quad (10)$$

$$\text{Therefore, 'time constant'} = \tau = \frac{1}{-\log_\epsilon n} \quad \ldots \quad (11)$$

For $n = \frac{1}{\epsilon}$, $\tau = 1$. In general, for τ to have integral values 1, 2, 3, ... s; $n = \frac{1}{\epsilon}, \left(\frac{1}{\epsilon}\right)^{\frac{1}{2}}, \left(\frac{1}{\epsilon}\right)^{\frac{1}{3}}, \ldots . \left(\frac{1}{\epsilon}\right)^{\frac{1}{s}}$.

Whitley Bay,
Northumberland
28th July, 1951

B. BERGER.

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 the titles of journals are taken from the World List of Scientific Periodicals. Titles that do not appear in this List are abbreviated in a style conforming to it.

ACOUSTICS AND AUDIO FREQUENCIES

016 : 534 **2309**

References to Contemporary Papers on Acoustics.—A. Taber Jones & R. T. Beyer. (*J. acoust. Soc. Amer.*, May 1951, Vol. 23, No. 3, pp. 377–385.) Continuation of 1812 of August.

534 + 621.395.61/.62](083.71) **2310**

Standards on Electroacoustics: Definitions of Terms, 1951.—(*Proc. Inst. Radio Engrs*, May 1951, Vol. 39, No. 5, pp. 509–532.) Reprints of this Standard, 51 IRE 6 SI, may be purchased while available from the Institute of Radio Engineers, 1 East 79 Street, New York 21, N.Y., at $1.00 per copy.

534.213-14 **2311**

Random Noise in an Attenuating Fluid Medium.—R. E. Roberson. (*J. acoust. Soc. Amer.*, May 1951, Vol. 23, No. 3, pp. 353–358.) "It is assumed that acoustic background noise is caused by a distribution of random 'white' noise sources whose physical mechanism is unspecified. A law expressing the amplitude-distance attenuation characteristic of the medium is also assumed. Several distributions of noise sources are considered: uniform volume distributions, uniform surface dipole distributions, and two mixed cases. The noise drop-off with frequency at a point below the surface is found for each case. For an infinite volume of noise sources, this drop-off is 6 db/octave at all frequencies. It is shown how this simple model can be generalized to other attenuation laws and other spatial and amplitude distributions of noise sources."

534.232 **2312**

The Emission of Sound by a Piston.—D. N. Chetaev. (*C. R. Acad. Sci. U.R.S.S.*, 21st Feb. 1951, Vol. 76, No. 6, pp. 813–816. In Russian.) Integral (1) determining the amplitude of the velocity potential of steady-state sound oscillations is considered, and the sound field at an arbitrary point P determined. A formula (5) determining the radiation resistance is also derived.

534.232 : 538.652 : 621.314.212 **2313**

Electroacoustic Transformation by means of Magnetostriction, with Special Reference to Radiation from Transformers.—H. H. Rust. (*Z. angew. Phys.*, Dec. 1950, Vol. 2, No. 12, pp. 487–491.) Magnetostriction curves, obtained with Ni, Fe and Fe/Si oscillators, and taking account of magnetostrictive hysteresis, indicate complex oscillations rich in harmonics. These results apply directly to transformers, which in normal operation must incidentally radiate such magnetostrictively excited mechanical waves, with adverse effects on the properties of any oil used for insulation. Suggestions are made for eliminating these harmful effects.

534.24 + 534.373 **2314**

Scattering and Absorption by an Acoustic Strip.—A. Levitas & M. Lax. (*J. acoust. Soc. Amer.*, May 1951, Vol. 23, No. 3, pp. 316–322.) An analytical method is described for determining the scattering and absorption of sound for a nonuniform-boundary case represented by the application to an infinite wall of a finite-width strip of soundproofing material.

534.24 **2315**

Focusing of Sound Waves by a Parabolic Reflector.—L. D. Rozenberg. (*Zh. tekh. Fiz.*, April 1950, Vol. 20, No. 4, pp. 385–396.) A mathematical investigation is presented of the sound field near the focus. Formulae are derived for determining the pressure at the focus, and the effects of increasing the aperture of the reflector on its focusing properties, while keeping the focal length constant, are examined. Methods are indicated for choosing the optimum focal length for a given aperture of the reflector and for determining the radius of the diffraction circle at the focus. Cases when the source of sound is at a finite distance from the reflector and not on the axis are also considered. Some experimental results are included.

534.24 **2316**

On the Nonspecular Reflection of Sound from Planes with Absorbent Bosses.—V. Twersky. (*J. acoust. Soc. Amer.*, May 1951, Vol. 23, No. 3, pp. 336–338.) The analysis developed in 9 of January for nonabsorbent surfaces is extended to the case of absorbent surfaces. The results for the two cases are compared; the effect of the finite impedance may be either to decrease or increase the radiation reflected at the specular angle.

534.24 **2317**

Sound Scattering of a Plane Wave from a Nonabsorbing Sphere.—R. W. Hart. (*J. acoust. Soc. Amer.*, May 1951, Vol. 23, No. 3, pp. 323–329.) An analytical treatment is developed. Consideration is restricted to the case where the acoustic properties of the sphere are not very different from those of the surrounding medium. See also 2139 of September (Hart & Montroll).

534.24 **2318**

On the Reflection of a Spherical Sound Wave from an Infinite Plane.—U. Ingard. (*J. acoust. Soc. Amer.*, May 1951, Vol. 23, No. 3, pp. 329–335.) Boundary conditions are given in terms of a normal impedance independent of the angle of incidence. The integral for the reflected wave is expressed in a form such that the wave can be considered as originating from an image source having a certain amplitude and phase; graphs for determining these values are given in terms of a 'numerical distance' which depends on the normal impedance and the position of the field point.

534.321.9 : 537.228.1 **2319**

Design of Variable Resonant Frequency Crystal Transducers.—W. L. Hall & W. J. Fry. (*Rev. sci. Instrum.*, March 1951, Vol. 22, No. 3, pp. 155–161.) A description of a system employing liquid mercury as a backing of continuously variable dimensions. The important aspects, viz., tight coupling of the crystal and mercury backing, and decoupling of the crystal and mercury from the supporting structure, are considered in detail. Construction procedure on a unit to cover the frequency range 40–80 kc/s is indicated. Experimental results on the magnitude of the electrical input impedance as a function of frequency and mercury-column length are given. The unit is compared with transducers having fixed resonance frequency.

534.414 **2320**

The Wavelength of a Spherical Resonator with a Circular Aperture.—H. Levine. (*J. acoust. Soc. Amer.*, May 1951, Vol. 23, No. 3, pp. 307–311.) An expression for the wavelength is derived in the form of an expansion exact as far as terms in $(a/R)^2$, where a is the aperture radius and R the sphere radius. A procedure for determining the wavelength approximately for a resonator of arbitrary shape is also described.

534.414 : 534.64 **2321**

Variation of the Resistance in the Resonator Neck with Intensity of Incident Sound.—R. K. Vepa. (*Sci. Culture*, April 1951, Vol. 16, No. 10, pp. 482–483.) Measurements were made on a resonator formed by a plate 0·65 cm thick with a 1·25-cm circular orifice, appropriately spaced from a rigid backing. Results are compared with earlier measurements using a thinner plate and smaller orifice.

534.771 **2322**

Upper Limit of Frequency for Human Hearing.—J. H. Combridge & J. O. Ackroyd: R. J. Pumphrey. (*Nature, Lond.*, 17th March 1951, Vol. 167, No. 4246, pp. 438–439.) Comment on 2959 of 1950 (Pumphrey) and author's reply.

534.78 **2323**

Effect of Delay Distortion upon the Intelligibility and Quality of Speech.—J. L. Flanagan. (*J. acoust. Soc. Amer.*, May 1951, Vol. 23, No. 3, pp. 303–307.) Speech articulation tests were made on an all-pass system capable of advancing or delaying one frequency band relative to the rest of the spectrum. Measurements were made at signal/noise ratios of 30 db and 0 db. The results indicate that maximum impairment of intelligibility occurs when the delays or advances are of the order of ¼ sec and when the band delayed or advanced is near the centre of the speech spectrum.

534.79 **2324**

Calculation and Measurement of the Loudness of Sounds.—L. L. Beranek, J. L. Marshall, A. L. Cudworth & A. P. G. Peterson. (*J. acoust. Soc. Amer.*, May 1951, Vol. 23, No. 3, pp. 261–269.) An equivalent-tone method is described, in which the spectrum of the sound is divided into frequency bands which are treated as pure tones in calculating their loudness. Calculations for bands of white noise and for complex tones are compared with subjectively obtained data. The agreement is good.

534.833.4 **2325**

Absorption of Sound by Resonant Panels.—G. G. Sacerdote & A. Gigli. (*J. acoust. Soc. Amer.*, May 1951, Vol. 23, No. 3, pp. 349–352.) The experimental determination of resonance frequency and absorption of resonators formed by plywood plates with uniformly spaced circular holes, placed at various distances from the wall is described. Measurements were made at normal incidence and with diffuse sound in a reverberant room. Moderate agreement between theoretical and experimental results is noted.

621.395.61/.62 **2326**

The Piezoelectric Sound Detector and its Electrical and Acoustic Equivalent Circuits.—F. A. Fischer. (*Arch. elekt. Übertragung*, Oct. 1950, Vol. 4, No. 10, pp. 435–436.) An equivalent electrical circuit is derived which is applicable for operation of the piezoelectric transducer as generator or as detector of sound. The paper is complementary to that noted in 2966 of 1950.

621.395.61/.62 **2327**

Post-War Developments in Electroacoustics [by Telefunken].—F. Bergtold. (*Telefunken Ztg*, Sept. 1950, Vol. 23, Nos. 87/88, pp. 106–110.) Apparatus described includes moving-coil microphone, pickup, sound-distribution system, loudspeaker and arrays, cinema installation, and house-communication system.

621.395.623.7 **2328**

Loudspeaker Damping.—A. Preisman. (*Audio Engng*, March & April 1951, Vol. 35, Nos. 3 & 4, pp. 22–23..38 & 24..45.) Loudspeaker characteristics are discussed theoretically, and an experimental method is described for determining the constants of the unit. The Q is determined from the shape of the impedance/frequency curve, and the source resistance for critical damping is calculated from the voice-coil and motional impedance at resonance. An alternative method based on consideration from a mechanical viewpoint is also presented.

621.395.625.2 **2329**

Gramophone Turntable Speeds.—G. F. Dutton. (*Wireless World*, June 1951, Vol. 57, No. 6, pp. 227–231.) Suitable speeds for microgroove recordings are considered in relation to public demand, record materials, groove spacing, needle size and distortion with different tangential velocities. Results are presented graphically, and a summary gives optimum speeds for different record diameters.

621.395.625.3 : 538.221 **2330**

Mixed Ferrites for Recording Heads.—Herr. (See 2445.)

621.396.645.37.029.4 **2331**

Independent Control of Selectivity and Bandwidth.—Villard. (See 2395.)

AERIALS AND TRANSMISSION LINES

621.39.09 **2332**

A New Solution of the Fundamental Problem of the Propagation of Electromagnetic Processes in a Multi-

Wire System.—N. A. Brazma. (*C. R. Acad. Sci. U.R.S.S.*, 1st Jan. 1951, Vol. 76, No. 1, pp. 41–44. In Russian.)

621.392.09 **2333**

Surface-Wave Transmission Line.—R. H. Nelson. (*Wireless Engr*, May 1951, Vol. 28, No. 332, p. 162.) Comment on 1300 of June (Barlow) and 563 of February (Rust).

621.392.22 **2334**

The Behaviour of Electromagnetic Waves in Highly Nonuniform Lines.—H. Meinke. (*Z. angew. Phys.*, Dec. 1950, Vol. 2, No. 12, pp. 473–478.) The significant characteristic of the wave field in the region of a nonuniformity is the appearance of a wedge-shaped intrusion, resulting from longitudinal field components, in the field pattern in the neighbourhood of the point of zero transverse electric field strength. This intrusion is calculated for the case of a field with constant curvature, and examples are given of the effects due to irregularities of arbitrary form.

621.392.26† : 538.561 **2335**

Theory of the Excitation of Oscillations in a Waveguide by means of a Linear Aerial.—A. I. Akhiezer & G. Ya. Lyubarski. (*Zh. tekh. Fiz.*, Sept. 1950, Vol. 20, No. 9, pp. 1049–1064.) One of the main problems in the theory of aerials is the determination of the current distribution in an aerial to which given electromotive forces are applied. It has been shown by Leontovich & Levin (2618 of 1945) that in the case of an aerial in unlimited space the problem is reduced to the solution of a linear integrodifferential equation. A study is here presented of the current distribution in a linear aerial mounted along the axis of a cylindrical waveguide. In this case it is necessary to solve an equation of the same type as for an aerial in unlimited space. No effective methods of solving this equation for an aerial of arbitrary dimensions are known. The discussion is therefore limited to the case of a sufficiently long and thin aerial and, using a method proposed by Leontovich & Levin (2618 of 1945), an approximate solution of the equation is found by expanding the current in a series of powers of the inverse logarithm of the ratio of the length to the radius of the aerial.

The following two cases are considered separately: (*a*) when the wavelength differs considerably from the critical wavelength of the waveguide; (*b*) when this difference is not great. The current distribution in a tuned aerial differs very much from that in an aerial in unlimited space. Simple formulae for determining the amplitudes of the waves excited in the waveguide are also derived.

621.392.26† : 621.39.09 **2336**

A Study of Asymmetrical Electromagnetic Waves from the Open End of a Circular Waveguide.—L. A. Vainshtein. (*C. R. Acad. Sci. U.R.S.S.*, 21st Sept. 1950, Vol. 74, No. 3, pp. 485–488. In Russian.) The methods used in 1283 of 1949 are not applicable to the case of asymmetrical electromagnetic waves, since in this case, owing to diffraction, two longitudinal components (1) and (2) of the electric vector appear at the open end, and the problem, therefore, cannot be reduced to a single integral equation. By using a generalized method similar to that presented in an earlier paper on sound radiation (*C. R. Acad. Sci. U.R.S.S.*, 1947, Vol. 58, p. 1957) an exact solution can nevertheless be found. A system of equations (7)–(10) is derived and methods are indicated for solving it.

621.392.26† : 621.39.09 **2337**

The Diffraction of Waves at the Open End of a Circular Waveguide with Diameter Greater than the Wavelength.—L. A. Vainshtein. (*C. R. Acad. Sci. U.R.S.S.*, 11th Oct. 1950, Vol. 74, No. 5, pp. 909–912. In Russian.) The physical meaning of the formulae (see 1283 of 1949 and 2336 above) determining the radiation field under the above conditions is discussed. Starting with the case of symmetrical waves, formula (7) determining the radiation field in the back half-space ($0<\theta<\pi/2$) is considered (θ being the angle between the Z axis and the radius vector of the field point); with increase of distance from the edge of the waveguide the primary cylindrical waves gradually become spherical. From a corresponding formula (12) for the front half-space ($\pi/2<\theta<\pi$) it follows that waves from different sections of the edge interfere with one another and produce a complex spherical wave. Similar results are also obtained in the case of asymmetrical waves, but the process of development is more complicated.

621.392.43 **2338**

The Exponential Line at Cut-off Wavelength and in the Stop Range.—A. Ruhrmann. (*Arch. elekt. Übertragung*, Oct. 1950, Vol. 4, No. 10, pp. 401–412.) The theory of current and voltage distribution at cut-off wavelength is discussed. The definition of characteristic impedance used in the pass range may be retained, but the transmission equations assume an indefinite form and require transformation. The cases of termination by characteristic impedance, short-circuiting and open-circuiting are considered separately; understanding of the mode of operation with complex termination is facilitated by reference to a circle diagram. On the basis of quadripole theory it is possible to define a characteristic impedance within the stop range, although no wave propagation is to be inferred from the equations or circle diagrams. Attenuation factor and transmission factor are defined, the concept of a wave propagation process analogous to that in the pass range being made possible by introducing complex parameters. See also 1583 of 1950.

621.392.5 : 681.142 **2339**

Magnetic Delay-Line Storage.—An Wang. (*Proc. Inst. Radio Engrs*, April 1951, Vol. 39, No. 4, pp. 401–407.) A number of magnetic cores are connected together to form a static magnetic delay line in which a series of binary digits can be stored and read out. The operation of this type of line is briefly described and carefully analysed, and the optimum operating conditions are derived. The effects of eddy-current loss and leakage inductance are considered, and criteria for stability of the system are discussed.

621.396.67 **2340**

Wrotham Aerial System: Part 1 — New Design of Slot-Radiator for V.H.F. Broadcasting.—C. Gillam. (*Wireless World*, June 1951, Vol. 57, No. 6, pp. 210–214.) An omnidirectional horizontal-polarization radiator with a gain of 9 db is obtained with 32 folded slots arranged in 8 tiers of 4 spaced uniformly round a vertical cylinder. The slots are cophased and fed by a branched transmission line with impedance-matching transformers, and are designed to handle simultaneously either three 25-kW f.m. transmissions or one 25-kW f.m. and one 18-kW a.m. transmission in the frequency band 87·5–95 Mc/s.

621.396.67 **2341**

The Aerial Installations for the [German] **Post-1945 High Power Transmitters.**—W. Berndt. (*Telefunken Ztg*, Sept. 1950, Vol. 23, Nos. 87/88, pp. 39–52.) Descriptions are given of the aerial installations for the medium- and long-wave broadcast and telegraphy transmitters described in 2565 below; a certain amount of improvisation was necessary. A feature common to all the installations is the spatial separation of transmitter and aerial system, the two being connected by h.f. cable.

621.396.67 **2342**

A Helix Theorem.—J. D. Kraus. (*Proc. Inst. Radio Engrs*, May 1951, Vol. 39, No. 5, p. 563.) For helical aerials of at least a few turns, with pitch angles of 10°–15°, it is postulated that "when the circumference of an axial or end-fire helix is about one wavelength (T_1 transmission mode dominant), there is a band of frequencies over which the phase velocity of wave propagation on the helix tends toward a value that makes the directivity a maximum."

621.396.67 **2343**

Radiation Properties of Spherical Antennas as a Function of the Location of the Driving Force.—P. R. Karr. (*Bur. Stand. J. Res.*, May 1951, Vol. 46, No. 5, pp. 422–436.) A theoretical analysis is made of the radiation from a conducting sphere fed at a narrow nonequatorial zone. Variations of radiation pattern, current distribution and input admittance with the latitude of the feed zone are studied. As long as the radius a of the sphere does not exceed $\lambda/2\pi$ the radiation conductance varies approximately as $\sin^4\theta_0$, where θ_0 is the colatitude of the feed zone. For $a > \lambda/2\pi$ the maximum value of radiation conductance may occur at values of θ_0 other than 90°.

621.396.67 **2344**

Slot Radiators.—N. A. Begovich. (*Proc. Inst. Radio Engrs*, May 1951, Vol. 39, No. 5, p. 508.) Correction to paper abstracted in 2711 of 1950.

621.396.67 **2345**

Biconical Electromagnetic Horns.—W. L. Barrow, L. J. Chu & J. J. Jansen. (*Proc. Inst. Radio Engrs*, April 1951, Vol. 39, No. 4, pp. 434–435.) Corrections to paper noted in 1404 of 1940.

621.396.67 : 538.566 **2346**

Techniques for Handling Elliptically Polarized Waves with Special Reference to Antennas: Part 1 — Transmission between Elliptically Polarized Antennas.—V. H. Rumsey. (*Proc. Inst. Radio Engrs*, May 1951, Vol. 39, No. 5, pp. 535–540.) A method of analysis is discussed which makes use of the impedance concept of transmission-line theory. It is shown that P, the ratio of two orthogonal tangential components of the electric field, is related to q, the ratio between the left- and right-handed circularly polarized components corresponding to the orthogonal components, in the same way as impedance is related to reflection coefficient. Representation of polarization on a transmission-line impedance chart is described, and solutions of various polarization problems in terms of impedance analogies are discussed.

621.396.67 : 538.566 **2347**

Techniques for Handling Elliptically Polarized Waves with Special Reference to Antennas: Part 2 — Geometrical Representation of the Polarization of a Plane Electromagnetic Wave.—Deschamps. (See 2418.)

621.396.67 : 538.566 **2348**

Techniques for Handling Elliptically Polarized Waves with Special Reference to Antennas: Part 3 — Elliptically Polarized Waves and Antennas.—Kales. (See 2419.)

621.396.67 : 538.566 **2349**

Techniques for Handling Elliptically Polarized Waves with Special Reference to Antennas: Part 4 — Measurements on Elliptically Polarized Antennas.—J. I. Bohnert. (*Proc. Inst. Radio Engrs*, May 1951, Vol. 39, No. 5, pp. 549–552.) Two methods for measuring polarization characteristics are outlined; one uses a rotating linear-polarization aerial, the other uses two circular-polarization aerials.

621.396.671.012 **2350**

Pattern Calculator for A.M.—G. R. Mather. (*Electronics*, April 1951, Vol. 24, No. 4, pp. 100–101.) A graphical method for calculation of the radiation pattern of two-tower directive arrays.

621.396.677 **2351**

The Electric and Magnetic Constants of Metallic Delay Media containing Obstacles of Arbitrary Shape and Thickness.—S. B. Cohn. (*J. appl. Phys.*, May 1951, Vol. 22, No. 5, pp. 628–634.) Methods of deriving the dielectric-constant and permeability of a metallic-obstacle medium are given; the equivalent shunt capacitive susceptance and series inductive reactance of the individual obstacles are also determined. By means of a correspondence established between an infinitely thin conducting obstacle and an aperture in an infinitely thin conducting wall, formulae are derived for different shapes of obstacle. The effect on the magnetic field of obstacles of moderate thickness is evaluated. For obstacles of arbitrary shape and thickness the constants can be determined by the electrolyte-tank method; this is demonstrated for a particular example.

621.396.677 **2352**

Directional Aerials for Radio Stations.—K. O. Schmidt. (*Fernmeldetech. Z.*, Feb. 1951, Vol. 4, No. 2, pp. 49–56.) Review and discussion of different types of radiators for use at wavelengths between 3 cm and 3 m.

621.396.677 : 621.396.11 **2353**

A Wide-Band Aerial System for Circularly Polarized Waves, Suitable for Ionospheric Research.—G. J. Phillips. (*Proc. Instn elect. Engrs*, Part III, May 1951, Vol. 98, No. 53, pp. 237–239.) The system described can be used to select, without readjustment, one or other of two waves circularly polarized in opposite senses and incident vertically, within the frequency range 2–6 Mc/s. Two mutually perpendicular horizontal dipoles are associated respectively with two phase-shifting LC lattice networks. E.m.f.s in the aerials, initially 90° out of phase, may be added in phase, thus giving selection of a circularly polarized component. A discrimination ratio of at least 12 : 1 between the components has been obtained.

621.396.677.001.4 **2354**

Reflecting Surface to Simulate an Infinite Conducting Plane.—S. J. Raff. (*J. appl. Phys.*, May 1951, Vol. 22, No. 5, pp. 610–613.) A finite reflecting surface which simulates an infinite plane is required for calibrating measurements of reflections back to a microwave transmitting aerial. The Fresnel-zone method of physical optics is used for the design calculations. Variational calculus is used to determine the optimum reflector shape for a given aerial pattern, reflector size, and range of aerial-to-reflector distance. Theoretical values are compared with results obtained on an example constructed for use at 25λ from a dipole aerial.

621.392 **2355**

Transmission Lines and Networks. [Book Review]—W. C. Johnson. Publishers: Mc-Graw Hill, New York, 1950, 361 pp., $5.00. (*Electronics*, April 1951, Vol. 24, No. 4, pp. 278, 280.) "Although basically a textbook for undergraduate students, the material covered should be of considerable interest to practicing engineers in both the power and communication fields."

CIRCUITS AND CIRCUIT ELEMENTS

621.3.011.6 : 621.317.726 **2356**

Calculation of CR Elements for the case of Varying Voltage and/or Nonlinear Resistances.—H. Elger. (*Arch.*

elekt. Übertragung, Oct. 1950, Vol. 4, No. 10, pp. 413–426.) Methods are presented for calculating the build-up conditions in CR elements when either steady or time-varying voltage is applied; consideration is given to potential-divider circuits. The effect of nonlinear resistances is investigated, as, e.g., in the demodulation of rectified modulated voltage with a square-law rectifier. Methods based on the theory are discussed for varying time constants within wide limits, and an example shows how the readings of a diode peak-voltage meter, operating at very high voltages, may be corrected for errors due to small discharges through the insulation between pulses.

621.3.015.7 : 621.387.4† **2357**

Single-Channel Analyzer.—J. E. Francis, Jr, P. R. Bell & J. C. Gundlach. (*Rev. sci. Instrum.*, March 1951, Vol. 22, No. 3, pp. 133–137.) An analyser for proportional and scintillation counters counts the number of pulses whose heights lie between E and $E + \Delta E$; ΔE is constant to within $\pm 1{\cdot}2\%$ for $E = 0 - 90$ V.

621.3.016.352 **2358**

Relation of Nyquist Diagram to Pole-Zero Plots.—H. F. Spirer. (*Proc. Inst. Radio Engrs*, May 1951, Vol. 39, No. 5, p. 562.) Comment on 1086 of May (Harman).

621.314.212 : 534.232 : 538.652 **2359**

Electroacoustic Transformation by means of Magnetostriction, with Special Reference to Radiation from Transformers.—Rust. (See 2313.)

621.314.3† **2360**

High-Gain Magnetic Amplifier.—R. Feinberg. (*Wireless Engr*, May 1951, Vol. 28, No. 332, pp. 151–155.) Self-excitation is an effective method of obtaining feedback in a transductor; high values of current amplification can be obtained by making the number of turns of the self-excitation winding sufficiently large in relation to the number of turns of the load winding. Turns relations for stable operation are discussed; when operating unstably the system may be used as an on-off trigger relay.

621.314.3† : 621.318.42 **2361**

Design of [magnetic-] **Amplifier Inductors with Series-Connected Ohmic Loads.**—E. Helmes. (*Arch. elekt. Übertragung*, Jan. 1951, Vol. 5, No. 1, pp. 39–46.) A formula is derived expressing the effective permeability and self-inductance and the transfer impedance between the coil and the load in terms of the core cross-section and the number of turns. Families of curves are plotted from measurements on (*a*) a two-element inductor with transformer-sheet core, (*b*) a three-element inductor with mumetal core. These curves can be applied to other core materials by changing the scale.

621.314.3† : [621.396.615.17 + 621.396.619.2 **2362**

The Use of Saturable Reactors as Discharge Devices for Pulse Generators.—W. S. Melville. (*Proc. Instn elect. Engrs*, Part III, May 1951, Vol. 98, No. 53, pp. 185–204. Discussion, pp. 204–207.) The development of materials used for saturable reactors is outlined; a magnetic material with rectangular hysteresis loop is required. Magnetic discharge devices can often replace electronic discharge devices; the merits of the two types are compared. The design and operation of saturable-reactor circuits for radar pulse-modulation and ignitron ignition are described.

621.315.592† + 621.314.632 **2363**

The Characteristics and some Applications of Varistors.—F. R. Stansel. (*Proc. Inst. Radio Engrs*, April 1951, Vol. 39, No. 4, pp. 342–358.) A tutorial paper reviewing the properties of semiconductor nonlinear circuit elements, with particular reference to those available commercially. The principles and limitations governing the use of these elements are summarized. The varying importance of the different parameters with different types of application is illustrated by short discussions of the design of voltage limiters, power rectifiers, h.f. modulators and compandors.

621.316.726 : 681.142 **2364**

Automatic Frequency Control.—J. M. M. Pinkerton. (*Electronic Engng*, April 1951, Vol. 23, No. 278, pp. 142–143.) Description of a method of controlling the clock pulse frequency in the storage system of a digital computer. The phase of a pulse which has travelled down an ultrasonic delay line is compared with that of a later pulse of the same series which has not been delayed, and the phase difference is used to derive a voltage for control of the frequency of the master oscillator producing the clock pulses, by means of a reactance valve.

621.316.86 **2365**

Pyrolytic Film Resistors: Carbon and Borocarbon.—R. O. Grisdale, A. C. Pfister & W. van Roosbroeck. (*Bell Syst. tech. J.*, April 1951, Vol. 30, No. 2, pp. 271–314.) A description of the production and structure of thin carbon films deposited on ceramics or fused silica by the pyrolysis of hydrocarbon vapours, and capable of providing resistors of high stability with resistances from a few ohms to tens of megohms. The incorporation of boron in the film results in a smaller temperature coefficient than that of many wire-wound resistors, and the negligible skin effect permits advantageous use of these film resistors at high frequencies. Resistance values up to $10^9\Omega$ have been obtained in the borocarbon type. See also 583 of March.

621.317.353.2.012.3 **2366**

Mixer Harmonic Chart.—T. T. Brown. (*Electronics*, April 1951, Vol. 24, No. 4, pp. 132, 134.) The chart facilitates identification of spurious frequencies resulting from beating of various harmonics of two inputs, the frequency of one being variable.

621.318.572 **2367**

A Three-State Flip-Flop.—A. D. Booth & J. Ringrose. (*Electronic Engng*, April 1951, Vol. 23, No. 278, p. 133.) With Type-6J6 and Type-6SN7 valves, particular values of cathode resistor were found to give three stable states in flip-flop units; an explanation is given of the circuit operation.

621.319.53 : 621.396.9 **2368**

High-Voltage Pulse Modulators for Radar Pulse Transmitters.—Tigler. (See 2431.)

621.385.3 : 546.289 **2369**

Duality as a Guide in Transistor Circuit Design.—R. L. Wallace, Jr, & G. Raisbeck. (*Bell Syst. tech. J.*, April 1951, Vol. 30, No. 2, pp. 381–417.) The properties of a transistor are compared with those of a vacuum-tube triode, and the relation between them is found to be such that by interchanging current and voltage a known vacuum-tube circuit can be transformed into one suitable for use with transistors. The necessary changes in the circuit elements are considered, and practical examples of these networks (or duals) are given. Circuits which permit the simultaneous use of vacuum tubes and transistors, such as the Doherty amplifier, are also discussed.

621.392 **2370**

The Potential Analogue Method of Network Synthesis.—S. Darlington. (*Bell Syst. tech. J.*, April 1951, Vol. 30, No. 2, pp. 315–365.) The method developed is based on analogy between the gain and phase of linear networks and the two-dimensional potential and stream functions produced by charges corresponding to the network singularities.

621.392 **2371**

The Synthesis of RC Networks to have Prescribed Transfer Functions.—H. J. Orchard. (*Proc. Inst. Radio Engrs*, April 1951, Vol. 39, No. 4, pp. 428–432.) A more general method than that of Guillemin (2462 of 1949) is described. The resulting network is in the form of a lattice, and is capable of providing any transfer function physically realizable by an RC network. The design procedure is simple. A numerical example is included.

621.392.4 **2372**

An Application of Equaliser Curves to the Design of Two-Terminal Networks.—P. W. Seymour. (*P.O. elect. Engrs' J.*, April 1951, Vol. 44, Part 1, pp. 31–32.) Description of a method for deriving the circuit constants of a 2-terminal network from available design data for a 4-terminal constant-impedance equalizer having an insertion-loss/frequency characteristic similar to the impedance/frequency characteristic required for the 2-terminal network.

621.392.5 **2373**

Fluctuations in a Linear System with Periodically Varying Parameters.—S. I. Borovitski. (*C. R. Acad. Sci. U.R.S.S.*, 11th Sept. 1950, Vol. 74, No. 2, pp. 233–236. In Russian.) A system is considered with parameters varying in accordance with the equation (1). Statistically, the behaviour of such a system can be represented by the Einstein-Fokker equation (2). With the aid of the main solution (3) of this equation the case of a statistically stable system is investigated; the spectrum of a perturbation consists of discrete lines on a continuous background. The discussion is illustrated by experimental curves obtained in analysing the output of a superregenerative receiver in the absence of a signal.

621.392.5 **2374**

Response Characteristics of Resistance-Reactance Ladder Networks.—R. R. Kenyon. (*Proc. Inst. Radio Engrs*, May 1951, Vol. 39, No. 5, pp. 557–559.) Generalized expressions for the transfer functions for resistance-reactance ladder networks are given and discussed in detail. The output voltage as a function of time is derived for the cases of unit-impulse and unit-step input voltage. Two methods are discussed for determining the output voltage for the case of an arbitrary input.

621.392.5 **2375**

Passive Pulse-Sharpening Circuits.—L. Reiffel. (*Rev. sci. Instrum.*, March 1951, Vol. 22, No. 3, pp. 214–216.) A circuit is described using a Ge damping diode in conjunction with a resonant circuit to shorten pulses obtained from a G-M counter or other pulse detector.

621.392.52 **2376**

Relations between Signals and Spectra.—K. Fränz. (*Arch. elekt. Übertragung*, Jan. 1951, Vol. 5, No. 1, pp. 10–14.) Theoretical proof of certain laws of filter theory. The major part of the energy of any signal of given duration is confined to a narrow spectrum whose limits are independent of the waveform. The connection between filter bandwidth and signal build-up is derived. The pass-band response curve of a filter with monotonic build-up must be roughly bell-shaped.

621.392.52 : 621.396.611.21 **2377**

Lattice-Type Crystal Filter.—R. Lowrie. (*Electronics*, April 1951, Vol. 24, No. 4, pp. 129–131.) Description of a 2-section filter incorporating eight crystals, with a pass band of width 3·9 kc/s centred at 2 Mc/s and a bandwidth of 12 kc/s at 60 db attenuation.

621.394/.396].6 **2378**

A New Colour-coded Wiring System.—N. G. Partridge. (*Electronic Engng*, April 1951, Vol. 23, No. 278, pp. 138–139.) The method is based on the consecutive numbering of all the items involved, whether wires, cableforms, chassis or racks, using the international colour-code system to enable the allotted number to be carried by each item in the form of coloured bands or labels.

621.394 .396].6 : 621.392.012 **2379**

Correlation of Circuit Diagram and Wiring Development of Electronic Systems.—A. W. Keen. (*Electronic Engng*, April 1951, Vol. 23, No. 278, pp. 144–145.)

621.396.6 : 061.4 **2380**

Trends in Components.—(*Wireless World*, May 1951, Vol. 57, No. 5, pp. 185–188.) A survey of the components and accessories shown at the annual private exhibition organized by the Radio and Electronic Component Manufacturers' Federation in London, April 1951.

621.396.611 + 621.317.7].029.63 **2381**

Circuits and Measurement Apparatus for the 30-cm Band.—Safa. (See 2479.)

621.396.611.21 **2382**

Some Notes on Overtone Crystals and Maintaining Oscillators operating in the Frequency Range of 33-55 Mc/s.—J. B. Supper. (*Proc. Instn elect. Engrs*, Part III, May 1951, Vol. 98, No. 53, pp. 240–247.) The terms 'minimum impedance' and 'inductive impedance' are proposed for identifying two general forms of maintaining circuit investigated, and the concept of 'impedance diameter' as a measure of crystal goodness is introduced. Measurements on different types of crystal and their behaviour in the inductive-impedance oscillator are recorded and discussed. The effect of plating area on crystals is considered, and attention is drawn to the improvement obtained by reducing the plating diameter below the present standard value.

621.396.611.21 **2383**

Amplitude of Vibration in Piezoelectric Crystals.—E. A. Gerber. (*Electronics*, April 1951, Vol. 24, No. 4, pp. 142..216.) The amplitude of vibration of a crystal has some influence on the crystal parameters. Simple expressions are derived relating amplitude to the r.f. current through the crystal and to the voltage across it, and comparison is made with the results of more general theories of a vibrating piezoelectric plate. Only thickness modes of vibration are considered. Experiments verifying the formulae are described.

621.396.611.3 **2384**

The Calculation of the Input Impedance of Coupled Oscillatory Circuits.—F. A. Fischer & U. John. (*Arch. elekt. Übertragung*, Jan. 1951, Vol. 5, No. 1, pp. 33–38.) Description of a method based on the representation of input impedance as a function of the difference between the damping and the resonance frequencies of the two circuits and the coupling factor. Curves and diagrams are plotted for a typical case. The basic formula is extended to the case of *n* coupled circuits.

621.396.611.4 **2385**

Application of the Method of Curvilinear Coordinates to the Calculation of a Π-type Cavity Resonator.—V. L. Patrushev. (*Zh. tekh. Fiz.*, June 1950, Vol. 20, No. 6, pp. 727–734.) Krasnushkin has studied the propagation of waves in waveguides of rectangular cross-section by using the method of normal waves. In a previous paper (*Bull. Acad. Sci. U.R.S.S., Sér. phys.*, 1948, Vol. 12, p. 684) the author applied this method to the investigation of the e.m. fields and natural frequencies of cavity resonators having rotational symmetry. Under certain conditions a Π-type cavity resonator can be regarded approximately as a coaxial line with capacitance loading and therefore belonging to this group of resonators. In the present paper it is shown that by introducing curvilinear coordinates a rigorous solution of the problem is possible in principle.

621.396.611.4 **2386**

Design of the Π-type Cavity Resonator.—V. L. Patrushev & O. V. Romanova. (*Zh. tekh. Fiz.*, July 1950, Vol. 20, No. 7, pp. 798–801.) A formula is derived for determining the length of the plunger used for tuning the resonator. The discussion is illustrated by two numerical examples which have been verified experimentally.

621.396.615.17 **2387**

Theory of the Symmetrical Multivibrator.—N. A. Zheleztsov. (*Zh. tekh. Fiz.*, July 1950, Vol. 20, No. 7, pp. 788–797.) The valve characteristic is assumed to consist of a number of linear sections. Analysis of the movement of the operating point along these sections makes it possible to determine the build-up of the discontinuous oscillations and to prove the singularity and stability of the discontinuous periodic solution.

621.396.645 **2388**

Application of the Properties of Newtonian Potentials to the Design of Frequency-Modulation Amplifiers.—P. Belgodère & A. Fromageot. (*Onde élect.*, Jan. 1951, Vol. 31, No. 286, pp. 18–32.) Use of a constant-gain amplifier to provide constant group-transmission time, (856 of 1950), requires that the width of the pass band be unnecessarily large. From a theoretical treatment an alternative method of design is derived. This has yet to be checked experimentally, in particular as regards tolerances on tuning frequencies and circuit parameters.

621.396.645 : 535.247.4 **2389**

A Balance Indicator with High Input Impedance using a Cathode Follower.—Dighton. (See 2513.)

621.396.645 : 621.317.083.4 **2390**

Sensitive Null Detector.—M. G. Scroggie. (*Wireless World*, May 1951, Vol. 57, No. 5, pp. 175–178.) Description of a selective a.f. bridge amplifier for use at frequencies between 50 and 1 500 c/s, with a 'magic eye', milliammeter, or telephones as the output indicator. An a.g.c. circuit permits a range of signal input of 10 μV to 10 V, and the time constant is such that a transient indication is given of a change in input at any signal level.

621.396.645 : 621.317.6 **2391**

The Determination of Amplifier Sensitivity with the Aid of the Noise Diode.—Squires. (See 2478.)

621.396.645.012.8 **2392**

Network Representation of Input and Output Admittances of Amplifiers.—F. W. Smith. (*Proc. Inst. Radio Engrs*, April 1951, Vol. 39, No. 4, p. 439.) Comment on 2194 of 1949 (Vallese).

621.396.645.211 **2393**

Maximum Output from a Resistance-Coupled Triode Voltage Amplifier.—J. M. Diamond. (*Proc. Inst. Radio Engrs*, April 1951, Vol. 39, No. 4, pp. 433–434.) Simple formulae are derived for optimum load resistance with respect to output voltage, and for maximum voltage swing obtainable.

621.396.645.36.029.4 **2394**

Bridge-Compensated Differential Amplifiers.—J. Labus. (*Arch. elekt. Übertragung*, Oct. 1950, Vol. 4, No. 10, pp. 437–440.) Sensitive push-pull amplifiers used for special purposes (e.g., electrocardiography) are liable to interference from a.c. fields at the input terminals. Several previously proposed circuits for eliminating this interference are briefly reviewed, and a method using a resistance-bridge network connected across the input is described, which suppresses the interfering voltages before they reach the grids of the first-stage valves and hence prevents the production of harmonics.

621.396.645.37.029.4 **2395**

Independent Control of Selectivity and Bandwidth.—O. G. Villard, Jr. (*Electronics*, April 1951, Vol. 24, No. 4, pp. 121–123.) The feedback circuit in a RC a.f. amplifier is designed to have constant amplitude/frequency but variable phase/frequency characteristics. The feedback is positive at the resonance frequency, negative at frequencies far from it. The complete circuit is shown for an amplifier with a constant percentage bandwidth/frequency variation and a choice of three bandwidths at any desired selectivity.

621.396.822 **2396**

Thermal Fluctuation of Charge in Linear Circuits.—E. A. N. Whitehead. (*Elliott J.*, March 1951, Vol. 1, No. 1, pp. 32–34.) Derives the usual expressions for the noise power developed across an impedance; the noise generators are considered as being in parallel with the various circuit components.

621.392 **2397**

Transmission Lines and Networks. [Book Review]—Johnson. (See 2355.)

621.392.025 **2398**

Alternating Current Circuits. [Book Review]—R. M. Kerchner & G. F. Corcoran. Publishers: J. Wiley & Sons, New York, 3rd edn 1950, 586 pp., $5.50. (*Proc. Inst. Radio Engrs*, April 1951, Vol. 39, No. 4, p.448.) ". . . One of the very best summaries of the elementary background of a.c. circuit analysis. . ."

GENERAL PHYSICS

534.01 + 538.56 **2399**

Theory of Waves and Oscillations in Nonhomogenous Discrete Structures.—P. E. Krasnushkin. (*Zh. tekh. Fiz.*, Sept. 1950, Vol. 20, No. 9, pp. 1065–1083.) A general discussion applicable to various discrete oscillating systems such as molecular chains of certain organic compounds, nonuniform waveguides and strings, the ionosphere, etc. The following two types of oscillations are met with in such systems: (*a*) 'collective' oscillations of sinusoidal form spread more or less uniformly over all elements of the system and (*b*) 'local' oscillations of exponential form in parts of the space occupied by the system. In order to investigate the nature of these two types of oscillations and the conditions of their appearance, consideration is given to the general case of a chain structure consisting of cells, each of which represents an oscillating system with one degree of freedom. The local oscillations approximate in frequency and shape to the natural oscillations of isolated cells and the collective oscillations appear as a result of the interaction between the local oscillations, depending on the degree of resonance coupling, a conception introduced by Mandelstam. The loosening of the coupling may result in the appearance of collective oscillations only within parts of the system and their penetration into other parts in the form of exponential 'tails'. Since the collective oscillations are essentially standing waves, the regions limiting them are called wave barriers. Three different types of these barriers are specified and their effect on the operation of the system is discussed.

535.12
2400

Wave Propagation in an Anisotropic Medium and the Corresponding Principal Directions.—M. Pastori. (*Nuovo Cim.*, 7th May 1949, Vol. 6, No. 3, pp. 187–193.) At any point within the medium at least three principal directions exist such that the sum of the squares of the velocities of the three corresponding wavefronts is constant.

535.215 : 538.221
2401

The Surface Photoelectric Effect in Ferromagnetics.—S. V. Vonsovski & A. V. Sokolov. (*C. R. Acad. Sci. U.R.S.S.*, 11th Jan. 1951, Vol. 76, No. 2, pp. 197–200. In Russian.) The anomalies in the photoelectric effect in ferromagnetics observed by Cardwell (*Phys. Rev.*, 1949, Vol. 76, p. 125) are discussed from the standpoint of the interaction between the outer s- and inner d-electrons, a concept developed by Vonsovski (2074 of 1947). The photoelectric current and the effective work function in ferromagnetics depend on the value of their spontaneous magnetization.

537.529
2402

A Review of Spark Discharge Phenomena.—F. M. Bruce. (*J. Brit. Instn Radio Engrs*, April 1951, Vol. 11, No. 4, pp. 121–135.) The Townsend and streamer theories are discussed; the latter requires further evidence for its full substantiation, while the range of application of the former is likely to be greatly increased. Results of investigations in progress will be of importance not only in the field of measurement over a wide variety of waveforms but also in the extended use of gaseous insulation. Consideration is given to the uniform-field gap for standardizing methods of measurement. H.f. breakdown is not discussed here; this was dealt with in a paper noted in 613 of March.

537.533.8
2403

Some Peculiarities of the Secondary-Electron Emission from Thin Films of Calcium Chloride.—V. N. Favorin. (*Zh. tekh. Fiz.*, Aug. 1950, Vol. 20, No. 8, pp. 916–922.)

537.562 : 537.311
2404

Convergence of the Chapman-Enskog Method for a Completely Ionized Gas.—R. Landshoff. (*Phys. Rev.*, 1st May 1951, Vol. 82, No. 3, p. 442.) A note relevant to 335 of February (Cohen, Spitzer & Routly).

537.71
2405

Generalized Electrical Formulas.—V. P. Hessler & D. D. Robb. (*Elect. Engng, N.Y.*, April 1951, Vol. 70, No. 4, pp. 332–336.) "A set of generalized electrical formulas is developed to which units of any absolute system may be applied. The generalization is accomplished with the aid of two additional constants, n and u. The general formulas may be reduced to the usual rationalized or unrationalized forms or to the Gaussian or Heaviside forms by the substitution of tabulated numerical values of the constants n and u in the general form."

538.221
2406

Single-Domain Structure in Ferromagnetics, and the Magnetic Properties of Finely Dispersed Substances.—E. Kondorski. (*C. R. Acad. Sci. U.R.S.S.*, 11th Sept. 1950, Vol. 74, No. 2, pp. 213–216. In Russian.)

538.221
2407

The Dependence of Magnetization Curves on Temperature and the Hysteresis Loop of High-Coercivity Alloys.—Ya. S. Shur & N. A. Baranova. (*C. R. Acad. Sci. U.R.S.S.*, 11th Sept. 1950, Vol. 74, No. 2, pp. 225–228. In Russian.)

538.24
2408

The Effect of Directed Stresses on the Shape of the Magnetization Curve in Strong Fields.—L. V. Kirenski & L. I. Slobodskoi. (*C. R. Acad. Sci. U.R.S.S.*, 21st Sept. 1950, Vol. 74, No. 3, pp. 457–459. In Russian.) A formula (1) expressing the intensity of magnetization is quoted, and relations between the various constants are discussed, particularly for the case when the elastic stresses in the sample are directed along the magnetizing field.

538.249
2409

On Certain Laws Governing Magnetic Viscosity.—R. V. Telesnin. (*C. R. Acad. Sci. U.R.S.S.*, 11th Dec. 1950, Vol. 75, No. 5, pp. 659–660. In Russian.)

538.249
2410

The Dependence of Magnetic Viscosity on Temperature.—R. V. Telesnin & E. F. Kuritsyna. (*C. R. Acad. Sci. U.R.S.S.*, 21st Dec. 1950, Vol. 75, No. 6, pp. 797–798. In Russian.)

538.56
2411

Applications of the Radiation from Fast Electron Beams.—H. Motz. (*J. appl. Phys.*, May 1951, Vol. 22, No. 5, pp. 527–535.) The radiation from beams of fast electrons passing through a succession of transverse electric or magnetic fields of alternating polarity is examined. The frequencies and the angular distribution of radiated energy are calculated; the coherence of the radiation is discussed. Several applications appear possible, one of which is the production of millimetre waves of considerable power. Another is the monitoring of the speed of electrons having energies up to 10^9 eV.

538.56 : 535.42
2412

Diffraction of Centimetre Electromagnetic Waves by Metal Disks.—H. Severin. (*Z. angew. Phys.*, Dec. 1950, Vol. 2, No. 12, pp. 499–505.) The diffraction phenomena observed along the axis of, and close up to, a conducting disk, for normal incidence of a plane wave, are compared with three approximate theoretical solutions. A wavelength of 10 cm and disks of thickness 2 mm and radius $0{\cdot}5\ \lambda$, $1\ \lambda$, $1{\cdot}5\ \lambda$ and $2\ \lambda$ were used. Best agreement with measurements is provided by a theory which assumes the disk to be covered with a layer of magnetic dipoles.

538.56 : 535.42
2413

On the Diffraction of a Plane Electromagnetic Wave by a Paraboloid of Revolution.—C. W. Horton & F. C. Karal, Jr. (*J. appl. Phys.*, May 1951, Vol. 22, No. 5, pp. 575–581.) A theoretical investigation of the diffraction by the convex surface of the paraboloid. Expressions are derived for the components of the incident, scattered and refracted waves. The case of a perfectly conducting paraboloid and a wave front perpendicular to the axis of rotation is considered. The variation of amplitude of the scattered wave with distance along axis of rotation is compared with the corresponding curve for a sphere of radius equal to the radius of curvature of the paraboloid at its nose.

538.56 : 535.42
2414

On the Diffraction of Electromagnetic Waves by Two Conducting Parallel Half-Planes.—M. G. Cheney, Jr, & R. B. Watson. (*J. appl. Phys.*, May 1951, Vol. 22, No. 5, pp. 675–679.) The diffraction produced by the edges of two half-planes, arranged one behind the other relative to the signal source, is investigated experimentally and theoretically. Agreement between the two sets of results is not very close.

538.561
2415

The Problem of the Excitation of Electromagnetic Oscillations.—B. Ya. Moyzhes. (*Zh. tekh. Fiz.*, June

1950, Vol. 20, No. 6, pp. 698–715.) A general method has been recently proposed by G. A. Grinberg ("Selected Problems of the Mathematical Theory of Electric and Magnetic Phenomena", published by the Academy of Sciences of U.S.S.R., 1948.) for solving a large group of problems in connection with the excitation of waveguides and other systems by a given distribution of currents or by slots for which the tangential component of the electric field is known. In this method one or, more generally, two independent equations are derived from Maxwell's equation. Each of these equations includes a scalar function (field component or a component of auxiliary function potentials) for which the boundary conditions have to be established separately. In the present paper this method is discussed in detail and applied to the cases of a cylindrical waveguide and a sectoral horn.

538.566 **2416**

Synthesis and Analysis of Elliptic Polarization Loci in Terms of Space-Quadrature Sinusoidal Components.—M. G. Morgan & W. R. Evans, Jr. (*Proc. Inst. Radio Engrs*, May 1951, Vol. 39, No. 5, pp. 552–556.) A mathematical analysis of elliptically polarized waves, by means of which the elliptic locus resulting from three mutually orthogonal component field vectors may be specified in terms of those vectors or the vectors specified in terms of the locus. The simpler case of two-component vectors is considered first.

538.566 : 535.43 **2417**

Electromagnetic Scattering from Spheres with Sizes Comparable to the Wavelength.—A. L. Aden. (*J. appl. Phys.*, May 1951, Vol. 22, No. 5, pp. 601–605.) The general formula for the back-scattering cross-section of a sphere is difficult to evaluate for a complex refractive index, owing to the lack of the necessary tables of Bessel functions. The evaluation may be carried out by transforming the formula by means of logarithmic derivative functions. Back-scattering cross-sections were measured for water spheres with sizes comparable to the wavelength (16·23 cm) using a standing-wave method. The water was contained in a thin hemispherical shell of dielectric, mounted on an aluminium disk. Very good agreement with theory was obtained.

538.566 : 621.396.67 **2418**

Techniques for Handling Elliptically Polarized Waves with Special Reference to Antennas: Part 2 — Geometrical Representation of the Polarization of a Plane Electromagnetic Wave.—G. A. Deschamps. (*Proc. Inst. Radio Engrs*, May 1951, Vol. 39, No. 5, pp. 540–544.) The polarization and amplitude of an elliptically polarized plane wave may be specified by three quantities which define the ellipse traced out by the field vector and which may be represented, according to a method used by Poincaré in optics, by the co-ordinates of a point on a sphere. Methods of solving problems using this concept are discussed.

538.566 : 621.396.67 **2419**

Techniques for Handling Elliptically Polarized Waves with Special Reference to Antennas: Part 3 — Elliptically Polarized Waves and Antennas.—M. L. Kales. (*Proc. Inst. Radio Engrs*, May 1951, Vol. 39, No. 5, pp. 544–549.) A complex vector algebra is presented, by means of which an elliptically polarized wave may be completely specified. The field at a point may be resolved into two space components, in general not in the same direction, having time variations in phase quadrature. Thus if the two space vectors are U_r and U_i, the complex vector given by $U_r + j\,U_i$ completely defines the field at the point. The algebraic properties of such vectors are discussed, and resolution of the field into orthogonal elliptically polarized components and the concept of phase are studied. Relations useful in measurements and aerial problems are given.

537.311.33 **2420**

Semi-Conductors. [Book Review]—D. A. Wright. Publishers: Methuen & Co., London, 130 pp., 7s. 6d. (*Wireless Engr*, May 1951, Vol. 28, No. 332, p. 164.) "Specially concerned with the theory of electron flow in semi-conductors, and across the boundary between them and either a metal or a vacuum. . . . This monograph will undoubtedly be of great use to students of the electron physics of solids."

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523 + 551 + 621.396.93 **2421**

Recent Work of the Radiophysics Division, C.S.I.R.O.—E. G. Bowen. (*Proc. Instn Radio Engrs, Aust.*, April 1951, Vol. 12, No. 4, pp. 99–108.) Developments in radio techniques for meteorology, astronomy and navigation are described. A digital computer has been designed and built by the Division to deal with its internal computing requirements.

523.72 : 621.396.822 **2422**

Radio Helioscopy.—M. Waldmeier. (*Naturwissenschaften*, Jan. 1951, Vol. 38, No. 1, pp. 1–4.) A survey based on papers presented at the General Assembly of the International Scientific Radio Union. Results of solar-radiation measurements within the wavelength range 1 cm–10 m are discussed, and the adequacy of theories so far advanced is examined. Most recent observations tend to support theories according to which the radiation disturbances are caused by coronal plasma oscillations excited by corpuscular rays or protuberances in motion.

523.746 : 538.12 **2423**

The Propagation of the Electromagnetic Field of Sunspots in the Sun's Atmosphere.—P. E. Kolpakov & Ya. P. Terletski. (*C. R. Acad. Sci. U.R.S.S.*, 11th Jan. 1951, Vol. 76, No. 2, pp. 185–188. In Russian.) Because of its electrical conductivity, the highly ionized atmosphere of the sun might be expected to act as a screen for the electromagnetic field of the sunspots. That indications of this field are nevertheless observed in the middle and upper layers of the sun's atmosphere is due to motion of the latter. A mathematical analysis of the propagation process, leading to the derivation of two differential equations (bottom of p. 187) is presented.

537.591 : [523.854 : 621.396.822 **2424**

Cosmic Rays as a Source of Galactic R.F. Radiation.—V. L. Ginzburg. (*C. R. Acad. Sci. U.R.S.S.*, 21st Jan. 1951, Vol. 76, No. 3, pp. 377–380. In Russian.) Galactic r.f. radiation cannot be explained by the thermal radiation of interstellar electrons. It is suggested that it may be produced by the radiation from the electronic component of cosmic rays travelling at relativistic velocities in the magnetic fields near and between the stars. On this assumption, and knowing the intensity of the magnetic field, it is possible to determine the concentration of the corresponding cosmic particles. Results of a detailed analysis are given separately for the cases of general galactic emission and emission from discrete sources. The possibility of cosmic-ray radiation in the magnetic field of the earth is also discussed. The results obtained are not conclusive.

550.38 **2425**

Geomagnetic Indices for the Period from 22nd Dec.

1950 to 31st March 1951.—(*Z. Met.*, April 1951, Vol. 5, No. 4, p. 123.) Observations made at Niemegk are presented in chart form.

551.510.535 **2426**

Sporadic E Movements on 21st June 1949.—N. C. Gerson. (*Tellus*, Feb. 1951, Vol. 3, No. 1, pp. 56–59.) An analysis of the reports from about 300 American radio amateurs of radio contacts made by reflection from sporadic-E regions. Between midnight and 0400 hours two clouds of E_s ionization were reported, which drifted westwards across the U.S.A. with average velocities of about 250 km/hr.

551.510.535 **2427**

Magneto-ionic Triple Splitting of Ionospheric Waves.—O. E. H. Rydbeck. (*Onde élect.*, Feb. & March 1951, Vol. 31, Nos. 287 & 288, pp. 70–81 & 153–156.) French version of paper noted in 1147 of May.

551.510.535 : 621.396.11 **2428**

The Effect of the Lorentz Polarization Term on the Vertical Incidence Absorption in a Deviating Ionosphere Layer.—J. M. Kelso. (*Proc. Inst. Radio Engrs*, April 1951, Vol. 39, No. 4, pp. 412–419.) Using the two parabola approximations and neglecting the effect of the earth's magnetic field, the influence of the Lorentz term on the apparent scale height and total absorption of a Chapman layer is calculated for a wave reflected in the layer. The absorption is increased by 4% or more when the Lorentz term is included. See also 638 of 1950.

551.594.25 **2429**

The Origin of the Electric Charge on Rain.—J. A. Chalmers. (*Quart. J. R. met. Soc.*, April 1951, Vol. 77, No. 332, pp. 249–259.) Previously measured values of charge on rain during periods when point discharge occurs can be accounted for quantitatively on reasonable assumptions in terms of the process of ion capture. The results show that separation of charge must operate at levels down to about 800 m, and the consequences of this are discussed in relation to theories of the process of separation of charge.

621.317.79 : 621.396.822 **2430**

High-Sensitivity High-Frequency Noise-Measurement Apparatus Calibrated Absolutely in kT_0 Units.—Röschlau. (See 2487.)

LOCATION AND AIDS TO NAVIGATION

621.396.9 : 621.319.53 **2431**

High-Voltage Pulse Modulators for Radar Pulse Transmitters.—H. Tigler. (*Arch. elekt. Übertragung*, Jan. & Feb. 1951, Vol. 5, Nos. 1 & 2, pp. 47–51 & 91–98.) Descriptive review of 9 different circuits for pulse generation using vacuum valves, thyratrons and spark discharge systems. Control of the spark gap and voltage multiplication by means of Marx circuits are discussed. This circuit and the spark discharge are specially suitable for short pulses at high power and high voltage.

621.396.93 + 523 + 551 **2432**

Recent Work of the Radiophysics Division, C.S.I.R.O.—Bowen. (See 2421.)

621.396.933 **2433**

A Source of Error in Radio Phase Measuring Systems.—R. Bateman, E. F. Florman & A. Tait. (*Proc. Inst. Radio Engrs*, April 1951, Vol. 39, No. 4, pp. 436–438.) Discussion on 2515 of 1950.

621.396.933 **2434**

A General Survey of Electronics in Air Transport.—C. H. Jackson. (*J. Brit. Instn Radio Engrs*, April 1951, Vol. 11, No. 4, pp. 139–155. Discussion, pp. 156–159.) Communications, navigation aids and aids to approach and landing are reviewed and related to standards of safety. Details of technique and function are not considered.

621.396.933 **2435**

Radio on the Airways.—(*Wireless World*, May 1951, Vol. 57, No. 5, pp. 199–202.) A general description of m.f. omnidirectional beacons and radio ranges, and also v.h.f. marker beacons, as used on the main air routes in Great Britain. The beacons and radio ranges, operating in the 200–400-kc/s band, provide airway entrance markers and 4-direction course indication respectively. Position information is given by the vertically radiating v.h.f. marker system.

621.396.9 **2436**

Radar Systems and Components. [Book Review]—Bell Telephone Laboratories. Publishers: D. Van Nostrand Co., New York, & Macmillan & Co., London, $7.50 or 56s. (*Engineering, Lond.*, 6th April 1951, Vol. 171, No. 4445, p. 392.) A collection of papers, covering the magnetron, the klystron, the resonant cavity, radar aerials, etc. Valuable for the specialist and for the general student.

MATERIALS AND SUBSIDIARY TECHNIQUES

338.987.4 : 621.396 .397].6 **2437**

Conservation of Critical Materials.—W. MacD. (*Electronics*, April 1951, Vol. 24, No. 4, pp. 84–87.) Discussion of design modifications to economize in the use of scarce metals while maintaining receiver performance. Television receivers are particularly considered.

537.311.33 **2438**

Electrical Properties of Grey Tin.—G. Busch, J. Wieland & H. Zoller. (*Helv. phys. Acta*, 15th Feb. 1951, Vol. 24, No. 1, pp. 49–62. In German.) Grey tin of high purity was prepared by prolonged cooling of spectroscopically pure metallic tin, and numerous alloys were made by adding small amounts of Al. Conductivity was determined by measuring the Q-factor of a coil with a core of grey-tin powder, at frequencies up to 30 Mc/s; at 0°C the value found was $5.10^3\ \Omega^{-1}\ \text{cm}^{-1}$. Hall effect and variation of resistivity with applied magnetic field were measured by conventional d.c. methods; remarkably large variations of resistivity were observed. The experiments show that grey tin is a semiconductor of high electrical conductivity, with properties very similar to those of Si and Ge.

537.311.33 **2439**

The Effect of Pressure on the Electrical Resistance of certain Semi-Conductors.—P. W. Bridgman. (*Proc. Amer. Acad. Arts Sci.*, April 1951, Vol. 79, No. 3, pp. 127–148.) Measurements made on Ge, Si and several oxides are described. The resistances of all the oxides decrease with rising temperature up to 200°C, but there is no common type of variation with change of pressure up to 50 000 kg/cm². The Ge and Si were investigated under hydrostatic conditions to 30 000 kg/cm² at room temperature only. Differences of behaviour as between *n*- and *p*-types are indicated and discussed.

537.311.33 **2440**

The Diffusion of the Current Carriers in Semiconductors with Mixed Conductivity.—V. A. Lashkarev. (*C. R. Acad. Sci. U.R.S.S.*, 11th Aug. 1950, Vol. 73, No. 5, pp. 929–932. In Russian.) It is shown that 'bi-polar' diffusion is due to a thermodynamically unbalanced state. The conditions are derived under which such a state is established, all excitation except thermal being excluded.

38.221 **2441**

Interaction between the *d*-Shells in the Transition Metals: Part 2 — Ferromagnetic Compounds of Manganese with Perovskite Structure.—C. Zener. (*Phys. Rev.*, 1st May 1951, Vol. 82, No. 3, pp. 403–405.) A discussion of the correlation between conductivity and ferromagnetism found by van Santen & Jonker (656 of March).

538.221 **2442**

Ferromagnetism in the Manganese-Indium System.—W. V. Goeddel & D. M. Yost. (*Phys. Rev.*, 15th May 1951, Vol. 82, No. 4, p. 555.) About 25 alloys were prepared, with Mn contents ranging from 3 to 91% by weight in steps of about 4%; over half (3 to 50% Mn) were found to show ferromagnetism, believed to be due to a single phase (Mn_2In). Alloys containing up to 49% Mn appear to be composed of In + Mn_2In, no eutectic being formed.

538.221 **2443**

An Investigation of the Magnetic Properties of Alloys of Manganese with Nickel and Cobalt.—F. Gal'perin. (*C. R. Acad. Sci. U.R.S.S.*, 1st Dec. 1950, Vol. 75, No. 4, pp. 515–518. In Russian.)

538.221 **2444**

An Investigation of the Magnetic Properties of Well Ordered Alloys.—F. Gal'perin. (*C. R. Acad. Sci. U.R.S.S.*, 11th Dec. 1950, Vol. 75, No. 5, pp. 647–650. In Russian.)

538.221 : 621.395.625.3 **2445**

Mixed Ferrites for Recording Heads.—R. Herr. (*Electronics*, April 1951, Vol. 24, No. 4, pp. 124–125.) Short discussion of the advantages of using ferrite materials instead of laminations in magnetic recording heads.

538.249 **2446**

Variation with Frequency of the Magnetic After-Effect in Powder Cores.—R. Feldtkeller & H. Hettich. (*Z. angew. Phys.*, Dec. 1950, Vol. 2, No. 12, pp. 494–499.)

546.431.82 **2447**

Jumps in the Conductivity of Barium Titanate.—N. A. Tolstoi. (*Zh. tekh. Fiz.*, Aug. 1950, Vol. 20, No. 8, pp.970–974.)

546.431.82 **2448**

X-Ray Investigations of the Ferroelectricity of Barium Titanate.—W. Känzig. (*Helv. phys. Acta*, 10th April 1951, Vol. 24, No. 2, pp. 175–216. In German.)

546.431.82 **2449**

The Nature of Electromechanical Oscillations in $BaTiO_3$ Ceramics.—N. A. Roi. (*C. R. Acad. Sci. U.R.S.S.*, 11th Aug. 1950, Vol. 73, No. 5, pp. 937–940. In Russian.) Discussion, with experimental curves, is presented for the following cases: (1) alternating field applied to non-polarized sample (quasi-electrostriction); (2) weak alternating field applied to weakly polarized sample (linearized quasi-electrostriction); (3) weak alternating field applied to strongly polarized sample (linear piezoelectric effect).

546.431.82 : 548.55 **2450**

Elastic and Electromechanical Coupling Coefficients of Single-Crystal Barium Titanate.—W. L. Bond, W. P. Mason & H. J. McSkimin. (*Phys. Rev.*, 1st May 1951, Vol. 82, No. 3, pp. 442–443.) A report of measurements made on large multi-domain single crystals.

548.0 : 537 **2451**

Ferroelectricity.—B. T. Matthias. (*Science*, 25th May 1951, Vol. 113, No. 2943, pp. 591–596.) A general discussion of known ferroelectric materials and an examination of explanatory theories that have been advanced.

549.514.51 **2452**

Zero-Temperature-Coefficient Quartz Crystals for Very High Temperatures.—W. P. Mason. (*Bell Syst. tech. J.*, April 1951, Vol. 30, No. 2, pp. 366–380.) Crystals with zero temperature coefficient of frequency were obtained by making measurements of a series of rotated Y-cuts in the thickness shear mode and a series of rotated X-cuts in the longitudinal length mode, and hence determining the orientation for AT-, BT-, CT- and DT-type crystals with low temperature coefficients passing through zero value at a prescribed temperature. Calculations are given for crystals operating at 200°C. An AT-type crystal was investigated experimentally, and the calculated results agreed reasonably well with measured values. The maximum temperature at which the temperature coefficient of an AT-type crystal can have zero value is 190°C.

620.193.21 : 679.5 **2453**

Outdoor Weather Aging of Plastics under Various Climatological Conditions.—S. E. Yustein, R. R. Winans & H. J. Stark. (*ASTM Bull.*, April 1951, No. 173, pp. 31–43. Discussion, p. 43.) A report on electrical and mechanical tests carried out on five types of clear transparent sheet plastics, six types of laminated material and five types of moulded terminal bars, after prolonged exposure to tropical, dry-desert, temperate, subarctic and arctic conditions.

621.3.015.5 : 621.315.61 **2454**

Electrical Breakdown over Insulators in High Vacuum.—P. H. Gleichauf. (*J. appl. Phys.*, May 1951, Vol. 22, No. 5, pp. 535–541.) Experimental investigations in the pressure range 5×10^{-3}–10^{-7} mm Hg are described.

621.315.61 : 539.23 **2455**

The Electric Tunnel Effect acróss Thin Insulator Films in Contacts.—R. Holm. (*J. appl. Phys.*, May 1951, Vol. 22, No. 5, pp. 569–574.) Previous calculations apply to either very weak or very strong electric fields. The important practical case of intermediate-strength fields is here considered. The image force is neglected, but it is shown how this can be allowed for approximately. Calculated values of tunnel resistivity are plotted against applied voltage for metallic and for semiconducting contact members. Some inadequacies of the theory are discussed.

621.315.61.011.5 **2456**

"Heat Developed" and "Powder" Lichtenberg Figures and the Ionization of Dielectric Surfaces produced by Electrical Impulses.—A. M. Thomas. (*Brit. J. appl. Phys.*, April 1951, Vol. 2, No. 4, pp. 98–109.) Some experiments on both types of figure are reported and their characteristics outlined. An explanation is suggested of the mode of formation of 'heat developed' figures which are associated with the state of the surface of certain kinds of solid dielectrics. 'Powder' figures are used to investigate the effect of repeated impulses of alternating polarity, and they show that the effect of a discharge of given polarity is not cancelled by a succeeding discharge of opposite polarity. The phenomena are discussed in relation to theories of surface breakdown and spark discharge.

621.315.612 **2457**

Contribution to the Study of Physico-Chemical Phenomena in the Ceramics Industry.—R. Lecuir. (*Ann. Radioélect.*, Jan. 1951, Vol. 6, No. 23, pp. 20–50.) A discussion of the forming and sintering of oxides not possessing the plasticity characteristic of clays, which therefore require organic additions to the mix varying according to the forming technique used. The effect of the state of aggregation of the initial powder and of the application of high pressures on the compactness of the product are examined, together with other factors affecting the amount of shrinkage on sintering and the sintering temperature required.

621.315.612.011.5 **2458**

Dielectric Losses in Ceramic Dielectrics and in Barium Titanate at High Frequencies.—A. L. Khodakov. (*Zh. tekh. Fiz.*, May 1950, Vol. 20, No. 5, pp. 529–533.) Measurements were made of tan δ at frequencies from 10 to 200 Mc/s and at temperatures from 15° to 180° C. In the case of $BaTiO_3$, tan δ decreases with temperature but remains practically constant within the frequency range specified.

621.318.2 **2459**

The Determination of the Optimum Parameters of Magnetic Systems with Permanent Magnets.—A. Ya. Sochnev. (*C. R. Acad. Sci. U.R.S.S.*, 1st Jan. 1951, Vol. 76, No. 1, pp. 65–68. In Russian.)

621.775 **2460**

Special Nickel-Iron Alloys prepared by Powder Metallurgy.—Thien-Chi N'Guyen & B. Michel. (*Ann. Radioélect.*, Jan. 1951, Vol. 6, No. 23, pp. 3–19.) The first of a series of articles on the manufacture of magnetic materials, particularly Ni-Fe alloys. A brief general discussion of ferromagnetism is presented. The superiority of powder-metallurgy techniques for producing these alloys is shown by a study of the surface texture and crystal structure of a sintered 50%-Ni alloy.

666.1.037.5 **2461**

The Physical Aspect of Glass-Metal Sealability in the Electronic Tube Industry.—G. Trébuchon & J. Kieffer. (*Glass Ind.*, April–June 1951, Vol. 32, Nos. 4–6, pp. 165–174..202, 240–247..255 & 290–295.) English translation of paper noted in 2253 of 1950, 135 of January and 929 of April.

666.2 : 549.623.5 **2462**

Glass/Mica Vacuum-tight Seals.—J. Labeyrie & P. Léger. (*Le Vide*, Jan. 1951, Vol. 6, No. 31, pp. 936–940.) See 1437 of 1950 (Labeyrie).

537.311.33 **2463**

Electrons and Holes in Semi-Conductors. [Book Review]—W. Shockley. Publishers: D. Van Nostrand Co., New York, 1950, 543 pp., $9.75. (*Proc. Inst. Radio Engrs*, April 1951, Vol. 39, No. 4, pp. 449–450.) "This excellent book might almost be said to consist of a set of three monographs of increasingly rigorous treatment. Part I, 'Introduction to Transistor Electronics', is entirely descriptive. . . Part II. . . is 'Descriptive Theory of Semiconductors'. . . In part III, 'Quantum Mechanical Foundations', many of the concepts presented in earlier portions of the book are subjected to rigorous examination."

549.514.51 : 621.396.611.21 **2464**

Quartz Vibrators and their Applications. [Book Review]—P. Vigoureux & C. F. Booth. Publishers: H.M. Stationery Office, London, 30s. (*J. Brit. Instn Radio Engrs*, April 1951, Vol. 11, No. 4, p. viii.) The properties of quartz, manufacture of crystals and applications in telecommunications, etc., are considered.

MATHEMATICS

512.831 **2465**

The Principle of Minimized Iterations in the Solution of the Matrix Eigenvalue Problem.—W. E. Arnoldi. (*Quart. appl. Math.*, April 1951, Vol. 9, No. 1, pp. 17–29.)

517.534 : 538.566 **2466**

On the Method of Saddle Points.—B. L. van der Waerden. (*Appl. sci. Res.*, 1951, Vol. B2, No. 1, pp. 33–45.) In attempting to solve the problem of radio propagation over a plane earth earlier investigators encountered a difficulty in evaluating the integral $\int_c we^{-\lambda u}dv$ by the method of steepest descents, because of the proximity of a pole to a saddle point. The problem is here transformed to one of integration in the u-plane. A solution is obtained in which the part of the integral corresponding to the pole can be readily separated out.

681.142 **2467**

Mechanized Reasoning. Logical Computors and their Design.—D. M. McCallum & J. B. Smith. (*Electronic Engng*, April 1951, Vol. 23, No. 278, pp. 126–133.)

681.142 **2468**

Visual Presentation of Binary Numbers.—E. H. Lenaerts. (*Electronic Engng*, April 1951, Vol. 23, No. 278, pp. 140–141.) By using a raster timebase, the pulses representing a number can be displayed as vertical deflections of the trace, or by a pattern of bright dots on a background of fainter dots representing the zeros. In a better method described the pulse train is superimposed upon the frame-deflection timebase, while the brightness of the spot is modulated by clock pulses timed to occur in all positions where a spot is possible. The pulses thus appear as short vertical lines with dots interposed which represent the zeros.

681.142 : 517.9 **2469**

Solution of a System of Linear Equations with a Slightly Unsymmetrical Matrix by using a Network Analyzer.—H. L. Knudsen. (*Trans. Dan. Acad. tech. Sci.*, 1950, No. 2, 16 pp. In English.) The method is based on iteration, only a few steps being necessary when the asymmetry is only slight. No equipment other than the network analyser is required.

681.142 : 621.316.726 **2470**

Automatic Frequency Control.—Pinkerton. (See 2364.)

MEASUREMENTS AND TEST GEAR

531.765 : 529.786 **2471**

Comparing Outputs from Precision Time Standards.—J. M. Shaull & C. M. Kortman. (*Electronics*, April 1951, Vol. 24, No. 4, pp. 102–107.) Description of equipment developed at the National Bureau of Standards for monitoring the time-keeping of a group of standard quartz clocks. The chronograph records time differences of two clocks to within 1 ms, using spark-generating equipment whose rate is controlled by one clock while the drum speed is governed by the other. A motor-driven switch connects each of several clocks in turn to the spark generator every 15 minutes, thus providing intercomparison data. The chronoscope uses a 3-in. c.r. tube, one clock and frequency divider being applied to produce a circular sweep with small fixed marker dots at 0·1-ms intervals, while a pulse from the circuit of the second clock produces a larger bright spot on the sweep. The chronograph and chronoscope are locked in time phase, so that observation of the position of the bright spot enables the time difference between the two clocks to be determined to within 20 μs. See also *Tech. Bull. nat. Bur. Stand.*, Jan. 1951, Vol. 35, No. 1, pp. 14–16.

535.322.4 : 546.217 **2472**

A Phase-Shift Refractometer.—C. W. Tolbert & A. W. Straiton. (*Rev. sci. Instrum.*, March 1951, Vol. 22, No. 3, pp. 162–165.) A description of apparatus for measuring small changes in the dielectric constant or refractive index of air by determining the phase change of a 9·375-kMc/s wave over a 3-ft path. The test path may be confined to a waveguide through which air is drawn, or it may be the space between two aerial systems. The waveguide system can be used to measure rapid changes in the refractive index with an error <1 part in 10^6; the aerial method has larger errors because of flexibility of supports and external reflections.

621.317.083.4 : 621.396.645 **2473**

Sensitive Null Detector.—Scroggie. (See 2390.)

621.317.335.3† + 621.317.374].029.64 **2474**

Measurement of Dielectric Constant and Losses of Solid Dielectrics by means of Waveguides.—G. D. Burdun. (*Zh. tekh. Fiz.*, July 1950, Vol. 20, No. 7, pp. 813–821.) Waves of the H_{01} mode are excited in a rectangular waveguide the end of which is closed by a piece of the dielectric under investigation. The distribution of the dielectric field intensity inside the waveguide is measured by means of a probe and indicator, and from these measurements the properties of the dielectric are determined. The theory of the method is discussed and the results of measurements with various dielectrics on wavelengths between 1·6 and 3·2 cm are presented. The accuracy of the method is to within about 1 or 2%.

621.317.35 : 621.397.5 **2475**

Notes on TV Waveform Monitor Frequency Response.—W. L. Hurford. (*Proc. Inst. Radio Engrs*, May 1951, Vol. 39, No. 5, pp. 562–563.) A comparison of monitors having (*a*) very wide response band, (*b*) the response specified in the I.R.E. Standards on Television (see 2035 of 1950) and (*c*) a sharp cut-off at twice the bandwidth of the I.R.E. curve. From the response of these devices to clean, sharp pulses and to pulses with spikes it is concluded that "the I.R.E. response is an excellent choice."

621.317.39 : [621.318.4 + 621.319.4 + 621.396.611.1 **2476**

Devices for the Measurement of the Temperature Coefficients of Coils, Capacitors and Oscillatory Circuits.—C. Schreck. (*Fernmeldetech. Z.*, Jan. 1951, Vol. 4, No. 1, pp. 30–36.) Review of the development of methods and apparatus necessitated by the continual demand for higher frequency-constancy; 'static' temperature coefficients (i.e., those related to external heating effects) have received more attention than 'dynamic' coefficients (related to internal heating effects).

621.317.4 : 621.317.755 **2477**

The Electron-Beam Ferroscope.—P. E. Klein. (*Arch. tech. Messen*, Feb. 1951, No. 181, pp. T23–T24.) Description and circuit details of a c.r.o. unit for displaying the magnetization curve of high-permeability iron-alloy samples. The voltage drop across a resistor in the test-circuit primary is applied to the X-plate amplifier; the output from the secondary is fed to the Y-plate amplifier either direct or through an integrating circuit. By means of a 3-position switch the waveform of the primary current, secondary voltage or B-H characteristic of the magnetic circuit can be displayed. Typical traces are shown.

621.317.6 : 621.396.645 **2478**

The Determination of Amplifier Sensitivity with the Aid of the Noise Diode.—W. K. Squires. (*Sylvania Technologist*, April 1951, Vol. 4, No. 2, pp. 35–37.) A measure of amplifier performance designated 'sensitivity factor' is introduced; it is expressed quantitatively as the ratio of standard noise output to actual noise output at maximum gain, and its use enables the gain and noise factor to be correlated. A method of measuring the sensitivity factor is described.

621.317.7 + 621.396.611].029.63 **2479**

Circuits and Measurement Apparatus for the 30-cm Band.—E. Safa. (*Onde élect.*, Jan. 1951, Vol. 31, No. 286, pp. 33–43.) Illustrated review outlining the characteristics of valves and associated coupling circuits and giving details of generators, wavemeter, wattmeter and curve tracer designed for the u.h.f. band.

621.317.715 : 621.396.611.33/.34 **2480**

Coupling of A.C. Galvanometer to A.C. Amplifier.—C. T. J. Alkemade & P. M. Endt. (*Appl. sci. Res.*, 1951, Vol. B2, No. 1, pp. 46–52.) A galvanometer with a 50-c/s magnetic field is considered. Untuned transformer coupling provides higher gain than capacitor coupling, but magnetic relaxation phenomena in the transformer cause slow changes in sensitivity which make capacitor coupling preferable.

621.317.725 **2481**

An Instantaneous Peak Voltmeter.—M. W. Tobin, H. Grundfest & R. L. Schoenfeld. (*Rev. sci. Instrum.*, March 1951, Vol. 22, No. 3, pp. 189–190.) The operation of the diode/capacitor peak-voltmeter circuit is discussed, and a circuit is described which provides a measurement of pulse amplitude unaffected by the amplitude of the previous pulse; this is used for investigating pulses of duration 0·1–20 ms at frequencies as low as 0·2 c/s.

621.317.725 **2482**

Audio-Frequency Valve Voltmeter.—S. Kelly. (*Wireless World*, June 1951, Vol. 57, No. 6, pp. 215–218.) Details are given of a self-calibrating, portable instrument designed for a voltage range of 1 mV to 10 V in four decades, with an input impedance $>10\ \text{M}\Omega$ across 10 pF and an output impedance $<500\ \Omega$.

621.317.726 : 621.3.011.6 **2483**

Calculation of CR Elements for the case of Varying Voltage and/or Nonlinear Resistances.—Elger. (See 2356.)

621.317.76 : 621.396.615 **2484**

An Instrument for Recording the Frequency Drift of an Oscillator.—W. W. Boelens. (*Philips tech. Rev.*, Jan. 1951, Vol. 12, No. 7, pp. 193–199.) The meter was designed for measuring the frequency variation of the local oscillator of f.m. receivers in the 88–108-Mc/s band. The reference frequencies, of which there are 10 in the band 80·4–118·5 Mc/s, are obtained from a 4·232-Mc/s crystal oscillator by a multiplication and mixing process. The frequency drift is indicated by the variation of a direct current which can be read on a d.c. meter or applied to a recording instrument.

621.317.79 : 621.3.018.78† : 621.396.61 **2485**

Measurement of Distortion in Broadcast Transmitters.—Müller. (See 2566.)

621.317.79 : 621.396.67 **2486**

A Phase Front Plotter for Testing Microwave Aerials.—C. A. Cochrane. (*Elliott J.*, March 1951, Vol. 1, No. 1, pp. 29–30.) A search aerial, servo-controlled via a r.f. phase discriminator, is used to find lines of constant phase, accurate to within about $\pi/16$ near the aerial. The search aerial is an open-circuited circular waveguide, suitable for wavelengths near 3·2 cm.

621.317.79 : 621.396.822 **2487**

High-Sensitivity High-Frequency Noise-Measurement Apparatus Calibrated Absolutely in kT_0 Units.—H. Röschlau. (*Arch. elekt. Übertragung*, Oct. 1950, Vol. 4, No. 10, pp. 427–434.) Requirements for a receiver to have high sensitivity, appropriate for the investigation of cosmic noise sources on a wavelength of 1·5 m, are examined. The choice of input valve and input circuit are discussed in detail, a cavity-resonator tank circuit being used on account of its high resonance resistance, together with a pentode with regeneratively coupled screen grid. Three alternative methods of performing the absolute calibration are described; a method using a noise diode Type SA102 was most exact and gave a value better than 0·05 kT_0 for the sensitivity. Details are given of the method used for coupling the receiver to the aerial array.

621.317.799† : 621.385.012 **2488**

Tube Characteristic Tracer using Pulse Techniques.—H. M. Wagner. (*Electronics*, April 1951, Vol. 24, No. 4, pp. 110–114.) A full description of an instrument designed primarily to obtain characteristic curves in the positive-grid region for small valves used at high pulse power levels. The curves are displayed on a c.r.o. screen and can be recorded photographically. See also 2576 of 1950 (Leferson) and 692 of March (Graffunder & Schultes).

621.396.615.015.7.001.4† **2489**

Radar Test Generator.—K. S. Stull. (*Electronics*, April 1951, Vol. 24, No. 4, pp. 93–95.) Circuit details and description of equipment providing triggered or free-running pulses of duration 0·25, 0·5 or 1·0 µs and also c.w. signals in the range 47–76 Mc/s, for testing wide-band circuits. The output voltage is variable from 0·1 µV to 0·1 V.

537.7 **2490**

Electrical Measurements and the Calculation of the Errors Involved: Part 1. [Book Review]—D. Karo. Publishers: Macdonald & Co., London, 1950, 191 pp., 18s. (*Nature, Lond.*, 12th May 1951, Vol. 167, No. 4254, p. 745.) ". . . an extremely useful book, particularly for finals students, research workers and engineers."

621.317.755 **2491**

Encyclopedia on Cathode-Ray Oscilloscopes and their Uses. [Book Review]—J. F. Rider & S. D. Uslan. Publishers: J. F. Rider, New York, 1950, 992 pp., $9.00. (*Electronics*, April 1951, Vol. 24, No. 4, pp. 282, 284.) "Although the authors explain in the foreword to this book that some readers having special interest may find that it has limited coverage, the book quite adequately backs up its title for the average reader. . . A novel aid. . . is a collection of synthesized waveform patterns, a total of 1580 extending over 79 pages. These are provided for those readers who do not have a harmonic wave analyzer available."

621.396.615.17 **2492**

Time Bases (Scanning Generators). [Book Review]—O. S. Puckle. Publishers: Chapman & Hall, London, 2nd edn, 387 pp., 30s. (*Electrician*, 6th April 1951, Vol. 146, No. 3799, p. 1131.) This edition includes a new chapter on Miller capacitance timebases, as well as many other modifications and additions to the original 1943 edition.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

620.179.16 **2493**

Nondestructive Testing of Large Forgings by an Ultrasonic Method.—W. Felix. (*Schweiz. Arch. angew. Wiss. Tech.*, April 1951, Vol. 17, No. 4, pp. 107–113.) An account of experience gained over a period of several years in the operation of actual tests, using pulse-type equipment.

621.316.7.076.7 + 621-526 **2494**

Control Systems and their Application to Industry.—W. R. Blunden. (*J. Instn Engrs Aust.*, April/May 1951, Vol. 23, Nos. 4/5, pp. 89–94.) A general survey, in which the different types of control system are classified both according to operation and application, and some of the more important types are described briefly.

621.317.083.7 : 551.508.1 **2495**

Cosmic-Ray Radiosonde and Telemetering System.—M. A. Pomerantz. (*Electronics*, April 1951, Vol. 24, No. 4, pp. 88–92.) Details of equipment comprising four G-M counters which trigger a multivibrator controlling the keying of an u.h.f. transmitter. Altitude and temperature are indicated by modulation intervals and modulation frequency.

621.365.5 **2496**

High-Frequency Generators and their Applications.—W. Burkhardtsmaier. (*Telefunken Ztg*, Sept. 1950, Vol. 23, Nos. 87/88, pp. 73–82.) Commercially available generators with outputs ranging from 1·5 to 70 kW are described and illustrated. Industrial heating applications are considered. A frequency of 400 kc/s is used for inductive heating, and 20 Mc/s for dielectric heating.

621.365.54† : 621.793 **2497**

Metal Evaporator uses High-Frequency Heating.—R. G. Picard & J. E. Joy. (*Electronics*, April 1951, Vol. 24, No. 4, pp. 126–128.) Difficulties due to non-wetting of filament and its reaction with the material to be evaporated are avoided by the use of a water-cooled h.f. heating coil. The method enables metals to be evaporated that would react with the usual tungsten or tantalum filament.

621.383 : 551.576 **2498**

Electronic Theory for Design of a Radar System using Light Waves (in particular for a Cloud Height Indicator).—A. Baude. (*Onde élect.*, Jan. & Feb. 1951, Vol. 31, Nos. 286 & 287, pp. 44–48 & 90–101.) The principle and theory of distance measurement by pulsed light waves are discussed. An illustrated description is given of two sets of apparatus developed for measurement of cloud height and estimation of layer thickness. At 1 500 m, error may be about 10 m. Clear echoes from above 10 km have been obtained despite intervening cloud layers.

621.384.611.2† **2499**

The 300 MeV Synchrotron at the Massachusetts Institute of Technology.—(*Engineer, Lond.*, 6th April 1951, Vol. 191, No. 4967, pp. 440–442.) A general account.

621.385.833 **2500**

Theory of the Three-Electrode Electrostatic Lens.—E. Regenstreif. (*Ann. Radioélect.*, Jan. & April 1951, Vol. 6, Nos. 23 & 24, pp. 51–83 & 114–155.) An expanded account of work noted in 1213, 1743 and 2314 of 1950, 1455 of June and 1742 of July.

621.385.833 **2501**

The Visualization of Atomic Distances by means of the Electron Microscope.—L. Wegmann. (*Helv. phys. Acta*, 15th Feb. 1951, Vol. 24, No. 1, pp. 63–71. In German.) By stopping out certain concentric lens zones it is theoretically possible to effect an improvement of the resolving power of uncorrected electron lenses sufficient to render visible the atomic lattices of crystals. The practical difficulties are explained briefly.

621.385.833 **2502**

Scattering Phenomena in Electron Microscope Image Formation.—C. E. Hall. (*J. appl. Phys.*, May 1951, Vol. 22, No. 5, pp. 655–662.)

621.385.833 **2503**

Electronoptical Theory of the Deflection of an Extended Electronoptical Image by means of Crossed Electrical Deflection Systems.—J. Himpan. (*Ann. Phys., Lpz.*, 15th Feb. 1951, Vol. 8, Nos. 5/8, pp. 405–422.) The discussion presented is valid for mutually perpendicular deflection systems with plates of any type and arrangement, and takes third-order aberrations into account. The conditions under which deflection in two dimensions can be performed without introducing distortion are established. For deflections greater than those permitted in the ideal case six different types of defect are recognized, viz., three types of overall image distortion and three of image-point deformation. Simple formulae are derived expressing these defects. Their magnitudes in particular cases are calculated by use of constants which can be determined from a few measurements.

621.385.833 **2504**

Calculation of the Optical Constants of Powerful Magnetic Electron Lenses.—W. Glaser. (*Ann. Phys., Lpz.*, 15th Feb. 1951, Vol. 8, Nos. 5/8, p. 423.) Corrections to paper noted in 703 of March.

621.385.833 : 537.533.72 **2505**

The Significance of the Concepts 'Focus' and 'Focal Length' in Electron Optics, and Strong Electron Lenses with Newtonian Image-Formation Equation.—W. Glaser & O. Bergmann. (*Z. angew. Math. Phys.*, 15th Nov. 1950, Vol. 1, No. 6, pp. 363–379.) The functions determining the relation between the position of the object and that of the image, for linear magnification, are more complex in the case of electron optics than in the case of light. These functions are approached, in the neighbourhood of two conjugate points, by the Newtonian osculating equation for the image function, including terms up to the fourth order. Each pair of conjugate points thus possesses corresponding foci, principal points and focal lengths resulting from the Newtonian equation. If the osculating cardinal elements are independent of the pair of conjugate points chosen, they characterize by themselves the formation of the image and are identical to the magnitudes defined in the usual way for light. Such fields, for which the image-formation equation of ordinary optics is strictly valid, are termed 'fields with Newtonian representation'.

A study is made of these strong fields and examples of them are given which approximate to the fields actually existing in electron lenses. In order to keep a physical significance for focal length, whatever the magnification, as close an approximation as possible to the empirical field must be obtained by one of the Newtonian type. A method for doing this is indicated and experimental methods of determining the focal points and focal lengths of such approximate fields are examined.

621.385.833 : 621.311.1 **2506**

Modern Low-Power High-Voltage Generators.—J. Vastel. (*Ann. Radioélect.*, Jan. 1951, Vol. 6, No. 23, pp. 84–94.) Various forms of generator suitable for supplying voltages in the range 30–80 kV, constant to within about 1 part in 100 000, are discussed and the causes of ripple in the output are analysed. Particular equipments described for supplying electron microscopes etc., use low- and high-frequency oscillators (600 c/s–46 kc/s), in association with voltage multipliers, giving output currents of about 100 μA and voltages of 60–80 kV.

621.386 **2507**

The Intensification of X-Ray Fluorescent Images.—W. S. Lusby. (*Elect. Engng, N.Y.*, April 1951, Vol. 70, No. 4, pp. 292–296.) Paper given at A.I.E.E. Winter General Meeting, New York, Jan. 1951. The intensifier has a Cs-Sb photocathode arranged close to the ZnS input screen, an accelerating voltage of 30 kV causing the emitted photoelectrons to impinge on an Al-backed Zn-CdS output screen of reduced size at the far end of the 17-in. tube. Requirements for medical and industrial applications are discussed. Experimental installations giving a brightness amplification of slightly over 100 times have been put into operation.

621.386.1 : 621.385 **2508**

Radiographic Examination of Electronic Valves.—H. B. van Wijlen. (*Philips tech. Rev.*, Jan. 1951, Vol. 12, No. 7, pp. 207–209.) The examination of the electrode structure of valves by means of X rays is described. A resolution of 6 μ is obtained with an image of approximately the same size as the object.

621.387.4† **2509**

A Secondary-Electron Photon Counter.—S. F. Rodionov & A. L. Osherovich. (*C. R. Acad. Sci. U.R.S.S.*, 21st Sept. 1950, Vol. 74, No. 3, pp. 461–463. In Russian.) A description is given of a photomultiplier device for counting 'visible' photons (λ = 3 600–6 500Å).A Sb-Cs photocathode developed by Kubetski is used, and light fluxes of the order of 10^{-14}–10^{-15} lumens can be measured.

621.387.4† **2510**

The Discharge Mechanism for Oversize Pulses in Counters with Vapour Filling.—H. Neuert. (*Ann. Phys., Lpz.*, 15th Feb. 1951, Vol. 8, Nos. 5/8, pp. 341–349.)

621.387.462† **2511**

Silver Bromide Crystal Counters.—K. A. Yamakawa. (*Phys. Rev.*, 15th May 1951, Vol. 82, No. 4, pp. 522–526.)

621.387.464† **2512**

The Scintillation Counter.—W. Hanle. (*Naturwissenschaften*, April 1951, Vol. 38, No. 8, pp. 176–185.) A survey of the development and applications of photomultiplier-type counters, with a list of 156 references.

621.396.645 : 535.247.4 **2513**

A Balance Indicator with High Input Impedance using a Cathode Follower.—D. T. R. Dighton. (*J. sci. Instrum.*, April 1951, Vol. 28, No. 4, pp. 101–102.) "The use of the cathode follower circuit with high value grid resistances is discussed and it is shown that a high-slope pentode can give an impedance conversion of 10^7. A cathode follower circuit suitable for a null-balance indicator for photometric work is described. The input impedance is 500 MΩ, the grid current about 10^{-10}A and the detection limit 1 to 2 mV change of grid potential. A simple method of compensating for slow variations of heater voltage is employed."

621.38 **2514**

Survey of Modern Electronics. [Book Review]—P. G. Andres. Publishers: J. Wiley & Sons, New York, 1950, 522 pp., $5.75. (*J. appl. Phys.*, May 1951, Vol. 22, No. 5, pp. 685–686.) "Written as a text for a short survey course for students in electrical engineering. . ."

621.365.54† **2515**

Induction Heating. [Book Review]—N. R. Stansel. Publishers: McGraw-Hill, London, 1949, 212 pp., 34s. (*Nature, Lond.*, 5th May 1951, Vol. 167, No. 4253, p. 700.) Of the nature of a handbook giving formulae and data relating to the electrical and thermal quantities involved.

PROPAGATION OF WAVES

538.566 **2516**

Is there a Zenneck Wave in the Field of a Radiator?—H. Ott. (*Arch. elekt. Übertragung*, Jan. 1951, Vol. 5, 621.396.61/.62 : 623.6 **2540**

Progress in Military (Land Forces) Radiocommunications.—Morand. (*Onde élect.*, Jan. 1951, Vol. 31, No. 286, pp. 3–17.) Description of equipment of American pattern selected for production in France from 1945 onward. Concise details and illustrations are given of (*a*) three short-range telephony intercommunication sets, two using f.m., the third ph.m. for use in vehicles; (*b*) two medium-range portable W/T-R/T sets using a.m.; (*c*) long-range equipment comprising a mobile station with trailer power unit; power is 250 W for R/T, 400 W for W/T, frequency range 2–18 Mc/s; a 4-channel multiplex mobile station with transmitter power 50 W, designed to replace land-line systems. The trend of technical development in miniaturization, tropicalization, etc., is outlined.

621.396.65 **2541**

Short-Range Communication by V.H.F. Radio.—(*G.E.C. Telecommun.*, 1946, Vol. 1, No. 2, pp. 61–79.) A summary of the factors governing the choice of systems for point-to-point and mobile services. Advantages of the v.h.f. band mentioned are the reduced noise level, the convenient size of aerial systems and the restriction of transmissions to the service area. Simplex, duplex and relay systems using both a.m. and f.m. are discussed, and a complete range of equipment for both fixed and mobile stations is described.

621.396.712 **2542**

150-kW Medium-Wave Broadcast Transmitter at Daventry.—(*Engineering, Lond.*, 27th April 1951, Vol. 171, No. 4448, pp. 506–507.) A description of the new British Third Programme transmitter and aerial system. The transmitter, which uses air-cooled valves, consists of two identical 100-kW units which can be paralleled. The fading-free area is increased by connecting the coaxial feeder across an insulator at a point 460 ft (about two-thirds of the height) up the mast radiator.

621.396.933 **2543**

A General Survey of Electronics in Air Transport.—Jackson. (See 2434.)

621.396.933 **2544**

The M.C.A. [Ministry of Civil Aviation] **V.H.F. Area Coverage Network: Audio Frequency Distribution.**—J. L. French. (*Electronic Engng*, April 1951, Vol.23, No. 278, pp. 146–148.) General description, with block diagram, of the a.f. equipment and its operation. See also 2545 below.

621.396.933 **2545**

The M.C.A. [Ministry of Civil Aviation] **V.H.F. Area Coverage Network: Provision of Transmitting Station Equipment.**—D. H. C. Scholes. (*Electronic Engng*, April 1951, Vol. 23, No. 278, pp. 148–150.) General description of the modified Type-T.1131 transmitter and its temperature-controlled crystal unit. See also 2544 above.

SUBSIDIARY APPARATUS

621-526 + 621.316.7.076.7 **2546**

Control Systems and their Application to Industry.—Blunden. (See 2494.)

621-526 **2547**

Servomechanisms with Linearly Varying Elements.—M. J. Kirby. (*Elect. Engng, N.Y.*, April 1951, Vol. 70, No. 4, p. 343.) Digest of paper presented at the A.I.E.E. Fall General Meeting, Oklahoma, 1950. An analytical method is presented for determining the stability of a servomechanism in which one or more elements vary linearly with time.

621.316.722 **2548**

High-Voltage Stabilization by means of the Corona Discharge between Coaxial Cylinders.—S. W. Lichtman. (*Proc. Inst. Radio Engrs*, April 1951, Vol. 39, No. 4, pp. 419–424.) 1950 I.R.E. National Convention paper. The design and performance of corona-discharge voltage-regulator valves for operation with currents of 10–200 μA at voltages between 700 V and 40 kV are described. The dependence of the mode of operation and efficiency on circuit parameters is discussed.

621.396.6.017.71.012.3 **2549**

Estimating Temerature Rise in Electronic Equipment Cases.—R. J. Bibbero. (*Proc. Inst. Radio. Engrs*, May 1951, Vol. 39, No. 5, pp. 504–508.) The discussion is concerned with airborne equipment. Charts are presented as an aid in calculating temperature rises. Corrections are given for pressure variations, and the effects of case colour, high aircraft speed, etc., are considered.

621.396.78† **2550**

Power Supplies for Large Transmitters.—H. Kropp. (*Fernmeldetech. Z.*, Jan. 1951, Vol. 4, No. 1, pp. 25–30.) Review of different types of rectifiers for low- and high-voltage supplies.

TELEVISION AND PHOTOTELEGRAPHY

621.397 **2551**

Cathode-Ray Picture Telegraphy.—F. Schröter. (*Telefunken Ztg*, Sept. 1950, Vol. 23, Nos. 87/88, pp. 111–118.) Inherent tube and circuit factors tending to reduce the resolution attainable in practice in a c.r. tube are discussed. For picture telegraphy, 'flying-spot'-type scanning systems appear to be most suitable, used in conjunction with a c.r. tube at the receiver. The electron-optical development is described of systems of this type having the following properties: ability to transmit directly, by reflection scanning, unprepared material such as manuscripts, drawings and photographs; elimination of mechanical aspects capable of affecting quality of transmission; the possibility of varying the aspect ratio and of emphasizing particular parts of the material; immediate readability at the receiver on a long-lag screen producing the same sharpness and co-ordinate fidelity as at the transmitter.

621.397.5 **2552**

B.B.C. Television.—T. H. Bridgewater. (*Electronic Engng*, April 1951, Vol. 23, No. 278, pp. 120–125.) A brief history of the outside-broadcasts section of the B.B.C. television service, together with a comparison of the characteristics and performance of the cable and radio links used to convey pictures from the pickup point to the main transmitter at Alexandra Palace. Future developments which will greatly extend the scope of such broadcasts are outlined. See also 752 of March.

621.397.5 : 535.62 **2553**

Quality of Colour Reproduction.—D. L. MacAdam. (*Proc. Inst. Radio Engrs*, May 1951, Vol. 39, No. 5, pp. 468–485; *J. Soc. Mot. Pict. Televis. Engrs*, May 1951, Vol. 56, No. 5, pp. 487–512.) A discussion of methods of evaluating the quality of colour reproduction in television in which, as in methods already used in colour photography, subjective judgments are compared with colour measurements made, e.g., on the I.C.I. system.

621.397.5 : 621.317.35 **2554**

Notes on TV Waveform Monitor Frequency Response.—Hurford. (See 2475.)

621.397.5 : 778.5 **2555**

American Television Film Recording Equipment.—R. B. Hickman. (*J. Televis. Soc.*, Oct./Dec. 1950, Vol. 6, No. 4, pp. 167–169.) The method used with the R.C.A. Kinephoto equipment to perform the conversion from the 30 frames/sec of U.S. television to the 24 frames/sec of standard motion-picture projection is described. Either an electronic or a mechanical shutter can be used to blank out the image during the pull-down interval. Details of exposure and processing of the film are given.

621.397.611.2 **2556**

Television Camera Tubes.—E. L. C. White & J. D. McGee. (*Wireless Engr*, May 1951, Vol. 28, No. 332, pp. 163–164.) Comment on 1264 of May (Bedford).

621.397.62 **2557**

Some Aspects of Single Side-Band Receiver Design.—W. M. Lloyd. (*J. Televis. Soc.*, Oct./Dec. 1950, Vol. 6, No. 4, pp. 135–149.) The discrepancies which appear in the response of the receiver to a unit step are discussed theoretically in relation to those features of the frequency characteristics which give rise to them. The experimentally obtained step-responses of two typical receivers are shown.

621.397.62 : 621.385.2 : 546.289 **2558**

An Analysis of the Germanium Diode as Video Detector.—Whalley, Masucci & Salz. (See 2589.)

621.397.62 : 621.396.67 **2559**

Indoor Television Aerial.—H. Page. (*Wireless World*, May 1951, Vol. 57, No. 5, pp. 168–170.) The aerial consists of a horizontal slot, about $\lambda/2$ long, in a vertical conducting sheet, with a gain of 4 db over a vertical $\lambda/2$ dipole. Details of construction and performance are given, showing negligible change of gain, impedance and radiation patterns for a 10% frequency change.

621.397.621.2 **2560**

Material-Saving Picture Tube.—L. E. Swedlund & R. Saunders, Jr. (*Electronics*, April 1951, Vol. 24, No. 4, pp. 118–120.) The use of an e.s. focusing system instead of the usual magnetic type economizes in alnico-5 and copper. Performance of the new electron gun is at least equal to that of the magnetic type and may even be the better.

621.397.645 **2561**

New Video Circuits in Modern TV Sets.—E. M. Noll. (*Radio-Electronics*, April 1951, Vol. 22, No. 7, pp. 26–27.) Video amplifier circuits used by a number of U.S. manufacturers are illustrated and briefly described.

621.397.645 **2562**

Shunt-Regulated Amplifiers.—V. J. Cooper. (*Wireless Engr*, May 1951, Vol. 28, No. 332, pp. 132–145.) Describes circuits used for modulating television transmitters, and designed to produce a substantially constant load large voltage swings regulated to ensure faithful reproduction. Numerous variants of the circuit are classified and analysed. Practical applications and experimental results are given.

621.397.8.08 **2563**

U.H.F. TV Propagation Measurements.—K. H. Cook & R. G. Artman. (*Tele-Tech*, March & April 1951, Vol. 10, Nos. 3 & 4, pp. 50–51, 93 & 52–54, 82.) Measurements of peak field intensity of the vision signal and observations of relative picture quality were made at 130 locations within 25 miles of the experimental transmitter at Kansas City under typical broadcasting conditions. Vision frequency was 507.25 Mc/s, radiated power 3.450 kW. Equipment is described and results are reported and discussed.

621.396.615.17 **2564**

Time Bases (Scanning Generators). [Book Review]—Puckle. (See 2492.)

TRANSMISSION

621.396.61 **2565**

The First High-Power Transmitters built since 1945 [in Germany].—K. Müller. (*Telefunken Ztg*, Sept. 1950, Vol. 23, Nos. 87/88, pp. 31–38.) Descriptions, with block diagrams and valve details, are given for the following: (*a*) 100-kW broadcast transmitter, 150–300 kc/s, at Königs Wusterhausen; (*b*) 5-/2-kW broadcast transmitters, 545–1 500 kc/s, for northwest Germany; (*c*) 20-kW broadcast transmitter, 545–1 500 kc/s, at Potsdam and Hanover; (*d*) 100-kW broadcast transmitter, 525–1 610 kc/s, at Berlin-Britz; (*e*) 30-kW telegraphy transmitter, 100–150 kc/s, at Bad Vilbel, Frankfurt/Main; (*f*) 60-kW telegraphy transmitter, 75–150 kc/s, also at Bad Vilbel. Innovations as compared with pre-1945 practice include: thoriated instead of plain tungsten cathodes in the directly heated valves; single-circuit cooling; ignitron protecting devices for transmitters of power > 20 kW; a simple thermostat control for the quartz crystals, giving frequency constancy to within 10^{-7} over periods of 24 hours.

621.396.61 : 621.317.79 : 621.3.018.78† **2566**

Measurement of Distortion in Broadcast Transmitters.—H. Müller. (*Telefunken Ztg*, Sept. 1950, Vol. 23, Nos. 87/88, pp. 53–66.) Apparatus for the measurement of nonlinear distortion, developed by Telefunken from 1946 onwards, is described. The filter method of measuring harmonic distortion is adequate for the general monitoring of transmission quality; for more stringent requirements, particularly when investigating the nature of the distortion and its frequency dependence towards the upper transmission-frequency limit, a two-tone method such as that of von Braunmühl (456 of 1935) is used enabling symmetrical and asymmetrical distortion to be separated. For the range 30–150 c/s a search-tone method is used, enabling the individual harmonics to be separated.

621.396.61 : 621.385.4 **2567**

A 30-Watt Transmitter for 430 Mc/s employing the Transmitting Valve QQE 06/40 (AX 9903).—(*Philips tech. Commun., Aust.*, 1951, No. 2, pp. 14–17.) See also 2062 of August (Dorgelo & Zijlstra).

621.396.78† **2568**

Power Supplies for Large Transmitters.—H. Kropp. (*Fernmeldetech. Z.*, Jan. 1951, Vol. 4, No. 1, pp. 25–30.) Review of different types of rectifiers for low- and high-voltage supplies.

VALVES AND THERMIONICS

537.533.8 **2569**

Secondary Electron Emission from Aluminium Oxide.—A. R. Shul'man & I. Yu. Rozentsveig. (*C. R. Acad. Sci. U.R.S.S.*, 21st Sept. 1950, Vol. 74, No. 3, pp. 497–500.) A report on an experimental investigation of the effect of temperature on the secondary-emission coefficient of Al_2O_3. No variation with temperature was observed.

537.534.8 **2570**

Positive Emission from Thermionic Cathodes.—K. H. Steigerwald. (*Z. angew. Phys.*, Dec. 1950, Vol. 2, No. 12, pp. 491–493.) Excitation of the fluorescent screen of an electron microscope was observed even when the negative voltage applied to the control electrode was sufficient to cut off electron emission from the cathode. The effect was traced to the emission of positive ions which release secondary electrons at the control electrode, the ion current being of the order of 10^{-9}–10^{-8} A for the tungsten hairpin cathode used, and occurring only in the range of cathode temperatures 1 300°K–1 800°K. The emission is thought to depend on a vaporization process.

537.58 **2571**

Elements of Thermionics.—W. E. Danforth. (*Proc. Inst. Radio Engrs*, May 1951, Vol. 39, No. 5, pp. 485–499.) A survey, intended primarily for workers in other fields, of the principal experimental and theoretical developments in thermionics. From a simple basis of statistical mechanics, relations are derived including the Richardson equation, the Schottky field-effect equation, and the Fowler equation for emission from a normal impurity semiconductor.

621.314.632 + 621.315.592† **2572**

The Characteristics and Some Applications of Varistors.—Stansel. (See 2363.)

621.385 **2573**

Reliability in Miniature and Subminiature Tubes.—P. T. Weeks. (*Proc. Inst. Radio Engrs*, May 1951, Vol. 39, No. 5, pp. 499–503.) The meaning of the term 'reliability' as applied to valves is discussed; reliability is found to be a function not only of valve design and quality, but also of the relation between valve ratings and the operating conditions and requirements. Specific features discussed include ruggedness, operating temperature, emission stability and life, and consideration is given to the general effect of reducing valve size.

621.385 **2574**

Valve Development by Telefunken since the Cessation of Hostilities (1945).—H. Rothe. (*Telefunken Ztg*, Sept. 1950, Vol. 23, Nos. 87/88, pp. 93–96.) On account of difficulties due to the condition of the various plants, no considerable development was possible till the second half of 1948. Since then the broadcast-receiver 11-series has been completed in glass-envelope valves and, for u.s.w. f.m. receivers, in metal-envelope valves. A new series of miniature valves ('pico' series) was marketed in mid-1949. Some half-dozen power valves intended for communications, broadcast transmitters and industrial generators are also very briefly described.

621.385 : 621.386.1 **2575**

Radiographic Examination of Electronic Valves.—van Wijlen. (See 2508.)

621.385 : 621.396.619.16 **2576**

Signal Retardation Electron Tubes with Delay Modulation.—É. Labin. (*Onde élect.*, Feb. 1951, Vol. 31, No. 287, pp. 82–89.) An electron beam, modulated in density by the signal to be retarded, is passed between a pair of deflecting plates before injection into a retarding chamber within which delay is effected by means of a magnetic field which causes the beam to follow a helical trajectory. The amount of the delay is dependent on the angle of entry into the chamber and this is controlled by the deflector plates, to which the required 'delay' modulation voltage is applied. In the non-modulated condition delay may be of the order of 100 μs; modulated, the maximum obtainable delay is an inverse function of the modulation frequency; at the limiting frequency, dispersion due to space charge may limit the output current to $<1\,\mu$A. This and other limitations of the method are discussed and illustrations are given of valves constructed to test the validity of the principle.

621.385.012 : 621.317.799† **2577**

Tube Characteristic Tracer using Pulse Techniques.—Wagner. (See 2488.)

621.385.029.6 **2578**

Pulse Technique in High-Power Valve Development.—A. M. Hardie. (*Metrop. Vick. Gaz.*, April 1951, Vol. 23, No. 386, pp. 350–360.) The trend of high-power valve development is discussed with particular reference to the future use of demountable valves on s.w. service. Design problems are enumerated. A recording technique is described for presenting positive-grid characteristics in a form suitable for engineering applications. Either point-by-point or photographic recording may be used; some examples of the latter are presented.

621.385.029.63/.64 **2579**

Amplification of the Traveling Wave Tube.—B. Friedman. (*J. appl. Phys.*, April 1951, Vol. 22, No. 4, pp. 443–447.) A simpler and more exact method is presented for solving the transcendental equation given by Chu & Jackson (3549 of 1948) for wave propagation in the helix of the travelling-wave valve. The propagation is considered as a perturbed form of that in the cold helix. The dependence of amplification factor on geometrical parameters and operating conditions is determined explicitly. The valve will not amplify if the d.c. beam current is too high.

621.385.029.63/.64 **2580**

Effect of Hydrostatic Pressure in an Electron Beam on the Operation of Traveling-Wave Devices.—P. Parzen & L. Goldstein. (*J. appl. Phys.*, April 1951, Vol. 22, No. 4, pp. 398–401.) Small velocity spreads in the electron beam appear to cause a decrease in gain and noise figure of a travelling-wave valve. The physical explanation is that the velocity spread introduces interactions between the electrons, in which the external circuit takes no part, this interaction being of the nature of a hydrostatic pressure.

621.385.029.63/.64 **2581**

Travelling-Wave Tubes with Dispersive Helices.—F. N. H. Robinson. (*Wireless Engr*, April 1951, Vol. 28, No. 331, pp. 110–113.) Oscillation occurs in travelling-wave-valve amplifiers when reflection takes place at mismatches between the ends of the helix and the external circuit. The difficulty of obtaining good matching over the wide frequency band of normal valves has led to the development of a dispersive helix, in which the phase velocity varies rapidly with frequency. This is achieved by making the diameter of the helix very small. Amplification then occurs for only a limited range of frequencies, over which correct termination of the helix is possible. By this means the beam current required to produce a given gain is much reduced. Noise factor is also comparatively low.

621.385.032.213 : 537.533.8 **2582**

Secondary-Emission Cathodes of High Stability.—B. D. Tazulakhov. (*Zh. tekh. Fiz.*, July 1950, Vol. 20, No. 7, pp. 773–787.) The preparation of cathodes possessing a high stability under high temperatures and heavy current loads was investigated experimentally. The requirements which the active and intermediate layers of the cathodes should satisfy are defined, and tables showing the properties of various suitable materials are given. The performance of complex BaO emitters deposited on Ag, Cu, Ni, nichrome, Mo and Ta is discussed in detail and experimental curves showing the secondary emission from these cathodes are plotted. It is claimed that in the production technique proposed, the thickness control of the emissive layer is much simpler than in the usual methods, where it is more of the nature of an art than of a technological process.

621.385.15 : 621.385.831 **2583**

Voltage-Controlled Secondary-Emission Multipliers.—A. J. W. M. van Overbeek. (*Wireless Engr*, April 1951, Vol. 28, No. 331, pp. 114–125.) Secondary-emission valves have, in some cases, a much shorter life than normal valves. This objectionable feature has been overcome by using a coating of Cs_2O on the dynodes and keeping their temperature below 180°C. The constructions of various experimental valves are shown and their characteristics described. A variable-μ valve and a very-high-slope valve with four stages of multiplication are shown. The use of grid dynodes is discussed. Some circuits in which secondary-emission valves offer specific advantages are described, including generators of sinusoidal and nonsinusoidal oscillations and trigger circuits.

621.385.16 : 537.312.5 **2584**

Magnetic Electron Multipliers for Detection of Positive Ions.—L. G. Smith. (*Rev. sci. Instrum.*, March 1951, Vol. 22, No. 3, pp. 166–170.) Two designs of 15-stage multipliers with crossed electric and magnetic fields are described. BeCu dynodes are used, of width $\frac{3}{8}$ in. for fields of 250–460 oersted and $\frac{1}{8}$ in. for fields of 300 1 100 oersted. From their performance it is concluded that a multiplier of this type could be designed to have a rise time between 10^{-11} and 10^{-10} sec.

621.385.2 **2585**

Effect of Variable Mass of the Electron on the Space-Charge Limited Current in a Diode.—S. Visvanathan. (*Canad. J. Phys.*, March 1951, Vol. 29, No. 2, pp.159–162.) "The change in the current-potential distribution due to the relativistic variation of the mass of the electron has been calculated by suitable series expansions in the case of a plane parallel diode and has been shown to be considerable in the case of large power tubes."

621.385.2 **2586**

The Transformation of Heat into Electrical Energy in Thermionic Phenomena.—R. Champeix. (*Le Vide*, Jan. 1951, Vol. 6, No. 31, pp. 936–940.) An experiment is described and theory is adduced showing that, in a thermionic diode, the standing current vanishes when the two electrodes are at the same temperature, independently of the composition of the electrodes. Practical suggestions are made for the design of a diode without standing current and for the determination of the actual source of emission of electrons from oxide-coated cathodes.

621.385.2 : [546.28 + 546.289 **2587**

Crystal Diodes.—R. W. Douglas & E. G. James. (*Proc. Instn elect. Engrs*, Part III, May 1951, Vol. 98, No. 53, pp. 157–168. Discussion, pp. 177–183.) The influence of small amounts of impurities on the electrical properties of semiconductors, and the mechanism of contact rectification are discussed. The processing of Ge and Si for use in crystal diodes is considered in the light of the theory. The design and performance of (*a*) a coaxial-type Si-crystal diode for use as a mixer at frequencies up to about 10 kMc/s, and (*b*) a wire-ended Ge-crystal diode are described. Particular attention is given to the frequency dependence of the rectification efficiency of the Ge diode and to its application as a replacement for the thermionic diode.

621.385.2 : 546.289 **2588**

A New High-Conductance Crystal Diode.—B. J. Rothlein. (*Sylvania Technologist*, April 1951, Vol. 4, No. 2, p. 44.) The experimental Ge diode described is made by applying to the whisker contact an amount of metal paste so small that it does not add appreciably to the capacitance.

621.385.2 : 546.289 : 621.397.62 **2589**

An Analysis of the Germanium Diode as Video Detector.—W. B. Whalley, C. Masucci & N. P. Salz. (*Sylvania Technologist*, April 1951, Vol. 4, No. 2, pp. 25–34.) Methods, including some rapid production-line tests, are discussed for the measurement of those characteristics of Ge diodes which are important in the detection of video signals. The forward and reverse conductances are assumed constant over the range of operation, and both loads with small and loads with large time constants are considered.

621.385.3 + 621.385.5 **2590**

Interelectrode Impedances in Triodes and Pentodes.—E. E. Zepler & S. S. Srivastava. (*Wireless Engr*, May 1951, Vol. 28, No. 332, pp. 146–150.) Bridge measurements of capacitance and conductance were made at 1 Mc/s and 32 Mc/s, and the values are plotted against mutual conductance. Discrepancies between observed capacitance variations and the values indicated by North's theory (1450 of 1936) are discussed, and explanations are advanced for some of the effects.

621.385.3 : 546.289 **2591**

Effect of Auxiliary Current on Transistor Operation.—H. J. Reich, P. M. Schultheiss, J. G. Skalnik, T. Flynn, & J. E. Gibson. (*J. appl. Phys.*, May 1951, Vol. 22, No. 5, pp. 682–683.) Transistor gain characteristics may be improved by the flow of direct current between auxiliary electrodes, one of which is placed as close as possible to the collector. The best improvement in current gain is obtained with relatively large spacing between emitter and collector.

621.385.3 : 546.289 **2592**

Crystal Triodes.—T. R. Scott. (*Proc. Instn elect. Engrs*, Part III, May 1951, Vol. 98, No. 53, pp. 169–177. Discussion, pp. 177–183.) The various forms of crystal triode developed up to date are reviewed. A brief résumé is given of the various materials proposed for the manufacture of these triodes, and the types of control used to modify their characteristics. Testing procedure is discussed. Applications and circuit design are dealt with briefly.

621.385.3.029.64 **2593**

Passive Feedback Admittance of Disc-Seal Triodes.—G. Diemer. (*Philips Res. Rep.*, Dec. 1950, Vol. 5, No. 6, pp. 423–434.) A discussion of the design of disk-seal triodes with a view to using the self-inductance of the grid wires to neutralize the feedback via the anode-cathode capacitance at microwave frequencies.

621.385.3.032.24 **2594**

Aspects in the Design and Manufacture of Planar Grids for Triodes at U.H.F.—W. J. Pohl. (*Electronic Engng*, March 1951, Vol. 23, No. 277, pp. 95–99.) Discussion, with calculations and application to practical manufacturing problems, of the relation between the grid dimensions and its ability to dissipate power. A recently developed method of producing planar ring-frame grids carrying highly tensioned wires, and the method of measurement of residual wire tension, are described. The most suitable material for constructing tensioned grid is tungsten wire with a copper coating of thickness about 15% of the radius of the wire.

621.385.4 : 621.396.61 **2595**

A 30-Watt Transmitter for 430 Mc/s employing the Transmitting Valve QQE 06/40 (AX 9903).—(*Philips tech. Comm., Aust.*, 1951, No. 2, pp. 14–17.) See also 2062 of August (Dorgelo & Zijlstra).

621.385.832.001.4 **2596**

A Note of Cathode-Resistance Stabilization of C.R.T. Gun Current.—H. Moss. (*Electronic Engng*, March 1951, Vol. 23, No. 277, pp. 111–112.) An expression defining the increase in current stability produced by an autobias cathode resistor is deduced, in terms of a stability factor S given by $2S = 7\ E_c/E_d - 5$, where E_c and E_d are the grid cut-off voltage and drive voltage respectively. Graphs are drawn of S against cathode resistance for various cut-off voltages.

621.386.7 **2597**

Centering the Cathode in a Demountable X-Ray Tube.—R. Fouret. (*C. R. Acad. Sci., Paris*, 30th April 1951, Vol. 232, No. 18, pp. 1651–1653.) Centering is performed while the tube is under vacuum, by an electrical method which consists of rendering minimum the capacitance between anode and cathode.

621.396.615.141.2 **2598**

On the Theory of the Anode Block of a Plane Magnetron.—S. D. Gvozdover & V. M. Lopukhin. (*Zh. tekh. Fiz.*, Aug. 1950, Vol. 20, No. 8, pp. 955–960.) A mathematical discussion is presented of magnetrons with anode blocks of the hole-and-slot (Fig. 1) and slot (Fig. 2) types. To simplify the discussion, the space occupied by the anode block is divided into the interaction space, which does not include the anode resonators, and the space which includes these resonators. Equations for the electromagnetic fields in the interaction space are derived and solved, and the natural frequencies of oscillation are determined by an approximate method in which the complex impedances of the interaction space are matched to those of the resonators. The discussion is limited to the case of two-dimensional (plane) magnetrons, and the end effects as well as the effects of couplings are neglected.

621.396.615.141.2 : 537.525.92 **2599**

The Space-Charge in a Magnetron under Static Cut-Off Conditions: Planar or Quasiplanar Magnetron.—G. A. Boutry & J. L. Delcroix. (*C. R. Acad. Sci., Paris*, 9th April 1951, Vol. 232, No. 15, pp. 1413–1415.) The two static cut-off states [1828 of 1950 (Delcroix & Boutry)] are compared: the total space charge is the same for the two cases. The condition of lower energy level of the electron gas corresponds to the Brillouin state. The cut-off surface is the same for the two cases.

621.396.615.141.2 : 537.525.92 **2600**

The Space-Charge in a Magnetron under Static Cut-Off Conditions: Cylindrical Magnetron.—J. L. Delcroix & G. A. Boutry. (*C. R. Acad. Sci., Paris*, 30th April 1951, Vol. 232, No. 18, pp. 1653–1655.) The two static cut-off states are compared (see 2599 above for corresponding consideration of planar magnetrons). The total space charge is different for the two cases; the condition of lower energy level of the electron gas corresponds to the 'bidromic' state. The cut-off surface is not generally in the same position for the two states.

621.396.615.141.2 : 537.525.92 **2601**

Analysis of Synchronous Conditions in the Cylindrical Magnetron Space Charge.—H. W. Welch, Jr, & W. G. Dow. (*J. appl. Phys.*, April 1951, Vol. 22, No. 4, pp. 433–438.) "In the multianode cylindrical magnetron there exist favored phase velocities of the electromagnetic wave around the interaction space between anode and cathode. These velocities are characteristic of the resonant system attached to the anode segments. In the oscillating magnetron the electronic space charge within the interaction space is presumed to maintain synchronism with one of these velocities. Certain of the conditions of synchronism which can be discussed analytically are treated in this paper. The results, although based on restrictive assumptions, can be used in the interpretation of magnetron operation and in predicting regions of efficient behaviour." See also 1282 of May (Welch).

621.396.615.142 **2602**

The Limiting Efficiency of Oscillation Generation by means of Velocity-Modulated Electron Beams in Drift-Space Valves with Fields of Finite Length.—R. Gebauer & H. Kosmahl. (*Z. angew. Phys.*, Dec. 1950, Vol. 2, No. 12, pp. 478–486.) The concept of the ideal efficiency is introduced; this quantity is a measure of the greatest possible amount of h.f. energy which can be extracted from the valve, neglecting the velocity modulation, and is hence also an indication of the quality of the focusing. The optimum length of drift-space for a given modulation depth and control-gap length is calculated. The focusing properties of infinitely short and finite-length fields are compared and found to be equivalent only for vanishingly small modulation. The relation of the practically attainable limiting efficiency to the ideal efficiency is defined; the practical limiting efficiency decreases with increase of modulation depth.

621.385.032.216 **2603**

Die Oxydkathode: 2. Teil — Technik und Physik. [Book Review]—G. Herrmann & S. Wagener. Publishers: J. A. Barth, Leipzig, 2nd edn 1950, 284 pp. (*Fernmeldetech. Z.*, Jan. 1951, Vol. 4, No. 1, p. 46.) A modern and exhaustive exposition of the subject. Vol. 1: 2985 of 1949.

MISCELLANEOUS

621.396 **2604**

Radio Progress during 1950.—(*Proc. Inst. Radio Engrs*, April 1951, Vol. 39, No. 4, pp. 359–396.) A survey based on material compiled by the 1950 Annual Review Committee of the I.R.E., and including 1084 references. The material is grouped under the following headings: antennas and waveguides; audio techniques; electroacoustics; sound recording and reproducing; circuit theory, electron tubes and solid-state devices; electronic computers; facsimile; industrial electronics; measurements; mobile radio; modulation systems; navigation aids; piezoelectric crystals; radio transmitters; receivers; standards on symbols; television system; video techniques; wave propagation.

621.396(083.72) **2605**

Standards on Abbreviations of Radio-Electronic Terms, 1951.—(*Proc. Inst. Radio Engrs*, April 1951, Vol. 39, No. 4, pp. 397–400.) Reprints of this Standard, 51 IRE 21 S1, may be purchased while available from The Institute of Radio Engineers, 1 East 79 Street, New York 21, N.Y., at $0.50 per copy.

621.396 Tesla **2606**

The Life and Work of Nikola Tesla.—A. Damianovitch. (*Bull. Soc. franç. Élect.*, Feb. 1951, Vol. 1, No. 2, pp. 85–99.) Lecture before the Société française des Électriciens, reviewing the pioneer work of Tesla in the field of a.c. and radio engineering.

621.39 **2607**

Electrical Engineers' Handbook—Electric Communication and Electronics. [Book Review]—H. Pender & K. McIlwain. Publishers: Chapman & Hall, London & J. Wiley, New York, 4th edn, 1345 pp., 68s. (*Electrician*, 9th March 1951, Vol. 146, No. 3795, p. 823.) An entirely rewritten edition, with contributions by 78 specialists. F.m. and pulse techniques in communications and radar are included for the first time. A bibliography is appended to each of the 23 sections.