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THE JOURNAL OF RADIO RESEARCH & PROGRESS

JANUARY 1953

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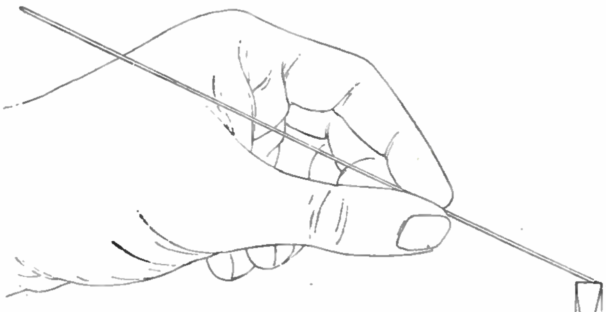

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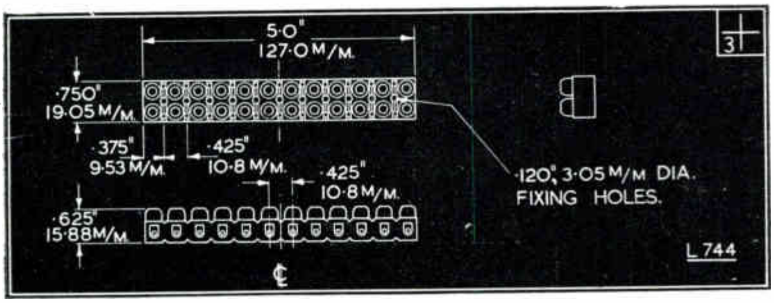

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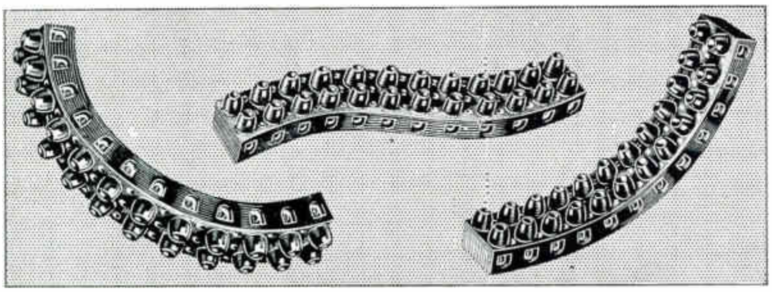

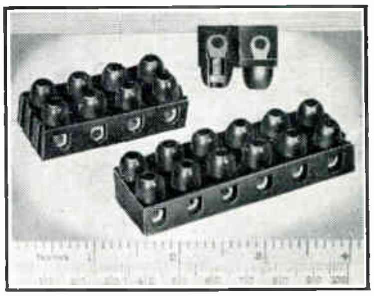

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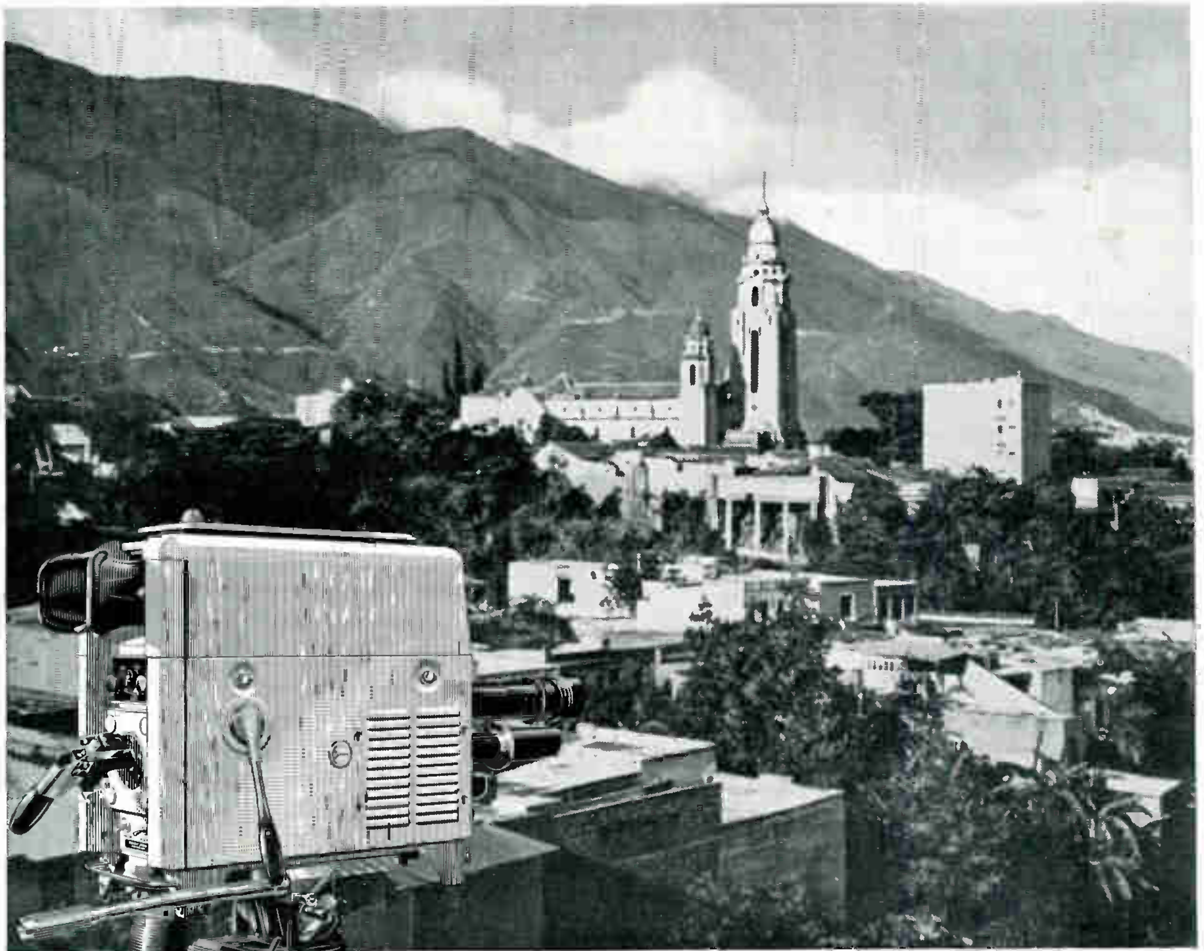

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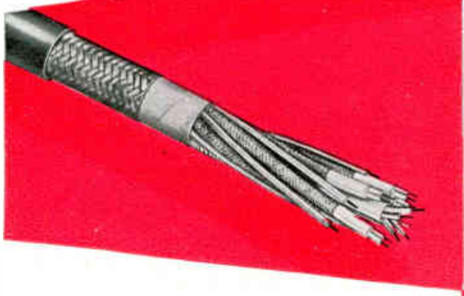

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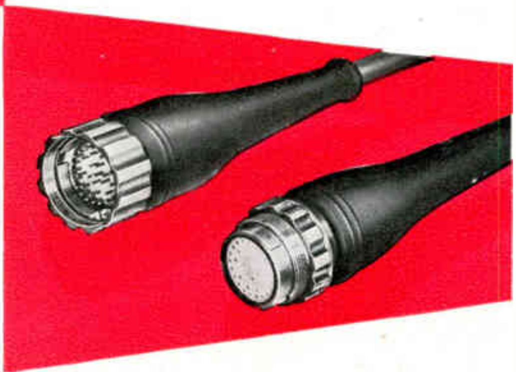

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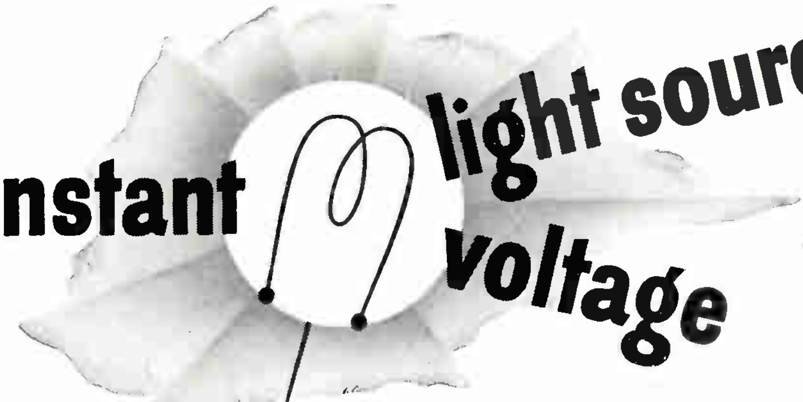

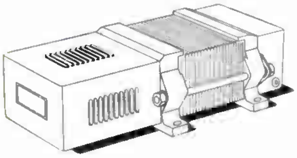

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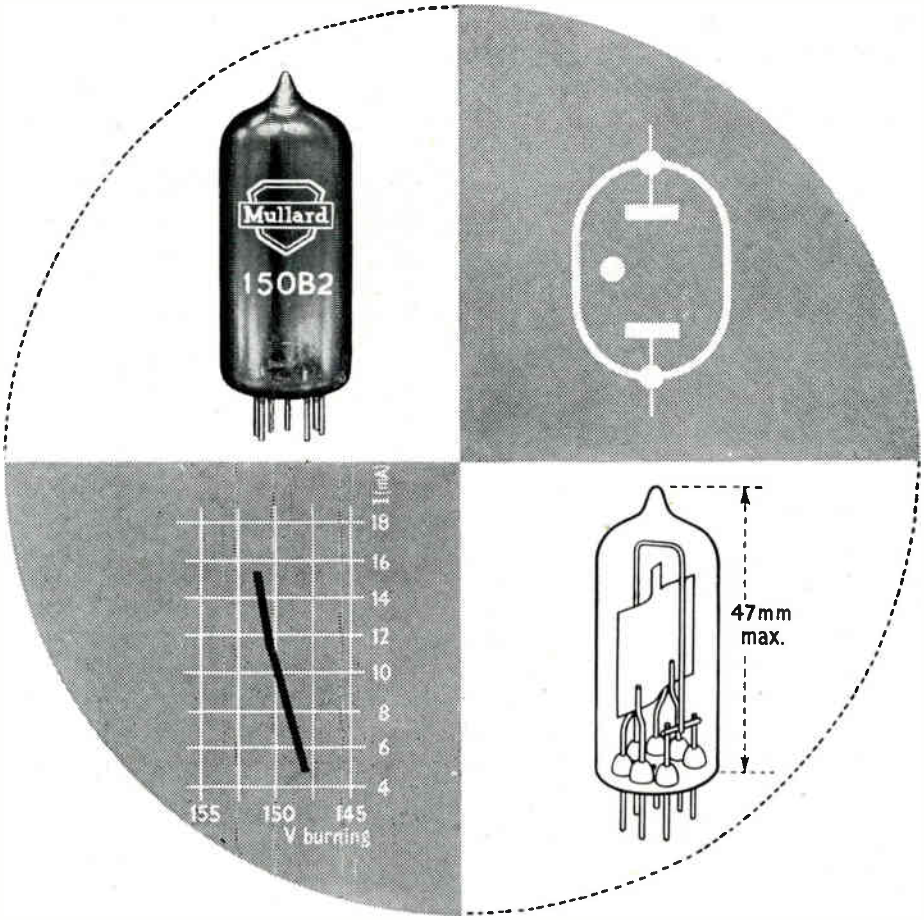

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WIRELESS ENGINEER

Vol. 30 JANUARY 1953 No. 1

Circuit Diagrams

IN 1948, after very lengthy discussions, the British Standards Institution issued B.S. 530: 1948 "Graphical Symbols for Telecommunications" and, in June of the same year, an article entitled "Improving Circuit Diagrams" by L. H. Bainbridge-Bell, who had taken a very active part in the discussions, appeared in *Electronic Engineering*. In a recent number of the *Proceedings of the Institution of Radio Engineers, Australia* (Sept. 1952, p. 345), there is a very interesting article entitled "The Utility Factor in Circuit Diagrams" by C. E. Williams, an Airways Engineer of the Department of Civil Aviation, who has had experience in teaching the use of relatively complex equipment. He says that great difficulty was experienced in working from commercially-produced diagrams and that, in many cases, it was found necessary to redraw completely whole circuits before they could be made to give their story to the students. He emphasizes that the draughtsman must know how the circuit works; the emphasis in his training should be placed on electronic engineering rather than upon mechanical drawing. If this layout man is going to think about the placement of each component, this will involve time, but this time should be regarded as an investment which ultimately will be repaid in full.

As an example of the subtle ways in which the desired impression can be created in the mind of the reader, the author considers Figs. 1 (a) and (b). Which should be used depends on the operation performed. Fig. 1 (a) gives the impression that the signal passes through the capacitor to the grid, and that the resistor is merely a grid leak, whereas Fig. 1 (b) gives the impression that the signal is applied across the capacitor and the resistor in series, and the voltage across the latter is applied to the grid. The man making the diagram should know enough about the action of the set to enable him to make the correct choice in this and other similar cases.

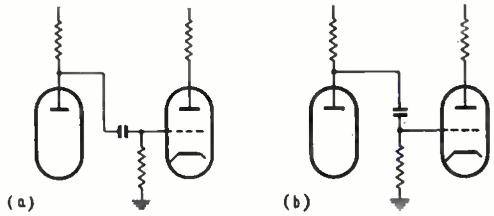
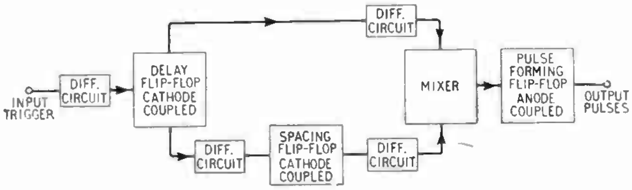


Fig. 1.

It is, however, in some of the less familiar fields that the advantages of standardization are most apparent. In the field of pulse techniques, for example, the most complex equipments can be broken down into not more than half a dozen basic types of sub-circuit. Once the operation of these basic sub-circuits is known and understood, tracing the operation of the complex circuit resolves itself into their recognition. The author takes as an example a pulse generator which is to produce two identical pulses of variable width, with variable spacing, the first pulse to be delayed by a variable time with respect to an external initiating trigger pulse. Fig. 2 shows a suitable block diagram, and the author says that even such diagrams are worthy of more thought than is usually given. Where the signal splits into two paths, these paths are shown in Fig. 2 as being of equal importance, and not one as a mere by-pass. Where the two signals are mixed, the two paths are shown physically as coming together and then continuing along a common path. Fig. 3, says the author, shows the circuit diagram as it would probably emerge from a typical drawing office, while Fig. 4 shows an alternative drawn around the block diagram. No effort has been made to put valve envelopes or components in line, but every effort has been made to standardize the sub-circuit layout.



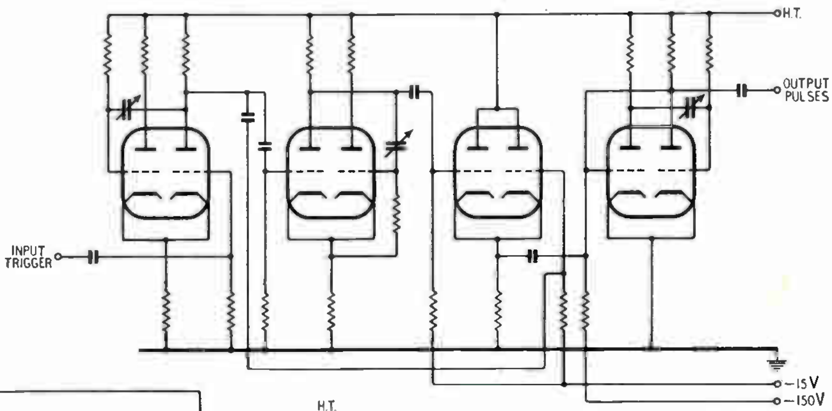
cut-off valve is on the left, so that its anode gives a negative pulse, and the anode of the right-hand valve a positive pulse. The author's rule for improved circuit diagrams is: adopt a standard layout for commonly used sub-circuits, and stick to this standard, even if deviation in some cases would lead to a neater diagram from an artistic point of view.

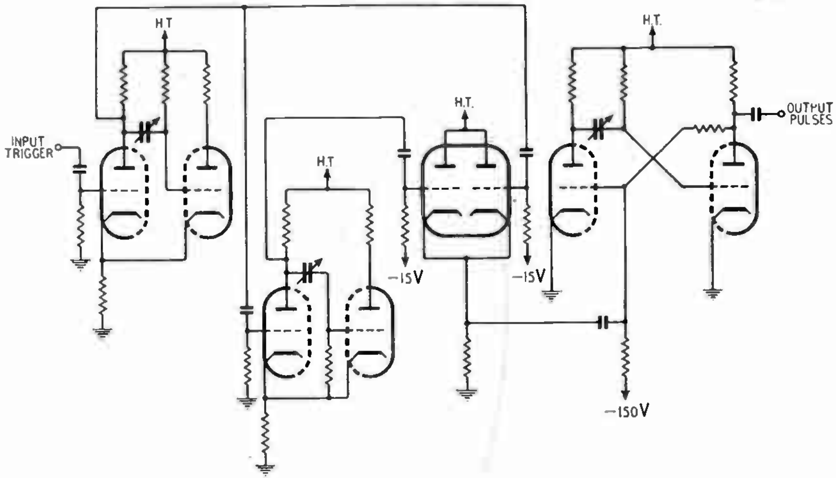
Another point stressed by the author is the avoidance of unnecessary lines in the diagram by omitting not only valve-heater wiring, but also high tension and bias lines, which, when there are several H.T. supplies and several bias voltages, can make the diagram chaotic. It will be seen

Fig. 2 (above).

Fig. 3 (right).

Fig. 4 (below).





that they are entirely omitted from Fig. 4, and the author's final rule is: include in the circuit diagram only those lines which carry signals.

The author speaks highly of the 1948 publication of the British Standards Institution and says that if all concerned with the preparation of electronics circuits were to read and thoroughly digest its contents the life of the users of the diagrams would be much happier. He suggests, however, that the Institution should consider forming an Electronic Circuits Standardization Committee which could ultimately publish a handbook of recommended practices. We suspect that by Institution he means the Australian I.R.E.

On account of its neatness most people prefer Fig. 3 until they try to explain how the circuit works; they then switch over to Fig. 4. The main points to notice in Fig. 4 are as follows: the circuits of the first two valves are practically identical; the third valve is a mixer, hence its symmetry; the fourth valve is an anode-coupled flip-flop, as indicated at once by the sloping cross-couplings; in all three flip-flops the normally

G. W. O. H.

FORMULAE FOR LADDER FILTERS

By H. J. Orchard, M.Sc.

SUMMARY.—Four related sets of explicit formulae for the elements of a basic low-pass structure are collected together in a common notation. Three of the sets have already been published, the fourth is new. They refer to operating conditions and loss responses of the type associated with the insertion-parameter theory introduced by Norton and Darlington.

1. Introduction

OVER the past 15 years three papers[1,2,3] have been published giving explicit formulae for the element values in a certain LC-ladder filter circuit. The circuit concerned has the configuration of a conventional 'constant-k' image-parameter low-pass filter, but the element values as fixed by these three sets of formulae refer to operating conditions and loss characteristics normally associated with the insertion-parameter theory due to Norton[1] and Darlington.[4] It is because the insertion-parameter theory leads to a more efficient network than does the image-parameter theory that these formulae are important. Without their use, the process of finding the element values in insertion-parameter filters is rather laborious, even with the best-organized computing schemes, and thus direct formulae for the elements are of considerable value.

The formulae so far given form, logically, three of a group of four. It is the purpose of this paper to give the remaining set of formulae and to exhibit all four sets, side-by-side, with a common notation, when their simplicity and similarity become immediately apparent. Attempts have been made to extend the formulae to more general filter circuits from which the existing formulae could be derived by a simple limiting process but, as yet, the results are rather limited; the problem will be described briefly, and it is hoped that the presentation will stimulate further research in this field.

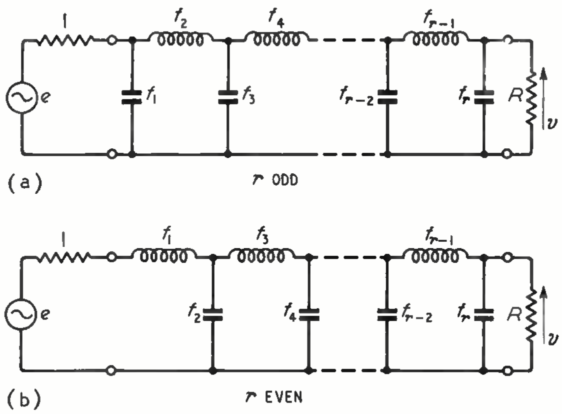


Fig. 1. *Form of network for 'r' odd (a) and even (b).*

MS accepted by the Editor, May 1952.

2. The Four Sets of Formulae

The network concerned is shown in Fig. 1; it is driven from a generator of unit internal resistance and e.m.f. e, and feeds a load of resistance R across which appears a voltage v. For both even and odd numbers of elements the circuit is arranged to terminate at the load end in a shunt capacitor. Following Belevitch[3] the element values are denoted by $f_1, f_2, \ldots f_r$.

For any prescribed shape of loss response from e to v which the network can provide, it is possible to set the value of R to one of two principal values; we can have

(i) the open-circuit case (i.e., with $R = \infty$) which may be used when the output voltage alone is of interest (e.g., when driving a valve), or

(ii) the terminated case, with R so chosen that the generator delivers the maximum possible power into R at the frequency (or frequencies) of minimum loss. This would be used when the output power is the quantity of interest.

Other values of R are, of course, possible, but they rarely have any practical significance.

Of the many filter-like loss characteristics which can be associated with the network of Fig. 1 there are two which, in recent years, have received much attention mainly because of their unique properties and mathematical simplicity; they are

(i) the Tchebyshev, or equal ripple, response shown in Fig. 2(a) and described by the pair of equations

$$A \text{ db} = 20 \log |e/v| = 10 \log (1 + t \cos^2 ru) + \text{const.} \quad \text{(1a)}$$

$$\omega = \cos u \quad \ldots \quad \ldots \quad \ldots \quad \text{(1b)}$$

and

(ii) the maximally-flat response shown in Fig. 2(b) and described by the equation

$$A \text{ db} = 20 \log |e/v| = 10 \log (1 + \omega^{2r}) + \text{const.} \quad \ldots \quad (2)$$

The values of the added constants in Equs. (1a) and (2) must be chosen to suit the value of R set for the network; they are easily determined but need not concern us here.

We are thus led to four distinct sets of element values for our network by associating either of these two responses with either of the two principal operating conditions. Explicit formulae for three of the four sets have already been published; these three formulae together with the new one recently discovered are most simply expressed by the recurrence relations given in Table 1.

For the remaining cases, the use of the recurrence relation prevents the formulae becoming unwieldy and, in numerical work, is preferable to any other form because the accuracy of the whole computation can be checked by referring to the special check formulae involving f_r and f_1.

TABLE 1

	Tchebyshev Behaviour	Maximally-Flat Behaviour
Open-circuit case	$f_n f_{n-1} = a_n a_{n-1}/d_{n-1}$ $f_1 = a_1/\gamma$ *Check* $f_r = r\gamma\sqrt{1+t}\, f_1/\sqrt{1+\gamma^2}$ for r odd $= r\gamma f_1/\sqrt{(1+t)(1+\gamma^2)}$ for r even	$f_n f_{n-1} = a_n a_{n-1}/c_{n-1}$ $f_1 = a_1$ *Check* $f_r = r f_1$
Terminated case	$f_n f_{n-1} = 4a_n a_{n-1}/b_{n-1}$ $f_1 = 2a_1/\gamma$ $R = 1$ for r odd $= \coth^2 rb/2$ for r even *Check* $f_r = R f_1$	$f_n f_{n-1} = 4a_n a_{n-1}$ $f_1 = 2a_1$ $R = 1$ for all r

$\gamma = \sinh b$ $\qquad$ $20 \log\coth rb = A_m$ db $= 10 \log (1 + t)$

$a_n = \sin (2n - 1)\pi/2r$ $\qquad$ $b_n = \gamma^2 + \sin^2 2n\pi/2r$

$c_n = \cos^2 n\pi/2r$ $\qquad$ $d_n = (\gamma^2 + \sin^2 n\pi/2r) \cos^2 n\pi/2r$

The element values obtained from the table will be for a network normalized both with regard to generator resistance and angular frequency at the edge of the pass range; to obtain the values for a network to be driven from a generator of resistance R_0 and to have a pass range up to an angular frequency ω_0, all resistances must be multiplied by R_0, all inductances by R_0/ω_0 and all capacitances by $1/R_0\omega_0$.

In addition, all the normal network and frequency transformations, for example, changing to the dual circuit and the use of the low-pass to band-pass transformation, can, of course, be applied. In particular, it may be mentioned that the dual of the open-circuit case is the short-circuit case where $R = 0$; in this event it is more customary to consider the network in reverse order; i.e., as driven from a zero impedance source and feeding into a resistive load.

In order to preserve the symmetry of the formulae, the elements in the maximally-flat terminated case have been quoted as a recurrence relation along with the remainder, although a moment's reflection will show that, in fact, they are given simply by

$$f_n = 2a_n \quad \ldots \quad (3)$$

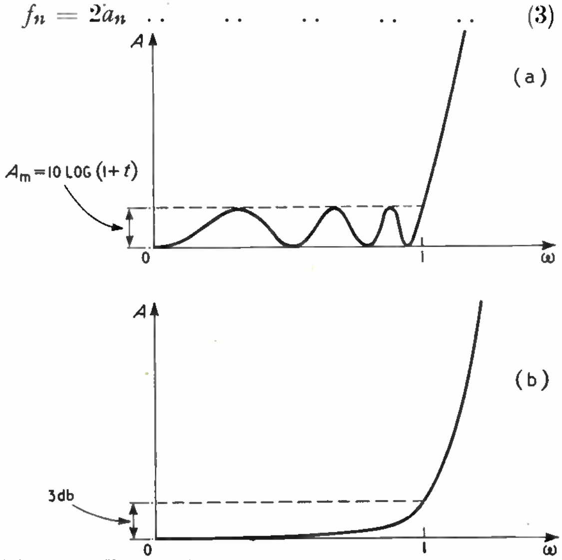


Fig. 2. *Form of characteristic, Tchebyshev* (*a*) *and maximally-flat* (*b*).

3. Discussion of Formulae

The formulae for the maximally-flat open-circuit case were given by Norton[1] in 1937 as a part of a paper dealing with junction filters; this paper is notable in that it was the forerunner of the insertion-parameter theory later expounded by Darlington.[4] Those for the maximally-flat terminated case appeared in 1951 in a paper by Bosse[2] who gave a very thorough treatment of this particular network; and finally the terminated Tchebyshev case was solved by Belevitch[3] in 1952. It was this last paper, with its comment that the formulae for the open-circuit Tchebyshev case had not been found, that stimulated the writer to look into the matter.

By collecting together all the known formulae in a common notation and from the knowledge that the maximally-flat case can always be derived by a limiting process from the Tchebyshev case,*

* This possibility was first pointed out to the writer by Dr. W. Saraga. It consists geometrically of letting the ellipse in the ω-plane, on which the roots of the power ratio lie, degenerate suitably into a circle; or, analytically, of renormalizing the frequency scale to $\omega' = \omega/\gamma$ and letting $\gamma \rightarrow \infty$.

a little intuition was sufficient to reveal the missing formulae for the open-circuit Tchebyshev case. No complete and formal proof for the formulae has actually been obtained although extensive numerical checks have been carried out along with a general examination of the problem and there appears to be no doubt as to their correctness.

A parent problem of much greater interest is to find the corresponding formulae for the terminated and symmetrical network having the configuration of an integral number of whole m-derived image-parameter low-pass filter sections and with the response given by

$$A \text{ db} = 10 \log \left[1 + t \, \mathrm{sn}^2(ruK_1/K;\, k_1)\right] \quad \text{(4a)}$$

$$\omega = \mathrm{sn}(u;\, k) \quad \text{(4b)}$$

This is a direct generalization of Equ. (1) in which the circular functions are replaced by Jacobian sn elliptic functions; it leads to a Tchebyshev behaviour in both pass-band and stop-band. Full details of this network are given by Darlington.[4] If these formulae could be found they would represent a considerable contribution to our filter-design technique, and in view of the simplicity of the existing formulae it is not unreasonable to hope that they may be.

4. Acknowledgment

Acknowledgment is made to the Engineer-in-Chief of the G.P.O. for permission to make use of the information contained in this paper.

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ANODE-FOLLOWER DERIVATIVES

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SUMMARY.—Low output-impedance stages characterized by anode output and feedback of the entire output voltage are derived from the basic 'anode-follower' by substitution of a valve impedance for the shunt resistor of the feedback path, or of a comparator stage for the entire input-feedback potential divider, and by replacement of the output valve by a series-connected push-pull pair

These developments suggest the possibility of obtaining an anode-follower analogue of each cathode-follower derivative, thereby increasing the number of circuit variants available for practical use.

1. Introduction

IN the design of electronic-circuit systems for use with signals or waveforms which occupy a wide frequency band it is frequently necessary to include stages having a very low output impedance together with high input impedance for buffering a source from a relatively low-impedance load, and similar purposes. Usually the cathode-follower or one of its derivatives is suitable but occasionally there is a need for low output impedance at a high mean-voltage level, or for a low output-impedance stage capable of providing an inverted copy of an existing signal. The former situation arises in anode modulation of an oscillator or carrier amplifier and in cathode modulation of a projection-type television picture tube; the latter occurs in audio and c.r. tube deflection amplifiers which are required to provide push-pull output. In such circumstances the anode-follower may be preferred. Although of early origin[1] this circuit was not widely employed until the war, when it came into general use as part of the self-balancing paraphase-type push-pull amplifier in radar display systems and cathode-ray test equipment.[2] More recently further consideration has been given to its original application to phase-splitting in push-pull audio amplifiers as an alternative to more conventional circuits.[3] The present paper is concerned chiefly with its low output-impedance property and with low output-impedance systems which may be derived from the basic anode-follower in an analogous manner to developments of the cathode-follower.

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2. Basic Anode-Follower Circuit

The anode-follower consists essentially of a normal (i.e., grid input–anode output) amplifier having sufficient anode–grid negative voltage feedback to decrease the stage gain to approximately unity, under which condition its output impedance is reduced to the order of $1/g_m$, as in the cathode-follower. Apart from its inversion property and high direct-voltage output level, the anode-follower is superior to the cathode-follower in certain applications in that its gain is not inevitably less than unity but may be set exactly to this value by manual adjustment of an element in the feedback path.

Since the anode-follower is a shunt type of negative-feedback amplifier its gain and output admittance[4] may be expressed in the forms:—

$$A = \frac{A(0)}{1 - A(0).\beta}$$

and

$$Y_o = \{1 - A(0).\beta\}.Y_o(0)$$

respectively, where

$A = A(\beta)$ = Stage gain of a feedback amplifier (considered as a function of β),

$A(0)$ = Stage gain of a feedback amplifier in the special case $\beta = 0$,

β = Feedback factor,

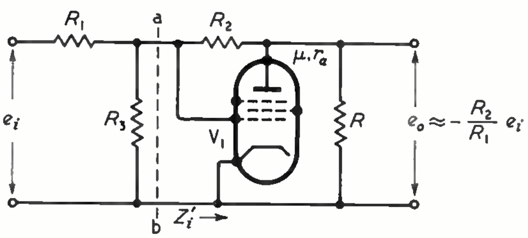
$Y_o = Y_o(\beta)$ = Output admittance of a feedback amplifier (considered as a function of β),

$Y_o(0)$ = Output admittance of a feedback amplifier in the special case $\beta = 0$.

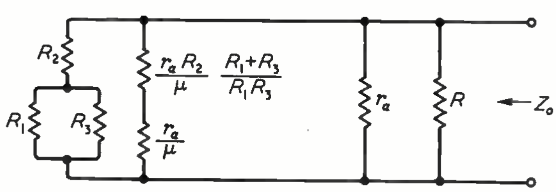
Therefore the gain-admittance product under feedback

$$A(\beta).Y_o(\beta) = A(0).Y_o(0)$$

which is independent of β. Under varying β, A and Y_o vary inversely so that gain may be exchanged for output admittance, as required. Thus when the anode-follower is used with a matched load its gain is not necessarily reduced to ½ (or less) as in the cathode-follower but, with a valve of sufficiently high g_m, may equal or exceed unity. A further advantage of the anode-follower over the cathode-follower is that, due to the ease with which β may be varied, the load matching adjustment is more conveniently made. The input-feedback potential-divider must, however, be designed as a frequency-compensated attenuator to allow wideband operation; moreover, the input impedance will generally be considerably less than that of a cathode-follower. Quadrature phase-shift over the feedback path must be avoided at extreme frequencies, otherwise the output impedance will have a substantial reactive component since the basic network is identical in configuration with that of the conventional electronic reactance circuit.



(a)



(b)

Fig. 1. *Basic circuit of anode-follower (a) and equivalent network of its output impedance (b).*

The circuit of the anode-follower is shown in skeleton form* in Fig. 1(a). Straightforward analysis shows that the input impedance of the right-hand portion of the network bisected at a, b, is:

$$Z'_i = \frac{r_a R + RR_2 + R_2 r_a}{r_a + (\mu + 1) R} \approx \frac{r_a R_2}{\mu R}; R \ll r_a, R_2; 1 \ll \mu$$

which tends to zero with increasing μ. Thus, provided μ is large, the input impedance of the entire stage reduces to

$$Z_i \approx R_1$$

Under this condition the gain is:

$$A = \frac{(r_a - \mu R_2)R}{\{r_a + (\mu + 1)R\}R_1}$$

$$\approx \frac{-\mu R_2}{(\mu + 1)R_1}; r_a \ll \mu R_2, r_a \ll (\mu + 1)R$$

$$\rightarrow -\frac{R_2}{R_1} \text{ as } \mu \rightarrow \infty$$

In the special case $R_1 = R_2$, $A = -1$ (anode-follower).

Using results developed for electronic reactance networks the exact (assuming linear operation) equivalent network of the output impedance may be obtained in the form shown in Fig. 1(b), from which the output impedance for the case $R_1 = R_2 = R_3 = P$ is:

$$Z_o = \frac{3 r_a P R}{(2R + 3P) r_a + (\mu + 3) PR}$$

$$\approx 3 \frac{r_a}{\mu}, 2R \ll 3P, 1 \ll \mu,$$

and tends to zero with increasing μ. The restriction on R_3 is unimportant because, with adequate μ, it occurs across a low-impedance node-pair and is necessary only to preserve grid-cathode continuity when the grid is capacitively coupled from the junction R_1/R_2.

Thus, in summary, the basic anode-follower

* For the sake of clarity all the circuit developments given have been reduced to skeleton form by showing direct couplings and by omitting grid bias, screen feed, decoupling and other secondary arrangements.

requires $\mu \gg 1$, $R_1 = R_2$, and with $R_1 = R_2 = R_3$ is characterized by $Z_i = R_1$, $Z_o = 3r_a/\mu$, $A = -1$.

3. Increase of Loop Gain

Where, as in certain computer applications, operator units are required whose gain is exactly -1 it is usual to define the upper limit of gain precisely at -1 by the use of a pair of equal, close-tolerance, high-stability resistors for R_1, R_2 and make the actual gain approach this limit as closely as desired by use of sufficient loop gain.[5] Three resistance-loaded high-μ pentodes connected in cascade are generally employed in place of V_1 of Fig. 1(a), as shown in Fig. 2(a), in order to obtain a total loop gain in excess of 10^6. This type of circuit is also used as an alternative to the single-valve anode- or cathode-follower when an exceptionally low output impedance is required, in which respect it is superior, at least in regard to loop gain, to the "super-" anode- and cathode-followers described in Section 5. A more direct comparison is afforded if V_1 and V_2 of Fig. 2(b) are replaced (at the expense of loop gain) by a cathode-coupled pair, as shown in Fig. 2(b), in which case the latter circuits,

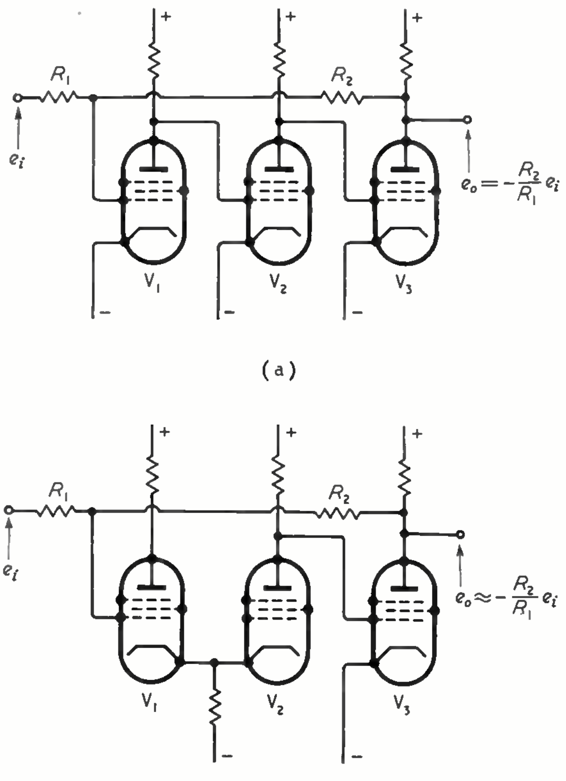


Fig. 2. *Precision anode-follower (a) and a more elaborate version (b) with a cathode-coupled pair substituted for V_1, V_2 of (a).*

which are typified by Fig. 5, would be preferred for their higher input impedance.

4. Electronic Feedback Impedance

An anode-follower will usually be driven by a voltage amplifier and a slight simplification occurs if e_i, R_1 of Fig. 1(a) are identified with the constant-voltage generator equivalent of the driver valve. The resultant circuit (see Fig. 3) has two

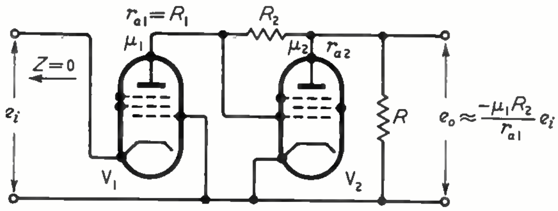


Fig. 3. *Replacement of R_1 of Fig. 1(a) by r_a of driver valve.*

interesting properties. With cathode drive to V_1 every node of the circuit is at a low impedance-level and the entire stage forms an excellent basis for a wideband amplifier, particularly if required to operate between cable impedances, as when a video pre-amplifier is required in television work. Moreover, since with μ_2 sufficiently large, the gain becomes $\mu_1 R_2/r_{a1}$, rather than $\mu_1 R_2/(r_{a1} + R_2)$ as would be obtained from a normal amplifier-follower cascade (with unit gain over the follower), the entire stage not only provides a low output impedance but also allows a larger fraction of μ of the driver valve to be realized at a given bandwidth. This result is useful because earthed-grid operation of V_1, and the low output-impedance obtaining at its anode by virtue of feedback over V_2, both allow the use of a triode for V_1. Grid and cathode may, of course, be transposed to obtain high input-impedance or avoid signal inversion: a double-triode of the common-cathode type may then be used, the low impedance allowing grid drive up to quite high frequencies, since anode-grid capacitance feedback (Miller effect) tends to zero with increase of μ_2.

5. Use of a Comparator Stage

The pair of resistors comprising the input-feedback attenuator may be regarded as a voltage comparator in that they provide at their junction a measure of the difference between the input and output voltages. With $R_1 = R_2$ the action of the amplifier is to drive the output towards equality with the input voltage by making their difference tend to zero. The essential requirement is comparison of e_i with e_o in order to obtain their difference $e_i - e_o$ and may be achieved more satisfactorily by the use of a comparator stage. Moreover, use of an amplifier valve in the comparison process generally allows loop gain to be increased, with the advantages already mentioned in Section 3.

In the simplest case e_i is applied to the grid of the comparator valve and e_o to its cathode, as shown in Fig. 4 (V_1). The difference voltage appears, after amplification, at the anode of the comparator and is coupled to the grid of the output valve. This circuit is superior in certain applications to the normal anode-follower since V_1 is a cathode-follower from the point of view of the input and the input impedance is therefore high.

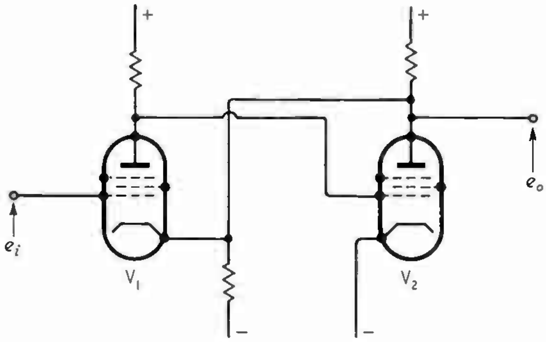


Fig. 4. *Substitution of valve input-output voltage comparator for R_1, R_2 of Fig. 1(a).*

A more flexible and convenient arrangement is obtained by the use of a cathode-coupled pair comparator[6] (see Fig. 5). Insertion of the cathode-follower between the output valve (V_3) anode and the comparator valve (V_1) cathode allows a higher loop gain to be realized since the impedance into which the output valve works is raised and does not decrease with increase of comparator μ. This circuit may be regarded as the anode-follower analogue[7] of the well-known "super" cathode-follower[8] or "cathode repeater" circuit.[9] The two circuits differ only by two transpositions; viz., change of output connection (and, therefore, of feedback connection) from anode to cathode of V_3, and shift of drive for V_3 from the second anode to the first anode of the cathode-coupled pair, thereby converting the basic circuit to anode-follower form while preserving negative-feedback loop gain. In linear operation they differ only to the extent that the gain of the cathode-coupled pair from one grid to the anode of the same half of the pair exceeds that to the opposite anode.[6]

With the notation of Fig. 5, and identical valves in the comparator stage (so that $\mu_1 = \mu_2 = \mu$ and $r_{a_1} = r_{a_2} = r_a$), the gain in the absence of feedback is:

$$A(0) = \frac{\mu\mu_3 R_1 R_3 \{r_a + (\mu + 1)R_2\}}{\{(r_a + R_1)R + (\mu + 1)(2r_a + R_1)R_2\}\,(r_{a3} + R_3)}$$

and the loop gain with feedback is:

$$A(0)\beta = \frac{-\mu(\mu + 1)\,\mu_3 R_1 R_2 R_3}{\{(r_a + R_1)\,R + (\mu + 1)(2r_a + R_1)R_2\}\,(r_{a3} + R_3)}$$

whence

$$-\beta^{-1} = \frac{r_a + (\mu + 1)\,R_2}{(\mu + 1)\,R_2} = 1 + \frac{r_a}{(\mu + 1)\,R_2}$$

which tends to unity with increase of comparator μ. Thus, for $A(0)$ large but μ finite, the output may slightly exceed the input because of feedback attenuation over the auxiliary cathode-follower V_2.

Ignoring R_3 and any external parallel load, the output impedance is given by

$$Z_o = r_{a3}/(1 - A(0).\beta)$$

With $r_a \ll (\mu + 1)R_2$ and $R_3 \ll r_{a3}$, the loop gain $A(0).\beta$ reduces to:

$$A(0).\beta \approx \frac{-\mu\mu_3 R_1 R_3}{(2r_a + R_1)r_{a3}}$$

giving:

$$Z_o \approx \frac{(2r_a + R_1)r_{a3}^2}{(2r_a + R_1)\,r_{a3} + \mu\mu_3\,R_1 R_3}$$

which simplifies under the condition $(2r_a + R_1)\,r_{a3} \ll \mu\mu_3 R_1 R_3$ to:

$$Z_o \approx \frac{(2r_a + R_1)r_{a3}^2}{\mu\mu_3 R_1 R_3}$$

As an example, with $r_a = 100$ kΩ, $R_1 = 100$ kΩ, $r_{a3} = 10$ kΩ, $R_3 = 10$ kΩ, $\mu = 100$, $\mu_3 = 20$, the output impedance would be approximately 15 Ω; i.e., 1/100 of the value which V_3 alone would give in the basic circuit.

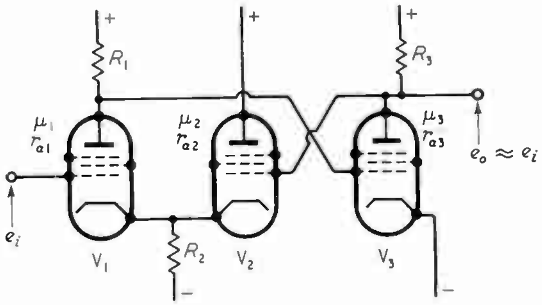


Fig. 5. *Use of cathode-coupled pair comparator ('super' anode-follower).*

6. Adaptation to Push-pull Output

Low output-impedance stages are generally required to handle large-amplitude signals and cannot then be treated accurately as linear circuits. In radar and television applications their response to fast large amplitude changes (transient response) is of particular importance. It is well known that, with appreciable load capacitance, the cathode-follower is capable of providing faster positive than negative output changes. For the same reason the anode-follower cannot operate equally quickly in either direction but, due to its different connection, is faster on negative output-changes.

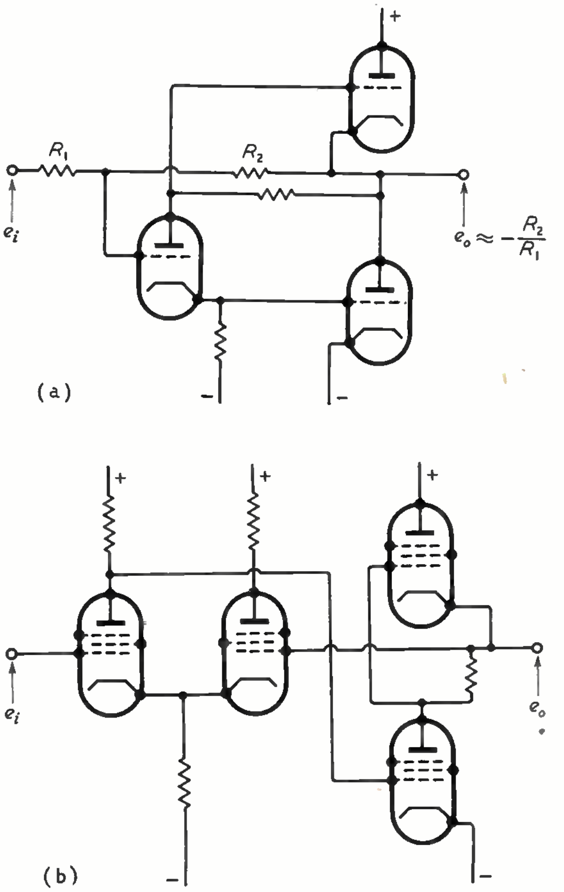


Fig. 6. *Anode-follower connection of single-ended output push-pull pair driven by conventional phase-splitter (a) and self-phase-splitting single-ended output push-pull pair with cathode-coupled input-output comparator (b).*

One remedy is to substitute a push-pull pair for the single output valve and arrange that one passes increasing current to the load on positive-going input changes while the other conducts more heavily on negative changes of input. For the usual requirement of single-ended output this method leads to the basic arrangement shown in Fig. 6(a), in which the upper output valve drives the load from its cathode and the lower valve from its anode. As a result, the stage as a whole is potentially equally fast in both directions. Push-pull drive to the two grids may be provided by a separate phase-splitter as shown in Fig. 6(a). Alternatively, the output pair may be arranged for single-ended drive as in Fig. 6(b) where the upper valve is driven by a voltage obtained by inserting an auxiliary load resistor in the anode lead of the lower output valve. The latter development may be regarded as the anode-follower analogue of the so-called "series-push-pull" or "stacked" cathode-follower. In both circuits the upper valve drives the load as a cathode-follower, while the lower output valve operates as an anode-follower. The difference lies in the choice of driving grid and consequent change of intervalve coupling.

7. Conclusion

A study of the circuits described, together with their variants, suggests that it is possible for each cathode-follower derivative to develop an anode-follower analogue. Although the performance of corresponding circuits is basically similar there are circumstances in which the anode-follower form has slight practical advantages. In the more advanced circuits both anode- and cathode-follower stages may with advantage be used together to provide the required performance in the most convenient manner.

Acknowledgments

The author is indebted to Drs. J. E. Best and E. L. C. White for helpful discussion of the developments described and to E.M.I. Research Laboratories, Ltd., for permission to publish the paper.

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RC CATHODE-FOLLOWER FEEDBACK CIRCUITS

By S. C. Dunn, M.Sc., A.M.I.E.E.

SUMMARY.—RC circuits which can be associated with a cathode-follower are considered. A distinctive feature of the circuits described is a voltage gain over a band of frequencies.

Introduction

THE advantages of utilizing resistors and capacitors as circuit elements are considerable, since they are readily available in numerous sizes and ratings. Whenever a network is required to give critical frequency discrimination the performance of the RC components must be aided and improved by the application of feedback techniques. In this paper some of the circuits which can be used with a cathode-follower are considered.

Circuit Features

The distinctive feature of these types of circuit is that, although dissipative elements are used, there is voltage gain over a band of frequencies. This is unusual, since most RL and RC networks, used as simple filters, possess no real pass-band and always exhibit some attenuation.

The method described here is to employ circuits which are known to possess the null property in transmission (i.e., at one frequency the transfer impedance is infinite) and grade the values of these components to give a more selective transmission. The normal relation between the grading ratios of the capacitors and resistors is then modified to produce a minimum in transmission instead of a null, and at the same time a phase reversal.

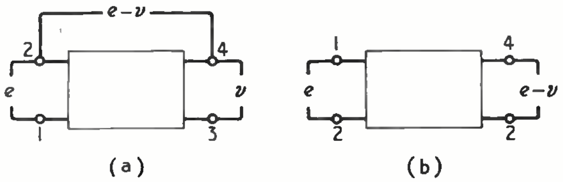


Fig. 1. *Rearrangement of 4-terminal network.*

Fig. 1(a) represents a quadripole which at one frequency gives an output shifted in phase by 180° with respect to the input. It is not necessary that terminals 1 and 3 be in direct connection as is usually the case in the common '3-terminal quadripole'. A voltage e applied at terminals 1, 2 produces an output v at terminals 3, 4. If we now choose terminal 2 as common to input and output, Fig. 1(b), for the same input e the output is $e - v$. The circuit gives, therefore, an output in phase with the input and greater in amplitude. We shall refer to circuits like those in Fig. 1(a) as prototype, and like those in Fig. 1(b) as derived circuits.

MS accepted by the Editor, March 1952.

A number of prototype circuits are shown in Fig. 2 and the corresponding derived circuits in Fig. 3. An interesting feature of these latter networks is that, with the exception of (b) they have a d.c. path between the output terminals and no such path between the input terminals. The prototypes consist of two groups. In the first group types (a) and (b) are the only members and have a monotonic response function with a limiting attenuation rate of 18 db per octave. Types (c) to (h) inclusive, form the other group and are usually null networks. By modifying the relationship between the elements, finite attenuation is substituted with a phase shift of 180°. In both groups the element values are graded to improve the attenuation/phase characteristics.

The properties of two-mesh circuits similar to Fig. 3(a) and (b) but with a phase shift of only 90° have been described by Longmire.[1] The common origin and salient properties of the circuits in the second group have been examined for the null property, and particularly for the symmetrical form, by Harris.[2]

Grading of Circuit Elements

It is often difficult to know how to taper the elements of a network to give a sharper response. The general objective is that successive meshes should not load those on the side of the generator. This is easy to apply in Fig. 2(a) and 2(c) but Fig. 2(e) proves to be more difficult. This problem can, however, be finally resolved by giving the elements general values and rearranging the equation for the null condition to expose unsuspected symmetry. This bridged-ladder circuit is redrawn in Fig. 4, the elements having the values shown. The null condition equation can be found by various ways and is given here as:—

$$\frac{1}{\frac{1}{a}\cdot\frac{pq}{p+q}} + \frac{1}{\frac{1}{b}\cdot\frac{sq}{s+q}} = \frac{1}{\frac{1}{d}\cdot(p+q+s)}$$

$$\frac{a}{p}\left(\frac{p}{q}+1\right)+\frac{b}{q}\left(\frac{q}{s}+1\right)=\frac{d}{p+q+s}$$

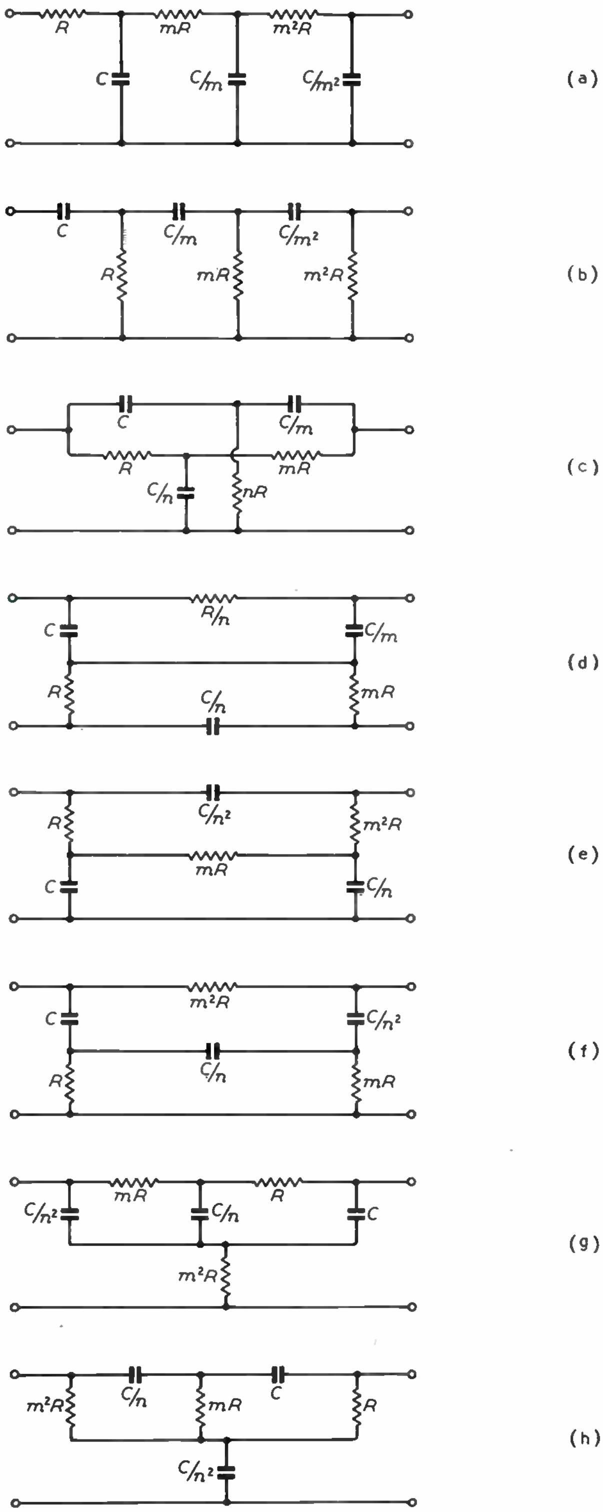


Fig. 2. *Prototype networks.*

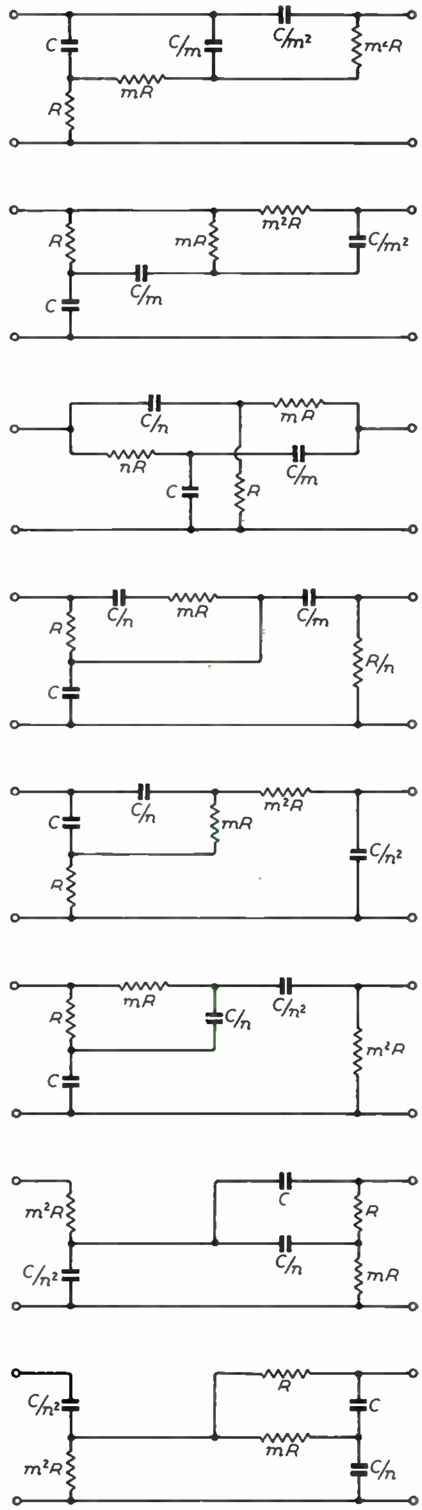


Fig. 3. *Derived networks.*

for $\frac{q}{p} \to \infty$, $\frac{s}{q} \to \infty$, $\frac{a}{p} + \frac{b}{q} = \frac{d}{s}$

$$\frac{a}{(p/q)} + b = \frac{d}{(s/q)}$$

now if $\frac{q}{p} = \frac{s}{q} = m$, $\frac{b}{a} = \frac{d}{b} = n$,

whence $\frac{(1/n)}{(1/m)} + 1 = \frac{n}{m}$ therefore $n = 1{\cdot}618\,m$.

A physical interpretation may now be envisaged, at least for the derived network. Fig. 3(e) is similar to Fig. 3(a) except that the last capacitor in 3(e) is returned to one input terminal while that

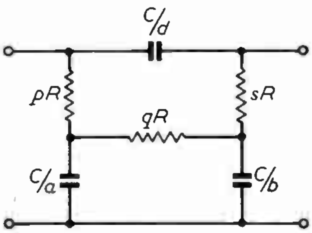


Fig. 4. *Network with generalized values.*

in 3(a) is connected to the other input terminal. The grading is now different for capacitors and for resistors. Fig. 2(f) may be treated in a similar way. If the circuits of Fig. 2(g) and Fig. 2(h) are derived by Tellegen's[3] graphical construction they will be found to exhibit the same properties as the parent connection but only when short-circuited on their output side and fed from a constant-current generator. However, the reciprocity theorem shows that if the input and output terminal pairs are interchanged as in Fig. 2 the normal working conditions may be restored.

Response Functions

For the circuit of Fig. 2(a), if $x = \omega CR$

$$\frac{v}{e} = \frac{1}{1 - (3 + 2/m)x^2 - jx[x^2 - (3 + 2/m + 1/m^2)]} \quad .. \quad (1)$$

Hence the response of the circuit of Fig. 3(a) is, for $m \to \infty$

$$\frac{v}{e} = 1 - \frac{1}{1 - 3x^2 - jx(x^2 - 3)} = \frac{-3x^2 + jx(x^2 - 3)}{1 - 3x^2 - jx(x^2 - 3)} \quad .. \quad .. \quad .. \quad .. \quad .. \quad (2)$$

which for $x^2 = 3$ yields:

$$\frac{v}{e} = 1{\cdot}125$$

A similar result holds for Fig. 3(b) when $x = \frac{1}{\omega CR}$

For the circuit of Fig. 2(c)

$$\frac{v}{e} = \frac{x[mnx^2 - n(m + nm)] - j[x^2n(n + mn) - mn]}{x[(mnx^2 - (m + n + mn + n^2 + mn^2)] - j\,[x^2(m + n + mn + n^2 + mn^2) - mn]} \quad .. \quad (3)$$

which for $x = 1$, $m \to \infty$

$$= \frac{n(n-1)}{n^2 + 1}$$

and this is a maximum when $n = \sqrt{2} - 1$.

whence $\frac{v}{e} = -\,0{\cdot}207$.

Thus for the circuit of Fig. 3(c), at the zero-phase-shift frequency,

$$\frac{v}{e} = 1{\cdot}207$$

For the circuit of Fig. 2(d)

$$\frac{v}{e} = \frac{x[mnx^2 - m(m + 1)] - j[m(m + 1)x^2 - mn]}{x[mnx^2 - (m + nm^2 + mn + m^2 + mn^2)] - j[(m + n + mn + nm^2 + mn^2)x^2 - mn]} \quad .. \quad (4)$$

whence dividing throughout by mn and letting $m \to 0$,

$$\frac{v}{e} = \frac{x(x^2 - 1/n) - j(x^2/n - 1)}{x[x^2 - (n + 1 + 1/n)] - j[(n + 1 + 1/n)\,x^2 - 1]} \quad (5)$$

and this will be real if

$$\frac{x^2 - n}{nx^2 - 1} = \frac{(1 + n + n^2)x^2 - n}{nx^2 - (1 + n + n^2)} \quad .. \quad .. \quad (6)$$

i.e., if $x = 1$,

whence $\frac{v}{e} = \frac{1 - n}{1 + n^2}$

and this is a maximum when $n = \sqrt{2} + 1$, yielding

$$\frac{v}{e} = -\,0{\cdot}207$$

For the circuit of 2(e)

$$\frac{v}{e} = \frac{x(m^3x^2 - m^2n) - j[(m^3 + m^2n)x^2 - n^3]}{x[m^3x^2 - (n^3 + nm^2 + mn^2 + n^2 + n + mn)] - j[x^2(m^3 + nm^2 + mn^2 + mn + m^2) - n^3]} \quad (7)$$

and at $x^2 = n/m$ there will be zero transmission if

$$\frac{m}{n} + 1 = \frac{n}{m} \quad \ldots \quad \ldots \quad \ldots \quad (8)$$

For m and n both tending to ∞, $n = 1{\cdot}618\ m$.

If we take the same value of x and write down the condition that v/e be purely real (i.e., that numerator and denominator have the same phase angle) we find the same result as above. Since it is possible that a different combination of x and m/n ratio may give an antiphase output, write $m/n = \alpha$ then if $m, n \to \infty$,

$$\frac{v}{e} = \frac{x(\alpha^3x^2 - \alpha^2) - j[x^2(\alpha^2 + \alpha^3) - 1]}{x[\alpha^3x^2 - (\alpha^2 + \alpha + 1)] - j[x^2(\alpha^3 + \alpha^2 + \alpha) - 1]} \quad (9)$$

which is real if

$$\frac{x^2(\alpha^3 + \alpha^2) - 1}{x^2\alpha^3 - \alpha^2} = \frac{x^2(\alpha^3 + \alpha^2 + \alpha) - 1}{x^2\alpha^3 - (\alpha^2 + \alpha + 1)} \quad (10)$$

i.e., if

$$x^4\alpha^4 + x^2\alpha^2(\alpha^2 + \alpha + 1) - (\alpha + 1) = 0 \quad (11)$$

Moreover if this last equation is true

$$\frac{v}{e} = \frac{x^2\alpha^3 - \alpha^2}{x^2\alpha^3 - (\alpha^2 + \alpha + 1)} \quad \ldots \quad \ldots \quad (12)$$

and this will be a negative quantity if

$$\frac{\alpha^3 + \alpha + 1}{x^2} > \alpha^3 > \frac{\alpha^2}{x^2} \quad \ldots \quad \ldots \quad (13)$$

Equation (11) is plotted in Fig. 5 as x vs. $1/\alpha$ because in this form it is found to approach closely a straight line. If a few trial values of x and α are taken from this graph and substituted in Equ. (12) the antiphase output is seen to go through a maximum value between $1/\alpha = 3$ and 4. The optimum values of x and α can be found by satisfying Equ. (11) and choosing values which maximize Equ. (12). Alternatively the literal solution of Equ. (11) can be substituted in Equ. (12) and the best value of α found. Either method is tedious. A very good approximation can be found by measuring the slope of the graph of Fig. 5 and substituting the value of $x\alpha$ in Equ. (12) and the resulting maximum-value problem is quite easy.

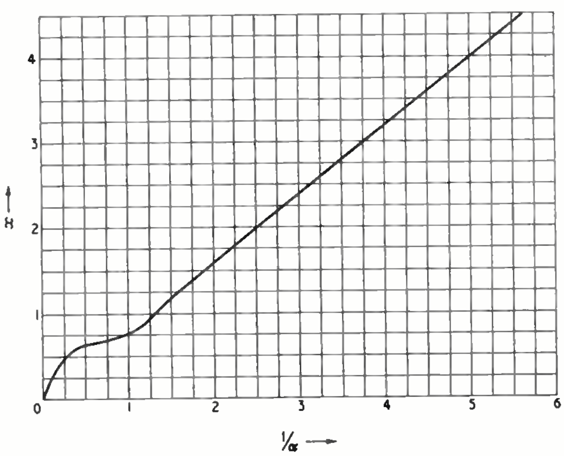


Fig. 5. *Plot of* $x^4\alpha^4 + x^2\alpha^2\ (\alpha^2 + \alpha + 1) - (\alpha+1) = 0$.

By measurement $x\alpha = 0{\cdot}8$ and substituting in Equ. (12)

$$\frac{v}{e} = \frac{\alpha(\alpha - 0{\cdot}64)}{\alpha^2 + 0{\cdot}36\alpha + 1}$$

$$\frac{d}{d\alpha}(v/e) = 0 \text{ when } \alpha^2 + 2\alpha - 0{\cdot}64 = 0$$

i.e., $\alpha = 0{\cdot}2806$, $x = 2{\cdot}85$.

A better approximation is obtained by accurately calculating a pair of x, α in the neighbourhood of these last values and substituting in the rigorous formula for the slope of the graph of Fig. 5 which is found as follows,

$$x^4\alpha^4 + x^2\alpha^2(\alpha^2 + \alpha + 1) - (\alpha + 1) = 0$$

$$x^4 + x^2 + x^2(1/\alpha) + x^2(1/\alpha)^2 - (1/\alpha)^3 - (1/\alpha)^4 = 0$$

$$\frac{d}{d(1/\alpha)}x = \frac{4 + 3\alpha - 2x^2\alpha^2 - x^2\alpha^3}{4x^3\alpha^3 + 2x\alpha(\alpha^2 + \alpha + 1)} \quad \ldots \quad (14)$$

For the circuit of Fig. 2(f)

$$\frac{v}{e} = \frac{x[m^3x^2 - (n^3 + n^2m + n^2 + mn)] - j[x^2(m + mn + mn^2) - n^3]}{x[m^3x^2 - (n^3 + m^2n + mn + mn^2 + n^2)] - j[x^2(m^3 + m^2 + m + mn^2 + mn) - n^3]} \quad \ldots \quad (15)$$

which for $n/m = \beta$, $m \to \infty$, $n \to \infty$

$$= \frac{x[x^2 - (\beta^3 + \beta^2)] - j[x^2\beta^2 - \beta^3]}{x[x^2 - (\beta^3 + \beta^2 + \beta)] - j[x^2(1 + \beta + \beta^2) - \beta^3]} \quad \ldots \quad \ldots \quad \ldots \quad (16)$$

which is real if

$$x^4(1 + \beta) - x^2\beta^2(1 + \beta + \beta^2) - \beta^4 = 0 \quad (17)$$

in which case

$$\frac{v}{e} = \frac{x^2 - (\beta^3 + \beta^2)}{x^2 - (\beta^3 + \beta^2 + \beta)} \quad \ldots \quad \ldots \quad (18)$$

and this is zero for $\beta = 0{\cdot}618$, $x = 0{\cdot}784$.

By plotting Equ. (17) in a suitable form we find $x = 1{\cdot}25$ in the region where the maximum negative output occurs. Substituting in Equ. (18) above

$$\frac{v}{e} = \frac{0{\cdot}563\beta^2 - \beta^3}{0{\cdot}563\beta^2 - (\beta^3 + \beta)}$$

and this is a maximum when

$$\beta = 0{\cdot}2812, \quad x = 0{\cdot}351.$$

For the circuit of Fig. 2(g), $v/e =$ .. (19)

$$\frac{x[m^3x^2 - m^2n] - j[m^3 + m^2n + m^3n + m^2n^2)x^2 - n^3]}{x[x^2m^3 - (m^2n^3 + m^2n^2 + m^2n + mn^2 + n^3 + mn^3)] - j[x^2(m^3n + m^2n^2 + m^2n + m^3 + mn^2) - n^3])}$$

which for $m/n = \gamma$, $m \to 0$, $n \to 0$, gives

$$\frac{v}{e} = \frac{x(x^2 - \gamma) - j[x^2(1 + \gamma) - \gamma^3]}{x[x^2 - (\gamma + \gamma^2 + \gamma^3)] - j[x^2(1 + \gamma + \gamma^2) - \gamma^3]} \quad \ldots \quad (20)$$

which is a real quantity when

$$x^4 + x^2(1 + \gamma + \gamma^2) - (\gamma^3 + \gamma^4) = 0 \quad (21)$$

in which case

$$\frac{v}{e} = \frac{x^2 - \gamma}{x^2 - (\gamma + \gamma^2 + \gamma^3)} \quad \ldots \quad (22)$$

and this is zero for $x = 1{\cdot}27$, $\gamma = 0{\cdot}618$.
In the region of maximum antiphase output $x = 0{\cdot}8$ whence

$$\frac{v}{e} = \frac{1 - 0{\cdot}64\gamma}{\gamma^2 + 0{\cdot}36\gamma + 1}$$

and this is a maximum when $x = 2{\cdot}85$, $\gamma = 0{\cdot}2806$

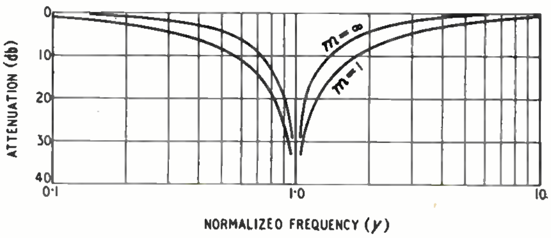


Fig. 7. *Insertion loss of parallel-T and series-π networks.*

The prototype circuits have the more strongly marked frequency characteristics and discussion of the different responses is best undertaken with reference to them. In any case, when the derived circuits are used as feedback elements a selective response reappears which is closely related to the original. Up till now the common variable has been frequency normalized so that a particular resistance and reactance are equal in magnitude at $x = 1$. So that each response may be conveniently plotted about the most interesting frequency a new variable is adopted which centres the zero- or reverse-phase-shift point in each diagram. Particulars of the value of this variable y are to be found in Appendix 1 for each of the networks.

For the circuit of Fig. 2(h), $v/e =$ .. (23)

$$\frac{x[m^3x^2 - (n^3 + n^2m + n^2m^2 + n^3m)] - j[x^2(n^2m^3 + n^2m + n^2m) - n^3]}{x[x^2m^3 - (m^2n + n^3m + n^3 + n^2m + m^2n^3)] - j[x^2(nm^3 + nm^2 + m^3 + mn^2 + m^3n^2 + m^2n^2) - n^3]}$$

which for $n/m = \delta$, $m \to 0$, $n \to 0$ is

$$\frac{v}{e} = \frac{x[x^2 - (\delta^2 + \delta^3)] - j[\delta^2x^2 - \delta^2]}{x[x^2 - (\delta + \delta^2 + \delta^3)] - j[x^2(1 + \delta + \delta^2) - \delta^3]} \quad (24)$$

and this will be real if,

$$(1 + \delta)x^4 - x^2(\delta^2 + \delta^3 + \delta^4) - \delta^4 = 0 \quad (25)$$

whence:

$$\frac{v}{e} = \frac{x^2 - (\delta^3 + \delta^2)}{x^2 - (\delta + \delta^2 + \delta^3)} \quad \ldots \quad (26)$$

which is zero for $x = 0{\cdot}784$, $\delta = 0{\cdot}618$ and gives maximum negative output when $\delta = 0{\cdot}2812$, $x = 0{\cdot}351$.

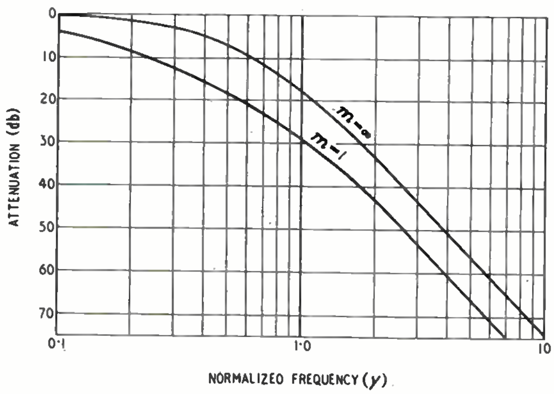


Fig. 6. *Response of 3-mesh ladder circuit.*

The effect of grading is shown for the Type (a) circuit in Fig. 6. Among the second group a symmetrical response is always obtained for a null network. In the case of (c) and (d)

$$\frac{v}{e} = \cos\theta/\theta \quad \ldots \quad (27)$$

where $\theta = \tan^{-1}\dfrac{2}{y - 1/y}$

while for (e) to (h) inclusive

$$\frac{v}{e} = \cos\psi/\psi \quad \ldots \quad (28)$$

where $\psi = \tan^{-1}\dfrac{2{\cdot}06}{y - 1/y}$

However, when the important step of modifying the grading ratio is taken, only the very symmetrical arrangements of (c) and (d) retain a symmetrical response. It is convenient when writing the response of these phase inverting circuits to include a displacement term showing the amount of inversion. For (c) and (d)

$$\frac{v}{e} = 1{\cdot}207 \cos \nu \underline{/\nu} - 0{\cdot}207 \quad \ldots \quad (29)$$

where

$$\nu = \tan^{-1} \frac{2{\cdot}828}{y - 1/y}$$

while for (e) to (h) inclusive,

$$\frac{v}{e} = \frac{1{\cdot}0855(y^2 - 1)(1{\cdot}96y - j)}{1{\cdot}96y(y^2 - 3{\cdot}1) - j(7{\cdot}56y^2 - 1)} - 0{\cdot}0855 \quad (30)$$

The numerical values of this last expression strike an average between the results obtained for the sub-groups alone. This is necessitated by the approximate calculation made above. For example, the value of δ comes out as $\delta = 9/32 = 0{\cdot}2812$, while that for α as $\alpha = -1 + \sqrt{41}/5 = 0{\cdot}2806$. It will be seen that the error is quite small. There is not much to choose between the various arrangements in the second group as regards the null connection but it will be seen that the symmetrical circuits offer advantages in the amount of antiphase output they provide. These features are displayed in Figs. 7 and 8.

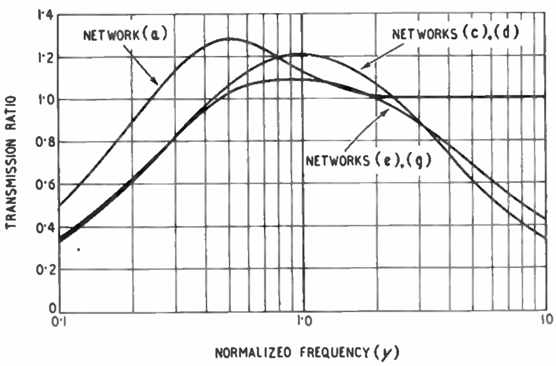


Fig. 8. *Derived circuit responses.*

Matrix Representation

Knowing the elements of the matrix of the original network it is convenient to be able to write down those of the derived network. To do this we shall first construct the equivalent-T of the original elements. With the usual notation.

$$Z_{01} = \frac{A}{C},\ Z_{s1} = \frac{B}{D},\ Z_{02} = \frac{D}{C},\ Z_{s2} = \frac{B}{A} \quad (31)$$

and the shunt element of the T is given by

$$Z_{sh} = \sqrt{Z_{02}(Z_{01} - Z_{s1})} = \frac{1}{C} \quad \ldots \quad (32)$$

The equivalent circuit is shown in Fig. 9(a) while Fig. 9(b) shows the derived circuit. Setting up the usual equations for this circuit we have

$$\begin{aligned} E_3 &= \frac{A}{A-1} E_4 + \frac{B}{A-1} I_4 \\ I_3 &= \frac{C}{A-1} E_4 + \frac{A + D - 2}{A-1} I_4 \end{aligned} \quad \ldots \quad (33)$$

The elements of the new matrix are thus easily formed from those of the old.

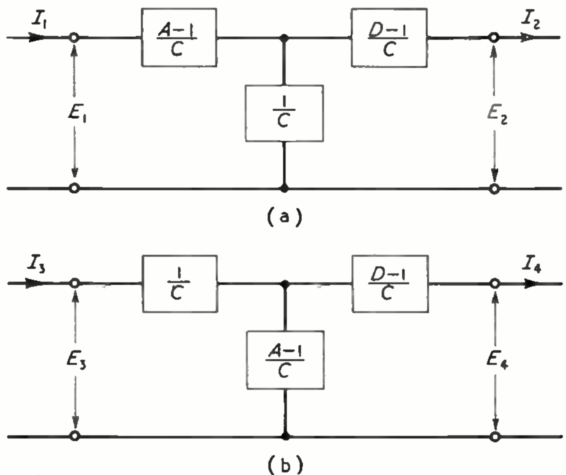


Fig. 9. Equivalent-T in matrix terms of (a) original, (b) derived circuit.

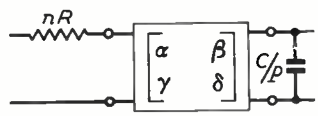
Effect of Termination on Response

At the frequencies at which these circuits are likely to be used the loading on the networks is of two kinds,

(a) capacitive across the output terminals,
(b) resistive in series with the input.

These elements, added to the circuits under examination, necessitate fourth-order determinants in finding the response function. Their character is such that an alternative method of evaluation is preferable. The loaded circuit is shown in Fig. 10 where α, β, γ, and δ are the elements of the matrix as found in the previous paragraph. The overall matrix of the circuit will now be

Fig. 10. *Loaded circuit as tandem connection of 3 quadripoles.*



$$\begin{bmatrix} 1 & n \\ 0 & 1 \end{bmatrix} \begin{bmatrix} \alpha & \beta \\ \gamma & \delta \end{bmatrix} \begin{bmatrix} 1 & 0 \\ j\frac{x}{p} & 1 \end{bmatrix} =$$

$$\begin{bmatrix} \alpha + n\gamma + j\frac{x}{p}(\beta + n\delta) & \beta + n\delta \\ \gamma + j\frac{x\delta}{p} & \delta \end{bmatrix}$$

The response function of the circuit is given by the reciprocal of the leading element of the matrix,

$$\text{i.e., } \frac{v}{e} = \frac{1}{\alpha + n\gamma + j\frac{x}{p}(\beta + n\delta)} \quad \ldots \quad \ldots \quad (34)$$

and if we consider the Type 3(b) circuit,

$$\alpha = \frac{1 - 3x^2 + jx(3 - x^2)}{-3x^2 + jx(3 - x^2)} \quad m \to \infty \quad (35)$$

$$\beta = \frac{(1 + m + m^2) - m^2x^2 + jx(1 + m + m^2)}{-3x^2 + jx(3 - x^2)} \quad \ldots \quad \ldots \quad (36)$$

which for $m \to \infty$

$$= m^2 \frac{(1 - x^2) + j2x}{-3x^2 + jx(3 - x^2)}$$

$$\gamma = \frac{-2x^2 - jx^3}{-3x^2 + jx(3 - x^2)} \; (m \to \infty) \quad \ldots \quad (37)$$

$$\delta = \frac{-x^2(3 + m^2) + jx(3 + 2m + m^2 - x^2)}{-3x^2 + jx(3 - x^2)} \quad \ldots \quad \ldots \quad (38)$$

whence

$$\frac{v}{e} = \frac{-3x^2 + jx(3 - x^2)}{1 - x^2(3 + 2n + \frac{2m^2 + 5n + m^2n}{p} + x^4\frac{n}{p} + jx\left[3 + \frac{m^2}{p} - x^2\left(1 + n - \frac{m^2 + 3n + m^2n}{p}\right)\right]} \quad (39)$$

For the present application interest is focused on the values of x at which the imaginary part of Equ. (39) is zero while the real part is greater than unity. It will be seen that a limiting solution for $m = \infty$ is not possible since the slightest terminal capacitance would destroy the network properties. However, for a special high value, $m = 10$, the phase condition yields

$$x^6\left(\frac{n}{p}\right) + x^4\left(\frac{100}{p} - 3n - 209\frac{n}{p}\right) + x^2\left(1 + 6n + \frac{300}{p} + 315\frac{n}{p}\right) - 3 = 0 \quad (40)$$

This amount of grading probably represents the ultimate amount which could be used in practice. The above equation enables a study to be made of the permissible terminal loading consistent with a required transmission factor.

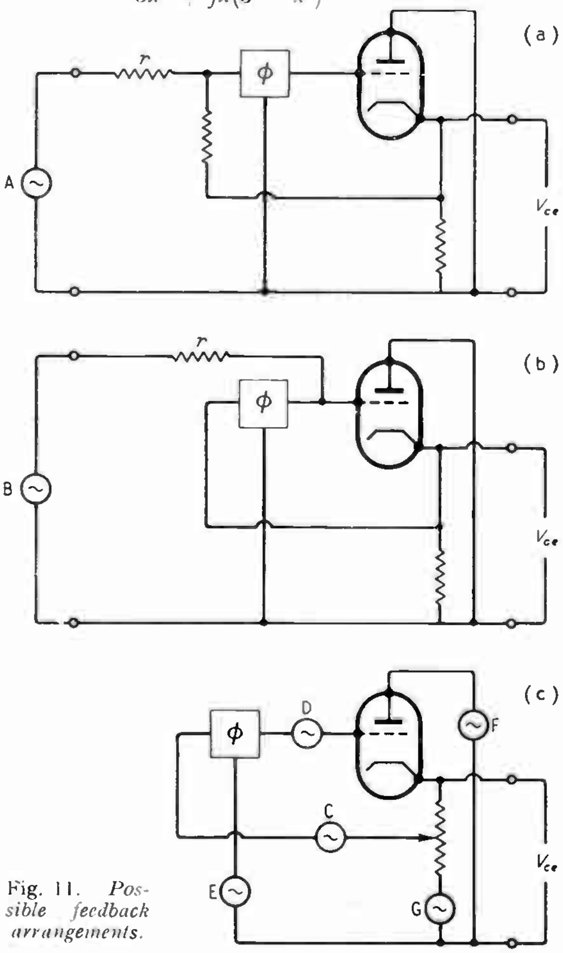


Fig. 11. *Possible feedback arrangements.*

The Circuits as Feedback Elements

In the same way that the prototype circuits can be connected between anode and grid of a valve to give an overall response function which is the inverse of the original, the derived circuits may be placed between cathode and grid. In order to be useful two precautions must be taken:

(a) the loop gain should be kept below unity to avoid the possibility of self-oscillation.

(b) the input must be applied at the proper place.

In Fig. 11 seven methods of inserting input are shown. Not all these methods are very practical, especially if they require an input with both terminals above earth. In cases A, B, F, G, the stability of the amplifier is ensured by correct choice of the value of a particular component, in C, D, E the input could be tapped down the cathode load. Except where otherwise stated, we shall assume that the input and output impedances of the feedback circuits are very large and that the circuit of Fig. 3(c) is taken as representative of the performance of all the circuits. This circuit has a response function which may be written

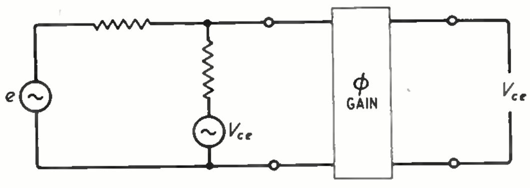
$$\frac{v}{e} = 1{\cdot}207 \cos\psi \; \underline{/\psi} \quad \ldots \quad \ldots \quad \ldots \quad (41)$$

where $\tan\psi = 0{\cdot}353(y - 1/y)$, or more conveniently,

$$\frac{v}{e} = \frac{1{\cdot}207}{1 + js} \text{ where } s = \tan \psi \quad .. \quad (42)$$

The equivalent circuit of Fig. 11(a) is shown in Fig. 12(a). If the transfer response of the feedback circuit is ϕ, then its combination with the cathode-follower may be represented as an ideal amplifier with infinite input impedance, zero output impedance and complex gain ϕ. In practice the finite input and output impedances will modify results in those cases where the design of the feedback network is controlled by stray capacitances. The input to the ideal amplifier consists of the original source and the voltage across the cathode load mixed in the two resistances shown. The effective input voltage is thus,

$$e\frac{\alpha}{\alpha+1} + v_{ce}\frac{1}{\alpha+1}$$



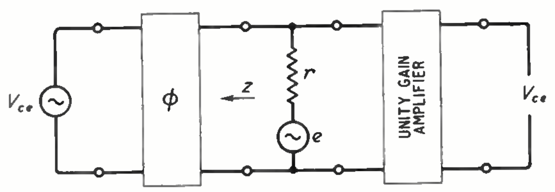


Fig. 12. *Equivalent circuits.*

The output of the ideal amplifier may be equated to the cathode-earth voltage:

$$\frac{\phi}{\alpha+1}(\alpha e + v_{ce}) = v_{ce}$$

$$\therefore \frac{v_{ce}}{e} = \frac{\alpha\phi}{1+\alpha-\phi} = A$$

$$\therefore A = \frac{1}{\dfrac{1+\alpha}{\alpha\phi} - \dfrac{1}{\alpha}} \quad .. \quad (43)$$

This may be quite easily computed since

$$\frac{1}{A} = \left(\frac{1+\alpha}{1{\cdot}207} - \frac{1}{\alpha}\right) + j\frac{s}{1{\cdot}207\alpha}$$

The gain factor A is plotted in Fig. 13 for various values of α. For α greater than 0·207 the system is stable but with smaller values self-oscillation will take place. Although the output voltage can never rise above a value fixed by the supply voltage and valve characteristics, in the stable region the gain can theoretically increase without limit. Since the overall response now resembles that of an LC acceptor circuit it is natural to associate with it a Q-value to describe its selectivity. Power considerations are not involved

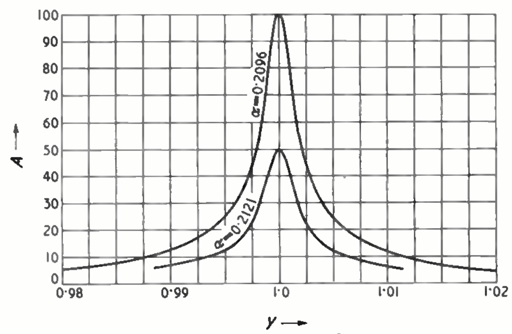


Fig. 13. *Feedback amplifier response.*

here so that the more complex definitions of Q are not applicable and we shall be content to define it as the ratio of the centre frequency to the total bandwidth at 3 db below maximum response. Maximum response is given here by

$$A_0 = \frac{1}{\left(\dfrac{1+\alpha}{1{\cdot}207\alpha} - \dfrac{1}{\alpha}\right)} = \frac{1{\cdot}207\alpha}{\alpha - 0{\cdot}207} \quad .. \quad (44)$$

The reciprocal of the value of s, say s_1, which satisfies $A^2 = \frac{1}{2} A_0^2$ .. .. .. (45) will be numerically equal to the value of Q according to our definition, since s_1 is the difference between the appropriate values of y satisfying Equ. (45) and y is normalized. The solution of Equ. (45) is

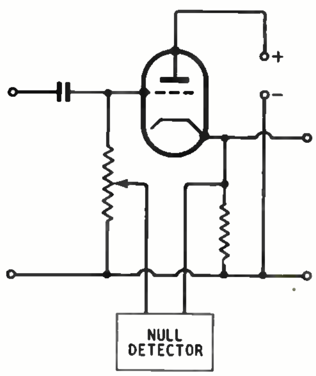


Fig. 14. *Rogers' circuit.*

$$\frac{1}{s_1} = Q = \frac{1}{\alpha - 0{\cdot}207} \quad .. \quad .. \quad (46)$$

and substituting in Equ. (44)

$$Q = \frac{A_0 - 1{\cdot}207}{1{\cdot}207 \times 0{\cdot}207} \quad .. \quad .. \quad (47)$$

which for large values of A_0 is equal to $4A_0$.

The attainment of large values of Q will obviously depend on the stability of the feedback circuit, mixer network and cathode follower, and should properly form a separate study. Nevertheless, it is of interest to examine the comparative performance of the other possible connections. The equivalent circuit of the arrangement of Fig. 11(b) is shown in Fig. 12(b), Z being the short-circuit impedance of the feedback network as seen from the grid. The input to the ideal amplifier (of unity gain) now consists of the sum of the 'potted-down' value of the input signal across the short-circuit impedance Z and the corresponding value of the open-circuit voltage from the ϕ-network appearing across the resistance r. It will be seen that r and Z fulfil the function of a mixing network here. As before,

$$V_{ce} = \frac{Z}{r + Z} e + \frac{r}{r + Z} \phi V_{ce} \quad \ldots \quad (48)$$

$$A = \frac{\beta}{1 + \beta - \phi}$$

where $\beta = Z/r$

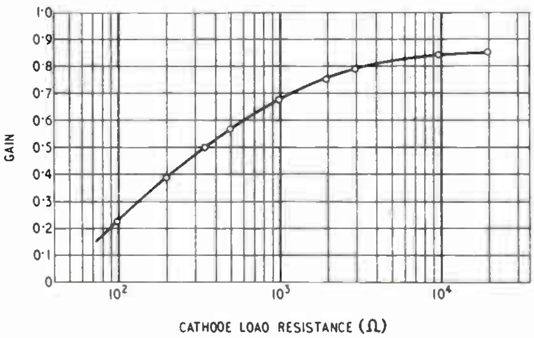


Fig. 15. *Effect of cathode load on gain.*

Using the previous method,

$$Q^2 = \left(\frac{1 + \beta}{\beta - 0{\cdot}207}\right)^2 - 2$$

and since $A_0 = \dfrac{\beta}{\beta - 0{\cdot}207}$,

for large gains $Q = 5{\cdot}83 A_0 \quad \ldots \quad \ldots \quad (49)$

In the circuit of Fig. 11(c) the input to the ideal amplifier (of gain ϕ) is the sum of the signal and a fraction γ of the output voltage. Here

$$A = \frac{1}{1/\phi - \gamma}$$

$$Q = 0{\cdot}83\, A_0 \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (50)$$

In the connection at D of the same figure the signal is injected after the output sample has been through the ϕ-network and

$$A = \frac{1}{1 - \gamma\phi}\,;\; Q = A_0. \quad \ldots \quad \ldots \quad (51)$$

In the arrangement E, where the signal is applied between the network common terminal and earth, the grid sees the signal as though coming from the prototype circuit of Fig. 2 and the output fraction as from the derived circuit.

$$\text{Hence } A = \frac{1 - \phi}{1 - \gamma\phi} \quad \ldots \quad \ldots \quad \ldots \quad (52)$$

so that for $\gamma = 1$, the entire circuit behaves as a zero-phaseshift all-pass device.

For the connection at F

$$A = \frac{1}{1 + r_a/R + \mu(1 - \phi\alpha)}$$

where r_a is the differential anode resistance of the valve and if $\alpha = 0{\cdot}83$

$$Q^2 = (\mu A_0 + 1)^2 - 2 \quad \ldots \quad \ldots \quad \ldots \quad (53)$$

In this case the performance depends closely on the valve properties and, since there is no likelihood of oscillation taking place, the maximum attainable selectivity is quite small. A similar situation obtains at G where

$$A = \frac{r_a}{R + r_a + \mu R\,(1 + \phi\alpha)}$$

$$Q^2 = \left(1 + \frac{\mu R A_0}{r_a}\right)^2 - 2 \quad \ldots \quad \ldots \quad (54)$$

An interesting feature of the above calculations is the fact that the Q-factor of the composite circuit is numerically several times larger than the gain required to provide the selectivity. This is in contrast to the fractional results obtained when the prototype networks are employed in the normal anode-to-grid connection.[4,5] It would be interesting to discover whether the quality-factor/gain ratios for the two arrangements bear a reciprocal relation to one another. It also remains to be seen whether the fact that equal selectivity can be obtained from the cathode-follower at a lower gain implies that such a connection is more stable.

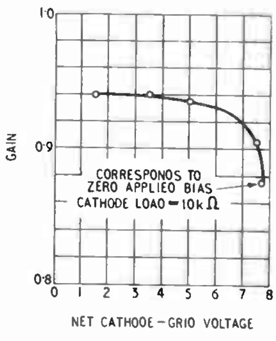


Fig. 16. *Effect of bias on gain.*

Experimental Work

The first aspect of the investigation concerned the measurement of the actual gain obtainable with a cathode-follower. It is commonly assumed that, if the cathode load is sufficiently high, then the gain is unity. There is a practical limit to the resistance tolerable between cathode and earth, and we are here interested in the precise magnitude of the defect from unity gain. Using a MHL4 valve ($\mu = 16$, $r_a = 6{,}500\ \Omega$) with zero bias, the variation of gain with cathode resistance was

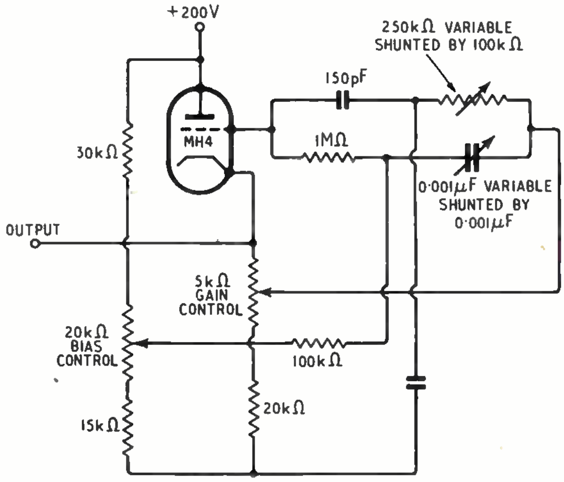


Fig. 17. 1000-c/s *oscillator.*

measured at 50 c/s, using the method, due to Rogers,[6] shown in Fig. 14. The detector consisted of a vibration galvanometer. The results, in Fig. 15, show that the performance is hardly adequate. Next, the effect of bias was examined, in this case on a MH4 valve ($\mu = 40$, $r_a = 11{,}500\ \Omega$). The input potentiometer in the measuring circuit, instead of being earthed at its lower end, was returned through a low-resistance source of positive voltage. The need for correct bias is shown in Fig. 16, where it will be seen that enough gain is available to exploit the properties of the feedback circuits. Since it is intended to devote further study to the use of the networks in selective amplifiers, the only further verification of theory attempted so far has been the construction of the 1,000-c/s oscillator of Fig. 17, in which the grading ratio is 10. An approximate measurement of the influence of supply voltage on frequency suggests that this is of the order of 25 parts per million per volt. It is expected that the use of other feedback circuits, notably that of Fig. 3(e), will enable RC oscillators to be made from standard components for much higher frequencies than hitherto.

Conclusion

A further technique has been examined for interconnecting valves and circuits. It has been realized for some time that the operational form of the transfer function of a four-terminal network can be inverted by using it as a feedback path in an amplifier. Here we have shown that similar and possibly improved results can be obtained by taking as the output terminals of the four-terminal network the usually-neglected pair and, as amplifier, a cathode-follower.

Since the first preparation of this paper as an internal report for A. C. Cossor, Ltd., a contribution from the synthesis point of view has been published by Epstein.[7]

The author is indebted to the Governors and Principal of Paisley Technical College for the use of the facilities of the Electrical Engineering Dept., and to Mr. B. C. Fleming-Williams, Director of Research, A. C. Cossor, Ltd., London, for material assistance in the preparation of the typescript.

APPENDIX 1.

Normalization of the Frequency Variable

The response of individual circuits was worked out in terms of the variable x where $x = \omega CR$. In many simple structures the interesting behaviour centres round the value $x = 1$. Here, however, the zero-phase-shift frequency of the derived network may occur at quite different values of x and a new variable, y, is introduced to shift the important region to around $y = 1$. The following definitions apply to the circuits of Fig. 3, the various types being given in brackets.

(a), (b)	$x = 1{\cdot}732y$
(c), (d)	$x = y$
(e), (g)	$x = 2{\cdot}85y$
(f), (h)	$y = 2{\cdot}85x$

Similarly the definitions for the primitive networks are

(a), (b)	$x = 1{\cdot}732y$
(c), (d)	$x = y$
(e), (g)	$x = 1{\cdot}272y$
(f), (h)	$y = 1{\cdot}272x$

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RESISTANCE-CAPACITANCE NETWORKS WITH OVER-UNITY GAIN

By W. Bacon, B.Sc., A.M.I.E.E. and D. P. Salmon, B.Sc.

SUMMARY.—Resistance-capacitance networks can be produced having a gain which is theoretically unlimited. Such circuits also show interesting impedance properties and, with their aid, RC oscillators may be designed in which the valve functions solely as a cathode follower.

Introduction

An interesting property of RC networks is their ability under certain conditions to produce an output voltage greater than the input voltage.

Such circuits have been described by Longmire[1] and Epstein.[2] Longmire applies a voltage V_{in} (Fig. 1) to a phase-retarding network R_1C_1. A second network R_2C_2, assumed of negligible loading effect, is connected across R_1. The resultant output voltage V_o is retarded in phase relative to the input and of greater magnitude. By feeding the voltage thus obtained to a similar network having phase-advancing properties, a resultant voltage can be obtained in phase with the input and of greater magnitude.

Epstein points out that any three-terminal network having an output with an antiphase component may be reconnected to give an output voltage greater than the input voltage. Thus, in Fig. 2(a), output may be taken from A and C instead of from C and B, and will then be greater than the input.

The gain obtained by the above methods is, however, small. Longmire obtains a ratio of output to input of 1·16, and Epstein seems to consider that the maximum gain which could ever be achieved would be 2.

The method of approach described below, however, yields networks which theoretically may have unlimited gain.

Over-Unity Gain from an *N*-Phase Network

Consider the voltage V_{AB} applied to the network of Fig. 3(b) or (c). By suitable choice of values X_1

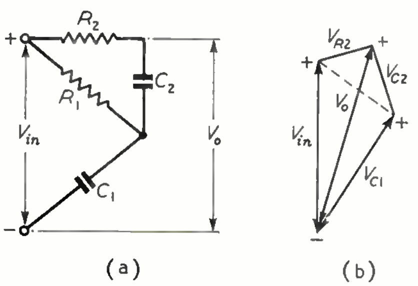


Fig. 1. *Double RC network (a) and vector diagram (b)*

MS accepted by the Editor, March 1952.

can be made to lie anywhere on the left-hand semi-circle and X_2 on the right-hand semi-circle. By using (*N*-2) networks on each side, where *N* is any even number, an *N*-phase system can be obtained, as shown in Fig. 4. If *N* is an odd number, the same result can be obtained with one more network.

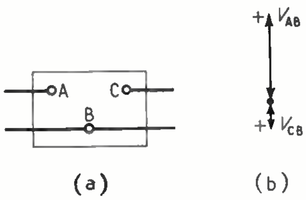


Fig. 2. *Three-terminal network (a) gives a voltage across AC greater than at AB when the voltage at CB is anti-phase to that at AB as shown at (b).*

Now suppose any two of the phases are applied to an RC network, as in Fig. 5. If the frequency is chosen so that the reactance of the capacitance is equal to the resistance, and if the values are high enough so that negligible loading is imposed on the previous circuit, then it will be seen that a new vector, V'_A, has been produced of greater magnitude than those from which it was derived and midway between them in phase. By proceeding thus for each pair of vectors a complete new *N*-phase system is obtained, the magnitude of each phase in which is greater than the original. Hence, considering input and taking output from any opposite pair of phases of the *N*-phase system, a circuit has been produced in which output is greater than input. The whole procedure may be repeated again giving still greater output voltage.

It will be seen that by this method an RC network can be made to produce a theoretically indefinitely large voltage gain. In practice, the gain obtainable is restricted by the increase in impedance which is necessary at each stage.

Four-Phase Network

The gain obtained will vary with the number of phases *N* into which the input is split, being greatest when *N* equals 4. Fig. 6 shows a possible 4-phase network. The resistors and capacitors R_1C_1 form the phase-splitting section; R_2C_2, the first voltage-multiplying network, and R_3C_3, the second voltage-multiplying network. Output is taken in phase from A and B or in quadrature from C and D. If no section had any appreciable loading effect on any other the overall gain would be 2. In practice, the gain will be reduced due to

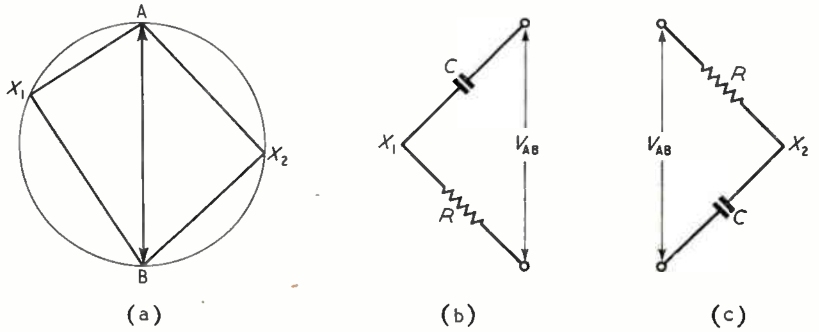


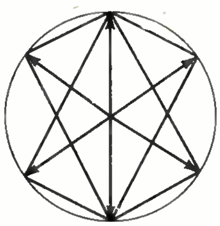
Fig. 3. *Basic networks (b) and (c) with vector diagram (a).*

the loading effect of one section upon another. Since, apart from the initial phase-shifting circuit, the network consists of a series of identical nets connected in tandem, a theory taking account of loading can be developed by working out for any net in general:

(1) Reduction in output voltage when loaded with any impedance, and
(2) The input impedance of the net under these conditions.

Consider a net (Fig. 7) with a 4-phase voltage applied through zero impedance and $R_1 = 1/\omega C_1$. Since the network is symmetrical there are fundamentally only two different currents, the others being obtained by multiplication by appropriate powers of the imaginary operator. Solving for I_1 and I_2, and using the results to obtain V_{on}, gives (see Appendix)

$$V_{on} = V \frac{\omega CZ(1-j)}{(1-j)+\omega CZ} = V \frac{\frac{Z}{R_1}(1-j)}{\frac{Z}{R_1}+(1-j)}$$



It will be seen that as $Z \rightarrow \infty$ this tends to $V(1-j)$ which is equal in magnitude to $\sqrt{2}$ times the voltage with no load.

Fig. 4. *Vector diagram of N-phase system.*

Input Impedance of Net

The above result also enables the input current to be calculated at each input point, and thus the input impedance for each phase. Thus, input impedance $= R_1/2\ (1-j)$.

The rather-surprising result is thus obtained that the input impedance of a net, measured between any input point and the common point, is independent of the load imposed on the net, provided this be symmetrical. As a result, each net can be dealt with separately and its output voltage will depend on the composition of the next

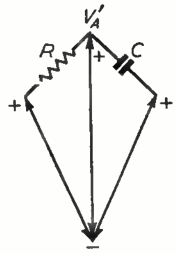


Fig. 5. *Two phases applied to an RC network.*

net, but not on the termination of the next net, a valuable simplification.

Output Voltage of Net Feeding into Second Net

Substituting the expression for the input impedance of a net in that obtained for the output voltage the result is obtained that:—
Output voltage equals input voltage

$$\times (1-j) \times \frac{1}{1 + R_1/R_2}$$

The output voltage of a net thus depends solely on the ratio R_1/R_2, and for the network to show greater-than-unity gain R_2/R_1 must be not less than a little under 3.

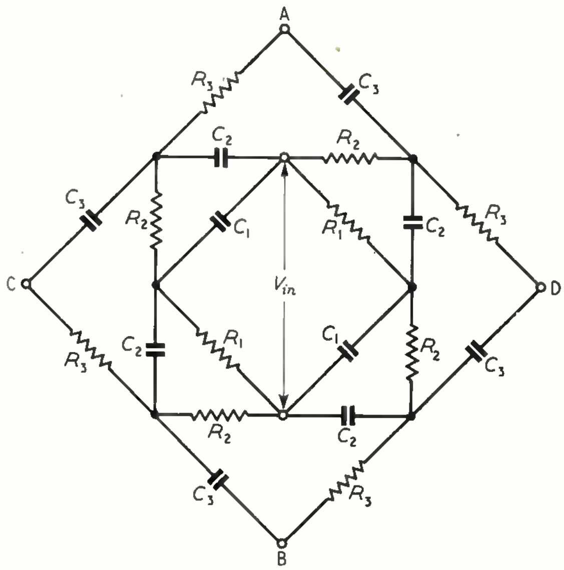


Fig. 6. *4-phase network. Values used experimentally were* $R_1 = 100\Omega$, $R_2 = 1\ k\Omega$, $R_3 = 50\ k\Omega$, $C_1 = 2{\cdot}5\ \mu F$, $C_2 = 0{\cdot}25\ \mu F$, $C_3 = 0{\cdot}005\ \mu F$.

Experimental Results

A network was built up with the values shown in Fig. 6. The following figures were obtained:

Theoretical gain at 640 c/s = 1·98
Measured gain at 640 c/s = 1·67

While the measured gain is less than the theoretical gain, it appears that figures approaching the calculated value may be obtained. Some discrepancy may be attributed to the tolerances of the components used.

The effect was tried of removing the outside network. This appeared to have no effect on the input impedance, thus confirming the result obtained theoretically.

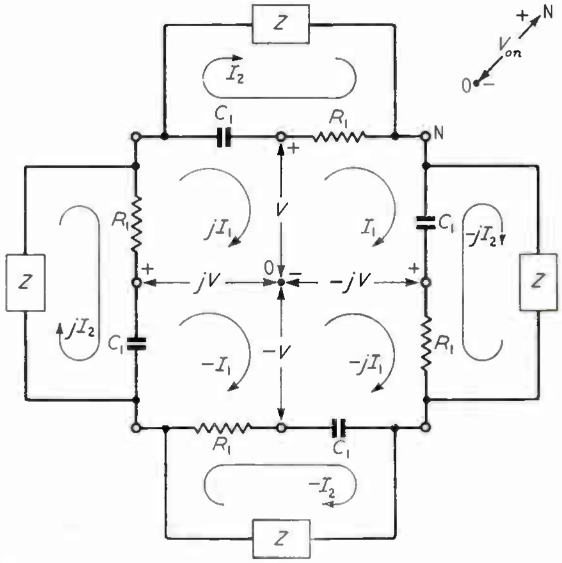


Fig. 7. *Circuit used for calculating input impedance to a 4-phase voltage.*

Construction of an Oscillator

These networks may be used in conjunction with a cathode follower to construct oscillators in which the valve has a gain of less than unity. Fig. 8 shows such a circuit built up by the writers. (The valve was operated at a rather higher anode voltage than that recommended by the manufacturers, and the circuit is, therefore, to be regarded as experimental.)

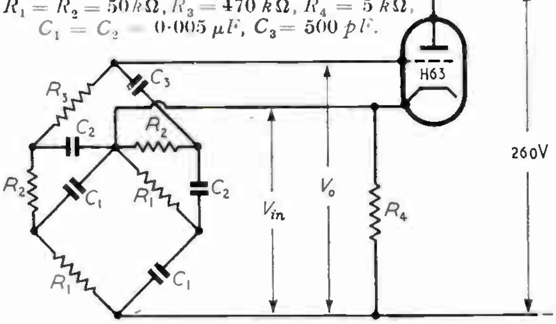
While complete networks lend themselves best to theoretical analysis, it is, in practice, often possible to simplify these with some loss of gain by omitting certain elements. The network in Fig. 8 is thus a degenerate version of Fig. 6.

The oscillator was found to have good waveform and stability, measured frequency agreeing closely with the calculated value.

Build-up of Oscillations

In order to examine the building up of oscillations, an experimental oscillator having a very long period of approximately 20 sec was constructed, and anode current observed from the instant of switching on. The curve of current against time thus obtained shows a sinusoidal oscillation superimposed on an exponentially changing current (Fig. 9).

Fig. 8. *Experimental oscillator circuit:* $R_1 = R_2 = 50\,k\Omega$, $R_3 = 470\,k\Omega$, $R_4 = 5\,k\Omega$, $C_1 = C_2 = 0{\cdot}005\,\mu F$, $C_3 = 500\,pF$.



There is one interesting feature of this curve which seems to show a departure from the usual ideas of oscillator behaviour. Instead of building up, the oscillation builds down—the final value being less than the initial. (It might be suggested that the oscillation was dying out exponentially, but observations over a period indicated that this was not so.) As a result the oscillator attains its final frequency and voltage after a very few cycles.

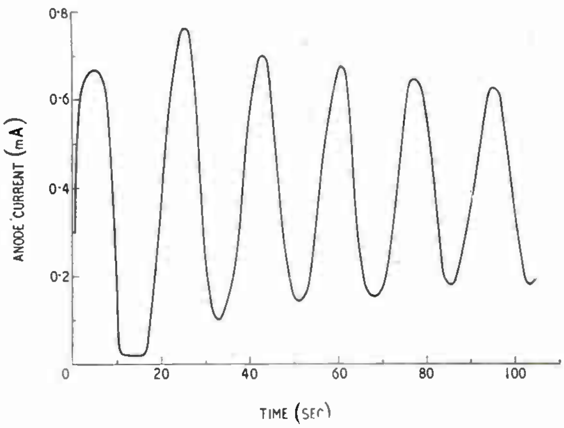


Fig. 9. *"Build-up" waveform of oscillator.*

Conclusion

The circuit described offers interesting possibilities where it is desired to synthesize a network with a gain slightly greater than unity.

Acknowledgment is made to the authorities of the Municipal College, Portsmouth, in whose laboratories the experimental part of this work was carried out.

APPENDIX

Input Impedance of Net and Output Voltage.

Considering the right-hand top corner in Fig. 7

$$V - (-jV) = (I_1 - I_2) R_1 + \left(I_1 - (-jI_2)\right)\frac{1}{j\omega C_1}$$

$$= I_1R_1 - I_2R_1 + \frac{I_1}{j\omega C_1} + \frac{I_2}{\omega C_1}$$

$$= I_1\left(R_1 + \frac{1}{j\omega C_1}\right) + I_2\left(\frac{1}{\omega C_1} - R_1\right)$$

Thus if $R_1 = \dfrac{1}{\omega C_1}$

$$V(1 + j) = I_1(R_1 - jR_1) = I_1R_1(1 - j)$$

Thus $I_1 = \dfrac{V}{R_1} \cdot \dfrac{(1+j)}{(1-j)} = \dfrac{jV}{R_1}$

Input impedance $= \dfrac{V}{I_1 - jI_1} = \dfrac{V}{I_1(1-j)}$

$$= \frac{V}{\dfrac{V(1+j)(1-j)}{R_1 \quad (1-j)}}$$

$$= \frac{R_1}{1+j} = R_1\frac{(1-j)}{2}$$

Also, considering the circuit for I_2

$$(I_2 - I_1)R_1 + (I_2 - jI_1)\frac{1}{j\omega C} + I_2Z = 0$$

and putting $\dfrac{1}{\omega C_1} = R_1$

$$I_2R_1 - I_1R_1 - jI_2R_1 - I_1R_1 + I_2Z = 0$$

$$\therefore I_2(Z + R_1(1 - j)) = 2I_1R_1$$

$$\therefore I_2 = \frac{2R_1}{Z + R_1(1-j)} \times \frac{V(1+j)}{R_1(1-j)}$$

$$= 2V \cdot \frac{(1+j)}{(1-j)} \cdot \frac{1}{Z + R_1(1-j)}$$

$$I_2 = \frac{2jV}{Z + R_1(1-j)}$$

$$\therefore (I_1 - I_2)R_1 = jV - \frac{2jVR_1}{Z + R_1(1-j)}$$

and $V_{on} = V - (I_1 - I_2)R_1 = V - jV + \dfrac{2jVR_1}{Z + R_1(1-j)}$

$$= V\frac{Z(1-j)}{Z + R_1(1-j)} = V\frac{\dfrac{Z}{R_1}(1-j)}{\dfrac{Z}{R_1} + (1-j)}$$

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CORRESPONDENCE

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

Current-Noise in Semi-conductors

SIR,—A review of the published work on the additional noise in semiconductors, which arises when a steady current flows, indicates that the nearest integral-power relationship is of the form $V^2_{df} \propto i^2/f$ where V^2_{df} is the mean-square fluctuation voltage at the terminals of a semiconductor which is carrying a steady current i, the fluctuation being measured in a narrow band df centred on the frequency f. This is particularly well illustrated by the work of Bernamont,[1] covering four decades of frequency, when due allowance is made for the presence of ordinary Johnson noise as well as current-noise.[2]

Moreover, if the current-noise is a modulation phenomenon, typified by $V^2 = i^2(\delta R)^2$, it follows from the balancing of 'dimensions' in this formula that V^2 must be proportional to i to the power of *two*, after which it is almost essential that it be proportional to f to the power of *minus one*.

The fact that V^2_{df} varies inversely with frequency, though it may not be fully proved that the power is precisely minus one, down to 0·01 c/s,[3] makes it difficult to explain the spectral distribution in terms of time-constants on the molecular scale. Moreover, a number of experiments have been carried out in the endeavour to distinguish between a surface effect and a volume effect, but in no case have they disproved the possibility of a volume effect.

It seems obvious to suggest that the current-noise is due to simple statistical fluctuation of the numbers of charge-carriers present in the bulk of the semi-conductor, but hitherto no such theory has been able to explain the inverse-frequency law. The earlier theories were based on the 'flicker' mechanism and, attributing separate current-fluctuations to the individual 'impurity centres' or the like, they predicted a spectral distribution of the form $V^2_{df} \propto 1/(1 + \omega^2\tau^2)$ where $\omega = 2\pi f$ and τ is the mean life of the statistically-distributed impurity centres. But in a volume effect one must first sum all the charge-carriers present to find a resistance (or conductance) and then find the noise from the mean-square fluctuation of this total number. If the fluctuation of number of charge carriers is assumed to be due to independent random (Poisson) processes of generation and of destruction of charge-carriers, the calculated spectral distribution is $V^2_{df} \propto 1/f^2$.

This is not the experimental spectrum, and the sum of two random processes does not lead to a stable average population—unless the constant mean rates of gain and loss are equal to an unimaginable degree, the population will either vanish or grow without limit. By introducing a mean re-combination rate which is a function of the number of carriers present, Surdin[4] overcame the latter difficulty, and also arrived at a spectral distribution somewhat similar to the now 'classic' relaxation-time expression $1/(1 + \omega^2\tau^2)$

The writer believes, however, that allowance for the *correlation between instantaneous gain and loss* which must also result from the influence of total number on the instantaneous re-combination rate will lead to a spectral distribution close to $1/f$. Preliminary calculations have shown (as is perhaps obvious) that the system with correlation will have a spectrum somewhere between the $1/f^2$ for a completely random variation in total number and the independence of frequency which is found in Johnson noise when the number is constant. The latter might be regarded as equivalent in effect to a limiting case in which the 'feedback' due to the dependence of re-combination on number present was so strong that any incipient variation was immediately cancelled.

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Electrical Engineering Department,
University of Birmingham.
30th October 1952.

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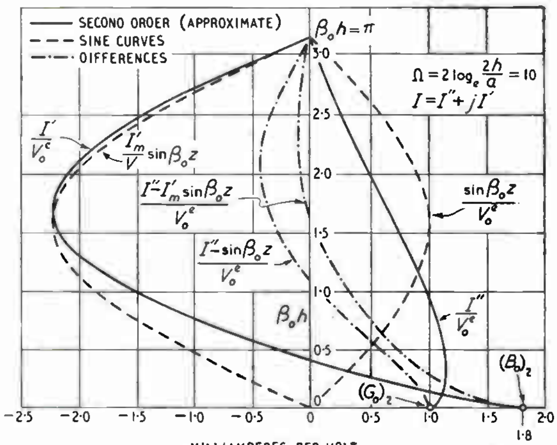
Cylindrical Aerials

Sir,—The following comments concern the interesting paper by B. Storm in the July issue of *Wireless Engineer*.

Storm's method of solving Hallén's integral equation consists in expressing the current under the sign of integration as a sum of a sinusoidal term and a trigonometric series which, since the current is an even function of z and must vanish at $z = h$, may include only odd cosine terms. The series of integrals is integrated term by term and the undetermined coefficients evaluated by satisfying the integral equation at a sufficient number of points along the antenna.



This is a valuable and fundamentally sound method, but it must be borne in mind that the number of terms required for an accurate determination of the input current depends greatly on the nature of the difference function which is to be represented by the odd cosine series. For some lengths of the aerial, notably those near resonance with lengths $2l$ near $\lambda/2$, $3\lambda/2$, etc., both components of current $I = I'' + jI'$ (where I'' is in phase with the driving voltage and I' in phase quadrature), are represented quite accurately by only a few terms. On the other hand, when $2l$ is near λ, 2λ, etc., the situation is different. This is illustrated in the figure where the theoretical distributions of $I''(z)$ and $I'(z)$ are shown. By subtracting appropriate sine curves, difference functions that are to be represented by an odd cosine series are obtained. It is seen that the normalized difference function for $I''(z) - \sin\beta z$ should require only two or three terms for a fairly accurate representation. On the other hand, the sharp peak in $I'(z) - I'_m \sin\beta z$ at $z = 0$ makes an accurate representation by an odd cosine series difficult near $z = 0$ unless very many terms are used. Indeed, even if the curve is well fitted for all other values of z, the value at $z = 0$ is bound to be much less accurate and it is this value which gives the susceptance B.

In order to correlate this discussion with the computations of Storm and those of King and Middleton, it is advantageous to study the admittances rather than the impedances. These are listed below together with the impedances. The third-order value of King has been determined by a new approach,[1] and the second-order values have been corrected to eliminate errors originally introduced from the use of the numerically obtained values of second-order functions as calculated by Bouwkamp.

λ/2-dipole: Storm

Calculated with 2 points:
$Z = 73{\cdot}3 + j42{\cdot}9$; $Y = (10{\cdot}2 - j5{\cdot}95) \times 10^{-3}$

Calculated with 3 points:
$Z = 81{\cdot}1 + j43{\cdot}7$; $Y = (9{\cdot}56 - j5{\cdot}15) \times 10^{-3}$

Calculated with 4 points:
$Z = 80{\cdot}7 + j44{\cdot}9$; $Y = (9{\cdot}46 - j5{\cdot}26) \times 10^{-3}$

Calculated with 5 points:
$Z = 82{\cdot}7 + j45{\cdot}4$; $Y = (9{\cdot}29 - j5{\cdot}10) \times 10^{-3}$

λ/2-dipole: King

First order	$Z = 75{\cdot}0 + j36{\cdot}0$;	$Y = (10{\cdot}8 - j5{\cdot}20) \times 10^{-3}$
Second order	$Z = 80{\cdot}8 + j43{\cdot}4$;	$Y = (9{\cdot}60 - j5{\cdot}16) \times 10^{-3}$
Third order	$Z = 81{\cdot}5 + j43{\cdot}4$;	$Y = (9{\cdot}56 - j5{\cdot}09) \times 10^{-3}$

λ/2-dipole: Storm

Calculated with 3 points:
$Z = 1313 - j1391$; $Y = (0{\cdot}359 + j0{\cdot}380) \times 10^{-3}$

λ/2-dipole: King

First order	$Z = 1560 - j1130$;	$Y = (0{\cdot}420 + j0{\cdot}304) \times 10^{-3}$
Second order	$Z = 1000 - j1350$;	$Y = (0{\cdot}354 + j0{\cdot}478) \times 10^{-3}$

For the $\lambda/2$-dipole the agreement between the Storm and the King values is excellent. The same is true for the values of G for the λ-dipole. On the other hand, for the susceptance B of the λ-dipole the agreement is very poor. It would appear, that owing to the sharp rise in the difference curve for the current $I'(z)$ near zero, three terms in an odd cosine series are quite inadequate to determine $I'(0)/V^e_0 = B$.

Perhaps it is well to add that recent measurements by Hartig[2] over a wide range of values of lengths and radii of the antenna are in very good agreement with the second-order King values. The greatest per-cent error is in the conductance near resonance where the third-order solution is found to agree excellently.

It is hoped that more extensive evaluations of the impedance will be made by Mr. Storm.

Ronold W. P. King.

Cruft Laboratory,
Harvard University,
Massachusetts, U.S.A.
21st November 1952.

[1] R. King, "An Alternative Method of Solving Hallén's Integral Equation and its Application to Antennas Near Resonance," Tech. Report No. 154, Cruft Laboratory, Harvard University (July 1952), *J. appl. Phys.*, January 1953 (to be published).
[2] E. O. Hartig, "Circular Apertures and their Effects on Half-Dipole Impedances," Tech. Report No. 107, Cruft Laboratory, Harvard University (June 1950).

Mutual Radiation Resistance of Aerials and Arrays

Sir,—In a paper[1] in the November *Wireless Engineer*, Knudsen calculates the mutual resistance of two parallel aerials using the Poynting vector method in the form first used by Van der Pol. In the course of the calculation a rather complicated integral appears [equation (19)] which is not evaluated directly but deduced from the known solution to the same aerial problem by the induced e.m.f. method.

The integral is almost identical to one I encountered some fifteen years ago in the calculation of the impedance of a horizontal dipole above an infinite ground plane. Not being aware at the time of the induced e.m.f. method the integral was investigated directly, and the integration carried out by an expansion, integration of terms, and subsequent summation. The details are given in a paper[2] published in 1939 and may be compared with Knudsen's equation (20). The same expansion technique may be used for integrating Knudsen's equation (25), although the subsequent series does not seem to be summable. In the particular case of $a = b$ his equation (25) gives

$$R_{12} = \frac{\zeta}{2\pi}\left\{\frac{1}{2}\int_0^{2\pi}\frac{1\text{-}\cos x}{x}\,dx + \sum_1^{\infty}(ka)^{2n}\,\frac{(-1)^n}{2n\,(|\underline{n})^2} + \right.$$

$$\left. + \sum_1^{\infty}(ka)^{2n}\,\frac{|\underline{2n-1}}{(|\underline{n})^3}\left[\left(\frac{d}{d\omega}\right)^n\cos\sqrt{\omega}\,\right]_{\omega=\pi^2}\right\}$$

This equation refers to the self-impedance of a cylindrical dipole of radius a. The technique used in (2) of summing the second infinite series by identification with a Taylor series expansion does not seem possible here owing to the altered form of the factorials. When $a \neq b$ the coefficient $\frac{|\underline{2n-1}}{(|\underline{n})^3}$ is replaced by $\frac{1}{2}\,|\underline{n-1}\sum_0^n\frac{(b/a)^{2r}}{(|\underline{r}\,|\underline{n-r})^2}$

which can be put in the form $\frac{1}{2n\,|\underline{n}}\sum_0^n({}^nC_r)^2\,(b/a)^{2r}$.

In (2) the coefficient ${}^nC_r$ was not squared, and it can be seen readily how the summation was completed using the binomial theorem, the Taylor expansion for $\cos\sqrt{\pi^2 + k^2a^2 + k^2b^2}$ and the cosine integral formula.

Looking back on the whole development, one is struck with the extreme power of the 'induced-e.m.f. method' over the 'Poynting vector method'—the various quantities such as $\sqrt{\pi^2+k^2a^2+k^2b^2}$ appear simply as geometrical distances occurring naturally in the aerial configuration, and do not have to be synthesized 'artificially.' In conclusion, it may not be out of place to draw attention to a paper by Bechmann[3] which does not appear to be as well known as it might be. He shows in a simple way that the integration of the Poynting vector over an infinite sphere can also be conducted over a surface in the immediate neighbourhood of the aerial itself. By interpreting the magnetic field, which appears in this formalism as simply the aerial current, the usual formulae of the induced-e.m.f. procedure are obtained, thus providing an interesting link-up of the two methods.

L. Lewin.

Enfield,
Middlesex.
17th November 1952.

[1] "Mutual Radiation Resistance of Aerials and Arrays," by H. Lottrup Knudsen, *Wireless Engineer*, Nov. 1952.

[2] "Radiation Resistance of a Horizontal Dipole above Earth." by L. Lewin, *Marconi Review*, April-June 1939.

[3] "On the Calculation of Radiation Resistance of Antennas and Antenna Combinations," by R. Bechmann, *Proc. Inst. Radio Engrs*, August 1931. (The original German article is to be found in Knudsen's reference 8.)

NEW BOOKS

Examples in Electrical Calculations

Admiralty B.R.158(52). Pp. 507 + xii, with 263 illustrations. H.M. Stationery Office, York House, Kingsway, London, W.C.2. Price 17s. 6d.

This supersedes the 1933 edition, and it consists of a collection of worked examples and carefully-graded sets of exercises covering the scope of many electrical examinations in the Navy. It also covers the Preliminary and Intermediate City and Guilds Examinations and the Ordinary National Certificate in Electrical Engineering. It must be emphasized, however, that it is not a textbook but deals solely with calculations; there is always sufficient descriptive matter and illustrations to enable the student to understand the principles involved in the examples. After setting out the principles, each chapter contains a number of problems worked out in detail and then a number of exercises, of which the numerical answers are given.

The book is divided into an introduction containing tables of units, symbols, etc., and 29 chapters; it concludes with some appendices dealing with the m.k.s. system, mathematical tables, vectors, graphs, differential coefficients, standard integrals, etc.,

The first nine chapters deal with electrical circuits and measurements and the laws of Ohm, Joule and Kirchhoff; then follow chapters on magnetism, electromagnetic induction, and capacitance. Chapters 13 to 15 deal with dynamos and chapters 16 to 19 with motors and starters. Then follow chapters on alternating-current theory, polyphase systems, transformers, alternators, synchronous and induction motors, conversion of a.c. to d.c., hot-cathode, mercury-arc and metal rectifiers, and finally a chapter on illumination. In the chapter on the speed control of d.c. motors and methods of braking, several pages are devoted to the metadyne generator.

The book is excellently produced and some of the vector diagrams and curves are in two or three colours. It will provide a wealth of material both to students and teachers. In a book of this type, nomenclature and symbols are almost bound to prove troublesome. In the list of quantities and symbols on page 1 there is no mention of electromotive force or potential difference; the explanation of the former omission is presumably due to the fact that in the book E is used for e.m.f. and for electric intensity and for illumination. Although potential difference is omitted from the list, it occurs, of course, often throughout the book, but its old-established and almost universally-recognized symbol p.d. occurs rarely; it is usually treated as a displaced person and labelled D.P. On two occasions, viz. on pages 6 and 16, the alternative P.D. or D.P. is given. So far as we can discover, no explanation is given of this queer twist; one would, however, expect the Navy to follow strictly the recommendations of the British Standards Institution.

G. W. O. H.

Electronic Measurements (2nd Edition)

By F. E. Terman, Sc.D., and J. M. Pettit, Ph.D. Pp. 707 + xiii. McGraw-Hill Publishing Co., Ltd., 95 Farringdon Street, London, E.C.4. Price 72s. 6d.

This book is, in reality, a second edition of the well-known "Measurements in Radio Engineering," but has had a change of title to make it "indicative of the increased scope of the new book, which now covers measurement fundamentals in numerous fields beyond conventional radio, including television, radar and other pulsed systems, microwaves, and a diversity of techniques of value to engineers in other areas who may use electronics in their instrumentation."

The change of title is hardly an improvement, for 'electronics' now means so many things that it means nothing. When one tries to find out what the authors mean by it one is baffled. They are not using it in its original sense of describing free electrons in vacuo or in gases (that is, valves, c.r. tubes and allied devices) for the book deals with moving-coil voltmeters, Wheatstone bridges, and the measurement of L, C and R. In some cases valves are not even loosely associated with these things. They are not using electronics to mean 'radio-like devices used for non-radio purposes', for a large part of the book is pure radio. One can only conclude that 'electronics' is used because it is a fashionable word.

It is surprising, too, to find that the authors apparently consider that television, radar and microwaves are not radio! Television need not be radio, of course, but radar essentially depends upon the transmission and reception of electromagnetic waves in space. One suspects that by radio the authors mean merely sound broadcasting!

In spite of its misleading title and the general carelessness in the use of words which the authors have shown in the preface, the book itself is much less open to criticism. It covers an enormous field and contains a great deal of useful information. Perhaps it is inevitable that the treatment should be rather brief, for if it were fuller many subjects would have had to be omitted. Many will find it desirable to supplement the information given from other sources, and the authors have evidently realized this for they include frequent references to papers giving fuller details. As a result of this, the book is much more suitable for the engineer's reference library than for the student.

The book has 15 chapters:—Voltage and Current; Power; Circuit Constants of Lumped Circuits; Circuit Constants in Systems Involving Distributed Constants; Measurement of Frequency; Waveform, Phase and Time-Interval Measurements; Characteristics of Triodes, Pentodes and Similar Tubes; Amplifier Measurements; Receiver Measurements; Antennas; Radio Waves; Laboratory Oscillators; Generators of Special Waveforms; Reactance and Resistance Standards and Devices Attenuators and Signal Generators.

The treatment is to a large extent descriptive. Where formulae are needed they are quoted in the form of final results rather than derived. It would have been helpful if the authors had stated what system of units they have employed; the units for some individual equations are stated, but not for all. It is only fair to say, however, that in very many cases the proper units are either obvious or unimportant.

W. T. C.

Radio and Television Manufacturing

Pp. 60 + iv. British Institute of Management, 8 Hill St., London, W.1. Price 20s.

This is a "case study data on productivity and factory performance prepared for the Mutual Security Agency, Productivity and Technical Assistance Division, by the United States Department of Labor." It includes statistics in tabular form from 16 manufacturers of radio and television sets.

Currents and Fields in Electrical Engineering

By Professor H. E. M. Barlow. Pp. 20 with 5 illustrations. H. K. Lewis & Co., Ltd., 136 Gower St., London, W.C.1. Price 5s.

ABSTRACTS AND REFERENCES INDEX

The Index to the Abstracts and References published throughout 1952 is in course of preparation and will, it is hoped, be available in February, price 3s. 9d. (including postage). As supplies are limited, our Publishers ask us to stress the need for early application for copies. Included with the Index is a selected list of journals scanned for abstracting, with publishers' addresses.

I.E.E. MEETINGS

14th January. "Printed and Potted Electronic Circuits", by G. W. A. Dummer, M.B.E., and D. L. Johnston, B.Sc.(Eng.).

20th January. Discussion on "Measurements of Magnetic Permeability" to be opened by A. J. King, D.Sc., M.Sc.Tech.

21st January. Discussion on "Some Difficulties in the Teaching of Electrical and Mechanical Resonance" to be opened by C. T. Baldwin, M.A., and J. C. Oakden, M.A., M.Sc.Tech.

26th January. Discussion on "The Relative Merits of Harmonic and Intermodulation Measurements for Assessing Distortion in Audio Equipment" to be opened by E. W. Berth-Jones.

These meetings will be held at the Institution of Electrical Engineers, Savoy Place, London, W.C.2, and, with the exception of the meeting on 21st January, which commences at 6 o'clock, will start at 5.30.

STANDARD-FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)

Values for November 1952

Date 1952 November	Frequency deviation from nominal: parts in 10^8 — MSF 60 kc/s 1029-1130 G.M.T.	Frequency deviation from nominal: parts in 10^8 — Droitwich 200 kc/s 1030 G.M.T.	Lead of MSF impulses on GBR 1000 G.M.T. time signal in milliseconds
1	−0·5	+3	−31·9
2	−0·5	+2	−31·6
3	−0·4	+3	−32·4
4	−0·4	+4	−33·1
5	−0·4	+3	−33·4
6	−0·3	+3	N.M.
7	−0·4	+4	−30·0
8	−0·3	+4	−29·4
9	−0·2	+4	−28·4
10	−0·1	+5	−28·9
11	−0·3	+5	−28·4
12	−0·2	+5	−27·7
13	−0·2	+5	−28·5
14	−0·2	−4	−27·3
15	0·0	−5	−26·7
16	−0·1	−4	−26·9
17	−0·1	−3	−26·2
18	0·0	−3	−25·6
19	−0·2	−3	−24·8
20	−0·1	−2	−23·7
21	−0·6	−1	−24·1
22	−0·5	−2	−23·8
23	−0·5	−2	−22·6
24*	−0·5	−1	−21·7
25	−0·5	−1	−21·6
26	−0·4	−1	−20·4
27	−0·5	−1	−21·3
28	−0·4	−2	−21·9
29*	−0·4	−2	N.M.
30*	−0·4	−1	N.M.

The values are based on astronomical data available on 1st December 1952.

N.M. = Not measured.

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

ABSTRACTS and REFERENCES

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

The abstracts are classified in accordance with the Universal Decimal Classification. They are arranged within broad subject sections in the order of the U.D.C. numbers, except that notices of book reviews are placed at the ends of the sections. U.D.C. numbers marked with a dagger (†) must be regarded as provisional. The abbreviations of journal titles conform generally with the style of the World List of Scientific Periodicals. An Author and Subject Index to the abstracts is published annually; it includes a list of journals abstracted, the abbreviations of their titles and their publishers' addresses.

ACOUSTICS AND AUDIO FREQUENCIES

016 : 534 1

References to Contemporary Papers on Acoustics.—R. T. Beyer. (*J. acoust. Soc. Amer.*, Sept. 1952, Vol. 24, No. 5, pp. 548–555.) Continuation of 3298 of 1952.

534 : 061.6 2

The Physikalisch-Technische Bundesanstalt, Brunswick.—M. Grützmacher. (*Ricerca sci.*, July 1952, Vol. 22, No. 7, pp. 1333–1347.) A description with special reference to the acoustics laboratories.

534 : 727.5 (68.01) 3

The Electro-acoustics Laboratory at the School for the Deaf, Worcester, C.P.—J. P. A. Lochner & A. Semmelink. (*Trans. S. Afr. Inst. elect. Engrs*, July 1952, Vol. 43, Part 7, pp. 212–221.) General description of the laboratory and its functions, with a more detailed description of the construction of the anechoic chamber.

534.231.3 : 621.3.011.21].001.362 4

Incomplete Analogy between Electrical and Acoustical Characteristic Impedances, and Consequences relating to Echoes in Continuously Stratified Media.—G. Eckart & P. Liénard. (*Acustica*, 1952, Vol. 2, No. 4, pp. 157–161. In French.) The acoustical characteristic impedance of a medium is defined by analogy with the corresponding electrical quantity. In the case of e.m. waves, there is no internal reflection in a continuously stratified medium if the characteristic impedance remains constant. This, however, is not true for sound waves, which should in all circumstances be reflected by a stratified atmosphere.

534.232 : 538.652 5

Magnetostriction Transducer Measurements.—H. J. Round. (*Radio tech. Dig., Édn franç.*, 1952, Vol. 6, No. 3, pp. 123–133.) Adaptation in French of paper noted in 1494 of 1952, with 14 additional references.

534.24 6

Acoustic Radiation Pressure of Plane Compressional Waves at Oblique Incidence.—F. E. Borgnis. (*J. acoust. Soc. Amer.*, Sept. 1952, Vol. 24, No. 5, pp. 468–469.) Simple analysis is given for a system comprising a sound beam of finite cross-section propagated in a nonviscous medium and incident on a plane reflector. For the special case of a reflector in the form of a wedge of angle 90°, the pressure due to a beam incident symmetrically on the edge of the wedge is independent of the coefficient of reflection.

534.321.9 7

Ultrasonic Velocity, Dispersion, and Absorption in Dry, CO_2-Free Air.—C. Ener, A. F. Gabrysh & J. C. Hubbard. (*J. acoust. Soc. Amer.*, Sept. 1952, Vol. 24, No. 5, pp. 474–477.) Measurements were made at 32°C and pressures p from 0·02 to 1 atm, using frequencies f of 2 and 3 Mc/s. The variation of dispersion and absorption with f/p is discussed. Changes in velocity, absorption and internal specific heat are interpreted as the result of the slowing of energy exchange between translational and rotational states.

534.6 : 534.321.9 8

New Method for the Visualization and Measurement of Ultrasonic Fields.—G. S. Bennett. (*J. acoust. Soc. Amer.*, Sept. 1952, Vol. 24, No. 5, pp. 470–474.) Details are given of a method depending on the ability of ultrasonic vibrations to accelerate chemical reactions, particularly the starch-iodine reaction. $BaTiO_3$ disks were used as ultrasonic sources, operating in a dilute solution of iodine, and glass plates coated with starch paste were used as detectors. Photographs are shown of near-field patterns obtained.

534.76 9

Masking of Tones by White Noise as a Function of the Interaural Phases of Both Components: Part 1—500 Cycles.—L. A. Jeffress, H. C. Blodgett & B. H. Deatherage. (*J. acoust. Soc. Amer.*, Sept. 1952, Vol. 24, No. 5, pp. 523–527.)

534.78 10

On the Effect of Frequency and Amplitude Distortion on the Intelligibility of Speech in Noise.—I. Pollack. (*J. acoust. Soc. Amer.*, Sept. 1952, Vol. 24, No. 5, pp. 538–540.) A comparison of the effects of noise and of frequency limiting on the intelligibility of speech (*a*) subjected to no amplitude distortion, and (*b*) subjected to infinite peak clipping.

534.78 11

Solid Sound.—L. G. Kersta. (*Bell Lab. Rec.*, Sept. 1952, Vol. 30, No. 9, pp. 354–357.) A method is described for preparing three-dimensional models showing the variation with time of the energy spectrum representing speech sounds. Models of the spoken words 'five' and 'nine' are illustrated.

534.845.1 **12**

Room-Acoustics Investigations with Directive Transmitters and Receivers.—E. Meyer & H. G. Diestel. (*Acustica*, 1952, Vol. 2, No. 4, pp. 161–166. In German.) The material whose absorption was to be measured completely covered one wall of a four-sided room. Measurements on three different materials were made in the range 500–1 000 c/s, using directive transmitters. Results for normal incidence are independent of the wall surface on which the material is fixed and agree with values obtained by means of Kundt's tube, and also qualitatively with results obtained at grazing incidence of the sound beam.

534.845.1 **13**

Long-Tube Method for Field Determination of Sound-Absorption Coefficients.—E. Jones, S. Edelman & A. London. (*J. Res. nat. Bur. Stand.*, July 1952, Vol. 49, No. 1, pp. 17–20.) Portable equipment has been developed by means of which the method previously described by London (1846 of 1950) can be used for the nondestructive testing of acoustic materials already installed; measurements are made at 512 c/s. The method enables the appearance of the materials to be correlated with their sound absorption coefficient; variations due to differences in plasterer workmanship are discussed.

534.846.4 **14**

Acoustic Correction in the Church of St. Anthony of Padua in Vienna X.—(*Radio Tech., Vienna*, July 1952, Vol. 28, No. 7, pp. 317–318.) The church has a reverberation time of 7 sec. The arrangement of five loudspeaker arrays to obviate echo is described.

534.87(204.1) : 621.396.822 **15**

Thermal-Noise Limit in the Detection of Underwater Acoustic Signals.—Mellen. (See 106.)

621.395.616 **16**

Electrical Input Resistance of the Capacitor Microphone.—H. Grosskopf. (*Arch. elekt Übertragung*, Aug. 1952, Vol. 6, No. 8, p. 351.) Comment on 18 of 1952 (Kirschner).

621.395.623.7 **17**

The Assessment of the Transient Distortion of Loudspeakers from the Frequency Response.—E. Seemann. (*Tech. Mitt. schweiz. Telegr.-TelephVerw.*, 1st April 1952, Vol. 30, No. 4, pp. 121–127. In German.) The transient distortion (crackles, etc.) produced by different loudspeakers can be compared by an objective method making use of the 'mean-value' frequency response curve proposed by Hentsch (2898 of 1951). The basic subjective measurements and the apparatus devised for obtaining the mean-value response curve are described. Examples show good agreement between the objective and subjective assessments.

621.395.623.7 **18**

Metal-Cone Loudspeaker.—F. H. Brittain. (*Wireless World*, Nov. 1952, Vol. 58, No. 11, pp. 440–443.) An account is given of development work on a high-quality metal-cone loudspeaker. A dip in the response curve at a frequency between 2 and 3 kc/s was traced to interference between vibrations from different parts of the cone and was eliminated by inserting a rigid 'bung' in the cone cavity. A peak in the response curve at about 8 kc/s was eliminated by slotting the cone circumferentially and bending the parts adjacent to the slots.

621.395.623.73 + 681.85].001.42 **19**

Objective Testing of Pickups and Loudspeakers.—K. R. McLachlan & R. Yorke. (*J. Brit. Instn Radio Engrs*, Sept. 1952, Vol. 12, No. 9, pp. 485–496.) Analytical and experimental techniques used to assess the performance of pickups and moving-coil loudspeakers are described. Analysis by electrodynamic analogies enables a general mathematical solution to be obtained of the problem of determining performance characteristics. Apparatus used to determine steady-state and transient response is described and examples of typical tests are given, together with details of the methods used for indicating and recording the mechanical vibrations of (*a*) the various components of pickups and (*b*) loudspeaker cones. The analysis for pickups affords an explanation of all the phenomena observed experimentally. A similar general agreement has not yet been reached for loudspeakers, the work on which is still in its initial stages.

621.395.625.3 **20**

The Mechanical Properties of Various Magnetic Recording Tapes and their Influence on the Quality of the Recording.—P. H. Werner. (*Tech. Mitt. schweiz. Telegr.-TelephVerw.*, 1st May 1952, Vol. 30, No. 5, pp. 173–180. In French and German.) Extraneous modulation may occur at frequencies up to 50 c/s due to slow variations of tape speed and at frequencies between about 1 and 3 kc/s due to longitudinal oscillations of the tape. Extension/tension curves are plotted for 14 commercial tapes. A formula is given relating the elastic modulus and the frequency of the longitudinal oscillations, and these frequencies are tabulated for metre lengths of the 14 tapes. The production of noise by friction between tape and rollers is discussed, and an experimental arrangement is described for investigating this effect. An indication is given of steps to be taken in the design and construction of the recording machine to reduce this noise.

621.395.625.3 **21**

Magnetic Recording in Film Production.—N. Leevers. (*J. Brit. Instn Radio Engrs*, Aug. 1952, Vol. 12, No. 8, pp. 421–427.) Description of equipment using twin-channel magnetic tape, one channel carrying pulse signals at picture frequency, or a multiple thereof, for synchronizing with the corresponding picture film.

621.395.625.3(083.74) **22**

Standardization of Magnetic-Recording Technique.—H. Schiesser & O. Schmidbauer. (*Frequenz*, Aug. 1952, Vol. 6, No. 8, pp. 222–229.) Proposals made at the 1950 Berne conference and at the 1951 C.C.I.R. meeting at Geneva are discussed, and methods of measurement of magnetic characteristics of recording and reproducing heads, and magnetic tapes, etc., are considered.

621.395.97 + 621.397.24 **23**

Relaying the Sound and Television Signals at the South Bank Site.—H. J. Barton-Chapple. (*J. Televis. Soc.*, April/June 1952, Vol. 6, No. 10, pp. 381–384.) A general description of arrangements during the Festival of Britain; reliability was the main factor determining the choice of amplifying equipment and associated control gear. Vision signals were distributed on 61·75 Mc/s and television sound at 58·25 Mc/s except for the Telekinema, where the sound was distributed at a.f.

621.396.645.371.029.3 : 621.395.667 **24**

Negative-Feedback Tone Control.—Baxandall. (See 64.)

AERIALS AND TRANSMISSION LINES

621.392 + 621.315.212].018.44 **25**

Mathematical Theory of Laminated Transmission Lines: Part 1.—S. P. Morgan, Jr. (*Bell Syst. tech. J.*, Sept. 1952, Vol. 31, No. 5, pp. 883–949.) The theory presented by Clogston (2908 of 1951) is extended and analysis is given for both parallel-plane and coaxial cables using laminated conductors, the present paper dealing mainly with the class of structure in which the conductors are separated by a space filled with a continuous dielectric. Lines of this type are termed Clogston-1 as opposed to the Clogston-2 type in which the whole space is filled with laminated material. The problems studied include: determination of the propagation constants and the fields of the various transmission modes; the choice of optimum dimensions for the lines; calculation of the frequency dependence of attenuation due to the finite thickness of the laminae; losses caused by dielectric mismatch in Clogston-1 lines and by nonuniformity of the laminae in Clogston-2 lines; and the effects of dielectric and magnetic dissipation. See also 2988 of 1952 (Black et al.).

621.392.21 **26**

Analysis of Multiconductor Systems with Transverse Electromagnetic Waves at High Frequencies.—H. J. von Baeyer & R. Knechtli. (*Z. angew. Math. Phys.*, 15th July 1952, Vol. 3, No. 4, pp. 271–286.) The transmission-line equations for current and voltage on each conductor are derived in a generalized form involving matrix symbols. Boundary conditions are considered and the generalized equations are applied in theory of the transmission-line directional coupler and in simplified theory of the folded dipole aerial.

621.392.26 **27**

The Relative Power-Carrying Capacity of High-Frequency Waveguides.—H. M. Barlow. (*Proc. Instn elect. Engrs*, Part III, Sept. 1952, Vol. 99, No. 61, p. 323.) Discussion on 1196 of 1952.

621.392.26 : [621.315.61 + 621.315.5 **28**

Dielectric and Metal Waveguides.—H. Kaden. (*Arch. elekt. Übertragung*, Aug. 1952, Vol. 6, No. 8, pp. 319–332.) An idealized planar type of waveguide is considered, for which a system of transcendental equations is derived; these apply to dielectric waveguides if the dielectric constant of the material is assumed preponderantly real with a small imaginary component which accounts for the losses. For metal waveguides the dielectric constant is assumed to be purely imaginary and of magnitude very large compared with that of free space. The results of the analysis indicate that for the transmission of centimetre waves the dielectric hollow waveguide is better than either the metal waveguide or the Goubau surface-wave line. Quantitative results are presented in two tables, the first giving comparative figures for the properties of dielectric hollow and solid waveguides, metal waveguides, and surface-wave transmission lines, of the same outside dimensions, for free-space wavelengths from 100 to 1 cm, the second comparing the properties of dielectric hollow and solid waveguides, and surface-wave transmission lines, with the same amount of insulating material, for the same range of wavelengths. Theory of the transmission of TM and TE waves in the general type of waveguide is given, the formulae for the type here considered being derived as special cases.

621.392.26 : 621.392.43 **29**

Broad-Band Matching with a Directional Coupler.—W. C. Jakes, Jr. (*Proc. Inst. Radio Engrs*, Oct. 1952, Vol. 40, No. 10, pp. 1216–1218.) Theoretical and experimental investigation of a method of using a directional coupler to cancel standing waves. The method gives a wide-band match which is independent of the relative locations of the directional coupler and the discontinuity causing the original mismatch. Good agreement was obtained between experimental results and theoretical curves.

621.392.43 **30**

Bandwidth of Quarter-Wave Sections.—E. G. Fubini & F. H. Rockett, Jr. (*Electronics*, Oct. 1952, Vol. 25, No. 10, pp. 138–139.) Charts are given from which can be read the bandwidth over which a prescribed degree of matching is obtainable, using either single- or double-$\lambda/4$ matching sections; the matched bandwidth is considerably greater in the latter case.

621.396.67 **31**

On the Theory of Antenna Beam Shaping.—A. S. Dunbar. (*J appl. Phys.*, Aug. 1952, Vol. 23, No. 8, pp. 847–853.) A theoretical investigation is made of the diffraction pattern produced by radiation from a given aperture, (*a*) for controlled variation of the amplitude distribution over the aperture with given uniform phase, and (*b*) for controlled variation of phase over the aperture with given uniform amplitude. The method used by Chu (*M.I.T. Research Laboratory of Electronics, Technical Report* 40, 1947) for calculating cylindrical reflectors is adapted to derive a general formulation for an amplitude distribution on a curved surface. The theory is applied to the design of progressive-phase aerials. Experimental results obtained with channel-guide and slot-array aerials support the theory.

621.396.67 **32**

Mutual Radiation Resistance of Aerials and Arrays.—H. L. Knudsen. (*Wireless Engr*, Nov. 1952, Vol. 29, No. 350, pp. 301–305.) A simple expression for the mutual resistance of two aerial arrays with known characteristics is derived on the basis of the Poynting-vector method. The application of the expression is demonstrated by the examples of (*a*) two parallel linear aerials, and (*b*) two concentric ring arrays.

621.396.67 : 621.316.54 **33**

Aerial Exchange.—(*Wireless World*, Nov. 1952, Vol. 58, No. 11, p. 444.) Description of equipment installed at one of the Admiralty's high-power transmitting stations for connecting any one of 10 transmitters to any one of 20 aerial systems. The switching system is partly motorized and comprises a semicylindrical structure of radius 14 ft and height about 16 ft, with travelling carriages, 11 moving horizontally and 20 others vertically.

621.396.67 : 621.392 : 621.397.61 **34**

Suspended Television Feeder.—(*Wireless World*, Nov. 1952, Vol. 58, No. 11, pp. 473–474.) Description of aerial feeder arrangements at the B.B.C. stations at Kirk O'Shotts and Wenvoe. See also 244 below.

621.396.67 : 621.396.823 **35**

Loop Aerial Reception.—G. Bramslev. (*Wireless World*, Nov. 1952, Vol. 58, No. 11, pp. 469–472.) An indication is given of the advantages of using a loop aerial for long-wave reception with an ordinary broadcast receiver, from the point of view of reducing interference from electrical apparatus. The loop is less responsive than the ordinary capacitive aerial to the locally produced electric fields mainly responsible for the interference. Design details are given for the loop and the coupling circuit to the receiver.

621.396.67 : 621.396.826 **36**

Electromagnetic Back-Scattering from Cylindrical Wires.—C. T. Tai. (*J. appl. Phys.*, Aug. 1952, Vol. 23, No. 8, pp. 909–916.) The problem dealt with previously by Van Vleck et al. (3035 of 1947) is here investigated using the variational method of Schwinger (*Phys. Rev.*, 1947, Vol. 72, p. 742). Different trial functions are used to determine the numerical values of the back-scattering cross section for broadside incidence. The boundary conditions for the currents at the ends of the wires are examined. See also 2716 of 1952 (Dike & King).

621.396.67 : [621.397.5 + 621.396.97.029.6 **37**

Combined Transmitting Aerials for Television and U.S.W. Broadcasting.—W. Berndt. (*Telefunken Ztg*, Aug. 1952, Vol. 25, No. 96, pp. 158–168.) The horizontal and vertical radiation diagrams of various simple unit arrangements of dipoles or slot aerials are discussed, and several aerial systems consisting of vertical assemblies of such units, with which television and sound signals can be transmitted simultaneously, are described, with illustrations of N.W.D.R. aerial systems at Witzleben, Berlin, and at Bielstein, Teutoburger Wald.

621.396.67.013.24 **38**

Impulse Electromagnetic Fields.—R. Kitai. (*Trans. S. Afr. Inst. elect. Engrs*, July 1952, Vol. 43, Part 7, pp. 200–211.) Expressions are developed for the fields of the Hertzian dipole and the magnetic dipole in terms of dipole moments. The expressions hold when the derivatives of the dipole moments are continuous for all values of time. By assuming that the moment M obeys the law $M \propto \tanh(kt)$, an insight into the nature of impulsive fields is obtained by considering the condition $k \rightarrow \infty$, when the function becomes a step function. General expressions for impulse fields are derived and the properties of such fields are illustrated by a worked-out example. Static, induction and radiation fields are found to have different shapes.

621.396.676 **39**

The Magnetic Dipole Antenna Immersed in a Conducting Medium.—J. R. Wait. (*Proc. Inst. Radio Engrs*, Oct. 1952, Vol. 40, No. 10, pp. 1244–1245.) Explicit expressions for the fields are derived for a magnetic dipole at the centre of a spherical insulating cavity in a conducting medium such as sea water. An expression for the total power radiated is given for the case when all displacement currents in the conducting medium are negligible.

621.396.677 **40**

Simultaneous Radiation of Odd and Even Patterns by a Linear Array.—C. B. Watts, Jr. (*Proc. Inst. Radio Engrs*, Oct. 1952, Vol. 40, No. 10, pp. 1236–1239.) A method is described for obtaining odd and even patterns from a broadside array, both patterns being free of minor lobes. An experimental array consisted of 18 uniformly spaced vertical slots in the narrower face of a horizontal rectangular waveguide about 105 ft long and of cross-section 41 × 78 in. The slots were end-loaded to bring them near $\lambda/2$ resonance for the operating frequency of 109·1 Mc/s. A hybrid junction was used to feed the signals for the odd and even patterns. Performance is limited to a comparatively narrow band of frequencies. Application is in connection with runway localizers for instrument landing.

621.396.677.012.12† **41**

Analysis of Microwave-Antenna Side-Lobes.—N. I. Korman, E. B. Herman & J. R. Ford. (*RCA Rev.*, Sept. 1952, Vol. 13, No. 3, pp. 323–334.) A simple method, based on Wheeler's theory of 'paired echoes' (3642 of 1939), is described which allows manufacturing tolerances to be expressed in terms of the side-lobe level for the majority of large microwave reflectors. The method is also useful for prediction of the radiation pattern of a reflector whose mechanical deviations from the perfect shape are known.

621.396.679 **42**

Design of Optimum Buried-Conductor R.F. Ground System.—F. R. Abbott. (*Proc. Inst. Radio Engrs*, Oct. 1952, Vol. 40, No. 10, p. 1160.) Correction to paper noted in 2729 of 1952.

CIRCUITS AND CIRCUIT ELEMENTS

621.3.012.2 **43**

Universal Circle Diagram for Certain Radio Systems.—J. Coulon. (*C. R. Acad. Sci., Paris*, 22nd Sept. 1952, Vol. 235, No. 12, pp. 608–609.) When the Q of a system (e.g. a Lecher line) is nearly constant over the pass band, circle diagrams similar to those previously obtained for quartz crystals (3160 of 1952) can be established.

621.314.7 **44**

Transistors: Part 3.—J. Malsch & H. Beneking. (*Arch. elekt. Übertragung*, Aug. 1952, Vol. 6, No. 8, pp. 333–346.) Discussion of (*a*) transistor equivalent circuits, (*b*) analogies between transistors and electronic valves, (*c*) practical transistor circuits, (*d*) noise, (*e*) operational frequency limits. Part 2: 1643 of 1952 (Malsch).

621.314.7 : 621.396.615 **45**

Transistor Oscillators.—E. A. Oser, R. O. Endres & R. P. Moore, Jr. (*RCA Rev.*, Sept. 1952, Vol. 13, No. 3, pp. 369–385.) Detailed discussion of various transistor circuits, including different types of sine-wave generator, relaxation oscillators, and a combination of the two giving self-quenching oscillations or stabilized frequency division.

621.316.726.029.64 **46**

Frequency Stabilization in the Microwave Range.—B. Koch. (*Arch. tech. Messen*, July & Sept. 1952, Nos. 198 & 200, pp. 155–158 & 203–204.) Descriptive review of different methods, including those using frequency-discriminator circuits and spectral-line systems.

621.316.86 **47**

Production Control of Printed Resistors.—W. H. Hannahs & J. W. Eng. (*Electronics*, Oct. 1952, Vol. 25, No. 10, pp. 106–109.) Factors affecting the reproducibility of resistors produced by the silk-screen process are discussed; those requiring especially careful control are carbon concentration, squeegee speed, temperature of application, curing schedule and application of protective coating.

621.318.57 **48**

Electronic Switching Elements for Communications Engineering.—K. Steinbuch. (*Fernmeldetech. Z.*, Aug. 1952, Vol. 5, No. 8, pp. 349–356.) A review of the characteristics of available glow-discharge tubes, thermistors, thyratrons, transistors, and multi-electrode counter tubes, with discussion of their possible applications in telephony circuits.

621.318.57 : 621.387.032.212 **49**

New Trigger Circuits for use with Cold Cathode Counting Tubes.—J. L. W. Churchill. (*J. Brit. Instn Radio Engrs*, Sept. 1952, Vol. 12, No. 9, pp. 497–504.) A description is given of several trigger circuits devised for use particularly with dekatron tubes [2066 of 1950 (Bacon & Pollard)]. The circuits can be used in the construction of a scaling unit which can subtract as well as add. The counting losses of such a unit when used on random pulses are considered.

621.318.572 : 621.387.4 **50**

High-Speed Counter uses Ternary Notation.—R. Weissman. (*Electronics*, Oct. 1952, Vol. 25, No. 10, pp. 118–121.) The basis of the counter is a flip-flop circuit modified to have three stable states by insertion of a diode coupling circuit between the cathodes. A nine-stage circuit is described, operating reliably up to counts of 175 000 per sec. The indicating system may use either one or two lights per stage.

621.319.4 **51**

R.F. Characteristics of Capacitors.—T. E. Clarke. (*Wireless World*, Nov. 1952, Vol. 58, No. 11, pp. 457–458.) Comment on 2740 of 1952 (Davidson).

621.319.47 **52**

Development of Vacuum Capacitors.—S. J. Borgars. (*Proc. Instn elect. Engrs*, Part III, Sept. 1952, Vol. 99, No. 61, pp. 307–315.) Vacuum capacitors are particularly suitable for h.f. h.v. operation in airborne radio equipment. An account is given of the development of two types, one having a capacitance of 50 pF within ±5% at a peak working voltage of 6 kV, the corresponding figures for the other being 100 pF and 8·5 kV. Electrical and mechanical test methods are outlined.

621.392.4 : 517.54 **53**

Resonance Characteristics by Conformal Mapping.—P. M. Honnell & R. E. Horn. (*Proc. Inst. Radio Engrs*, Oct. 1952, Vol. 40, No. 10, pp. 1211–1215.) The expression $f(\lambda) = a\lambda + b + c/\lambda$, where $\lambda = \sigma + j\omega$, is termed the 'resonance function'. It represents the impedance of the series-connected LRS circuit ($S = 1/C$) or the admittance of the parallel-connected $CG\Gamma$ circuit ($\Gamma = 1/L$). Conformal mapping of the λ-plane on to the $f(\lambda)$-plane gives a figure which illustrates the meaning of the resonance function for complex frequencies. Typical examples illustrate the application of the figure to the parallel $CG\Gamma$ circuit and to one of its intrinsic generalizations. The mapping of the reciprocal functions $1/f(\lambda)$ and $1/\{1 + f(\lambda)\}$ is also considered.

621.392.5 : 621.396.645 **54**

Networks with Maximally Flat Delay.—W. E. Thomson. (*Wireless Engr*, Nov. 1952, Vol. 29, No. 350, p. 309.) Corrections to paper noted in 3375 of 1952.

621.392.5.018.7 **55**

Waveform Computations by the Time-Series Method.—N. W. Lewis. (*Proc. Instn elect. Engrs*, Part III, Sept. 1952, Vol. 99, No. 61, pp. 294–306.) The time-series method greatly reduces the work of computation in the solution of waveform or transient-response problems of cascade-connected linear quadripoles. Practical procedures for dealing with time series are described and illustrated by numerical calculations relating to waveform tests on a 100-mile coaxial-cable television link with a nominal upper cut-off frequency of 3 Mc/s.

621.392.52 **56**

The Numerical Calculation of Filter Circuits with Generalized Parameters, using Modern Theory with Special Attention to Cauer's Work.—V. Fetzer. (*Arch. elekt. Übertragung*, Aug. 1952, Vol. 6, No. 8, pp. 350–351.) Corrections to paper abstracted in 1545 of 1952.

621.392.52 **57**

The Transmission Range of Two-Circuit Band-Pass Filters, particularly for Large Bandwidths.—H. Meinke. (*Fernmeldetech. Z.*, Aug. 1952, Vol. 5, No. 8, pp. 362–371.) Graphical methods of calculation are developed which give results with accuracy adequate for practical purposes. The methods are applicable to unsymmetrical and wide-band filters, and also to narrow-band symmetrical filters.

621.392.52 **58**

Synthesis of Narrow-Band Direct-Coupled [waveguide] **Filters.**—H. J. Riblet. (*Proc. Inst. Radio Engrs*, Oct. 1952, Vol. 40, No. 10, pp. 1219–1223.) A general synthesis procedure is described that is based on an approximate first-order equivalence between direct-coupled and λ/4-coupled filters. The transmission characteristics computed for a 6-cavity filter with an overall Q of 37·6 are in excellent agreement with measurements.

621.392.52 **59**

A Frequency-Eliminating Transconductance Bridge.—W. C. Michels & R. C. Barbera. (*Rev. sci. Instrum.*, June 1952, Vol. 23, No. 6, pp. 293–295.) Description of a bridge circuit which completely eliminates a single frequency component and also partially suppresses a band of frequencies whose width can be varied within wide limits by suitable choice of circuit parameters. Applications to ripple elimination and to band-elimination amplifiers are discussed.

621.392.52 : 621.396.621 : 621.396.822 **60**

Optimum Filters for the Detection of Signals in Noise.—Zadeh & Ragazzini. (See 205.)

621.392.52 : 621.396.621 : 621.396.822 **61**

The Detection of a Sine Wave in the Presence of Noise by the Use of a Nonlinear Filter.—Slattery. (See 206.)

621.392.6 **62**

A General Network Theorem, with Applications.—B. D. H. Tellegen. (*Philips Res. Rep.*, Aug. 1952, Vol. 7, No. 4, pp. 259–269.) It is proved that, in a network configuration with branch currents i satisfying the node equations and branch voltages v satisfying the mesh equations, $\Sigma\, iv$, summed over all branches, is zero. Using this result it is possible to prove the energy theorem and the reciprocity relation of networks, and to show that if arbitrarily varying voltages are applied to a $2n$-pole network at rest, the difference between the electric and the magnetic energy at any instant depends only on the admittance matrix of the $2n$-pole network and not on the particular network used to realize this matrix.

621.394/.396].6 : 003.63 **63**

The Utility Factor in Circuit Diagrams.—C. E. Williams. (*Proc. Instn Radio Engrs, Aust.*, Sept. 1952, Vol. 13, No. 9, pp. 345–349.) The importance is stressed of laying out circuit diagrams so that the working of the circuit can be easily and clearly grasped. The common type of drawing-office diagram may look neat, but usually requires to be radically rearranged to indicate circuit functions. It is suggested that circuit diagrams should be drawn on the same general plan as block diagrams, a standard lay-out being used for common sub-circuits, and only lines carrying signals being included.

621.395.667 : 621.396.645.371.029.3 **64**

Negative-Feedback Tone Control.—P. J. Baxandall. (*Wireless World*, Oct. & Nov. 1952, Vol. 58, Nos. 10 & 11, pp. 402–405 & 444.) The circuit described provides independent control of bass and treble response without switching. The RC arrangement on each side of a bass-control potentiometer P_1 is made symmetrical so that it may be combined in the feedback line with a treble-control circuit, the potentiometer of which has an earthed centre-tap. At medium and high frequencies P_1 is effectively shorted. With both potentiometers at a middle setting, response is level.

621.396.611.1 **65**

Parallel-Tuned Circuit Periodically Switched to a Direct-Current Source.—L. J. Giacoletto. (*RCA Rev.*, Sept. 1952, Vol. 13, No. 3, pp. 386–416.) Circuit phenomena with the switch (*a*) closed, (*b*) open, are analysed in turn, linear solutions being obtained. The complete solution is then obtained by matching the boundary conditions for the two cases. This solution includes sinusoidal, sawtooth, and complex waveforms, dependent on circuit parameters and switching period. Elimination of dissipative circuit elements simplifies the general solution, with resulting clarification of the operation of the circuit. Tests with a mechanically switched circuit verified the theory, the entire gamut of waveforms being obtained by variation of circuit parameters and switching period.

621.396.611.3 : 621.392.26 **66**

On the Scattering Matrix of Symmetrical Waveguide Junctions.—A. E. Pannenborg. (*Philips Res. Rep.*, April, June & Aug. 1952, Vol. 7, Nos. 2–4, pp. 131–157, 169–188 & 270–302.) Fundamental properties of the scattering matrix are derived. Tomonaga's theory for lossless resonant structures is extended to include the effect of losses. The structural symmetry of microwave circuits is discussed, junctions consisting of two parallel sections of rectangular waveguide with one side common being particularly considered. The theory of directional couplers with such symmetry is developed. An attenuator and standard matching transformer, each having a directional coupler as a basic unit, are described. Other forms of resonant coupling are discussed.

621.396.611.4 **67**

Natural Electromagnetic Oscillations in a Rectangular Cavity with Walls of Finite Conductivity.—R. Müller & E. Ruch. (*Z. angew. Phys.*, July 1952, Vol. 4, No. 7, pp. 254–258.) In the limiting case of infinite wall conductivity, the TE and TM self-oscillations have the same frequency. This does not hold for finite wall conductivity, in which case linear combinations of the TE and TM oscillations, with the same frequency, exist. These combinations, termed 'matched' oscillations, represent to a first approximation the natural oscillations of the actual resonator. The determination of the characteristics of these matched oscillations is based on the solution of a simple geometrical problem, from which the damping and mistuning are calculated.

621.396.615 **68**

Oscillator Systems including Elements having Inertia.—N. Minorsky. (*C. R. Acad. Sci., Paris*, 22nd Sept. 1952, Vol. 235, No. 12, pp. 604–605.) Analysis is given for oscillator circuits, both with and without valves, including elements whose resistance varies with temperature.

621.396.645 **69**

Amplification and Bandwidth of Video Amplifiers.—F. J. Tischer. (*Arch. elekt. Übertragung*, Aug. 1952, Vol. 6, No. 8, pp. 309–315.) The characteristics of the ideal amplifier with constant amplification and constant phase transit time throughout the pass band are briefly reviewed, and the principal types of amplifier are analysed, with particular reference to amplification and bandwidth. The theoretical maximum values of these parameters are determined for the various types and compared with the values obtained in practice and with one another. Arrangements using distributed amplification and having very wide pass bands are considered briefly.

621.396.645 : 621.315.61 **70**

A Mathematical Analysis of a Dielectric Amplifier.—L. A. Pipes. (*J. appl. Phys.*, Aug. 1952, Vol. 23, No. 8, pp. 818–824.) The fundamental principles of operation of dielectric amplifiers are outlined and a basic circuit with resistive load is analysed in detail. It is assumed that the hysteresis curve of the dielectric is represented by a hyperbolic-sine function. Expressions are calculated for the steady-state input and output currents, and the time constant of the amplifier is estimated from consideration of the transient response.

621.396.645.35 : 621.317.3 **71**

High-Gain D.C. Amplifiers.—Kandiah & Brown. (See 172.)

621.396.645.35 : 621.318.435.3 **72**

The Design of a Practical D.C. Amplifier based on the Second-Harmonic Type of Magnetic Modulator.—S. W. Noble & P. J. Baxandall. (*Proc. Instn elect. Engrs*, Part II, Aug. 1952, Vol. 99, No. 70, pp. 327–344. Discussion, pp. 344–348.) Description of the development of an amplifier based on the work of Williams & Noble (152 of 1951), including details of (*a*) a 1·5-kc/s oscillator with second-harmonic distortion $<0{\cdot}0005\%$, (*b*) a low-pass filter to prevent power from the oscillator reaching the d.c. source, (*c*) a phase-sensitive rectifier, (*d*) switching and overall d.c. negative-feedback arrangements.

621.396.645.37 : 621.387.424 **73**

A Circuit for the Limitation of Discharge in G-M Counters.—W. C. Porter & W. E. Ramsey. (*J. Franklin Inst.*, Aug. 1952, Vol. 254, No. 2, pp. 153–163.) A two-valve feedback amplifier is used to limit the discharge to a small part of the total length of the wire, full sensitivity being restored in about 1 μs.

GENERAL PHYSICS

537.224 **74**

Electrets.—J. Euler. (*Z. Ver. dtsch. Ing.*, 1st June 1952, Vol. 94, No. 16, pp. 481–483.) General discussion of the properties of electrets. Applications in electrometers and in apparatus for detecting the presence of radioactive radiation are noted.

537.291 + 538.691 **75**

The Motion of Charged Particles in the Magnetic Field of a Linear Current, and the Electric Field of a Cylindrical Capacitor.—V. M. Kel'man & I. V. Rodnikova. (*Zh. eksp. teor. Fiz.*, Dec. 1951, Vol. 21, No. 12, pp. 1364–1369.) The particle trajectories for such a system are determined. Under certain conditions the system may be used for focusing beams of charged particles.

537.291 + 538.691] : 537.122 **76**

An Integrable Case of Electron Motion in Electric and Magnetic Field.—H. Poritsky & R. P. Jerrard. (*J. appl. Phys.*, Aug. 1952, Vol. 23, No. 8, pp. 928–930.) The case is considered of a two-dimensional electric field, with the potential given by $V = A + B\ (x^2 - y^2)/2$, applied together with a uniform magnetic field in the z direction. The path of an electron is an ellipse whose centre moves in a hyperbola: the electron may drift into regions of higher or lower potential.

537.291 + 538.691] : 537.122 **77**

Calculation of Plane Electron Trajectories in Particular Electric and Magnetic Fields by means of Complex Vector Loci.—H. Kleinwächter. (*Arch. elekt. Übertragung*, Aug. 1952, Vol. 6, No. 8, pp. 315–318.) Further examples of the method previously described (4449 of 1939) include (*a*) trajectories in a time-variable homogeneous magnetic field, (*b*) circular paths in a suitable e.m. field, as for the betatron.

537.311.33 **78**

A Note on the Theory of Semiconductors.—P. T. Landsberg. (*Proc. phys. Soc.*, 1st Aug. 1952, Vol. 65, No. 392A, pp. 604–608.) Two different theories are obtained, depending on whether the free energy due to the spin of electrons in impurity centres is or is not taken into account. The second of these alternatives is essentially that adopted by Wilson (*Proc. roy. Soc. A*, 1931, Vol. 133, p. 458 & Vol. 134, p. 277), and has been used almost universally, though the first, which is that adopted by Mott & Gurney (*Electronic Processes in Ionic Crystals*, 2nd edn 1948, p. 157), can be applied to degenerate semiconductors and is to be preferred.

537.311.4 : 621.396.822 **79**

A Theory of Contact Noise.—R. L. Petritz. (*Phys. Rev.*, 1st Aug. 1952, Vol. 87, No. 3, pp. 535–536.) The outlines of a theory of contact noise are presented, based on the idea that it is due to temperature fluctuations in the neighbourhood of the contact. Detailed theory is being prepared for publication.

537.311.62 + 538.52 **80**

The Apparent High-Frequency Resistance of a Conducting Layer of Finite Width parallel to an Infinite Plane Conductor. Opposing Field and Induced Currents.—A. Colombani & M. Gourceaux. (*C. R. Acad. Sci., Paris*, 22nd Sept. 1952, Vol. 235, No. 12, pp. 605–608.) Formulae are derived for the case where the conducting strip comprises a winding with a number of separate turns.

537.311.62 + 538.52 **81**

Currents Induced in a Plane Conductor of Great Width by a Plane Strip of Small Width carrying a High-Frequency Current. Approximate Formulae giving the Variations of Resistance and Self-Inductance of the Inductor.—M. Gourceaux & A. Colombani. (*C. R. Acad. Sci., Paris*, 29th Sept. 1952, Vol. 235, No. 13, pp. 650–652.) Continuation of analysis noted in 80 above.

537.323 : 546.28 **82**

Thermoelectric Measurements on *p*-Type Silicon.—J. Savornin & F. Savornin. (*C. R. Acad. Sci., Paris*, 18th Aug. 1952, Vol. 235, No. 7, pp. 465–467.) The thermo-electric power, positive with respect to Cu, is about 700 μV/1°C for material of purity 99·85%, falling to about 550 μV/1°C for 99·4% purity and rising again to 590 μV/1°C for 98% purity.

537.523/.527].029.6 **83**

High-Frequency Electrical Breakdown of Gases.—W. P. Allis & S. C. Brown. (*Phys. Rev.*, 1st Aug. 1952, Vol. 87, No. 3, pp. 419–424.) Theory is presented which is applicable to any gas over a wide range of pressure. Experimental results for H are in agreement with the theory.

537.523.4 : 538.639 **84**

Sparking Potentials in a Transverse Magnetic Field.—J. M. Somerville. (*Proc. phys. Soc.*, 1st Aug. 1952, Vol. 65, No. 392B, pp. 620–629.) By taking account of the distribution of electron free paths, and of electron re-capture by the cathode, a theory is developed which is in better agreement with observation than that of Valle (*Nuovo Cim.*, 1950, Vol. 7, p. 174).

537.525.72 : 538.63 : 621.396.822 **85**

A Resonance Phenomenon in Electrodeless H.F. Gas Discharges with Superposed Magnetic Field.—A. Lindberg, H. Neuert & H. Weidner. (*Naturwissenschaften*, Aug. 1952, Vol. 39, No. 16, pp. 374–375.) Experiments in atmospheres of H_2 and Ar are described which show that the resonance effect previously noted [614 of 1951 (Koch & Neuert)] is associated with a sharp drop between two peaks in the intensity of the emitted light; it occurs always at the same value of the field strength.

537.71 **86**

Dimensions and Units.—M. Berger. (*C. R. Acad. Sci., Paris*, 20th Oct. 1952, Vol. 235, No. 16, pp. 872–874.) By considering the expression for power in terms of the dimensions M, L and T on the one hand and in terms of current and voltage on the other hand, it follows that quantity of electricity may be regarded as having the dimensions L^3. Using this result, all the other electrical magnitudes can be expressed in terms of M, L and T with integral indices. The separate question of the best system of units is discussed briefly; the M.K.S. system fits well with the above dimensional system.

538.2 **87**

Some Post-War Developments in Magnetism.—L. F. Bates. (*Proc. phys. Soc.*, 1st Aug. 1952, Vol. 65, No. 392A, pp. 577–594.) Presidential address to the Physical Society, May 1952. The extension of knowledge of the ferromagnetic domain through the work of Néel and others and the systematic use of the Bitter powder-figure techniques is surveyed, and the main resonance phenomena and experiments on the diffraction of neutrons by antiferromagnetic crystals are described.

538.566 : 535.42 **88**

Diffraction of Electromagnetic Waves by a Half-Plane.—B. N. Harden. (*Proc. Instn elect. Engrs*, Part III, Sept. 1952, Vol. 99, No. 61, pp. 229–235.) The observed intensity and phase values of the field near the edge of a thin high-conductivity sheet are compared with Sommerfeld's rigorous solution for an infinitely thin perfectly conducting half-plane, good agreement being obtained. Results for sheets of different thicknesses and conductivities show the importance of these factors in determining the field near the diffracting edge.

538.566 : 537.562 **89**

Experimental Demonstration in the Laboratory of the Existence of Magneto-hydrodynamic Waves in Ionized Helium.—W. H. Bostick & M. A. Levine. (*Phys. Rev.*, 15th Aug. 1952, Vol. 87, No. 4, p. 671.)

538.566.2 **90**

Reciprocity of the Transmissive Properties of Any Multiple Layer for Electromagnetic Waves.—C. v. Fragstein. (*Optik, Stuttgart*, 1952, Vol. 9, No. 8, pp. 337–359.) A full analytical proof is given that the transmission of a perpendicularly incident plane e.m. wave through a multiple layer consisting of any number of plane-parallel absorbing or nonabsorbing components is the same in both directions if the first and last media have the same absorption index, i.e. if $\chi_1 = \chi_n$. If χ_1 and χ_n are different, the ratio of the transmission factors in the two directions is $(1 + \chi_1^2) / (1 + \chi_n^2)$. Treatment is simplified by applying quadripole theory, but this does not give a closed expression for the overall transmission factor. The validity of the Kirchhoff inversion principle is illustrated.

538.569.4.029.65 : 546.21 **91**

Line-Breadths of the Microwave Spectrum of Oxygen.—R. S. Anderson, W. V. Smith & W. Gordy. (*Phys. Rev.*, 15th Aug. 1952, Vol. 87, No. 4, pp. 561–568.) The microwave absorption band for O_2 consists mainly of 25 lines around the wavelength 5 mm, with spacings of only a few hundred Mc/s. The intensity of the band depends on the variation of line widths with pressure; this variation is expressed in terms of the 'line-breadth parameter', defined as the half-width, measured at half intensity, of an absorption line at a pressure of 1 atm. Determinations of this parameter made with a Zeeman modulation spectrograph are reported and discussed, and an interpretation of the self-broadening collision process is developed. Measurements are also reported for broadening due to collisions with foreign molecules.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.72 + 523.85] : 621.396.822 **92**

Radio-Astronomy.—M. Ryle & J. A. Ratcliffe. (*Endeavour*, July 1952, Vol. 11, No. 43, pp. 117–125.)

A general review of the subject, with a description of methods of measurement and discussion of the theory of galactic and solar r.f. radiation.

523.72 : 621.396.822.029.65 **93**

Solar Outbursts at 8·5-mm Wavelength.—J. P. Hagen & N. Hepburn. (*Nature, Lond.*, 9th Aug. 1952, Vol. 170, No. 4319, pp. 244–245.) Report of results obtained at the Naval Research Laboratory, Washington, D.C., with an aerial having an aperture of 24 in. and beam width of 1·1°, and correlation with observations at 3-cm wavelength and with the occurrence of solar flares. Typical records are reproduced. The 8·5-mm bursts are in general of smaller amplitude and of much shorter duration than the bursts observed on greater wavelengths.

523.78 : 523.72 **94**

Radio Observations of the Solar Eclipses of September 1, 1951, and February 25, 1952.—J. F. Denisse, E. J. Blum & J. L. Steinberg. (*Nature, Lond.*, 2nd Aug. 1952, Vol. 170, No. 4318, pp. 191–192.) Preliminary report of results which were confirmed during the February eclipse by observation at Marcoussis, near Paris, and at Dakar, French West Africa. See also 1282 of May (Bosson et al.).

523.841.11 : 621.396.822 **95**

Radio-Frequency Radiation from Tycho Brahe's Supernova (A.D. 1572).—R. H. Brown & C. Hazard. (*Nature, Lond.*, 30th Aug. 1952, Vol. 170, No. 4322, pp. 364–365.) Observations at a frequency of 158·5 Mc/s revealed a radio source whose coordinates agree closely with those of the supernova of 1572. Data for the supernovae of A.D. 1054 and 1572 are compared.

523.85 : 621.396.822 **96**

Line Emission from Interstellar Material in the Radio-Frequency Range.—R. Lüst. (*Naturwissenschaften*, Aug. 1952, Vol. 39, No. 16, pp. 372–374.) Historical review of published results, including particular reference to the hydrogen emission line at 1·42 kMc/s.

523.854 : 621.396.822] : 523.165 **97**

On the Possible Relation of Galactic Radio Noise to Cosmic Rays.—G. W. Hutchinson. (*Phil. Mag.*, Aug. 1952, Vol. 43, No. 343, pp. 847–852.) The possibility is considered that cosmic rays are accelerated in regions of the galaxy with intermediate particle density ($\sim 10^9/cm^3$). A conservative estimate of the magnetic fields in such regions would lead to a radio-noise flux of the observed order of magnitude; the observed spectrum could easily be produced.

523.99 : 523.8 : 621.396.822 : 523.755 **98**

Occultation of a Radio Star by the Solar Corona.—K. E. Machin & F. G. Smith. (*Nature, Lond.*, 23rd Aug. 1952, Vol. 170, No. 4321, pp. 319–320.) Interferometers of high resolving power were used to reduce the amplitude of the record from the undisturbed sun during observations of the radio star in Taurus as it passed near the sun's southern limb. A decrease of amplitude occurred on both 38 Mc/s and 81·5 Mc/s when the angular separation of star and sun was as great as ten times the angular radius of the visible disk, the amplitude decrease being greatest when the angular separation approached its minimum value. The results will be discussed in a later paper. See also 1281 of 1952.

537.226.2 : [546.212 + 551.311.234.5 **99**

On the Dielectric Constant of the Water in Wet Clay.—L. S. Palmer. (*Proc. phys. Soc.*, 1st Sept. 1952, Vol. 65, No. 393B, pp. 674–678.) An explanation is put forward of the variation with moisture content of the permittivity of soil samples previously observed by Cownie & Palmer (2793 of 1952). It is suggested that relatively dry clay consists of closely packed water-coated particles in an air matrix, while relatively wet clay consists of particles uniformly distributed in a water matrix, the effective permittivity of the water increasing from the value for 'bound' water (about 3) to that for 'free' water (about 80) as the percentage of water is increased.

550.385 : 535.13 **100**

Maxwell's Equations for Three-Layered Media and their Application to the Theory of the Movement of a Magnetic Storm.—M. Matschinski. (*Rev. sci., Paris*, March/April 1952, Vol. 90, No. 3316, pp. 91–103.) Solutions of the characteristic equations for ideal and for real three-layered media are given. Application is made to discussion of the characteristics of the marginal layers and calculation of the velocity of propagation of the elements of a magnetic storm. Other possible applications of the theory given are noted.

551.510.535 **101**

Ionospheric Disturbance of 23rd–28th February 1952.—H. Siedentopf & A. Behr. (*Naturwissenschaften*, Aug. 1952, Vol. 39, No. 16, pp. 377–378.) A short account of observed variations of night-sky brightness, F_2-layer limiting frequency, radiocommunication with North America, and radiosonde measurements of air temperature. A solar origin for the disturbance cannot be assigned with certainty.

551.510.535 **102**

A Self-Consistent Calculation of the Dissociation of Oxygen in the Upper Atmosphere.—H. E. Moses & Ta-You Wu. (*Phys. Rev.*, 15th Aug. 1952, Vol. 87, No. 4, pp. 628–632.) A model is considered similar to that previously proposed (129 of 1952) but with the requirement of radiative energy balance abandoned and with a particular temperature distribution assumed. Calculations indicate a much narrower dissociation region, occurring at a lower altitude.

551.510.535 : 621.396.11 **103**

The B.B.C. Ionospheric-Storm-Warning System.—T. W. Bennington & L. J. Prechner. (*B.B.C. Quart.*, Summer 1952, Vol. 7, No. 2, pp. 107–119.) The various data on which the B.B.C. warning system is based are noted and a statistical review is presented of the correlation between actual and forecast ionospheric conditions in the period 1947–1950. The accuracy of the forecasts during this period was of the order of 60–70%.

551.594.11 : 523.78 **104**

Observations of the Electric Field of the Atmosphere at Khartoum during the Total Solar Eclipse of 25th February 1952.—A. Dauvillier. (*C. R. Acad. Sci., Paris*, 20th Oct. 1952, Vol. 235, No. 16, pp. 852–854.) No effect due to the eclipse was observed in measurements of the electric field near the ground. This result supports that obtained by Chauveau in 1912 but disagrees with more recent observations, e.g. those made by Sucksdorff in Finland (2561 of 1946).

551.594.6 : 621.396.65 **105**

Correlation between the Mean Level of Atmospherics and the Degree of Intelligibility in a Kilometre-Wave Radio Link.—F. Carbenay. (*C. R. Acad. Sci., Paris*, 11th Aug. & 29th Sept. 1952, Vol. 235, Nos. 6 & 13, pp. 423–425 & 652.) Measurements made on 26th–27th June 1952, at the Laboratoire National de Radioélectricité, using a wavelength of 11 km, are reported. The signalling speed was 40 words/min and the signal strength at the receiver 340 μV/m. A vertical aerial was used, and atmospherics were received simultaneously with the signal. The numbers of wrong or indecipherable figures and letters received are indicated on a graph of mean field strength of atmospherics plotted against hour of day. Good correlation is observed.

LOCATION AND AIDS TO NAVIGATION

534.87(204.1) : 621.396.822 **106**
Thermal-Noise Limit in the Detection of Underwater Acoustic Signals.—R. H. Mellen. (*J. acoust. Soc. Amer.*, Sept. 1952, Vol. 24, No. 5, pp. 478–480.) Experimentally found spectra of sea noise, for different sea states, are compared with the thermal-noise spectrum for an ideal medium as derived from classical statistical mechanics; the difference between the two is a function of frequency and sea state. The intensity of the smallest plane-wave signal detectable against the noise background by a linear reversible hydrophone is determined in terms of the directivity ratio, the bandwidth and the operating noise factor.

621.396.9 **107**
Radio Echoes and Lightning.—V. G. Miles. (*Nature, Lond.*, 30th Aug. 1952, Vol. 170, No. 4322, pp. 365–366.) Report of effects observed on the p.p.i. screen of radar equipment, with a vertically directed beam, during passage of a thunderstorm overhead, when echoes were obtained coincident with lightning flashes.

621.396.9 : 523.531 **108**
Radio-Echo Observations of the Major Night-Time Meteor-Streams.—G. S. Hawkins & M. Almond. (*Mon. Not. R. astr. Soc.*, 1952, Vol. 112, No. 2, pp. 219–233.) An account of determinations made during 1946–1951 of the radiant coordinates and meteor velocities of the Perseids, Geminids and Quadrantids. Data are included for five other streams.

621.396.9 : 629.13 **109**
Search Radar for Civil Aircraft.—P. L. Stride. (*J. Brit. Instn Radio Engrs*, Aug. 1952, Vol. 12, No. 8, pp. 445–460.) Consideration of the basic design problems of equipment intended primarily for (*a*) detection of storm clouds, (*b*) avoidance of high ground, (*c*) navigation by 'map painting', indicates that a pulse power of 10 kW at 3-cm wavelength, with a beam width of 6°, is adequate. Roll and pitch stabilization of the scanning axis is essential. A general description is given of suitable equipment, with details of the scanner and its stabilized mounting, the servo amplifier, the transmitter-receiver, the indicator, and the control unit. Results of performance tests are illustrated by typical displays.

MATERIALS AND SUBSIDIARY TECHNIQUES

533.5 : 678.14-415 **110**
Use of Membranes in Vacuum Technique.—Prugne & P. Garin. (*Le Vide*, May 1952, Vol. 7, No. 39, pp. 1197–1199.) The fitting of thin rubber membranes, either flat or corrugated, in pressure taps and valves is described.

535.37 : 621.397.621.2 **111**
A Single-Component White Luminescent Screen for Television Tubes.—F. A. Kröger, A. Bril & J. A. M. Dikhoff. (*Philips Res. Rep.*, Aug. 1952, Vol. 7, No. 4, pp. 241–250.) The development of a new phosphor, (Zn, Cd)S-Ag-Au-Al, is described. This exhibits white luminescence when excited by ultraviolet, X or cathode rays. Efficiency and current saturation are about the same as for the commonly used sulphide mixtures.

537.226 : 546.431.82 **112**
The Effect of the Polarizing Field on the Value of the Dielectric Constant and Dielectric Losses of $BaTiO_3$.—E. V. Sinyakov, E. A. Stafiychuk & L. S. Sinegubova. (*Zh. eksp. teor. Fiz.*, Dec. 1951, Vol. 21, No. 12, pp. 1396–1402.) A report on an experimental investigation, the main conclusions of which are as follows: (*a*) a constant electric field decreases the dielectric constant and loss angle particularly within the ferroelectric range of temperatures near the Curie point, and (*b*) the distorting action of the field displaces the Curie point towards higher temperatures.

537.226 : 621.396.677 **113**
Isotropic Artificial Dielectric.—C. Süsskind. (*Proc. Inst. Radio Engrs*, Oct. 1952, Vol. 40, No. 10, p. 1251.) Comment on 2523 of 1952 (Corkum).

537.311.33 **114**
The Physical Mechanism of [conduction] Phenomena in Complexes of Electronic Semiconductors.—J. Martinet. (*C. R. Acad. Sci., Paris*, 20th Oct. 1952, Vol. 235, No. 16, pp. 874–876.) Observed I/V characteristics are presented for various semiconductors formed by compacting and sintering iron-oxide powders; these characteristics are linear when plotted on a double-logarithmic scale. The investigation covers only low values of applied voltage. The results justify the hypothesis that the current travels in the same way through an agglomerate as through a potential barrier.

537.311.33 **115**
On the Variations of Lattice Parameters of some Semi-conducting Oxides.—L. D. Brownlee & E. W. J. Mitchell. (*Proc. phys. Soc.*, 1st Sept. 1952, Vol. 65, No. 393B, pp. 710–716.) The investigations made by Verwey et al. (905 of 1951) on Ni(Li)O systems have been extended to $Fe_2(Ti)O_3$ systems and to reduced Mg_2TiO_4.

537.311.33 : 546.289 **116**
Impurity Effects in the Thermal Conversion of Germanium.—W. P. Slichter & E. D. Kolb. (*Phys. Rev.*, 1st Aug. 1952, Vol. 87, No. 3, pp. 527–528.) Experiments on the growth of single crystals of Ge of very high purity have revealed no thermal acceptors as a consequence of the process of growth. Heat treatment of crystals after immersion in very dilute $CuSO_4$ solution resulted in extensive conversion. The results in general support the conclusion that conversion is associated with the rapid diffusion of Cu into the Ge.

537.311.33 : 546.289 **117**
Contact Properties of *p*-type Germanium.—J. W. Granville, H. K. Henisch & P. M. Tipple. (*Proc. phys. Soc.*, 1st Aug. 1952, Vol. 65, No. 392B, pp. 651–652.) Typical characteristics are shown (*a*) for an unformed W point contact on etched and on polished samples of *p*-type Ge, (*b*) for contacts between two samples of *p*-type Ge. Curves for the *n*-type material from which the *p*-type material was prepared by heat shock are included for comparison.

537.311.33 : 546.289 **118**
Area Contacts on Germanium.—J. W. Granville & H. K. Henisch. (*Proc. phys. Soc.*, 1st Aug. 1952, Vol. 65, No. 392B, pp. 650–651.) The I/V relations of large-area contacts, prepared by evaporation in vacuo of Au on to *n*-type Ge, are qualitatively similar to those for point contacts. No significant differences are found with contacts of Au, Ag or Cu and no improvement of the rectifying properties could be achieved by processes of the type used for 'forming' point contacts. Observation results are shown graphically, together with results for formed and unformed W point contacts.

537.311.33 : 546.289 **119**
Copper as an Acceptor Element in Germanium.—C. S. Fuller & J. D. Struthers. (*Phys. Rev.*, 1st Aug. 1952, Vol. 87, No. 3, pp. 526–527.) Report of results indicating that Cu is a surface impurity responsible for

the 'thermal conversion' of Ge [see also 2235 of 1952 (Fuller et al.)]. Experiments on Si at 1 100°C show that Cu diffuses into it at a rate comparable with that for Ge, and that an increase of hole conductivity occurs.

537.311.33 : 546.817.221 **120**

The Conductivity and Hall Coefficient of Sintered Lead Sulphide.—E. H. Putley. (*Proc. phys. Soc.*, 1st Sept. 1952, Vol. 65, No. 393B, pp. 736–737.) Values measured over a range of temperatures are shown in graphs and compared with values obtained previously on single crystals. The variation of the Hall coefficient is substantially the same for sintered specimens as for single crystals of comparable purity. The variation of conductivity is similar for the two types down to the temperature of minimum conductivity; below this temperature intergranular boundaries affect the conductivity of the sintered specimens.

537.311.33 : 621.314.7 **121**

On the Distribution of Transistor Action.—T. H. Tønnesen. (*Proc. phys. Soc.*, 1st Sept. 1952, Vol. 65, No. 393B, pp. 737–739.) Results are reported of experiments on a larger number of semiconductors prepared in the form of filaments; about 40 of the materials are considered likely to show transistor action.

537.311.33 : 621.396.822 **122**

Electrical Noise in Semiconductors.—H. C. Montgomery. (*Bell Syst. tech. J.*, Sept. 1952, Vol. 31, No. 5, pp. 950–975.) A survey is made of the noise characteristics of Ge diodes and triodes, in the light of existing theories. Experiments with single-crystal Ge filaments carrying current are described; the results indicate that the noise is produced by variations in the concentration of the minority carrier (i.e. holes in n-type material, electrons in p-type). Noise voltages in adjacent portions of a filament were quantitatively correlated with the life time and transit time of the minority carriers, and the effect of a magnetic field on the noise was found to agree with the calculated changes of the life time of the minority carriers.

538.221 **123**

Magnetic Viscosity under Discontinuously and Continuously Variable Field Conditions.—R. Street, J. C. Woolley & P. B. Smith. (*Proc. phys. Soc.*, 1st Sept. 1952, Vol. 65, No. 393B, pp. 679–696.)

538.221 **124**

Vibralloy—a New Ferromagnetic Alloy.—M. E. Fine. (*Bell Lab. Rec.*, Sept. 1952, Vol. 30, No. 9, pp. 345–348.) An alloy of Fe with 43% Ni and 9% Mo, developed from elinvar, has elastic and magnetic properties combining to make it highly suitable for vibrating reeds.

538.221 **125**

The Magnetoresistance of Ferromagnetic Al-Si-Fe Alloys.—R. Parker. (*Proc. phys. Soc.*, 1st Aug. 1952, Vol. 65, No. 392B, pp. 616–620.) By extending the treatment previously given (3018 of 1951), an equation is derived for the saturation magnetoresistance of some mixed ferromagnetic alloys as a function of composition and temperature. Results calculated from this equation are in good agreement with measured values for various Al-Si-Fe alloys.

538.221 : 621.314.2.042.143 **126**

The Change of Shape of the Hysteresis Loop of Transformer Sheet exhibiting Magnetic Creep.—R. Feldtkeller. (*Z. angew. Phys.*, Aug. 1952, Vol. 4, No. 8, pp. 281–284.) Observations are reported on sheets of carbon-containing Si-Fe alloy subjected to a weak sinusoidally varying magnetic field. Immediately after switching on the field the hysteresis loop exhibits a marked constriction at the centre; the constriction disappears after some hours at room temperature and more quickly at higher temperatures. The presence of the constriction corresponds with the existence of short-lived overtones in opposite phase to the stable overtones corresponding to the points of the hysteresis loop. The constriction results from the peculiar statistical nature of the Barkhausen jumps in the demagnetized material.

538.221 : 669.862.5.721 **127**

Ferromagnetism of Certain Gadolinium-Magnesium Alloys.—F. Gaume-Mahn. (*C. R. Acad. Sci., Paris*, 4th Aug. 1952, Vol. 235, No. 5, pp. 352–354.)

538.221 : 681.142 **128**

The Behavior of Rectangular-Hysteresis-Loop Magnetic Materials under Current-Pulse Conditions.—E. A. Sands. (*Proc. Inst. Radio Engrs*, Oct. 1952, Vol. 40, No. 10, pp. 1246–1250.) An account of an investigation of the effect of variation of magnetic and physical parameters on the time needed to change a magnetic toroid from a condition of residual flux to that of saturation.

538.632 : 546.431-31 **129**

The Hall Effect in Single Crystals of Barium Oxide.—E. M. Pell. (*Phys. Rev.*, 1st Aug. 1952, Vol. 87, No. 3, pp. 457–462.) Measurement results obtained by an a.c. method for the temperature range 400–800°K are presented. Conduction was predominantly n-type.

538.652 **130**

Investigations of the Variation of the Partial Processes in Magnetostriction.—E. Bailitis, C. Hagen & H. H. Rust. (*Z. angew. Phys.*, Aug. 1952, Vol. 4, No. 8, pp. 284–291.) Observations confirm that when a ferromagnetic material is magnetized, both reversible and irreversible changes of length occur, the overall longitudinal magnetostriction representing a resultant effect. Three separate component variations are isolated, designated respectively as the main, remanence and inertia components; on superposition these give the overall characteristic found by Nagaoka. Materials investigated include invar, indilatans extra, superinvar, Si-Cr-Fe, Ni, Ni-Mn and silver steel. The three magnetostriction components differ widely between these materials. Making some simplifying assumptions, a distribution function for the arrangement of the Weiss moments is derived which provides a satisfactory explanation of the experimental results. Discontinuous variation of length is observed, when the magnetizing force is decreased, for values less than the coercive force.

538.652 : 538.221 **131**

Magnetostriction of Cobalt Ferrites as a Function of Composition.—R. Vautier. (*C. R. Acad. Sci., Paris*, 4th Aug. 1952, Vol. 235, No. 5, pp. 356–358.) Non-oriented Co ferrites have high negative magnetostriction, while oriented samples exhibit a slight positive effect in the direction of orientation and larger negative effects perpendicular to the orientation direction. Measurement results are shown graphically for CoO contents of 35–48%.

538.652 : 538.221 **132**

Temperature Variation of the Magnetostriction of a Cobalt Ferrite.—R. Vautier. (*C. R. Acad. Sci., Paris*, 11th Aug. 1952, Vol. 235, No. 6, pp. 417–419.)

539.231 : 546.48-31 **133**

Electrical Conductivity and Structure of Sputtered CdO Layers.—G. Helwig. (*Z. Phys.*, 19th Aug. 1952, Vol. 132, No. 5, pp. 621–642.) The specific conductivity depends not only on the sputtering time, but also on the electrical power used in the process and on the oxygen content of the N-O or Ar-O mixtures. Test methods and results are described.

539.24 : 537.311.31 : 546.92 **134**

Electrical Conduction of Thin Films of Platinum covered with a Dielectric Layer by Evaporation in a Vacuum.—C. Feldman & B. Vodar. (*C. R. Acad. Sci., Paris*, 11th Aug. 1952, Vol. 235, No. 6, pp. 414–417.) The effect on the resistance of the Pt film of covering it with a layer of SiO_2 is to increase the variation with temperature and to decrease the variation with applied electric field. See also 3126 of 1952 (Feldman).

546.482.21 **135**

Single Synthetic Cadmium Sulfide Crystals.—S. J. Czyzak, D. J. Craig, C. E. McCain & D. C. Reynolds. (*J. appl. Phys.*, Aug. 1952, Vol. 23, No. 8, pp. 932–933.) A procedure is described which results in the growth of single crystals in the form of hexagonal prisms, starting with chemically pure CdS powder.

621.315.61 : 537.222.5 **136**

Characteristics and Measurement of Brush-Discharge Ionization in Dielectrics.—D. Renaudin. (*Bull. Soc. franç. Élect.*, Aug. 1952, Vol. 2, No. 20, pp. 431–435.) Local ionization at comparatively low voltages due to impurities in the dielectric is discussed and qualitative methods of measurement are noted. A quantitative measurement of mean ionization density is made by a comparative method, using the recurrent rapid discharge of a capacitor with controlled ionization. Results of such measurements on h.v. transformers, some of which were oil-filled, are shown.

621.315.612 : 546.882/.883]-3 **137**

Niobate and Tantalate Dielectrics.—E. Wainer & C. Wentworth. (*J. Amer. ceram. Soc.*, 1st Aug. 1952, Vol. 35, No. 8, pp. 207–214.) The ceramic and electrical properties of the niobates and tantalates of the elements of the first and second groups are described in detail, with numerous tables and diagrams. The materials include members of the perovskite crystal system which are ferroelectric and have Curie points in the range 350–475°C. The materials $NaNbO_3$ and $NaTaO_3$ are of particular interest. The results obtained indicate that $Cd_2Nb_2O_7$ is the ceramic and dielectric analogue of $SrTiO_3$, and that the binary compound $NaNbO_3.Cd_2Nb_2O_7$ is the analogue of $BaTiO_3$, except that its high dielectric constant and piezoelectric properties are maintained over a much wider range of temperature. Study of these compounds has indicated means for extending the temperature range of usefulness of perovskite-type ceramics for capacitor and electromechanical applications.

621.315.616.96 **138**

Some Physical Constants of Araldite.—P. André. (*Le Vide*, May 1952, Vol. 7, No. 39, pp. 1200–1202.) Dielectric characteristics are shown as functions of temperature and frequency.

621.317.011.5 : [546.212 + 612.1 **139**

A Comparison of the Dielectric Behaviour of Pure Water and Human Blood at Microwave Frequencies.—H. F. Cook. (*Brit. J. appl. Phys.*, Aug. 1952, Vol. 3, No. 8, pp. 249–255.) Methods of measuring the complex dielectric constant at frequencies from 1·7 kMc/s to 24 kMc/s are described. The results for water at temperatures in the range 0–60°C are analysed in relation to the Debye and the Cole-Cole dispersion equations, and the possibility that the dispersion is characterized by a narrow spectrum of relaxation times is discussed. Results for whole blood are given for the temperature range 15–35°C; the observed dispersion is attributed entirely to water relaxation. 41 references.

621.396.611.21 **140**

Some Characteristics of Quartz Crystals.—J. Coulon. (*Rev. gén. Élect.*, Aug. 1952, Vol. 61, No. 8, pp. 373–380.) Operational characteristics related to two particular points on the resonance curve of a quartz crystal are discussed, and a method for determining the constants of quartz crystals, making use of a circle diagram, is described. See also 2549 and 2550 of 1952.

666.1.037 : 621.3.032.7 **141**

Techniques of Sealing by Optical Polishing.—J. Bleuze & P. Dussaussoy. (*Le Vide*, May 1952, Vol. 7, No. 39, pp. 1182–1190.) In the basic technique fusion occurs between polished glass surfaces at a steady temperature of about 500°C. A steady low pressure is maintained during the complete process. Either flat or curved surfaces are used. Enamel or colloidal Ag may be applied before sealing. The technique also applies to metals and ceramics. The application in valve manufacture is described, with details of the processes of surfacing the glass and mounting the electrodes. The characteristics of some types of valve have been much improved by adoption of this method of sealing.

666.1.037.5 **142**

A High-Conductivity Glass-to-Metal Seal.—J. C. Turnbull. (*RCA Rev.*, Sept. 1952, Vol. 13, No. 3, pp. 291–299.) A method is described for plating kovar with high-conductivity metals (Cu and Cr) before sealing to glass. This procedure reduces the h.f. heating of the seals.

666.22 : 546.244-31 **143**

Tellurite Glasses.—J. E. Stanworth. (*J. Soc. Glass Tech.*, Aug. 1952, Vol. 36, No. 171, pp. 217–241.) A detailed account of the work noted in 2834 of 1952.

MATHEMATICS

517.941.91 **144**

An Integral Variant associated with the Wave Equation.—F. H. van den Dungen. (*C. R. Acad. Sci., Paris*, 8th Sept. 1952, Vol. 235, No. 10, pp. 532–533.)

517.948 **145**

Solution of Systems of Linear Equations by Minimized Iterations.—C. Lanczos. (*J. Res. nat. Bur. Stand.*, July 1952, Vol. 49, No. 1, pp. 33–53.) The general principles previously developed (1418 of 1951) are applied to the solution of large systems of linear algebraic equations.

681.142 **146**

A Universal Unit for the Electrical Differential Analyzer.—R. Tomovich. (*J. Franklin Inst.*, Aug. 1952, Vol. 254, No. 2, pp. 143–151.)

681.142 **147**

Automatic Programme Planning for Programme-Controlled Computers.—H. Rutishauser. (*Z. angew. Math. Phys.*, 15th July 1952, Vol. 3, No. 4, pp. 312–313.) Methods for using the computer itself to determine the programme for a given problem have been worked out; details are described in a booklet to be published soon.

681.142 **148**

Digital to Analog Converter.—M. Miller, B. L. Waddell & J. Patmore. (*Electronics*, Oct. 1952, Vol. 25, No. 10, pp. 127–129.) Data in digital form, e.g. on punched cards, are converted into direct voltages; the basic elements of the equipment comprise timer, two temporary storage units, two d.c. converters and a control panel.

681.142 : 517.392 149
Development of a Product Integrator.—P. Germain. (*HF, Brussels*, 1952, Vol. 2, No. 3, pp. 69–75.) The two functions whose product is to be integrated are represented by curves drawn on the two halves of a cylindrical drum. Photoelectric equipment gives the ordinate of each curve for any particular value of the variable x, and electrical pulse methods are used to obtain the product of the ordinates and to integrate the successive products for n equidistant values of x.

681.142 : [621.392.26 + 621.396.611.4 150
The Solution of Waveguide and Cavity-Resonator Problems with the Resistance-Network Analogue.—G. Liebmann. (*Proc. Instn elect. Engrs*, Part III, Sept. 1952, Vol. 99, No. 61, pp. 316–319.) Digest only. See 2839 of 1952.

681.142 : 621.396.6 : 511.124 151
A High-Accuracy Time-Division Multiplier.—E. A. Goldberg. (*RCA Rev.*, Sept. 1952, Vol. 13, No. 3, pp. 265–274.) Equipment for use in analogue-type computers is described which produces a train of rectangular pulses whose amplitude is proportional to one variable, and whose duration is proportional to another variable. The average or d.c. component of the pulse train is then proportional to the product of the two variables. Accuracy to within 0·01% of full scale is achieved by use of (*a*) a feedback system for establishing accurate timing, (*b*) steep-fronted switching pulses, (*c*) an electronic switch of predictable performance independent of valve characteristics, (*d*) precision resistors and reference voltages.

MEASUREMENTS AND TEST GEAR

531.765 : 621.396.615.17 152
Sawtooth-Current Generator with Long Sweep Time for Recording of Time Intervals.—H. Lueg & E. Oberhausen. (*Arch. tech. Messen*, July 1952, No. 198, pp. 145–146.) Description, with detailed circuit diagram, of equipment based on the Miller integrator [347 of 1949 (Briggs)], with a linear time scale up to 20 sec and with a frequency constancy, under normal operating conditions, to within ±1%. For a sweep of 100 sec, time errors up to 3% may occur. With the addition of a univibrator and recorder, time intervals are registered directly as ordinates.

621.3.018.41(083.74) : 621.317.361 : 529.77 153
The Estimation of Absolute Frequency in 1950–1951.—H. M. Smith. (*Proc. Instn elect. Engrs*, Part II, Aug. 1952, Vol. 99, No. 70, pp. 407–409; *ibid.*, Part III, Sept. 1952, Vol. 99, No. 61, pp. 320–321.) Summary only. Measurements at Greenwich of the mean annual fluctuation of the period of the earth's rotation show a diminution of 40% in amplitude compared with the mean of published values for the period 1934–1949. A table is given which shows the mid-monthly deviations from the nominal frequency of the frequency standards at Abinger (2), Greenwich (1), Dollis Hill (G.P.O.) (4), and Teddington (N.P.L.) (1) for 1950 and 1951. A criterion of the quality of these standards is furnished by the mean absolute value of the tabulated deviations over a period. For five of the standards the criterion is about 1 part in 10^9 and for the other three about 2 parts in 10^9 per month per month.

621.317 : 061.3 154
Electrical Engineering Measurement Technique.—(*Electronica*, 2nd Aug. 1952, Vol. 5, No. 102, pp. 121–127.) Report of papers and discussions at a conference held at Delft in May 1952.

621.317.029.6 : 621.396.621.54 155
Principles and Applications of Converters for High-Frequency Measurements.—D. A. Alsberg. (*Proc. Inst. Radio Engrs*, Oct. 1952, Vol. 40, No. 10, pp. 1195–1203.) The use of heterodyne methods enables measurements to be made over wide ranges of frequency, with the reference standards operating at a fixed frequency. The accuracy of such measurements depends on the performance characteristics of the transducers or converters used. Design principles are outlined for maximum linearity and dynamic range of converters and for minimum zero corrections. These principles are applied in equipment for point-by-point and sweep measurements of delay, phase, impedance, and transmission characteristics.

621.317.328.029.63 156
600-Mc/s Field-Strength Meter.—A. C. Gordon-Smith. (*Wireless Engr*, Nov. 1952, Vol. 29, No. 350, pp. 306–308.) Receiving equipment is described suitable for the accurate comparison of field strengths over the frequency range 500–700 Mc/s. For measuring modulated signals, the equipment comprises a crystal frequency changer, a 30-Mc/s i.f. amplifier incorporating a piston attenuator, and a 1-kc/s selective amplifier followed by rectifier and d.c. meter. For measuring unmodulated signals the a.f. amplifier is omitted. Calibration was performed both by the field-radiation method and by the direct-injection method; good agreement was obtained between the two methods.

621.317.332 : 621.396.615.141.2 157
Conductance Measurements on Operating Magnetron Oscillators.—M. Nowogrodzki. (*Proc. Inst. Radio Engrs*, Oct. 1952, Vol. 40, No. 10, pp. 1239–1243.) The conductance terms in formulae for the equivalent circuit of a magnetron oscillator can be obtained from measurements of the Q-factor of the magnetron in the oscillating and non-oscillating conditions. The 'operating' Q-factors are determined from measurements of the variations of output power and oscillation frequency caused by a specified load mismatch. Experimental data are quoted.

621.317.34 : 621.315.212 158
Measurement of the Characteristics of a Cable for Radio-Frequency Transmission.—H. Vigneron. (*HF, Brussels*, 1952, Vol. 2, No. 3, pp. 77–80.) Theoretical formulae are derived which are less general, but which are both easier to obtain and to manipulate than those given by Hontoy (3169 of 1952). Hontoy's notation is used. A simpler method of measurement is also described, based on an acceptable approximation in which a spiral on a Smith's diagram is replaced by a circle. The method also furnishes a supplementary constant of the cable.

621.317.341.029.62 159
Measurement of Transmission-Line Attenuation.—(*Tech. News Bull. nat. Bur. Stand.*, Sept. 1952, Vol. 36, No. 9, pp. 133–135.) The method applies to balanced unscreened transmission lines. A sliding polystyrene fitting holds a small pickup loop at a constant distance from the resonant line, both ends of which are shorted. A rectifier and filter attached to the fitting are connected to a galvanometer measuring the standing waves along the line. Attenuation is approximately equal to $\coth^{-1}\alpha$, where α is the s.w.r. Matching difficulties are avoided; accuracy to within 1% is attainable.

621.317.35 160
Methods of obtaining Amplitude-Frequency Spectra.—A. E. Hastings. (*Rev. sci. Instrum.*, July 1952, Vol. 23, No. 7, pp. 344–346.) The frequency components of an arbitrary function of short duration are obtained from the discrete spectrum which results when the function

is repeated at a fixed rate. The analysis is presented, and alternative practical arrangements are described involving optical or c.r. scanning or recording on magnetic tape.

621.317.35 : 519.272.119 **161**

Device for Computing Correlation Functions.—A. E. Hastings & J. E. Meade. (*Rev. sci. Instrum.*, July 1952, Vol. 23, No. 7, pp. 347–349.) An analogue device is described in which the original time function is recorded on magnetic tape and is reproduced by two pickups with the desired time separation. An electrodynamic wattmeter is used to multiply the outputs together.

621.317.351 **162**

Cathode-Ray-Tube Beam Intensifier.—R. W. Rochelle. (*Electronics*, Oct. 1952, Vol. 25, No. 10, pp. 151–153.) To obtain the very high operating speed required for examining small portions of pulse edges, the intensifier circuit uses a hard valve for gating, together with two thyratrons for shutting off the beam. Timing is controlled by use of appropriate lengths of delay cable.

621.317.353.2/.3 **163**

Double-Tone and Intermodulation Methods of Distortion Measurement: Differences between the Results obtained, and their Causes.—H. Müller. (*Telefunken Ztg*, Aug. 1952, Vol. 25, No. 96, pp. 142–148.) Conversion of the results obtained by one method into those given by the other is only found possible if the equipment under test has the same frequency response characteristic both for the primary tones used in the double-tone method and for the distortion products. Discussion of transit distortion and load distortion is based on consideration of the characteristics of the circuit equivalent of a distorting network, consisting of a quadripole producing nonlinear distortion connected between two other quadripoles producing only linear distortion. Analysis and measurement results show that neither of the two methods furnishes correct values in all cases. The intermodulation method is recommended for use in the lower third of the a.f. range and the double-tone method for the upper part of the range.

621.317.4 : 621.397.621 **164**

Field Plotting in Deflection-Yoke Design.—Sieminski. (See 256.)

621.317.444 **165**

Magnetic-Field Measurements with the Iron-Cored Magnetometer by the Harmonic Method.—R. Kühne. (*Arch. tech. Messen*, Aug. 1952, No. 199, pp. 175–178). Discussion of basic principles of the method and review of practical equipment. 28 references.

621.317.7 : 061.4 **166**

Electrical Measurement Technology.—W. Hunsinger. (*Z. Ver. dtsch. Ing.*, 1st July 1952, Vol. 94, No. 19, pp. 619–626.) Review of the measurement equipment shown at the 1952 Technical Fair, Hanover, with short descriptions of selected exhibits and 147 references to relevant papers.

621.317.725 + 621.317.79.018.78 **167**

Voltage and Distortion Meters without Valves, as Examples of Rectifier-Type Instruments.—K. Hagenhaus. (*Frequenz*, Aug. 1952, Vol. 6, No. 8, pp. 217–222.) Description of (*a*) a rectifier-capacitor-bridge type of voltmeter with ranges covering 1–500 V and with scale errors not exceeding 2% from 30 c/s to 300 Mc/s, (*b*) a distortion meter with a range of 0·5–30%, suitable for measurement of hum and harmonic distortion of equipment for transmission of speech or music, using fixed frequencies of 160, 180, 3 000 and 5 000 c/s.

621.317.733 : 621.311.6 **168**

Precision Voltage Source.—W. J. Cunningham. (*Wireless Engr*, Nov. 1952, Vol. 29, No. 350, p. 309.) Comment on 3505 of 1952 (Attree).

621.317.733.029.54/.55 **169**

Impedance Bridges for the Megacycle Range.—H. T. Wilhelm. (*Bell Syst. tech. J.*, Sept. 1952, Vol. 31, No. 5, pp. 999–1012.) Three bridges designed for precision measurements on networks and components for wide-band coaxial-line transmission systems are described in detail: (*a*) general-purpose 20-Mc/s unit operating both as admittance and series-impedance bridge and covering a reactance range from a few ohms to nearly one megohm; (*b*) 5-Mc/s Maxwell inductance bridge for the range $0{\cdot}001\mu$ H–10 μH; (*c*) 10-Mc/s admittance bridge especially for determining the temperature coefficients and frequency characteristics of small capacitors up to 200 pF, with an accuracy to within 0·01 pF. Standards having a range of several decades are provided.

621.317.734 **170**

Two Electronic Resistance or Conductance Meters.—L. B. Turner. (*Proc. Instn elect. Engrs*, Part III, Sept. 1952, Vol. 99, No. 61, p. 322.) Digest only. See 2853 of 1952.

621.317.761 + 621.396.621.54 **171**

Direct-Reading Frequency-Measurement Equipment for the Range 30 c/s–30 Mc/s.—L. R. M. Vos de Wael. (*Onde élect.*, Aug./Sept. 1952, Vol. 32, Nos. 305/306, pp. 351–356.) See 2855 of 1952.

621.396.645.35 : 621.317.3 **172**

High-Gain D.C. Amplifiers.—K. Kandiah & D. E. Brown. (*Proc. Instn elect. Engrs*, Part II, Aug. 1952, Vol. 99, No. 70, pp. 314–326. Discussion, pp. 344–348.) Critical review of various methods of measuring small currents and voltages, with discussion of the use of negative feedback, d.c./a.c. conversion by means of a contact modulator, magnetic modulators, and correction of zero drift in direct-coupled amplifiers.

621.396.645.35 : 621.318.435.3 **173**

The Design of a Practical D.C. Amplifier based on the Second-Harmonic Type of Magnetic Modulator.—Noble & Baxandall. (See 72.)

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

531.552 : 621.396.9 **174**

Radar Measurement of the Velocity of Projectiles.—B. Koch. (*Onde élect.*, Aug./Sept. 1952, Vol. 32, Nos. 305/306, pp. 357–371.) Detailed account of a method of using the Doppler-Fizeau effect to derive a continuous indication of the velocity of a projectile. Wavelengths in the dm or cm range are used, the transmitting and receiving equipment being preferably located close to the projectile path. In the case of cannon, the equipment is installed either directly behind or in front of the gun. Typical records are reproduced and numerous measurements quoted which show the accuracy to be at least equal to that given by classical methods.

537.533.72 : 538.691 **175**

Two-Directional Focusing with Short Uniform Magnetic Fields.—C. Reutersward. (*Ark. Fys.*, 14th Aug. 1952, Vol. 4, Parts 1/2, pp. 159–171.) Analysis with application to ion lenses.

621.314.3† : 621.316.728 176

Power Control with Magnetic Amplifiers.—E. L. Harder. (*Electronics*, Oct. 1952, Vol. 25, No. 10, pp. 115–117.) The use of magnetic amplifiers to control high-power industrial equipment and other devices is described.

621.317.3.029.5 : 677 177

High-Frequency Measurement Methods in the Textile Industry.—H. Locher. (*Bull. schweiz. elektrotech. Ver.*, 9th Aug. 1952, Vol. 43, No. 16, pp. 653–658.)

621.365.54† : 061.4 178

Equipment for Inductive Heating.—W. Stuhlmann. (*Z. Ver. dtsch. Ing.*, 1st July 1952, Vol. 94, No. 19, pp. 653–654.) Review of exhibits at the 1952 Technical Fair, Hanover.

621.365.54† : 621.935 179

High-Frequency Tempering of the Points of the Teeth of a Band-Saw.—F. P. Pietermaat & R. Antoine. (*HF, Brussels*, 1952, Vol. 2, No. 3, pp. 53–54.) Illustrated description of the arrangements for inductive heating of the teeth, using a 1-Mc/s generator, and subsequent quenching in an oil shower.

621.383 : 621.96 180

An Electronic Tracing Head for Oxygen Cutting.—H. E. Newton. (*Metrop. Vick. Gaz.*, Sept. 1952, Vol. 24, No. 398, pp. 227–231.) Description of pantograph-type apparatus including travel motor, steering motor and optical system with photoelectric control causing the tracing head to follow the outline of a drawing in Indian ink on white paper.

621.384.611 181

The Oak Ridge 86-Inch Cyclotron.—R. S. Livingston. (*Nature, Lond.*, 9th Aug. 1952, Vol. 170, No. 4319, pp. 221–223.) Description, with operating details, of an accelerator giving an internal proton beam of intensity $>$ 1 mA at 24 MeV.

621.384.612 182

The Electron Synchrotron.—S. E. Barden. (*Metrop. Vick. Gaz.*, Aug. 1952, Vol. 24, No. 397, pp. 207–217.) An outline of synchrotron theory, with some details of the 375-MeV synchrotron built for Glasgow University.

621.384.612.1† 183

Cyclotron.—Please note that the number 621.384.611 will be used in future for cyclotrons in place of 621.384.612.1† used hitherto.

621.384.62 184

An Electrostatic Generator for 1 MV.—N. Forsberg & P. Isberg. (*Ark. Fys.*, 5th Aug. 1952, Vol. 3, No. 6, pp. 519–524.) Description of a non-pressurized van de Graaff generator constructed for nuclear research at the Royal Institute of Technology, Stockholm. The linear accelerator used with the generator is also described.

621.385.833 185

Measurement of the First and Second Derivatives of the Axial Field of a Powerful Magnetic Lens.—P. Gautier. (*C. R. Acad. Sci., Paris*, 4th Aug. 1952, Vol. 235, No. 5, pp. 361–364.)

621.385.833 186

New Type of Electrostatic Immersion Objective with High Resolving Power.—A. Septier. (*C. R. Acad. Sci., Paris*, 29th Sept. 1952, Vol. 235, No. 13, pp. 652–654.) A system comprising essentially a plane cathode followed by a unipotential lens was investigated in order to obtain a very strong extractor field. A resolution of the order of 100 mμ was obtained.

621.385.833 187

Resolving Power of the Electrostatic Immersion Objective.—A. Septier. (*C. R. Acad. Sci., Paris*, 22nd Sept. 1952, Vol. 235, No. 12, pp. 609–611.) For a particular metallurgical electron microscope the value found in practice for the resolving power is better than the theoretical value.

621.385.833 188

The Focal Properties and Spherical-Aberration Constants of Aperture Electron Lenses.—M. M. MacNaughton. (*Proc. phys. Soc.*, 1st Aug. 1952, Vol. 65, No. 392B, pp. 590–596.)

621.385.833 189

Investigation of Magnetic Lenses having the Axial Field $H(0, z) = \gamma/z^n$.—U. F. Gianola. (*Proc. phys. Soc.*, 1st Aug. 1952, Vol. 65, No. 392B, pp. 597–603.)

621.385.833 : 061.3 190

German Society for Electron Microscopy. Fourth Annual Conference.—T. Mulvey. (*Nature, Lond.*, 16th Aug. 1952, Vol. 170, No. 4320, pp. 271–273.) Summaries are given of some of the 80 papers read, selected to indicate the general scope of the conference held at Tübingen, June 1952. See also 3205 of 1952.

621.387.464 : 621.396.822 191

Discrimination against Noise in Scintillation Counters.—R. J. T. Herbert. (*Nucleonics*, Aug. 1952, Vol. 10, No. 8, pp. 37–39.)

621.396 : 623.5 192

Acoustic Firing Error Indicator.—M. C. Eliason & W. G. Hornbostel. (*Electronics*, Oct. 1952, Vol. 25, No. 10, pp. 98–101.) Equipment enabling useful information to be obtained from misses as well as hits comprises two radio transmitters mounted on the airborne target and radio receivers located near the gun. The transmitters are associated with special flat-response capacitor microphones which pick up the shock waves from passing bullets; the receivers indicate magnitude and sense of the error.

621.791.3 : 621.316.7.076.7 193

Electronic Control of Soldering Machines.—J. Dusailly. (*Tech. mod., Paris*, May 1952, Vol. 44, No. 5, pp. 141–143.) Description of equipment using thyratrons for timing and heating-current control of soldering operations.

621.791.7 : 621.317.313 194

Measurement of the Effective Values of Welding Currents.—R. Peretz. (*HF, Brussels*, 1952, Vol. 2, No. 3, pp. 55–67.) Two methods of measurement are described, one using a series manganin resistor and the other a magnetic amplifier, with electronic equipment for measuring the short welding time. Single-phase currents up to 20 kA, with thyratron control, were measured. Results obtained by the two methods were in good agreement.

PROPAGATION OF WAVES

538.566 195

The Nonexistence of Sommerfeld's Surface Wave.—P. Poincelot. (*C. R. Acad. Sci., Paris*, 4th Aug. 1952, Vol. 235, No. 5, pp. 350–352.) A source of error in Sommerfeld's analysis is noted, from which it is concluded that the surface wave does not exist. See also 2892 of 1949 (Kahan & Eckart).

538.566 : 551.510.535 **196**

The Mean Velocity of the Energy in a Nonuniform Ionized Absorbing Medium with Slowly Varying Parameters [stratified medium].—H. Arzeliès. (*C. R. Acad. Sci., Paris*, 11th Aug. 1952, Vol. 235, No. 6, pp. 421–423.) Theory is developed without assuming the propagated wave to be transverse and the electric and magnetic vectors both linearly polarized. Two parameters are introduced, characterizing respectively the phase velocity and the wave attenuation; these depend not only on the location of the point considered, but also on the angle between the phase velocity and the planes of stratification. There are two simple sets of polarization conditions possible: the electric vector is linearly polarized and the magnetic vector elliptically polarized, or vice versa. The first case arises if the wave penetrating the medium has its electric vector parallel to the strata. Simple formulae are derived for the mean energy velocity, the Poynting vector, and the energy density.

621.396.11 : 519.2 **197**

Fundamentals of Probability Theory.—G. Bangen & H. W. Fastert. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, 1952, Special No: Bases for planning of u.s.w. networks, pp. 35–44.) An outline of the principal features of the calculus of probability, with particular reference to its application to the phenomena of the propagation of e.m. waves. Section A deals with the one-dimensional random variable, with explanation of the characteristics of normal distribution and extension to random variables dependent on two parameters. Section B considers the case of two-dimensional random variables and explains the concept of correlation and also some special formulae.

621.396.11 : 537.562 **198**

The Propagation of Electromagnetic Signals in an Ionized Gas.—N. G. Denisov. (*Zh. eksp. teor. Fiz.*, Dec. 1951, Vol. 21, No. 12, pp. 1354–1363.) An exact solution of the problem in terms of Lommel's functions of two variables is obtained. Using an integral representation of these functions, an asymptotic form of the solution in terms of Fresnel integrals is derived which describes the field of the main part of the signal which has passed over sufficiently large distances. For this case relatively simple formulae are obtained for determining the envelope of the oscillations of the field; this makes the solution more convenient for practical application.

621.396.11 : 551.510.535 **199**

Calculation of Sky-Wave Field Strength.—K. Rawer. (*Wireless Engr*, Nov. 1952, Vol. 29, No. 350, pp. 287–301.) The first account in English of the method used by the French Service de Prévision Ionosphérique Militaire, in which the different paths actually followed in ionospheric propagation are considered separately and their effects combined. Absorption and blanketing by the lower ionosphere layers, and the geometrical optics of the reflection layer are taken into account. Comparison is made with the C.R.P.L. method. See also 2812 of 1951.

621.396.11 : 551.510.535 **200**

The B.B.C. Ionospheric-Storm-Warning System.—Bennington & Prechner. (See 103.)

621.396.11.029.55 **201**

Scatter Sounding: a New Technique in Ionospheric Research.—O. G. Villard, Jr, & A. M. Peterson. (*Science*, 29th Aug. 1952, Vol. 116, No. 3009, pp. 221–224.) See also 2579 of 1952.

621.396.11.029.55 **202**

Marked Deterioration of Radio-Propagation Conditions recently observed on Intercontinental Circuits using Decametre Waves.—J. Maire. (*Ann. Radioélect.*, July 1952, Vol. 7, No. 29, pp. 221–224.) After a relatively calm period of several years, a long succession of ionospheric storms commenced towards the middle of August 1950, which seriously affected communications on the Paris-New York circuit. Whereas in 1948 only two frequencies (10 and 18 Mc/s) or at most three (7, 10 and 18 Mc/s) sufficed for uninterrupted communication, five frequencies (5, 7, 10, 15 and 18 Mc/s) were found necessary during the winters 1950–51 and 1951–52. The relation of the effect to sunspot activity and geomagnetic variations is discussed. Both annual and 27-day-period variations have been noted. Since the minimum phase of solar activity is being approached, a diminution of the trouble is probable in the coming years.

621.396.11.029.64 : 621.396.812.3 **203**

Some Aspects of Microwave Fading on an Optical Path over Sea.—A. G. Bogle. (*Proc. Instn elect. Engrs*, Part III, Sept. 1952, Vol. 99, No. 61, pp. 236–240.) Observations of the fading of 3·26-cm and 9·2-cm signals transmitted across Cook Strait, New Zealand, were made for about 15 months during 1949 and 1951. Most of the equipment used was identical with that used in the Cardigan Bay tests described by Megaw (518 of 1947). Two principal types of fading were distinguished: (*a*) 'roller' fading in which broad maxima were separated by narrow deep minima at intervals ranging from a few minutes to about an hour; (*b*) slow shallow fading, usually associated with 'scintillation' and sometimes persisting for as long as six hours. From the evidence available it is concluded that the deep fading is caused by the random coincidence of two conditions, (*a*) phase opposition of the direct and indirect signals, which is dependent on the variation of the effective gradient of the modified refractive index over the path, (*b*) equality of the two signal strengths, which depends on the distribution of discontinuities of refractive index along the path, attributed to turbulence.

RECEPTION

621.396.621 : 621.396.619.11/.13 **204**

Circuit Technique of the New Valves for A.M./F.M.—D. Hopf & H. Bock. (*Funk-Technik, Berlin*, Aug. 1952, Vol. 7, No. 16, pp. 433–434..441.) Design details for an efficient a.m./f.m. receiver using two Type-ECH81 valves and one each of Types EF85, EABC80 and EL41.

621.396.621 : 621.396.822 : 621.392.52 **205**

Optimum Filters for the Detection of Signals in Noise.—L. A. Zadeh & J. R. Ragazzini. (*Proc. Inst. Radio Engrs*, Oct. 1952, Vol. 40, No. 10, pp. 1223–1231.) An optimum predetection filter is defined as one which maximizes the difference between the signal and noise components of the output in terms of a suitable difference function. In a special case, this definition leads to the criterion used by North (3419 of 1944) and yields filters which maximize the signal/noise ratio at a specified instant of time. North's theory of such filters is extended to the case of nonwhite noise and finite-observation-time filters. Explicit expressions for the pulse responses of such filters are developed and two practical examples are considered.

621.396.621 : 621.396.822 : 621.392.52 **206**

The Detection of a Sine Wave in the Presence of Noise by the Use of a Nonlinear Filter.—T. G. Slattery. (*Proc. Inst. Radio Engrs*, Oct. 1952, Vol. 40, No. 10, pp. 1232–1236.) Basic theory of the design of nonlinear filters for the detection of signals in noise is reviewed, and an account is given of the construction of such a filter for the detection of a sine wave of unknown frequency in the presence of random noise.

621.396.621.53.029.51/.55 **207**
The All-Wave Receiver, Type E 103 AW/4.—H. Behling. (*Telefunken Ztg*, Aug. 1952, Vol. 25, No. 96, pp. 185–193.) Description of a double-heterodyne receiver covering from 100 kc/s to 30 Mc/s in seven ranges with adequate overlap. As regards quality of reproduction, sensitivity and selectivity, the receiver is intermediate between a good broadcasting receiver and a high-quality commercial receiver. Performance data for the s.w. band are compared with results for a special s.w. receiver of the 'Köln' type.

621.396.622 : 621.396.619.11 **208**
Linear Detection of Amplitude-Modulated Signals.—H. Hellerman & C. R. Cahn. (*Proc. Inst. Radio Engrs*, Oct. 1952, Vol. 40, No. 10, p. 1251.) Analysis leading to an expression for the output of a type of linear detector suitable for a.m. signals, in a form which clearly indicates the theoretical limitations of the detection process.

621.396.823 : 621.396.67 **209**
Loop Aerial Reception.—Bramslev. (See 35.)

STATIONS AND COMMUNICATION SYSTEMS

621.396 : 656.22(43) **210**
Radio Communication Technique used on German Railways.—M. Finck & F. Pepping. (*Telefunken Ztg*, Aug. 1952, Vol. 25, No. 96, pp. 176–184.) An account of developments starting with the period 1920–1924, when long-wave telegraphy systems were used for communication between district head offices and also for communication with relief trains. Modern u.s.w. equipment is described.

621.396.1.029.6 : 061.3 **211**
European V.H.F. Broadcasting.—(*Wireless World*, Oct. 1952, Vol. 58, No. 10, pp. 433–434.) Summary of plans considered at the Stockholm conference, June 1952, for the allocation of frequency bands for television and sound. See also 3560 of 1952.

621.396.41 **212**
A Time-Division Multiplex Terminal.—O. E. Dow. (*RCA Rev.*, Sept. 1952, Vol. 13, No. 3, pp. 275–290.) Detailed description of the equipment for a 22-channel p.a.m. system with channel transmission bands from 100 c/s to 3·4 kc/s. Each channel is sampled 8333 times per second by means of an electronic distributor. The resulting a.m. pulses of both plus and minus polarities are used for f.m. of the r.f. carrier. A method of noise reduction and a crosstalk-balancing circuit are described and performance details of the system are given.

621.396.61.029.62 : 621.396.8 : 519.2 **213**
Simplified Method of Determining Service Probability and its Application to the Planning of U.S.W. Networks.—R. Gressmann & K. H. Kaltbeitzer. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, 1952, Special No: Bases for planning of u.s.w. networks, pp. 3–17.) The F.C.C. statistical treatment of field strength is outlined and its suitability for application to the conditions existing in Germany is critically discussed. The logarithmic-normal statistical field-strength distribution is explained. Practical examples illustrate the application of the statistical method for estimating the service area of a transmitter (*a*) with no interference, (*b*) with a single interfering signal, (*c*) with several interfering signals. Only simple mathematics is used. A map shows the estimated service areas of various television transmitters in Western Germany.

621.396.61.029.62 : 621.396.8 : 519.2 **214**
Method for Determining the Service Probability in the Field of a Transmitter affected by Any Number of Interfering Transmitters.—J. Grosskopf. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, 1952, Special No: Bases for planning of u.s.w. networks, pp. 18–34.) A detailed explanation is given of the statistical method of service-area estimation described in the Report of the Ad Hoc Committee, F.C.C., Washington, D.C. The effects of interfering fields are taken into account, and since the F.C.C. Report omits the greater part of the mathematical basis for the numerous formulae given, sufficient mathematics is included here to make the formulae intelligible.

621.396.65 : 551.594.6 **215**
Correlation between the Mean Level of Atmospherics and the Degree of Intelligibility in a Kilometre-Wave Radio Link.—Carbenay. (See 105.)

621.396.65 : 621.396.5 **216**
Radio Relay Stations of the TD-2.—W. L. Tierney (*Bell Lab. Rec.*, Aug. 1952, Vol. 30, No. 8, pp. 326–332.) Illustrations and brief descriptions of typical stations along the 2 992-mile circuit.

621.396.65.029.6 **217**
Microwave Radio Links.—A. T. Starr & T. H. Walker. (*Proc. Instn elect. Engrs*, Part III, Sept. 1952, Vol. 99, No. 61, pp. 241–255. Discussion, pp. 289–293.) Propagation phenomena are discussed in relation to the design of multichannel radio links, and a detailed treatment is given of a f.m. link, thermal and intermodulation noise being calculated in terms of the system parameters and the frequency deviation determined for minimum total noise. Pre-emphasis is discussed and the bandwidth necessary to meet specified requirements is determined. Statistical problems are not considered, and all used channels are assumed to have signals of equal amplitude. The mathematical analysis is given in appendices.

621.396.65.029.62 : 621.3.018.78 **218**
Multipath Distortions on a F.M. U.S.W. Beam Link and their Effect on its Operation.—H. J. Fründt. (*Telefunken Ztg*, Aug. 1952, Vol. 25, No. 96, pp. 149–157.) Investigations of a radio link between Berlin and the Harz mountains, operating in the 60-Mc/s band with a frequency swing of ± 150 kc/s, showed multipath-transmission effects with a transit-time difference of about 1·7 μs and an amplitude ratio of about 0·15, with corresponding fluctuations of crosstalk between the telephony channels. In spite of this the noise level was 6 N below and the crosstalk 5·7 N below the signal level. Multipath effects were not observed on a similar link between Berlin and the Harz mountains (2614 of 1952). This operates in the same frequency band and has a frequency swing of ± 500 kc/s, but the two paths are several kilometres apart and have different ground profiles.

621.396.65.029.64 : 621.396.61/.62 **219**
Microwave Techniques for Communication Links.—G. King, L. Lewin, J. Lipinski & J. B. Setchfield. (*Proc. Instn elect. Engrs*, Part III, Sept. 1952, Vol. 99, No. 61, pp. 275–288. Discussion, pp. 289–293.) An account of theoretical and practical work in connection with the development of microwave components for the frequency band 3·6–4·2 kMc/s, including hybrid-T junctions, wave-guide horns for feeding parabolic reflectors, low-reflection crystal mounting, aerial matching devices, etc. Measurement techniques required to ensure satisfactory performance are outlined and a detailed description is given of a precision microwave signal generator and a piston attenuator suitable for measurements on microwave circuits and components.

621.396.65.029.64 : 621.396.61/.62 **220**

Circuit Technique in Frequency-Modulated Microwave Links.—H. Grayson, T. S. McLeod, R. A. G. Dunkley & G. Dawson. (*Proc. Instn elect. Engrs*, Part III, Sept. 1952, Vol. 99, No. 61, pp. 256–274. Discussion, pp. 289–293.) Discussion of circuit problems in wide-band microwave radio links is illustrated by reference to the development of links operating at frequencies around 4 kMc/s for 180-channel telephony or for television. Modulating and demodulating circuits with very nearly linear characteristics, and i.f. amplifiers with uniform time-delay, are described, together with the techniques required to assess their performances. The optimum type of wide-band low-noise i.f. amplifier consists of earthed-grid triodes in cascade. High-level i.f. microwave mixers, and a.f.c. systems, are described and the special problems of television video-frequency circuits are discussed.

621.396.71 + 621.397.7] (71) **221**

The Radio Canada Building in Montreal.—A. Frigon. (*B.B.C. Quart.*, Summer 1952, Vol. 7, No. 2, pp. 100–106.) General description of the arrangements of the broadcasting and television sections in the adapted 12-storey building in Dorchester Street West, Montreal.

621.396.712(434.1) **222**

Transmitter Installation on the Hoher Meissner, North Hesse, Germany.—P. Eich. (*Telefunken Ztg*, Aug. 1952, Vol. 25, No. 96, pp. 194–197.) Outline description of the station equipment, which at present includes a medium-wave 20-kW transmitter, and a 10-kW u.s.w. transmitter feeding an 8-stack double-slot aerial. Programmes are transmitted over an u.s.w. link from the Feldberg.

621.396.712.2/.3 **223**

Auxiliary Circuits in a Studio.—W. Roos. (*Tech. Mitt. schweiz. Telegr.-TelephVerw.*, 1st April 1952, Vol. 30, No. 4, pp. 140–144. In German.) Description of control system and circuits installed at the Zürich studios enabling actors playing in separate studios to hear studio effects and speeches made in the other studios as necessary. Circuits for communication between sound engineer and actors are also described.

621.396.73 **224**

New Equipment for Outside Broadcasts.—S. D. Berry. (*B.B.C. Quart.*, Summer 1952, Vol. 7, No. 2, pp. 120–128.) Description of amplifier and associated equipment, comprising 4-channel mixer, Type-OBA/9 amplifier and programme meter (two per set), distribution and general control unit, loudspeaker amplifier and isolating amplifier, and power unit, together with loudspeaker, spares and accessories. Special features of the various units are noted.

621.396.931 **225**

Investigations on Train Radiotelephony.—H. Kobierski. (*Telefunken Ztg*, Aug. 1952, Vol. 25, No. 96, pp. 169–175.) An account of experiments carried out in a specially equipped coach on the electrified line from Nuremberg to Ratisbon. Frequencies in the 80-Mc/s and 160-Mc/s bands were used; their relative advantages and disadvantages are discussed. The results favour the use of a diversity transmission system in which the frequency separation between the two transmitters is about 30 kc/s (twice the f.m. deviation). The receivers are automatically tuned to the transmitter giving the stronger signal.

621.396.933 : 061.4 **226**

Air Radio Developments.—(*Wireless World*, Oct. 1952, Vol. 58, No. 10, pp. 397–400.) Review of airborne and ground equipment on view at the exhibition organized by the Society of British Aircraft Constructors, Farnborough, September 1952.

SUBSIDIARY APPARATUS

621-526 **227**

The Stability of a Multiple Linear Servo System.—F. H. Raymond. (*C. R. Acad. Sci., Paris*, 1st Sept. 1952, Vol. 235, No. 9, pp. 508–510.)

621.311.69 : 534.113 **228**

Recent Developments in Vibrators and Vibrator Power Packs.—J. H. Mitchell. (*J. Brit. Instn Radio Engrs*, Aug. 1952, Vol. 12, No. 8, pp. 431–444.) Discussion of the Grade-1 type of vibrator which is expected to give at least the 1 000-hour life required for Services equipment. Reasons are given for specializing mainly on synchronous split-reed separately driven types. An outline is given of extensive investigations of contact phenomena, vibrator design, and the circuits fed by the vibrator. A range of vibrators is described with outputs from a few milliwatts to over 200 W. Very high conversion efficiencies have been obtained.

621.314.63 : 546.824-3 **229**

Titanium-Dioxide Rectifiers.—R. G. Breckenridge & W. R. Hosler. (*J. Res. nat. Bur. Stand.*, Aug. 1952, Vol. 49, No. 2, pp. 65–72.) A detailed account of work carried out up to the present on the TiO_2-film rectifiers previously noted (3567 of 1952). A feature of great practical importance is the improvement in performance with increase of temperature. For a particular steam-treated sample the forward current increased, while the backward current decreased slightly, as the temperature was increased up to 140°C. Heating to 200°C, however, causes irreversible damage. Differences of characteristics observed with different counter-electrode metals are discussed, a marked change of properties being noted when Zn was used. Further investigations are in progress.

621.316.722 : 621.316.86 **230**

Voltage Regulators using Nonlinear Elements.—N'Guyen Thien-Chi & J. Suchet. (*Ann. Radioélect.*, July 1952, Vol. 7, No. 29, pp. 189–198.) A simple method of determining the parameters of circuits providing reference voltages working into a constant load is described, and is illustrated by calculations for arrangements using either a single element or a pair of the nonlinear resistors manufactured by the Compagnie générale de T.S.F. (3364 of 1952). Power limitations restrict the field of application of this type of voltage regulator.

621.316.722 : 621.384.5 **231**

How to Design V.R. Tube Circuits.—R. C. Miles. (*Electronics*, Oct. 1952, Vol. 25, No. 10, pp. 135–137.) Families of supply-voltage/supply-series-resistance curves for various voltage-regulator tubes are used as the basis for designing stabilized direct-voltage supply circuits.

TELEVISION AND PHOTOTELEGRAPHY

621.397 : 061.4 **232**

The [Television] **Society's Annual Exhibition.**—(*J. Televis. Soc.*, Jan./March & April/June 1952, Vol. 6, Nos. 9 & 10, pp. 326–329 & 390–393.) Brief descriptions are given of television and associated equipment exhibited at Century House, London, December 1951.

621.397.24 + 621.395.97 **233**

Relaying the Sound and Television Signals at the South Bank Site.—Barton-Chapple. (See 23.)

621.397.3 **234**

The Energy Spectrum of the Television Image.—F. Winckel. (*Arch. elekt. Übertragung*, Sept. 1952, Vol. 6, No. 9, pp. 385–387.) The investigations of Mertz & Gray

(1934 Abstracts, p. 568) are discussed. The gaps in the energy spectrum of the television image are evaluated by analysing picture content, taking the particular case of a vertical bar rotated through 360°. The width of the video-frequency sideband is shown to be proportional to the tangent of the angle of rotation. The gaps in the spectrum tend to become filled as the number of scanning lines decreases; interlacing turns this to advantage.

621.397.331.2 235

A Note on the Design of Constant-Resistance Cathode-Ray Deflection Circuits.—R. C. Webb. (*RCA Rev.*, Sept. 1952, Vol. 13, No. 3, pp. 335–343.) Description of a horizontal-deflection circuit which is particularly useful with small television pickup tubes like the vidicon, and allows several hundred feet of transmission line to be interposed between the amplifier valve and the deflection coil, which forms part of a network that appears as a pure resistance matching the line impedance. The efficiency of the system is low, but the resulting transient disturbances in video circuits are small.

621.397.335 : 535.623 236

Frame Synchronization for Color Television.—D. Richman. (*Electronics*, Oct. 1952, Vol. 25, No. 10, pp. 146–150.) Analysis indicates that the present F.C.C. monochrome synchronizing waveform is able to provide reliable triggered frame synchronization with the N.T.S.C. colour system [1750 of 1952 (Hirsch et al.)]. A suitable circuit is described.

621.397.5 237

Present Stage of Development of Television.—V. K. Zworykin. (*Ricerca sci.*, Oct. 1952, Vol. 22, No. 10, pp. 1893–1927.) A survey covering choice of television standards, tube developments, colour systems and applications in industry. 30 references.

621.397.5 : 535.623 238

Gamma Correction in Constant-Luminance Color-Television Systems.—S. Applebaum. (*Proc. Inst. Radio Engrs*, Oct. 1952, Vol. 40, No. 10, pp. 1185–1195.) Analysis of the effects of precorrection of the red, green and blue image coordinates to provide unity overall gamma.

621.397.5 : 535.623 239

Colour Television.—A. Karolus. (*Z. angew. Phys.*, Aug. 1952, Vol. 4, No. 8, pp. 300–320.) A survey paper reporting developments to date, especially in the U.S.A. 63 references.

621.397.5 : 535.623 240

A Summary of Recent Advances in 'Dot-Sequential' Color TV Systems.—B D. Loughlin. (*Proc. nat. Electronics Conf., Chicago*, 1951, Vol. 7, p. 361.) Summary only. See 826 of 1952.

621.397.5 : 535.88 241

An Experimental System for Slightly Delayed Projection of Television Pictures.—P. Mandel. (*Proc. Inst. Radio Engrs*, Oct. 1952, Vol. 40, No. 10, pp. 1177–1184.) Description of a system comprising a flying-spot scanner for pictures on 35-mm film, a microwave relay transmitting the picture signals to the receiving station, where the picture is reproduced on the screen of a c.r. tube, photographed on synchronously driven 16-mm film, which is developed, fixed and dried in 60 sec and then run through a standard cinema projector for display on a large screen. Tests carried out in Paris indicate that the loss in fine detail is about 10%. The slight delay of 60 sec should not be objectionable.

621.397.5 : 621.396.65 242

British Television Relay Network.—(*Elect. Commun.*, Sept. 1952, Vol. 29, No. 3, pp. 171–178.) A short description of the design and characteristics of the various links in this network, with special emphasis on the features of the r.f. relay chains.

621.397.5(41) : 061.3 243

The Convention on 'The British Contribution to Television'.—(*Onde élect.*, Aug./Sept. 1952, Vol. 32, Nos. 305/306, pp. 372–388.) Summaries of the papers presented at the convention, with an introduction by General Leschi. See also 2629 of 1952.

621.397.61 244

B.B.C. Television Transmitting Stations.—(*Engineer, Lond.*, 15th Aug. 1952, Vol. 194, No. 5038, pp. 227–229.) Some details are given of the high-power equipment brought into use at the Kirk o' Shotts station in August 1952. The vision transmitter uses low-level modulation, the output power of 50 kW being obtained with a pair of water-cooled triodes, Type BW.165, in the final stage. The sound transmitter is of conventional design using class-B modulation, with a carrier output of 18 kW at 100% modulation. The transmission line feeding the common aerial at the top of the 750-ft mast consists of a 5-in. Cu tube with an axial inner conductor of locked-coil wire rope with outer wires of high-conductivity Cu, the rope having a 2-ton load at the lower end. This construction reduces echo effects. Power supplies are noted and a map shows estimated field-strength contours. See also 2640 of 1952.

Details are also given of the equipment of the new station at Wenvoe, near Cardiff, which commenced transmissions in August 1952, using medium-power transmitters. The vision transmitter has a peak-white output of 5 kW, modulation being effected at the 500 W level; the sound transmitter has an output of 2 kW and uses a class-B modulator. As at Holme Moss and Kirk o' Shotts, a combining unit is used with a single transmission line feeding the common aerial at the top of the 750-ft mast. Programmes will ultimately be transmitted from London via a new coaxial cable, but until this is ready the transmissions will be routed over an experimental v.h.f. link between London and Cardiff.

621.397.61(43) 245

The First [German] **10-kW Television Transmitter.**—(*Funk-Technik, Berlin*, 1st Aug. 1952, Vol. 7, No. 15, pp. 396–397.) Technical details of the directional transmitter shortly to be brought into operation in Nikolassee, Berlin, to serve Berlin and Western Germany. The frequency range is 174–216 Mc/s, with a video-frequency bandwidth of 5 Mc/s.

621.397.61(494) 246

The Suitability of the Dôle as a Site for a Television Transmitter.—H. Laett. (*Tech. Mitt. schweiz. Telegr.-TelephVerw.*, 1st May 1952, Vol. 30, No. 5, pp. 167–172. In German.) For a station to serve an area including Lausanne and Geneva, the 'Combe gelée' plateau at about 1 500 m altitude, on the Swiss side of the Dôle, was considered likely to be a suitable site. Tests were made over the period July–October 1951, using a B.B.C. television transmitter (carrier power 500 W, frequency 62·25 Mc/s) with a wide-band directional aerial giving horizontal polarization, and pulse equipment for making reflection measurements. Field-strength distribution at height 3 m is shown on a map; the value at height 10 m is on the average 7 db below that at 3 m. It is estimated from the measurements that using a 5-kW transmitter and two aerials directed respectively towards Lausanne and Geneva, more than half a million people could be served; within the service zone the coverage would be 84%.

621.397.61.029.6 : 621.317.7 247

Ultra-High-Frequency Television Monitor.—F. D. Lewis. (*Proc. nat. Electronics Conf., Chicago*, 1951, Vol. 7, pp. 440–448.) Description of equipment for carrier-frequency monitoring of the video transmitter, and both carrier-frequency and modulation monitoring of the f.m. sound transmitter. A single precision reference crystal is used for independent monitoring of both carrier frequencies. A new harmonic generator using Ge diodes has been developed for the u.h.f. stage of the crystal-controlled multiplier chain. Heterodyne methods are applied to derive signals suitable for operating cycle-counter types of frequency-deviation meters.

621.397.611/.621].2 248

Scanning-Current Linearization by Negative Feedback.—A. W. Keen. (*Proc. Inst. Radio Engrs*, Oct. 1952, Vol. 40, No. 10, p. 1215.) Summary only. See 1752 of 1952.

621.397.611.2 249

Performance of the Vidicon, a Small Developmental Television Camera Tube.—B. H. Vine, R. B. Janes & F. S. Veith. (*Proc. nat. Electronics Conf., Chicago*, 1951, Vol. 7, pp. 449–453.) See 2914 of 1952.

621.397.62 250

Television A.G.C. Circuit.—G. F. Johnson. (*Wireless World*, Oct. 1952, Vol. 58, No. 10, pp. 424–426.) Description of a simple system of stabilizing mean brightness by maintaining a constant difference of average voltage between the grid of the first i.f. amplifier and the grid of the synchronizing-signal separator.

621.397.62 251

Problems of Television Projection-Type Receivers.—E. Schwartz. (*Elektrotech. Z., Edn B*, 21st Sept. 1952, Vol. 4, No. 9, pp. 249–253.) Effects associated with the c.r. tube, including halation, reflection from the tube walls, effect of screen thickness on luminous intensity, secondary and X-ray emission, and thermal loading of screen, are discussed and the optics of projection systems and the characteristics of picture screens are considered briefly.

621.397.62 252

An Economical Sync Clipper of Unusual Noise Immunity.—M. Marks. (*Proc. nat. Electronics Conf., Chicago*, 1951, Vol. 7, pp. 352–360.) Impulse noise strong enough to disturb the operation of most types of synchronizing circuit is rendered innocuous by the synchronizing-signal clipper described. A single conventional receiver valve is operated so that noise pulses, distinguished from desired signals by their higher peak amplitude, produce no output. Synchronization pulses, on the contrary, are clipped at top and bottom as usual and fed to the horizontal and vertical sweep systems. Practical circuits are described.

621.397.62 : 535.623 253

Synchronous Demodulator for Color TV.—R. B. McGregor. (*Electronics*, Oct. 1952, Vol. 25, No. 10, pp. 214 .. 230.) In the N.T.S.C. system the colour information is carried at a subcarrier frequency, the hue and saturation information being embodied respectively in the modulation of two subcarrier components with a mutual phase separation of 90°. The two signals are separated by applying to the suppressor grid of the detector valve an oscillating voltage of the same frequency as that of the subcarrier applied to the control grid, and in phase with one of the components.

621.397.62 : 621.396.622.72 254

A Constant-Input-Impedance Second Detector for Television Receivers.—W. K. Squires & R. A. Goundry. (*Proc. nat. Electronics Conf., Chicago*, 1951, Vol. 7, pp. 362–368.) See 2059 of 1952.

621.397.621 255

Faulty Interlacing.—W. T. Cocking. (*Wireless World*, Nov. 1952, Vol. 58, No. 11, p. 457.) Comment on 2918 of 1952 (Patchett), stressing that the first requirement for correct interlacing is that the frame synchronizing pulses applied to the sawtooth-wave generator shall be identical in successive frames for such time as they are capable of influencing the generator.

621.397.621 : 621.317.4 256

Field Plotting in Deflection-Yoke Design.—E. Sieminski. (*Electronics*, Oct. 1952, Vol. 25, No. 10, pp. 122–126.) Simple apparatus including a probe coil is described for measuring the field of magnetic deflection yokes for television tubes; the test procedure makes it easy to relate a difference of tube performance to a particular design variation.

621.397.621.2 257

Using C.R. Tubes with Internal Pole Pieces.—C. V. Fogelberg, E. W. Morse, S. L. Reiches & D. P. Ingle. (*Electronics*, Oct. 1952, Vol. 25, No. 10, pp. 102–105.) By including part of the magnetic focusing system inside the c.r. tube, the focusing-energy requirements can be reduced and the focusing field better controlled. Various examples of internal pole pieces are illustrated and methods of mounting them in television tubes are described.

621.397.645 + 621.397.61 258

Ekco Amplifier/Converter. New TV Retransmitting Equipment.—(*Electrician*, 22nd Aug. 1952, Vol. 149, No. 3871, pp. 537–538.) Short description of equipment comprising high-gain receivers for the sound and vision channels, together with frequency-changer units, low-level amplifiers, sound and vision transmitters, and a mains-operated power unit. Such equipment provides good reception over some 150 square miles in fringe or shadow areas. Provision is made for retransmission on either the British, American or European line standards.

621.397.7 + 621.396.71] (71) 259

The Radio Canada Building in Montreal.—Frigon. (See 221.)

621.397.813 260

Modulation Distortion in Television Reception, and the Possibility of its Compensation.—F. Kirschstein & H. Bödeker. (*Fernmeldetech. Z.*, Aug. 1952, Vol. 5, No. 8, pp. 357–361.) The deformation of television signals transmitted by a s.s.b. system is analysed, use being made of data furnished by Kell & Fredendall (2063 of 1949) on transient phenomena in vestigial-sideband transmission. The analysis was checked by tests carried out on the Feldberg experimental transmitter. A simple practical compensation arrangement is described which uses an RC circuit, with a time constant of 0·25 μs, as a frequency-dependent feedback unit in the cathode circuit of the video-amplifier output valve. This restores the rectangular form of the distorted wave from the demodulator.

TRANSMISSION

621.396.61 : 621.396.97 261

Automatic Broadcasting: Remote Control and Automatic Monitoring of Third Programme Transmitter.—R. W. Leslie & C. Gunn-Russell. (*Wireless World*, Nov. 1952, Vol. 58, No. 11, pp. 449–451.) See 2924 of 1952.

621.396.619.231.018.783† **262**

The Problem of Distortion in Anode-Voltage-Modulated Transmitters.—W. T. Runge. (*Telefunken Ztg*, Aug. 1952, Vol. 25, No. 96, pp. 135–142.) In the final h.f. stage, conversion distortions are small because the external resistance is sufficiently high compared with the valve internal resistance determined by the mutual conductance and penetration factor. Load distortion, however, cannot be neglected. A decrease of these distortions can be obtained by reducing the internal resistance of the control stage. Distortions in the modulation transformer are discussed by considering its equivalent circuit, and the nonlinear distortion factors are calculated. In the final stage of the modulator, nonlinear distortion also occurs, a 5% deviation of the amplitude of the fundamental wave producing a distortion of about 1%, a value which also results if the phases of the two half-waves differ by about 0·7°. Certain requirements can be deduced for the preliminary modulation stages; a modulator satisfying these requirements is described. Application of the knowledge gained in these investigations to the design of the 20-kW medium-wave transmitter installed on the Hoher Meissner resulted in the distortion being very small. Measurement results are quoted.

VALVES AND THERMIONICS

537.533 : 621.396.619 **263**

Analysis of Modulated Electron Beams.—W. W. Cannon & L. E. Bloom. (*Proc. nat. Electronics Conf., Chicago*, 1951, Vol. 7, pp. 59–63.) Description of experimental equipment for analysing the velocity and density distribution of an electron beam modulated by a 3-kMc/s signal.

621.314.7 + 621.396.622.63 **264**

Industrial Applications of Semiconductors: Part 5—Crystal Valves.—A. Lindell & G. M. Wells. (*Research, Lond.*, July 1952, Vol. 5, No. 7, pp. 317–323.) The construction and operating characteristics of point-contact and junction-type Ge and Si diodes and triodes are reviewed.

621.314.7 : 53.01 **265**

The Physics of Transistors.—E. Billig. (*Brit. J. appl. Phys.*, Aug. 1952, Vol. 3, No. 8, pp. 241–248.) Some of the fundamental concepts of the solid state which are required for an understanding of transistor action are discussed, electron energy levels, contact potential and potential barriers, rectification, extrinsic and intrinsic semiconductors, and transistor action being considered.

621.383.27 : 621.387.464 **266**

Two New Photomultipliers for Scintillation Counting.—M. H. Greenblatt, M. W. Green, P. W. Davison & G. A. Morton. (*Nucleonics*, Aug. 1952, Vol. 10, No. 8, pp. 44–47.) Detailed description of a high-efficiency tube, Type H-5037, with a large-area photocathode, and of a high-gain tube, Type 4646, with a large output current. Typical applications are discussed briefly.

621.384.5 : 621.316.722 **267**

The Characteristics of Some Miniature High-Stability Glow-Discharge Voltage Regulator Tubes.—G. Grimsdell : F. A. Benson. (*J. sci. Instrum.*, Sept. 1952, Vol. 29, No. 9, p. 301.) Comment on 1772 of 1952 and author's reply.

621.385 : 537.525.92 **268**

Space-Charge-Wave Propagation in a Cylindrical Electron Beam of Finite Lateral Extension.—P. Parzen. (*Elect. Commun.*, Sept. 1952, Vol. 29, No. 3, pp. 238–242.) Reprint. See 2371 of 1952.

621.385 : 621.392.21 **269**

Transmission-Line Tubes.—V. J. Fowler. (*Proc. nat. Electronics Conf., Chicago*, 1951, Vol. 7, pp. 318–330.) Proposals are advanced for the development of a new type of wide-band amplifier valve whose construction is based on that of amplifiers with distributed amplification, which are here termed 'chain amplifiers'.

621.385.004(083.75) **270**

General Considerations in Regard to Specifications for Reliable Tubes.—C. R. Knight & K. C. Harding. (*Proc. Inst. Radio Engrs*, Oct. 1952, Vol. 40, No. 10, pp. 1207–1210.) Proposals are made for a specification accurately describing the characteristics of the finished product and including details of inspection and acceptance tests. The lot-acceptance system, in conjunction with an adequate sampling procedure, is considered preferable to 100% screening.

621.385.004.15 **271**

Concerning the Reliability of Electron Tubes.—M. A. Acheson & E. M. McElwee. (*Proc. Inst. Radio Engrs*, Oct. 1952, Vol. 40, No. 10, pp. 1204–1206; *Sylvania Technologist*, April 1951, Vol. 4, No. 2, pp. 38–40.) Analysis of various causes of valve failure, serving to indicate the direction of research and development programmes for production of reliable valves.

621.385.004.15 **272**

Technique of Trustworthy Valves.—E. G. Rowe. (*Proc. Inst. Radio Engrs*, Oct. 1952, Vol. 40, No. 10, pp. 1166–1177.) Condensed reprint. See 1776 of 1952.

621.385.029.6 **273**

Low-Noise Travelling-Wave Tube.—A. G. Peifer, P. Parzen & J. H. Bryant. (*Proc. nat. Electronics Conf., Chicago*, 1951, Vol. 7, pp. 314–317; *Elect. Commun.*, Sept. 1952, Vol. 29, No. 3, pp. 234–237.) Outline description of a valve operating in the range 4·2–5·2 kMc/s, with a gain of 15 db, an output power of 0·5 mW, and a noise figure of 10 db. A new version under construction will have a gain of about 30 db, with a comparable noise figure.

621.385.029.6 **274**

Some Recent Developments in Travelling-Wave Tubes for Communication Purposes.—J. H. Bryant, T. J. Marchese & H. W. Cole. (*Proc. nat. Electronics Conf., Chicago*, 1951, Vol. 7, pp. 299–303; *Elect. Commun.*, Sept. 1952, Vol. 29, No. 3, pp. 229–233.) Basic features of travelling-wave valves are reviewed and typical performance characteristics and applications are considered. A new type of valve is described which is of rugged construction and is specially designed for ease of handling. The operating frequency range is 5·9–7·1 kMc/s, gain 25 db, power output 10 W and operating voltage 1·2 kV, this voltage permitting a reduction of overall length. An experimental permanent-magnet system for this type of valve is shown.

621.385.029.6 **275**

Experimental Low-Noise Amplifying Valve. — G. Convert. (*Ann. Radioélect.*, July 1952, Vol. 7, No. 29, pp. 225–234.) Theory serving as a basis for the development of a low-noise travelling-wave valve is presented, and an experimental valve is described with an optimum noise factor of 10 db and a gain of 14 db. The gain is a linear function of the cube root of the beam current. Variation of the noise factor as a function of the voltage of the second anode and as a function of the beam current is shown in curves. Pertinent theory has previously been published (2397 of 1952).

621.385.029.62/.64 276

Generalities on Travelling-Wave-Valve Feedback Self-Oscillators. Theory of the Reflex Travelling-Wave Valve.—M. Denis. (*Ann. Radioélect.*, July 1952, Vol. 7, No. 29, pp. 169–188.) The properties of travelling-wave-valve oscillators with internal or external e.m. feedback are reviewed, and the advantages of using high-dispersion transmission lines to obtain satisfactory operation of such oscillators are discussed. Theory indicates that optimum performance should be obtained with a line having minimum abnormal dispersion, i.e. one in which the group velocity is negative and equal to the velocity of light. Such lines are best used in oscillators of the type termed 'carcinotron' [3616 of 1952 (Guénard et al.)]. A tentative theory is developed for the reflex type of travelling-wave valve, using a method of exposition similar to that of Bernier (2974 of 1947) for the normal type. Reflex valves have a wide range of electronic tuning and are easily controlled by adjustment of anode and repeller voltages. Results obtained with an experimental reflex valve confirm the theoretical predictions except as regards power output, which was much lower than expected.

621.385.029.63/.64 277

Low-Noise Traveling-Wave Amplifier.—R. W. Peter. (*RCA Rev.*, Sept. 1952, Vol. 13, No. 3, pp. 344–368.) Design considerations for the electron gun and the circuit of wide-band low-noise travelling-wave amplifiers are discussed, and detailed theory is given of the 'three-region' low-noise gun, which is the essential factor in the design of such amplifiers. The construction of a demountable travelling-wave valve is described and measurements on experimental valves are presented which show the dependence of the noise factor on various valve parameters. The type of travelling-wave amplifier developed can compete with the best crystal-mixer amplifiers as regards noise factor. The best performance obtained with a 500-V wide-band amplifier operating at frequencies near 3 kMc/s was a noise factor of 8·5 db with a 15-db gain.

621.385.029.63 278

Amplification by Space-Charge Waves in an Electron Beam acted on by Crossed Electric and Magnetic Fields.—R. Warnecke, H. Huber, P. Guénard & O. Doehler. (*C. R. Acad. Sci., Paris*, 18th Aug. 1952, Vol. 235, No. 7, pp. 470–472.) It has previously been shown [333 of 1950 (Warnecke et al.)] that amplification can result from the interaction of two beams, with different velocities, in crossed fields. Amplification is also found possible with a single beam in which a high space-charge density introduces differences of velocity. Theory for small-amplitude signals [1022 of 1951 (Warnecke et al.)] shows that the gain is proportional to the relative change of velocity, which is calculable in terms of known parameters. Experimental arrangements are described with which a power output of 20 W, a gain of 15 db and an efficiency of 5% have been obtained at a frequency of 1·2 kMc/s.

621.385.029.64 279

The Excitation of Electromagnetic Fields by Current Waves.—H. Kleinwächter. (*Arch. elekt. Übertragung*, Sept. 1952, Vol. 6, No. 9, pp. 376–378.) Theory is developed from Maxwell's equations to show that e.m. fields corresponding to a plane or H_{01} wave can be excited by travelling current waves. The appropriate transverse current wave with phase velocity greater than the velocity of light is produced by a multielement waveguide arrangement as described in 2075 of 1952. Space-charge effects are not considered, hence the formulae apply only for weak fields and low-density beams.

621.385.029.64 280

Experimental Verification of Small-Signal Theory for the Travelling-Wave Valve with Helix.—H. Schnitger & D. Weber. (*Arch. elekt. Übertragung*, Sept. 1952, Vol. 6, No. 9, pp. 369–376.) Measurements were made to determine the extent to which wave velocity, helix characteristic impedance and valve gain are affected by enclosing the helix in a glass tube, which is a common practice. From the values of wave velocity found experimentally, the dependence of the gain on current and voltage variations was calculated and compared with measured values; good agreement was obtained. In the experimental wide-band valves used, no extra attenuation was necessary to prevent self-oscillation. The measurements covered the wavelength range 10·5–30 cm.

621.385.032.213 : 621.396.822 281

Space-Charge Reduction of Low-Frequency Fluctuations in Thermionic Emitters.—T. B. Tomlinson. (*J. appl. Phys.*, Aug. 1952, Vol. 23, No. 8, pp. 894–899.) A report of experiments to investigate the difference between the space-charge reduction factor for flicker noise and for shot noise. An indirect method is described, using a triode (*a*) connected as a retarding-field diode, and (*b*) with normal connections. By making one pair of measurements at low frequency and another pair at high frequency it is shown that the reducing effect of space charge is greater for flicker than for shot noise.

621.385.032.213.1 282

The Transient Behaviour of Thermionic Filaments with Temperature-Limited Emission.—F. H. Hibberd. (*J. sci. Instrum.*, Sept. 1952, Vol. 29, No. 9, pp. 280–283.) The time lag of a filament in responding to variations of heating current is investigated; methods are described for calculating and for measuring its magnitude. The value of the thermal time constant depends on whether the power is supplied at constant voltage or at constant current, being somewhat greater in the former case. For W filaments its value may lie between about 0·03 and 0·3 sec.

621.385.032.213.1 : 536.2 283

Temperature Distribution at the End of a Hot Wire.—J. A. Prins, J. Schenk & J. M. Dumoré. (*Appl. sci. Res.*, 1952, Vol. A3, No. 4, pp. 272–278.) Analysis is presented leading to a more general solution than that obtained by Clark & Neuber (1024 of 1951). See also 375 of 1952 (Fischer).

621.385.032.216 284

Spectral Dependence of Thermionic Emission with Activation from (Ba-Sr)O Cathodes over the Visible Region.—T. Hibi & K. Ishikawa. (*Phys. Rev.*, 15th Aug. 1952, Vol. 87, No. 4, pp. 673–674.) A 750-W lamp was used with filters as a monochromatic light source to illuminate the cathode incorporated in a test diode. Results are shown in curves and discussed in relation to trapping levels.

621.385.032.216 285

The Decay and Recovery of the Pulsed Emission of Oxide-Coated Cathodes.—R. M. Matheson & L. S. Nergaard. (*J. appl. Phys.*, Aug. 1952, Vol. 23, No. 8, pp. 869–875.) Full paper. Summary abstracted in 2080 of 1950.

621.385.032.216 286

Industrial Applications of Semiconductors: Part 6—Oxide Cathodes.—S. Wagener. (*Research, Lond.*, Aug. 1952, Vol. 5, No. 8, pp. 355–362.) The properties of the semiconducting oxide coating are reviewed and performance levels achieved in modern valves are quoted. The structure and formation of the interface semiconductor are discussed and its effect on valve performance and life is considered.

621.385.032.44 **287**

On Extending the Operating Voltage Range of Electron-Tube Heaters.—J. Kurshan. (*RCA Rev.*, Sept. 1952, Vol. 13, No. 3, pp. 300–322.) A survey of possible means of increasing the present voltage tolerance limits of ± 10% to ± 20% indicates that, even with a heater material having an extremely high temperature coefficient of resistance, ±16% is a fundamental physical limit.

621.385.12 : 621.318.572 **288**

Electronic Switching.—E. A. R. Peddle. (*Wireless World*, Oct. & Nov. 1952, Vol. 58, Nos. 10 & 11, pp. 421–423 & 465–468.) Part 1: Principles of the use of cold-cathode gas-discharge tubes. Part 2: Applications of the cold-cathode gas-discharge triode.

621.385.15 **289**

Theoretical and Experimental Investigation of Dynamic Secondary-Electron Multipliers.—H. Beneking. (*Z. angew. Phys.*, July 1952, Vol. 4, No. 7, pp. 258–267.) Design formulae for electron multipliers are derived and are confirmed by experiments which, among other things, indicate a value of 2–3 eV for the emission energy of the secondary electrons. Data for practical applications are presented graphically and the production and characteristics of multiplier tubes with Mg-MgO secondary-emissive anodes are described.

621.385.2 **290**

High-Frequency Diode Admittance with Retarding Direct-Current Field.—K. S. Knol & G. Diemer. (*Philips Res. Rep.*, Aug. 1952, Vol. 7, No. 4, pp. 251–258.) Formulae are developed for the susceptance of a planar diode with a negative anode voltage. The terms due to reflected electrons (total-emission susceptance) and electrons reaching the anode (exponential susceptance) are evaluated separately. Results obtained agree qualitatively with experimental findings.

621.385.2 : 537.525.92 **291**

The Space-Charge Smoothing Factor: Part 2.—C. S. Bull. (*Proc. Instn elect. Engrs*, Part III, Sept. 1952, Vol. 99, No. 61, pp. 319–320.) Digest only. Analysis showing a decided difference between fluctuation phenomena in diodes and in resistors. The mean-square deviation $\overline{\epsilon^2}$ of the conductance of a resistor is shown to be zero, so that even for signals as small and as rapidly changing as thermal fluctuations, its conductance is the same as that obtained by ordinary measurements. For diodes, experimental results indicate a value of $\overline{\epsilon^2}/N$ of about 0·5, where N is the number of fluctuating elements regarded as constituting the diode. In an ordinary case N is of the order of 10^4. Part 1: 2058 of 1951.

621.385.2.032.21 : [537.212 + 537.525.4 **292**

Relation between the Phenomenon of Sparking and the Electric Field at the Surface of a Cathode.—H. Bonifas. (*Bull. Soc. franç. Élect.*, Oct. 1952, Vol. 2, No. 22, pp. 553–554.) Determination of the e.s. field at the surface of the cathode, using equations previously developed (2084 of 1952), enables an explanation to be given of the mechanism of sparking and of the detachment of cathode active material in rectifying diodes. Discussion indicates that in order to avoid these effects, the cathode surface should be as smooth as possible, with uniform emissivity and low transverse resistivity.

621.385.3/.4 **293**

High-Transconductance Tubes for Broad-Band Telephone System Uses.—G. T. Ford & E. J. Walsh. (*Proc. nat. Electronics Conf., Chicago*, 1951, Vol. 7, pp. 304–313.) See 1165 of 1952.

621.385.5 **294**

Screen Dissipation of Pentodes.—S. C. Dunn. (*Wireless Engr*, Nov. 1952, Vol. 29, No. 350, pp. 309–310.) It is suggested that the usual information and characteristics supplied by valve manufacturers should be supplemented by a statement of optimum screen voltage. The influence of this parameter on the available working area in the I_a-V_a plane is discussed.

621.387 **295**

Investigation of Gas-filled Valves for Industrial Applications.—C. Biguenet & M. Vassilian. (*Le Vide*, May 1952, Vol. 7, No. 39, pp. 1191–1196.) Thyratron action is analysed and the suitability of Xe as a gas filling is illustrated.

621.387.42 : 621.318.57 **296**

Gas-Filled Counter and Switching Valves.—H. Harmuth. (*Elektrotech. u. Maschinenb.*, 1st July 1952, Vol. 69, No. 13, pp. 310–313.) Description of a neon-filled counter tube essentially similar to that of Lamb & Brustman (266 of 1950), and of suitable decade-counter and switching circuits using such valves.

621.396.615.141.2 **297**

The Possibility of Generating Millimetre Waves with Pulsed Multislot Magnetrons of the Rising-Sun Type.—W. Praxmarer. (*NachrTech.*, Sept. 1952, Vol. 2, No. 9, pp. 277–282.) Analysis is presented which leads to the production of diagrams determining oscillation range, operation conditions, and also mechanical dimensions. A second oscillation range is found below 5-mm wavelength, and with suitable mechanical design it appears possible to reach a wavelength of 0·8 mm.

621.396.615.141.2 : 621.317.332 **298**

Conductance Measurements on Operating Magnetron Oscillators.—Nowogrodzki. (See 157.)

MISCELLANEOUS

621.3 : 061.4 **299**

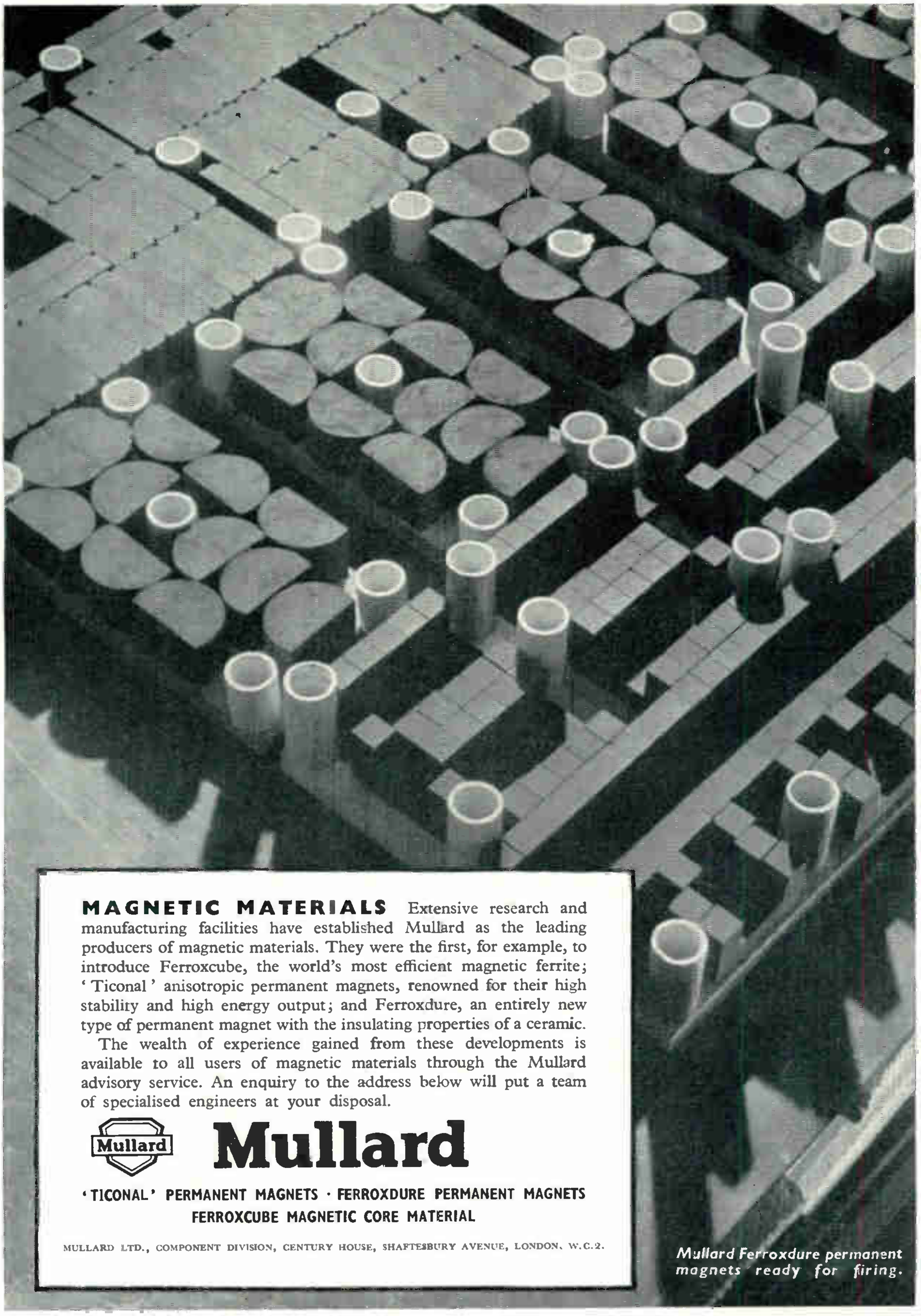
Radio Technology and Electroacoustics.—W. Althans. (*Z. Ver. dtsch. Ing.*, 1st July 1952, Vol. 94, No. 19, pp. 632–636.) Review of equipment shown at the 1952 Technical Fair, Hanover, including radio and television receivers, high-power h.f. transmitting valves, portable R/T sets, navigation aids, magnetic recorders, u.h.f. measurement equipment, etc.

621.39 : 061.4 **300**

Radio Exhibition Review.—(*Wireless World*, Oct. 1952, Vol. 58, No. 10, pp. 384–397.) Detailed discussion of trends in the design of television and broadcast receivers, c.r. tubes and valves, as exemplified in exhibits at the 19th National Radio Exhibition in September 1952. Other equipment reviewed includes a new design of metal-cone loudspeaker and a stencil-cutting system for reproducing pictorial matter. See also 3539 of 1952.

621.396/.397 : 061.4 **301**

The British National Radio Exhibition.—L. Carduner. (*Audio Engng*, Oct. 1952, Vol. 36, No. 10, pp. 12..81.) An American's impressions of the show.



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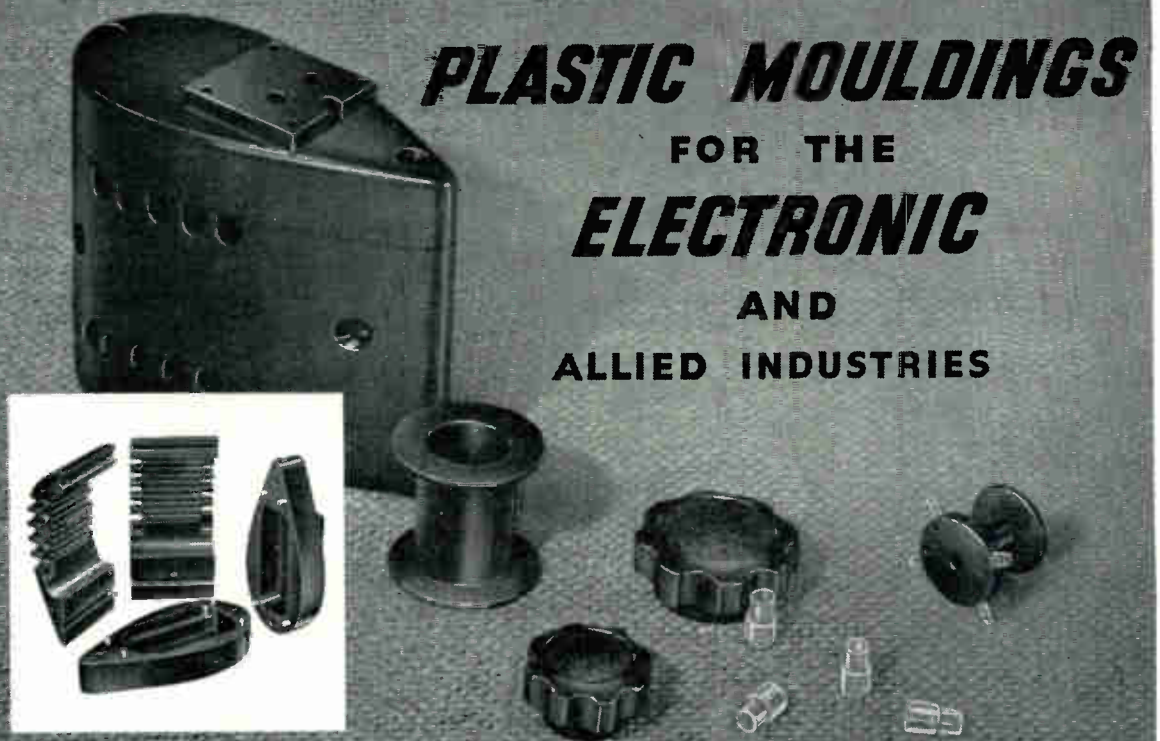

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