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DECEMBER 1953

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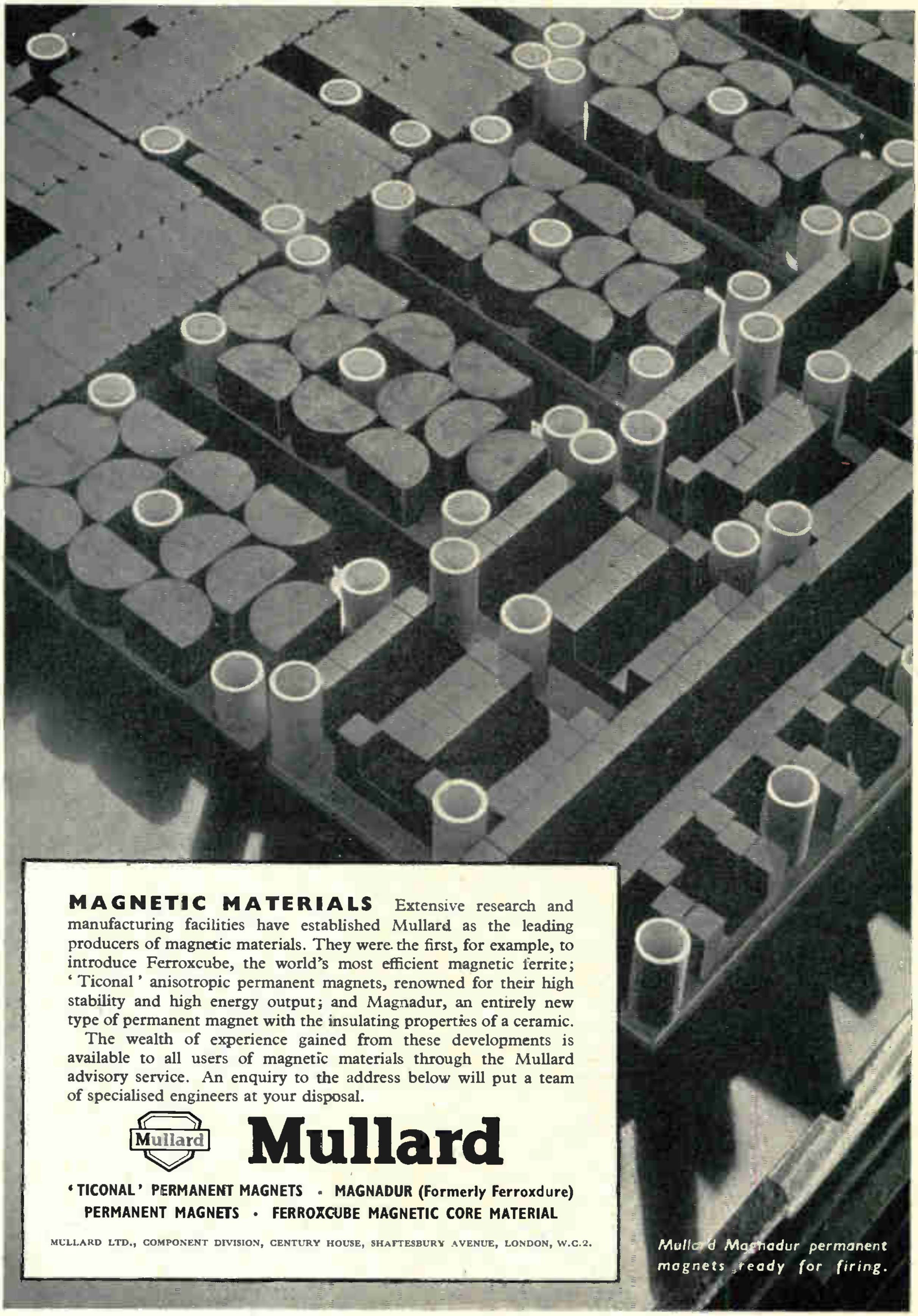

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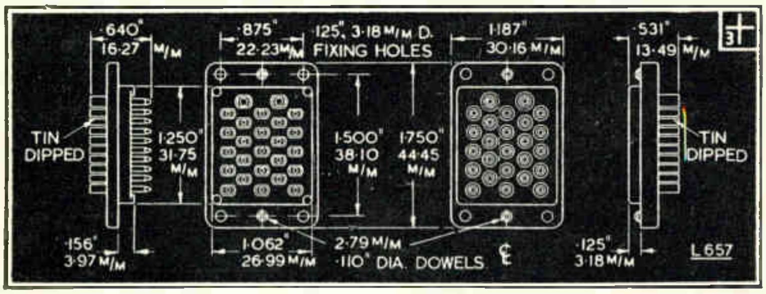

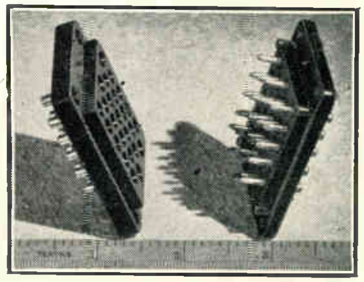

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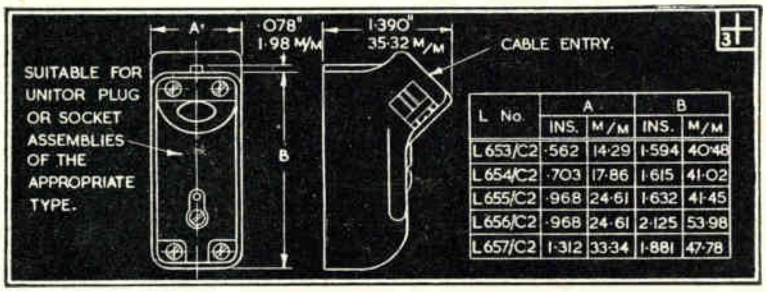

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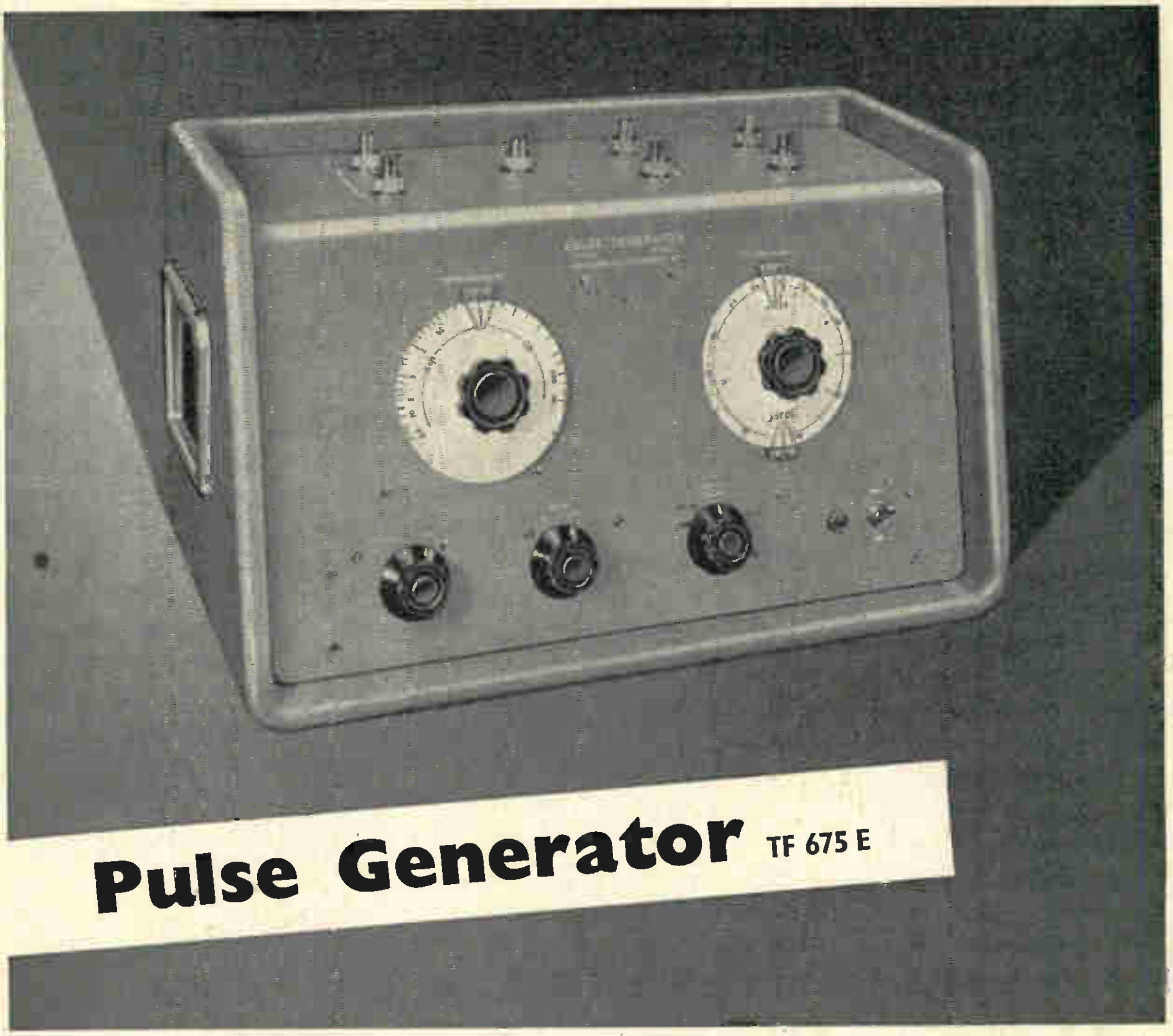

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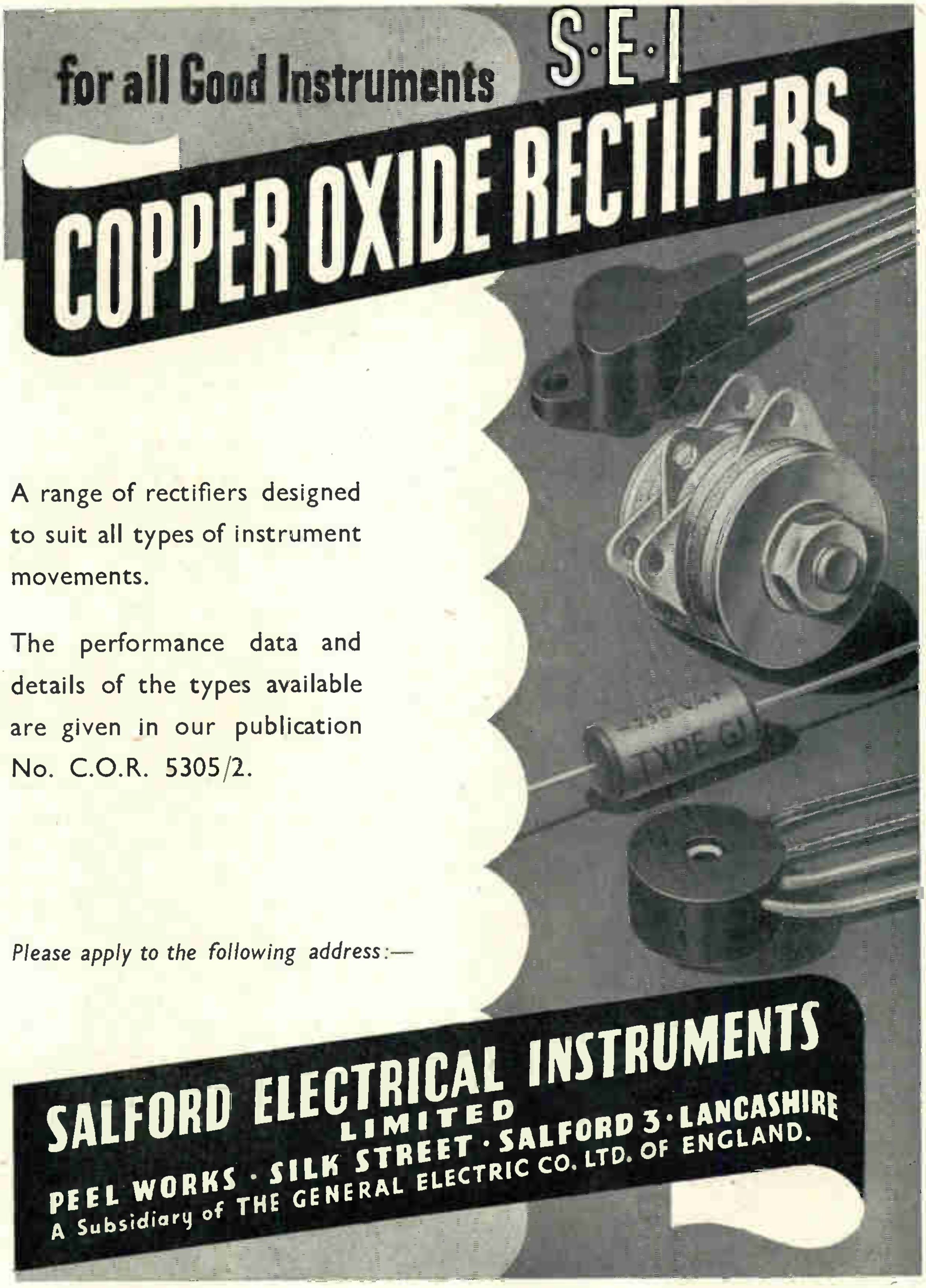
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WIRELESS ENGINEER

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What is the Meaning of Total Rationalization?

IN the September Editorial we referred to an article entitled "Dimensional Analysis, Units and Rationalization" by R. Vermeulen, which had appeared in *Philips Research Reports*. In an appreciative letter the author has called our attention to an article by Max Landolt and Jan de Boer in the *Revue Générale de l'Electricité* of December 1951 entitled "Quelle est la signification de la rationalisation totale?" which we have found to be a translation from a report which they submitted in January 1952 to the Committee on "Electric and magnetic magnitudes and units" of the International Electrotechnical Commission." After studying this in both languages we feel inclined to repeat the sentence with which we concluded the September Editorial, viz. "It shows that in some quarters there is still much uncertainty and confusion."

Beginning with the fundamental statement that a physical quantity or magnitude = a measure number × a unit, the authors discuss the question whether, in the total rationalization of a system of units, the changes should be made in the units themselves or in the physical quantities; they say that they favour the latter procedure. In symbols

$$G = \{G\}_a\,[G]_a \text{ and } G = \{G\}_b\,[G]_b$$

in which G is the quantity (grandeur), $[G]_a$ the unit of the quantity and $\{G\}_a$ a pure number. If the unit is changed from $[G]_a$ to $[G]_b$ then the number will be changed from $\{G\}_a$ to $\{G\}_b$, where

$$\frac{\{G\}_b}{\{G\}_a} = \frac{[G]_a}{[G]_b}.$$

Now the question, to which the authors devote fifteen pages of discussion, is whether the process of rationalization should involve a change in $[G]$ or in G. We disagree entirely with the following statement: "Often the defining equation of a quantity contains a numerical factor as for instance the quantity-equation defining the kinetic energy. Kinetic energy is defined as $\frac{1}{2}mv^2$, the introduction of the factor $\frac{1}{2}$ in this defining equation is a matter of pure convention. If one omits this factor, one defines another quantity, and the conservation law would read in that case; the sum of potential energy and half the kinetic energy is constant. So in fact one may often define more physical quantities corresponding to the same physical phenomenon." This is surely fundamentally wrong; starting with the body of mass m at rest, the work done by a force in accelerating it to a velocity v, and therefore stored in the body as kinetic energy can be proved by any engineering student to be equal to $\frac{1}{2}mv^2$, and this is the energy dissipated as heat if the body, in the form of a bullet, strikes a target and is thus brought to rest. If the authors wish to introduce the energy quantity mv^2 they should not call it the kinetic energy, as this can only lead to confusion. It is as if one defined the height of a man as the length of his shadow on a wall, cast by a source of light as far from him on one side as the wall is on the other. This might be a useful quantity, but calling it the height of the man could only lead to confusion. When new quantities are introduced, new names should be given to them; similarly with units. We admit that the general conception of many of the electromagnetic quantities is by no means so definite as that of kinetic energy or height, but we feel sure that any attempt to change what is now well established as the meaning of the strength of the magnetic field would lead to great confusion, especially as we have already four systems of units in which it can be expressed.

The authors cause much confusion by applying the name "oersted" both to the unrationalized and to the rationalized unit of magnetic-field strength. They state that the field intensity at the centre of a ring 1 m diameter carrying a current of 1 A is

1 A/m in the rationalized Giorgi system, or
0·001 Oe in the rationalized c.g.s.e.m. system, or
$4\pi/1000$ Oe in the non-rationalized c.g.s.e.m. system.

They are thus applying the name Oersted to two units differing by a ratio of 4π to 1. The name was originally given to the unrationalized unit and, perhaps unfortunately, is almost universally so applied; to apply it also to the rationalized unit is a dangerous procedure. Fortunately the unit in the rationalized Giorgi system is 1 A/m. They might have distinguished between the two oersteds by referring to the rationalized one as a "ratoe."

The relation between the units of H in the different systems is as follows:—

unrat. c.g.s.e.m.....1 oersted $= 1000/4\pi$ A/m
rat. c.g.s.e.m.....4π oersteds $= 1000$ A/m
unrat. m.k.s.....10^{-3} oersted $= 1/4\pi$ A/m
rat. m.k.s...$4\pi \times 10^{-3}$ oersted $= 1$ A/m.

To calculate the value of H at the centre of the ring 1 m diameter our older readers would write $H = 2\pi I/10r = 2\pi/(10 \times 50) = 4\pi/1000$ oersted, which we see is exactly equal to 1 A/m. Since the rat. c.g.s. unit is equal to 4π oersteds, the number of units will be reduced in this ratio (i.e., to 0·001 unit) *but not to* 0·001 *oersted.*

On page ten of the Report the authors state with respect to this example that "Direct calculation using the definition of the ampere and the oersted gives 1 A/m = 1/1000 Oe." This as we have seen is incorrect but would be correct if written 1 A/m = 1/1000 ratoe, where 1 ratoe = 4π Oe. The authors are aware of this and say that it would be better to have two different names and two different symbols as proposed by Giorgi and Blondel "but as it can be expected that the rational magnetic-field intensity will be used more and more, the irrational field intensity will gradually disappear and no special name is needed."

The most satisfactory thing about this report is that, although the authors expressed their preference for the rationalization of the quantities and not the units, they finally give a Table of units agreeing, so far as we can see, in every detail with the generally recognized rationalized units, both Giorgi and c.g.s.e.m., except that their oersted is a rationalized oersted. In both the English and French versions they give 1 C/Nm^2 as the Giorgi unit of the dielectric constant ϵ. This is presumably a mistake for 1 C^2/Nm^2 in which C = coulomb, N = newton and m = metre.

Since writing the above we have received a copy of an article by Professor H. König of Berne entitled "Über die Mehrdeutigkeit des Grössenbegriffs" which appeared in the Bulletin of the Swiss Elektrotechnische Verein for 1950. The author discusses the different meanings that can be attached to the concept of magnitude or quantity. He distinguishes between the point of view of the *Synthetiker*—the systematic, theoretical thinker who says that electric-field strength electrostatically defined is not equal to the same field strength electromagnetically defined—and the *Realist* or practical man who says that field strength is field strength, and the number of units is inversely proportional to the size of the unit. We hope that this will act as a deterrent to any reader thinking of becoming a synthetiker.

G. W. O. H.

INDEXES

The Index for 1953 to the editorial pages of *Wireless Engineer* will be included in the March 1954 issue, in which there will also be the Index to Abstracts and References published during 1953, and a list of journals scanned for abstracting, with their publishers' addresses.

The March 1954 issue, which will include the normal editorial pages, as well as the Indexes, will be priced at 6s. The Index pages will be detachable for binding with the 1953 volume.

ELECTRONIC A.C. POTENTIOMETER

By L. Tasny-Tschiassny, Dr. Tech., M.I.E.Aust.

(Electrical Engineering Department, University of Sydney, Sydney, Australia.)

1. Introduction

THE increasing importance of the accurate determination of the phase angle between two alternating voltages is evident from the large number of relevant papers published recently.[1–12] A simple, quickly-working, and accurate electronic potentiometer serving this purpose can be built from standard radio components. A phase-angle range of 360° is covered unequivocally. A voltage range between a few millivolts and several hundred volts and an accuracy of a fraction of 1% can be easily obtained with the ordinary precautions as to screening, leakages, etc.

The potentiometer compares the magnitudes of an unknown voltage and an adjustable and known fraction of a reference voltage. The quick determination of phase angles by this procedure is explained in Section 3. The comparison is carried out with the aid of an electronic square-law signal current, the ratio of the percentage change of the alternating voltage to the percentage change of the needle deflection becomes less than 0·5, the figure holding true for a zero-signal zero-deflection square-law detector.

The operation of the potentiometer is independent of the frequency, the value of the frequency need not even be known.

2. Voltage-Divider Circuit

Fig. 1 shows the principle of the voltage-divider circuit. The component values correspond to two models successfully operated in the audio-frequency range. If the balancing switch is in the 'Work' position, the three leads to the 2-pole 6-position 'Selector Switch' and marked '$k_1k_2V_x$', '$c_1c_2V_r$', and 'uV_r' respectively are on potentials to E corresponding to their markings. The operation of the 'Quadrant Switch' changes the phase of the reference voltage by 180°. The 5-kΩ uncalibrated

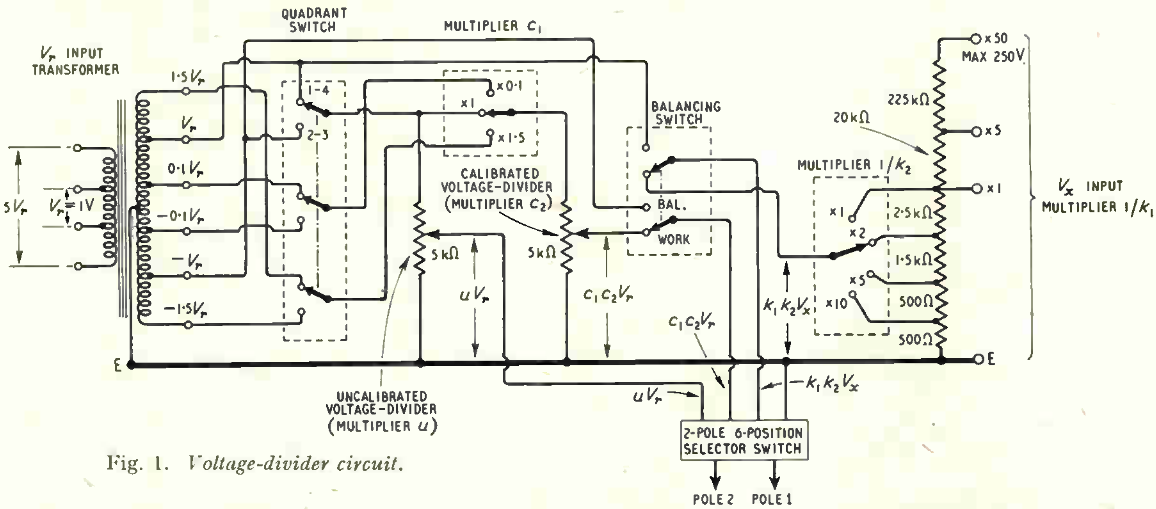

Fig. 1. *Voltage-divider circuit.*

detector working near the full range of its d.c. indicating meter. If the latter is properly damped, for instance, by the parallel connection of a capacitor of high capacitance,* and the change-over switch is of quick action and operated several times, minute differences in the deflections, such as 0·2 %, can be comfortably detected. If the d.c. meter is backed off to a negative zero-

* The author used an electrolytic capacitor of 500 μF capacitance and 12 V d.c. peak voltage. The small inverse voltage that may occur during the operation does not affect the capacitor.

MS accepted by the Editor, March 1953

voltage divider is a carbon potentiometer, to one side of which a 50-kΩ carbon variable resistor is connected in parallel to obtain a finer control. The phase shift introduced by it is negligible. The calibrated voltage divider is a three-stage Kelvin-Varley slide[13] of 5 kΩ total resistance, consisting of 11 resistors of 500 Ω, 11 resistors of 100 Ω, and a linear potentiometer of 200 Ω, all wire-wound. Small, and even different, phase angles of the individual resistances are irrelevant, as long as the scalar accuracy of the voltage division is not affected. The calibrated voltage divider can be made accurate to 0·1 % of the full voltage without any particular difficulty. For small unknown voltages the 'Multiplier' $c_1 = 0{\cdot}1$ increases by one the number of digits read-off on the calibrated voltage divider. The Multiplier $c_1 = 1{\cdot}5$ may be necessary when the phase-angle between the unknown and reference voltages is between 60° and 120° or between 240° and 300°. The pole connections of the Selector Switch are shown in Table 1.

TABLE 1

Pole Connections of the Selector Switch.

Position	1	2	3	4	5	6
Pole 1 ..	uV_r	$k_1k_2V_x$	E	uV_r	$c_1c_2V_r$	E
Pole 2 ..	E	E	$c_1c_2V_r$	$k_1k_2V_x$	$k_1k_2V_x$	E

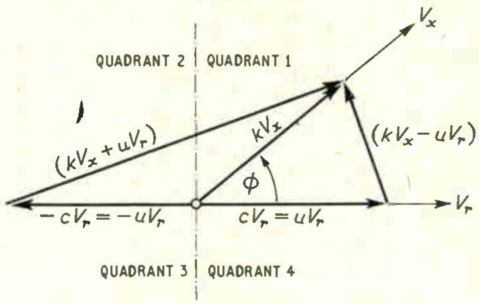

Fig. 2. *Vector diagram of voltages.*

If the drain caused by the reference voltage input transformer and/or the 250-kΩ voltage divider for V_x is inadmissibly high, one or two impedance converters, such as cathode followers or other 100 % feedback amplifiers, can be inserted before the relevant input terminals. If the frequency is so high that the voltage ratio at the tappings of an iron-core transformer can no longer be relied upon, an electronic phase inverter in conjunction with resistive voltage dividers can be used instead of the transformer.

3. Principle of Magnitude and Phase Measurements

By suitable selection of the values c_1, k_1 and k_2 and proper adjustment of the values c_2 and u the magnitudes of the potentials to E of the leads '$k_1k_2V_x$', '$c_1c_2V_r$', and 'uV_r' are made equal. The accessories with the aid of which this is done are described in Section 4. Then

$$\frac{|V_x|}{|V_r|} = \frac{c_1c_2}{k_1k_2} \quad \ldots \quad \ldots \quad (1)$$

With the aid of an adding network (see Section 4) and with the aid of the Quadrant Switch, Fig. 1, voltages $(k_1k_2V_x \pm uV_r)$ to E are produced. The magnitude of the smaller of these voltages (Fig. 2) is, with the aid of the adjustments c_1 and c_2, made equal to a known part $(c_1'c_2'V_r)$ of the reference voltage. Then

$$\phi = 2\sin^{-1}\left(\frac{c_1'c_2'}{2c_1c_2}\right)$$

Tables correlating once and for all $2\sin^{-1}(c_1'c'_2/2c_1c_2)$ and $(c_1'c_2'/c_1c_2)$ can be compiled. The position of the Quadrant Switch for the smaller of the two voltages $(k_1k_2V_x \pm uV_r)$ determines whether the voltage V_x is, with reference to the voltage V_r, either in quadrants 1 or 4 or in quadrants 2 or 3. The decision regarding the actual quadrant can be made by a modification of the adding network in a way that voltages $(k_1k_2V_x \pm uV_r\epsilon^{j\omega\alpha})$ and $(k_1k_2V_x \pm uV_r\epsilon^{-j\omega\alpha})$ are formed where α is a small angle not necessarily known. If the Quadrant Switch is always left in the position in which the angle between $k_1k_2V_x$ and $\pm uV_r$ is between 0° and 90° or 270° and 360°, the switch producing the phase shifts α and $(-\alpha)$ can be marked once and for all 'Quadrants 1-3' and 'Quadrants 2-4', because the relative magnitudes of the voltages $(k_1k_2V_x \pm uV_r\epsilon^{j\omega\alpha})$ and $(k_1k_2V_x \pm uV_r\epsilon^{-j\omega\alpha})$ make this distinction. Preferably the positions are marked in such a way that the markings refer to the smaller voltage.

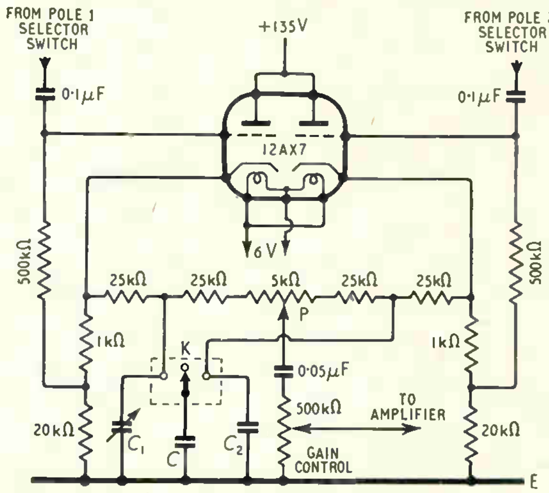

Fig. 3. *Adding and phase-shifting network.*

4. Adding and Phase-Shifting Network

Fig. 3 shows the principle of the adding and phase-shifting network with component values suiting those of Fig. 1. Its performance will be analyzed mathematically in Section 6. Two cathode followers operated from a regulated power supply and with d.c.-heater supplies form the link between the output of the Selector Switch and the network proper. Their input impedance is high enough not to affect the correctness of the voltage division by the voltage-divider network of Fig. 1. C, with alternative values of 10,000 pF, 2,000 pF, 500 pF and 100 pF, according to the frequency, is the phase-shifting capacitor producing an apparent phase shift of the reference voltage by an angle α or $(-\alpha)$, according to whether the key K is turned to the right or to the left-hand side. C_1 and C_2 are small trimmer capacitors. The output of the 0·5-MΩ gain-control potentiometer leads through a suitable low-hum amplifier (with a good short time stability of gain) into the square-law detector.

5. Notes Regarding Harmonics

If the phase angle between two voltages is defined as the phase angle between their fundamentals, a square-law detector considerably reduces the influence of a moderate percentage of harmonics. For instance, the influence of 5% harmonics on the detector deflection is only 0·25%.* If the phase angle between the unknown and reference voltages is small, the percentage of harmonics in the difference voltage may become considerable. One method to avoid errors is the insertion of a low-pass filter between the amplifier and detector. Another way is to determine, for a certain frequency once and for all, the phase angle—preferably a few degrees—introduced by a certain phase-shifting capacitor C (Fig. 3). By its insertion the phase angle between two voltages is increased by a known amount and the percentage content of harmonics in the difference-voltage reduced.

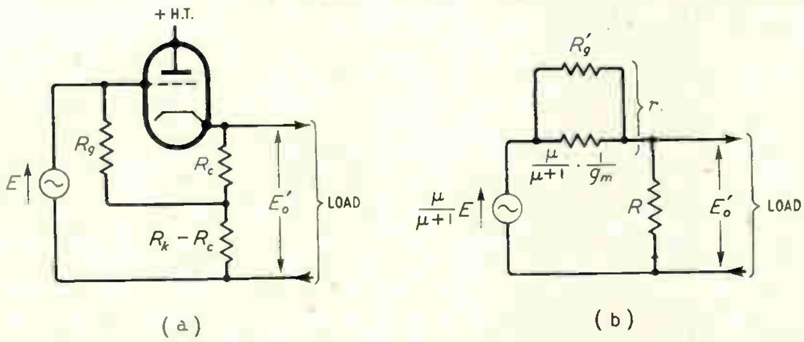

Fig. 4. *Equivalent circuit of a cathode follower.*

6. Analysis of the Cathode-Follower and Adding Network. Balancing

The author is unaware of any published analysis of the cathode-follower circuit Fig. 4(a) in which the value of the grid leak R_g is finite. It can be shown that the a.c. behaviour of the circuit can be represented by the equivalent circuit of Fig. 4(b) in which

$$R_g' = \frac{\mu}{\mu+1}\cdot\left(R_g\cdot\frac{R_k}{R_k-R_c}+R_c\right) \quad \ldots \quad (3)$$

$$R = \frac{R_k(1+R_c/R_g)-R_c^2/R_g}{1-(1/\mu).[(R_k-R_c)/R_g]} \quad \ldots \quad \ldots \quad (4)$$

It is well known that for cathode followers there is little change of $\mu/(\mu+1)$ even for a considerable change of μ, if μ is high, because $d(\mu/\overline{\mu+1}) = d\mu/\mu^2$. If, for instance, $\mu = 100$, a 10% change of μ (i.e., a change exceeding the values met with in practice) will produce only 0·1% change of $\mu/(\mu+1)$. Hence, since $R_g \gg R_k - R_c$, R_g' and R can be considered as being constants for all practical purposes. Consequently, the only quantities in Fig. 4(b) that may vary appreciably during the life of a valve, are g_m and r, i.e., the resistance of the parallel combination of $\mu/\{(\mu+1)g_m\}$ and R_g'.

Fig. 5 is the equivalent circuit of the adding network, if the two cathode followers are replaced by their equivalent circuits (the subscripts 1 and 2 are added to the symbols) and if the trimmer capacitors C_1 and C_2 and the phase-shifting capacitor C are left out. The analysis of Fig. 5 for the output voltage E_0 leads to

$$E_0 = \frac{E_1\dfrac{\mu_1}{\mu_1+1}\cdot\dfrac{1}{r_1+Z_1(1+r_1/R_1)} + E_2\dfrac{\mu_2}{\mu_2+1}\cdot\dfrac{1}{r_2+Z_2(1+r_2/R_2)}}{\dfrac{1}{Z_0}+\dfrac{1+r_1/R_1}{r_1+Z_1(1+r_1/R_1)}+\dfrac{1+r_2/R_2}{r_2+Z_2(1+r_2/R_2)}} \quad \ldots \quad \ldots \quad (5)$$

If, in this equation, all impedances are assumed to be purely resistive, the multipliers of E_1 and E_2 can be made equal by proper adjustment of Z_1 and Z_2; i.e., of the 5-kΩ potentiometer P (Fig. 3). Then E_0 is proportional to $(E_1 + E_2)$. If one of the voltages E_1 or E_2 is zero, E_0 is proportional to E_2 or E_1 respectively. In the positions 1, 2, and 3 of the Selector Switch individual voltages, and in the positions 4 and 5 sums or differences of individual voltages, are measured. In position 6 no voltage is applied.

If Z_1 and Z_2 are great compared with r_1 and r_2, the influence of changes of g_{m_1} and g_{m_2} from

* Caution is necessary if the voltages may contain even harmonics and the detector responds to the r.m.s. value of one half-wave. The author employed a full-wave square-law detector.

initial values, reflected by changes of r_1 and r_2 is small. For the component values given in the figures, a maximum error of less than 0·1% can be expected.

The presence of the capacitors C_1, C_2 and C can be accounted for by a star-delta transformation. Its result is the insertion of series inductances in the branches Z_1 and Z_2 and of high, almost purely capacitive, reactances in parallel with the branches R_1, R_2, and Z_0. For small phase-shifts the influence of the changes of Z_1 and Z_2 on equation (5) is equivalent to phase shifts of the voltages E_1 and E_2, whereas the influences of the changes of R_1, R_2, and Z_0 are negligible. Hence the capacitor C can be used as a phase-shifting component and the capacitors C_1 and C_2 as trimmers to compensate for unwanted small inductances and capacitances.

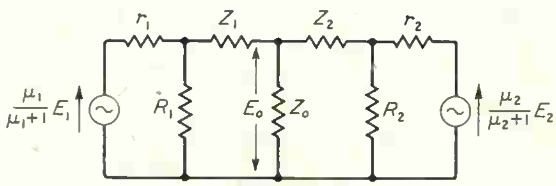

Fig. 5. *Equivalent circuit of the adding network.*

The balancing of the two numerator terms of equation (5) is carried out by applying two voltages $E_1 = -E_2$ to the cathode-follower inputs and adjusting the 5-kΩ potentiometer P and the trimmer capacitor c_1 to a minimum of the voltage E_0 as detected by the detector at full gain. The 'Bal.' position of the Balancing Switch and position 5 of the Selector Switch serve this purpose. Then the gain is reduced and, as a check, equal deflections of the detector must result for the positions 2 and 3 of the Selector Switch.

The input impedance Z_i of the cathode-follower circuit of Fig. 4(a) can be approximately computed by

$$Z_i \approx \frac{R_g}{1-(1-R_c/R_k)\,(E_0'/E_i)} \qquad \ldots \quad (6)$$

where E_0' is the output voltage under load. For the component values given, Z_i is negligible as a shunt to the voltage-divider circuits. The direct grid current of a valve 12AX7 under the operating conditions in question is between 0·2 and 1 μA, usually not exceeding 0·6 μA. For the weak a.c. signals applied, the reduction of the effective a.c. input impedance resulting from variations of the direct grid current during one a.c. cycle can be neglected.

7. Summary of the Operation of the Potentiometer

After some experience the work with the potentiometer is very rapid. Before a long series of measurements the potentiometer is balanced, as explained in Section 6; even after several hours work re-balancing proves to be unnecessary.

The measurements proper in the 'Work' position of the Balancing Switch comprise the following steps, in which the gain control is always adjusted to approximately the same deflection of the d.c. meter of the square-law detector near its full scale deflection.

(a) Select values k_1 and k_2 such that $k_1k_2V_x$ (position 2 of the Selector Switch) is just less than Vr (position 3 of the Selector Switch for $c_1 = 1$ and $c_2 = 1{\cdot}000$).

(b) Adjust u and c_2 to equal deflections in the positions 1, 2, and 3 of the Selector Switch. Record the values k_1, k_2, c_1, and c_2 and apply equation (1).

(c) With the Selector Switch in position 4 bring the Quadrant Switch into its position producing the smaller deflection.

(d) Adjust c_1 and c_2 to values c_1' and c_2', so that equal deflections result in the positions 3 and 4 of the Selector Switch. Apply equation (2) and the tables.

(e) In position 4 of the Selector Switch operate the key K. (Fig. 3). The quadrant is given by the figure appearing in both the lettering of the position of the key K which produces the smaller deflection, and the lettering of the Quadrant Switch.

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NOISE FACTOR OF CONVENTIONAL V.H.F. AMPLIFIERS

By N. Houlding, B.Sc. Tech.

(*Concluded from p. 290, November issue*)

4. Experimental Results

As far as is possible, the figures quoted are representative of average valves, except that particular curves may not be strictly typical of the average. For a narrow-band amplifier particularly, the optimum value of source conductance is dependent on the input coil losses and this should be remembered in making comparisons. Except where mentioned, *the results are with the input circuit tuned to resonance at mid-band*, including the space charge capacitance, and the wideband results were obtained with no additional capacitance in the grid circuit. All quoted results are for amplifiers of mid-band frequency 45 Mc/s.

4.1. *Accuracy of Equivalent Circuit.*

Fig. 12 shows curves of n against source resistance for four triodes. Experimental results are not in exact agreement with the theoretical expressions given in 2.6 and it has been concluded that there is a significant dicrepancy in the value of source resistance due to lead effects which accounts for much of the apparent error. It has been found that the theoretical formulae are very useful for calculating the effect of changes in any single parameter, and the simple formulae given in 2.4 have been found to be very accurate.

As might be expected, the results for wideband amplifiers in which the effect of induced grid noise is less due to the lower impedance of the input circuit are in better agreement with theory, and Table 2 shows the agreement between calculated figure of merit and noise factor in amplifiers 12 Mc/s wide at —3 db overall. In the table C_{strays} was taken as 2 pF.

TABLE 2

Valve type	Figure of merit $(C_{gs} + 2C_{ag} + C_{strays})\ R_{equ}$	Source resistance (ohms)	Noise factor (ratio) $B = 12$ Mc/s
Experimental	7,200	675	3·6
Experimental	6,550	675	3·2
CV139	5,050	360	2·5
6AK5 (triode)	4,500	650	2·4
Experimental	2,800	675	2·1
Experimental	2,100	675	2·0

4.2. *Variation over the Band and Feedback Effects*

Results for a common-grid circuit are shown in Fig. 13. The circuit used an unequal-Q coupling between the first two stages to minimize the deficiencies of the common-grid circuit. The input circuit was tuned to minimum noise output which should coincide with input resonance (excluding susceptance due to feedback) if the phase angle of the mutual conductance is 180°. The overall variation of u is appreciably increased by the second stage noise. The full-band noise factor calculated from the curves is 0·5 db greater than the mid-band value, and direct measurement confirmed this.

This particular circuit serves to illustrate the fallacy of using feedback to obtain the frequency response. With a capacitance of 5 pF added between grid and cathode, the input circuit frequency response was still very wide, but the full-band noise factor was increased by 0·15 db, showing the dependence of noise factor on the source impedance excluding feedback. Furthermore, a CV139 common-grid circuit for a 15-Mc/s bandwidth will give a satisfactory frequency response with a source resistance of 1,000 ohms (minimum n), yet the optimum full-band noise factor is obtained with 350 ohms and the noise factor is about 3 db worse with a 1,000-ohm source, due to the large effect of the increase in noise at the edges of the frequency band.

For one particular amplifier it was found that feedback damping was not altogether a disadvantage. The unit had a bandwidth for the first CV139 stage (input and anode circuits) of 4 Mc/s at 1 db. The CR value of the source was such that a maximum value of only 400 ohms at the grid could be used with a neutralized circuit. It was found that the same noise factor could be obtained with a common-grid circuit and an 800-ohm source at the cathode as with a common-cathode circuit using 470-ohm anode damping. However, the common-grid circuit used an unequal-Q anode circuit which was extremely critical both in alignment and need for low-loss primary coil, and use of a similar circuit in the input of the common-cathode stage would have shown the advantages of the common-cathode.

With a CV139 common-grid circuit, it was observed that a damping resistance in the anode, of the order of 20,000 ohms, caused a slightly greater increase in noise factor than with the same resistance in the grid circuit. It is greater because the effect of the second stage noise (considering the resistance as part of the first stage) is probably slightly worse with the anode damping, for the power gain is reduced and the term $(N_2 - 1)$ is not reduced as much. It was also found that damping in the anode circuit of the second stage of two CV139 common-grid circuits in cascade was at least as serious as the same damping in the input circuit. The effect of this damping in the anode of the second common-grid stage could be minimized by making the coupling between first and second stages a step-down transformer.

An example of utilizing feedback to give input damping is a circuit used for measurement of the noisiness of crystal mixers. It is desirable to use a circuit with low noise factor so that the experimental discrimination is good, and at the same time make the circuit such that both the power gain and noise factor are not critically dependent on source resistance. By using a common-cathode circuit with cathode lead inductance feedback, the input conductance was made equal to that value required for optimum noise factor. The noise factor was degraded slightly and the optimum source conductance was increased. The cathode-lead inductance required also produced a significant input susceptance, but this was just sufficient to balance the detuning susceptance required for optimum noise factor. (See 4.3.)

4.3. *Asymmetry*

The improvement obtained by capacitance detuning is shown in Fig. 12. With a common-grid circuit the effect is small, due no doubt to the fact that normal tuning to minimum noise output is slightly capacitive.*

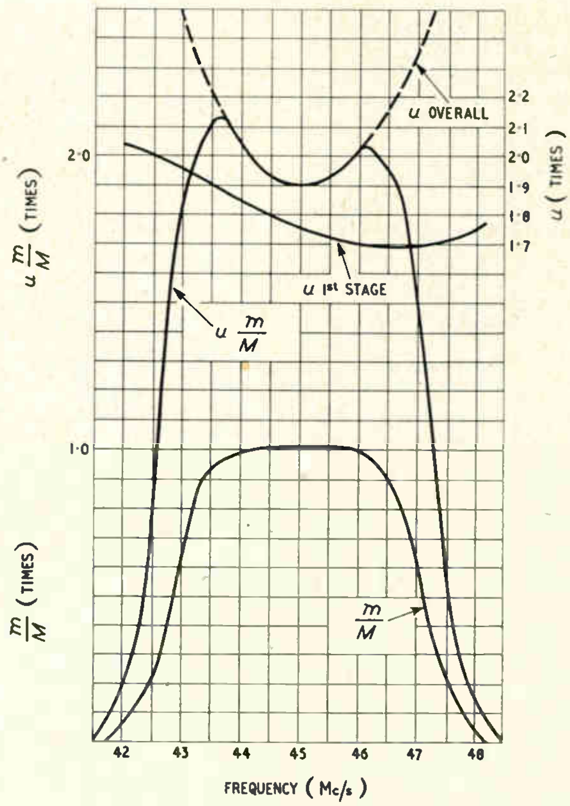

Fig. 13. *Variation of single-frequency noise factor over the response for CV139 common-grid circuit. CV139 single-stage feeding CV1091 pentode with unequal-Q coupling (3·9kΩ secondary damping).*

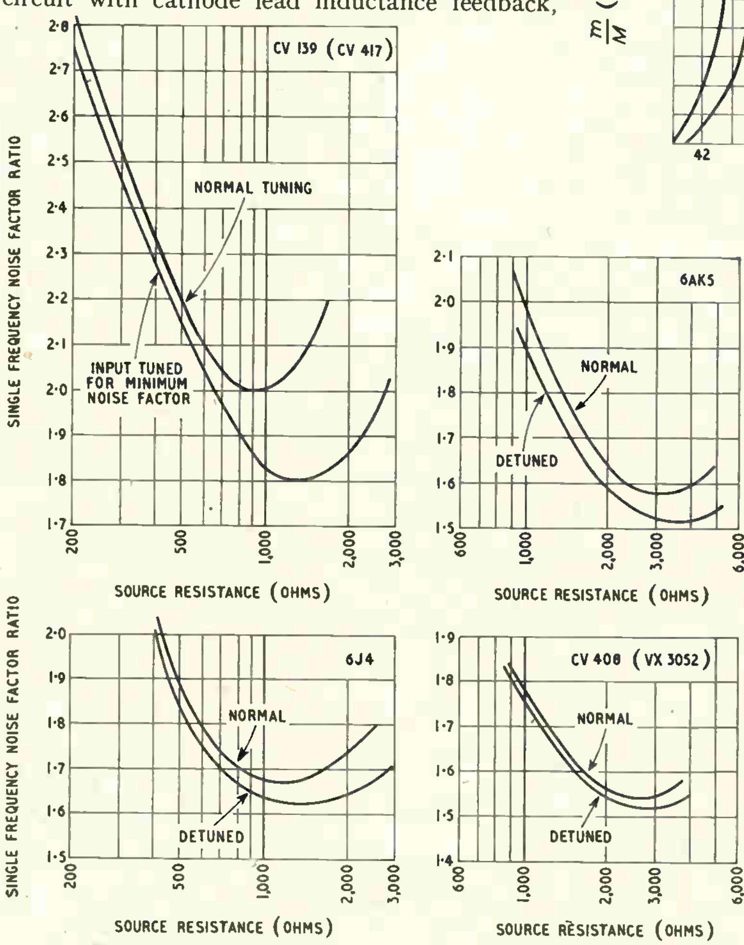

Fig. 12. *Effect of source resistance on mid-band noise factor.*

* Because of transit time the phase angle of the mutual conductance is less than 180°.

Table 3 gives results for a CV139 and also emphasizes the feedback effect of anode–cathode capacitance with the common-grid circuit.

With the common-cathode circuit the effect is more marked. Particular care was taken to check that faulty neutralization was not masking any true asymmetry by adjusting this with anode current flowing and feeding signal from anode to grid circuits. This gave almost exactly the same neutralization setting as the normal procedure. It was found that detuning the neutralizing circuit by the appropriate amount gave minimum detuned noise factor when the input circuit was tuned for a maximum signal in the normal way. For this, it is important to ensure that the anode circuit is resistive at mid-band, otherwise the feedback effect is more complex, and this should be remembered when designing a unit with stagger-tuned circuits.

Table 4 gives the magnitude of the effect with the same four triodes of Fig. 12 in a common-cathode circuit. The capacitance change required has been measured for a range of source resistance values and is constant at a given frequency. For most valves the change in capacitance due to space charge is about 1½ times the detuning required, so that tuning the input circuit with the valve biased to cut off gives approximately the correct 'detuned' condition (at 45 Mc/s). A more accurate detuning procedure can easily be established for any finished design.

The improvement in full-band noise factor is less. With a CV139 the improvement is about 0·3 db for 4-Mc/s bandwidth and 0·1 db for 15-Mc/s bandwidth.

The effect has also been observed with pentodes. With suitable cathode-lead inductance feed-

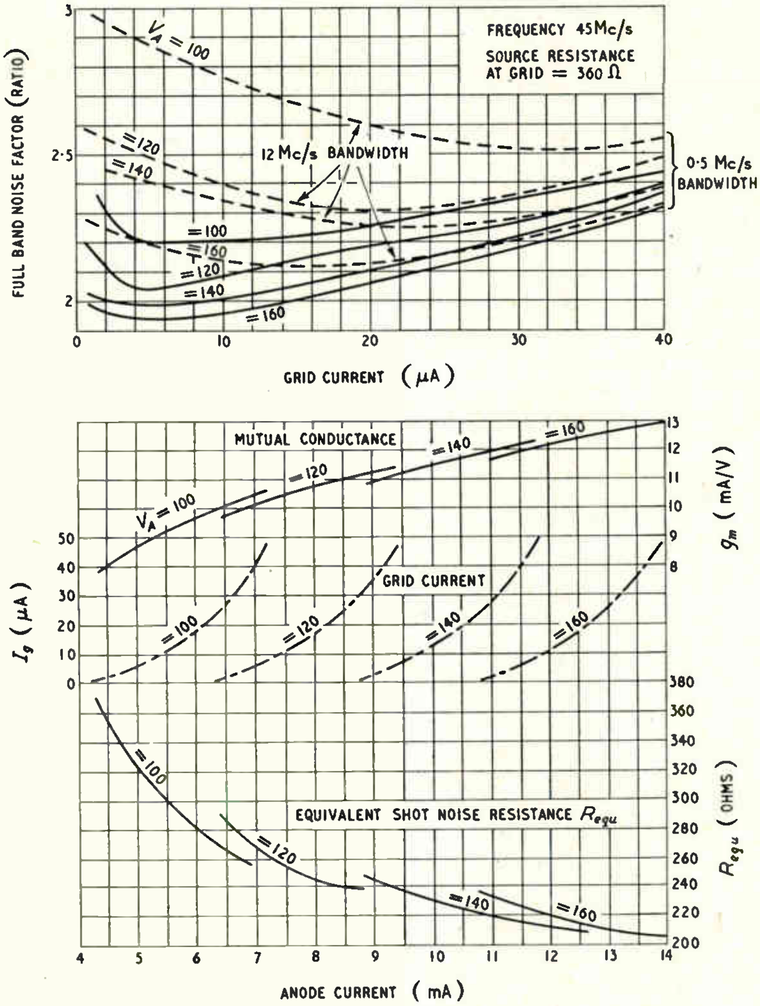

Fig. 14. *Performance of CV139 biased for grid current.*

TABLE 3

Asymmetry of Noise Factor with CV139 Common-Grid at 45 Mc/s. (Source resistance at cathode 1,000 ohms)

Tuning point	Capacitance increase at cathode (pF)	Change in mid-band noise factor (db)
Minimum noise	0	0
Minimum noise factor	+ 0·8	— 0·1
Maximum signal	+ 4·3	+ 0·65

TABLE 4

Asymmetry of Noise Factor with Common-Cathode Circuits at 45 Mc/s

Valve	Change in single-frequency noise factor by detuning (db)	Change in capacitance (pF)	Source resistance (ohms)
CV139	0·4	2·3	1,000
6AK5	≯0·2	0·5	3,000
6J4	0·2	1·1	1,300
CV408	≯0·2	1·0	2,500

back on a CV139 the asymmetry is small. It is coincidental that the inductance required to give an input conductance of 1,250 micro-mhos gives about −2-pF capacitance due to feedback.

4.4. *Abnormal Valve Operation*

Apart from input-circuit detuning, it has been found possible to obtain improvement in noise factor by operating the valve very close to grid current. With a valve such as the 6AK5 with very consistent characteristics, it is possible to take advantage of this and use a cathode-bias resistor as low as 56 ohms with 100 V anode supply (particularly if there is also a voltage-dropping resistor in the anode supply). With other valves there may be the danger that a high proportion of the samples will run into grid current and give a poor noise factor when operating with high source resistance. The improved performance is due to the reduction in shot noise and increase in mutual conductance (hence reduction in second stage noise). The optimum source resistance is lowered.

The effect is most marked in a wideband amplifier in which reduction in shot noise is very beneficial, and with a CV139 in a wideband amplifier it is even advantageous to bias so that grid current is taken, but with reduced anode voltage so as not to exceed the anode dissipation rating. With a source resistance at the grid of 370 ohms and overall bandwidth of 12 Mc/s at −3 db, the optimum grid current was 25 μA for a wide range of anode voltage and current—see Fig. 14. The best performance, obtained by operating at 2·5 W anode dissipation in a common-cathode circuit, gave 0·7 db improvement in full-band noise factor on that obtained with normal operation (250 V, 10 mA). The result is particularly significant in view of the additional noise due to grid current which would account for a term of 0·2 (0·4 db) in noise factor. No information on valve life under these conditions has been obtained.

4.5. *Summarized Results*

In Table 5, the total input circuit coil loss for the CV139 and 6J4 triodes was equivalent to a grid–cathode conductance of about 60 micro-mhos and about 35 micro-mhos with the low capacitance 6AK5 and CV408. A tapped auto-transformer was used and this gives less input circuit attenuation than a double-wound coil.

The results obtained with a CV138 pentode using a special input circuit are included to show the benefit obtained with such a circuit, although it is not recommended, and in a wideband amplifier it is preferred to operate the valve in grid-current with a simple input circuit.

The results quoted in the table for the neutralized triode were obtained with the inductance-neutralized circuit [Fig. 8(b)]. Measurements on a wideband amplifier using the capacitance-neutralized circuit feeding a CV138 pentode gave an improvement of 0·1 db over the inductance-neutralized circuit and it is estimated that a further 0·1 to 0·2 db would be obtained with a triode second stage.

In comparing these results with those obtained by Wallmann, Macnee and Gadsden[5] at 30 Mc/s, the following points should be borne in mind:—

They used an unequal-Q input circuit and, moreover, probably tuned for minimum noise factor.

The 6AK5 was operated by them at optimum conditions found experimentally to be 105 V and 70 ohms bias resistance.

The operating conditions for the triode results given in Table 5 were:—

CV139	250 V	10 mA
6J4	110 V	10 mA
6AK5 (CV580)	110 V	8 mA
CV408	150 V	10 mA

For all results, no capacitance was added in the grid circuit which would have simulated conditions with a crystal mixer. This is the only fair comparison of valve performance, but as improvements in valve design lead to lower capacitances, it will become more essential to minimize the capacitance of the mixer or other input circuit feed.

With the capacitance-neutralized circuit, it has been found that the best detuned noise factor is obtained if the neutralizing capacitance is detuned to compensate for the input circuit detuning. The practical procedure is then:—

(a) Neutralize with zero anode current using c.w. signal.

(b) With anode current flowing, peak the input circuit at mid-band, then detune capacitively so that signal is reduced by an amount dependent on the grid impedance. This has to be established by experiment, but with optimum source impedance is in the region of 15% (in volts) for most valves.

(c) Detune the neutralizing capacitance (increase) to give maximum signal with anode current flowing. The input circuit will now be peaked at the same setting as left in (b).

With most valves, the anode–grid capacitance is sufficiently consistent for the neutralizing capacitor to be fixed, and the only tuning required with change of valve is to peak up the input circuit.

Numerous experiments have been made with the 6AK5 (CV850) to establish the best compromise for operating conditions, and conflicting results have been obtained. In some tests, operation with 100 V and 56 ohms cathode-bias resistor gave excellent performance, but later work has indicated that there are probably ageing effects which result in valves running into grid current under these conditions. It is safer to operate with 120-ohms bias resistor, and at about 140 V 15 mA most valves give slightly better performance than at 100 V 10 mA. The source impedance giving minimum mid-band noise factor is lowered with operation at high current and low bias. For an overall bandwidth of 2 Mc/s at —3 db, most samples of this valve type give a noise factor of 2·0 db or better, using an input coil having a resonant resistance of 35,000 ohms and anode circuit load resistance of 1,000 ohms.

With the CV408 (Osram A1714), the optimum source resistance is about 2,500 ohms. The spread in performance is rather greater than with the 6AK5, although the average performance is certainly better. It is not possible to quote results for a truly average valve, and operation at high anode current gives a degradation with the majority of samples, although a few do give excellent performance with low bias and high anode current. A fixed capacitance-neutralized circuit is practical on the basis of tests with about 100 samples, and a cathode bias resistance of 180 ohms seems the best compromise. Some circuit difficulties have been encountered with both inductance and capacitance neutralization, and it has been found necessary to take particular care to avoid common earth paths and to return the 'getter' terminal to the output earth and the spare pin to the input earth. Unless such precautions are taken, some form of parasitic oscillation may occur, or there may be reduction of the bandwidth of the first stage.

TABLE 5. Results for Different Valves and Circuits

Valve and circuit	Equivalent shot noise resistance (ohms)	Minimum mid-band noise factor first stage (only)		Overall receiver bandwidth 4 Mc/s		Overall receiver bandwidth 12 Mc/s		
		db	Source resistance at grid (ohms)	First-stage noise factor (db)	Overall noise factor (db)	First-stage noise factor (db)	Overall noise factor (db)	Source resistance at grid (ohms)
Common-cathode								
CV139	370	2·8	1,000	3·0	3·1	3·6	4·0	400
6J4	Not measured	2·2	1,300	2·4	2·5		3·9	400
6AK5 triode connected ..	580	1·9	2,500	2·1	2·3		4·0	700
CV138 triode connected ..	Not measured	2·9	1,000	—	—		4·0	400
CV408	450	1·7	2,500	1·9	2·1		3·2	650
Common-grid CV139 with unequal-Q to CV138 pentode		2·8	1,000	3·0	3·5	4·0	7·5	Not measured
2 stages of common-grid CV139 with unequal-Q coupling to CV138 pentode		2·8	1,000	3·0	3·3			
2 stages of common-grid CV139 with single-circuit coupling to negative feed-back pentode pair. 1,200-ohm damping on circuit feeding third stage ..		2·8	1,000				6·0	350
CV138 pentode single-circuit input	550 (shot) 430 (partition sic.)	4·3	2,000					
CV138 pentode unequal-Q input						4·8	5·0	Not measured
6AK5 pentode single-circuit input	770 (shot) 850 (partition)	3·8	2,500					
EF50 pentode single-circuit input		5·0	1,200	5·2	5·3		About 9	

Fig. 15 shows the full-band noise factor for the CV139*†, 6AK5 and 6J4†. The simple type of input circuit included no additional capacitance and the source conductance for wider band circuits was adjusted to the condition mentioned in 3.3. Valve operating conditions are the same as listed above for the results in Table 5. The 6J6 gives almost as good a performance as the 6J4, but few valves have been tested. The sub-miniature CV465 is worse than the 6AK5 but the design is not yet sufficiently established for average results to be quoted.

5. Conclusion

The best noise factor which can be obtained with a given valve can only be found by experiment, but the general principles of circuit design are well established.

The available power concept is extremely useful in obtaining an understanding of circuit behaviour, but there is still something lacking in our detailed knowledge of valve noise.

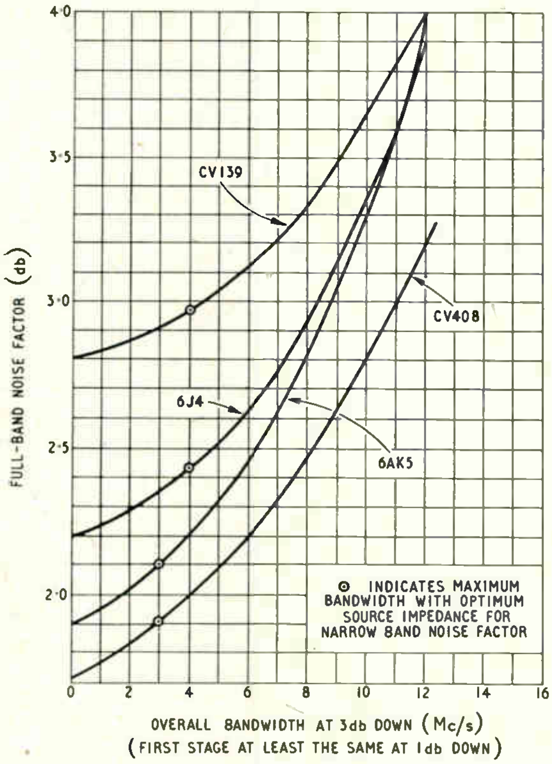

Fig. 15. *Noise factor for different bandwidths with common-cathode triode circuits (frequency 45 Mc/s).*

* The CV139 is now superseded by CV417, because of effects due to secondary emission from the glass bulb caused by electrons which escape from the electrode structure.

† The CV139 and 6J4 were designed for common-grid operation and therefore have excessive input capacitances in common-cathode circuits. The same structures without the grid shields, etc., would give considerable improvement in wideband amplifiers.

Circuit development (and possibly further valve development) is required to establish tolerances which will enable neutralized circuits to be used without readjustment on valve replacement, and to give minimum noise factor with symmetrical frequency response.

In practical systems, the performance obtained will be disappointing as compared with the published results for the valve, unless particular care is taken to keep the capacitances small in the input circuit, to use a high Q coil in the input, and to ensure that the noise of the first stage is really swamping that from the rest of the amplifier. With the best valves, the input circuit loss is extremely significant in narrow-band amplifiers.

There is still scope for further improvement in the noise performance of valves, and it is believed that the economic limit of this has not yet been reached, for such improvement, however small, is more easily made available to a large number of equipments than is the equivalent improvement in transmitters.

Acknowledgments

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The author wishes to thank those of his colleagues, past and present, who have contributed to the experimental work, and particularly to thank Mr. C. P. Fogg for encouragement.

APPENDIX 1

The earliest definition of noise factor in 1941 was in terms of available powers (E. G. James and J. B. Clegg*). Since, in practice, the receiver detector is a voltage-operated device, the available power definition is not in strict agreement with actual measurement and an R.E.M.C. Panel has defined† noise factor in terms of mean-square noise output voltage relative to that part of the output noise voltage which is derived from the source. For a multistage receiver this leads to cumbersome expressions in terms of voltage amplification ratio and input and output impedances, etc.

At the detector, the impedance $(R + jX)$ includes the output impedance of the last stage and the input impedance of the detector. If the noise in series with this, due to receiver noise, is $V_r^{2\Sigma}$ per c/s, and due to amplified noise from the source is $V_a^{2\Sigma}$ per c/s, we get

For available powers $$N' = \frac{\int \frac{V_r^{2\Sigma} df}{4R} + \int \frac{V_a^{2\Sigma} df}{4R}}{\int \frac{V_a^{2\Sigma} df}{4R}}$$

If R is constant over the frequency response, or if $V_r^{2\Sigma}$ and $V_a^{2\Sigma}$ are similar functions of frequency, then

* For a war-time committee.

† The definitions of receiver characteristics prepared by this Inter-Service Committee form the basis of new British Standards to be issued soon.

$$N' = \frac{\int V_r^2\Sigma df + \int V_a^2\Sigma df}{\int V_a^2\Sigma df} = N$$

Hence, if the output circuit is of much greater or much smaller bandwidth than overall, the two definitions are identical. In normal practice the difference is small and does not justify the use of correct formulae in place of equations (1) to (5).

TABLE 6

Circuit	Power gain excluding output loss	Output impedance excluding output circuit loss	Power gain including output loss	Noise factor including output loss
(*a*) Common-cathode	$\frac{\mu^2 R_A}{r_a}$	r_a	$\frac{\mu^2 R_A R_L}{r_a(r_a + R_L)}$	$N + \frac{r_a^2}{\mu^2 R_L R_A}$
(*b*) Common-grid	$\frac{(\mu + 1)R_A}{r_a + (\mu + 1)R_A}$	$r_a + (\mu + 1)R_A$	$\frac{(\mu + 1)^2 R_A R_L}{\{r_a + (\mu + 1)R_A\} \times \{R_L + r_a + (\mu + 1)R_A\}}$	$N + \frac{\{r_a + (\mu + 1)R_A\}^2}{(\mu + 1)^2 R_A R_L}$
(*c*) Common-anode	$\frac{\mu^2 R_A}{(\mu + 1)r_a}$	$\frac{r_a}{\mu + 1}$	$\frac{\mu^2 R_A R_L}{r_a\{r_a + (\mu + 1)R_L\}}$	$N + \frac{r_a^2}{\mu^2 R_L R_A}$

APPENDIX 2

The effect of variation over the band was illustrated for a receiver with constant gain over the response, for which

$$N = \frac{\int umdf}{mB} = \frac{\int udf}{B}$$

Normally m is not constant over the pass-band and cannot be eliminated from the expression.

Then if:—

(a) either or both m and u are symmetrical about mid-band, and

(b) u is monotonic and increasing on both sides of mid-band frequency,

it can be shown that:*

$$\int_{-\infty}^{+\infty} umdf > M \int_{-B/2}^{+B/2} udf$$

In other words, the full-band noise factor is greater than it would be if the response were constant over a frequency range equal to the noise bandwidth.

APPENDIX 3

The effect of the increase in single-frequency noise factor at the edges of the band may modify the limiting sensitivity of the receiver. That is to say, with the same values of full-band noise factor for two receivers which are identical in all respects except that u is constant over the pass-band in one and not in the other, the masking effect of noise after demodulation may be different. It is unlikely that the effect will be at all significant, however, since the noise spectrum after rectification (without signal) will be of similar type for both receivers (i.e., with increase at low frequencies). Since the detection of weak signals on a cathode-ray tube is probably dependent on the change in the noise spectrum caused by the signal, it is difficult to forecast the sense of any difference in sensitivity.

* P. M. Woodward, R.R.E.

APPENDIX 4

The more detailed expressions for the three triode circuits, with resistive input and output circuits and no reactive feedback due to inter-electrode capacitances or lead inductances, are shown in Table 6.

Other expressions for the triode circuits are:

Input resistance of common-grid

$$= \frac{R_L + r_a}{\mu + 1}$$

(The effect of C_{ak} may be obtained by replacing r_a and μ by the appropriate complex expressions.)

Input resistance of common-cathode with resistance R_N between grid and anode

$$= \frac{R_N R_L + R_N r_a + R_L r_a}{r_a + (\mu + 1) R_L}$$

Output resistance of common-cathode

$$= \frac{r_a (R_N + R_A)}{R_N + r_a + (\mu + 1) R_A}$$

Power gain of common-cathode

$$= \frac{(\mu R_N - r_a)^2 R_A}{\{r_a + R_N + (\mu + 1) R_A\} (R_N + R_A) r_a}$$

APPENDIX 5

The effect of input circuit damping is to increase the noise factor $(1 + G_D/G_A)$ times its value with a source of $(G_D + G_A)$, as shown very simply in 2.4.

This agrees with (9a), for we have

$$n = \left(1 + \frac{G_D}{G_A}\right)\left\{1 + \frac{tG_T}{G_D + G_A} + \frac{R_{equ}(G^2 + S^2)}{G_A + G_D}\right\}$$
$$= 1 + \frac{G_D}{G_A} + \frac{tG_T}{G_A} + \frac{R_{equ}(G^2 + S^2)}{G_A}$$

If G_A gives optimum noise factor, then to a close approximation the total change is in the ratio $(1 + G_D/G_A)$.

For anode grid resistance with a common-cathode triode, the equivalent circuit is shown in Fig. 16(a) (for $R_N > r_a$).

It is easier to calculate the effect of R_N on noise factor by using the fact that it is approximately the same as with the same value of anode-grid resistance for a common-grid circuit. The increase in noise factor for the common-cathode triode can therefore very easily be calculated using the expression in the table in Appendix 4 for the common-grid triode.

This gives
$$\Delta n = \frac{\{r_a + (\mu + 1)R_A\}^2}{(\mu + 1)^2 R_N R_A}$$
$$\approx \frac{G_N}{G_A} + \frac{2G_N}{g_m} + \frac{G_N G_A}{g_m^2}$$
Therefore
$$\Delta n = \frac{G_N}{G_A} + \frac{2G_N}{g_m}$$

This is not so great an increase as for the same damping between grid and cathode, but as an approximate rule it can be considered so with practical values of G_N and G_A.

$\therefore$ Effective shot noise resistance for the common-grid is given by
$$R'_{equ} = R_{equ}\frac{\mu^2}{(\mu + 1)^2}$$

For induced grid noise, it is obvious that the contribution due to this is unchanged relative to noise from the source impedance with the common-grid connection.

For the common-anode connection, the modification to the induced grid noise contribution can be calculated in the same way as for shot noise, that is by comparison with the ratio of available power gains of the common-anode and common-cathode circuits.

Assuming neutralization of the grid-cathode capacitance (as would be essential in practice), calculation of the available power output due to induced grid noise gives:—

Common-anode induced grid noise power
$$\approx \frac{(g_m R_A - 1)^2 i^{2\Sigma} r_a}{4(\mu + 1)^2}$$
Common-cathode induced grid noise power
$$\approx \frac{i^{2\Sigma} R_A^2 \mu^2}{4r_a(\mu + 1)}$$
where $i^{2\Sigma}$ = induced grid noise current, grid to cathode.

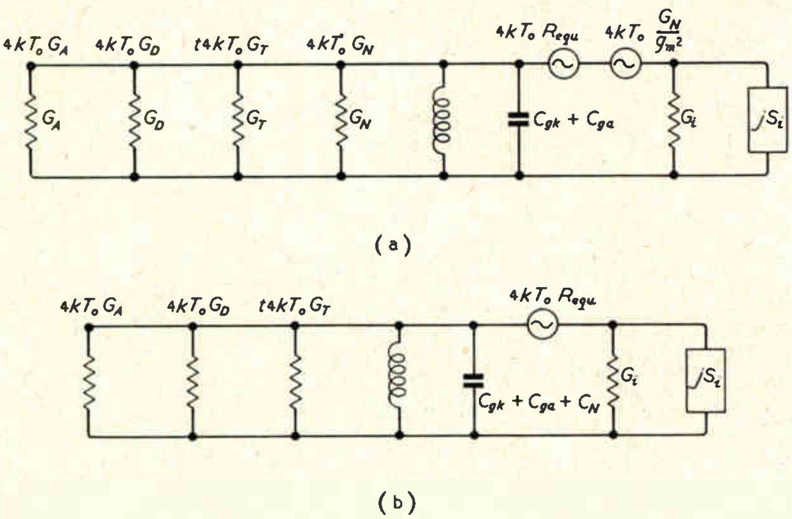

Fig. 16. *Equivalent noise circuit of neutralized triode.* (*a*) *Equivalent of Fig.* 8 (*b*) (G_N = *Shunt conductance of neutralizing coil;* $G_i + jS_i$ = *input admittance due to feedback.* $4k\,T_o\frac{G_N}{g_m^2}$ *is phase-related to current generator* $4kT_oG_N$ *so that addition is in anti-phase when input circuit is resistive).* (*b*) *Equivalent of Fig.* 8 (*d*).

APPENDIX 6

The true equivalent shot noise resistance for the three triode circuits may be obtained from the expressions for output impedance, gain, etc.

For common-cathode:
$$\frac{\text{Available Shot Noise Power}}{\text{Available Power Gain}} = \frac{i_n^{2\Sigma} r_a^2}{4\mu^2 R_A}$$

The same expression is obtained for the common-anode.

For the common-grid, the shot noise generator $i_n^{2\Sigma}$ between anode and cathode develops voltages in the grid-cathode circuit which are amplified in phase opposition to those developed directly in the anode-grid circuit.

Since $(a - b)^2 = a^2 + b^2 - 2(a^2b^2)^{\frac{1}{2}}$, the available noise power at the output is given by:—
$$\frac{i_n^{2\Sigma}}{4}\left[r_a + (\mu + 1)R_A + \frac{(\mu + 1)^2 R_A^2}{r_a + (\mu + 1)R_A} - 2(\mu + 1)R_A\right]$$
$$= \frac{i_n^{2\Sigma}}{4}\left[\frac{r_a^2}{r_a + (\mu + 1)R_A}\right]$$
Hence
$$\frac{\text{Available Noise Power}}{\text{Available Power Gain}} = \frac{i_n^{2\Sigma} r_a^2}{4(\mu + 1)^2 R_A}$$

Hence Ratio
$$\frac{\text{Common Anode}}{\text{Common Cathode}} = \frac{(g_m R_A - 1)^2}{g_m^2 R_A^2(\mu + 1)}$$

But ratio of power gains is $\frac{1}{\mu + 1}$,

$$\therefore \frac{\text{Induced Grid Noise Contribution, Common Anode}}{\text{Induced Grid Noise Contribution, Common Cathode}} \approx \left(1 - \frac{1}{g_m R_A}\right)^2$$

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BAND-PASS FILTERS

Synthesis of Constant-Resistance Types for a Prescribed Response

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SUMMARY.—The response of the multistage constant-resistance network is the product of the responses of its individual stages, similar to the multistage amplifier. Thus band-pass filters can be designed to have a maximally-flat or quasi-Tchebyshev response exactly like the corresponding band-pass amplifier. This type is suited for wide band filters as well as for multiplexers.

1. Multistage Filters

THE staggered multistage band-pass amplifier has been calculated to give a maximally-flat response,*[1] and later by Linnebach[2] to give a quasi-Tchebyshev response.† The calculation of the band-pass amplifier is much easier than that of the usual reactive filter because of the multiplicative property of the response due to the isolation of the stages by amplifier valves. In the case of the amplifier each stage represents one factor of the factorized overall response. On the other hand, the reactive filter has to be treated as a whole. The maximally-flat filter has been generally solved.[3] The quasi-Tchebyshev, however, has been only lately generally solved by the author.[4,5] The calculation of the latter type, as well as its adjustment, is not an easy matter in the case of a filter having a large number of circuit-elements.

The constant-resistance network makes possible cascaded stages as in the case of the multistage amplifier. The resulting network, however, suffers from the following disadvantages: (1) Two resonant circuits are needed for each zero or pole of the insertion loss, as compared with one circuit in case of amplifier or reactive filters; (2) There is a constant insertion loss to be added to the selective loss of the corresponding reactive filter. However, in special cases the constant-resistance filter may be advantageously used; e.g., as a wide-band filter with a large number of circuit-elements and quasi-Tchebyshev response. The corresponding reactive filter is in this case more complicated to calculate and to adjust. But the main advantage of the constant-resistance band-pass filter is the simplicity with which several such filters can be connected, on the input side, in series or in parallel (Fig. 1), for the input impedance of each is independent of frequency and equal to R. A multiplexer consists thus simply of several such filters, having different pass-bands, and connected in series or in parallel as shown in Fig. 1.

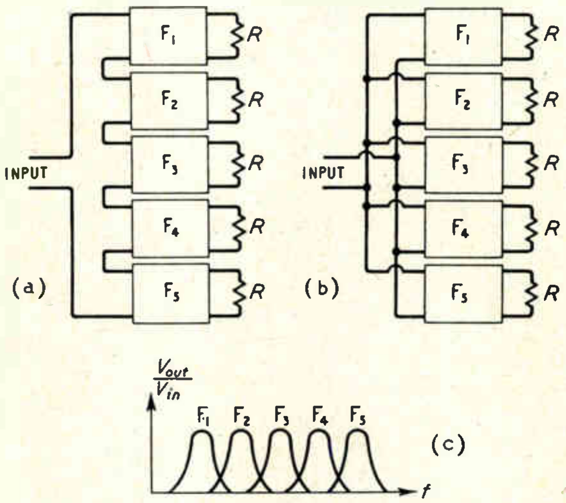

Fig. 1. *Multiplexer composed of five constant-resistance band-pass filters: (a) Series connection, (b) Parallel connection, (c) Response curves of individual filters.*

The two main properties of the constant-resistance network (Fig. 2) are:

(1) Input impedance $= R$.

(2) Voltage ratio $\left(\dfrac{V_{in}}{V_{out}}\right) = 1 + \dfrac{Z}{R}$

$$= 1 + \frac{R'}{Z'} \quad \ldots \quad \ldots \quad (1)$$

Let Z be a series resonant circuit and Z' be a parallel one (Fig. 3).
Then:

$$Z = j\omega L + \frac{1}{j\omega C} = jRQ\left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right)$$

$$\approx j\frac{2RQ}{\omega_0}(\omega - \omega_0) \quad \ldots \quad \ldots \quad (2)$$

* The maximally-flat response is defined as:

$$\left|\frac{V_{in}}{V_{out}}\right|^2 = K^2\left\{1 + \left(k\frac{\omega - \omega_0}{\omega' - \omega_0}\right)^{2n}\right\}.$$

† The quasi-Tchebyshev band-pass response is defined as:

$$\left|\frac{V_{in}}{V_{out}}\right|^2 = K^2\left\{1 + h\,T_n^2\left(\frac{\omega - \omega_0}{\omega' - \omega_0}\right)\right\},$$

where T_n is the Tchebyshev polynomial of the first kind and nth degree. It possesses a Tchebyshev response only in the pass-band.

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$$\left(\frac{V_{in}}{V_{out}}\right) = \left[1 + j\frac{2Q}{\omega_0}(\omega - \omega_0)\right] \quad .. \quad (3)$$

$$\text{Loss } L = 20 \log \left|\frac{V_{in}}{V_{out}}\right|$$

$$= 10 \log \left[1 + \left\{\frac{2Q}{\omega_0}(\omega - \omega_0)\right\}^2\right] \text{ db} \quad .. \quad (4)$$

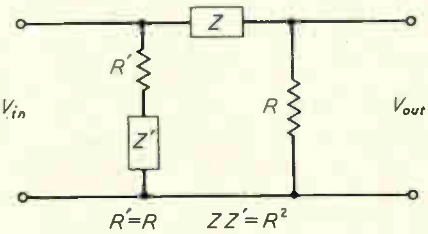
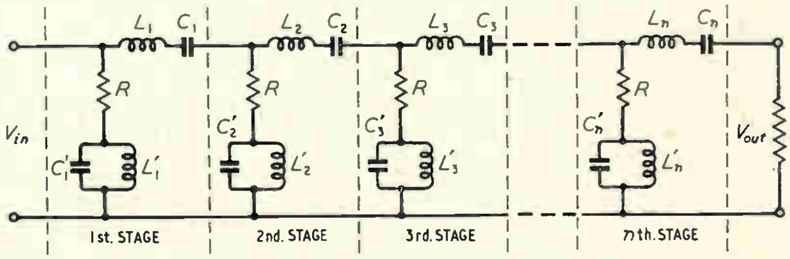
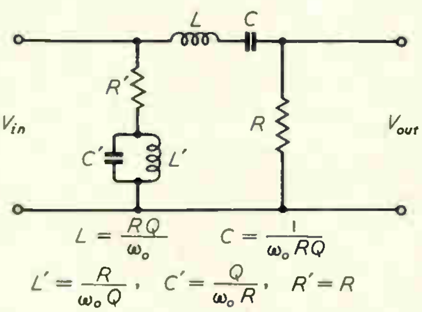


Fig. 2 (above). *Constant-resistance network.*

Fig. 3 (below). *Constant-resistance band-pass filter.*

Fig. 4 (right). *Multistage constant-resistance band-pass filter.*





L has a zero at ω_0. Now a similar network can be substituted for R, and this can be continued for, say, n stages (Fig. 4). The final voltage ratio is the product of the corresponding n factors.

Even poles of the loss L can be realized by making Z a parallel resonant circuit and Z' a series one (Fig. 5). We have then:

$$\left(\frac{V_{in}}{V_{out}}\right) = 1 + \frac{1}{j\frac{2Q}{\omega_0}(\omega - \omega_0)} \quad .. \quad .. \quad (5)$$

In Fig. 3 we note that $L/L' = C'/C = Q^2$. Thus for large values of Q the circuit-elements have either extremely high or extremely low values. Hence this type of network is only suitable for wideband filtering.

We note further that the error in the approximate relations (2) and (3) is small even for values of Q as small as 3. If ω' and ω'' are the half-power points then:

$$\omega' = \omega_0\sqrt{1 + \frac{1}{4Q^2}} + \frac{\omega_0}{2Q},$$

$$\omega'' = \omega_0\sqrt{1 + \frac{1}{4Q^2}} - \frac{\omega_0}{2Q}$$

the approximate values being correspondingly $\omega_0 + \frac{\omega_0}{2Q}$ and $\omega_0 - \frac{\omega_0}{2Q}$. The error is approximately $\frac{\omega_0}{8Q^2}$; its value for $Q = 5$ is $\frac{\omega_0}{200}$ and for $Q = 3$ is $\frac{\omega_0}{72}$.

The voltage-ratio of the network of Fig. 4 is given by:

$$\left[\frac{V_{in}}{V_{out}}\right] = \prod_{v=1}^{n}\left[1 + j\frac{2Q_v}{\omega_{0v}}(\omega - \omega_{0v})\right] =$$

$$\prod_{v=1}^{n}\left[1 + j\frac{2Qv}{\omega_{0v}}(\omega_0 - \omega_{0v}) + j\frac{2Q_v}{\omega_{0v}}(\omega - \omega_0)\right]$$

$$.. \quad .. \quad .. \quad (6)$$

where ω_{0v} and Q_v correspond to the vth stage and ω_0 to the mid-band frequency (Figs. 6 and 7).

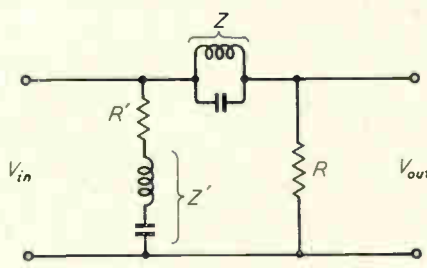


Fig. 5. *Circuit to provide an attenuation pole.*

2. Maximally-Flat Response

$$\text{Let } S = \left|\frac{V_{in}}{V_{out}}\right|^2 = K^2\left\{1 + \left(k\frac{\omega - \omega_0}{\omega' - \omega_0}\right)^{2n}\right\}$$

$$.. \quad .. \quad (7)$$

Comparing with Fig. 6 we have:

$$k = \sqrt[2n]{\left|\frac{V_{out\ max}}{V_{out\ min}}\right|^2 - 1} \quad \ldots \quad \ldots \quad (8)$$

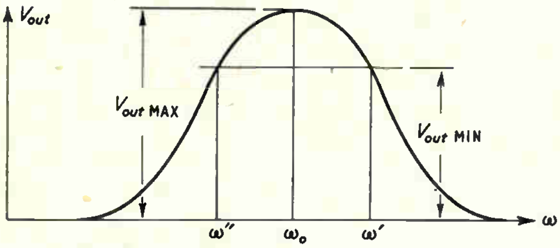


Fig. 6. *Maximally-flat band-pass response.*

We shall now find $R = \left(\dfrac{V_{in}}{V_{out}}\right)$ from $S = \left|\dfrac{V_{in}}{V_{out}}\right|^2$. Let ω_v be a zero of S.

Then: $$k\frac{\omega_v - \omega_0}{\omega' - \omega_0} = \sqrt[2n]{-1} = \sqrt[2n]{e^{j(\pm 2v - 1)\pi}} = e^{j(\pm 2v - 1)\pi/2n} = e^{j(\pm 2v - 1)\psi}$$

where $v = 1, 2, 3, \ldots, n$, and $\psi = \dfrac{\pi}{2n}$. The zeros of $R = \left(\dfrac{V_{in}}{V_{out}}\right)$ may therefore be taken as ω_v,

where $$k\frac{\omega_v - \omega_0}{\omega' - \omega_0} = e^{j(2v-1)\psi} = e^{j\theta_v} \quad \ldots \quad (9)$$

where $$\theta_v = (2v - 1)\psi \quad \ldots \quad \ldots \quad \ldots \quad (10)$$

Therefore $\left(\dfrac{V_{in}}{V_{out}}\right)$ may be expressed as the product:

$$\left(\frac{V_{in}}{V_{out}}\right) = Kj^n \prod_{v=1}^{n}\left[k\frac{\omega - \omega_0}{\omega' - \omega_0} - e^{j\theta_v}\right]$$

$$= \left[K\prod_{v=1}^{n}\sin\theta_v\right]\prod_{v=1}^{n}\left\{1 - j\cot\theta_v + j\frac{k}{\sin\theta_v}\frac{\omega - \omega_0}{\omega' - \omega_0}\right\}$$

$$= \frac{K}{2^{n-1}}\prod_{v=1}^{n}\left\{1 - j\cot\theta_v + j\frac{k}{\sin\theta_v}\frac{\omega - \omega_0}{\omega' - \omega_0}\right\} \quad \ldots \quad \ldots \quad (11)$$

Expressions (6) and (11) are identical when $K = 2^{n-1}$, $\cot\theta_v = -\dfrac{2Q_v}{\omega_{0v}}(\omega_0 - \omega_{0v})$ and $\dfrac{k}{\sin\theta_v}\dfrac{1}{(\omega' - \omega_0)} = \dfrac{2Q_v}{\omega_{0v}}$. We deduce therefore the following:

(1) The mid-band loss $L_0 = 6(n - 1)$ db. (12)

(2) The Q of the vth stage is given by:

$$Q_v \approx \sqrt[2n]{\left|\frac{V_{out\ max}}{V_{out\ min}}\right|^2 - 1} \times \frac{1}{\sin(2v - 1)\dfrac{\pi}{2n}}\frac{\omega_0}{(\omega' - \omega'')} \quad \ldots \quad \ldots \quad (13)$$

(3) The detuning of the vth stage is given by:

$$(\omega_{0v} - \omega_0) = \frac{1}{\sqrt[2n]{\left|\dfrac{V_{out\ max}}{V_{out\ min}}\right|^2 - 1}} \times \frac{(\omega' - \omega'')}{2}\cos(2v - 1)\frac{\pi}{2n} \quad \ldots \quad \ldots \quad (14)$$

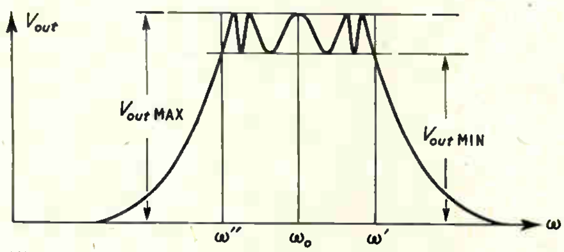


Fig. 7. *Five-stage quasi-Tchebyshev band-pass response.*

3. Quasi-Tchebyshev Response

Let: $$S = \left|\frac{V_{in}}{V_{out}}\right|^2 = K^2[1 + h^2 T_n^2(\Omega)] \quad (15)$$

where $\Omega = \dfrac{\omega - \omega_0}{\omega' - \omega_0}$

and $T_n(\Omega) = \cos n.\cos^{-1}\Omega = \cosh n.\cosh^{-1}\Omega =$ Tchebyshev polynomial of the first kind and nth degree.

From Fig. 7 we have:

$$h = \sqrt{\left|\frac{V_{out\ max}}{V_{out\ min}}\right|^2 - 1} \quad \ldots \quad \ldots \quad (16)$$

We modify expression (15) into

$$S = \left|\frac{V_{in}}{V_{out}}\right|^2 = K^2K'^2[(1 - d^{2n})^2 + 4d^{2n}T_n^2(\Omega)]$$
$$= K^2K'^2[(1 + d^{4n}) + 2d^{2n}T_{2n}(\Omega)] \quad \ldots \quad \ldots \quad (17)$$

Therefore $K'(1 - d^{2n}) = 1$ and $2K'd^n = h$.

Hence $K' = \frac{1}{2}(\sqrt{1 + h^2} + 1) = \dfrac{1}{(1 - d^{2n})}$ and

$$d^{2n} = \frac{\sqrt{1 + h^2} - 1}{\sqrt{1 + h^2} + 1} \quad \ldots \quad \ldots \quad (18)$$

We shall now determine the zeros of S. Let $\Omega_v = \cos\phi_v$ be a zero. Hence

$$(1 + d^{4n}) + 2d^{2n}\cos 2n\phi_v = 0$$

Therefore $e^{j4n\phi_v} + 2De^{j2n\phi v} + 1 = 0$

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

$$j2n\phi_v = \log_e(-D \pm \sqrt{D^2 - 1})$$
$$= \log_e|-d^{\pm 2n}| + j\pi(2v - 1)$$
$$\therefore \phi_v = (2v - 1)\psi \pm j \log_e d;$$

where $v = 1, 2, 3, \ldots n;\ \psi = \frac{\pi}{2n}$.

The zeros of $S = \left|\frac{V_{in}}{V_{out}}\right|^2$ are therefore:

$$\Omega_v = \cos \phi_v = \cos\left[(2v - 1)\psi \pm j \log_e d\right]$$
$$= \frac{1}{2d}\left[(1 + d^2)\cos\theta_v \pm j(1 - d^2)\sin\theta_v\right]$$

where $\theta_v = (2v - 1)\frac{\pi}{2n}$; $v = 1, 2, 3, \ldots n$.

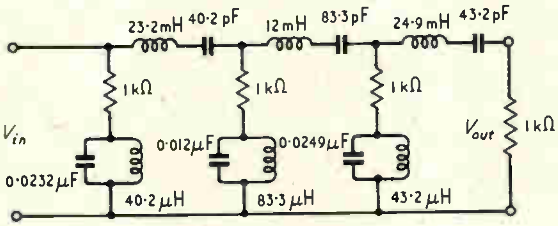


Fig. 8. *Maximally-flat constant-resistance band-pass filter of Example 1.*

The zeros of $R = \left(\frac{V_{in}}{V_{out}}\right)$ are therefore:

$$\Omega_v = \frac{1}{2d}\left[(1 + d^2)\cos\theta_v + j(1 - d^2)\sin\theta_v\right] \quad (19)$$

R may therefore be expressed as the following product:

$$R = \left(\frac{V_{in}}{V_{out}}\right) = KK'(2d)^n j^n \times$$
$$\prod_{v=1}^{n}\left\{\Omega - \frac{1 + d^2}{2d}\cos\theta_v - j\frac{1 - d^2}{2d}\sin\theta_v\right\} \quad (20)$$

The factor $(2d)^n$ is due to the equality of the coefficients of Ω^{2n} in the equation $S = |R|^2$.

$$\left(\frac{V_{in}}{V_{out}}\right) = \left\{KK'(1 - d^2)^n \prod_{v=1}^{n}\sin\theta_v\right\} \times$$
$$\prod_{v=1}^{n}\left\{1 - j\left(\frac{1 + d^2}{1 - d^2}\right)\cot\theta_v + j\left(\frac{2d}{1 - d^2}\right)\frac{1}{\sin\theta_v}\left(\frac{\omega - \omega_0}{\omega' - \omega_0}\right)\right\}$$
$$= \left\{\frac{KK'(1 - d^2)^n}{2^{n-1}}\right\}\prod_{v=1}^{n}\left\{1 - j\left(\frac{1 + d^2}{1 - d^2}\right)\cot\theta_v + j\left(\frac{2d}{1 - d^2}\right)\frac{1}{\sin\theta_v}\left(\frac{\omega - \omega_0}{\omega' - \omega_0}\right)\right\} \quad \ldots \quad (21)$$

By comparing (6) and (21) we deduce:

(1) Mid-band loss $= 20 \log K$

$$= 20 \log \frac{2^{n-1}(1 - d^{2n})}{(1 - d^2)^n}\ \text{db.} \quad \ldots \quad (22)$$

(2) The Q of the vth stage is given by:

$$Q_v \approx \left(\frac{2d}{1 - d^2}\right)\left(\frac{\omega_0}{\omega' - \omega''}\right)\frac{1}{\sin(2v - 1)\frac{\pi}{2n}}$$

(3) The detuning of the vth stage is given by:

$$(\omega_{0v} - \omega_0) = \left(\frac{1 + d^2}{4d}\right)(\omega' - \omega'')\cos(2v - 1)\frac{\pi}{2n} \quad \ldots \quad \ldots \quad (24)$$

4. Typical Examples

Example (1) A maximally-flat constant-resistance filter is to be calculated having:

$n = 3$, $\omega_0 = 10^6$, $(\omega' - \omega'') = 10^5$,

$$\left|\frac{V_{out\ max}}{V_{out\ min}}\right| = 2 \text{ and } R = 1000\ \Omega.$$

We have $k = \sqrt[2n]{\left|\frac{V_{out\ max}}{V_{out\ min}}\right|^2 - 1} \doteq 1{\cdot}2$,

$$\psi = \frac{\pi}{2n} = 30°,\ \theta_1 = 30°,\ \theta_2 = 90°,\ \theta_3 = 150°.$$

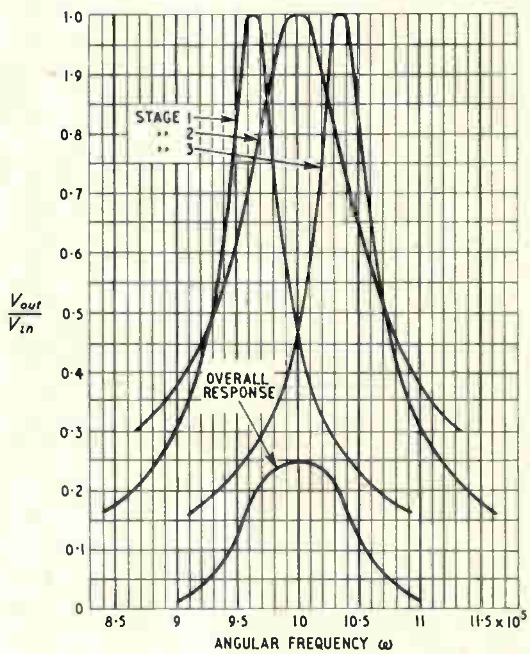


Fig. 9. *Response of maximally-flat band-pass filter of Example 1 and Fig. 7.*

$$Q_v = k \frac{\omega_0}{\omega' + \omega''} \frac{1}{\sin\,\theta_v} = \frac{12}{\sin\,\theta_v};$$

$$Q_1 = 24,\ Q_2 = 12,\ Q_3 = 24.$$

$$(\omega_{0v} - \omega_0) = \frac{\cos\,\theta_v}{2k}(\omega' - \omega'')$$

$$= 4{\cdot}17 \times 10^4 \cos\,\theta_v$$

$$\omega_{01} = 1{\cdot}0361 \times 10^6,\quad \omega_{02} = 10^6,\quad \omega_{03} = 9{\cdot}638 \times 10^5.$$

$$L_v = \frac{RQ_v}{\omega_{0v}},\qquad C_v = \frac{1}{\omega_{0v} R Q_v},$$

$$L'_v = \frac{R}{\omega_{0v} Q_v},\qquad C_v' = \frac{Q_v}{\omega_{0v} R}$$

The values of the circuit-elements are given in Fig. 8. The mid-band loss is $6(n - 1) = 12$ db. (voltage ratio $4:1$). The responses of the individual stages, as well as the overall response (which is equal to their product) are shown in Fig. 9.

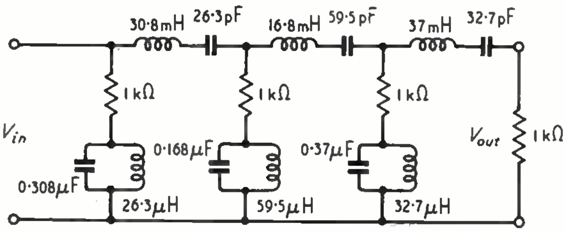


Fig. 10. *Quasi-Tchebyshev constant-resistance band-pass filter of Example 2.*

Example (2) A quasi-Tchebyshev constant-resistance filter is to be calculated with: $n = 3$,

$$\omega_0 = 10^6,\ (\omega' - \omega'') = 2 \times 10^5,\ \left|\frac{V_{out\ max}}{V_{out\ min}}\right| = \sqrt{2},$$

and $R = 1000\,\Omega$.

We have: $h = \sqrt{\left|\frac{V_{out\ max}}{V_{out\ min}}\right|^2 - 1} = 1$

$$d^{2n} = \frac{\sqrt{1 + h^2} - 1}{\sqrt{1 + h^2} + 1},\ d^6 = 0{\cdot}1715,\ d = 0{\cdot}7454.$$

$$\text{Mid-band loss} = 20\ \log \frac{2^{n-1}(1 - d^{2n})}{(1 - d^2)^n}$$

$$= 20\ \log 37{\cdot}7 = 31{\cdot}5 \text{ db.}$$

$$Q_v = \left(\frac{2d}{1 - d^2}\right) \frac{\omega_0}{\omega' - \omega''} \frac{1}{\sin\,\theta_v} = \frac{16{\cdot}8}{\sin\,\theta_v};$$

$$Q_1 = 33.6,\ Q_2 = 16{\cdot}8,\ Q_3 = 33{\cdot}6.$$

$$(\omega_{0v} - \omega_0) = \frac{1 + d^2}{4d}(\omega' - \omega'')\cos\,\theta_v$$

$$= 10{\cdot}4 \times 10^4 \cos\,\theta_v$$

$$\omega_{01} = 1{\cdot}09 \times 10^6,\quad \omega_{02} = 10^6,\quad \omega_{03} = 9{\cdot}1 \times 10^5.$$

The filter is shown in Fig. 10. The responses of the individual stages, as well as the overall response (which is equal to their product), are shown in Fig. 11.

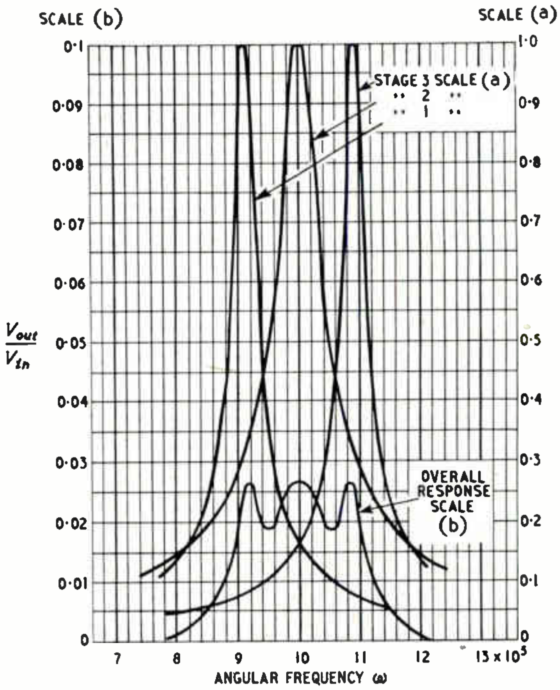


Fig. 11. *Response of quasi-Tchebyshev band-pass filter of Example 2 and Fig. 9.*

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ELECTRON MULTIPLIER VALVE

Comparison with the Conventional R.F. Amplifier

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SUMMARY.—A detailed comparison of the performance of an electron multiplier with that of a conventional valve is given. The conclusion is that whenever bandwidth requirements are high the use of multipliers may be beneficial. Arguments against making the overall multiplication ratio larger than approximately 10 are presented. H.F. limitations due to times of flight and signal-to-noise ratio are discussed.

1. Introduction

It has been known for a long time that the use of secondary-electron multiplication can considerably increase the performance of receiving valves.[1] It will be shown below that the electron multiplier has a field of utility in a considerable range of frequencies and bandwidth just outside those available for an ordinary valve. It is proposed to estimate the extent of this useful range by considering the limitations imposed by irreducible capacitances, electron time of flight and signal-to-noise ratio.

The multiplier which we shall compare with a valve is schematically represented in Fig. 1. The input structure (i.e., cathode, control grid, and screen grid) is essentially identical with the corresponding part of an r.f. pentode. The primary electron beam is directed on to the secondary emitting surface and the secondaries are collected by a grid which plays the part of an anode. It should be pointed out that electrode capacitances in a valve and a multiplier of this type can be made very nearly equal, so that the increase in mutual conductance due to multiplication need not be offset by any increase in capacitance.

2. Transit-Time Limitations

The quantity which is of importance for the design of wideband amplifiers is the product of the gain, G, and bandwidth, B. This product was shown by Hansen[2] to be:*—

$$G.B = \frac{g}{2\pi C} \qquad \ldots \quad \ldots \quad \ldots \quad (1)$$

where C is the total capacitance determining the frequency of the tuned anode circuit and g the mutual conductance. If g can be increased by secondary-emission multiplication the value of the product $G.B$ can be increased in the same proportion. This would seem to open nearly unlimited possibilities since, with a multi-stage multiplier, multiplication ratios of the order of 10^3 or more could be contemplated. This, however, is not the case. We shall show that an overall multiplication ratio greater than about 10 is not practical. The limitation arises because the value $G.B$ for the multiplier, although proportional to the secondary multiplication ratio, δ, cannot be increased beyond a limit imposed by the electron time of flight τ between cathode and control grid. On account of this finite time of flight the valve cannot be operated at frequencies larger than $1/\tau$.

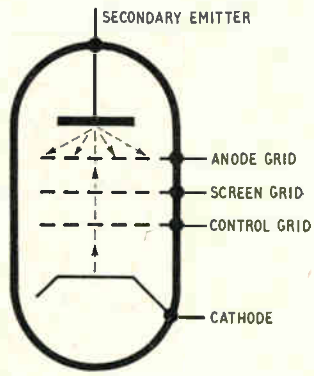

Fig. 1. *Schematic representation of electron multiplier.*

Replacing the valve by its equivalent diode, we can express the value of $G.B$ in terms of the equivalent anode voltage V and distance d. The maximum value of $G.B$ will be obtained if we use the minimum irreducible capacitance C_m of the diode considered as a plane parallel capacitor. By using Child's Law[3] we obtain

$$(G.B)_{max} = \frac{g}{2\pi C_m} = 0{\cdot}7 \times 10^7 \frac{V^{\frac{1}{2}}}{d} \quad \ldots \quad (2)$$

We now compare this result with an expression[4] for the time of flight τ:—

$$\tau = \int_0^d \frac{dx}{v(x)} \qquad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (3)$$

which again by Child's Law may be shown to be:—

$$\tau = 0{\cdot}5 \times 10^{-7} \frac{d}{V^{\frac{1}{2}}} \qquad \ldots \quad \ldots \quad \ldots \quad (4)$$

* The units employed in this paper are:— A, V, F, Ω, sec, c/s, cm.

MS accepted by the Editor, March 1953

From (2) and (4) we see that

$$(G.B)_{max} \approx \frac{1}{\tau} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (5)$$

Thus the maximum value of $(G.B)$ is of the order of the reciprocal of the electron time of flight between cathode and grid.

In general we shall have an expression:—

$$G = m \frac{1}{B} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (6)$$

where

$$0 < m < m_0 = \frac{g}{2\pi C_m} \quad \ldots \quad \ldots \quad (7)$$

Equation (6) corresponds to a family of hyperbolae of which two are shown on Fig. 2. The curve passing through B and C corresponds to the limiting case (5) and the valve cannot operate at the gain and bandwidth corresponding to any point lying above this curve. Suppose, however, that the actual capacitance C which we should have taken into account is $\alpha > 1$ times larger than the limiting capacitance C_m. The hyperbola passing through A and D corresponds to this case. If we choose now any point between this curve and the co-ordinate axes, the valve should be able to supply the corresponding bandwidth and gain. For practical reasons it is not worth while, and is even disadvantageous under certain conditions, to use smaller gains than 2 to 3. This lower limit corresponds to G_0 in Fig. 2. Let us also introduce a certain maximum gain G_m and thus define the 'working area' of the valve as given by G_0 AD G_m. The maximum bandwidth available is that given by the point A:—

$$B_A = \frac{m_0}{\alpha} \frac{1}{G_0} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (8)$$

Since we cannot increase the bandwidth any more by reducing the capacitance, the only other alternative is to increase m_0 by increasing the slope of the valve by electron multiplication. Allowing the current to increase δ times we can increase the slope in the same ratio.[1] Hence the maximum bandwidth attainable will be:—

$$B_{max} = \frac{m_0 \delta}{\alpha} \frac{1}{G_0} \quad \ldots \quad \ldots \quad \ldots \quad (9)$$

There is no point, however, in making δ larger than α since the maximum value of m cannot exceed m_0 on account of the time of flight of the electron in the cathode to grid space.

We arrive thus at the following conclusion: if we want to increase the performance of a valve we should include between its control grid and anode an electron multiplier stage with a multiplication ratio

$$\delta_{useful} = \alpha = \frac{C}{C_m} \quad \ldots \quad \ldots \quad \ldots \quad (10)$$

This will increase the gain obtainable for a particular bandwidth to a maximum value which is determined by the electronic time of flight and expressed by (5).

We want now to obtain an estimate of the magnitude of δ_{useful} for a practical case. Suppose we add a multiplier stage to the r.f. pentode 6AM6. C_m is a quantity which is difficult either to measure or to calculate since it comprises only the capacitance of the cathode to control grid under the assumption that the equivalent anode is in the plane of the grid. We can, however, use the limiting condition

$$\frac{g}{2\pi C_m} \approx \frac{1}{\tau} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (11)$$

provided we can calculate τ. If we use formula (4) for this purpose, uncertainty arises as to the value of the equivalent anode voltage since the valve is normally run with the grid negative. The capacitance C is equal to the capacitance of the grid to all electrodes of the valve (input capacitance) plus the stray capacitances of the circuit and the capacitance of the preceding valve.

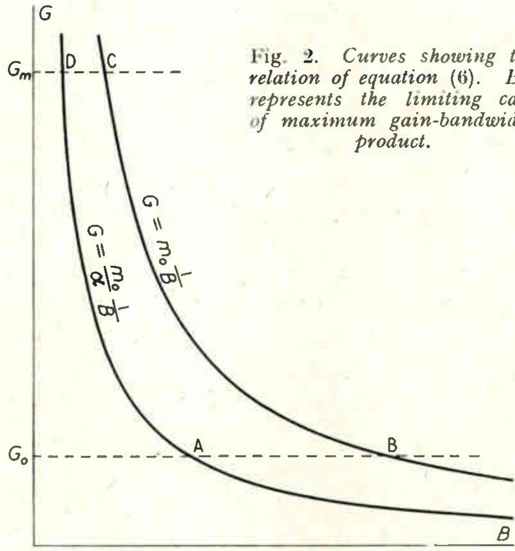

Fig. 2. *Curves showing the relation of equation (6). BC represents the limiting case of maximum gain-bandwidth product.*

From (4), (10) and (11) we obtain:—

$$\delta_{useful} = \frac{4\pi C V}{gd} \times 10^7 \quad \ldots \quad \ldots \quad (12)$$

Using the following values for the 6AM6:—

$C = 2\,C_{input} = 14 \times 10^{-12}$ F

$V = 0{\cdot}2$ V

$g = 6 \times 10^{-3}$ A/V

$d = 1{\cdot}5 \times 10^{-2}$ cm

we obtain:— $\delta_{useful} \approx 9$

Hence, by the use of electron multiplication, we can increase the value of $G.B$ by a factor of the order of 10. This proves the assertion made above that a multistage multiplier with a very large overall multiplication ratio would, from the point of view of the $G.B$ product, be no better than a single-stage valve with $\delta = 10$.

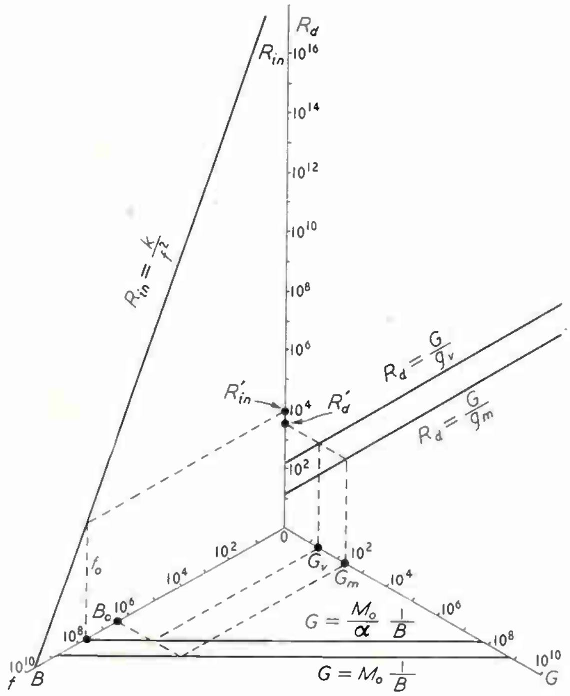

Fig. 3. *Diagram for computing the relative performance of an ordinary valve and a multiplier type.*

3. Performance at High Frequencies

(a) *Limitations due to Input Impedance.*

It is well known that as the frequency increases the gain obtainable from a valve gradually decreases. We have seen in Section 2 that there is a maximum gain for any particular value of bandwidth. We must remember, however, that for a given bandwidth we can have different values of the carrier frequency which can also affect the performance of the valve. In other words a valve may lose performance even if it is operated well within the limits of the $G.B$ product if the frequency is too high. The quantity m, sometimes called the figure of merit, is therefore not a unique basis for classification of valves.

There are thus two questions to be answered: (a) given a certain bandwidth B and frequency f—can a particular valve provide any useful gain under these conditions, and (b) in what respect is a multiplier better than a conventional valve?

The maximum gain for a given bandwidth is given by (6). To obtain this gain we have to use a damping resistance given by:—

$$R_d = \frac{G}{g} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (13)$$

The damping resistance R_d is composed of the anode load R_L and the input impedance of the next valve R_{in} taken in parallel:—

$$R_d = \frac{R_L R_{in}}{R_L + R_{in}}$$

Hence:—

$$R_L = \frac{R_d R_{in}}{R_L - R_{in}}$$

Since R_L must be positive, the condition necessary to obtain the gain G is

$$R_d < R_{in} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (14)$$

The input impedance is inversely proportional to the square of the frequency:—

$$R_{in} = \frac{k}{f^2} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (15)$$

where k is a constant determined by the geometry of the valve. Using (6), (13), (14) and (15) the condition for operation is

$$\frac{k}{m} g > \frac{f^2}{B} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (16)$$

To compare the performance of a valve with that of a multiplier made by addition of a single multiplication stage ($\delta = 10$ say) we shall consider the case of the pentode 6AM6 and the Birmingham television transmitter ($f_0 = 70$ Mc/s; $B_0 = 3$ Mc/s).

In Fig. 3 the three co-ordinate axes have their origin at 0. We can look upon the system of co-ordinates as rectangular and viewed in perspective or as a plane system of oblique axes. Consider first the axes G and B. The two straight lines correspond to the two hyperbolae of Fig. 2 plotted on a logarithmic scale and determine the maximum gains G_v and G_m that could be obtained by using respectively a valve and a multiplier at a bandwidth B_0. These gains can be obtained by using a suitable damping resistance R_d defined by equation (13). Denoting by g_v and g_m the slopes of the valve and multiplier respectively and plotting (13) on a logarithmic scale, we obtain the two straight lines in the system of axes $R_d - G$. It is obvious that the damping resistance in both cases must be the same (although the gain of the multiplier will be 10 times larger, its slope is also increased in the same ratio). We thus find a resistance R_d' which must be used in order to obtain maximum gain both for the valve and the multiplier, compatible with the $G.B$ limitation.

Consider now the axis $R_{input} - f$. The line plotted in this system corresponds to (15). The constant k was determined from a single measurement of input impedance ($R_{input} = 10^4\,\Omega$ at $f = 45$ Mc/s). To a frequency f_0 corresponds an input impedance R_m'. From condition (14) it follows that unless the point R_d' falls below R'_{in} on the $R_{input}\, R_d$ axis we cannot obtain the prescribed gain at the particular frequency and bandwidth.

It follows from these considerations that if a multiplier has the same value of the constant k as a valve (same input impedance–frequency characteristic) there exists for both the same limiting frequency (for a fixed bandwidth) at which both will cease operation. It should be stressed, however, that under possible operating conditions the multiplier will always have a larger gain by a factor δ. This is a very important advantage, especially when the bandwidth required is high.

We have assumed above that the input impedance of a multiplier is the same as that of a valve possessing the same input structure. This is a reasonable assumption if we look upon input impedance as being due to the finite time of flight from cathode to grid and to the inductance of cathode leads.

(b) *Frequency Limitations due to the Multiplier Section.*

We have hitherto discussed the limitations imposed upon high-frequency performance due to the input structure and therefore common to the conventional valve and a multiplier. We concluded that the existence of a finite time of flight τ in the cathode to grid space and certain irreducible capacitances of the valve will limit the useful multiplication ratio δ and the maximum frequency at which the tube can operate.

We now want to know whether this frequency limit will be further reduced by the events in the multiplier section. The problem of frequency response is exceedingly complicated and it does not seem rewarding to do more than discuss it in a qualitative way.

Consider an electron starting from the input (say the plane of the screen grid) and flying through the anode to the secondary emitter and causing the emission of δ secondaries. This series of events will produce a pulse of current in the anode circuit, the duration of which is equal to:—

$$\tau_0 = \tau_1 + \tau_2 + \tau_3 + \tau_4$$

where:—

- τ_1 — transit time from screen grid to anode.
- τ_2 — transit time from anode to secondary emitter.
- τ_3 — time delay in secondary emission.
- τ_4 — transit time from secondary emission to anode.

Let us first assume that these transit times are constant and equal for all electrons. If the multiplication ratio δ is also equal for all electrons, the pulses due to them will be identical, differing only in phase. The total anode current may be obtained by summing all these pulses.

Suppose now that the signal frequency is small compared with $1/\tau_0$. The shape of the pulse will not greatly influence the low-frequency components and the gain of the valve will be determined by δ.

In some types of multipliers (like the orbital type described by Bull and Atherton[5]), τ_1 may be very much larger than $\tau_2 + \tau_3 + \tau_4$ on account of large screen grid to anode distance. During τ_1 we shall thus have a slow rise of current in the anode circuit. If the frequency is considerably larger than $1/\tau_1$ this induced current will be very nearly averaged to zero and the effect of τ_1 will be to delay the signal. As an electron passes the anode towards the secondary emitter, it produces a current in the opposite direction to that which was flowing while the electron was approaching the anode. The anode thus appears to lose a charge just before it starts acquiring δ of them in virtue of multiplication. It follows that at certain frequencies, comparable with $1/(\tau_2 + \tau_3 + \tau_4)$, the gain will be determined by $\delta - 1$. For still higher frequencies, when the number of oscillations during the time $\tau_2 + \tau_3 + \tau_4$ becomes large, the gain will fall to zero.

An exact solution of the problem would require a knowledge of the shapes of the induced pulses and their Fourier analysis. Moreover, the assumption which we made at the beginning about the constancy of the various times of flight is at best only an approximation. τ_1 and τ_2 will vary on account of different thermal emission velocities and path lengths. τ_3 and τ_4 will also be statistically distributed on account of the fluctuation in secondary-emission delay times and initial velocities.

As mentioned above, the experimental approach to the problem of the frequency response of multipliers is the only one possible. Experiments, however, cannot separate effects due to input structure from those described in this section. Which part of the valve is responsible for the initial fall of gain will depend upon the value of the ratio:—

$$\frac{\tau}{\tau_2 + \tau_3 + \tau_4}$$

where τ is the cathode to grid time of flight. Assuming that τ is given by (4) with the values of V and d corresponding to the 6AM6, we have $\tau \approx 2 \times 10^{-9}$ seconds. In absence of space charge between secondary emitter and anode the time of flight is given by :—

$$\tau_4 = \frac{10^{-7}d'}{5{\cdot}9\,V^{\frac{1}{2}}}$$

where d' is the secondary emitter to anode clearance, and V' the voltage difference. Putting $d' = 0{\cdot}05$ cm and $V' = 100$ V, we obtain $\tau_4 \approx 10^{-10}$. Since $\tau_2 < 4$, the fall of gain could be due to the multiplier section only if τ_3 (secondary-emission delay time) were much longer than 10^{-9} seconds. This appears to be unlikely.* We should expect, therefore, that with the above approximate clearances there would be no serious limitations on high-frequency performance due to the multiplier section.

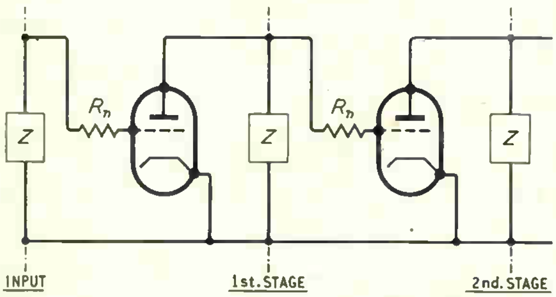

Fig. 4. *Three-stage high-frequency amplifier.*

4. Noise Considerations

We must now complete the comparison between a valve and a multiplier by considering the noise aspect.

It is well-known that secondary-emission multiplication increases the shot noise of the primary current by a certain factor. For the sake of the present discussion it will be sufficient to assume that the equivalent noise resistance R_n' for a multiplier is given by:—

$$R_n' = \gamma R_n \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (17)$$

where γ is a constant larger than unity and R_n is the equivalent noise resistance of a pentode of similar input structure to the multiplier.

Fig. 4 represents a three-stage amplifier for high frequencies. In this schematic representation Z is the load impedance which is nearly equal to the input impedance of the valves. If we let G be the gain per stage, v_s the r.m.s. value of the signal voltage and g the mutual conductance of the valves, then:—

$$G = g\,|Z|$$

The output voltage after n stages of amplification is:—

$$V_s = G^n v_s \quad \ldots \quad \ldots \quad \ldots \quad (18)$$

To obtain the signal-to-noise ratio we have to compute the output noise voltage. At the input the noise mean-square voltage due to circuit resistances is:—

*See McKay: Advances in Electronics, I (N.Y.1948), p. 66.

$$4\,k\,T\,R_z\,B$$

where R_z is the resistive component of Z. After one stage of amplification this voltage becomes:—

$$4\,k\,T\,B\,(R_z + R_n)\,G^2 + 4\,k\,T\,B\,R_z$$

After two stages:—

$$\{4\,k\,T\,B\,(R_z + R_n)\,G^2 + 4\,k\,T\,B\,R_z\}\,G^2 + 4\,k\,T\,B\,R_z$$

and so on. Summing the geometrical series for n stages, we obtain the output mean-square noise voltage:—

$$V_n^2 = 4\,k\,T\,B\left\{R_nG^{2n} + \frac{1 - G^{2n+2}}{1 - G^2}R_z\right\} \quad \ldots \quad (19)$$

Hence the signal-to-noise ratio is:—

$$\frac{V_s^2}{V_n^2} = \frac{G^{2n}v_s^2}{4\,k\,T\,B\left\{R_nG^{2n} + \dfrac{1 - G^{2n+2}}{1 - G^2}R_z\right\}} \quad \ldots \quad (20)$$

We are now interested in the comparison of signal-to-noise ratios for amplifiers using alternatively valves and multipliers.

Let us consider two specific examples. Suppose first that we want to make an amplifier with an overall gain of 120 db and a bandwidth of approximately 3 Mc/s. From Section 3 it follows that we can use the 6AM6 pentodes with a gain per stage of approximately 10. The amplifier, therefore, must have six stages. From (20) the signal-to-noise ratio is:—

$$\frac{V_s^2}{V_n^2} = \frac{v_s^2}{4\,k\,T\,B\,(R_z + R_n)}$$

if we neglect 1 against G^2. We can obtain the same overall gain of 120 db by using four stages of multipliers with a gain per stage of 32. Since 3·2 is a reasonable value of δ such an amplifier seems to be a practical proposition. The signal-to-noise ratio will be in this case:—

$$\frac{V_s^2}{V_n^2} = \frac{v_s^2}{4\,k\,T\,B\,(R_z + \gamma R_n)}$$

As we see, the use of multipliers will cut down the number of stages but will, at the same time, decrease the signal-to-noise ratio. The question whether it is permissible to use multipliers cannot be answered in general without the knowledge of the values of γ, v_s and the required signal-to-noise ratio. It seems that if the signal-to-noise ratio is the only limitation (the valves being otherwise operated well within the limits of their performance set up in Section 3) the use of multipliers may not be advantageous.

Let us consider now the second example. Suppose that the operating conditions of the valves are such that the maximum gain per stage is too small to be of any use. In this case we obtain the signal-to-noise ratio from (20) by taking the limit as G tends to unity:—

$$\underset{G\to 1}{\mathrm{Lt}} \frac{V_s^2}{V_n^2} = \frac{v_s^2}{4\,k\,T\,B\,\{R_z(1+n)+R_n\}}$$

In this case we clearly do not get any benefit from the amplifier and the signal-to-noise ratio decreases as the number of stages is increased. A better performance may be obtained by the use of multipliers which can still supply a gain δ times larger. Substituting δ for G in (20) and γR_n for R_n we obtain:—

$$\frac{V_s^2}{V_n^2} = \frac{\delta^{2n} v_s^2}{4\,k\,T\,B\left\{\gamma R_n \delta^{2n} + \dfrac{1-\delta^{2n+2}}{1-\delta^2} R_z\right\}}$$

In this case as we increase the number of stages the signal-to-noise ratio will not tend to zero as before but to a fixed value given by:—

$$\underset{n\to\infty}{\mathrm{Lt}} \frac{V_s^2}{V_n^2} = \frac{v_s}{4\,k\,T\,B\left\{\gamma R_n + R_z \dfrac{\delta^2}{\delta^2-1}\right\}} \quad \ldots \quad \ldots \quad \ldots \quad (21)$$

If the value of signal-to-noise ratio given by (21) is not smaller than the required one, we can thus, at least in principle, amplify the signal by using a sufficiently large number of stages.

We can summarize these results as follows:—

(a) If the valves are operated well within the limits of their performance and the signal-to-noise ratio *cannot exceed a certain specified value* then the reduction of the number of stages by using multipliers may not be permissible on account of the increased noise.

(b) If the operating conditions are such that the valves cannot supply any useful gain, we have to use multipliers. The signal-to-noise ratio will tend to the value given by (21) as we increase the number of stages.

(c) It does not appear possible to give a general ruling in favour of the multipliers. The decision to use them must be based on the following considerations:—

(i) Determination of the maximum possible gain per stage for a set of operating conditions (frequency and bandwidth).

(ii) Possibility of obtaining the required overall gain and signal-to-noise ratio in a practical number of stages.

Acknowledgment

The author is indebted to Mr. B. Wilkinson, Managing Director of Electronic Tubes, Ltd., for permission to publish this paper. The arguments presented in Section 2 are essentially due to Dr. C. S. Bull. To him and to Mr. E. Grodszinsky the author is indebted for many discussions.

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NEW BOOKS

The M.K.S. System of Units: A Guide for Electrical Engineers

By T. McGreevy, M.Sc.Tech., M.I.E.E. Pp. 283 + xii. Sir Isaac Pitman & Sons, Ltd., Parker Street, Kingsway, London, W.C.2. Price 21s.

The first thing expected of a guide is that he shall know where he is going himself; the next is that he shall choose a route within the capacity of his charges; and finally he must inspire sufficient confidence for them to follow him. The author fulfils his undertakings in all these respects, and this book is a most useful contribution to the growing literature on the teaching of electricity in terms of rationalized m.k.s. units. Two classes of reader have been considered, the practising electrical engineer, and the teacher of whole-time and part-time classes in the electrical-engineering departments of technical colleges. The radio engineer has not been specifically catered for, and the reason is clear; the advantages of the m.k.s. system in field theory have been realized for so long that he has, so to speak, been sitting on the summit for years and needs neither encouragement to make the climb nor the assistance of a guide en route. Nevertheless, as physical data still emerge originally in c.g.s. units, the careful discussion of the relation between these and the m.k.s. units, and the conversion tables in the appendix, make this a generally useful reference book; however well you know your bearings, you are always better off with a good map.

In the first five chapters, mainly for the engineer whose chief concern is in applying standard formulae, the numerical changes in going from one system to another are explained, prefaced by an account of the relationships involved. The remainder of the book is largely for teachers. It would be hard to persuade anyone who has not wrestled with elementary students that merely changing the system of units can lead to clearer teaching; yet this is indeed the case. In the c.g.s. system most of the sums come right even if you confuse B and H—so a proper treatment seems needlessly pedantic to the beginner. In the m.k.s. system things have to be worked out properly from the start. While agreeing with the author that the point magnetic pole is a bad starting-point for a system of measurement, the reviewer feels a little kinder towards it as a concept; it is just as respectable as a magnetic quantum number, and just about as tangible. The m.k.s. system could readily be used with the old teaching sequence; but once it becomes urgent to distinguish B and H a much better sequence is possible, using the basic electrical unit, the ampere, as the logical starting-point. The real beauty of the system is that it casts no impediment in the way of a course with a sound practical foundation which both satisfies the teacher's conscience and is also manifestly practical to the student.

There are several ways of doing this. One is to base the treatment on the mechanical force between circuit-elements, which has the disadvantage, for the engineer who is concerned with the magnetic properties of materials, that H hardly comes into the scheme at all.

Another is what may be called the 'uniform-field' approach, which probably commends itself most to the teacher of pure physics or the radio student whose main interest later on is in field theory. Mr. McGreevy introduces H as the concept which is described vectorially in advanced work as the gradient of a scalar function of position. Associated with a current I in N parallel conductors is the magnetomotive force NI; this is the line-integral of H. A very sound first step, though the term 'm.m.f.-gradient' for H is perhaps unhappy. The experimental starting point is the measurement of B using the fluxmeter, and an ingenious method is suggested for measuring H within an iron anchor-ring with an air-gap.

In the later chapters the more important formulae for instruments, machines, and power calculations are derived from first principles. In the chapter on telecommunications, the only cases directly considered are conversions of purely acoustic quantities into m.k.s. units. The starting-point in electrostatics is the mechanical force between the plates of a capacitor, again a direct relationship to simple experiment.

Some earlier writers gave undue prominence to recondite dimensional arguments, side-tracking the main issues and laying themselves open to retorts in kind. The author does not do this, of course, and in the chapter on dimensions (which in a work as comprehensive as this was probably included for completeness only) points out that checking by units is often much more profitable than trying to check a formula by dimensions. This is, indeed, felt by many people to be a valuable feature of the m.k.s. unit system in which it can so easily be done.

Teachers will welcome the careful attention to practical teaching problems—in particular, the different approaches suggested for part-time and whole-time students—and also the references in the text which embrace all the really important English publications on m.k.s. units. The book can be thoroughly recommended as an introduction and guide to the system.

G. R. N.

Electron Optics

By O. KLEMPERER. Pp. 471 + xii with 168 illustrations. Cambridge University Press, 200 Euston Road, London, N.W.1. Price 50s.

This is one of the Cambridge Monographs on Physics; it is a second edition but the first edition, published in 1939, when the subject of electron optics was at an early stage of development, was a small tract, a fraction of the size of the present volume. The author, who is Assistant Professor at the Imperial College, is a well-known authority on the subject, having not only published many papers but also taken out a number of patents. The object of the book is not only to introduce the student to a specialized subject, but also to present the research worker or designer of electron-optical gear, with the basic principles and the most essential quantitative information.

The first five of the twelve chapters give an outline of the principles of electron lenses, both electrostatic and magnetic, field plotting and ray tracing; two chapters are then devoted to lens errors, which are discussed from an essentially experimental point of view, and a chapter to the effects of space charge. Chapters IX and X deal with various emission systems and line focusing. Chapter XI deals very fully with deflecting fields, both electric and magnetic, and in the final chapter the author discusses the application of electron optics in industry and research.

The book concludes with a very extensive bibliography of about 500 references. It is a pity that more care was not taken with the Tables of units in the appendix, in which Joules and Maxwells are always printed with capital letters and amperes, coulombs, gauss, newtons and henrys with small letters; also Ω is not the symbol for resistance; this should be R, and Ω should be in the next column as an abbreviation or symbol for the ohm. The statement that a pressure of a newton per square metre is equal to 10 dynes/cm^2 (or 10^5 bar or $7{\cdot}5 \times 10^{-3}$ mm Hg) is wrong; a bar $= 10^6$ dynes/cm^2 and therefore 10 dynes/cm^2 $= 10^{-5}$ bar. These are minor details which detract little from the value of the book, which can be thoroughly recommended to anyone interested in electron optics.

G. W. O. H.

Faster than Thought

Edited by B. V. BOWDEN. Pp. 416 + xix. Sir Isaac Pitman & Sons, Ltd., Parker Street, Kingsway, London, W.C.2. Price 35s.

A foreword is contributed by the Earl of Halsbury, and 23 authors have assisted the editor, the majority of them being concerned with only one chapter. They are listed, together with the chapters to which they contributed, on page xv. Their collective achievement is remarkable. The fundamental information about basic principles and circuits, the way in which a typical modern digital computing machine is built of these basic circuits, performance, maintenance and 'programming' are all there, and yet the reader not concerned with the detailed operation is given a clear and often humorous account of the historical development of such machines, a description of important British machines, a discussion of actual and potential applications of these machines to logical problems, technology, Government calculations, commerce, and even to chess, draughts and "Nim."

The first chapter is historical; the life and work of Charles Babbage are prominent. This is fitting because Babbage visualized most of the ideas needed for modern large-scale digital computers, but valves and 'electronics' make it very much easier now than in Babbage's lifetime to carry out these ideas practically. The frontispiece is a portrait of Ada Augusta, Countess of Lovelace and daughter of Lord Byron, one of the few contemporaries who really understood what Babbage was trying to do.

Circuit components and the organization of a typical machine are discussed in the second and third chapters; only here are the basic units required for the construction of a digital computer considered in detail. Elsewhere the machine can be regarded as capable of performing addition, subtraction, multiplication, division, and some logical operations on data supplied to it in the right form. The machine must also be able to store such data and to set out the results of its operations suitably. Failures apparently occur remarkably seldom; performance and maintenance are considered in Chapter 4. The most complicated mathematical operation can be 'broken down' into a number of fundamental operations; a 'convenient working-list' of such fundamental operations is given in Chapter 5. Translating the problem into 'machine language' is called 'programming,' and this often requires a large proportion of the total effort.

The principal British digital calculating machines are discussed in Chapters 6 to 13, namely the University of Manchester computing machine (M.A.D.M.), the Cambridge 'Electronic Delay Storage Automatic Calculator' (E.D.S.A.C.), the National Physical Laboratory (Teddington) 'Automatic Computing Engine' (A.C.E.), the Harwell electronic digital computer, the Telecommunications Research Establishment (Malvern) parallel electronic digital computer, the Imperial College (London) computing engine (a small relay machine of parallel type), the Royal Aircraft Establishment (Farnborough) sequence-controlled calculator, and various machines at the Birkbeck College (London) computation laboratory. American machines are only briefly reviewed in Chapter 14 because they are adequately discussed elsewhere.

Applications in various fields are considered in Chapters 15–24. In some cases, such as accountancy, there are processes which are simple and many times repeated; these could be taken over by a machine entirely. The way in which the Manchester machine could be used to determine the wages, P.A.Y.E. deductions, etc., for a factory employing about 4,000 people making light electrical equipment of all kinds has been worked out. In meteorology, however, the art of forecasting could never be reduced to a mechanical process, but the machine can be used to simplify our description of the weather. Complicated equations are available, for example, to tell us how much water is condensed out as cloud droplets when air is lifted. Solutions for several cases can be obtained mechanically and represented on a diagram which can tell the forecaster all he wishes to know about likely condensation in a few minutes.

In Chapter 25 the performance of a machine at chess, draughts and "Nim" are considered; the scores of actual games of chess and draughts played by machines against men are included. Finally, in Chapter 26, the performance of a machine is compared with that of men with outstanding computational ability like Professor Aitken, and the limitations of a machine's powers of 'thought' are discussed.

The advantage of these machines is their ability to perform suitable operations in a few microseconds; their limitation is that they can only 'talk their own language,' and do what has been built into them. They cannot, at chess for example, take an overall view of the situation.

Considerable re-organization of modern life is required if we are to make adequate and efficient use of the potentialities of these machines. This book can be thoroughly recommended because it gives us a clear overall picture of how these machines work, when we should use them, and when we shall be wiser to call in the human expert. Sufficient further detail concerning basic circuits, individual machines, etc., is also available for the minority who require it.

J. W. H.

Die Laplace-Transformation und ihre Anwendung

By Paul Funk, Hans Sagan and Franz Selig. Pp. 106 + vii. Franz Deuticke, 4 Helferstorferstrasse, Wien 1, Austria. Price 16s. 8d.

Théorie des Circuits Impulsionnels

By H. Borg. Pp. 193 + viii. Editions de la Revue d'Optique, 165 rue de Sevres; 3 and 5 Boulevard Pasteur, Paris 15e, France.

Les Redresseurs en Simple Alternance

By J. Lecorguillier. Pp. 158. Editions Eyrolles, 61 boulevard Saint-Germain, Paris 5e, France. Price 1,850 francs.

Les Régimes Transitoires dans les Réseaux Electriques

By Paul Poincelot. Pp. 132. Gauthier-Villars, 55 Quai des Grands-Augustins, Paris, France.

Les Lignes a Retard et leur Utilisation

By Gaston Potier. Pp. 102 + viii. Gauthier-Villars, 55 Quai des Grands-Augustins, Paris, France.

These two books belong to the series of monographs "Les Filtres Electriques," edited by Pierre David.

Osram Valve Manual

Part 2.—Transmitting and Industrial Valves Engineering Data. Pp. 217. The General Electric Co. Ltd., Magnet House, Kingsway, London, W.C.2. Price 10s.

TECHNICAL PUBLICATIONS

Ferroxcube

Pp. 46. Mullard, Ltd., Century House, Shaftesbury Avenue, London, W.C.2.

After an opening chapter on magnetic ferrites, this booklet gives details of the electrical, magnetic and mechanical properties of Ferroxcube, together with design data for high-Q coils and a list of standard sizes of cores.

The Permanent Effect of Water on Varnished Coils

By H. R. Heap, B.Sc. E.R.A. Technical Report A/T 138. Pp. 8 with 7 illustrations. The British Electrical and Allied Industries Research Association, Thorncroft Manor, Dorking Road, Leatherhead, Surrey. Price 7s. 6d.

Measurement of Radio Interference in the Frequency Range 0·15 to 30 Mc/s. A Portable Measuring Set.

By S. F. Pearce and D. C. G. Smith. E.R.A. Publication M/T 116. Pp. 9 + 5 figures. The Electrical Research Association, Thorncroft Manor, Dorking Road, Leatherhead, Surrey. Price 10s. 6d.

Measurement of Radio Interference in the Frequency Range 0·15 to 30 Mc/s. A Mains Isolating Unit.

By J. Miedzinski and S. F. Pearce. E.R.A. Publication M/T117. Pp. 7 and 11 figures. The Electrical Research Association, Thorncroft Manor, Dorking Road, Leatherhead, Surrey. Price 12s. 6d.

Measurements of Atmospheric Noise at High Frequencies 1945-1951

By F. Horner, M.Sc., A.M.I.E.E. Radio Research Special Report No. 26. Pp. 40 + iv. Published for the Department of Scientific & Industrial Research by H.M. Stationery Office, York House, Kingsway, London, W.C.2. Price 1s. 9d.

Copper in Instrumentation

Pp. 152. Copper Development Association, Kendals Hall, Radlett, Herts.

Proceedings of the Western Computer Conference

Pp. 231. Published by the Institute of Radio Engineers, Inc., 1 East 79 Street, New York 21, N.Y., U.S.A. Price $3.50.

The conference was held by the Joint I.R.E.-A.I.E.E.-A.C.M. Computer Conference Committee at Los Angeles, California, on 4th–6th February 1953. The proceedings contain the following papers:—

Keynote and Luncheon Addresses.

"An Evaluation of Analog and Digital Computers," Panel Discussion.

"Commercial Applications—The Implication of Census Experience," J. L. McPherson.

"Payroll Accounting with Elecom 120 Computer," R. F. Shaw.

"Automatic Data Processing in Larger Manufacturing Plants," M. E. Salveson and R. G. Canning.

"Requirements of the Bureau of Old-Age and Survivors Insurance for Electronic Data Processing Equipment," E. E. Stickell.

"The Processing of Information-Containing Documents," G. W. Brown and L. N. Ridenour.

"Airplane Landing Gear Performance Solutions with an Electronic Analog Computer," D. W. Drake and H. W. Foster.

"The Equivalent Circuits of Shells used in Airframe Construction," R. H. MacNeal.

"Analog-Digital Techniques in Autopilot Design," W. T. Hunter and R. L. Johnson.

"Applications of Computers to Aircraft Dynamic Problems," B. Hall, R. Ruthrauff and D. Dill.

"The Snapping Dipoles of Ferroelectrics as a Memory Element for Digital Computers," C. F. Pulvari.

"Magnetic Reproducer and Printer," J. C. Sims, Jr.

"An Improved Cathode Ray Tube Storage System," R. Thorensen.

"Nonlinear Resistors in Logical Switching Circuits," F. A. Schwertz and R. T. Steinback.

"New Laboratory for Three-Dimensional Guided Missile Simulation," Louis Bauer.

"A New Concept in Analog Computers," Lee Cahn.

"A Magnetically Coupled Low-Cost High-Speed Shaft-Position Digitizer," A. J. Winter.

"The Solution of Partial Differential Equations by Difference Methods using the Electronic Differential Analyzer," R. M. Howe and V. S. Haneman.

"The Nordsieck Computer," Arnold Nordsieck.

BUREAU OF STANDARDS' PUBLICATIONS

Tables of Normal Probability Functions

Applied Mathematics Series 23. Pp. 344. Price $2.75.

Electrodeposition Research

Proceedings of the N.B.S. Semicentennial Symposium on Electrodeposition Research, held at the N.B.S., on 4th-6th December, 1951. Circular 529. Pp. 129. Price $1.50.

Reference Data for Orienting Quartz Plates by X-Ray Diffraction

By Catherine Barclay and Leland T. Sogu. Circular 543. Pp. 7. Price 15 cents.

Table of Dielectric Constants and Electric Dipole Moments of Substances in the Gaseous State.

By Arthur A. Maryott and Floyd Buckley. Circular 537. Pp. 29. Price 20 cents.

The four above publications can be obtained from the Government Printing Office, Washington 25, D.C., U.S.A.

TELEVISION SOCIETY'S EXHIBITION

The annual exhibition of the Television Society will be on 7th (6–9 p.m., members only), 8th (12–9 p.m.) and 9th (10 a.m.–9 p.m.) January 1954. It will be held in the Electrical Department of King's College, Strand, London, W.C.2, and free admission tickets are obtainable from the Television Society, 164 Shaftesbury Avenue, London, W.C.2.

MEETINGS

I.E.E.

14th December. "Will Transistors Oust Receiving Valves?" Discussion to be opened by E. H. Cooke-Yarborough.

5th January. "A Scaling Unit Employing Multi-Electrode Cold-Cathode Tubes" and "A Sensitive Pulse Trigger Circuit with a Stable Threshold", by K. Kandiah, M.A. "The Development of a Neutron Spectrometer for the Intermediate Energies", by F. S. Goulding, B.Sc., J. C. Hammerton, M.A., M. G. Kelliher, B.A., M.Sc., A. W. Merrison, B.Sc., and E. R. Wiblin, M.A.

These meetings will be held at the Institution of Electrical Engineers, Savoy Place, Victoria Embankment, London, W.C.2, and will commence at 5.30.

BRIT.I.R.E.

9th December. A Symposium of Vibration Methods of Testing: "Vibration Generators", by H. Moore; "Stroboscopes", by F. M. Savage; "Strain Gauges", by P. Jackson, and "Electronic Aids to Vibration Measurement", by R. K. Vinycomb, B.Sc.

6th January. "Some Factors in the Engineering Design of the V.H.F. Multi-channel Telephone Equipment", by W. T. Brown.

To be held at the London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1, at 6.30 p.m.

THE INSTITUTE OF NAVIGATION

18th December. "The Reduction of Sea and Rain Clutter in Marine Radars", by J. Croney; "A Racon Beacon for Reception by Civil Marine Radars", by C. Randall-Cook; "A Portable Beacon for Identifying Ships on Harbour Radars", by A. L. P. Milwright. To be held at the Royal Geographical Society, 1 Kensington Gore, London, S.W.7, commencing at 5 p.m.

STANDARD-FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)

Values for October 1953

Date 1953 October	Frequency deviation from nominal: parts in 10^8		Lead of MSF impulses on GBR 1000 G.M.T. time signal in milliseconds
	MSF 60 kc/s 1429-1530 G.M.T.	Droitwich 200 kc/s 1030 G.M.T.	
1	N.M.	—5	+ 7·3
2	—1·1	—5	+ 6·0
3	—1·1	—4	N.M.
4	—1·0	—4	N.M.
5	—1·0	—4	— 1·1
6	—1·0	—4	— 3·6
7	—1·0	—4	— 5·8
8	—1·0	—3	— 8·1
9	—1·0	—3	— 9·6
10	—1·0	—4	N.M.
11	N.M.	—4	N.M.
12	—0·9	—3	—14·6
13	—1·0	—1	—16·8
14	—0·9	—2	—17·6
15	N.M.	—2	N.M.
16	—1·2	—3	—19·3
17	—1·3	—2	N.M.
18	—1·2	—2	N.M.
19	—1·0	—2	—24·0
20	—0·9	—2	—25·5
21	—0·9	—1	—25·6
22	—0·9	—2	—28·0
23	—1·0	0	—29·1
24	—0·9	—2	N.M.
25	—0·9	0	N.M.
26	—0·8	—1	—32·6
27	—0·8	0	—34·5
28	—0·8	0	—37·3
29	—0·8	—1	—38·3
30	—0·8	0	—39·2
31	—0·8	0	N.M.

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

N.M. = Not Measured.

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

534.141 + 534.143] : 534.322.1 **3461**

The Relations between Electrical and Mechanical Production of Musical Sounds. The Question of Tone-Colour/Interval Circles.—R. Bierl. (*Z. angew. Phys.*, June 1953, Vol. 5, No. 6, pp. 231–237.)

534.15 **3462**

Study and Representation of a Complex Musical Tone.—A. Moles. (*Funk u. Ton*, June 1953, Vol. 7, No. 6, pp. 277–287.) German version of 1219 of May.

534.24 : 526.956.5 **3463**

Reflection of Sound in the Ocean from Temperature Changes.—R. R. Carhart. (*J. appl. Phys.*, July 1953, Vol. 24, No. 7, pp. 929–934.) Coefficients r and r' for 'sharp' and 'smooth' reflection are distinguished, according as the thickness of the layer in which a velocity change occurs is or is not negligible. Theory for the two cases is developed and values of r and r' as functions of angle of incidence corresponding to practical conditions are calculated.

534.6 : 621.395.625.3 **3464**

Storage Methods of Sound Measurement.—H. Etzold. (*Funk u. Ton*, June 1953, Vol. 7, No. 6, pp. 307–315.) Equipment is described which uses an A.E.G. Type-KL15 magnetophone as the storage unit. Details are given of the means adopted for providing a linear decibel scale for the intensity of the recorded sounds.

534.833.4 **3465**

Absorption Coefficient of Acoustic Materials.—R. Lamoral. (*Onde élect.*, July 1953, Vol. 33, No. 316, pp. 461–467.) Values of the absorption coefficient α of two materials were determined from reverberation-time measurements, using Sabine's formula. The frequencies used ranged from 125 c/s to 4 kc/s. The results obtained for samples of different surface area, or for samples of the same total area consisting of various numbers of separate pieces, differ so much that comparison between the values of α for different materials can only be reliable if the necessary measurements are carried out under specified conditions.

534.833.4 **3466**

Testing Sound-Absorbing Materials in a Reverberating Room.—B. D. Tartakovski & M. M. Efrussi. (*C. R. Acad. Sci. U.R.S.S.*, 21st Jan. 1952, Vol. 82, No. 3, pp. 373–376. In Russian.)

534.84 **3467**

Collation of the Reverberation Times of Berlin Sound-Recording Studios and Auditoria.—H. Dippner & H. J. Zemke. (*Frequenz*, March 1953, Vol. 7, No. 3, pp. 71–81.) Measurements have been made in churches, concert halls, film theatres and studios. Reverberation-time/frequency curves are shown with details of the size and seating capacity for each of the 52 cases dealt with, and many photographs of interiors.

534.84 : 534.861.1 : 621.396.712.3 **3468**

Architectural Acoustics of the Cologne Broadcasting Centre.—L. Müller. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, 1953, Vol. 5, Nos. 5/6, pp. 87–97.) A detailed account, with numerous illustrations, is given of the treatment of the floors, walls and ceilings of the various studios to obtain acoustical characteristics suitable for the differing requirements for chamber music, full orchestra or organ, speech, dance music, etc. Methods of acoustic damping in the foyer and corridors, by means of absorbing materials, resonators and diffusers of various types, are also described.

621.395.623.7 : 534.843 **3469**

The Loudspeaker in the Home.—P. J. Walker. (*J. Brit. Instn Radio Engrs*, July 1953, Vol. 13, No. 7, pp. 377–380.) A nonmathematical discussion of the problems of sound reproduction in relation to the acoustics of ordinary rooms. The concept of the ideal headphone is introduced and its limitations are indicated.

621.395.623.8 **3470**

Public-Address Systems in Generating Plants.—S. C. Bartlett. (*Trans. Amer. Inst. elect. Engrs*, 1951, Vol. 70, Part II, pp. 1804–1810.) For another account see 615 of March.

621.395.625.3 **3471**

Studies on Magnetic Recording: Part 1—Introduction.—W. K. Westmijze. (*Philips Res. Rep.*, April 1953, Vol. 8, No. 2, pp. 148–157.) A survey of the principles and historical development of methods of magnetic recording.

AERIALS AND TRANSMISSION LINES

621.315.212 : 621.315.615 3472

Liquid-Dielectric Radiofrequency Coaxial Cables.—R. M. Soria, C. C. Camillo & J. G. Krisilas. (*Proc. nat. Electronics Conf., Chicago*, 1952, Vol. 8, pp. 469–480.) An account is given of research in connection with the development of flexible coaxial cables using a liquid dielectric, driven through the cable under pressure, to obtain increased heat dissipation. Mineral oil and a silicone oil were found suitable. The results obtained show that a cable of the liquid-dielectric type can be constructed to operate with an input power 8 times that for the equivalent standard RG cable, and with satisfactory operation at high ambient temperatures and high altitudes.

621.392.09 3473

Surface-Wave Propagation over a Coated Conductor with Small Cylindrical Curvature in Direction of Travel.—K. Horiuchi. (*J. appl. Phys.*, July 1953, Vol. 24, No. 7, pp. 961–962.) Analysis indicates departures to be expected from the results obtained for a plane conductor [2097 of 1951 (Attwood)]. The effect of a bend in a surface-wave transmission line is to produce attenuation and to increase the field intensity on the outside of the bend.

621.392.21 : 621.317.343 3474

Characteristic Impedance of Rectangular Coaxial Transmission Lines.—Y. A. Omar & C. F. Miller. (*Trans. Amer. Inst. elect. Engrs*, 1952, Vol. 71, Part I, pp. 81–89.) The characteristic impedance of several coaxial rectangular transmission lines was measured by a standing-wave method at about 430 Mc/s. The results obtained enabled a relation between line dimensions and characteristic impedance to be developed from that for confocal elliptical lines.

621.392.21.017 3475

Effect of Losses in H.F. Transmission Lines.—S. Albagli. (*Onde élect.*, June 1953, Vol. 33, No. 315, pp. 270–273.) In transmission-line calculations where it is necessary to take account of losses, it is often assumed that the characteristic impedance is real. This is admissible as a first approximation, but it can lead to considerable errors in certain cases. To the second approximation the characteristic impedance is real only if the metal and dielectric loss coefficients are equal. Comparison is made between results obtained on this supposition and those obtained on the assumption that dielectric losses are negligible. In some cases the results differ by 100%.

621.392.211 3476

Propagation of Electromagnetic Waves along an Infinite Helical Slit.—S. A. Vakin. (*C. R. Acad. Sci. U.R.S.S.*, 1st May 1952, Vol. 84, No. 1, pp. 37–40. In Russian.) The field, inside and outside the cylindrical surface containing the slit, is determined by assuming a particular distribution of potential along the slit. A transcendental equation for the propagation constant is derived and solved for four particular values of δ, the helix pitch angle. The retardation of the wave, τ, which is a function of the propagation constant, is ~ 1 over a range of values of r_0, the radius of the cylinder. This theoretical result was confirmed experimentally.

621.392.26 3477

Matching Discontinuities in Waveguides.—J. C. Parr. (*Wireless Engr*, Oct. 1953, Vol. 30, No. 10, pp. 243–249.) Simple systematic procedure is described for matching both waveguide measuring equipment and other waveguide sections. The technique of four-screw tuning is used, and theory is outlined to show how it is possible to obtain any desired value of reflection factor. The different requirements for matching terminating sections, insertion devices or generators are discussed. Modifications appropriate for cases of large mismatch are indicated. The distance between tuning screws should be the smallest convenient multiple of $\lambda/8$.

621.392.43 : 621.396.67 3478

A Solution for a Practically Frequency-Independent Transition between a H.F. Coaxial Cable and a Balanced H.F. Transmission Line.—H. Graziadeï. (*Fernmeldetech. Z.*, July 1953, Vol. 6, No. 7, pp. 311–319.) Feeder arrangements for rhombic transmitter aerials are considered. The theory of a wide-band unbalance/balance impedance-matching transformer is given; design and applications are considered. A transformer consisting of a reactive loop and an exponential line in parallel with the output of a coaxial line has been designed to give a nearly level response over the 10–60-m band for a 60Ω/500Ω line impedance ratio.

621.396.67 3479

Polystyrene and Lucite Rod Antennas.—G. von Trentini. (*J. appl. Phys.*, July 1953, Vol. 24, No. 7, pp. 960–961.) Measurements were made with tapered rods thinner than those used by Horton & McKinney (1204 of 1952); values of gain obtained were somewhat higher and effects due to dielectric losses less accentuated.

621.396.67.012.71 3480

V.H.F. Aerial Radiation Pattern Measurements.—E. G. Hamer. (*Electronic Engng*, Oct. 1953, Vol. 25, No. 308, pp. 427–431.) Various measurement techniques are discussed and an indication is given of conditions to be satisfied to obtain polar diagrams free from errors due to induction fields, surface wave and variation of Brewster angle. Automatic machines for plotting radiation patterns are described.

621.396.674 3481

Loop Aerials for Portable Broadcast Receivers.—E. G. Beard. (*Philips tech. Commun., Aust.*, 1953, No. 2, pp. 16–18.) Theoretical considerations lead to a suggested design in which a small loop aerial is surrounded by a number of coplanar closed metallic loops, the induced e.m.f.'s in which are all transferred to the central aerial, thus giving high efficiency.

621.396.677 3482

Radiation Conductance of Axial and Transverse Slots in Cylinders of Elliptical Cross Section.—J. Y. Wong. (*Proc. Inst. Radio Engrs*, Sept. 1953, Vol. 41, No. 9, pp. 1172–1177.) Formulae and graphs are presented for the radiation field and the radiation conductance. The degree of curvature of the cylinder surface has a considerable influence on the radiation conductance for the transverse slot.

621.396.677 3483

Grids as Circuit Elements for Electromagnetic Waves in Space.—G. von Trentini. (*Z. angew. Phys.*, June 1953, Vol. 5, No. 6, pp. 221–231.) An account of experimental investigations of the characteristics of grid lenses similar to that previously described (937 of April), but using different types of grid-wire loading.

621.396.677 3484

Microwave Wide-Angle Scanner.—J. Brown. (*Wireless Engr*, Oct. 1953, Vol. 30, No. 10, pp. 250–255.) A study is made of microwave lenses of the type comprising a sphere whose refractive index varies with distance from the centre. In previously described lenses of this type [e.g. 2723 of 1950 (Jones)] scanning is performed by moving the feed over the lens surface. Mechanical problems can be simplified by arranging for the feed to move on a smaller sphere concentric with the lens. A suitable law is derived for the variation of refractive index in this case.

621.396.677 **3485**

Experimental Verification of the Metal-Strip Delay-Lens Theory.—S. B. Cohn. (*J. appl. Phys.*, July 1953, Vol. 24, No. 7, pp. 839–841.) Measurements made on five metal-strip delay structures, three in the form of waveguide elements and two free-space arrays, are described. Only one of four suggested formulae for refractive index is in good agreement with experiment over the full practical range. This is based on a transmission-line equivalent circuit taking proximity effects into account.

621.396.677 **3486**

Excitation and Radiation Properties of Microwave Lenses.—K. Hurrle. (*Fernmeldetech. Z.*, July 1953, Vol. 6, No. 7, pp. 332–337.) The conditions for the optimum excitation of a lens and the resulting radiation pattern are derived. Experimental results are given for a dielectric lens and a path-length lens [3058 of 1949 (Kock)] operating at a wavelength of 6·1 cm.

621.396.677 : 535.326 : 621.317.3 **3487**

Microwave Measurements on Metallic Delay Media.—Cohn. (See 3652.)

621.396.677.011.21 **3488**

Input Impedance of Folded-Dipole Antennas.—R. E. Beam & P. Andris. (*Proc. nat. Electronics Conf., Chicago*, 1952, Vol. 8, pp. 678–691.) A general expression for the input impedance is derived in terms of the self- and mutual-radiation impedances and transmission-line impedances of the two conductors forming the folded dipole. Integral-equation methods are used to determine approximate values of the radiation impedances for sinusoidal current distribution. Experimental and theoretical curves representing the resistive and reactive components of the input impedance as functions of frequency are given for two folded dipoles, one with both conductors of $\frac{7}{8}$-in. Cu tube and the other with fed elements of $\frac{3}{8}$-in. tube, the rest being $\frac{7}{8}$-in. In some cases the gap capacitance at the feed point was taken into account in the calculations.

621.396.677.029.63 **3489**

Helical-Beam Antenna Performance.—E. F. Harris. (*Commun. Engng*, July/Aug. 1953, Vol. 13, No. 4, pp. 19–20 . . 45.) Helical aerials for wide-band point-to-point radio communication in the 450–470-Mc/s, 890–960-Mc/s and 1·75–2·11-kMc/s bands are described and design details given. The aerial gains are 13, 16·5 and 20 db respectively, using a single helix, two helices or four helices. If helices of opposite sense are used for transmission and reception, a discrimination of about 20 db is obtained between the direct beam and the ground-reflected beam.

621.396.677.1 : 523.72 : 621.396.822 **3490**

New Techniques in Radio Astronomy.—J. D. Kraus & E. Ksiazek. (*Electronics*, Sept. 1953, Vol. 26, No. 9, pp. 148–152.) A broadside array of 24 pairs of helical aerials with axes parallel, mounted on a pivoted ground screen, has been installed at the Ohio State University. Each helix is 10 ft long, 15 in. in diameter, and has 10 turns. A beam width of 1·2° is obtained at 250 Mc/s. The receiving equipment is described and typical records are shown.

621.396.677.5 : 621.3.042.12 **3491**

The Receiving Loop with a Hollow Prolate Spheroidal Core.—J. R. Wait. (*Canad. J. Technol.*, June 1953, Vol. 31, No. 6, pp. 132–137.) The case of a loop with a solid spheroidal core has previously been considered (1917 of July). The relative gain is calculated for a loop wound symmetrically round the centre of a hollow shell of ferromagnetic material with permeability μ of 20, 50, 200 and 500. Neglecting core losses, a hollow core of moderate length is more efficient than a solid core of the same mass, particularly for high values of μ.

CIRCUITS AND CIRCUIT ELEMENTS

621.3(083.74) **3492**

The Standardization of Symbols and the Arrangement of Electronic Circuit Diagrams.—L. H. Bainbridge-Bell. (*J. Brit. Instn Radio Engrs*, July 1953, Vol. 13, No. 7, pp. 339–347. Discussion, pp. 347–353.) Arrangement of circuit diagrams to give a clear indication of the operation is advocated.

621.3.011.22.025 **3493**

Realization of Alternating-Current Resistance.—U. Kirschner. (*Funk u. Ton*, June 1953, Vol. 7, No. 6, pp. 298–306.) Methods of realizing resistance functions by arrangements of the partial-fractions or the continued-fractions type are described. Possible arrangements are shown in two tables.

621.3.012.11 **3494**

Geometrical Transformation of Impedance Diagrams.—H. Briner & W. Graffunder. In 2576 of September please change the last word in the abstract to 'described'.

621.3.066.6 **3495**

The Effect of Inductance on Fine Transfer between Platinum Contacts.—J. Warham. (*Proc. Instn elect. Engrs*, Part I, July 1953, Vol. 100, No. 124, pp. 163–168.) An investigation was made with Pt contacts in a 6-V circuit breaking currents of the order of 1 A; the circuit inductance was varied from 0·5 to 10 μH. Measurements of the volume transferred and examination of the surface structure indicate that there are two types of transfer: (*a*) true bridge transfer, independent of inductance, and (*b*) short-arc transfer, dependent on the inductance.

621.3.066.6 **3496**

The Behaviour of Metallic Contacts at Low Voltages in Adverse Environments.—A. Fairweather. (*Proc. Instn elect. Engrs*, Part I, July 1953, Vol. 100, No. 124, pp. 174–182.) The problem discussed is that of obtaining a metal-to-metal contact mechanically when the contacts are contaminated by dust or grease or are coated with films produced by adsorption, tarnishing or corrosion. Various types of contact are considered. Principles are outlined which provide a basis for design and testing; a new technique is described for the continuous dry lubrication of sliding contacts in mechanical and electrical systems.

621.3.066.6 : 621.314.58 **3497**

Long-Life Contacts for Unidirectional Currents of 1–20 Amperes.—A. L. Allen. (*Proc. Instn elect. Engrs*, Part I, July 1953, Vol. 100, No. 124, pp. 158–162.) An experimental investigation was made of fundamental physical phenomena concerned in the operation of contacts in vibratory converters; the influence of voltage and current on the direction of migration of material was studied. Long life was obtainable with Pt contacts or with contacts of dissimilar metals, but on heavy duty an adequate life was obtained only with W contacts in a low-oxygen atmosphere.

621.3.066.6 : 621.396.822 **3498**

Noise of Metal Contacts.—F. A. P. M. Theunissen. (*Appl. sci. Res.*, 1953, Vol. B3, No. 3, pp. 201–208.) Measurements were made of the small voltage fluctuation produced by the passage of direct current through ball contacts. Current values from 1 to about 50 mA were used, and the influence of contact resistance, pressure and temperature was investigated, for steel, copper and gold contacts. A possible mechanism accounting for the results is suggested.

621.314.2 **3499**

Design of Unequal-*Q* Double-Tuned Transformers.—S. Deutsch. (*Trans. Amer. Inst. elect. Engrs*, 1952, Vol. 71, Part I, pp. 314–320.) Equations and design curves are given which should simplify calculations of certain of the transformer characteristics.

621.314.22.015.7 **3500**

A Turns Index for Pulse Transformer Design.—H. W. Lord. (*Trans. Amer. Inst. elect. Engrs*, 1952, Vol. 71, Part I, pp. 165–168.) Full paper. See 1250 of May.

621.314.222 : 621.314.6 **3501**

Transient Conditions in a Transformer Supplying Energy to a Half-Wave Rectifier Circuit.—P. N. Martin. (*Trans. Amer. Inst. elect. Engrs*, 1951, Vol. 70, Part II, pp. 1468–1479.) Analysis is given for a simplified equivalent circuit, leakage reactance and core loss being neglected. Experimental results confirm the theory.

621.314.263 **3502**

The Magnetic-Cross Valve.—H. J. McCreary. (*Trans. Amer. Inst. elect. Engrs*, 1951, Vol. 70, Part II, pp. 1868–1874. Discussion, pp. 1874–1875.) Further details are given of the device previously described (2727 of 1950) and of its practical applications, with circuit diagrams and illustrations of the 'power ringers' converting 60-c/s power to 30 c/s and 20 c/s respectively.

621.314.3† **3503**

On the Mechanics of Magnetic-Amplifier Operation.—R. A. Ramey. (*Trans. Amer. Inst. elect. Engrs*, 1951, Vol. 70, Part II, pp. 1214–1222. Discussion, pp. 1222–1223.)

621.314.3† **3504**

Magnetic Amplifiers of the Balance Detector Type — their Basic Principles, Characteristics, and Applications.—W. A. Geyger. (*Trans. Amer. Inst. elect. Engrs*, 1951, Vol. 70, Part II, pp. 1707–1718. Discussion, pp. 1718–1720.)

621.314.3† **3505**

Predetermination of Control Characteristics of Half-Wave Self-Saturated Magnetic Amplifiers.—H. Lehmann. (*Trans. Amer. Inst. elect. Engrs*, 1951, Vol. 70, Part II, pp. 2097–2103. Discussion, p. 2103.)

621.314.3† **3506**

Bibliography of Magnetic Amplifier Devices and the Saturable-Reactor Art.—J. G. Miles. (*Trans. Amer. Inst. elect. Engrs*, 1951, Vol. 70, Part II, pp. 2104–2123.) A very comprehensive bibliography of 901 published books, articles, and patents, for the period 1887 to mid-1951.

621.314.3† **3507**

On the Control of Magnetic Amplifiers.—R. A. Ramey. (*Trans. Amer. Inst. elect. Engrs*, 1951, Vol. 70, Part II, pp. 2124–2128.)

621.314.3† **3508**

Steady-State and Transient Analysis of an Idealized Series-Connected Magnetic Amplifier.—L. A. Pipes. (*Trans. Amer. Inst. elect. Engrs*, 1951, Vol. 70, Part II, pp. 2129–2133. Discussion, pp. 2133–2135.)

621.314.3† **3509**

Series-Connected Magnetic Amplifier with Inductive Loading.—T. G. Wilson. (*Trans. Amer. Inst. elect. Engrs*, 1952, Vol. 71, Part I, pp. 101–110.) Full paper. See 338 of February.

621.314.3† **3510**

The Effective Feedback Ratio of Magnetic Amplifiers.—L. A. Finzi, G. F. Pittman, Jr, & H. L. Durand. (*Trans. Amer. Inst. elect. Engrs*, 1952, Vol. 71, Part I, pp. 157–164.)

621.314.3† **3511**

Magnetic Amplifiers of the Self-Balancing Potentiometer Type.—W. A. Geyger. (*Trans. Amer. Inst. elect. Engrs*, 1952, Vol. 71, Part I, pp. 383–395.) For a digest see *Elect. Engng, N.Y.*, April 1953, Vol. 72, No. 4, p. 294.

621.314.3† **3512**

High-Speed Magnetic Amplifier.—L. J. Johnson. (*Elect. Mfg*, Nov. 1952, Vol. 50, No. 5, pp. 98–101 . . 324.)

621.314.7 : 621.3.015.3 **3513**

Transient Analysis of Junction Transistor Amplifiers.—W. F. Chow & J. J. Suran. (*Proc. Inst. Radio Engrs*, Sept. 1953, Vol. 41, No. 9, pp. 1125–1129.) In analysing transistor operation, transit-time effects must be taken into account; this can be done to a sufficiently close degree of approximation by including in the equivalent circuit an *RC* network in series with an idealized delay line. Only simple Laplace transforms are then involved in the response calculations. The theory is supported by experimental results.

621.314.7 : 621.3.015.7 **3514**

Pulse Response of Junction Transistors.—N. H. Enenstein & M. E. McMahon. (*Trans. Inst. Radio Engrs*, June 1953, No. PGED-3, pp. 5–8.) The case of low load resistance is considered theoretically and experimental results are shown graphically. The response characteristics are discussed. Good qualitative and fair quantitative agreement was found between predicted characteristics and those determined from pulse and a.c. measurements.

621.314.7 : 621.318.57 **3515**

The Phase-Bistable Transistor Circuit.—R. H. Baker, I. L. Lebow, R. H. Rediker & I. S. Reed. (*Proc. Inst. Radio Engrs*, Sept. 1953, Vol. 41, No. 9, pp. 1119–1124.) Properties of transistor switching circuits are discussed. A unit is described which comprises a commutating ring of two transistors each connected as a one-shot multivibrator, the two being alternately triggered by a series of clock pulses. The arrangement is used in a binary counter, with a clock-pulse frequency of 500 kc/s.

621.314.7 : 621.318.57 **3516**

A Transistors in Trigger Circuits.—S. Walter. (*Proc. Inst. Radio Engrs*, Sept. 1953, Vol. 41, No. 9, p. 1190.) A simple circuit using a single transistor is described.

621.314.7 : 621.396.615.029.3 **3517**

Low-Distortion Transistor Audio Oscillator.—P. G. Sulzer. (*Electronics*, Sept. 1953, Vol. 26, No. 9, pp. 171–173.) Design details of an oscillator operating at any one of the four frequencies 20, 200, 2 000 and 20 000 c/s, and including 3 *p-n-p* junction transistors. Series resistors are used to stabilize emitter currents, and a tungsten-filament lamp to control amplitude. Maximum available output is 1 V. The low-noise characteristics are ascribed to the very narrow operating bandwidth.

621.314.7 : 621.396.645 **3518**

The Common-Collector Transistor Amplifier at Carrier Frequencies.—F. R. Stansel. (*Proc. Inst. Radio Engrs*, Sept. 1953, Vol. 41, No. 9, pp. 1096–1102.) Formulae are derived for the a.f. operating parameters of the grounded-collector ('common-collector') circuit. The variation of the transistor current-amplification factor α with frequency is investigated. The effect of operating conditions on the cut-off frequency f_c (i.e., the frequency at which α drops to $1/\sqrt{2}$ of its a.f. value) is discussed. Modified formulae are hence derived which are valid up to about $2f_c$. Experimental results support the theory presented. Point-contact, *n-p-n*-junction and early *p-n-p*-alloy transistors were studied.

621.314.7 : 621.396.645 **3519**

Complementary-Symmetry Transistor Circuits.—R. D. Lohman. (*Electronics*, Sept. 1953, Vol. 26, No. 9, pp. 140–143.) Under normal bias conditions, the current which flows in each lead to an electrode of a *p-n-p* transistor is the negative of the corresponding current in an *n-p-n* transistor; this is termed static symmetry. The polarity of an input signal that will increase conduction in a *p-n-p* transistor is the opposite of that for an *n-p-n* transistor; this is termed dynamic symmetry. Circuits are described that use (*a*) static symmetry, (*b*) dynamic symmetry, (*c*) both types of symmetry. These include a stabilized direct-coupled class-A amplifier with a voltage gain of 660, a pulse circuit which converts a 0·25-V half sine-wave to 20-V pulses, a class-B power output circuit using four transistors with overall feedback and giving a gain of about 30 db when working into a 16-Ω load, and a television vertical-deflection system with a class-B output stage connected directly to the deflection yoke. See also 2583 of September (Sziklai).

621.314.7 : 621.396.645 **3520**

Transistor Theory and Transistor Circuits.—H. E. Hollmann. (*Arch. elekt. Übertragung*, July 1953, Vol. 7, No. 7, pp. 315–327.) Simple theory for transistors is developed from that for thermionic valves on the basis of the duality principle and by comparison with the aperiodic retarding-field valve. Semidynamic rather than static characteristics are obtained, by biasing the collector by means of a resistor. These characteristics can be converted to the usual static ones by a coordinate transformation. Parameters are measured by means of a transistor bridge equivalent to that used for obtaining the dynamic characteristics of a thermionic valve. The parameters found are related to the constant-current values of the parameters hitherto used. Simple formulae are derived for the voltage and power amplification of grounded-base and grounded-emitter circuits. Use of the negative static resistance of a grounded-emitter transistor for measuring the resonance resistance of oscillatory circuits is discussed.

621.314.7 : 621.396.645.36.029.4 **3521**

Push-Pull Transistor Amplifiers.—J. I. Missen. (*Wireless World*, Oct. 1953, Vol. 59, No. 10, pp. 467–470.) A description is given of a practical a.f. power amplifier using point-contact transistors and developed from the class-B thermionic-valve amplifier by application of the duality principle. This amplifier can deliver $>$ 400 mW at $<$ 10% harmonic distortion.

621.314.7.012.8 **3522**

Equivalent-Circuit Diagrams of the Transistor.—W. Klein. (*Frequenz*, March 1953, Vol. 7, No. 3, pp. 59–60.) Different representations for transistors are derived from the equivalent circuit of the nonreversible quadripole (3031 of 1952):–(*a*) a two-valve circuit; (*b*) a one-valve circuit with a lossless transformer.

621.314.7.012.8 **3523**

Transistors: Theory and Application: Part 7—Equivalent Transistor Circuits and Equations.—A. Coblenz & H. L. Owens. (*Electronics*, Sept. 1953, Vol. 26, No. 9, pp. 156–161.) Relations between transistor parameters, circuit elements and performance are derived by application of Kirchhoff's mesh equations for the transistor equivalent circuit. Part 6: 3444 of November.

621.316.726.078.3 **3524**

Theory of A.F.C. Synchronization.—W. J. Gruen. (*Proc. Inst. Radio Engrs*, Sept. 1953, Vol. 41, No. 9, p. 1171.) Correction to paper noted in 3214 of November.

621.316.84 : 539.23] : 621.317.727.029.63 **3525**

Fabrication of Radio-Frequency Micropotentiometer Resistance Elements.—L. F. Behrent. (*J. Res. nat. Bur. Stand.*, July 1953, Vol. 51, No. 1, pp. 1–9.) Stable low-resistance disk-type resistors are required for the Bureau of Standards r.f. micropotentiometer previously described (1712 of 1951). Various processes for forming thin-film resistors are considered. Three methods are described by which it is possible to produce resistors stable to within ± 1%, viz., (*a*) high-temperature firing, (*b*) evaporating and plating, (*c*) use of disks cut from sheets of carbon film deposited on bakelite. A chemical-reduction process is under investigation.

621.318.4 **3526**

Copper Eddy-Current Losses in Coils with Carbonyl-Iron and Ferrite Cores.—J. Brackmann & J. Frey. (*Frequenz*, July 1953, Vol. 7, No. 7, pp. 185–191.) The design of low-loss short multilayer coils and long single-layer coils is considered. Because core losses are much reduced with modern ferrite cores, the relative importance of the copper eddy-current losses is greater. These losses and their dependence on the coil and core parameters are discussed in detail.

621.318.435 **3527**

Saturable Reactors with Inductive D.C. Load: Part 1—Steady-State Operation.—H. F. Storm. (*Trans. Amer. Inst. elect. Engrs*, 1952, Vol. 71, Part I, pp. 335–343.)

621.319.45 : 539.23 : 537.311.33 **3528**

Distribution of Conductivity within Dielectric Films on Aluminium.—J. E. Lilienfeld & C. Miller. (*J. electrochem. Soc.*, May 1953, Vol. 100, No. 5, pp. 222–226.) Frequency and formation-voltage characteristics were determined for a capacitor with an anodized Al anode, Al cathodes and electrolyte consisting of boric acid and borates, in the range 20 c/s–10 kc/s. An equivalent circuit is given and a model is suggested consisting of a 2-layer dielectric film, one layer, adjacent to the metal surface, having low conductivity, low power loss, independent of frequency, but dependent on the peak formation voltage, the other film with opposite properties.

621.385.3 : 621.392.5 **3529**

Triode Transformation Groups.—A. W. Keen. (*Wireless Engr*, Oct. 1953, Vol. 30, No. 10, pp. 238–243.) The properties of the six networks including respectively the six possible transmission paths through a triode valve (1321 of 1951) are shown to be interrelated in such a way that, given the characteristics of any one of the circuits, those of the remainder are obtainable by a routine transformation process.

621.392 **3530**

Potential Analog Network Synthesis for Arbitrary Loss Functions.—E. S. Kuh. (*J. appl. Phys.*, July 1953, Vol. 24, No. 7, pp. 897–902.) A method of network design is described in which the appropriate potential problem is formulated on the basis of a given loss function by introducing continuous charge distribution on the complex-frequency plane. The technique of charge quantization is used to find the natural modes of the network function.

621.392.012 3531

Feedback Theory — Some Properties of Signal Flow Graphs.—S. J. Mason. (*Proc. Inst. Radio Engrs*, Sept. 1953, Vol. 41, No. 9, pp. 1144–1156.) The equations characterizing transmission systems may be represented by networks of directed branches, termed 'signal-flow graphs'; the branches terminate at nodes, which represent repeater stations. The topology of these diagrams is studied with particular reference to feedback circuits. The technique is illustrated by application to specific circuit design problems.

621.392.4 3532

Two Theorems on Two-Terminal Electrical Networks.—F. Reza. (*C. R. Acad. Sci., Paris*, 10th Aug. 1953, Vol. 237, No. 6, pp. 429–430.) Given that the impedance of the network is a rational function $Z(S)$ of the complex frequency S, and putting $Z(S) = P(S)/Q(S)$, it is shown that

$$Z_n(S) = \frac{d^n P(S)\ dS^n}{d^n Q(S)\ dS^n}$$

for networks consisting of (*a*) reactances only, and (*b*) resistances together with reactances of one kind.

621.392.43 3533

Modified Exponential Line as Improved Transforming Element with High-Pass Characteristic.—H. H. Meinke. (*Arch. elekt. Übertragung*, July 1953, Vol. 7, No. 7, pp. 347–354.) For a given length of exponential line there is a critical frequency above which the transformation property becomes effective; at frequencies only a little above critical the transformation is only approximate. This drawback can be avoided by designing the line so that the characteristic impedance does not vary exactly exponentially. Approximate line equations are derived for this case and practical forms of such lines are discussed; the influence of constructional defects is examined.

621.392.5 3534

A General Definition of Pass Band.—S. Colombo. (*C. R. Acad. Sci., Paris*, 10th Aug. 1953, Vol. 237, No. 6, pp. 427–429.)

621.392.5 3535

***RLC* Lattice Networks.**—L. Weinberg. (*Proc. Inst. Radio Engrs*, Sept. 1953, Vol. 41, No. 9, pp. 1139–1144.) A method for synthesizing open-circuited lattice networks to have a required transfer impedance is based on the partial-fraction expansion of the given function. No mutual inductances are used, and series resistance is associated with all the self-inductances, so that low-Q coils may be used. A numerical example is worked out.

621.392.5 3536

Synthesis of Paralleled 3-Terminal *RC* Networks to provide Complex Zeros in the Transfer Function.—P. F. Ordung, G. S. Axelby, H. L. Krauss & W. P. Yetter. (*Trans. Amer. Inst. elect. Engrs*, 1951, Vol. 70, Part II, pp. 1861–1867.) An improvement on the method given by Guillemin (2462 of 1949) is described which requires fewer paralleled networks. The desired transfer ratio is converted into the sum of several transfer ratios, each of which can be realized with a single ladder network. A method of synthesis for these networks in terms of a common driving-point-admittance function is described and theory developed for their connection in parallel. See also 656 of March (Ordung et al.).

621.392.5 3537

The Dual-Input Parallel-T Network.—C. F. White & K. A. Morgan. (*Proc. nat. Electronics Conf., Chicago*, 1952, Vol. 8, pp. 588–597.) The frequency of maximum attenuation in this type of circuit is a function of the ratio of the two input voltages. Various potentiometer-drive arrangements can be used for external tuning over a wide frequency range. Experimental results for a circuit with dual-potentiometer input are in good agreement with theory.

621.392.5 3538

Synthesis of the Transfer Function of 2-Terminal-Pair Networks.—R. Kahal. (*Trans. Amer. Inst. elect. Engrs*, 1952, Vol. 71, Part I, pp. 129–134.) Full paper. See 1263 of May.

621.392.5 : 621.3.015.3 3539

The Possibility of defining the Width of the Pass Band in terms of the Distortion of a Transient.—S. Colombo. (*C. R. Acad. Sci., Paris*, 17th Aug. 1953, Vol. 237, No. 7, pp. 455–457.) Analysis using the definition previously given (3534 above) indicates that it is impossible to define the distortion of a transient by a single build-up time.

621.392.5 : 621.318.435 3540

An Approximate Graphical Analysis of the Steady-State Response of Nonlinear Networks.—S. Duinker. (*Philips Res. Rep.*, April 1953, Vol. 8, No. 2, pp. 133–147.) An approximate graphical analysis is given for the case when one of the circuit parameters is varied gradually. Transients and the generation of subharmonics are not dealt with. The networks investigated consist of a nonlinear iron-cored inductor with a.c. and d.c. magnetization, in series with a linear resistor and linear capacitor. The results of the analysis can be applied to the investigation of the effects of resistive or reactive loads of magnetic amplifiers, and the utilization of jump phenomena (ferroresonance effects) in switching devices.

621.392.5.012.11 : 621.392.43 3541

A Theorem on the Impedance-Transforming Properties of Reactive Networks.—L. Storch. (*J. appl. Phys.*, July 1953, Vol. 24, No. 7, pp. 833–838.) Two circle transformations, derived from the impedance-transformation properties of a 4-terminal network, are used to develop a 'circle-locus' theorem which considerably simplifies analysis of impedance-matching networks. It is applicable to any linear reactive network with any arbitrary reference impedance not lying on the reactance axis.

621.392.5.015.3 3542

An Approximate Method of Obtaining the Transient Response from the Frequency Response.—J. R. Wait. (*Canad. J. Technol.*, June 1953, Vol. 31, No. 6, pp. 127–131.) The method is based on expressing the transient response $A(t)$ in terms of the real part of the frequency-response function, which is plotted on a logarithmic frequency scale and approximated by a series of straight-line segments. $A(t)$ is calculated from the slopes of these segments.

621.392.52 3543

Concerning the Minimum Number of Resonators and the Minimum Unloaded Q needed in a Filter.—M. Dishal. (*Trans. Inst. Radio Engrs*, June 1953, No. PGVC-3, pp. 85–117.) The number of resonator elements required to meet given specifications of filter characteristics, and the lowest no-load values of Q which the resonators may have, can be deduced from the formulae and graphs given. The specifications include the 'accept' and 'reject' bandwidths, the peak/valley ripple ratio and response in the 'reject' band of the selective circuit. Both constant-k and m-derived types of filter are considered, and numerical examples are worked out.

621.392.52.012.3 **3544**

Wave-Filter Characteristics by a Direct Method.—R. C. Taylor & C. U. Watts. (*Trans. Amer. Inst. elect. Engrs*, 1952, Vol. 71, Part I, pp. 96–100.) Full paper. See 961 of April.

621.392.52.029.3 : 621.396.645.371 **3545**

Audio-Frequency Filters using Negative Feedback.—T. Janisz. (*Electronics*, Sept. 1953, Vol. 26, No. 9, pp. 222 . . 228.) The operation is described of a two-valve amplifier with a series-resonance circuit across the cathode impedance of each of the two valves, and a parallel-resonance circuit in the anode circuit of the first valve. A band-pass filter with a flat-topped response curve is obtained by adjusting the resonance frequency of the cathode circuit of the second valve to be intermediate to the cathode-circuit and anode-circuit resonance frequencies of the first valve.

621.392.6 **3546**

Nonlinear Multipoles.—L. A. Zadeh. (*Proc. nat. Acad. Sci., Wash.*, April 1953, Vol. 39, No. 4, pp. 274–280.) A conceptual scheme is outlined which is intended to be used in conjunction with theories of oriented graphs and with machine computers for the analysis and design of nonlinear networks. A system of classes of nonlinear multipoles is defined, the classes being characterized by their responses to a particular set of pulses. See also 2598 of September.

621.396.6 : 061.4(443.611) **3547**

The 17th National Radio-Components Exhibition.—(*Onde élect.*, June 1953, Vol. 33, No. 315, pp. 287–293.) Review of a selection of exhibits in the sections allotted to valves, miscellaneous components and accessories, and measurement instruments at the exhibition held in Paris, 27th February–3rd March 1953.

621.396.6.002.2 **3548**

Printed Circuits and the Automatic Factory.—R. A. Gerhold. (*Proc. nat. Electronics Conf., Chicago*, 1952, Vol. 8, pp. 481–488.) Discussion of the application of the auto-sembly technique [355 of February (Danko)] to quantity production of a wide variety of electronic equipment, with substantial reduction of costs and economy in skilled personnel.

621.396.611.1 **3549**

Frequency Feedback.—H. E. Hollmann. (*Proc. nat. Electronics Conf., Chicago*, 1952, Vol. 8, pp. 577–587.) Discussion of a phenomenon which occurs in all resonant systems possessing a relation between resonance frequency and amplitude. In such systems there may be an external feedback channel, or internal feedback due to nonlinear reactors or to nonlinear resistors combined with fixed reactors, as in the so-called *RX* modulators. The phenomenon shows itself in asymmetrical deformation of the resonance curve. Applications of the effect in a.m.-f.m. conversion, and in f.m.-a.m. conversion for frequency-shift reception, are noted.

621.396.611.1 : 621.396.619.13 **3550**

A Method of Evaluation of the Quasi-Stationary Distortion of F.M. Signals in Tuned Interstages.—J. J. Hupert. (*Proc. nat. Electronics Conf., Chicago*, 1952, Vol. 8, pp. 445–461.) A graphical approach to the problem of designing low-distortion f.m. circuits, making use of the normalized derivative of the phase/frequency function, is shown to be preferable to that using the phase/frequency curve itself. The principles are applied to the determination of the distortion factor of a single tuned-circuit interstage and of a cascade arrangement of three stagger-tuned resonant circuits.

621.396.611.1.015.3 **3551**

Transient Phenomena in an Oscillatory Circuit with Variable Inductance.—P. G. Gorodetski. (*Zh. tekh. Fiz.*, Oct. 1952, Vol. 22, No. 10, pp. 1687–1692.) The operation of the circuit is discussed and a formula (2a) is derived for determining the variation of the current in the circuit. A numerical example is given.

621.396.611.21 **3552**

High-Frequency Crystal Units for use in Selective Networks, and their Proposed Application in Filters suitable for Mobile-Radio Channel Selection.—D. F. Ciccolella & L. J. Labrie. (*Trans. Inst. Radio Engrs*, June 1953, No. PGVC-3, pp. 118–128.) The design of bevelled AT-type quartz plates for use in band-pass crystal units is described. The reduction of unwanted modes of vibration, including thickness-shear overtone modes, is considered. Approximate electrical data for plated circular AT-type quartz plates with all unwanted modes reduced to a level 30 db below the response of the main thickness-shear mode are tabulated. The design is considered of a crystal filter with a mid-band frequency of 1·552 Mc/s and bandwidths of 30 kc/s and 40 kc/s respectively at the 6-db and 100-db points.

621.396.611.21 **3553**

Thickness Vibrations of Piezoelectric Crystal Plates.—R. Bechmann. (*Arch. elekt. Übertragung*, July 1953, Vol. 7, No. 7, pp. 354–357.) Addendum to work noted in 357 of February.

621.396.611.31 : 621.318.42 **3554**

A Simple Method of Coupling Toroidal Coils.—R. R. Darden, Jr. (*Proc. nat. Electronics Conf., Chicago*, 1952, Vol. 8, pp. 618–620.) Theory is given of a method of coupling suitable for the low values of coupling coefficient usually required for i.f. transformers, f.m. discriminators and in some types of filter. The coupling coils, with very few turns, are wound on a separate toroidal core, each being connected in series with one of the toroidal coils to be coupled. In a simplified version a coupling coil of one or two turns, connected in series with one toroidal coil, is wound on top of the other toroidal coil. Such a transformer may be accommodated in a ¾-in. cube with the two toroids close together with no shield between them.

621.396.611.4 : 621.315.212 **3555**

Electromagnetic Waves in Hollow Metal Cylinders with Circular Cross-Section.—R. Müller. (*Arch. elekt. Übertragung*, July 1953, Vol. 7, No. 7, pp. 341–346.) The object of the paper is to coordinate the treatment of the subject and to clarify some points. The electric and magnetic lines of force are mapped in perspective. Other diagrams show the distribution of Poynting vector over the cross-section and the currents in the walls. Wave propagation in concentric lines is investigated.

621.396.615 : 517.93 **3556**

Nonlinear Electromechanical Systems.—G. Cahen. (*Rev. gén. Élect.*, June 1953, Vol. 62, No. 6, pp. 277–293.) By an extension of Liénard's graphical method (1928 Abstracts, p. 469) only two or three coplanar curves need be traced for a complete topological study of the stability conditions for systems represented by equations of the types

$$\ddot{x} + \dot{x}^2 a(x) + \dot{x}b(x) + c(x) = 0$$
$$\ddot{x} + f(\dot{x}) + \dot{x}b(x) + r(x) = 0.$$

Results of a detailed analysis are applied to the case of a filtered oscillator (1279 of May and 1624 of June). A much shorter version of the first part of this paper was given in *C. R. Acad. Sci., Paris*, 3rd Nov. 1952, Vol. 235, No. 18, pp. 1003–1005.

621.396.615.17 **3557**

A Study on the Triggering of a Plate-Coupled Multivibrator by Negative Pulses.—S. K. Sen & B. K. Bhattacharyya. (*Indian J. Phys.*, Dec. 1952, Vol. 26, No. 12, pp. 597–616.) Experimental observations showed (*a*) that voltage waveforms at different electrodes were markedly influenced by the form of the input pulses, (*b*) that triggering action depended largely on the input-pulse amplitude, (*c*) that triggering might not take place for either very small or very large pulse amplitudes, depending on circuit arrangement, but larger amplitudes caused greater difficulties. Analysis of the circuit response to pulses of various durations and amplitudes gave results supporting the observations.

621.396.615.17 **3558**

Design of Triode Flip-Flops for Long-Term Stability.—J. O. Paivinen & I. L. Auerbach. (*Trans. Inst. Radio Engrs*, June 1953, Vol. EC-2, No. 2, pp. 14–26.) Description of an analytical method of design based on considerations of d.c. stability, limiting tolerances in respect of voltage and component values being taken into account initially. An Eccles-Jordan circuit with injection diodes is considered as a general case and equations are derived expressing the operating conditions in the grid and the anode circuits. Solution of these equations gives appropriate key values. The method is applied to three special cases and a numerical example is given.

621.396.615.17 : 621.396.619.13 : 621.317.083.7 **3559**

Wide-Band Data Transmitter.—D. J. Gray, V. P. Gurske & W. E. Morrow. (*Electronics*, Sept. 1953, Vol. 26, No. 9, pp. 168–170.) A f.m. oscillator uses a double-phantastron circuit [2478 of 1948 (Close & Lebenbaum)] for a linear data-recording system. Signals in the range 0–5 kc/s can be handled with an accuracy to within 1%. Full circuit details are given.

621.396.615.18 **3560**

Stable Frequency Dividers using Thyrite Elements.—W. L. Hughes. (*Proc. nat. Electronics Conf., Chicago*, 1952, Vol. 8, pp. 730–733.) Locked-oscillator frequency dividers using nonlinear thyrite elements as load impedors are described. With slight circuit modifications, a stable locking action is obtainable for division by any number between 1 and 10. No circuit adjustments are required as the valves age. Dividers have been constructed for frequencies up to 1 Mc/s.

621.396.616 : 534.112.001.8 **3561**

Vibrating-Wire High-*Q* Resonator.—A. W. Dickson & W. P. Murden. (*Electronics*, Sept. 1953, Vol. 26, No. 9, pp. 164–167.) Resonator units are described in which a wire of high tensile strength is stretched between long pole-pieces maintaining a strong transverse magnetic field. Resonance occurs when the frequency of current sent through the wire coincides with the natural frequency of the system. Theory is given and the equivalent electrical circuit analysed. With a tungsten wire mounted in vacuo, a Q value of 1 840 has been obtained at 2·21 kc/s. Applications of such devices in filters and for frequency control are noted.

621.396.619.11/.13 **3562**

Spurious Frequency Modulation in Signal Generators.—T. P. Flanagan. (*Marconi Instrumentation*, June 1953, Vol. 4, No. 2, pp. 24–28.) The chief cause of f.m. in an a.m. circuit is the variation of valve input capacitance with modulation. Methods of reducing this effect are examined. Two circuits effective in reducing f.m. at v.h.f. are described: (*a*) an untuned class-A amplifier with grid modulation; (*b*) a voltage-dividing network containing a crystal diode suitably biased to act as resistance.

621.396.645 **3563**

Valve Matching using Resistors.—H. V. Harley. (*Wireless World*, Oct. 1953, Vol. 59, No. 10, pp. 488–493.) Desired characteristics can be obtained by including resistors in series and/or parallel with a valve; formulae for the effective parameters are tabulated for several simple combinations. Matching of a.c. characteristics may in practice be accompanied by reasonable matching of the overall d.c. characteristics; a circuit is shown using the latter effect for determining resistances required for matching two valves.

621.396.645 : 621.385.3.029.6 **3564**

A General Circle Diagram for the Input Admittance of a Grounded-Grid Disk-Seal Triode.—E. Willwacher. (*Fernmeldetech. Z.*, July 1953, Vol. 6, No. 7, pp. 328–331.) The method given takes account of the transconductance and anode-cathode capacitance of the triode. The effect of the transconductance phase, which depends on the electron transit time, is shown in the diagram. A numerical example is given. See also 3245 of November.

621.396.645.012 **3565**

Cathode-Follower Design Charts.—N. O. Sokal. (*Electronics*, Sept. 1953, Vol. 26, No. 9, pp. 192–194, 196.) Output-impedance/input-voltage charts are shown for 9 commonly used valve types. Design procedure is outlined.

621.396.645.015.75 **3566**

On the Faithful Reproduction of the Flat Top of a Pulse in a High-Fidelity Pulse Amplifier.—B. K. Bhattacharyya. (*Indian J. Phys.*, Jan. 1953, Vol. 27, No. 1, pp. 39–54.) The pulse amplifier analysed comprises a pentode with *RC* coupling circuits. Two methods of determining anode current, and hence amplifier response, are applied, the one based on the overall-amplification reduction due to voltages across the parallel *RC* screen circuit and the *RC* cathode circuit, the other based on the superposition principle. Correction of pulse-top distortion is achieved by the anode-supply *RC* decoupling circuit; the conditions for satisfactory correction are derived.

621.396.645.029.3 **3567**

A Modern Broadcasting Pre-amplifier.—Abadie & Blondé. (*Onde élect.*, July 1953, Vol. 33, No. 316, pp. 468–472.) Description of a simple 2-stage feedback amplifier using Type-EF40 low-noise pentodes and suitable for use with sound-reproduction equipment. Distortion is low, gain (40 db) practically uniform from 30 c/s to 15 kc/s, and pickup of stray fields effectively negligible.

621.396.645.371.081.75 **3568**

Harmonic Distortion and Negative Feedback.—R. O. Rowlands. (*Wireless Engr*, Oct. 1953, Vol. 30, No. 10, pp. 261–262.) Author's reply to comment by Kerr (3253 of November).

621.396.662.2 : 538.221 **3569**

Use of Ferromagnetic Materials in Electronic Tuning of Radiofrequency Components.—S. Stiber. (*Proc. nat. Electronics Conf., Chicago*, 1952, Vol. 8, pp. 462–468.) The change of the incremental permeability of a core of ferrite material due to a secondary superimposed magnetic field is discussed. The effect is utilized in the design of toroidal inductors whose inductance can be varied by altering the current in an auxiliary winding. Inductors of this type are incorporated in a receiver tunable from 500 kc/s to 3 Mc/s, with a bandwidth of 11 kc/s at 500 kc/s and of 22 kc/s at 3 Mc/s. The power required in the auxiliary tuning coil is < 2 W. A variable inductor has been developed whose volume is < 1 in.^{3}, weight slightly > 1 oz, and tuning range, for 2 W applied power, of

7 : 1 in the frequency band 500 kc/s–3 Mc/s, falling to 2 : 1 in the band 25–50 Mc/s. See also *Electronics*, July 1953, Vol. 26, No. 7, pp. 186–188.

GENERAL PHYSICS

530.145 : 531.51 : 538.1 **3570**

The Relation of the Quantum Theory to the Theories of Gravitation and Electromagnetism, and an Application to the Theory of the Electron.—H. T. Flint & E. M. Williamson. (*Z. Phys.*, 25th June 1953, Vol. 135, No. 3, pp. 260–269.)

534.01 **3571**

On an Iterative Method for Nonlinear Vibrations.—R. E. Roberson. (*J. appl. Mech.*, June 1953, Vol. 20, No. 2, pp. 237–240.) Analysis of a method giving results analogous to those obtained by Duffing's method.

535.22 + 621.317.029.6] : 061.3 **3572**

High-Frequency Electrical Measurements. Conference in Washington. D.C.—Essen. (See 3651.)

537.212 **3573**

A New Property of 2-Dimensional Fields.—A. D. Moore. (*Trans. Amer. Inst. elect. Engrs*, 1952, Vol. 71, Part I, pp. 343–346. Discussion, pp. 346–347.) Mapping of the field inside a nearly enclosed area is facilitated by working on the correlative field, whose derivation is explained. An analytical proof of the correlative property, based on the Schwartz-Christoffel transformation, is given in an appendix. The proof is due to W. R. Smythe. For a digest see *Elect. Engng, N.Y.*, April 1953, Vol. 72, No. 4, p. 290.

537.213 + 538.123] : 517.947.42 **3574**

The Derivation of Vector Potential from Tables for Scalar Potential.—J. J. Smith. (*Trans. Amer. Inst. elect. Engrs*, 1952, Vol. 71, Part I, pp. 169–174.) Full paper. See 2285 of August.

537.228 : 535.37 **3575**

Field Strength and Temperature Studies of Electroluminescent Powders in Dielectric Media.—S. Roberts. (*J. opt. Soc. Amer.*, Nov. 1952, Vol. 42, No. 11, pp. 850–854.) Measurements were made on electroluminescent cells using ZnSSe: Cu phosphor dispersed in a thermoplastic matrix. The intensity of luminescence depends critically on the field in the phosphor, and hence on its dielectric constant, but does not depend on the dielectric properties of the matrix except in so far as the latter provides a means of supporting a strong field in the phosphor. Little change of luminescent intensity was observed when the temperature was varied from $-100°$ to $+50°$C, polystyrene being used as matrix.

537.228 : 535.37 **3576**

Dielectric Changes of Electroluminescent Phosphor during Illumination.—S. Roberts. (*J. opt. Soc. Amer.*, July 1953, Vol. 43, No. 7, pp. 590–592.) Measurements were made of the dielectric properties of electroluminescent films composed as described previously (3575 above), for different values of applied voltage at 1 000 c/s and for various wavelengths and intensities of illumination. The dielectric constant depends markedly on the illumination, the magnitude of the variations depending critically on the applied voltage. The effect exhibits a maximum at about 4046 Å.

537.311.31 : 539.23 **3577**

Development of the Electrical Resistance of Thin Films of Platinum subjected to a Relatively High Direct Voltage.—M. Erny. (*C. R. Acad. Sci., Paris*, 3rd Aug. 1953, Vol. 237, No. 5, pp. 387–389.) Measurements made on films of different thicknesses, before and after the application of voltages of 300, 400 or 500 V, indicate that the resistance is increased, decreased or unchanged depending on the amount of heat developed, which in turn depends on the initial resistance and hence on the thickness.

537.311.32 **3578**

On Pre-exponential Factors in Formulae for Ionic Conductivity in Solids.—Y. Haven & J. H. van Santen. (*Philips Res. Rep.*, Dec. 1952, Vol. 7, No. 6, pp. 474–477.) If, in formulae for equilibrium constants and rate constants, energy is assumed to be proportional to temperature, the factor preceding the exponential term is not increased by a factor $\exp(\alpha/k)$, α being a constant and k Boltzmann's constant, but by a factor involving entropy.

537.52 : 621.3.066.6 **3579**

Initiation of Discharges at Electrical Contacts.—F. L. Jones. (*Proc. Instn elect. Engrs*, Part I, July 1953, Vol. 100, No. 124, pp. 169–173.) The experimental investigation described deals with discharges between low-power contacts, with special attention to the cases of make and break at medium voltage. The discharges were found to be initiated by a cold-field emission of electrons governed by the nature of the surface tarnish layers. The influence of the gas atmosphere was examined. A theory of the mechanism is outlined.

537.533 : 538.691 **3580**

On the Nonoptical Theory of Focusing in Rotating Magnetic Fields.—P. I. Tsukkerman. (*Zh. tekh. Fiz.*, Nov. 1952, Vol. 22, No. 11, pp. 1843–1847.) The difficulties arising in the application of optical methods to electron optics are pointed out. The general theory of the focusing action of static magnetic fields developed by Grinberg, which is based on nonoptical methods, is applied to the case of paraxial electron beams in rotating magnetic fields.

537.533.8 **3581**

Measurement of Secondary Electron Emission.—H. Gobrecht & F. Speer. (*Z. Phys.*, 25th June 1953, Vol. 135, No. 3, pp. 331–348.) The sources of error in secondary-electron emission determinations are discussed in detail. Retarding-voltage curves, obtained experimentally, are analysed and a method of calculating the proportion of reflected primary electrons in the secondary electron current is given for the case when the target and spherical collector are made of the same material. For a Cr-Ni alloy and a primary potential of 1 kV, this proportion is 22%.

537.533.9 : [538.56.029.65 + 535.212 **3582**

Experiments on Radiation by Fast Electron Beams.—H. Motz, W. Thon & R. N. Whitehurst. (*J. appl. Phys.*, July 1953, Vol. 24, No. 7, pp. 826–833.) The design of a magnet system for an undulator, in which the field direction alternates periodically along the electron path [2411 of 1951 (Motz)], is described. A 100-MeV electron beam from a linear accelerator was passed through the system; light radiated by the beam was observed and its plane of polarization was determined. Using a small linear accelerator with good bunching action and beam energy 3 MeV, radiation at wavelengths below 1·9 mm was observed; peak power output was > 1 W. Millimetre waves were also generated in the accelerator tube.

537.582 : 518.4 **3583**

Graphs for a Rapid Calculation of the Work Function of Thermionic Emission.—C. G. J. Jansen & R. Loosjes. (*Philips Res. Rep.*, April 1953, Vol. 8, No. 2, pp. 81–90.) A set of graphs is given for the saturation current density (10^{-10}–10^{2} A/cm^2) as a function of absolute temperature (300–2 400°K), with the work function (0·6–8 V) as parameter. The use of the graphs is explained.

537.71.081.5 **3584**

Dimensional Analysis, Units and Rationalization.—R. Vermeulen. (*Philips Res. Rep.*, Dec. 1952, Vol. 7, No. 6, pp. 432–441.) Careless manipulation of dimensional formulae can lead to fallacious results. The trouble is due to the use of the multiplication sign for two different kinds of products having very different physical meanings. Once these are clearly distinguished, and the physical meaning considered at every stage in the proceedings, the difficulties encountered, even with rationalized units, are overcome.

538.11 **3585**

Antiferromagnetism.—H. Labhart. (*Z. angew. Math. Phys.*, 15th Jan. 1953, Vol. 4, No. 1, pp. 1–24.) A survey paper. The distinction between antiferromagnetism and ferrimagnetism is indicated; the former is found mainly in simple inorganic compounds, the latter in complex compounds. Problems discussed concern the interaction between adjacent dipoles, the different possible arrangements within the various lattices, the magnetic properties, and the phenomena of the transition from order to disorder. 88 references.

538.114 **3586**

A Theory of Ferromagnetism.—L. Pauling. (*Proc. nat. Acad. Sci., Wash.*, June 1953, Vol. 39, No. 6, pp. 551–560.) A theory is formulated that, when applied to iron, making use only of spectroscopic data for the Fe atom, gives a value of 2·20 Bohr magnetons for the saturation magnetic moment per iron atom and a value of 1 350°K for the Curie temperature. These values are in reasonable agreement with the experimental values of 2·22 magnetons and 1 043°K.

538.24 **3587**

The Calculation of the Magnetizing Force.—A. A. Halacsy. (*Trans. Amer. Inst. elect. Engrs*, 1952, Vol. 71, Part I, pp. 90–95.) Full details of the method described in 1662 of 1950.

538.56 : 535.421 **3588**

The Electromagnetic Properties of a Plane Grating.—V. M. Lopukhin & V. S. Nikol'ski. (*Zh. tekh. Fiz.*, Oct. 1952, Vol. 22, No. 10, pp. 1599–1605.) A plane grating is considered which consists of a number of parallel conducting planes between which electron streams pass, and its waveguide properties are discussed, taking into account the velocity scatter of the electrons. A general solution of the problem is given for an arbitrary type of scatter function. The velocity scatter narrows the range of average velocities of the streams for which amplification is possible and displaces it towards the higher velocity values; the maximum amplification is also reduced.

538.566 : 535.135 **3589**

Propagation of Electromagnetic Waves in a Transparent Periodically Stratified Space.—C. Dufour & A. Herpin. (*Rev. d'Optique*, June 1953, Vol. 32, No. 6, pp. 321–348.) Equations derived for the general case are applied to the special case of a pile of double layers of equal optical thickness.

538.566 : 535.42 **3590**

Transmission of Electromagnetic Waves through Pairs of Parallel-Wire Grids.—W. E. Groves. (*J. appl. Phys.*, July 1953, Vol. 24, No. 7, pp. 845–854.) An analytical solution is obtained for the power transmission coefficient of a double-grid system. Transmission coefficients are plotted as functions of the intergrid spacing for different wire spacing and angles of tilt. The curves are compared with those obtained by measurement at λ 3·2 cm using wire of radius 0·0875 cm. Agreement is good until the wire spacing approaches λ. Reasons for this divergence are noted.

538.566 : 537.562 **3591**

Plasma Oscillations in Crossed Electric and Magnetic Fields.—A. I. Akhiezer & R. V. Polovin. (*Zh. tekh. Fiz.*, Nov. 1952, Vol. 22, No. 11, pp. 1794–1802.) An idealized case is considered and the specific structure which retards the e.m. waves is replaced by a medium with an effective dielectric constant greater than unity. It is also assumed that the constant charge and current of the electron beam are compensated by the charge and current of ions which do not participate in the h.f. oscillations. With these assumptions, conditions are determined under which the electron beam becomes unstable, the fluctuations of the velocity and density of the beam increasing indefinitely.

538.566.2 **3592**

Extension of Fermat's Principle to Group Propagation Time.—P. Poincelot. (*C. R. Acad. Sci., Paris*, 3rd Aug. 1953, Vol. 237, No. 5, pp. 382–384.) Continuation of work noted in 3420 of 1952. It is shown by analysis that the difference between the group propagation times for two real paths joining the same two points is zero.

538.569.4 **3593**

Amplification of Microwave Radiation by Substances not in Thermal Equilibrium.—J. Weber. (*Trans. Inst. Radio Engrs*, June 1953, No. PGED-3, pp. 1–4.) Methods are discussed for the production of a nonequilibrium energy distribution, and the amount of amplification which may possibly be obtained by this method is calculated.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.72 : 621.396.822 : 621.396.677.1 **3594**

New Techniques in Radio Astronomy.—Kraus & Ksiazek. (See 3490.)

523.746 **3595**

Extrapolation of Sunspot/Climate Relationships.—S. W. Visser. (*J. Met.*, June 1953, Vol. 10, No. 3, pp. 232–233.) Consideration of the periodicity of sunspot-number variations is based on taking the maximum year as the middle one of three successive maximal years. The next maximum should occur in 1959 and should be a strong one.

538.56 : 550.37 **3596**

Nonradiative Natural Oscillations of a Conducting Sphere surrounded by a Layer of Air and an Ionospheric Sheath.—W. O. Schumann. (*Z. Naturf.*, Feb. 1952, Vol. 7a, No. 2, pp. 149–154.) A theoretical investigation having special reference to e.m. oscillations of the earth, and relevant to the study of atmospherics. The dependence of the oscillation frequencies on the air-layer thickness and on the plasma properties is analysed. The lowest natural frequency to be expected for earth oscillations is about 11 c/s. Oscillations of low frequencies are reflected at the inner surface of the plasma, while higher-order oscillations may penetrate right through the plasma and radiate into space.

538.56 : 550.37 **3597**

Damping of the Natural Electromagnetic Oscillations of the System Earth-Air-Ionosphere.—W. O. Schumann. (*Z. Naturf.*, March/April 1952, Vol. 7a, Nos. 3/4, pp. 250–252.) Continuation of work noted in 3596 above. An approximate formula is derived for the damping of the lowest-frequency oscillations, from which the effective height of the air layer and the effective conductivity of the ionosphere can be determined.

551.510.535 **3598**

Electron Density in the Upper Atmosphere and Interpretation of the *h' f* Curves of Ionosphere Virtual Height.

—F. Mariani. (*Ann. Geofis.*, Jan. 1953, Vol. 6, No. 1, pp. 21–45.) For the case of vertical incidence and the geomagnetic latitude of Rome, an expression is developed for the group refractive index using Appleton's formula. Absorption is neglected. The optical path of e.m. waves totally or partially reflected or transmitted through an ionosphere layer is determined as a function of frequency for various cases. The results are applied in conjunction with experimental *h'f* curves to derive linear relations giving directly the thickness of the layer and its height above the ground. Cases of single and superposed layers are treated.

551.510.535 : 621.3.087.4 **3599**

Instrumentation for Measuring Changes in Phase of Ionospheric Echoes.—R. E. Jones. (*Rev. sci. Instrum.*, June 1953, Vol. 24, No. 6, pp. 433–436.) Changes in carrier phase of ionosphere pulse echoes are determined from a comparison of the transmitted signal, stored by a 'memory' circuit, with the echo signal received. The phase variation is presented, in the instrument described, on a c.r.o. and a continuous photographic record is obtained. The equipment used is similar in principle to that of Findlay (397 of 1952). Results of E-region observations at 150 kc/s have been published by Davids (516 of February).

551.510.535 : 621.3.087.5 : 621.396.11 **3600**

COZI Communication-Zone Indicator.—L. C. Edwards. (*Electronics*, Aug. 1953, Vol. 26, No. 8, pp. 152–155.) A general description, with block diagram of timing and indicating circuits, is given of a low-power oblique-incidence ionosphere sounder designed to indicate skip distances and communication zones from 500 to 2 000 miles. Operation is on any one of six pre-set frequencies in the range 5–32 Mc/s, peak pulse power 600–900 W, pulse duration 0·5–2·5 ms and repetition frequency 20/sec. The one-hop skip distance is obtained directly from the time lag of the backscatter from the ground after single reflection from the ionosphere. Typical oscillograms show the increase of skip distance as the carrier frequency is increased in steps from 7 to 30 Mc/s, the complete series of 12 records, two at each frequency, being obtained in 8 min.

LOCATION AND AIDS TO NAVIGATION

621.39.001.11 : [621.396.9 + 621.397.5 **3601**

The London Conference, September 1952: Part 2—Radar and Television.—J. Loeb. (*Onde élect.*, July 1953, Vol. 33, No. 316, pp. 478–481.) A review of some of the papers presented at the conference on 'Applications of the theory of information', dealing with the detection of signals in the presence of noise, the use of autocorrelation (theory and practice), the compression of television signals, and facsimile. Part 1 : 3702 below.

621.396.9 : 519.21 **3602**

On the Statistical Theory of Detection of a Randomly Modulated Carrier.—W. M. Stone. (*J. appl. Phys.*, July 1953, Vol. 24, No. 7, pp. 935–939.) "A method is outlined for estimating the probability of detection for a pulsed radar, assuming a randomly modulated carrier. For a square-law detector, closed forms for the moments of the distribution of the envelope are presented in terms of three different choices of the distribution of carrier amplitude, thus leading to an Edgeworth series representation of the desired probability. At least one choice of distribution of the carrier amplitude leads to closed forms for the moments for the case of a linear detector. Curves of probability of detection vs signal-to-noise power ratio are constructed and cross checked by a method of numerical integration."

MATERIALS AND SUBSIDIARY TECHNIQUES

531.788.7 **3603**

New Developments in the Production and Measurement of Ultra-high Vacuum.—D. Alpert. (*J. appl. Phys.*, July 1953, Vol. 24, No. 7, pp. 860–876.) Limitations of conventional ionization gauges are discussed, in particular X-ray effects, differences of sensitivity for different gases, and the pumping action of the gauge itself. Techniques and apparatus are described by which pressures down to 10^{-10} mm Hg are achieved without chemical getters, special traps or refrigerants; these include a Bayard-Alpert gauge (2785 of 1950) with special electrode arrangement for minimizing X-ray effects, two designs of a manipulative vacuum valve for very low pressures, an absolute manometer, methods of leak detection, and a power supply system for intermittent operation of the gauge.

533.5 : 537.525.5 **3604**

Clean-Up of Helium Gas in an Arc Discharge.—M. J. Reddan & G. F. Rouse. (*Trans. Amer. Inst. elect. Engrs*, 1951, Vol. 70, Part II, pp. 1924–1929.)

535.215.1 **3605**

On the General Criterion [for a choice] **between the 'Barrier Layer' and 'Concentration' Theories of Photoconductivity.**—S. M. Ryvkin. (*Zh. tekh. Fiz.*, Oct. 1952, Vol. 22, No. 10, pp. 1693–1695.) The criterion proposed by Gibson (3157 of 1951) is stated to be insufficient; the choice between the 'barrier layer' and 'concentration' mechanisms should be decided by special experiments for each particular material.

535.215.1 : 537.533.7 **3606**

An Electron-optical Method of investigating Electromagnetic Fields and its Application to Investigation of the Internal Photoelectric Effect.—V. S. Vavilov. (*Zh. tekh. Fiz.*, Oct. 1952, Vol. 22, No. 10, pp. 1644–1657.) A method similar to that used in optics by Foucault and Toepler is proposed. With the aid of the apparatus described a quantitative investigation of the drift of photocurrent carriers in an electric field in a semiconductor is possible. Observations are made of the redistribution of the potential along the samples when photoconduction is produced in them.

535.37 **3607**

Intrinsic Efficiencies of Phosphors under Cathode-Ray Excitation.—A. Bril & H. A. Klasens. (*Philips Res. Rep.*, Dec. 1952, Vol. 7, No. 6, pp. 401–420.) Measurements were made of the efficiencies of some 30 phosphors and of the light outputs for thin phosphor layers on aluminized or plain screens. For sulphide mixtures the maximum light output on the glass side was only 25% of the total light emitted. Maximum gain when aluminized screens are used is less than that predicted by theory, because of absorption in the Al layer.

535.37 **3608**

New Phosphors for Flying-Spot Cathode-Ray Tubes.—A. Bril & H. A. Klasens. (*Philips Res. Rep.*, Dec. 1952, Vol. 7, No. 6, pp. 421–431.) The following phosphors are recommended: (*a*) for monochrome television, Ce-activated phosphors, particularly $2CaO.Al_2O_3.SiO_2$-Ce, when the glass used does not blacken under c.r. or soft X-ray bombardment, (*b*) for the red component in colour-television tubes, Bi-activated phosphors, and (*c*) for ultraviolet flying-spot microscropy, ZrP_2O_7 phosphor.

535.37 : 546.472.21 **3609**

Copper-Activated Zinc-Sulfide Phosphors with Yellow and Red Emission.—H. C. Froelich. (*J. electrochem. Soc.*, June 1953, Vol. 100, No. 6, pp. 280–288.) Long-wave emission was obtained by firing in an atmosphere of H_2S and using high activator concentration. The 6 700-Å and 5 800-Å bands are discussed.

535.372 **3610**

Alkaline-Earth Orthophosphate Phosphors.—K. H. Butler. (*J. electrochem. Soc.*, May 1953, Vol. 100, No. 5, pp. 250–255.) The method of preparation and the emission spectra of orthophosphates of Ca, Sr, Ba and their mixtures are described. The primary activator used was Sm. Variation of the Mn secondary-activator content in certain of the phosphors results in a variation of the deep red colour emitted.

537.311.33 **3611**

The Transport of Added Current Carriers in a Homogeneous Semiconductor.—W. van Roosbroeck. (*Phys. Rev.*, 15th July 1953, Vol. 91, No. 2, pp. 282–289.) Transport equations are derived which show the dependence of (*a*) diffusivity and local drift velocity on concentration, and (*b*) the decay-time function on recombination. Carrier depletion with electrical neutrality may occur. For known total current density, the transport problem is determined by the continuity equation alone. In the case of small added-carrier concentration, the continuity equation assumes a mathematically simpler form, and relations are derived for Ge, which give the drift-velocity ratio, diffusivity and group mobility in terms of resistivity and temperature.

537.311.33 **3612**

Preliminary Data on the Relations between Lattice Defects and Debye R.F. Absorption in Iron Oxides.—For 'R. Freymann' please read 'R. Freymann, R. Rohmer & B. Hagene' in 2328 of August.

537.311.33 **3613**

Thermoelectric Measurements on Copper Oxide at High Temperatures.—H. Müser & H. Schilling. (*Z. Naturf.*, Feb. 1952, Vol. 7a, No. 2, pp. 211–212.) A note on experiments made to determine what types of semiconductor mechanism are possible in copper oxide at high temperatures.

537.311.33 : 537.312.6 **3614**

Determination of the Resistance/Temperature Characteristics of Bulk Semiconductors.—R. B. McQuistan. (*Proc. nat. Electronics Conf.*, *Chicago*, 1952, Vol. 8, pp. 525–534.) Description, with detailed circuit diagrams, of equipment using pulses of rectified current for heating the test sample and, in the intervals between pulses, displaying on a c.r.o. the resistivity and temperature of the sample. Graphs thus obtained for Ge, Si and Te agree with those obtained by conventional methods. See also *Electronics*, June 1953, Vol. 26, No. 6, pp. 150–155.

537.311.33 : 539.23 : 546.289 **3615**

Semiconducting Films.—W. M. Becker & K. Lark-Horovitz. (*Proc. nat. Electronics Conf.*, *Chicago*, 1952, Vol. 8, pp. 506–509.) Thin films of Ge deposited on quartz plates by thermal dissociation of GeH_4 were of *p*-type at room temperature. Hall-constant and resistivity curves for the temperature range 77·5–900°K have quite a different shape from those for bulk Ge and the carrier mobilities are lower than usually found for bulk Ge.

537.311.33 : [546.28 + 546.289 **3616**

Diffusion of Lithium into Germanium and Silicon.—C. S. Fuller & J. A. Ditzenberger. (*Phys. Rev.*, 1st July 1953, Vol. 91, No. 1, p. 193.) Results are given of measurements at temperatures between 450° and 1 000°C, using the method previously described [2822 of 1952 (Fuller)].

537.311.33 : 546.289 **3617**

Experimental Confirmation of Relation between Pulse Drift Mobility and Charge-Carrier Drift Mobility in Germanium.—M. B. Prince. (*Phys. Rev.*, 15th July 1953, Vol. 91, No. 2, pp. 271–272.) Experimental data on drift mobilities of minority carriers in Ge are brought into agreement with theoretical predictions by distinguishing between group velocity and particle velocity of a pulse of minority carriers. Corrected high-temperature measurements of electron drift mobility are consistent with the theoretical prediction $\mu = A\,T^{-3/2}$. The experimentally determined value of A is $2{\cdot}0 \times 10^7$.

537.311.33 : 546.289 : 621.317.335.3.029.64 **3618**

Microwave Measurements on Germanium Semiconductors.—F. A. D'Altroy & H. Y. Fan. (*Proc. nat. Electronics Conf.*, *Chicago*, 1952, Vol. 8, pp. 522–524.) Measurements at 3-cm wavelength for a pure Ge sample and for one doped with Sb gave conductivities of $0{\cdot}0266\ \Omega^{-1}\text{cm}^{-1}$ and $0{\cdot}063\ \Omega^{-1}\text{cm}^{-1}$ respectively, the corresponding d.c. values being 0·025 and 0·055. The dielectric constants were respectively 15·4 and 15·6.

537.311.33 : 546.86.48 **3619**

Impurity and Intrinsic Semiconduction of Intermetallic Compounds: Part 1.—E. Justi & G. Lautz. (*Z. Naturf.*, Feb. 1952, Vol. 7a, No. 2, pp. 191–200.) Compounds of Cd and Sb are considered. Experimental and theoretical investigations show that in CdSb the Brillouin zones are filled, hence this material is an intrinsic semiconductor. By adding small amounts of other metals impurity semiconductors are produced; these are examined in respect of variation of resistance with composition at room temperature. Properties distinguishing the intermetallic-compound semiconductors from other types are indicated.

537.311.33 : 621.314.63 **3620**

Relaxation Processes in Rectifying Pairs of Semiconductors.—A. V. Joffe & A. F. Joffe. (*Zh. tekh. Fiz.*, Feb. 1953, Vol. 23, No. 2, pp. 209–224.) In order to investigate the slow processes which occur in a rectifying pair of semiconductors as a result of a contact or applied potential difference, and which gradually alter the resistance of the layers, experiments were conducted with $Cu_2O + TiO_2$ pairs. These experiments are described in detail, with numerous graphs and tables. A theoretical interpretation of the results is given; the primary cause of the observed phenomena is concluded to be the movement of ions in the semiconductors. Calculations of the resistance, based on this assumption, show good agreement with the experimental results.

538.221 **3621**

Magnetic Viscosity of Ni-Zn Ferrites.—E. F. Kuritsyna. (*C. R. Acad. Sci. U.R.S.S.*, 1st May 1952, Vol. 84, No. 1, pp. 45–46. In Russian.) Three samples, of permeability μ_0 600, 650 and 1 600, respectively, were investigated at 80 and 293°K and the last sample also at 195°K. Two maxima of relaxation time were observed, at field strengths corresponding to the points of inflexion of the hysteresis curve.

538.221 : 534.372 **3622**

The Origin of Damping in High-Strength Ferromagnetic Alloys.—A. W. Cochardt. (*J. appl. Mech.*, June 1953, Vol. 20, No. 2, pp. 196–200.) The damping of vibrations is shown to be related mainly to the effect of magnetostriction.

538.221 : 536.631 **3623**

The Specific-Heat Discontinuity in Antiferromagnets and Ferrites.—L. N. Howard & J. S. Smart. (*Phys. Rev.*, 1st July 1953, Vol. 91, No. 1, pp. 17–19.)

538.221 : 538.652 **3624**

Dynamical Physical Parameters of the Magnetostrictive Excitation of Extensional and Torsional Vibrations in Ferrites.—C. M. van der Burgt. (*Philips Res. Rep.*, April 1953, Vol. 8, No. 2, pp. 91–132.) Following a survey of the four sets of magnetostriction equations for given adiabatic conditions and arbitrary depolarization, the complex nature of the dynamic and magnetoelastic constants is discussed. The stress-sensitivity constant and the magneto-mechanical coupling coefficient of several NiZn (ferroxcube IV) materials, determined experimentally, are of the same order as for metallic magnetostrictive materials, but mechanical Q factors up to 15 000 at 50 kc/s have been obtained. The variation of elastic and magnetic lag with frequency and biasing polarization is discussed and the experimental correlation between conductivity and elastic and magnetic losses is explained.

538.632 : [669.245.25 + 669.127.5 **3625**

Hall Effects of the Cobalt Nickel Alloys and of Armco Iron.—S. Foner & E. M. Pugh. (*Phys. Rev.*, 1st July 1953, Vol. 91, No. 1, pp. 20–27.)

539.24 : 537.311.3 **3626**

Conductivity of Thin Films in a Longitudinal Magnetic Field.—E. Koenigsberg. (*Phys. Rev.*, 1st July 1953, Vol. 91, No. 1, pp. 8–9.) Chambers' expression for the conductivity of a thin metal sample in a longitudinal field (2762 of 1950) is evaluated for thin films for two sets of boundary conditions.

546.226.161-1 : 621.317.333.6 **3627**

The Electric Strength of Sulfur Hexafluoride at Radio Frequencies.—J. W. Gibson & C. F. Miller. (*J. electrochem. Soc.*, June 1953, Vol. 100, No. 6, pp. 265–271.) Measurements were made at 60 c/s and in the range 2–16 Mc/s at pressures of 5–20 lb/in.². At 20 lb/in.² the electric strength is 390 V/mil and is constant in a uniform field at frequencies up to 16 Mc/s. The effect of electrode material and of irradiation from a Hg-arc lamp were also investigated.

546.47-31 : [535.343.2 + 537.311.3 **3628**

Optical and Electrical Properties of ZnO Single Crystals with Excess of Zn.—E. Scharowsky. (*Z. Phys.*, 25th June 1953, Vol. 135, No. 3, pp. 318–330.) The dependence of electrical conductivity and light absorption on the excess of Zn was investigated.

546.74 : 621.38 **3629**

Nickel in Electronics.—K. Jackson & R. O. Jenkins. (*Metallurgia, Manchr*, June 1953, Vol. 47, No. 284, pp. 277–282.) A survey is given of the properties of Ni and its alloys and of their applications in electronic valve manufacture. Gas liberation from Ni and the effects of impurities are discussed and the choice of material for various valve components is considered in detail.

621.314.63 **3630**

The Dynamical Theory of Solid Rectifiers: Part 1—Limits of Applicability of the Static Theory.—A. I. Gubanov. (*Zh. tekh. Fiz.*, Nov. 1952, Vol. 22, No. 11, pp. 1803–1813.) From an analysis of nonstationary diffusion equations, limits are determined for the applicability of the static theory of solid rectifiers previously proposed (2745 of 1951 and 424 of 1952) for the case of a rectifier operating with an alternating voltage of the order of a few volts. Increase of the frequency affects first of all the establishment of an equilibrium between the numbers of mobile and fixed charges, and this effect becomes noticeable at the mains frequency of 50 c/s and above. The redistribution of mobile charges in the barrier layer of the rectifier is, however, independent of frequency up to about 1 Mc/s.

621.314.63 **3631**

The Dynamical Theory of Solid Rectifiers: Part 2—Frequency Characteristic.—A. I. Gubanov. (*Zh. tekh. Fiz.*, Nov. 1952, Vol. 22, No. 11, pp. 1814–1826.) The effect is considered of the frequency of an alternating voltage, of the order of a few volts, applied to a solid rectifier, on the distribution of fixed charges in the barrier layer. The distribution of potential is calculated as well as conduction and displacement currents as determined by the frequency. The lowering of the rectification factor in the case of an alternating current is due to the resulting displacement current, which increases with frequency. A numerical example is given for the case of a Cu_2O rectifier, and curves are plotted showing the frequency dependence of the backward and displacement currents of the rectifier, and the variation of these currents with time during the negative half-cycle.

621.315.5.015.5 **3632**

Recent Information on Electric Breakdown.—K. Konstantinowsky. (*Elektrotech. u. Maschinenb.*, 15th May 1953, Vol. 70, No. 10, pp. 224–230.) Results recently published on various aspects of breakdown of insulating materials, cable insulation in particular, are reviewed.

621.315.61 **3633**

Electrical and Physical Properties of IN-420: a New Chlorinated Liquid Dielectric.—A. J. Warner. (*Trans. Amer. Inst. elect. Engrs*, 1952, Vol. 71, Part I, pp. 330–335; *Elect. Commun.*, June 1953, Vol. 30, No. 2, pp. 118–123.) Particularly suitable as an impregnant for capacitors to be used at temperatures up to 125°C.

621.315.616.96 **3634**

New Electrical Insulating Materials.—P. Jolivet. (*Rev. gén. Élect.*, June 1953, Vol. 62, No. 6, pp. 267–275.) The chemical structure, preparation, properties and applications of nine organic solids are detailed: five vinyl resins, aniline formaldehyde, and ethoxylene, alkyd and silicone resins.

621.318 + 621.315.612 + 548.55 **3635**

Modern Trends in Communication Materials.—L. A. Thomas. (*J. Brit. Instn Radio Engrs*, July 1953, Vol. 13, No. 7, pp. 356–360.) Ferroelectric dielectrics, magnetic core materials and synthetic crystals are discussed; crystal valves are not dealt with.

621.318.134 **3636**

Losses in Ferroxcube Rods and Tubes.—H. van Suchtelen. (*Electronic Applic. Bull.*, July 1952, Vol. 13, No. 7, pp.109–114.) Approximate formulae are derived for the loss angle for rods and tubes with windings over the whole or a short part of their length. The results indicate that, for high quality, tube cores and short coils on long rods should not be used. Measurements on ferroxcube IV B rods and tubes confirm the calculations.

666.19 : 621.392.5 **3637**

Vitreous Silica for Ultrasonic Delay-Line Applications.—E. S. Pennell. (*Proc. nat. Electronics Conf., Chicago*, 1952, Vol. 8, pp. 799–810.) The vitreous silica for ultrasonic delay lines must be free from strain or other source of birefringence and must be clear and homogeneous. Tests with polarized light of material from various sources are described. The material Ultrasil obtained from the Heraeus works, Hanau, Germany, was found to be of particularly good quality.

666.29 : 621.387.424.032.79 **3638**

Enamel Seal for Thin Mica Windows in Glass Tubes.—R. Meunier & M. Bonpas. (*Le Vide*, May 1953, Vol. 8, No. 45, pp. 1342–1343.) Description of a process developed for Geiger-Müller tubes.

MATHEMATICS

512.831 : 621.3 **3639**

On Matrix Algebra and the Application of 'Normalized Magnitudes' in Electrical Engineering.—H. Frühauf. (*NachrTech.*, June 1953, Vol. 3, No. 6, pp. 263–268.) The methods are illustrated by calculations of the band-pass characteristics of coupled tuned circuits.

517.9 **3640**

Matrix Solution of Equations of the Mathieu-Hill Type.—L. A. Pipes. (*J. appl. Phys.*, July 1953, Vol. 24, No. 7, pp. 902–910.) The equations discussed are encountered in radio circuit and propagation theory.

518.5 : 621.3.018.75 **3641**

A Graphical Spectrum Analyzer for Pulse Series.—H. P. Raabe. (*Proc. Inst. Radio Engrs*, Sept. 1953, Vol. 41, No. 9, pp. 1129–1138.) Description of theory and operation of a mechanical device similar to a slide rule, for presenting graphically the frequency spectrum of an infinite series of pulses having amplitude modulation but no frequency modulation.

681.142 **3642**

The Approximation of Arc-Tangent ω with a Linear Electrical Network.—D. L. Finn. (*Proc. nat. Electronics Conf., Chicago*, 1952, Vol. 8, pp. 435–444.)

681.142 **3643**

A Stabilized Electronic Multiplier.—C. D. Morrill & R. V. Baum. (*Proc. nat. Electronics Conf., Chicago*, 1952, Vol. 8, pp. 710–715.) Description of a time-division multiplier for obtaining the product of two or more variable voltages to within 0·1%.

681.142 **3644**

Computing Machines.—R. Bird. (*Electronic Engng*, Oct. 1953, Vol. 25, No. 308, pp. 407–410.) Computers for business calculations are considered and methods are described for converting information on punched cards into the binary system, and vice versa.

681.142 **3645**

Serial Digital Adders for a Variable Radix of Notation.—R. Townsend. (*Electronic Engng*, Oct. 1953, Vol. 25, No. 308, pp. 410–416.) Methods are described by means of which business data, such as £. s. d. figures, can be handled in a binary-system computer.

681.142 **3646**

A Magnetic-Drum Digital-Storage System.—B. F. C. Cooper. (*Proc. Instn Radio Engrs, Aust.*, July 1953, Vol. 14, No. 7, pp. 169–177.) This magnetic-drum system was designed as auxiliary store for the C.S.I.R.O. Mark I computer previously described [1050 of April (Beard & Pearcey)]. It has a capacity of 1 024 'words' of 20 binary digits and its rotation rate of 6 000 r.p.m. gives an average access time of 5 ms for random references. Associated basic circuits are described.

681.142 **3647**

Hidden Regenerative Loops in Electronic Analog Computers.—L. G. Walters. (*Trans. Inst. Radio Engrs*, June 1953, Vol. EC-2, No. 2, pp. 1–4.) Detailed analysis showing how coupling due to energy-storing elements gives rise to regenerative loops, the gain of which determines the system stability.

681.142 : 016 **3648**

Storage Systems in Arithmetical Computers.—R. Dussine. (*Onde élect.*, July 1953, Vol. 33, No. 316, pp. 453–455.) Annotated references are given to 21 papers dealing with systems based on pulse counting, or on the use of (*a*) ultrasonic or electrical delay lines, (*b*) c.r. tubes, (*c*) magnetic materials.

681.142 : 621.385.832.082.72 **3649**

The Scotch-Plaid Raster.—J. Kates & V. G. Smith. (*Trans. Amer. Inst. elect. Engrs*, 1951, Vol. 70, Part II, pp. 1480–1484.) Substantial advantages for digit storage on a c.r. tube screen are derived from use of a raster with nonuniform spacing, the raster resembling the stripes of a Scotch plaid. Theory is given for two types, the 'minimum tolerance' and the 'uniform probability' raster.

681.142 : 621.392.5 **3650**

Electrical Delay Lines for Digital Computer Applications.—J. R. Anderson. (*Trans. Inst. Radio Engrs*, June 1953, Vol. EC-2, No. 2, pp. 5–13.) The maximum storage capacity of commercially available lumped-parameter and distributed-parameter delay lines is about 23 and 15 pulses respectively, regardless of total delay time. An analysis indicates that dissipation in inductive elements is the chief limiting factor. Insertion loss can be reduced and storage capacity increased to about 32 pulses by using as inductive elements low-permeability, high-Q, Ni-Zn ferrites around straight single conductors. The design of such units is described.

MEASUREMENTS AND TEST GEAR

621.317.029.6 + 535.22] : 061.3 **3651**

High-Frequency Electrical Measurements. Conference in Washington, D.C.—L. Essen. (*Nature, Lond.*, 11th July 1953, Vol. 172, No. 4367, pp. 52–54.) A brief account of some of the papers read at the third biennial conference organized by the A.I.E.E., I.R.E. and N.B.S. and held during 14th–16th January 1953.

621.317.3 : 535.326 : 621.396.677 **3652**

Microwave Measurements on Metallic Delay Media.—S. B. Cohn. (*Proc. Inst. Radio Engrs*, Sept. 1953, Vol. 41, No. 9, pp. 1177–1183.) Determinations have been made of the refractive index of microwave-lens delay media comprising alternate polyfoam spacers and thin polystyrene sheets printed with patterns of conducting areas. The measurement equipment and method are described and necessary correction formulae are given. Results are presented in the form of graphs useful for design purposes.

621.317.321 **3653**

Voltmeter Loading Again.—R. A. Wiersma. (*Wireless World*, Oct. 1953, Vol. 59, No. 10, pp. 499–500.) A simple method is given for determining the true voltage in a high-resistance circuit when using a single universal meter.

621.317.333.6 : 546.226.161-1 **3654**

The Electric Strength of Sulfur Hexafluoride at Radio Frequencies.—Gibson & Miller. (See 3627.)

621.317.335.029.65 **3655**

Dielectric Constant Measurements at 8·6-mm Wavelength.—P. Hertel, Jr, A. W. Straiton & C. W. Tolbert. (*J. appl. Phys.*, July 1953, Vol. 24, No. 7, pp. 956–957.) Values of permittivity and conductivity were determined from measurements of the attenuation and phase shift of 8·6-mm waves passed through the dielectric. Values obtained for water, ethyl alcohol and soil are compared with the results of other workers.

621.317.335.3 + 621.317.374].029.6 **3656**

Measurements of the Loss Angle and Permittivity of Solid Dielectrics in the Region of Ultra-Short and Decimetric Waves.—K. A. Vodop'yanov & B. I. Vorozhtsov. (*Zh. tekh. Fiz.*, Nov. 1952, Vol. 22, No. 11, pp. 1877–1880.) Various methods developed by Soviet physicists are reviewed and optimum conditions for their application are indicated.

621.317.336 : 621.385.029.64/.65 : 513.647.1 **3657**

Helix Impedance Measurements using an Electron Beam.—Watkins & Siegman. (See 3753.)

621.317.336.1 : 621.314.222 **3658**

A Meter for Measuring the Coefficient of Coupling of I.F. Transformers.—E. A. Saunders & G. R. Cooper. (*Proc. nat. Electronics Conf., Chicago*, 1952, Vol. 8, pp. 609–617.) A method is described for measuring the coupling coefficient k in terms of the change in the anti-resonance frequency of the primary coil and tuning capacitor with the secondary short-circuited and then on open circuit. A variable-frequency oscillator of the type described by Crosby (2156 of 1946), including the primary circuit, is set to zero beat with a fixed-frequency oscillator, the transformer secondary being short-circuited. Measurements of the beat frequency obtained when the transformer secondary is on open circuit are made by means of a single-shot multivibrator giving a series of pulses, of constant amplitude and duration, which are fed to a d.c. microammeter calibrated to read k directly from 0·01 to 0·4, the accuracy being to within 3%. The frequency range covered is from about 300 kc/s to 1 Mc/s secondary resonance frequency.

621.317.342 : 621.392.5 **3659**

Measurement of Group-Delay Time in Networks.—A. van Weel. (*Philips Res. Rep.*, Dec. 1952, Vol. 7, No. 6, pp. 467–473.) Measurement of the phase variation impressed on a l.f. modulation of a h.f. carrier during passage through an unknown network is made by introducing the phase variation into a l.f. oscillatory circuit and measuring the resultant frequency variation. Under stated conditions there is a linear relation between the group-delay variation and the frequency variation. The choice of amplifier delay time (t) and frequency (f) depends on the sensitivity required and the maximum group-delay variation to be measured. For the instrument described $t = \sim 70$ μs, $f = 30$ kc/s, and the sensitivity $\sim 0{\cdot}01$ degree; group-delay variations between 10^{-9} and 10^{-6} sec can be measured.

621.317.361 : 621.396.619.13 **3660**

Direct Measurement of Frequency of Modulated F.M. Transmitters by Pulse Counting.—H. M. Schmidt. (*Frequenz*, July 1953, Vol. 7, No. 7, pp. 203–209.) Two instruments are described capable of indicating the mean frequency of a f.m. signal (with a frequency deviation of $\pm$ 75 kc/s) with an absolute accuracy within $\pm$ 50 c/s. The received signal is heterodyned to a lower frequency, and the count is made in the one case by means of a continuously integrating device, and in the other by means of a circuit which averages the count over an adjustable period.

621.317.382.029.6 **3661**

Mismatch Errors in Microwave Power Measurements.—R. W. Beatty & A. C. Macpherson. (*Proc. Inst. Radio Engrs*, Sept. 1953, Vol. 41, No. 9, pp. 1112–1119.) Expressions are derived for error due to mismatch when an u.h.f. or microwave power meter is calibrated by comparison with a standard power meter, e.g. a bolometer device. The comparison may be made by (*a*) alternate connection to a stable power source, (*b*) simultaneous connection to a T or magic-T junction, or (*c*) alternate connection to a T or magic-T junction. Expressions are also derived for the mismatch error when using a calibrated power meter in the cases where (*a*) the power meter is directly connected to the source, (*b*) an attenuator is interposed, and (*c*) a directional coupler is interposed.

621.317.7 : 621.314.63 **3662**

Harmonic-Insensitive Rectifiers for A.C. Measurements.—R. L. Frank. (*Proc. nat. Electronics Conf., Chicago*, 1952, Vol. 8, pp. 495–505.) Measurements with a rectifier-type instrument of the amplitude of the fundamental of an alternating current may be in error if harmonics are present. The average type of rectifier circuit, with conduction over 180°, is insensitive to even harmonics and the error due to odd harmonics decreases as the order of the harmonic increases. By using a rectifier circuit giving conduction over 120° of phase angle, insensitivity to third harmonics is achieved. The performance of 120°-conduction and 180°-conduction circuits for various harmonic levels is analysed. For nonsinusoidal waveforms the 120°-conduction circuit gives a close approximation to the true r.m.s. value. Applications in a general-purpose a.c. voltmeter and precision triple-balanced phase comparator are described.

621.317.7 : 621.392.43 **3663**

Balance Measurements on Balun Transformers.—O. M. Woodward, Jr. (*Electronics*, Sept. 1953, Vol. 26, No. 9, pp. 188–191.) Balun transformers can be rated in terms of the percentage ratio of the balanced-mode components of load current to total load current. The balance comparator is designed to provide the measurement equipment required. It comprises a generator, the balun under test and a variable load. The load consists of two short slotted coaxial lines, with carbon-resistor inner conductors, arranged at right angles to each other. The indicator has a single-turn loop which is rotated to the positions at which its indication is proportional only to either the unbalanced or the balanced load-current components. Short symmetrical connection between the balun under test and the comparator terminals is essential. Results of measurements on baluns for about 60 and 700 Mc/s are discussed.

621.317.7 : 621.396.619.13 **3664**

The Design of Measuring Equipment for the Determination of Circuit Performance of F.M. Systems.—A. G. Wray. (*J. Brit. Instn Radio Engrs*, July 1953, Vol. 13, No. 7, pp. 363–375.) The measuring equipment is considered under the following headings: (*a*) generators, which may be either of the beat-frequency or the harmonic type; (*b*) detection equipment, in particular deviation meters including low-distortion high-stability discriminators, the pulse-counter type being recommended; (*c*) f.m. station monitor; (*d*) spectrum analysers with c.r.o. display and sweep-frequency oscillator, the sweep produced either electronically or mechanically.

621.317.7.083.4 **3665**

The Requirements and Design for a D.C. Null Detector.—F. L. Maltby. (*Trans. Amer. Inst. elect. Engrs*, 1951, Vol. 70, Part II, pp. 1876–1881. Discussion, pp. 1881–1883.) Equipment is described in which a synchronous reversing switch is used to convert the error direct voltage into an alternating voltage at the mains frequency. This alternating voltage is amplified and applied to the control winding of a 2-phase balancing motor whose reference winding is energized continuously at the mains frequency. Design requirements are discussed and circuit details given.

621.317.715 : 621.314.58 **3666**

The Induction Galvanometer, a Sensitive Instrument Converter.—R. W. Gilbert. (*Trans. Amer. Inst. elect.*

Engrs, 1951, Vol. 70, Part II, pp. 1121–1126. Discussion, p. 1126.) Detailed description of the construction and operation of an instrument consisting essentially of a conventional permanent-magnet moving-coil galvanometer with the addition of means for injecting an a.c. component of magnetic flux into the permanent-magnet field. Deflection of the coil due to d.c. produces a proportionate alternating voltage in the coil, which is amplified by a special frequency-shift amplifier. A commercial design for 200-kc/s conversion is illustrated, for which the conversion gain, defined as the ratio of energy developed in the moving-coil impedance ($\sim$300 Ω) to the d.c. energy producing it, may be as high as 4×10^{10}. When properly adjusted, the galvanometer deflection for full output is estimated as about 5 seconds of angle.

621.317.755.015.7 **3667**
Pulse Oscillograph.—W. Eckardt. (*NachrTech.*, June 1953, Vol. 3, No. 6, pp. 250–257.) Description of equipment for pulse widths of 1–300 μs and pulse repetition frequencies up to 10^5/sec.

621.317.761.029.64/.65 **3668**
Accurate Frequency Meters for the Range 22–37 kMc/s.—Bonnet. (*Onde élect.*, June 1953, Vol. 33, No. 315, pp. 259–269.) A description is given of the construction and method of use of two meters of the cavity-resonator type, differing only in the diameters of the cavities and the spacing of the holes for coupling to a waveguide. Preliminary calibration was effected by means of the NH_3 absorption lines. Accurate measurements make use of harmonics of a 15·275-Mc/s quartz crystal, a c.r.o. method being adopted for identification of the harmonics and of the oscillation modes used, which are mainly H_{011}, H_{012} and H_{013}. Measurement accuracy is of the same order as that to which the quartz-crystal frequency is known.

621.317.783/.784 : 621.316.313 **3669**
An Electronic Wattmeter.—W. B. Boast. (*Proc. nat. Electronics Conf., Chicago*, 1952, Vol. 8, pp. 716–724.) Description of the method of operating conventional electrodynamometers at 10 kc/s from specially designed amplifiers, for making accurate measurements of power and reactive power on the a.c.-network analyser at Iowa State College.

621.317.79 : 621.396.611.21 **3670**
A Review of Methods for Measuring the Constants of Piezoelectric Vibrators.—E. A. Gerber. (*Proc. Inst. Radio Engrs*, Sept. 1953, Vol. 41, No. 9, pp. 1103–1112.) The equivalent-circuit representation of piezoelectric vibrators is shown, and the fundamental parameters indicated. Measurement methods are considered under two headings, viz., routine and laboratory. For routine measurements, methods are preferred in which the crystal under test constitutes the frequency-controlling element in the circuit of the test oscillator. To obtain the greater precision required with laboratory measurements, bridge circuits are preferred.

621.396.615.029.63/.64 **3671**
An S-Band Sweep Generator and Test Set.—J. H. Kluck & R. E. Larson. (*Proc. nat. Electronics Conf., Chicago*, 1952, Vol. 8, pp. 823–830.) The generator provides a single sweep from 2·6 to 3·6 kMc/s or a 100-Mc/s sweep whose centre frequency can be anywhere within the range 2·6–3·8 kMc/s. The frequency sweep is obtained by mixing the outputs of two K-band klystrons. Square-wave modulation equipment, and a detector system consisting of a wide-band bolometer, tuned amplifier and c.r.o., are provided to give a visual indication of the frequency response of S-band components. Typical oscillograms obtained with the equipment are reproduced.

621.396.615.17.015.7 **3672**
A Wide-Range Pulse Generator for Laboratory Applications.—R. W. Frank. (*Proc. nat. Electronics Conf., Chicago*, 1952, Vol. 8, pp. 811–822.) Description, with schematic circuit diagrams, of a pulse generator, developed by the General Radio Co., providing pulses of duration from 0·05 μs to 0·1 sec, with rise and fall times of 0·025 μs. The pulses are derived from a linear saw-tooth wave which is also available. A delay system for phasing applications can be calibrated over the range 1 μs–0·15 sec. The sweep circuit gives an amplitude sufficient for full deflection of the beam of a standard 5-in. c.r. tube. Any output impedance up to 600 Ω can be chosen.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

532.137 : 621-526 **3673**
Design and Construction of a Viscometer with Electronic Servomechanism.—P. Jung. (*HF, Brussels*, 1953, Vol. 2, No. 7, pp. 189–196.) Operation of the Couette viscometer is improved by incorporating a servomechanism.

621.317.083.7 **3674**
A High-Speed Telemetering System with Automatic Calibration.—W. E. Phillips. (*Trans. Amer. Inst. elect. Engrs*, 1951, Vol. 70, Part II, pp. 1256–1260. Discussion, pp. 1260–1261.) Description of equipment using frequency bands of 20–25 c/s and 80–100 c/s for signal transmission. At the receiving end a converter produces a direct voltage proportional to the frequency of the incoming signal, which operates a recorder.

621.317.083.7 : 551.508.11 **3675**
A Radar Sonde Installation.—(*Electronic Engng*, Oct. 1953, Vol. 25, No. 308, p. 416.) Brief description of fully automatic equipment for the British Meteorological Office, consisting of ground interrogator and computing station working in conjunction with airborne transponders. For a somewhat fuller account see *Elect. J.*, 28th Aug. 1953, Vol. 151, No. 9, pp. 651–652.

621.365.54† : 016 **3676**
Bibliography on the Present-Day Technology of H.F. Generators for Induction Heating.—J. Reboux. (*Onde élect.*, July 1953, Vol. 33 No. 316, pp. 456–460.) A list, with notes, of 35 publications dealing with (*a*) the basic theory and practice of h.f. induction heating, (*b*) design of suitable generators, (*c*) practical applications.

621.384.612/.613 **3677**
Automatic Ejection in Betatrons and Synchrotrons.—L. W. von Tersch & R. L. Doty. (*Proc. nat. Electronics Conf., Chicago*, 1952, Vol. 8, pp. 703–709.)

621.384.612 **3678**
Origin of the 'Strong-Focusing' Principle.—E. D. Courant, M. S. Livingston, H. S. Snyder & J. P. Blewett. (*Phys. Rev.*, 1st July 1953, Vol. 91, No. 1, pp. 202–203.) Note pointing out that N. Christophilos, in an unpublished manuscript prepared in 1950, proposed an accelerator on the same lines as that described in 1454 of May.

621.384.622.1 **3679**
A Portable 250-Kilovolt Accelerator.—T. A. Bergstralh, K. L. Dunning, E. Durand, C. H. Ellison, H. K.

Howerton & W. Slavin. (*Rev. sci. Instrum.*, June 1953, Vol. 24, No. 6, pp. 417–419.)

621.385.833 **3680**

Angular Aberrations in Sector-Shaped Electromagnetic Lenses for Focusing Beams of Charged Particles.—E. G. Johnson & A. O. Nier. (*Phys. Rev.*, 1st July 1953, Vol. 91, No. 1, pp. 10–17.)

621.385.833 : 061.3 **3681**

Electron Optics. Symposium in London. — O. Klemperer. (*Nature, Lond.*, 11th July 1953, Vol. 172, No. 4357, pp. 61–62.) A brief report on the Physical Society symposium on 'Recent Research in Electron Optics' held during 15th–16th May 1953.

621.387.424 **3682**

A Battery-Operated Geiger-Müller Counter.—G. Hepp. (*Philips tech. Rev.*, June 1953, Vol. 14, No. 12, pp. 369–376.)

621.389 : 621.396.645.35 : 541.132.3 **3683**

Electronic Instrumentation applied to the Chemical Industry, with special reference to pH Measurement.—G. Hitchcox. (*J. Brit. Instn Radio Engrs*, Aug. 1953, Vol. 13, No. 8, pp. 401–411.) The design of the direct-reading continuous-service pH meter is discussed; this is essentially a stable d.c. amplifier, either direct-coupled or with conversion to a.c.; two typical commercial instruments are described.

621.397.9 **3684**

Measurement of the Size-Distribution of Spray Particles.—L. K. Wheeler & E. S. Trickett. (*Electronic Engng*, Oct. 1953, Vol. 25, No. 308, pp. 402–406.) A phototelegraph transmitter is used to scan spot patterns produced by a spray, and the resulting pulses are passed to a discriminating device.

681.26 : 621.38 : 681.178 **3685**

Ultrarapid Electronic Balance.—A. Jeudon. (*Ann. Télécommun.*, June 1953, Vol. 8, No. 6, pp. 190–196.) Description of feedback equipment producing a current proportional to the applied force up to a maximum weight of 30 g, with application to detection of overweight letters.

PROPAGATION OF WAVES

538.566 **3686**

The Nonexistence of Sommerfeld's Surface Wave.—P. Poincelot. (*Ann. Télécommun.*, June 1953, Vol. 8, No. 6, pp. 206–211.) See 195 of January.

538.566 : 537.311.3 **3687**

General Method of Determining the Conductivity of Heterogeneous Ground. — M. Argirovic. (*Ann. Télécommun.*, June 1953, Vol. 8, No. 6, pp. 212–224.) A critical propagation zone is reached when the value of Sommerfeld's 'numerical distance' is $> e$, the base of Napierian logarithms. Study of the critical point leads to two very simple expressions for the attenuation factor and to a curve differing slightly from that of Sommerfeld near the critical point. From the rule for the sum of numerical distances (distances $< e$), a method is developed for the direct determination of ground conductivities, and zones of equal conductivity, from measured values of field strength and range. A graphical method of evaluating the required quantities is described. The method has been applied to calculations of night-attenuation curves. Comparison of the calculated values of ground conductivity with measured values for particular transmission paths shows satisfactory agreement.

621.396.11 **3688**

Radio Propagation over a Flat Earth across a Boundary Separating Two Different Media.—P. C. Clemmow. (*Phil. Trans. A*, 15th June 1953, Vol. 246, No. 905, pp. 1–55.) A theoretical investigation is made of the propagation of vertically polarized waves, assuming in the first place an earth model consisting of an infinitely thin perfectly conducting half-plane lying in the surface of an otherwise homogeneous earth. The resulting boundary-value problem is solved for a plane wave inclined at an arbitrary angle, and the solution for a line source is obtained in the form of an integral equation. The case when transmitter and receiver are both at ground level is considered in detail; when the distance between them is large the solution becomes simplified, giving results in agreement with those of Feinberg (1902 of 1947). The possibility of an increase of field strength just beyond the boundary is confirmed. A different approximation is obtained when the transmitter and receiver are elevated; this is used to indicate the validity of height-gain factors. For the usual practical case of two regions with arbitrary complex permittivity, approximate boundary conditions are introduced into the analysis. Attenuation and phase curves are given for a numerical example, the attenuation results being in good agreement with those reported by Millington (2307 of 1949).

621.396.11 **3689**

Predictions of Optimum Working Frequency in Washington and in Madrid.—R. Gea Sacasa. (*Rev. Telecomunicación, Madrid*, June 1953, Vol. 9, No. 32, pp. 2–20.) Predictions made by the 'Spanish Method' (3536 of 1952) and based on data for Madrid are compared with predictions made by the C.R.P.L. for a number of circuits in the region of latitude 40°N in North America and for two circuits linking the U.S.A. and Europe. The main points of agreement and disagreement are discussed. It is suggested that when correction is made for seasonal variation of critical frequencies, the sunspot cycle may be found to have no effect on ionospheric propagation.

621.396.11 : 551.510.535 **3690**

Studies on Ionospheric Absorption.—B. Chatterjee. (*Indian J. Phys.*, Dec. 1952, Vol. 26, No. 12, pp. 585–596.) The factors causing ionospheric absorption are briefly reviewed and the method applied in the systematic observations instituted at Calcutta is described in some detail. Typical graphs showing the variation of absorption with frequency are discussed. There is a marked increase in absorption near the critical frequencies of the layers. When magneto-ionic splitting occurs, the extraordinary component usually suffers greater absorption. When an E_s layer is present, certain F echoes are more highly attenuated. Very high absorption on all s.w. frequencies on particular nights is explained as a result of the formation of a sporadic-D layer.

621.396.11 : 551.510.535 **3691**

The Calculation of the Path of a Radio-Ray in a Given Ionosphere.—A. H. de Voogt. (*Proc. Inst. Radio Engrs*, Sept. 1953, Vol. 41, No. 9, pp. 1183–1186.) A general formula is derived giving the electron density in the ionosphere as a third-degree function of height. The form of the distribution curve can be brought into coincidence with previously proposed distribution curves by inserting appropriate values for the formula constants, in accordance with ionospheric soundings. Exact calculations can be made of transmission time, greatest height reached, and distance travelled on the earth's surface, for a given angle of departure of a ray. Details of the calculations are available on request from the PTT-Radio-Service, Scheveningseweg 6, The Hague.

621.396.11 : 551.510.535 : 621.3.087.5 **3692**

COZI Communication-Zone Indicator.—Edwards. (See 3600.)

621.396.11 : 621.317.353.3† **3693**

Solar Influence in Gyro-interaction Experiments.—G. Righini & G. Godoli. (*Ann. Geofis.*, Jan. 1953, Vol. 6, No. 1, pp. 11–19.) Statistical analysis of data collected during the period 1948–1950 shows that it is exceedingly improbable that there is any solar influence on gyro-interaction.

621.396.11.029.63 : 551.5 **3694**

Ultra-short Waves and Meteorology.—J. Dufour. (*Bull. tech. Suisse rom.*, 16th May 1953, Vol. 79, Nos. 9/10, pp. 121–126.) Discussion of tropospheric propagation at wavelengths below 10 m, with particular reference to atmospheric refraction and general meteorological conditions. Observations on Swiss radio links are described.

621.396.812.3.029.62 **3695**

Fading and Loss of Contact on Ultrashort Waves due to the Presence of Inversion Layers.—J. Voge. (*C. R. Acad. Sci., Paris*, 31st Aug. 1953, Vol. 237, No. 9, pp. 491–493.) On the basis of calculations it is concluded that either the direct or the ground-reflected ray may be suppressed over a direct-visibility path, depending on the heights of the path-terminal points relative to the inversion layer. The types of fading or interference resulting in the different cases are indicated.

RECEPTION

621.3.087.4 : 621.396.8 : 551.594.6 **3696**

Intelligibility Recorder for Radiotelegraph Signals.—S. Silleni. (*Ann. Geofis.*, Jan. 1953, Vol. 6, No. 1, pp. 137–153.) A recorder has been developed for atmospheric-noise measurements on single-channel A1 or F1 type transmissions at 10–50 bauds; it can be adapted for other noise measurements. Three voltages are derived. The first (q_S) is proportional to the mean received energy. The second (q_T) represents the energy that would be received in the absence of atmospheric noise. The third (q_B) represents the energy-error level above which an error will be counted, this level being determined from statistical considerations. These three voltages are applied to a circuit which determines the differences $\pm (q_T - q_S) - q_B$. If either difference is positive, a pulse signal is passed to the error-counting circuit actuating the pen recorder. A second counter circuit resets both counters to zero when 1 000 signals have been received. Block and circuit diagrams are shown and explained.

621.396.621.54 **3697**

Design for a Communications Receiver.—W. H. Segrott. (*Short Wave Mag.*, April & May 1953, Vol. 11, Nos. 2 & 3, pp. 85–96 & 154–160.) A description is given, with complete circuit details, of a receiver for the 7-, 14-, and 21-Mc/s bands, designed so that a minimum of test equipment is needed in the construction and alignment. Ganged tuning units are not used. Intermediate frequencies of 5 Mc/s, 465 kc/s and 50 kc/s are used, selectivity and signal/noise ratio being comparable with those of a high-grade commercial receiver. With two converters for the 28-Mc/s and v.h.f. bands, the receiver comprises 9 units, one of which includes a c.r. tube monitor and signal-strength meter.

621.396.621.54.029.51 **3698**

Converter for 200 kc/s.—C. B. Raithby. (*Wireless World*, Oct. 1953, Vol. 59, No. 10, p. 487.) The converter circuit described comprises a triode-hexode mixer with a crystal-controlled oscillator; crystal frequencies ranging from 3 to 10 Mc/s have been found satisfactory. Reception on 200 kc/s can be obtained with any reasonably well screened s.w. receiver.

621.396.622 : 519.272 **3699**

Measurements of Detector Output Spectra by Correlation Methods.—L. Weinberg & L. G. Kraft. (*Proc. Inst. Radio Engrs*, Sept. 1953, Vol. 41, No. 9, pp. 1157–1166.) Correlation technique is applied experimentally to determine the power-density spectra of the outputs of linear and square-law detectors. A digital correlator [1237 of 1951 (Singleton)] is used to obtain values of the input and output autocorrelation functions for inputs of filtered noise either alone or combined with a sine wave; these values are compared with the curves calculated from theory.

STATIONS AND COMMUNICATION SYSTEMS

621.39.001.11 **3700**

The Equivalence of Optimum Transducers and Sufficient and Most Efficient Statistics.—G. W. Preston. (*J. appl. Phys.*, July 1953, Vol. 24, No. 7, pp. 841–844.) A statistical treatment of the problem of recovering the original form of a signal which has become mixed with noise, while preserving information in the Shannon sense. A transducer is discussed which is the physical analogue of a statistic, in the Fisher sense; the properties of the optimum transducer are equivalent to the statistical properties of 'sufficiency' (preservation of information) and 'efficiency' (maximization of fidelity). The case of a Gaussian signal mixed with Gaussian noise is considered in detail; the maximum-likelihood estimate of the signal is derived and its physical analogue is identified as the Wiener smoothing filter with infinite lag.

621.39.001.11 **3701**

Correlation versus Linear Transforms.—D. A. Bell: M. J. E. Golay. (*Proc. Inst Radio Engrs*, Sept. 1953, Vol. 41, No. 9, p. 1187.) Discussion on 1477 of May.

621.39.001.11 : 061.3 **3702**

The London Conference, September 1952: Part 1—Applications of the Theory of Information.—A. Fromageot. (*Onde élect.*, July 1953, Vol. 33, No. 316, pp. 473–478.) A review of the subjects discussed which concern telephony and telegraphy. Part 2: 3601 above.

621.39.001.11 : 517.43 **3703**

Analysis of Signals and of Linear Transmission Operators.—H. Angles d'Auriac. (*H F, Brussels*, 1953, Vol. 2, Nos. 5, 6 & 7, pp. 119–138, 162–172 & 179–187.) Analytical methods of Fourier and Dirac and their application to the determination of the propagation constants of a linear transmission system are explained. An alternative method is to represent the operation of the system by a 'linear transmission operator' which may be reduced to elementary operators of 'filtering' and 'echo' types. The practical significance and application of the method in the case of signals with limited spectrum or finite duration are discussed; an examination of the requirements of a test signal for checking transmission systems indicates the suitability of the 'raised-cosine' signal for this purpose.

621.395.44 : 621.315.052.63 **3704**

Transmission Considerations.—G. E. Burridge & A. S. G. Jong. (*Trans. Amer. Inst. elect. Engrs*, 1951, Vol. 70, Part II, pp. 1335–1340. Discussion, p. 1340.) Full paper. Summary noted in 1102 of 1952.

621.395.44 : 621.315.052.63 : 621.395.822 **3705**

A Study of Carrier-Frequency Noise on Power Lines: Parts 1 & 2.—R. C. Cheek & J. D. Moynihan. (*Trans. Amer. Inst. elect. Engrs*, 1951, Vol. 70, Part II, pp. 1127–1133 & 1325–1334. Discussion, pp. 1133–1134.) Theoretical aspects of the noise problem are discussed and measurement techniques are described. The results of measurements of noise levels on representative power-transmission lines with operating voltages from 23 to 230 kV are presented and discussed. Impulsive noise with repetition rates from 60 to 180/sec was found predominant.

621.396.4 **3706**

Channel-Spacing Considerations in the 154–174-Mc/s Band.—H. E. Strauss. (*Trans. Inst. Radio Engrs*, June 1953, No. PGVC-3, pp. 44–57.)

621.396.41 **3707**

A Report on Channel-Splitting Demonstrations conducted in Syracuse.—N. H. Shephard. (*Trans. Inst. Radio Engrs*, June 1953, No. PGVC-3, pp. 58–66.)

621.396.41 **3708**

Commercial Experience with 160-Mc/s-20-kc/s Equipment.—D. E. Noble. (*Trans. Inst. Radio Engrs*, June 1953, No. PGVC-3, pp. 67–70.) Report of tests carried out on a 20-kc/s split-channel system.

621.396.41.001.42 **3709**

Field Test of Split-Channel 50-Mc/s Systems.—W. M. Rust, Jr. (*Trans. Inst. Radio Engrs*, June 1953, No. PGVC-3, pp. 32–35.)

621.396.41.001.42 **3710**

Operational Experience with a Split-Channel 50-Mc/s System.—J. S. Stover. (*Trans. Inst. Radio Engrs*, June 1953, No. PGVC-3, pp. 36–37.)

621.396.5 : 621.396.8 **3711**

Comparison of Mobile-Radio Transmission at 150, 450, 900 and 3 700 Mc/s.—W. R. Young, Jr. (*Trans. Inst. Radio Engrs*, June 1953, No. PGVC-3, pp. 71–84.) Reprint. See 823 of March.

621.396.619.16 **3712**

Deltamodulation, a Method of P.C.M. Transmission using the 1-Unit Code.—F. de Jager. (*Philips Res. Rep.*, Dec. 1952, Vol. 7, No. 6, pp. 442–466.) Theory of the method is developed, particularly in its relation to information theory. See also 2603 of 1952 (Schouten et al.).

621.396.65 : 621.396.93 **3713**

Frequency Economy in Mobile-Radio Bands.—K. Bullington. (*Trans. Inst. Radio Engrs*, June 1953, No. PGVC-3, pp. 4–27.) Reprint. See 1797 of June.

621.396.65 : 621.396.931 **3714**

Technical Considerations governing the Choice of Channel Spacing in Mobile Communication Bands.—D. M. Heller. (*Trans. Inst. Radio Engrs*, June 1953, No. PGVC-3, pp. 28–31.)

621.396.65 : 621.396.933 **3715**

Multi-Station Air-to-Ground Communications.—(*Wireless World*, Oct. 1953, Vol. 59, No. 10, p. 500.) A v.h.f. area coverage scheme for New Zealand is described, with control centre at Wellington, alternative control centre at Paraparaumu, transmitting and receiving stations at Mount Egmont and Colonial Knob, and two-way radio links between the last-mentioned station and the others. Each link carries six speech channels for modulating ground-to-air transmitters, an engineers' speech channel, and audio tones for switching and remote control. The link transmissions are frequency modulated, with subcarriers for the ground-to-air traffic amplitude modulated on the s.s.b. system. The ground-to-air transmissions are amplitude modulated.

621.396.712.3 **3716**

Technical Arrangements in the Cologne Broadcasting Centre.—O. Bero. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, 1953, Vol. 5, Nos. 5/6, pp. 98–108.) Illustrated description, with block circuit diagrams, of the distribution system linking the various studios and control rooms, and of the control and operating equipment.

621.396.712.3 : 621.395.625.2/.3].004.5 **3717**

Supervision of Electroacoustic Equipment in the Cologne Broadcasting Centre.—F. Enkel. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, 1953, Vol. 5, Nos. 5/6, pp. 109–114.) General description of the arrangements for testing and maintenance of microphones, magnetophones and amplifiers.

621.396.931 **3718**

The [Belgian] Post-Office Public Ground Mobile Radiotelephone Service.—P. Bouchier. (*HF, Brussels*, 1953, Vol. 2, No. 7, pp. 173–178.) A general description of the organization and equipment, with discussion of the frequencies used and the selective calling and secrecy arrangements. Tests made in the Brussels area are reported; the influence of the nature of the terrain is indicated.

621.396.619.001.11 **3719**

Harmonics, Sidebands and Transients in Communication Engineering, as studied by the Fourier and Laplace Analyses. [Book Review]—C. L. Cuccia. Publishers: McGraw-Hill, London, 1952, 465 pp., 76s. 6d. (*Nature, Lond.*, 4th July 1953, Vol. 172, No. 4366, pp. 5–6.)

SUBSIDIARY APPARATUS

621-526 **3720**

Synthesis of Feedback Control Systems by means of Pole and Zero Location of the Closed Loop Function.—M. R. Aaron. (*Trans. Amer. Inst. elect. Engrs*, 1951, Vol. 70, Part II, pp. 1439–1445. Discussion, pp. 1445–1446.) A mathematical description of the desired frequency response, based on the location of the poles and zeros of the closed loop function, is given which ensures that the specifications of velocity or acceleration error constant, bandwidth, and relative stability will be met and which provides for acceptable system transient response. Algebraic manipulation yields the transfer function of the main-loop and subsidiary feedback-loop corrective networks required to meet the specifications.

621-526 **3721**

Servomechanism Transient Performance from Decibel/log-Frequency Plots.—H. Harris, Jr, M. J. Kirby & E. F. von Arx. (*Trans. Amer. Inst. elect. Engrs*, 1951, Vol. 70, Part II, pp. 1452–1459. Discussion, p. 1459.)

621-526 **3722**

Sampled-Data Control Systems Studied through Comparison of Sampling with Amplitude Modulation.—W. K. Linvill. (*Trans. Amer..Inst. elect. Engrs*, 1951, Vol. 70, Part II, pp. 1779–1786. Discussion, pp. 1787–1788.)

621-526 : 621.3.016.352 **3723**

Stability of Varying-Element Servomechanisms with Polynomial Coefficients.—M. J. Kirby & R. M. Giulianelli.

(*Trans. Amer. Inst. elect. Engrs*, 1951, Vol. 70, Part II, pp. 1447–1450. Discussion, p. 1451.)

621-526 : 621.314.3† **3724**

An Improved Magnetic Servo Amplifier.—C. W. Lufcy, A. E. Schmid & P. W. Barnhart. (*Elect. Engng, N.Y.*, April 1953, Vol. 72, No. 4, p. 308.) Digest only.

621.316.072.1/.2 **3725**

Current and Voltage Regulators.—J. Pottier & A. Lavaitte. (*Onde élect.*, July 1953, Vol. 33, No. 316, pp. 442–452.) A review of the principles of various types of regulator utilizing a reference standard. Systems for low and high power, and for direct and alternating current or voltage, are considered.

621.316.722.1 **3726**

An Electromechanical A.C. Line-Voltage Stabilizer.—D. M. Murray & N. L. Kusters. (*Trans. Amer. Inst. elect. Engrs*, 1951, Vol. 70, Part II, pp. 1741–1748.) Description of a stabilizing unit having a short recovery time comparing favourably with that of commercial electronic devices. A temperature-limited diode in a bridge circuit is used to detect the line-voltage variations, and is followed by a phase inverter and a thyratron driving circuit.

621.352.12 **3727**

Recent Developments in the Field of Electric Cells.—J. Pernik. (*Rev. gén. Élect.*, June 1953, Vol. 62, No. 6, pp. 294–303.) Improvements in Leclanché-type cells are reviewed and the construction and properties of (*a*) Ag-Mg cells with AgCl depolarizer, (*b*) C-Zn cells with HgO depolarizer, are described.

621.355.9 **3728**

The Silver-Zinc Accumulator.—C. L. Chapman. (*Electronic Engng*, Oct. 1953, Vol. 25, No. 308, pp. 422–423.) Construction details are described and performance figures given. Types are available with capacities from 0·75 to 60 AH, the corresponding weights ranging from $\frac{3}{4}$ oz to 1 lb. 12 oz. The temperature limits of operation are discussed.

TELEVISION AND PHOTOTELEGRAPHY

621.39.001.11 : [621.396.9 + 621.397.5 **3729**

The London Conference, September 1952: Part 2—Radar and Television.—Loeb. (See 3601.)

621.397 **3730**

A High-Speed Direct-Scanning Facsimile System.—C. R. Deibert, F. T. Turner & R. H. Snider. (*Trans. Amer. Inst. elect. Engrs*, 1952, Vol. 71, Part I, pp. 115–121.) Full paper. See 841 of March.

621.397.5 : 621.39.001.11 **3731**

Economy of Bandwidth in Television.—D. A. Bell. (*J. Brit. Instn Radio Engrs*, Sept. 1953, Vol. 13, No. 9, pp. 447–470.) The nature of television signals and proposed methods of improving the picture-quality/bandwidth ratio are examined in the light of communication theory; some of the less obvious features of the spectrum generated by conventional scanning methods are discussed. The offset-carrier and 'tête-bêche' systems for reducing shared-channel interference are described. Devices for reducing the bandwidth required to accommodate the signal or for increasing the amount of information transmitted with the existing bandwidth are reviewed. 41 references.

621.397.61(410) **3732**

Television in Great Britain.—K. H. Deutsch. (*Funk u. Ton*, June 1953, Vol. 7, No. 6, pp. 288–297.) A general account, dealing with technical transmission data, programme production, studio arrangements, transmitters, and cable and radio links connecting different points of the system.

621.397.611.2 **3733**

New Developments in the Image Iconoscope.—J. C. Francken & H. Bruining. (*Philips tech. Rev.*, May 1953, Vol. 14, No. 11, pp. 327–335.) See 2167 of July (Bruining).

621.397.62 **3734**

Intercarrier-Sound Television Receivers.—A. Boekhorst. (*Philips tech. Commun., Aust.*, 1953, No. 2, pp. 3–15.) Reprint. See 857 of March.

621.397.62 : 621.314.2 **3735**

I.F. Amplifiers with Bifilar-Wound Coils for Television Receivers.—W. Taeger. (*Frequenz*, March 1953, Vol. 7, No. 3, pp. 57–59.) Equivalent-circuit analysis of bifilar i.f. transformers (2138 of 1950), showing how the coupling coefficient can be calculated.

621.397.62 : 621.316.729 **3736**

The Operation of Slow-Acting Deflection-Synchronization Circuits in Television Receivers.—G. W. Kijakowski. (*NachrTech*, June 1953, Vol. 3, No. 6, pp. 269–274. German translation of paper in *Radiotekhnika*, 1952, No. 6.) Analysis is given of the process of frequency and phase control of a sine-wave generator by synchronization pulses. The optimum circuit parameter values can be determined from expressions given for the sensitivity of the system to interference by single pulses and the conditions for rapid aperiodic restoring of synchronism. Experiments confirmed the theory.

621.397.62 : 621.385.832 **3737**

A Short-Length Direct-View Picture-Tube.—J. L. H. Jonker. (*Philips tech. Rev.*, June 1953, Vol. 14, No. 12, pp. 361–367.) Various known methods are discussed for reducing the length of c.r. tubes for television receivers without decreasing the size of the picture. A description is given of an experimental tube with its neck bent through an angle $> 90°$. A permanent magnet is used to bend the beam correspondingly. The usual scanning-deflection angle of 65° is used.

621.397.62 : 621.397.335 **3738**

Flywheel Synchronization.—B. T. Gilling. (*Wireless World*, Oct. 1953, Vol. 59, No. 10, p. 495.) A note of modifications made to the circuit previously described (1499 of May) consequent on the incorporation of a multivibrator sawtooth generator.

621.397.62.002 **3739**

The Production of Television Receivers.—F. Allen. (*J. Brit. Instn Radio Engrs*, Aug. 1953, Vol. 13, No. 8, pp. 383–398.) An account is given of labour-saving methods in use in a particular factory for the quantity production of television components; testing, assembly and packing processes are described.

621.397.621.2 : 535.623 **3740**

Rotating-Screen Color TV Tube.—I. Rehman, E. Singer & C. S. Szegho. (*Electronics*, Sept. 1953, Vol. 26, No. 9, pp. 214 . . 222.) To overcome the appreciable light absorption of colour-filter wheels used in front of the c.r. tube in sequential systems, a continuously pumped projection tube has been developed within which is mounted a tricolour segmented fluorescent disk revolving at 1 800 r.p.m. The disk has a diameter of $7\frac{1}{2}$ in. Phosphors with decay-time constants of the order of $2{\cdot}7 \times 10^{-6}$ sec are required if high resolution is to be maintained; those used for flying-spot c.r. tubes are suitable. Measures to increase available light output are considered.

621.397.621.2 : 621.316.86 **3741**

Frame-Deflection Circuit with VDR [voltage-dependent] **Resistor.**—H. H. van Abbe & A. Boekhorst. (*Electronic Applic. Bull.*, July 1952, Vol. 13, No. 7, pp. 101–108.) The V/I relation for the particular type of resistor used is given by $V = cI^{\beta}$, where β ranges from 0·16 to 0·30. Such a resistor is included in the network coupling the frame oscillator to the output stage, to linearize the sawtooth current supplied to the deflection coils; its operation is frequency-independent. A complete frame-deflection circuit diagram is shown.

621.397.645.026.445 **3742**

An Experimental 100-kW Television Output Stage.—D. Zaayer. (*Philips tech. Rev.*, June 1953, Vol. 14, No. 12, pp. 345–357.) A detailed account is given of experimental work on an output stage required to have a response level to within 1 db over a frequency band of width 6 Mc/s, for 625-line transmissions. High-level modulation is used in a push-pull grounded-grid circuit tunable from 48 to 68 Mc/s and preceded by a 20-kW driver stage, two sub-driver stages and a low-power variable-frequency oscillator.

621.397.645.36 **3743**

Push-Pull Amplifiers for Increased Video Output.—A. Newton. (*Electronics*, Sept. 1953, Vol. 26, No. 9, pp. 228 . . 236.) Improved linearity and higher output are among the advantages of push-pull amplifiers used as video output stages. Design problems involved are discussed, particularly d.c. restoration. Three alternative circuits are shown.

621.397.645.371.029.4/.5 **3744**

Special Amplifiers for Television.—J. Sánchez-Cordovés. (*Rev. Telecomunicación, Madrid*, June 1953, Vol. 9, No. 32, pp. 30–36.) Wide-band amplifiers for television transmitters are considered. The particular types discussed are variants of the cathode repeater and the shunt-regulated cathode-follower for high-level modulators.

621.397.8 **3745**

The Relation between Transient Response and Picture Quality in Television Transmission Systems.—J. Müller. (*Fernmeldetech. Z.*, July 1953, Vol. 6, No. 7, pp. 320–324.) The subjective limits of the perception and the tolerance of various forms of distortion were related to the measured transient response of the system. The tolerable limits of the transient response characteristics, measured as the distortion of a rectangular wave, are tabulated. See also 2063 of 1949 (Kell & Fredendall).

TRANSMISSION

621.396.61 : 621.396.712 **3746**

Modern Transmitter for A.M. Broadcasting.—A. A. McKenzie. (*Electronics*, Sept. 1953, Vol. 26, No. 9, pp. 130–136.) Illustrated description of the 50-kW transmitter at WNEW, New York, a new station which operates 24 hours a day on 1·13 Mc/s. Various techniques are used to ensure uninterrupted operation.

VALVES AND THERMIONICS

621.314.7 : 546.289 **3747**

Bandwidth Limitation of Junction Transistors.—A. J. W. M. van Overbeek & F. H. Stieltjes. (*Wireless Engr*, Oct. 1953, Vol. 30, No. 10, p. 261.) Theory given by Steele (881 of March) makes it possible to calculate a fundamental limit to the wide-band figure-of-merit of junction transistors, if the output capacitance is not taken into account. It is calculated that the separation between emitter and collector corresponding to a figure-of-merit of 1 mA/V per pF is 3 μ for a *p-n-p* transistor and 4·4 μ for an *n-p-n* transistor.

621.383.27 **3748**

Photomultipliers and the Detection of Nuclear Particles.—H. Dormont & E. Morilleau. (*Le Vide*, May 1953, Vol. 8, No. 45, pp. 1344–1352.) Properties required in photomultipliers for scintillation counting are stated. Measuring procedure is indicated for verifying that conditions in respect of gain, matching of photocathode to crystal, and stability are satisfactory. The construction and characteristics are described of a photomultiplier designed by the Laboratoires d'Électronique et de Physique appliquées; its performance is close to that of the R.C.A. Type-5819.

621.385.004.15 **3749**

Reliability of Filamentary Subminiature Tubes.—R. Wood. (*Proc. nat. Electronics Conf., Chicago*, 1952, Vol. 8, pp. 562–567.) Discussion of design factors resulting in improved reliability as regards maintenance of characteristics during life and under conditions of shock or vibration.

621.385.029.6 : 621.396.615.14 **3750**

Traveling-Wave-Tube Oscillators.—H. R. Johnson & J. R. Whinnery. (*Trans. Inst. Radio Engrs*, June 1953, No. PGED-3, p. 24.) Correction to paper abstracted in 1872 of June.

621.385.029.6.016.22 **3751**

Factors affecting Traveling-Wave-Tube Power Capacity.—C. C. Cutler & D. J. Brangaccio. (*Trans. Inst. Radio Engrs*, June 1953, No. PGED-3, pp. 9–23.) The amount and distribution of attenuation in the circuit limit the power output capability of a travelling-wave valve. Empirical design criteria, based on experimental results, are given. These specify the attenuation distributions resulting in a minimum length of active circuit and a minimum gain consistent with maximum efficiency. A beam efficiency of 11% and overall anode efficiency of $\sim$20% were obtained in a valve operating at 4 kMc/s and giving 8 W output power. Efficiency as a function of the gain, space charge and attenuation is considered.

621.385.029.64/.65 **3752**

Starting Currents in the Backward-Wave Oscillator.—L. R. Walker. (*J. appl. Phys.*, July 1953, Vol. 24, No. 7, pp. 854–859.) In this type of oscillator an electron stream interacts with a particular spatial harmonic of one of the transmission modes of a periodic structure such as that used in a travelling-wave amplifier [547 of 1952 (Millman)]. A formula is derived for the starting current, taking account of space-charge effects. The analysis shows clearly that backward-wave oscillation is an interference phenomenon.

621.385.029.64/.65 : 513.647.1 : 621.317.336 **3753**

Helix Impedance Measurements using an Electron Beam.—D. A. Watkins & A. E. Siegman. (*J. appl. Phys.*, July 1953, Vol. 24, No. 7, pp. 917–922.) Measurements were made on a helix in a specially constructed travelling-wave valve. The impedance for the fundamental mode was less by a factor of 0·8–0·3 than that calculated on the basis of Pierce's sheath model. Observations of other modes verify Sensiper's analysis which predicts that certain values of phase constant cannot apply in the case of a single-wire helix. A relation between the impedance for one space-harmonic component and that for its fundamental is derived. The operation of the system as a backward-wave oscillator, continuously tunable over the range 1·5–4·3 kMc/s, is described.

621.385.029.64 **3754**

A Method for the Reduction of the Noise-Factor of Travelling-Wave Valves.—H. Schnitger & D. Weber. (*Fernmeldetech. Z.*, July 1953, Vol. 6, No. 7, pp. 302–310.) The electron beam passes through an auxiliary delay line before entering the main delay line. Due to the interaction between the beam and the wave in the auxiliary delay line, a velocity-modulation noise component is added to the density-modulation noise component; these components cancel one another partially or completely, if the auxiliary delay line is designed correctly. A numerical example is given to illustrate the application of the formulae derived; at λ 7·2 cm the noise factor is 8·5. It is theoretically possible to reduce the noise factor to about 1·2 by this method.

621.385.032.216 **3755**

Characteristic Shifts in Oxide-Cathode Tubes.—W. P. Bartley & J. E. White. (*Trans. Amer. Inst. elect. Engrs*, 1952, Vol. 71, Part I, pp. 43–49.) Full paper. See 3599 of 1952.

621.385.1 **3756**

A Rubber-Membrane Model for Tracing Electron Paths in Space-Charge Fields.—G. A. Alma, G. Diemer & H. Groendijk. (*Philips tech. Rev.*, May 1953, Vol. 14, No. 11, pp. 336–344.) A rubber-membrane method is described in which the effect of space charge on the field is simulated by applying pressure from underneath the membrane. Experiments illustrating various electron-optical phenomena are discussed and the application of the method in the design of a microwave triode is indicated.

621.385.13 **3757**

An Approximation to the Three-Halves Power Law for a Finite Cathode in a Uniform Field.—I. I. Levintov. (*C. R. Acad. Sci. U.R.S.S.*, 21st Aug. 1952, Vol. 85, No. 6, pp. 1247–1250. In Russian.)

621.385.2/.3 : 621.396.822.029.6 **3758**

The Noise of Electronic Valves at Very High Frequencies: Part 1 — The Diode. Part 2 — The Triode.—G. Diemer & K. S. Knol. (*Philips tech. Rev.*, Dec. 1952, Vol. 14, No. 6, pp. 153–164 & Feb. 1953, Vol. 14, No. 8, pp. 236–244.) A more detailed discussion than that noted in 907 of March (Diemer).

621.385.3 : 621.365.54/.55† **3759**

The Design of High-Power Vacuum Tubes for Industrial Heating Applications.—H. D. Doolittle. (*Trans. Amer. Inst. elect. Engrs*, 1951, Vol. 70, Part II, pp. 1934–1937.)

621.385.832 **3760**

Direct-Viewing Memory Tube.—S. T. Smith & H. E. Brown. (*Proc. Inst. Radio Engrs*, Sept. 1953, Vol. 41, No. 9, pp. 1167–1171.) The storage tube described by Haeff (319 of 1948) is modified by using a mesh-type storage screen in combination with a fluorescent screen at relatively high potential. Electrons passing through positively charged areas of the mesh are accelerated to produce an intensified bright image of the charge pattern. Contrast, writing speed, resolution, persistence, and the provision of an electrical output are discussed.

621.385.832 : 535.371.07† **3761**

Monolayer Fluorescent Screens.—L. R. Koller. (*J. opt. Soc. Amer.*, July 1953, Vol. 43, No. 7, p. 620.) Details are given of a method of preparing c.r.-tube screens of monomolecular thickness; the phosphor grains are first coated with a water repellent, e.g., a silicone.

621.385.832 : 621.318.572 **3762**

A Decade Counter Tube for High Counting Rates.—A. J. W. M. van Overbeek, J. L. H. Jonker & K. Rodenhuis. (*Philips tech. Rev.*, May 1953, Vol. 14, No. 11, pp. 313–326.) A description of the tube [3614 of 1952 (Jonker et al.)] is given, and a new circuit is discussed by means of which 30 000 pulses per sec can be counted.

621.387 : 621.318.572 **3763**

The Principles and Method of Operation of some Modern Gas-filled Counter Tubes.—A. B. Thomas. (*J. Brit. Instn Radio Engrs*, Aug. 1953, Vol. 13, No. 8, pp. 414–419; *Proc. Instn Radio Engrs, Aust.*, Aug. 1952, Vol. 13, No. 8, pp. 311–315.) Four types of discharge possible in gas-filled tubes are discussed; the conditions under which each occurs are indicated. The methods of operation of the remtron, dekatron and nomotron are described.

621.396.615.141.2 **3764**

Currents and Space Charges in the Magnetron.—K. Fritz. (*Arch. elekt. Übertragung*, July 1953, Vol. 7, No. 7, pp. 338–340.) Analysis is developed starting from the equation of motion of the optimum-path electron previously derived (2952 of 1952). From consideration of the radial component of electron motion a theoretical upper limit for the anode current is found; this value is nearly attained in operation. Experimental investigation of the tangential component is more difficult. The theoretical results obtained do not depend on electron bunching processes, i.e., only the direction and velocity of the electrons are involved. The theory is applicable both to ordinary and to inverted magnetrons.

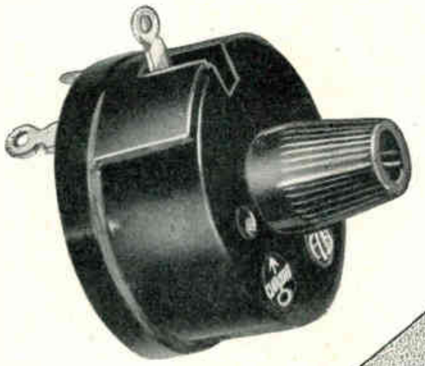
MISCELLANEOUS

621.39 : 061.4 **3765**

Radio Show Review.—(*Wireless World*, Oct. 1953, Vol. 59, No. 10, pp. 446–461.) Report on design tendencies revealed at the 20th National Radio Exhibition (3163 of October). Main attention is devoted to television receivers, but sound receivers, sound reproducers, valves and industrial electronic applications are also discussed. See also *Wireless Engineer*, Oct. 1953, Vol. 30, No. 10, pp. 255–260.

ABSTRACTS AND REFERENCES INDEX

The Index to the Abstracts and References published throughout 1953 is in course of preparation and will be included with the March 1954 issue of *Wireless Engineer*, which will be priced at 6s. for this issue only. Subscribers will obtain the Index automatically since it is now a part of the journal and there is no need for them to make separate application for it. A selected list of journals scanned for abstracting, with publishers' addresses, will be included.



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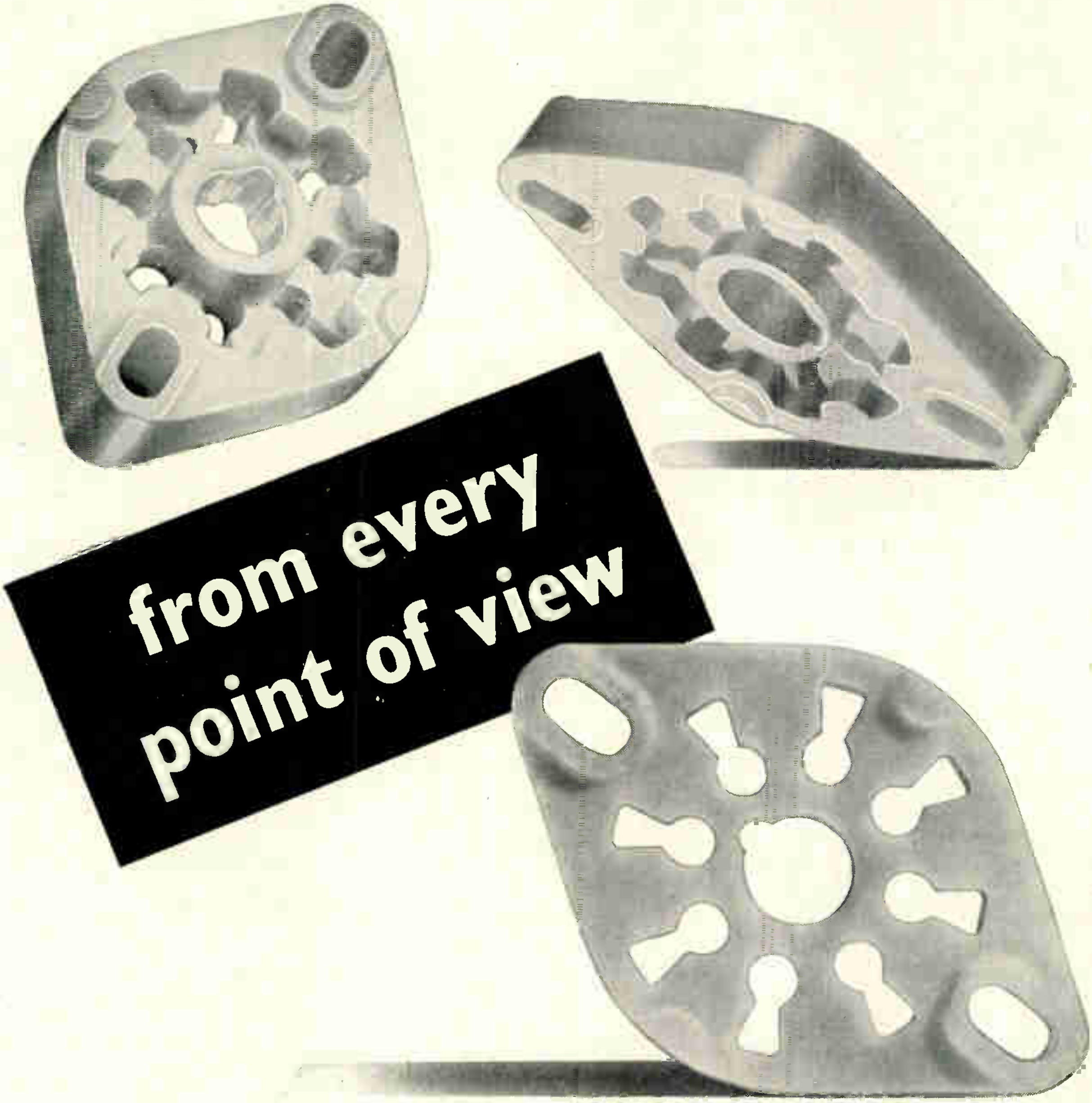

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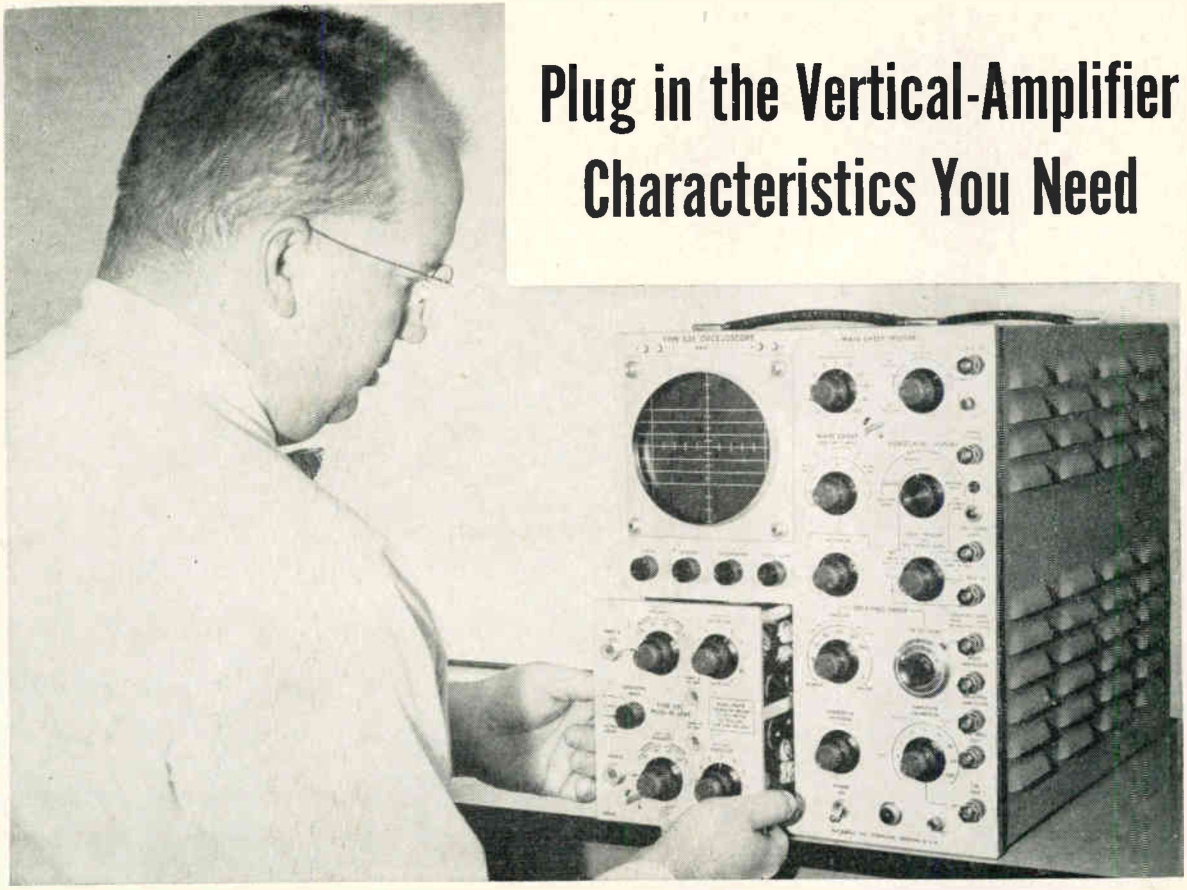

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.1	500	350	2	½	CP465	2/2
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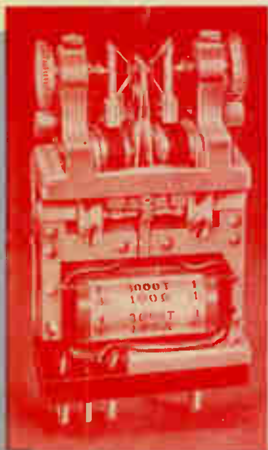
Their high standard of performance makes them suitable for use where signal impulses of varying time duration must be repeated with the utmost accuracy, for instance in telegraphy, measurement, protection and tele-control schemes.

The relay operates successfully direct from valves, the barrier layer type of photo cell, and also from thermo-couples. In the amplification of very weak D.C. signals, the relay's almost perfect contact performance makes it suitable for the inversion of the D.C. into A.C. to permit electronic amplification. *Three Main Types are available.*

TYPE 3

has maximum speed : maximum contact pressure and sensitivity : minimum transit time. Dimensions : 120 mm. × 61 mm. × 33 mm. Weight 625 gm.



TYPE 4

has longer contact travel than the Type 3, with medium speed and sensitivity. Dimensions : 89 mm. × 56 mm. × 25 mm. Weight 368 gm.

TYPE 5

miniature relay with outstanding performance for size. Rugged design of exceptional thermal stability. Dimensions : 56 mm. × 37 mm. × 19 mm. Weight 140 gm.

Relays can be supplied with either solder tag or plug base.

Note.—All dimensions quoted exclude connection terminals.

Complete details and performance data will be willingly supplied on request by the Manufacturers :

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