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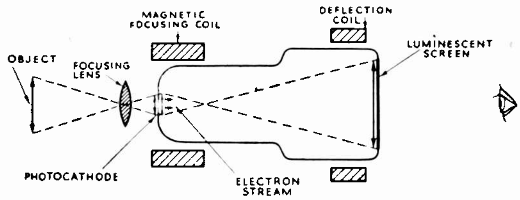
This is an Image Converter

The action of an Image Converter is achieved by first focusing a projected image on to the photocathode; the resultant stream of electrons is then accelerated, and finally focused on to the luminescent screen. The conversion of radiations to electron beams by means of the image converter enables their deflection and modulation by electromagnetic means and thus opens up important new possibilities to designers of electronic equipments. It can, for example, be used for:

- ★ Image magnification, image reduction and intensification, ultra high speed stroboscopes and electronic spectrography.
- ★ Wavelength Conversion i.e. Infra-red or Ultra-violet to visible.
- ★ Studying Ultra-fast Transient Phenomena of the order of 10^{-7} to 10^{-8} seconds—with time base, the gated tube acts as an ultra high-speed camera shutter.

AN IMAGE CONVERTER IS AN ELECTRON-OPTICAL DEVICE WHICH CONVERTS LIGHT OR NEAR LIGHT RADIATIONS FROM A GIVEN SUBJECT INTO A VISIBLE IMAGE ON A LUMINESCENT SCREEN.



PRINCIPLE OF OPERATION

The Image Converters listed below are the first of a comprehensive range at present being developed by Mullard. For full technical information on these tubes, please write to the Communications and Industrial Valve Department.

TYPE:	ME1200AG	ME1201AG	ME1202CA
Description	Visible image converter	Grid-controlled visible image converter	Infra-red image converter
Focusing and Deflection	Magnetic	Magnetic	Magnetic
Photocathode	Caesium-antimony	Caesium-antimony	Caesium-oxidised silver
Sensitivity (At 2,700°K)	20 μA/lumen.	20 μA/lumen.	15 μA/lumen.
Luminescent screen colour	green	green	rapid decay blue
Max. anode-cathode voltage	6 KV.	6 KV.	6 KV.
Max. grid-cathode voltage	—	6 KV.	—
Max. grid-anode voltage	—	6.1 KV.	—
Linear magnification of image	3 times	3 times	1
Screen resolution	200 lines/cm.	200 lines/cm.	200 lines/cm.
Typical operation: Va-k	6 KV.	6 KV.	5 KV.
Vg-k	—	3 KV.	—
Vg-k for extinction of image	—	—100 V.	—

Variants of these tubes with different photocathodes and luminescent screens are being developed

Mullard

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MVT 103 REV (R)

WIRELESS ENGINEER

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Saw-Tooth Generators

THE saw-tooth voltage generator, so widely used as a time base in radar, television and oscilloscopes, has a defective mode of operation which is quite well known but which has received little or no mention in print. There are many types of saw-tooth generator but, in our experience, all of them can exhibit the fault in greater or lesser degree.

This fault is a tendency to produce a saw-tooth waveform in which successive cycles are not identical. The differences between cycles are not random, however, but themselves change regularly in a way which is related to the saw-tooth repetition frequency.

In its commonest form, only alternate saw-teeth are identical so that if the waveform is displayed on an oscilloscope it is necessary to show two cycles of the wave to obtain a clean single trace. If successive cycles are superimposed, as when an attempt is made to show one cycle only, a double trace is obtained.

Somewhat more rarely there are continuous differences between several successive cycles. Thus, the second may differ from the first, and the third from both the first and second, while the fourth, fifth and sixth cycles may be identical with the first, second and third. In this case, a clean single trace is obtained only when the oscilloscope is adjusted to show three (or a multiple of three) traces.

The chief obvious difference between the cycles is in amplitude and, in an extreme case, one cycle may be no more than one-half the amplitude of the next. The defective operation is then very obvious. However, it can be so small that it is barely detectable with an ordinary oscilloscope display, but this is not to say that it is unimportant in some applications.

If good interlacing is to be secured in a television picture, for instance, successive cycles of the frame saw-tooth generator must not differ from one another by more than about 0·05 per cent. It is obvious, therefore, that the presence of this effect in even a very small degree will very seriously affect interlacing. We have previously pointed this out[1] and shown that the correct synchronization of the generator, so that each cycle is correctly initiated by a regularly recurring fly-back, is no guarantee of a correctly interlaced picture.

The effect most commonly occurs in generators of the multivibrator type for it appears to require the presence of two RC time constants of similar orders of magnitude. It can, however, also occur with the blocking oscillator, which has only one essential RC time constant but, in our experience, only when there is a second such time constant present, such as an anode decoupling circuit.

We have been unable to devise a satisfactory explanation of the effect and, apart from the reference already quoted, we have never seen any mention of it in print. In the course of conversation we have found that a good many engineers have encountered it but no one to whom we have so far spoken has been able to say why it occurs. We suspect that the effect may not be a defect in the ordinary sense of the word but that it may be inherent in saw-tooth generators and that its apparent presence or absence is purely a matter of degree.

In view of the widespread use of saw-tooth generators we feel that it is strange that this effect has apparently received no attention and we suggest that it is one which would well repay investigation.

W. T. C.

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[1] "Interlacing", by W. T. Cocking, *Wireless World*, April 1947, p. 124.

SHUNT-REGULATED AMPLIFIERS

Television Modulator Applications

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SUMMARY.—The difficult conditions encountered when modulating television transmitters at high level has resulted in the development of several interesting amplifiers which are described here under the general title of Shunt-Regulated Amplifiers.

It is shown that these amplifiers have improved performance as regards amplitude linearity, frequency response and conversion efficiency. Some types have the properties of positive, zero or negative output impedance, positive, infinite or negative input impedance and are stabilized against supply voltage variations.

Practical circuits and experimental results are given.

LIST OF SYMBOLS

A	Transfer factor or 'gain.'
i_{a1}	Current fluctuation in valve 1.
i_{a2}	Current fluctuation in valve 2.
i_L	Current fluctuation in load.
k	Attenuation or amplification factor.
r_{a1}	Anode a.c. resistance of valve 1.
r_{a2}	Anode a.c. resistance of valve 2.
V_{g1}	Grid-cathode voltage fluctuation in valve 1.
V_{g2}	Grid-cathode voltage fluctuation in valve 2.
V_{a1}	Anode-cathode voltage fluctuation in valve 1.
V_{a2}	Anode-cathode voltage fluctuation in valve 2.
V_i	Input voltage fluctuation.
V_o	Output voltage fluctuation.
V_r	Supply voltage fluctuation.
Z_1, Z_2, etc.	Impedances.
Z_L	Load impedance.
Z_i	Input impedance; i.e., the ratio of fluctuation of input voltage to the resulting fluctuation of input current.
Z'_i	Input impedance at terminals other than the normal input terminal of the amplifier.
Z_o	Output impedance; i.e., the ratio of a fluctuation of output voltage consequent upon a fluctuation of output current to the fluctuation of output current.
Z_{o1}	Output impedance of valve 1.
Z_{o2}	Output impedance of valve 2.

Introduction

IN general terms a high-level video-frequency modulator must have a wide frequency response, and must be capable of maintaining this response across a load consisting of a large shunt capacitance and a shunt resistance which, due to the non-linear flow of grid current will vary from a very high value to a very low value with the amplitude of the signal.

In the particular case of a 50-kW transmitter, grid modulated in its final amplifier, the modulator must produce with good linearity and good frequency response a voltage output of 1,100 V peak-to-peak across the load presented by the grid-modulated amplifier. This consists of a capacitance of 500 pF shunted by a resistance, due to the grid-current load, falling from a resistance of infinity at zero volts to a resistance that may be as low as 400 ohms at 1,100 V peak-to-peak.

Also, whereas the reactive loading is a direct function of frequency, the grid-current loading is a function of amplitude and can be a maximum at any frequency at which full amplitude components are present.

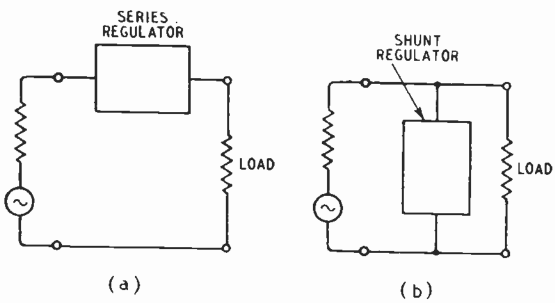


Fig. 1. *Basic regulator circuits; series (a) and shunt (b).*

1. General Approach

We may conceive the general problem as consisting of two parts.

1. Providing the required voltage excursion into a substantially constant load.
2. Providing a regulating system connected in association with the output terminals of 1 to ensure that the voltage excursions across the load are a faithful reproduction of the excursions across the output terminals of 1 despite the load variations of current, either resistive or reactive.

Two ways of fulfilling these requirements suggest themselves and are illustrated in Fig. 1. Various series arrangements (a) are possible but none appears particularly attractive for the problem in hand, but there are some useful shunt forms (b).

The performance of any such system is completely specified by the input and output impedances Z_i and Z_o and the transfer factor; i.e., the 'gain.'

The output impedance Z_o is defined as the ratio

MS accepted by the Editor, February 1951

of a change of output voltage to the initiating change of output current, while the input imped-

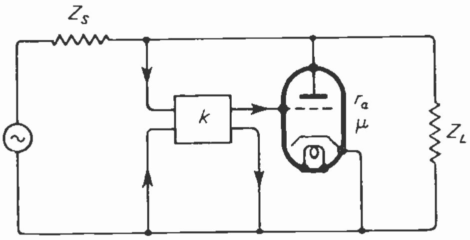


Fig. 2. *Circuit of shunt-regulator Type (a).*

ance Z_i is the ratio of an input voltage fluctuation to the consequent input current fluctuation. The transfer factor is the ratio of a voltage fluctuation at the output to the initiating voltage fluctuation at the input.

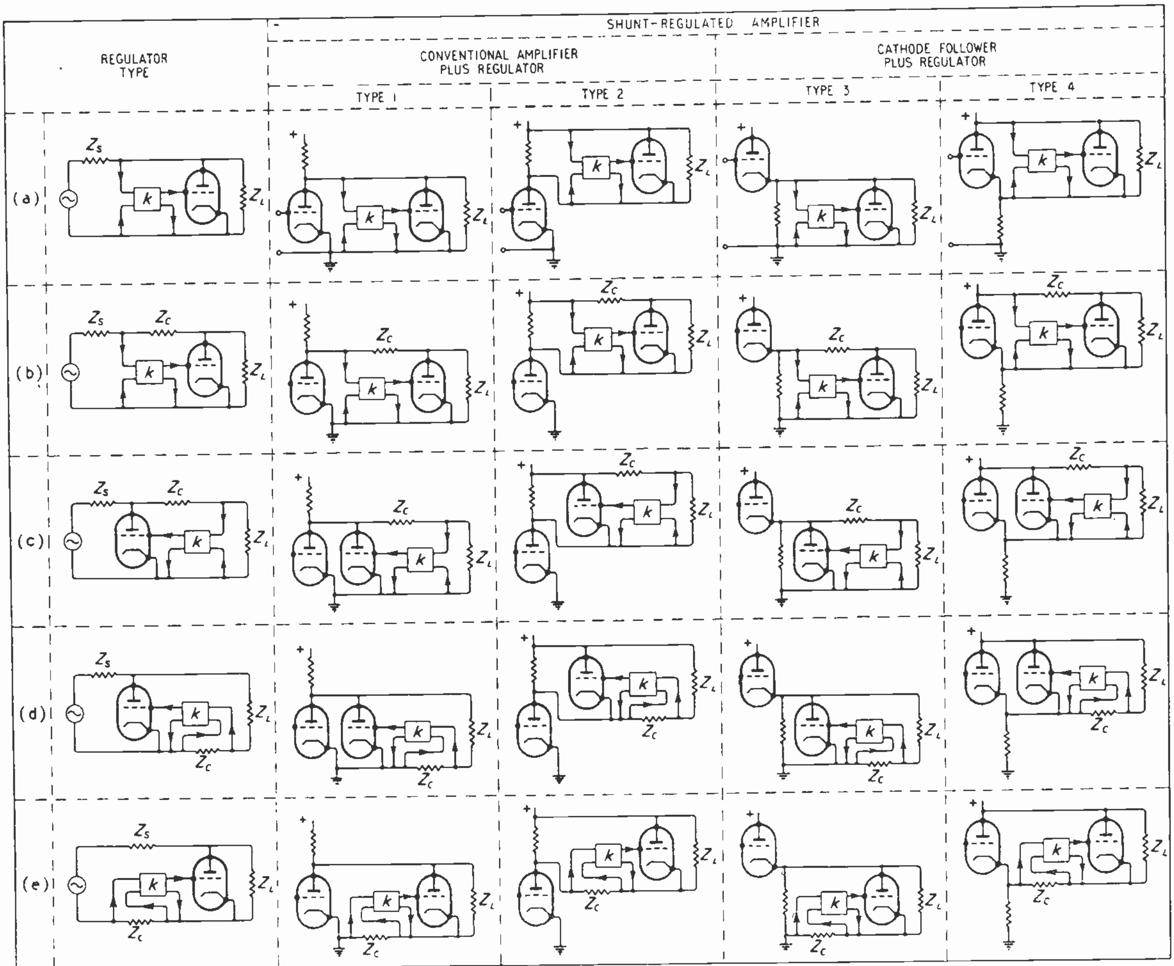
For the purpose in hand we need to find regulators with transfer factors approaching unity and with output impedances significantly lower than the source impedances.

In Table 1 are shown the basic types of shunt regulator in association with amplifier and cathode-follower sources. This table is not intended to be exhaustive, but to typify the approach, and an examination of these combinations reveals interesting possibilities.

2. Shunt-regulated Amplifier Type 1(a)

As a start, let us consider briefly the system shown in Table 1 as Type 1(a). The circuit is re-drawn in Fig. 2 with the amplifier represented by the source generator and impedance Z_S. Expressions for Z_i, Z_o and 'gain' A can be obtained in the usual way by writing the circuit equations and solving for these quantities. As the analysis is straightforward only the results are given here.

TABLE 1



In these diagrams the rectangle k represents an amplifier or attenuator of output/input voltage ratio k with or without a phase reversal according to the requirements of the particular circuit.

For the circuit of Fig. 2 this procedure gives:—

$$Z_i = Z_s + \frac{Z_L r_a}{r_a + Z_L (1 + \mu k)} \quad \cdots \quad \cdots \quad (1)$$

$$Z_o = \frac{Z_s r_a}{r_a + Z_s (1 + \mu k)} \quad \cdots \quad \cdots \quad (2)$$

$$A = \frac{Z_L r_a}{Z_L\{r_a + Z_s (1 + \mu k)\} + Z_s r_a} \quad \cdots \quad (3)$$

When the two valves of the regulated amplifier are alike and the anode loading resistor is high, $Z_s \approx r_a$ and then

$$Z_o = \frac{r_a}{2 + \mu k} \quad \cdots \quad \cdots \quad \cdots \quad \cdots \quad (4)$$

$$A = \frac{Z_L}{r_a + Z_L (2 + \mu k)} \quad \cdots \quad \cdots \quad (5)$$

It is, therefore, clear that if $Z_L = r_a$ the gain becomes $A = 1/(3 + \mu k)$ while if $Z_L = \infty$, the gain is $A = 1/(2 + \mu k)$, so that if A is to approach unity k must be fractional and negative. Even when this is so, the improvement in Z_o is not good, and when Z_o is made low by reducing the negative numerical value of k, gain is sacrificed.

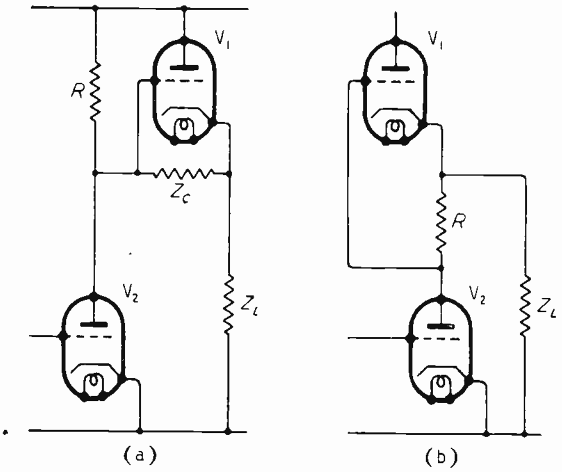


Fig. 3. *Circuit of shunt-regulator Type 2(e); general form (a), simplified form for similar valves (b).*

The single conventional amplifier on its own with $Z_L = r_a$ has a gain of $\mu/2$. If we make $k = -2/\mu$ this is tantamount to connecting the two valves in parallel and we find, as we should expect, that in this case $A = 1$.

The result $Z_o = \infty$, obtained by substituting $k = -\mu/2$ in the above formula, is anomalous because the input voltage to the regulator is then independent of the output and the relation derived for the regulator as such cannot be applied.

What is revealing, however, is that the combination of an amplifier with this simple regulator can equally well be affected as in Table 1, type 2(a) with substantially the same results. In this case the arrangement may be further simplified, for the valves may carry equal mean currents, while the current fluctuations available at the output terminals are approximately the same as those obtainable from the two valves in parallel. As a practical application therefore, the series arrangement of valves may be more attractive than the more conventional parallel arrangement. This was foreseen by Rudolf Urtel[1] in 1935, and was suggested in a different form by N. H. Clough[2] in 1940.

As a general rule in examining practical possibilities, we shall concentrate on the types in which the valves are in a series d.c. connection.

3. Shunt-regulated Amplifier Type 2(e)

The type next to be considered is shown in Table 1 as Type 2(e) since this arrangement shows very considerable advantages over normal technique. Since A can approach unity with k unity and Z_c small, we can simplify the arrangement to the form of Fig. 3(a). Now if Z_c approaches infinity the regulator valve V_1 is a conventional cathode-follower and supplies substantially all the load demand of current. On the other hand, if $Z_c = 0$ the regulator valve acts merely as a resistance and makes no individual contribution of current to the load, all the load now being transferred to the first valve. The intermediate conditions are interesting.

Since the excitation of valve V_1, and therefore its contribution of current to the load, are directly proportional to the current in its grid-to-cathode impedance Z_c, the two valves V_1 and V_2 can be made to share the non-linear resistive load currents and we may conceive the valve V_1 as a current 'booster.'

From Fig. 3(a) we can say that V_1 supplies current in accordance with the demands of the load, and it does this because of the current-monitoring impedance Z_c. Also, as the signal current output of the valve V_2 increases, the current in the valve V_1 tends to decrease and a partial current-balancing effect occurs.

A further simplification is possible where the valves are similar, for then their mean currents can be made equal; the circuit can then be reduced to that of Fig. 3(b). The output impedance of this arrangement lies between the a.c. resistance r_a of the valve V_2 and $1/g_m$ of the valve V_1 depending upon the value of the resistor R. The operation of this system can be explained quantitatively by reference to the characteristics of the valves which, for simplicity, are taken to be similar.

3.1 *Graphical Analysis.*

Typical characteristics for type ACM3 valves are shown in Fig. 4; assuming a given instantaneous current flowing through the system and a given h.t. voltage, we can derive the grid-cathode excitation of the valve V_1 by taking the product of the current and the resistance R. This settles the voltage drop V_{a1} across the valve V_1 because both current and grid voltage are known and, therefore, the voltage V_{a2} across the valve V_2. Current and voltage relations in the valve V_2 are then known and the voltage excursions for the two valves can be drawn.

Assuming an h.t. supply of 2,500 V and $R = 100\ \Omega$, then $V_{a2} = 2{,}500 - iR - V_{a1}$.

TABLE 2

i	iR	V_{a1}	V_{a2}
0·2	20	530	1950
0·4	40	980	1480
0·6	60	1420	1020
0·8	80	1820	600

The anode excursion of valve V_1 is shown in Fig. 4 by the plot ABC and that of valve V_2 by plot DBE. It is clear from these plots that the valve V_2 works into a normal load condition determined by the slope of the line DBE and the effective anode load is $r_a + (\mu + 1)R$.

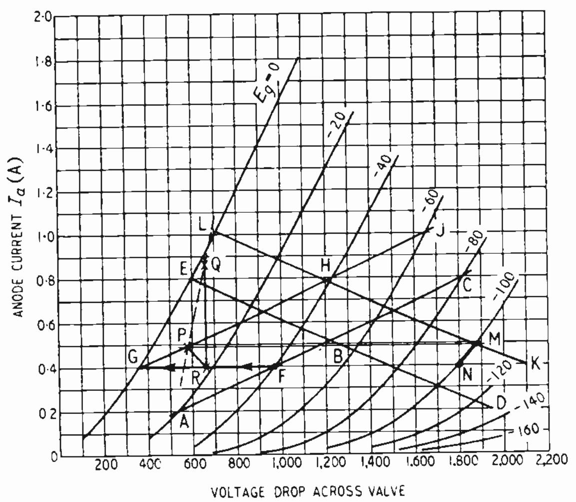


Fig. 4. *Characteristics of ACM3 valve with load lines drawn for a shunt-regulated amplifier Type 2(e).*

If now the lower limit of current excursion is decided, say point F on curve ABC, this value of current can be made to coincide with the $E_g = 0$ characteristic (for class A operation) for the valve V_1 by adding bias. This moves point F to point G and transposes both load lines to the positions shown by the plots GHJ and KHL. It is approximately equivalent to increasing the h.t. voltage by μ times the bias applied to effect the transposition. In this condition the a.c. resistance of the valve V_1 with its feedback resistor remains constant at the previous value of $r_a + (\mu + 1)R$ while the d.c. resistance is appreciably reduced. The linearity of the voltage excursion is also considerably improved by this transposition as can be readily seen in Fig. 4.

When working into a reactive load as in Fig. 5, the capacitor C is charged by a rising voltage at the output terminal Y. This corresponds to a falling current in valve V_2. With the valve V_1 replaced by the conventional resistor the total charging current must flow in the circuit of valve V_2 and the maximum charging rate is determined by the magnitude of the current change in V_2. With V_2 operative, however, the voltage at the point Y tends to rise more rapidly since additional charging current can be supplied by the valve V_1, which acts partially as a cathode follower.

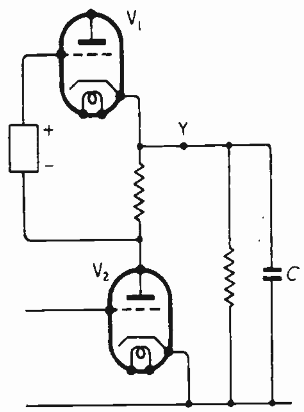


Fig. 5. *Shunt-regulated amplifier Type 2(e) with capacitive load circuit.*

The output impedance and the sharing of the load-current demands by the two valves V_1 and V_2 can be found quantitatively from the usual form of circuit analysis. The results only are quoted here. The output impedance is easily shown to be

$$Z_o = \frac{r_{a1}\,(r_{a2} + R)}{r_{a2} + R(\mu_1 + 1) + r_{a1}} \quad \ldots \quad \ldots \quad (6)$$

and is equivalent to two current generators in parallel of output impedances

$$r_{a2} + R\ (\text{Valve } V_2) \text{ and } \frac{r_{a1}(r_{a2} + R)}{r_{a2} + (\mu + 1)R}\ (\text{Valve } V_1)$$

and the non-linear resistive and reactive load currents are supplied from the two equivalent generators in the inverse ratio of the generator impedances.

For the ACM3 valves taken to illustrate the system, the ratio becomes approximately 4 to 1 for a value of R of 100 ohms; i.e., 4/5 of the load current is supplied by the 'booster' valve. The output impedance, using the same numerical values, is about 150 ohms and this reduced output impedance may be demonstrated graphically with the aid of Fig. 4.

Assume that a load has been applied at the output terminal Y and a voltage drop occurs causing the current in the resistance R and the valve V_2 to fall. Let this fall of current be 0·095 A from point M to point N. The drop of voltage at the anode of valve V_2 is then 75 V, and so the drop of voltage at the cathode of valve V_1 is $75+100\times0{\cdot}095=84{\cdot}5$ V. The 9·5-V drop across the resistor R changes the grid-cathode voltage of valve V_1 from -10 to $-\frac{1}{2}$ V.

The new current in valve V_1 is thus approximately 0·88 A as shown by the construction PRQ the point Q being 84·5 V higher than point P and at a grid–cathode voltage 9·5 V more positive. The current flowing into the load is thus

$$QR = 0{\cdot}88 - 0{\cdot}4 = 0{\cdot}48 \text{ A}$$

and the output impedance is

$$\frac{84{\cdot}5}{0{\cdot}48} = 176\ \Omega.$$

Therefore, the load division ratio between valves is

$$= \frac{0{\cdot}48 - 0{\cdot}095}{0{\cdot}095} = \frac{0{\cdot}375}{0{\cdot}095} = 4/1.$$

This checks roughly with the theoretical figures obtained from the foregoing equation by substituting values of r_a and μ obtained from the valve characteristic at the appropriate points; i.e., $r_{a1}=600\,\Omega$, $r_{a2}=800\,\Omega$, $\mu=16{\cdot}5$, $Z_o\approx171\,\Omega$, load division $\approx 4{\cdot}1$. The agreement is within the tolerance of estimation from the curves.

Regarded as a single equivalent valve, the characteristics for $E_g = -100$ V pass through the points PQ. The usual limitations associated with current cut-off with reactive loading are reduced since the load lines are raised well away from the zero-current region. It is interesting to note that the voltage obtainable at the anode of the valve V_2 is higher than at the cathode of valve V_1 by the iR drop in the resistor R. In practice, therefore, one or other output terminal can be used depending upon the relative importance of output impedance and output voltage.

The gain of the shunt-regulated amplifier of this type is given by

$$\frac{V_{out}}{V_{in}} = \frac{\mu_2(r_{a1} + \mu_1 R)}{r_{a1} + r_{a2} + (\mu + 1)R} \qquad .. \quad (7)$$

which for the practical values already taken is approximately $\frac{3}{4}\mu$.

A direct comparison between the properties of two ACM3 valves arranged as above and arranged conventionally as two valves in parallel to give the same gain is interesting and is given in Table 3.

4. Shunt-regulated Amplifier Type 3(e)

The same form of regulator [Type (e)] but associated with a cathode-follower

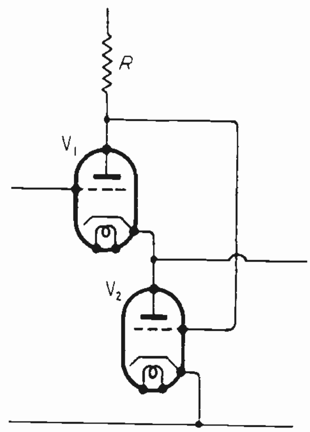


Fig. 6. *Shunt-regulated cathode-follower Type 3(e).*

TABLE 3

	Conventional	Shunt Regulated
ACM3 taken as $\mu=15$, $r_a=600\,\Omega$		
Valves	2 ACM3 in parallel	2 ACM3 in series
Gain	11	11
Output Voltage (p-p)	1000	1000
Resistor	825 Ω	100 Ω
Watts dissipated in resistor	2·1 kW	64 W
Input capacitance (approx.)	400 pF	200 pF
Departure from amplitude linearity in response to a sawtooth waveform	6 % approx.	1 % approx.
H.T. required	2900 V	2500 V
Mean current	1·6 A	0·8 A
H.T. power input	4640 W	2000 W
Output impedance	228 Ω	175 Ω
Current fluctuation in supply for 1000 V swing	1·2 A	0·6 A

source, as in Table 1, arrangement 3(e) is also of particular interest.

Simplified as for Type 2(e) we get the arrangement of Fig. 6, the current-monitoring resistor being transferred for convenience to the anode side of the cathode-follower source valve to get a fractional value of k and a phase reversal. The output impedance is

$$Z_o = \frac{(r_{a1}+R)r_{a2}}{(\mu_1+1)r_{a2}+r_{a1}+[(\mu_1+1)\mu_2+1]R} \qquad (8)$$

and is composed of $\frac{r_{a1}+R}{\mu_1+1}$ due to the cathode-follower valve, in parallel with

$$\frac{r_{a2}(R+r_{a1})}{r_{a1}+[(\mu_1+1)\,\mu_2+1]R},$$

and in the limit when $R \gg r_{a1}$

$$Z_o \text{ becomes } \frac{1}{(\mu_1+1)g_{m2}}.$$

In practice this order of improvement is not possible, but very worth-while reductions of output impedance are easily achieved.

Taking a pair of ACM3 valves, as before, with 100 ohms as the control resistor the output impedance becomes approximately 10 ohms as compared with $1/g_m$ for this valve of approximately 36 ohms. As for the amplifier arrangement, the operation can be studied graphically. This is done in Fig. 7 for two ACM3 valves.

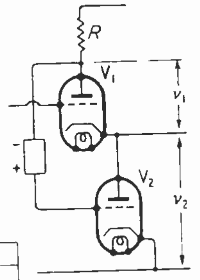
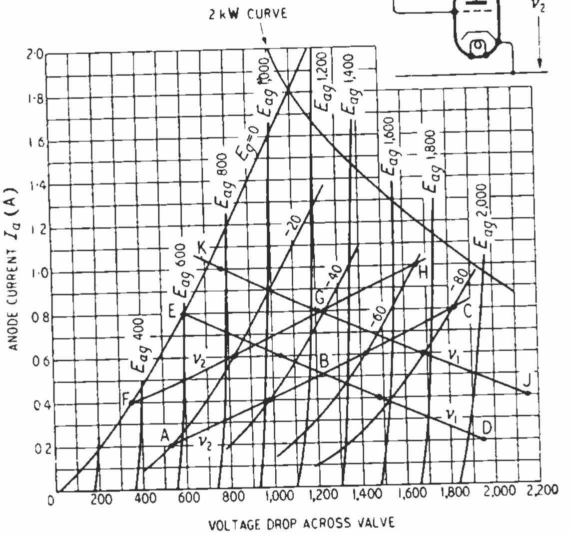


Fig. 7. *Characteristics of ACM3 valve with load lines for a shunt-regulated cathode-follower Type 3(e).*



As before, assuming a given instantaneous current flowing through the system with no load connected to the output terminals, and a given h.t. supply voltage, we can derive the grid-cathode excitation of the valve by taking the product of the current and the resistance R. This settles the voltage drop across the valve V_2 and, therefore, the voltage across the valve V_1. Current and voltage excursions for the two valves can then be drawn.

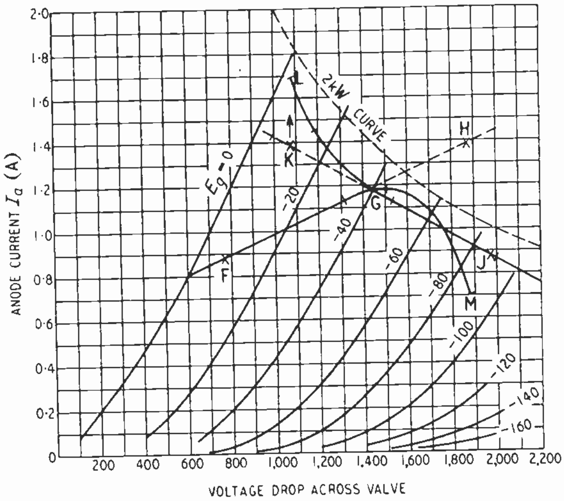


Fig. 8. *Characteristics of ACM3 valve with curved load lines due to grid current in the following stage; shunt-regulated cathode-follower Type 3(e).*

This is demonstrated in Fig. 7 by the excursions ABC and DBE. As before, the addition of positive bias to the valve V_2 will transpose the excursions to FGH and JGK. It should be noted that the positive bias is only relative to the signal circuit and that, in fact, the potential of the grid of the valve V_2 will be very much less positive than the anode of the valve V_1 when all the mean d.c. components are added.

The load lines drawn in Fig. 7 show the excursions when there is no current demand by the load. If such an arrangement is used to grid modulate a transmitter, the non-linear grid current loading changes the shape of the 'load lines' as in Fig. 8. The ordinate amplitude HM represents the share of grid current delivered by the 'regulator' valve; the ordinate KL represents the share of grid current delivered by the cathode-follower valve, the total grid current being represented by the vertical distance between L and M.

It will be seen that the curvature of the load line FGM has been used to raise the mean current of the valves to permit the maximum possible margin for reactive current.

A comparison in Table 4 is made between the shunt-regulated cathode-follower described and a conventional cathode-follower designed to give the same output voltage swing with the same reactive-current handling capacity.

5. Shunt-regulated Amplifier Type 2(e/e)

Another interesting arrangement is derived from the combination of the shunt-regulated amplifier Type 2(e) described, and a further stage of shunt-regulation of the basic Type (e) of Table 1. This is shown in Fig. 9(a) and for optimum conditions the amplifier k must be phase-reversing and have fractional gain. Excitation for the additional regulator of this phase and magnitude can be obtained from an impedance in the anode circuit of the first regulator valve as shown in Fig. 9(b).

Further simplification is possible. The voltage excursions across the valves V_3 and V_2 differ only by the relatively-small voltage developed across the impedance Z_2. We may, therefore, combine valves V_3 and V_2 and the only substantial difference will be the feedback introduced by the connection from the impedance Z_1 to the input grid. This simplification is made in Fig. 9(c).

When the load Z_L is applied, feedback proportional to that part of the load current taken through the valve V_1 is applied to the input

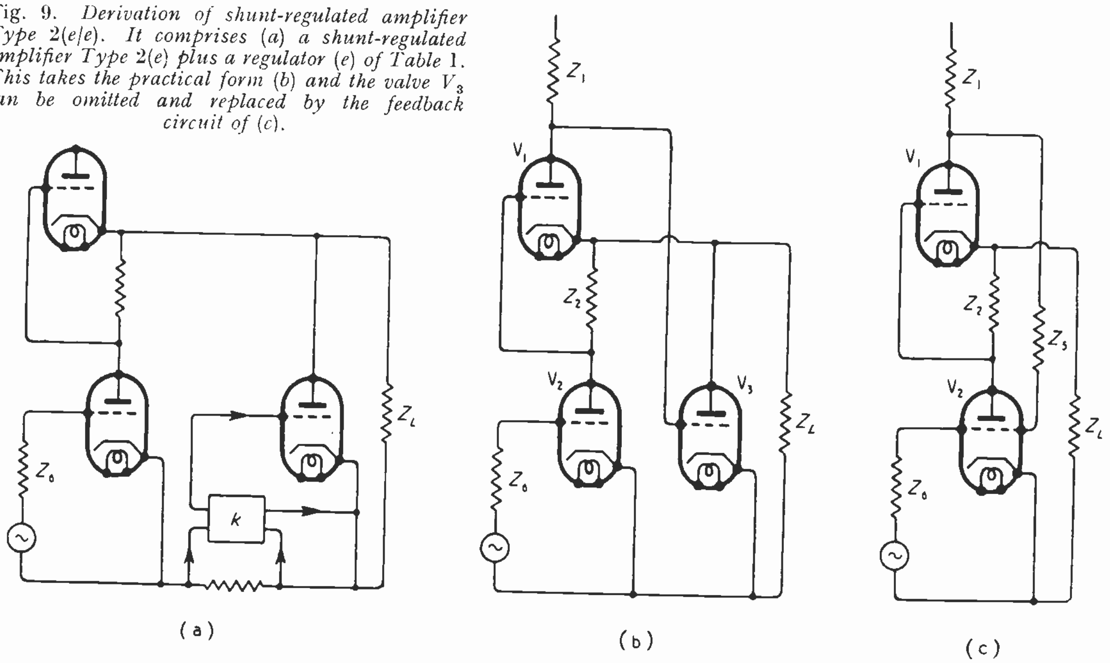


Fig. 9. *Derivation of shunt-regulated amplifier Type 2(e/e). It comprises (a) a shunt-regulated amplifier Type 2(e) plus a regulator (e) of Table 1. This takes the practical form (b) and the valve V_3 can be omitted and replaced by the feedback circuit of (c).*

TABLE 4

	Conventional	Shunt Regulated
ACM3 taken as $\mu = 14$, $r_a = 600\ \Omega$		
Valves	2 ACM3 in parallel	2 ACM3 in series
Output voltage (p-p)	1000	1000
Gain	0·91	0·91
Resistor	875 Ω	100 Ω
Watts dissipated in resistor	5 kW	140 W
Output impedance	18 Ω	10 Ω
Mean current	2·4 A	1·2 A
H.T. required	3480 V	3000 V
H.T. power input	8·3 kW	3·6 kW

terminal through the network composed of Z_5 and Z_6.

This feedback can be used with advantage to increase the input impedance of the stage. The arrangement of Fig. 9(c) will have the improved linearity of the Type 2(e) shunt-regulated amplifier already described and is capable of the same graphical analysis. The only differences in the graphical analysis are the displacement of the load lines by the additional voltage drop in the additional impedance Z_1 and the need for computing the input excitation in view of the feedback now operating.

In addition, this circuit may have positive, zero or negative output impedance, positive, infinite or negative input impedance and also possesses the property of inverting impedances between output and input terminals. Thus with the system adjusted to give substantially zero output impedance we may add capacitance across the output terminals and reflect negative capacitance across the input terminals, thus improving frequency response in the driving source.

This property arises from the fact that the feedback operating on the input terminal is positive and dependent on load current. If, therefore, the load takes a large capacitive current, the voltage feedback will be leading in phase and produce the effect of negative capacitance at the input terminals.

Examination of the formulae shows that the specific attainment of zero output impedance is dependent upon constancy of the valve parameters and, in practice, an amplifier adjusted to zero output impedance at one point on its characteristics will depart from zero impedance during a voltage excursion. It is, however, true to state that the departure from constancy will be little different from that experienced in any other sort of amplifier, and that a change from, say, 1 ohm to 2 ohms does not detract from the usefulness of being able to achieve an output impedance of this order.

In practice this property is equally useful in dealing with the regulation due to non-linear grid current loading and in dealing with heavy reactive loading.

The properties of this arrangement are summarized below, reference being made to the circuit shown in Fig. 9(c).

The output impedance is given by:—

$$Z_o = \frac{(r_{a2}+Z_2)(r_{a1}+Z_1)(Z_5+Z_6) + (r_{a2}+Z_2)r_{a1}Z_1 - \mu_1\mu_2 Z_1 Z_2 Z_6}{(r_{a2}+Z_2)(Z_5+Z_6) + (r_{a2}+Z_2)Z_1 + \mu_2 Z_1 Z_6 + (r_{a1}+Z_1)(Z_5+Z_6) + Z_1 r_{a1} + \mu_1 Z_2(Z_5+Z_6) + \mu_1 Z_1 Z_2} \quad (9)$$

where μ_1, r_{a1} refer to V_1 and μ_2, r_{a2} refer to V_2, and this is zero when

$$(r_{a2}+Z_2)(r_{a1}+Z_1)(Z_5+Z_6) + (r_{a2}+Z_2)r_{a1}Z_1 = \mu_1\mu_2 Z_1 Z_2 Z_6 \quad \ldots \quad (10)$$

The output impedance varies rapidly with change of feedback ratio k where $k = \dfrac{Z_6}{Z_5+Z_6}$

The division of loading between the two valves is in inverse proportion to the output impedance of each branch of the system and is in the ratio:—

$$\frac{i_{a2}}{i_{a1}} = \frac{\mu_2 Z_1 k(Z_5+Z_6) + (Z_1+r_{a1})(Z_5+Z_6) + r_{a1}Z_1}{\mu_1 Z_2(Z_5+Z_6) + \mu_1 Z_1 Z_2 + (r_{a2}+Z_2)(Z_5+Z_6+Z_1)} \quad \ldots \quad (11)$$

where i_{a1} and i_{a2} are the anode currents of V_1 and V_2 respectively.

This rises with increase of k if all other values are kept constant and is dependent significantly on the values Z_1 and Z_2.

The input impedance is:—

$$Z_i = \frac{Z_L+Z_o}{Z_L}\left[\frac{(r_{a2}+Z_2)(Z_5+Z_6) + (r_{a2}+Z_2)Z_1 + (r_{a1}+Z_1)(Z_5+Z_6) + r_{a1}Z_1 + \mu_1 Z_2(Z_5+Z_6) + \mu_2 Z_1 Z_6 + \mu_1 Z_1 Z_2}{(r_{a1}+r_{a2}+(\mu_2+1)Z_1 + (\mu_1+1)Z_2 + \{(r_{a1}+Z_1)(r_{a2}+Z_2) - \mu_1\mu_2 Z_1 Z_2\}/Z_L}\right] - Z_6 \quad (12)$$

and can be infinite by selection of the load impedance.

In general the input impedance decreases with increase of k and increases with reduction of Z_L to the point where it becomes infinite.

The condition for infinite input impedance is usually where $Z_L < r_a$.

If the reactive component of the load fulfils the condition for negative input reactance while the resistive component of the load maintains a satisfactory input resistance then the overall effect on the input circuit is to improve its frequency response, the demand of reactive current by the input circuit now being supplied from the output. A practical example of this technique is given later.

The gain is given by the relation:—

$$A = \frac{V_o}{V_i} = \frac{Z_L}{Z_L + Z_o}\left[\frac{Z_1\{r_{a2} + Z_2 - \mu_2(r_{a1} + \mu_1 Z_2)\} - \mu_2 Z_5(r_{a1} + Z_1 + \mu_1 Z_2)}{(Z_5 + Z_6)\{r_{a1} + Z_1 + r_{a2} + Z_2(\mu_1 + 1)\} + Z_1(r_{a1} + r_{a2} + Z_2 + \mu_1 Z_2 + \mu_2 Z_6)}\right] \quad (13)$$

and in practical cases may be of the order of $\mu/2$ or greater for the zero Z_o condition.

Gain falls with increase of k where $k = \frac{Z_6}{Z_5 + Z_6}$.

The stability against supply-voltage variations, that is, the response at output terminals to a fluctuation at the h.t. terminals, is:—

$$\frac{V_o}{V_r} = \frac{Z_L}{Z_L + Z_o}\left[\frac{(r_{a2}+Z_2)(Z_5+Z_6) - \mu_2 Z_6(r_{a1}+\mu_1 Z_2)}{(r_{a2}+Z_2)(Z_5+Z_6+Z_1)+(r_{a1}+Z_1)(Z_5+Z_6)+r_{a1}Z_1+\mu_1 Z_2(Z_5+Z_6)+\mu_2 Z_1 Z_6+\mu_1 Z_1 Z_2}\right] \quad (14)$$

and is positive when

$$(r_{a2} + Z_2)(Z_5 + Z_6) > \mu_2 Z_6(r_{a1} + \mu_1 Z_2)$$

is zero when

$$(r_{a2} + Z_2)(Z_5 + Z_6) = \mu_2 Z_6(r_{a1} + \mu_1 Z_2)$$

and is negative when

$$(r_{a2} + Z_2)(Z_5 + Z_6) < \mu_2 Z_6(r_{a1} + \mu_1 Z_2).$$

These properties are plotted in Fig. 10 against changes of feedback ratio for a particular practical case using ACM3 valves.

The excitation required at the grid of valve 2 can be derived from the curves if a feedback ratio $\frac{Z_6}{Z_5 + Z_6}$ is assumed.

Thus if Z_6=500 ohms and Z_5=1500 ohms the feedback voltage is 0·25 of the voltage appearing across Z_1. The input generator voltage is approximately $\frac{Z_5}{Z_5 + Z_6}$ of the consequent voltage appearing at the grid.

The gain is computed from the curves in the manner shown in Table 5. If the values Z_5= 1500 Ω, Z_6 = 500 Ω, $r_{a1} = r_{a2}$ = 600 Ω, $Z_1 = Z_2$ = 100 Ω, μ = 14 are substituted in equation (13) we get gain=7·1.

It should be noted that this gain is computed relative to the generator e.m.f. and thus the gain for the two stages will be μ(driver stage valve) × gain as above, and the gain from the grid terminal will be approximately 4/3 times the above or approximately $9\frac{1}{2}$.

If the load is non-linear, as for example when it is a class B stage, the grid current will be supplied according to the load division relation i_{a2}/i_{a1}.

The load sharing of 1 A of non-linear resistive current is illustrated in Fig. 11 on the voltage excursion curves for a particular set of conditions. As before, the excursion loci are shown transposed by biasing to give the maximum possible mean current to deal with the elliptical loci of reactive components of load in order to extend the frequency response at a given amplitude of signal.

Since both valves carry the same mean current, yet both deliver current to the load, the conversion efficiency of such stages is approximately doubled as compared with a stage of comparable performance using valves in parallel.

TABLE 5

i_a	0·4	0·6	0·8	A
$i_a Z_1$	−40	−60	−80	V
Feedback	−10	−15	−20	V
Change of feedback between 0·4 A and 0·6 A, and 0·6 A and 0·8 A	−5		−5	V
ΔV_g required	−50		−40	V
Generator	55 × 4/3 = 73		45 × 4/3 = 60	V
Then from the curves				
V_o	2070+40	1560+60	1110+80	V
ΔV_o	490		440	V
Gain	490/73 6·7		440/60 7·3	

The improved linearity of amplitude response is, in practice, substantial and is achieved without loss of gain.

6. Shunt-regulated Amplifier Type 3(e/e)

The cathode-follower version of the double-shunt regulated pair is equally interesting and is derived as before by adding a further stage of shunt-regulation to the shunt-regulated cathode-follower described. The metamorphosis is shown in Fig. 12(a), (b) and (c).

Circuit analysis provides the following equations which reveal the properties of the circuit:—

$$Z_o = \frac{(r_{a1}+Z_1)(r_{a2}+Z_2)(Z_3+Z_4)+(r_{a1}+Z_1)(r_{a2}Z_2)-\mu_1\mu_2 Z_1 Z_2 Z_4}{(r_{a1}+Z_1)(Z_3+Z_4+r_{a2})+(\mu_1+1)[(r_{a2}+Z_2)(Z_3+Z_4)+r_{a2}Z_2+\mu_2 Z_1(Z_3+Z_4)]-\mu_1 Z_4(r_{a2}+\mu_2 Z_1)} \qquad (15)$$

and this is zero when $(r_{a1}+Z_1)(r_{a2}+Z_2)(Z_3+Z_4)+(r_{a1}+Z_1)r_{a2}Z_2 = \mu_1\mu_2 Z_1 Z_2 Z_4$.

In practice it is usually desirable to make $Z_3 + Z_4$ large compared with r_{a2} and to make r_{a2} large

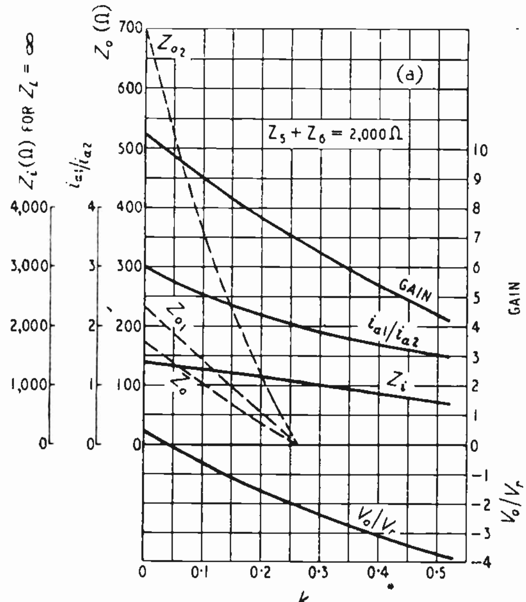
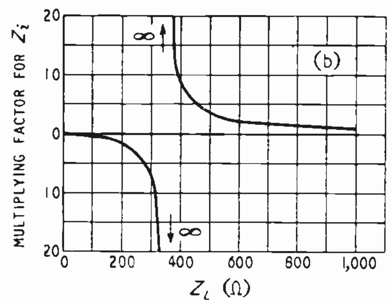
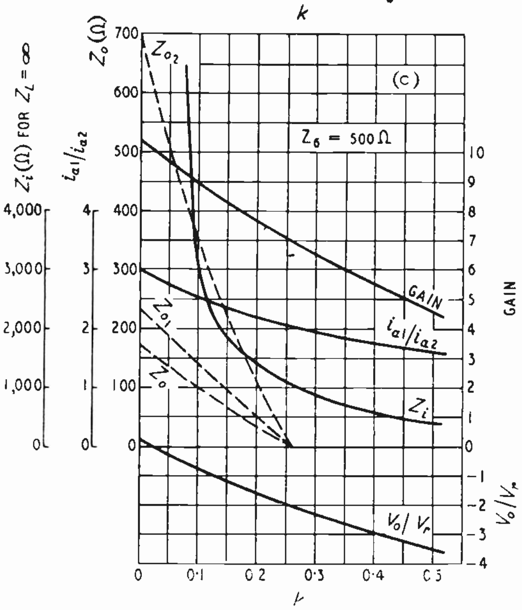
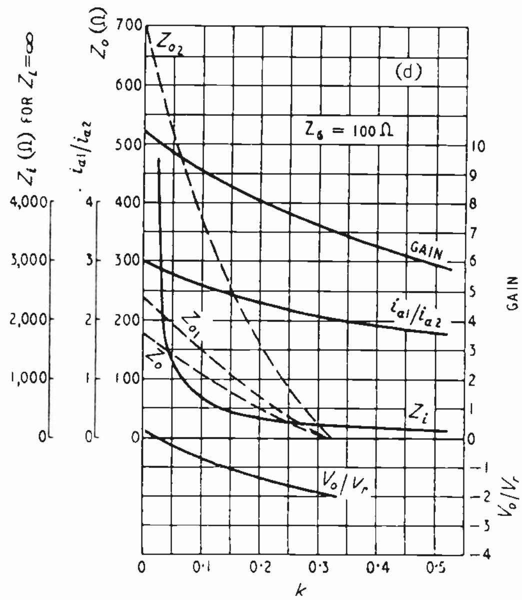


Fig. 10. *Characteristics of shunt-regulated amplifier Type 2(e/e); $\mu = 14$, $r_{a1} = r_{a2} = 600\Omega$, $Z_1 = Z_2 = 100\Omega$, $k = Z_6/(Z_5 + Z_6)$. The performance for three different values of Z_6 is given in (a), (c) and (d), while (b) gives the multiplying factor for a finite load impedance.*







compared with Z_2. In this case the above simplifies to

$$(r_{a1} + Z_1)(r_{a2} + Z_2) \approx \mu_1\mu_2 Z_1 Z_2 k$$

where k is the feedback ratio $\dfrac{Z_4}{Z_3 + Z_4}$.

In applying this circuit, a knowledge of how the load divides between the valves is necessary. If a fluctuation of load current is supplied by the two valves in the ratio i_{a1}/i_{a2} this division of load is given by the relation

$$\frac{i_{a1}}{i_{a2}} = -\frac{(\mu_1 + 1)[(r_{a2} + Z_2)(Z_3 + Z_4) + r_{a2}Z_2] - \mu_1 r_{a2} Z_4}{(r_{a1} + Z_1)(r_{a2} + Z_3 + Z_4) + \mu_2 Z_1 (Z_3 + Z_4)\{1 + \mu_1(1 - k)\}} \quad \ldots \quad (16)$$

the minus sign showing that as i_{a1} increases, i_{a2} decreases.

The gain of the system is given by the expression,

$$A = \frac{(r_{a1} + Z_1)r_{a2} + \mu_1 [(r_{a2} + Z_2)(Z_3 + Z_4) + r_{a2}Z_2 + \mu_2 Z_1 (Z_3 + Z_4)] - \mu_1 Z_4 (\mu_2 Z_1 + r_{a2} + Z_2)}{(r_{a1} + Z_1)(Z_3 + Z_4 + r_{a2}) + (\mu + 1)\{(r_{a2} + Z_2)(Z_3 + Z_4) + r_{a2}Z_2 + \mu_2(Z_3 + Z_4)Z_1\} - \mu_1 Z_4 (r_{a2} + \mu_2 Z_1)} \times \frac{Z_L}{Z_L + Z_o} \quad \ldots \quad (17)$$

which for practical cases (where $Z_3 + Z_4 \gg r_{a2}$; and $r_{a2} \gg Z_2$ and where $Z_L \gg Z_o$)

$$\text{becomes} \approx \frac{(1 - k)\mu}{(1 - k)\mu + 1}$$

$$\text{where } k = \frac{Z_4}{Z_3 + Z_4}.$$

The input impedance is $Z_i = Z'_i - Z_4$ where

$$Z'_i = \frac{Z_L + Z_o}{Z_L}\left[\frac{(r_{a1} + Z_1)(Z_3 + Z_4 + r_{a2}) + (\mu_1 + 1)\{(r_{a2} + Z_2)(Z_3 + Z_4) + r_{a2}Z_2 + \mu_2 Z_1 (Z_3 + Z_4)\} - \mu_1 Z_4 (\mu_2 Z_1 + r_{a2})}{(r_{a1} + Z_1) + (r_{a2} + Z_2) + \mu_1 Z_2 + \mu_2 Z_1 + \{(r_{a1} + Z_1)(r_{a2} + Z_2) - \mu_1\mu_2 Z_1 Z_2\}/Z_L}\right] \quad \ldots \quad (18)$$

and can be simplified by making $Z_3 + Z_4 \gg r_{a2} \gg Z_2$ as before and with $Z_L \gg Z_o$.

$$Z'_i \approx \frac{(\mu + 1)\{r_{a2} + (\mu + 1)Z_1\}(Z_3 + Z_4)}{(r_{a1} + Z_1)(r_{a2} + Z_2)(1 + 1/Z_L) - \mu_1\mu_2 Z_1 Z_2/Z_L}$$

and its dependence on the load is obvious. Values can be chosen to make Z_i infinite.

It is likewise possible and practicable to make the resistive component of input impedance infinite while the reactive component of input impedance is negative. This results in an improvement of response by the same mechanism as was described for Type 2(e/e).

The response at the output terminals to a fluctuation applied at the h.t. supply terminals is given by:—

$$\frac{V_o}{V_r} = \left[\frac{(r_{a2} + Z_2)(Z_3 + Z_4) + \mu_2 Z_1 (Z_3 + Z_4) + r_{a2}Z_2 - \mu_2 (r_{a1} + Z_1)(Z_3 + Z_4) - \mu_1\mu_2 Z_2 Z_4}{(r_{a1} + Z_1)(Z_3 + Z_4 + r_{a2}) + (\mu + 1)\{(r_{a2} + Z_2)(Z_3 + Z_4) + r_{a2}Z_2 + \mu_2 Z_i (Z_3 + Z_4) - \mu_1 Z_4\}(r_{a2} + \mu_2 Z_1)}\right]\frac{Z_L}{Z_L + Z_o} \quad (19)$$

and for most practical cases where low output impedance is required this ratio is not small but is between ½ and 1. What is interesting, however, is that the ratio is negative; i.e., the fluctuation at the supply terminal appears at the output terminal in the opposite phase. If a driving amplifier is supplied from the same d.c. source, complete cancellation of the supply fluctuation is possible.

It is interesting to apply these formulae to a practical case. Assume the two valves of a pair are ACM3 valves with $\mu_1 = \mu_2 = 14$, $r_{a1} = r_{a2} = 600\,\Omega$. Then for zero output impedance $Z_4 = 520\,\Omega$ and $Z_3 = 1480\,\Omega$. Substituting in the gain equation gives gain $= 0{\cdot}88$, while the approximate formula give $0{\cdot}91$.

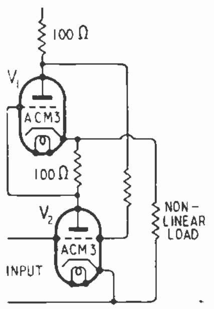
The division of load is given by equation (16) and is $0{\cdot}54$, Z_1 and Z_2 being each $100\,\Omega$.

For $Z_L = \infty$, the input impedance is $Z_i = 12{,}100\,\Omega$, and to obtain an input impedance $Z_i = \infty$, the load impedance must be $Z_L = 350\,\Omega$. As for the other cases, a simple graphical analysis is possible, from which it is possible to confirm the properties arrived at in the theoretical analysis in a manner similar to that already illustrated for the other types.

These properties are plotted in Fig. 13 against changes of feedback ratio for a particular practical case using ACM3 valves.

7. Combinations of Shunt-regulated Amplifiers and Cathode-followers

An examination of the formulae derived for the response of these systems to supply fluctuations shows that the amplifier and cathode-follower versions reproduce the supply fluctuations in opposite phases at reduced level. It is thus possible, in practice, to proportion the operating conditions so that a shunt-regulated amplifier, in combination with a shunt-regulated cathode-follower, will attenuate supply fluctuations to zero at the output terminal. Even if this precise adjustment is not made, the reduction of the effect of supply fluctuations is sufficient to reduce the degree of supply-voltage stabilization required.



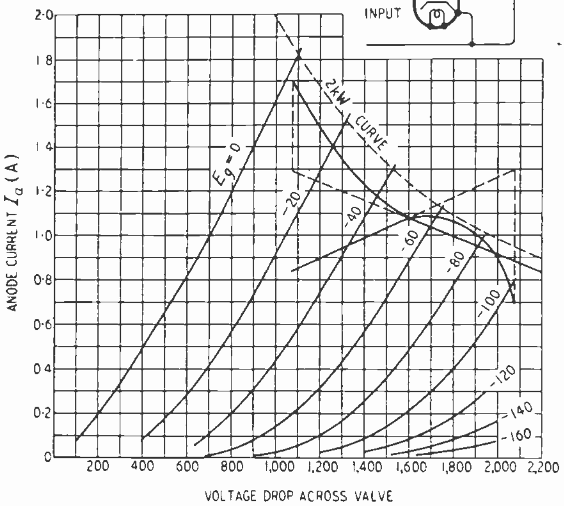


Fig. 11. *This diagram illustrates the load-sharing properties of the Type 2(e/e) regulator.*

Applying the numerical values to r_{a1}, μ_1, Z_1, etc., used to illustrate the properties, we find the following responses to supply fluctuations.

Type	Response
2(e)	0·26
3(e)	−0·25

Therefore, a combination of a shunt-regulated amplifier and a shunt-regulated cathode-follower will, with practical values of regulating impedance, give almost complete isolation from supply fluctuations.

In the case of Types 2(e/e) and 3(e/e) shunt-regulated amplifiers in combination, the balancing out of supply-voltage fluctuations is also possible and the necessary conditions can be read from the curves of Figs. 10(a) and 13(a).

8. Practical Applications

As a demonstration of the impedance inverting properties of the shunt-regulated amplifier Fig. 14 shows the response of the arrangement of Fig. 15 for different values of capacitance added across the output terminals.

The circuit shown is exactly that of the experimental model and does not represent a good finished design. It will be noted that the feedback from the lower output valve has been taken via the amplifier-coupling circuit for practical convenience and that no attempt has been made to deal with low-frequency response, the sole purpose of the experiment being to verify the impedance inverting property at the higher frequencies. It should also be noted that no attempt has been made in the circuit to achieve either good intrinsic

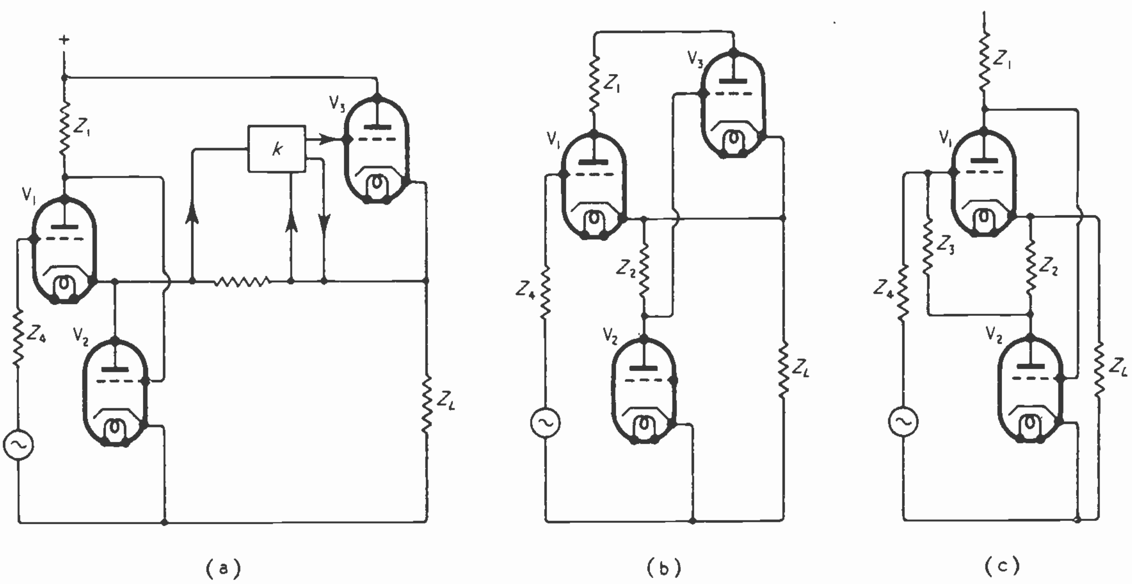
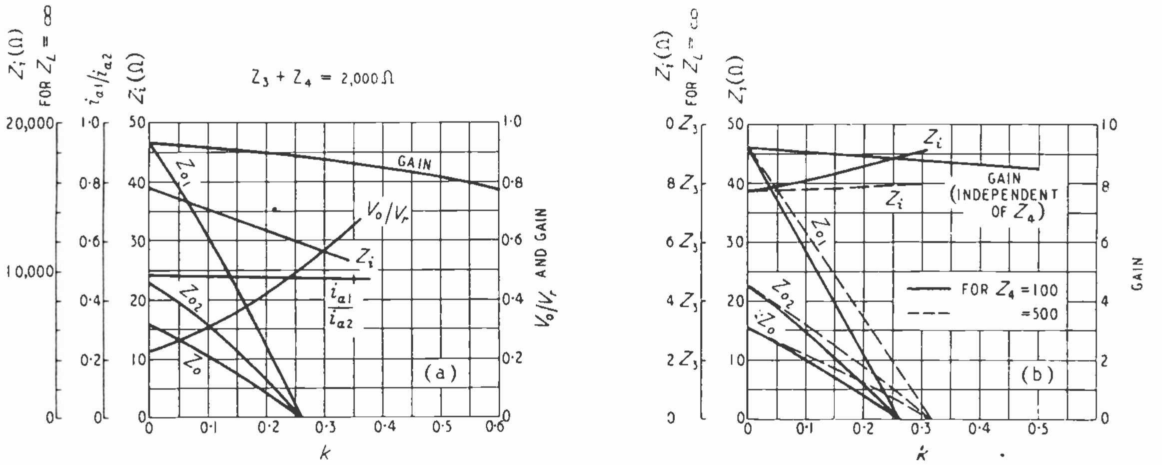


Fig. 12. *Derivation of shunt-regulated cathode-follower Type* 3(*e/e*). *It comprises a shunt-regulated cathode-follower* 3(*e*) *plus a regulator* (*e*) *of Table* 1. *The third valve in* (*b*) *can be replaced by a feedback network* (*c*).



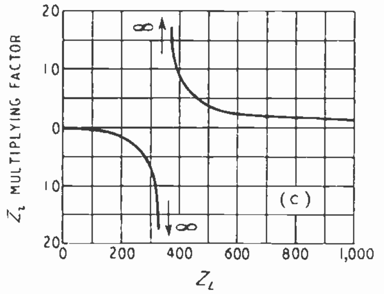


Fig. 13. *Characteristics of shunt-regulated cathode-follower Type* 3(*e*/*e*); $\mu = 14$, $r_{a1} = r_{a2} = 600\Omega$, $Z_1 = Z_2 = 100\Omega$. *The performance with different values of* Z_4 *is given in* (*a*) *and* (*c*) *with the input-impedance multiplying factor in* (*b*).

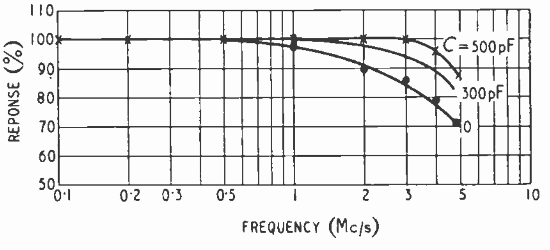


Fig. 14 (right). *Effect of load capacitance on the frequency response of the shunt-regulated amplifier of Fig.* 15.

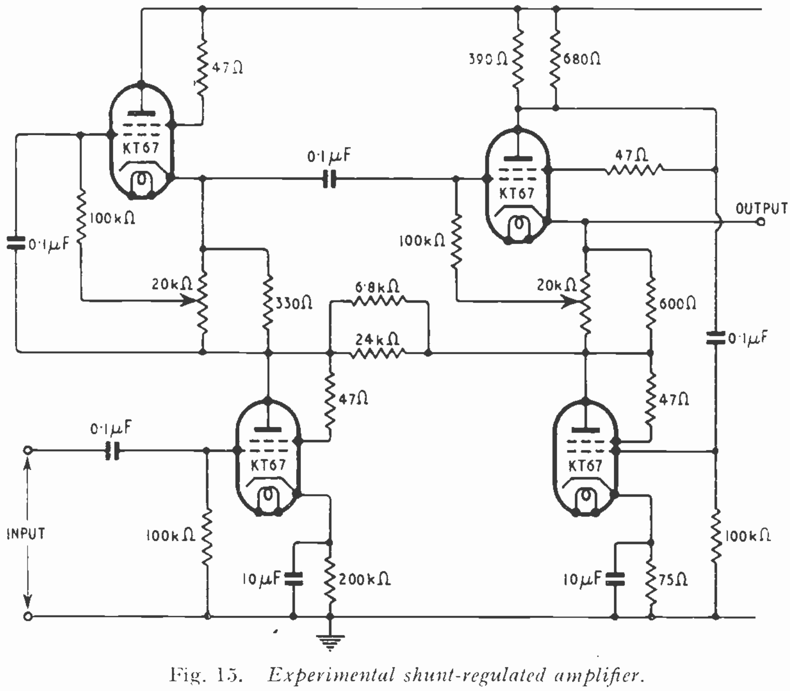


Fig. 15. *Experimental shunt-regulated amplifier.*

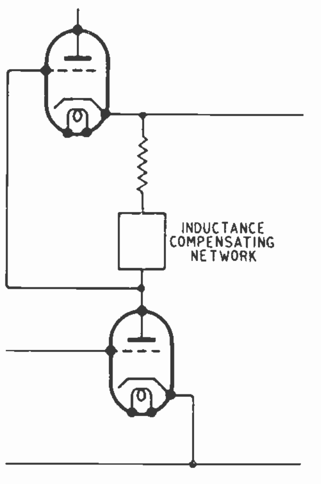


Fig. 16 (above). *Method of inductance compensating the feedback resistor.*

frequency response or good response on the feedback loop. It was considered satisfactory, however, that this test under these conditions adequately verified the theory.

A combination of the techniques described in this article has been incorporated at low and high levels in a recently completed modulator for a 50-kW television transmitter and a comparison of its performance with the estimated performance of a conventional modulator is made in Table 6.

TABLE 6

	Conventional (Estimated)	Shunt Regulated (Measured)
Total Number of ACM3 valves (including stabilization)	14	11
Frequency response	± 5% to 4 Mc/s	± 5% to 7 Mc/s
Total power input including h.t. and fil.	90 kW	30 kW
Rise time of output step with 0·07-μsec rise input step	0·14 μsec 5% overshoot	0·09 μsec* no overshoot
Max. voltage swing	1250	1250
Non-linearity for transmitter grid current of 3 A	7-10%	<1%

* On the assumption that the input step is exponential this gives an indicial response of <0·05 μsec.

9. Frequency Compensation in Shunt-regulated Amplifiers

This is outside the scope of the present introductory article, but as an indication of what can be done with inductance compensating the feedback resistor (Fig. 16), the change of response is shown in Fig. 17 for compensation in this one circuit for maximal flatness. Further compensation is possible by addition of series inductances as shown in Fig. 18.

The natural frequency response of the Type 2(e) shunt-regulated amplifier is typified by curve A, Fig. 17, and is not a natural RC law.

An inductance included with the coupling resistor corrects the law and produces, for all practical purposes, a natural RC law. This is shown in curve B. This first degree of compensation removes substantially all the load capacitance from the input valve, the RC law being determined by the inherent capacitance in the circuit. Further compensating networks on the same two-terminal systems produces the curve C, Fig. 17.

With inductance compensation taken to the limit, the 50-kW modulator quoted had in its laboratory form a response flat ±5% to 11 Mc/s.

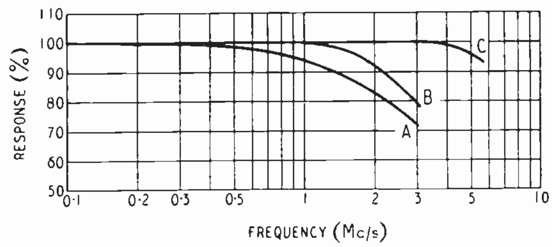


Fig. 17 (above). *Frequency response of shunt-regulated amplifier with one stage of inductance compensation. Curve A is the natural response of the Type 2(e) circuit while curve B shows the effect of adding one stage of inductance compensation. The effect of further compensation is shown by curve C.*

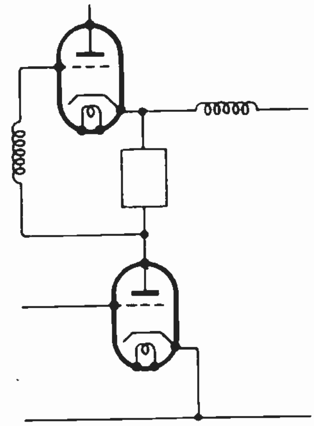


Fig. 18 (right). *Further inductance compensation can be applied in the manner shown here.*

Acknowledgment

The author is indebted to Messrs. Marconi's Wireless Telegraph Company, Ltd., for permission to publish this paper.

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[2] British Patent No. 526418; N. H. Clough.
A search of the literature shows that amplifier arrangements have been used with valves in series d.c. connection. The following references are given.
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INTERELECTRODE IMPEDANCES IN TRIODES AND PENTODES

By E. E. Zepler, Ph.D., M.Brit.I.R.E. and S. S. Srivastava, M.Sc.

THE grid-cathode capacitance in triodes or pentodes is known to increase when the valves become conducting. The effect, which is connected in a fairly complicated way with the mutual conductance and with the transit times cathode to grid and grid to anode or grid to screen grid respectively, has been discussed in several previous papers.[1,2,3,4] It appears that no formula given is in satisfactory agreement with observed facts. In the following, measurements of capacitance and conductance between electrodes in a triode and in a pentode are described, and a theory is suggested which tends to explain some of the apparent discrepancies.

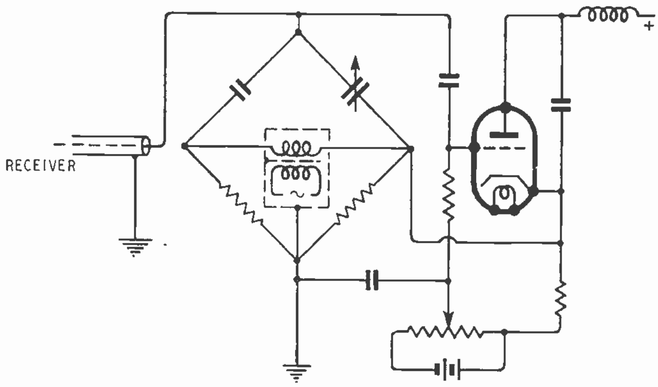

Fig. 1. *Method of measuring* $\delta(C_{gk} + C_{ga})$.

Method of Measurement

Measurements were carried out at 1 Mc/s and 32 Mc/s. At 1 Mc/s conductances between grid and cathode were negligible with the values of grid bias chosen, and a modified type of Wheatstone bridge described in a previous paper was employed (Fig. 1). For the sake of simplicity $(C_{gk} + C_{ga})$ was measured, which had been found to differ by only $1/\mu$ of its value from C_{gk}[1]. At 32 Mc/s a twin-T bridge was used (Fig. 2). Experimental difficulties in the latter case were surprisingly few, no doubt due to the earth being common to signal generator and receiver. The capacitances measured were the same at 1 Mc/s and 32 Mc/s within the accuracy of measurement.

The results obtained are shown in Figs. 3 and 4. In order to isolate the effects due to transit time it is necessary to find the influence of feedback on the grid-cathode impedance. It is well known that, at 32 Mc/s, the effect of feedback on the grid-cathode capacitance is negligibly small when anode and cathode are short-circuited by a capacitor, but the grid-cathode conductance can be very much affected. The conditions may be seen from Fig. 5 which shows a sketch of the leads going from the electrodes to the valve base. When a voltage E is applied between grid and cathode, an anode current approximately equal to $g_m E$ is produced which induces a voltage $g_m E j\omega M_1$ in the grid circuit, M_1 being the mutual inductance between the anode loop *ak* and the grid loop *gk*. This causes an input conductance $G_1 = \omega^2 g_m C_{gk} M_1$. Usually this formula is given in the form $G_1 = \omega^2 g_m C_{gk} L$, where L is the inductance of the cathode lead. It should be realized, however, that the inductance of the cathode lead has no real significance and that the relevant factor is the mutual inductance of the two loops.[5]

Similarly, the capacitance from grid to anode is the cause of a negative input conductance of magnitude $G_2 = \omega^2 g_m C_{ga} M_2$, M_2 being the mutual inductance between the loops *ak* and *ag*. In order to find the mutual inductances required a valve was broken and, at the points where the leads joined the electrodes, cathode, grid and anode were connected by a metal strip. The mutual inductances of the loops concerned could now easily be measured. This procedure seems justified since it may be assumed that the magnetic fields from currents in and between the electrodes are negligible. Furthermore, the magnitudes of C_{gk} and C_{ga} were measured and thus, for given values of g_m, the input conductances due to feedback could be calculated. An experimental check on these calculations was obtained by inserting a small known inductance of 0·036 microhenry subsequently in the cathode and anode leads and measuring the change in input conductance. The known inductance consisted of two parallel wires shorted at the end and at right angles to the valve leads. Insertion of this inductance for fixed values of g_m but different values of V_a showed, at the same time, that the value of C_{gk} to be inserted in the formula is that of the 'hot' capacitance; i.e., the cold capacitance plus the increase shown in Fig. 3. The results derived showed that the positive feedback through C_{ga} slightly exceeded the negative feedback through C_{gk}.

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The corrected curves showing the input conductance due to transit time only are given in Fig. 6.

It is of interest to compare the results of Figs. 3 and 6 with the formulae given by North.* We neglect, for the sake of simplicity, terms which are of second order under the conditions experienced here; then

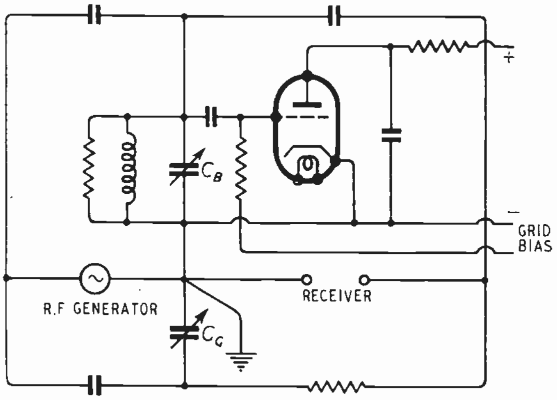

Fig. 2. *Twin-T circuit for determining* $\delta(C_{gk} + C_{ga})$ *and* $\delta(G_{gk} + G_{ga})$.

$$\delta C = \delta(C_{gk} + C_{ga}) = \frac{C_{gk}(cold)}{3} + \frac{2}{3} g_m \tau_2$$

$$G = G_{gk} + G_{ga} = \frac{g_m}{180}\,(9\omega^2\tau_1{}^2 + 44\omega^2\tau_1\tau_2)$$

These formulae apply to plane parallel electrodes and are valid only under conditions of space charge limitation, when i_a is proportional to $V^{3/2}$, and where V is approximately equal to $V_g + V_a/\mu$. The transit time from cathode to grid is τ_1, and that from grid to anode (or screen grid respectively) is τ_2. Rearranging the second expression and again neglecting second-order terms we obtain

$$G = 2g_m f^2 \tau_1{}^2 (1 + 3{\cdot}3\sqrt{V/V_a})$$

The cathode of the 6J5 is a circular cylinder while grid and anode are of a flat oval shape. We may assume that the bulk of the emission takes place where the cathode is nearest the anode so that the assumption of plane parallel electrodes is a fair approximation. The formulae take no account, however, of the fact that the conditions alter appreciably along the surface of the cathode in the direction of its axis. Due to the shielding effect of the grid wire μ varies very much along the surface of the cathode. Measurements gave a μ of 20 for points midway between the grid wires and more than 60 for points in line with the wires.[6] The importance of this may be seen from an example. With g_m=2 mA/V we consider two conditions under which measurements were taken; viz., $V_a = 43$ V and $V_a = 215$ V. In the first case V_g was $-0{\cdot}6$ V, in the second case it was $-8{\cdot}5$ V. If we disregard contact potential we see that with $V_a = 43$ V practically the whole of the cathode emits, while with V_a=215 V emission takes place only at those points where μ<25; i.e., only at points not shaded by the grid wires. This explains, in our opinion, the great differences in δC and G, for fixed values of g_m but different supply voltages, as shown in Figs. 3 and 6.

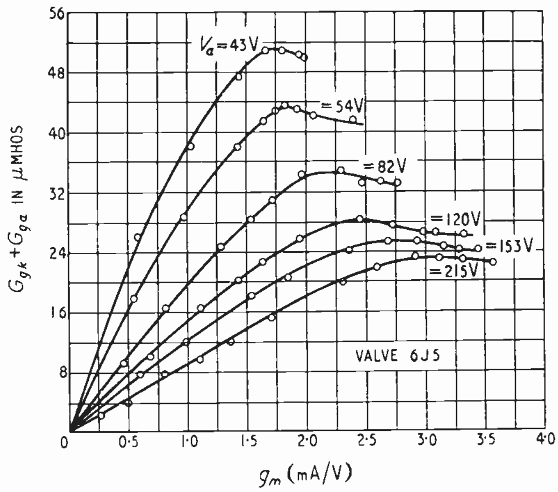

Fig. 4. *Variation of conductance with mutual conductance for a triode.*

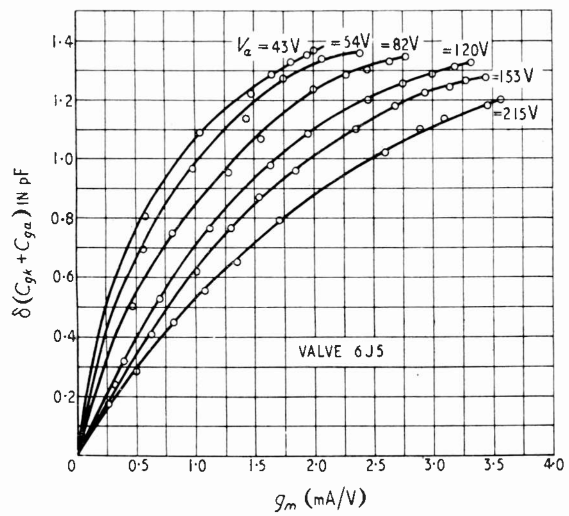

Fig. 3. *Variation of capacitance with mutual conductance for a triode.*

If we try to explain these differences from the different transit times between grid and anode (according to theory τ_1 is given if g_m is fixed), the

* For this reference we are indebted to R. H. Booth, *Wireless Engineer*, June 1949, p. 211.

ruled out. The anode was connected with the suppressor grid, and all the other electrodes were for r.f. at screen-grid potential (Fig. 10). Then the screen grid may be considered as a source of electrons the velocity of which is determined by the screen-grid potential. Since under normal conditions the anode current of a pentode is 70-80% of the total current we may assume that a similar fraction of the emission current enters the space between screen and suppressor grid before eventually reaching the screen grid. Some of the results obtained are shown in Fig. 11.

For an approximate theoretical derivation we neglect interaction between the electrons and assume plane parallel electrodes. The screen grid is the zero plane in the space between screen and suppressor grid and x is the distance from this plane. V_1 is the potential of the screen grid and V_2 that of the suppressor grid. The electrons leave the zero plane with velocity $v_0 = \sqrt{(2eV_1/m)}$, where e and m are charge and mass of the electrons respectively. From the equation for the movement of an electron $mdv/dt = -eE$ we obtain $v = v_0 - eEt/m$ and $x = v_0 t - eEt^2/2m$ if $x=0$ at $t=0$.

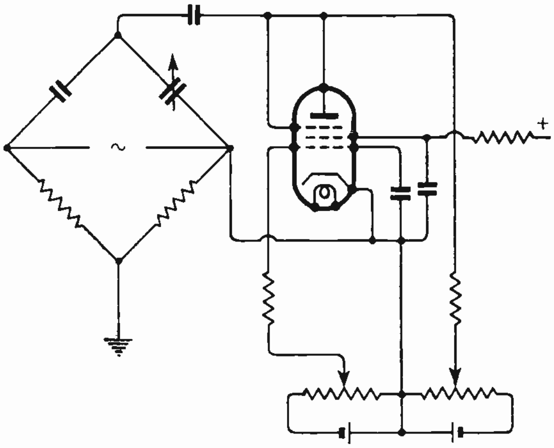

Fig. 10. *Method of measuring suppressor-grid to screen-grid capacitance.*

Substituting t, we obtain between v and x the relation $v = \pm\sqrt{(v_0^2 - 2eEx/m)}$. The maximum distance from the screen grid reached by the electrons is $x_0 = mv_0^2/2eE$. If i is the current density between screen grid and suppressor grid due to electrons travelling in one direction, the charge density σ is $\dfrac{i}{v} = \dfrac{i}{\sqrt{(v_0^2 - 2eEx/m)}}$. The charge induced on the suppressor grid by the electrons in the space between screen and suppressor grid is found by integration, using the fact known from image theory that a charge q, between two parallel plates of distance D, and at distance x from plate 1, induces on plate 1 a charge $q(D-x)/D$ and on plate 2 the charge qx/D. Hence the total charge induced on the screen grid is

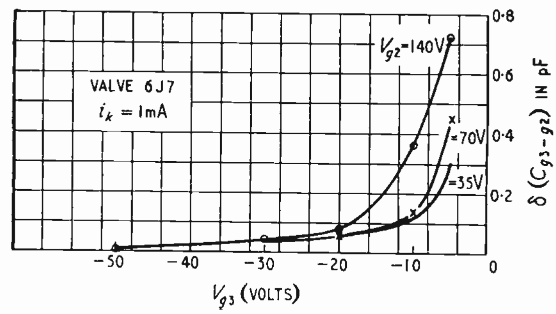

Fig. 11. *Variation of suppressor-screen capacitance with suppressor bias.*

$$\frac{1}{D}\int_0^{x\,(max)} \sigma x dx = \frac{i}{D}\int_0^{\frac{mv_0^2}{2eE}} \frac{xdx}{\sqrt{v_0^2 - \dfrac{2eEx}{m}}}$$

$$= 0{\cdot}95\, iD\left(\frac{m}{e}\right)^{1/2} \frac{V_1^{3/2}}{(V_1 - V_2)^2}$$

$$\therefore\ C = \frac{dq}{dV_2} = 1{\cdot}9\, iD\left(\frac{m}{e}\right)^{1/2} \frac{V_1^{3/2}}{(V_1 - V_2)^3},\ \text{all}$$

quantities being in e.m.u., or

$$C\ (\text{in pF}) = 0{\cdot}6 \times 10^9 iD\left(\frac{m}{e}\right)^{1/2} \frac{V_1^{3/2}}{(V_1 - V_2)^3},\ \text{where}$$

i is in amperes, V in volts, e in coulombs, m in grammes and D in cm. Substituting the values $V_1 = 70$ V, $V_2 = -30$ V, $i = 1$ mA, $D = 0{\cdot}25$ cm gives $C = 0{\cdot}67 \times 10^{-2}$pF. Comparison with Fig. 11 shows that the agreement with practical results is not very good. While the approximations made might account for the differences with very negative suppressor grid, the formula fails to explain the steep rise in capacitance on approaching the cut-off point and the increase in capacitance with increasing screen-grid voltage, as shown in Fig. 11. It is intended to carry on the investigations with the view to elucidating the causes for these discrepancies.

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HIGH-GAIN MAGNETIC AMPLIFIER

Theory of the Self-excited Transductor

By R. Feinberg, Dr. Ing.

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SUMMARY.—Self-excitation is an effective method of obtaining feedback in a transductor. A self-excited transductor may be designed to give a high value of current amplification by making the number of turns, $2N_f$, of the self-excitation winding sufficiently large in relation to the number of turns, $2N_l$, of the load winding. Transductor operation is stable, and thus suitable for amplifier action, when $N_f < N_{f,crit}$, but unstable and, therefore, unsuitable for amplifier action when $N_f \geqslant N_{f,crit}$ where, to a first approximation, $N_{f,crit} = \frac{1}{2}N_l$ for the parallel-transductor and $N_{f,crit} = N_l$ for the series-transductor. A self-excited transductor with $N_f \geqslant N_{f,crit}$ may be used to function like an on-off trigger relay.

SELF-EXCITED transductors are used in practice in high-gain magnetic amplifiers. The ordinary transductor was discussed in a previous paper[1]. In this paper the analysis is extended to the transductor with self-excitation.

Circuit Arrangement

The principle of a circuit with self-excited transductor is shown in Fig. 1. The a.c. power supply 1 provides a current i_l to the load 2 connected in series with the load winding 3a of the self-excited transductor 3 and a four-arm bridge rectifier 4.

The self-excited transductor[2] is a modification of the ordinary transductor[1]. A further winding, the self-excitation winding 3c (Fig. 1), is added to magnetize the transductor cores analogous to the control winding 3b. The self-excitation winding is magnetically coupled with the control winding but has, like the control winding, no coupling with the load winding.

The rectified current i_f of the load circuit passes from the d.c. terminals of rectifier 4 through the self-excitation winding 3c of the transductor, see Fig. 1. Magnetic excitation of the transductor cores by i_f either aids the control current i_c, thus giving positive feedback, or opposes it, in that case giving negative feedback.

Calculating the Transductor Characteristic

Calculation is based on Fig. 2(a) for the parallel-transductor and on Fig. 2(b) for the series-transductor. The assumptions for calculation are (a) zero load on the transductor (i.e., zero impedance in the load circuit), (b) zero impedance in the control circuit, (c) ideal transductor properties (i.e., a transductor without leakage flux and no iron nor copper losses), (d) ideal rectifier characteristic (i.e., zero resistance in the forward direction and infinite resistance in the reverse direction), and (e) sinusoidal a.c. supply voltage v.

The voltage drop across the combined self-excitation coils is zero because they and the control coils are tightly coupled, with zero impedance in the control circuit. Let the a.c. supply voltage, v in Fig. 2, for convenience be represented by the expression

$$v = -V\sqrt{2}\sin\omega t \qquad (1)$$

Because of zero voltage drop across the combined self-excitation coils and across the rectifier the instantaneous values of magnetic flux density, b_1 and b_2, in core 1 and core 2 of the transductors (Fig. 2) are, with the respective notation of Fig. 3 and of equation (1),

$$b_1 = B_0 + B\sqrt{2}\cos\omega t \qquad (2)$$

$$\text{and } b_2 = -B_0 + B\sqrt{2}\cos\omega t \qquad (3)$$

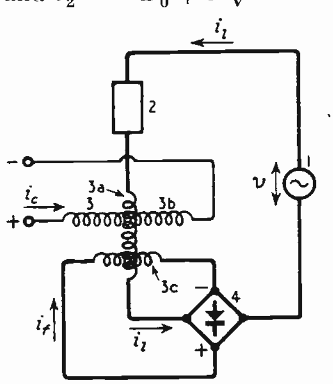

Fig. 1. *Principle of a high-gain magnetic amplifier with self-excited transductor.* 1. *a.c. power supply;* 2. *load;* 3. *transductor;* (*a*), *load winding,* (*b*), *control winding,* (*c*), *self-excitation winding;* 4. *feedback rectifier.*

where B_0 and $-B_0$, respectively, denote the mean values of magnetic flux density in the cores 1 and 2 and $B\sqrt{2}$ the peak value of the alternating components, see Fig. 3.

Assuming uniform cross-section of the transductor cores the magnetizing forces, h_1 and h_2, in each core are expressed, according to Fig. 3, by the respective relations

$$h_1 = H_0 + \sum_{n=1}^{\infty} H_n\sqrt{2}\cos n\omega t \qquad (4)$$

$$\text{and } h_2 = -H_0 - \sum_{n=1}^{\infty} H_n\sqrt{2}\cos n(\omega t - \pi) \qquad (5)$$

Let l denote the effective length of magnetic path

* Now with Ferranti Ltd., Manchester.

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in a core. Thus, the ampere-turns, θ_1 and θ_2, corresponding to h_1 and h_2, respectively, are for core 1, with the arrow direction of excitation adopted in Fig. 2,

$$\theta_1 = h_1 \cdot l = \Theta_0 + \theta_{1p} \quad \ldots \quad \ldots \quad \ldots \quad (6)$$

and for core 2, with the arrow direction of excitation in Fig. 2,

we have $I_l \equiv I_f$ and, therefore, from equation (10)

$$I_c N_c = H_0 . l - I_l N_f \quad \ldots \quad \ldots \quad \ldots \quad (11)$$

To calculate I_l let us first consider the waveform of i_l. The waveform of i_l is unaffected by self-excitation, see the oscillograms of i_l and of i_f in Fig. 4 for the parallel-transductor [Fig. 2 (a)], and in Fig. 5 for the series-transuctor [Fig.

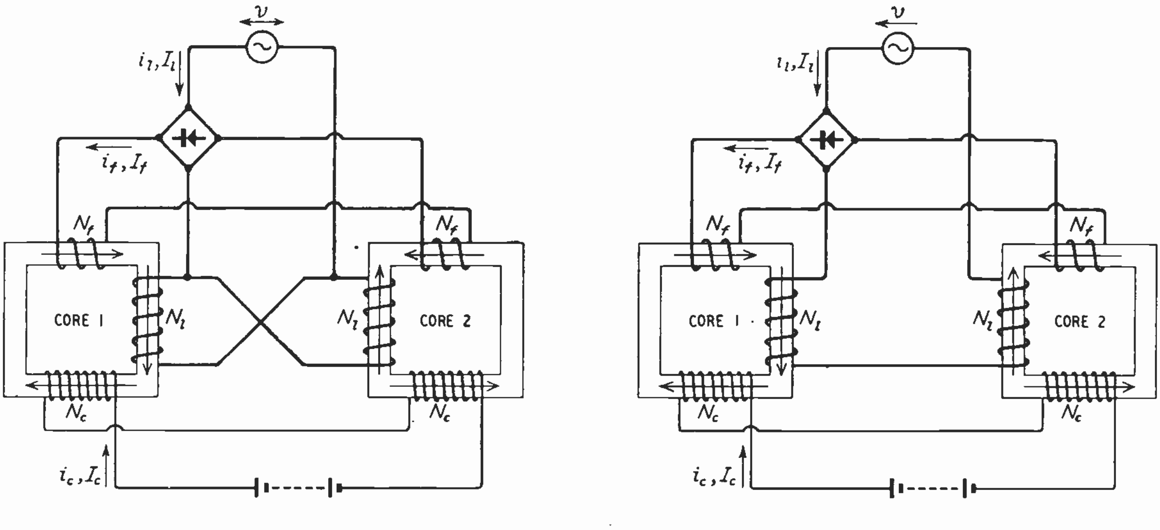


Fig. 2. *Transductor arrangement for calculation; (a) parallel-transductor, (b) series-transductor.*

$$\theta_2 = h_2 \cdot l = -\Theta_0 + \theta_{2p} \quad \ldots \quad \ldots \quad (7)$$

$$\text{where } \Theta_0 = H_0 . l \quad \ldots \quad \ldots \quad \ldots \quad (8)$$

means the d.c. component of ampere-turns in each core, $\theta_{1p} = l \sum_{n=1}^{\infty} H_n \sqrt{2} \cos n\omega t$ the a.c. component in core 1 and $\theta_{2p} = -l \sum_{n=1}^{\infty} H_n \sqrt{2} \cos n(\omega t - \pi)$ the a.c. component in core 2.

Let I_c be the mean value of current in the control winding and I_f the mean value of current in the self-excitation winding. Then, with the notation of Fig. 2,

$$\Theta_0 = I_c N_c + I_f N_f \quad \ldots \quad \ldots \quad \ldots \quad (9)$$

where N_c is the number of turns of a control coil and N_f the number of turns of a self-excitation coil. From equation (9), with equation (8),

$$I_c N_c = \Theta_0 - I_f N_f = H_0 l - I_f N_f \quad \ldots \quad (10)$$

In transductor calculation it is usually convenient to express the magnitude of load current, I_l, not as r.m.s. value but as the mean value of the rectified form of i_l. Since i_f is the rectified form of i_l (Figs. 1 and 2) and $I_f = \frac{1}{\pi} \int_0^{\pi} i_f \, d\omega t$,

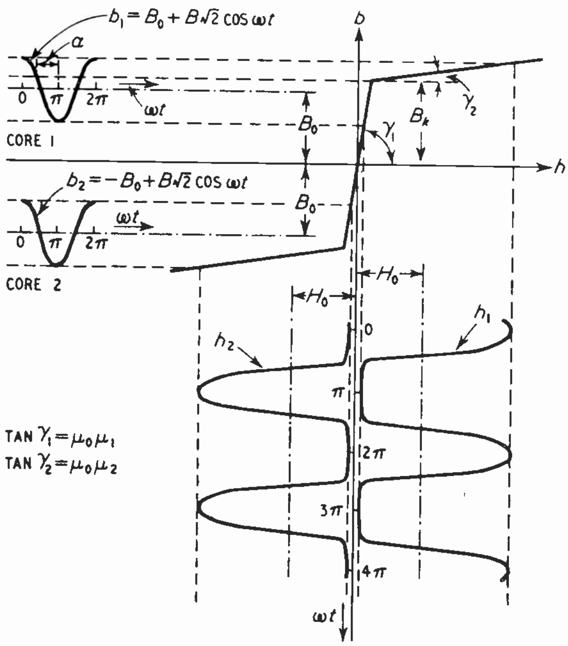
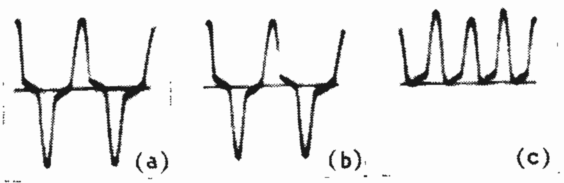


Fig. 3. *Magnetic flux densities and magnetizing forces in the transductor cores with a sinusoidal transductor voltage.*

2 (b)]. (The slight asymmetry of i_f in the oscillograms is caused by some asymmetry of the coils of the experimental transductor.) The mathematical expression for I_l is, therefore, independent of any effect from self-excitation in the transductor.

The mathematical procedure for calculating I_l and H_0 is analogous to the calculation of the



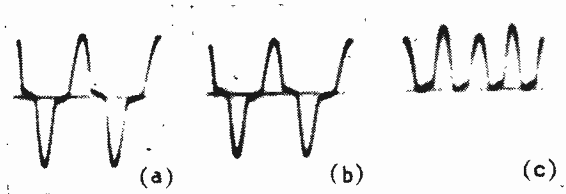


Fig. 4 (above). *Oscillograms of currents in a parallel-transductor,* Fig. 2 (a).

Fig. 5 (below). *Oscillograms of currents in series-transductor,* Fig. 2 (b).

(a) *load current* i_1, *in transductor without self-excitation;*
(b) *load current* i_1, *in transductor with self-excitation;*
(c) *feedback current* i_f, *in transductor with self-excitation.*

corresponding values for the ordinary transductor[1]. Thus we have with the notation of Fig. 3

$$H_0 = \frac{1}{2\pi}\int_0^{2\pi} h_l \, d\omega t$$

$$= \frac{1}{\mu_0\mu_2}\left[B\sqrt{2}\left(\frac{\pi-\alpha}{\pi}\cos\alpha + \frac{\sin\alpha}{\pi}\right) + \frac{\mu_2}{\mu_1}\left(B_k + B\sqrt{2}\frac{\alpha\cos\alpha - \sin\alpha}{\pi}\right)\right] \quad (12)$$

and $$I_l = k_i\frac{l}{N_l}\,\frac{2\sqrt{2}}{\pi}\left(H_1 - \frac{1}{3}H_3 + \frac{1}{5}H_5 - + \ldots\right) \quad (13)$$

where for the parallel-transductor, Fig. 2(a),

$$k_i = 2 \quad \ldots \quad (14)$$

and for the series-transductor, Fig. 2(b),

$$k_i = 1 \quad \ldots \quad (15)$$

For H_n (n=1, 3, 5 . . . .) we have, with the notation of Fig. 3,

$$H_n\sqrt{2} = \frac{1}{\pi}\int_0^{2\pi} h_1 \cos n\omega t \, d\omega t \quad \ldots \quad (16)$$

and consequently, for $n = 1$,

$$H_1 = \frac{B}{\mu_0\mu_2}\left[1 - \frac{2\alpha - \sin 2\alpha}{2\pi}\left(1 - \frac{\mu_2}{\mu_1}\right)\right] \quad (17)$$

and for $n = 3, 5, 7 \ldots.$

$$H_n = \frac{B}{\mu_0\mu_2}\,\frac{2^n\cos n\alpha \, . \, \sin\alpha - 2\sin n\alpha \, . \, \cos\alpha}{n(n^2-1)\pi}\left(1 - \frac{\mu_2}{\mu_1}\right) \quad \ldots \quad (18)$$

The auxiliary quantity α is defined, see Fig. 3, by the relation

$$\cos\alpha = \frac{B_0 - B_k}{B\sqrt{2}} \quad \ldots \quad (19)$$

where B_k denotes the magnetic flux density at the knee of the magnetization characteristic.

For graphical presentation of the transductor characteristic it is convenient to introduce the dimensionless quantity y, representing the load ampere-turns,

$$y = \frac{1}{k_i}\frac{\pi}{2}\frac{\mu_0\mu_2}{B_k l} I_l N_l \quad \ldots \quad (20)$$

and the dimensionless quantity x, representing the control ampere-turns,

$$x = \frac{\mu_0\mu_2}{B_k l} I_c N_c \quad \ldots \quad (21)$$

or, with equations (11) and (20),

$$x = \frac{\mu_0\mu_2}{B_k} H_0 - k_i\frac{2}{\pi}\frac{N_f}{N_l} y \quad \ldots \quad (22)$$

The quantity $\mu_0 = 0.4\pi \times 10^{-6}$ henry/metre [see equations (12), (17), (18), (20), (21), (22)] denotes the absolute permeability of free space expressed in the rationalized form of the m.k.s. system of units, $\mu_1 = (\tan\gamma_1)/\mu_0$ [see equations (12), (17), (18)] is the relative permeability of the high-permeability region of the magnetization characteristic, and $\mu_2 = (\tan\gamma_2)/\mu_0$ [see equations (12), (17), (18), (20), (21), (22)] the relative incremental permeability of the saturation region (see Fig. 3).

The value of B [see equations (2), (3), (12), (17), (18), (19) and Fig. 3] is related to the r.m.s. value V of voltage across the transductor [see equation (1)] by the expression

$$B = k_v\frac{V}{\omega A N_l} \quad \ldots \quad (23)$$

where, for the parallel-transductor, Fig. 2(a),

$$k_v = 1 \quad \ldots \quad (24)$$

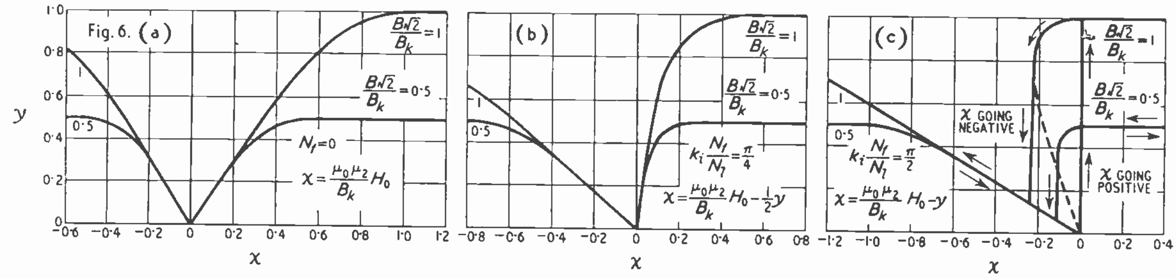
and for the series-transductor, Fig. 2(b),

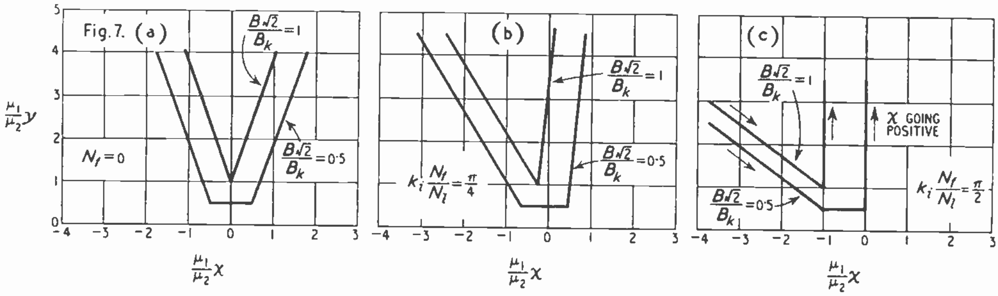
$$k_v = 0{\cdot}5 \quad \ldots \quad (25)$$

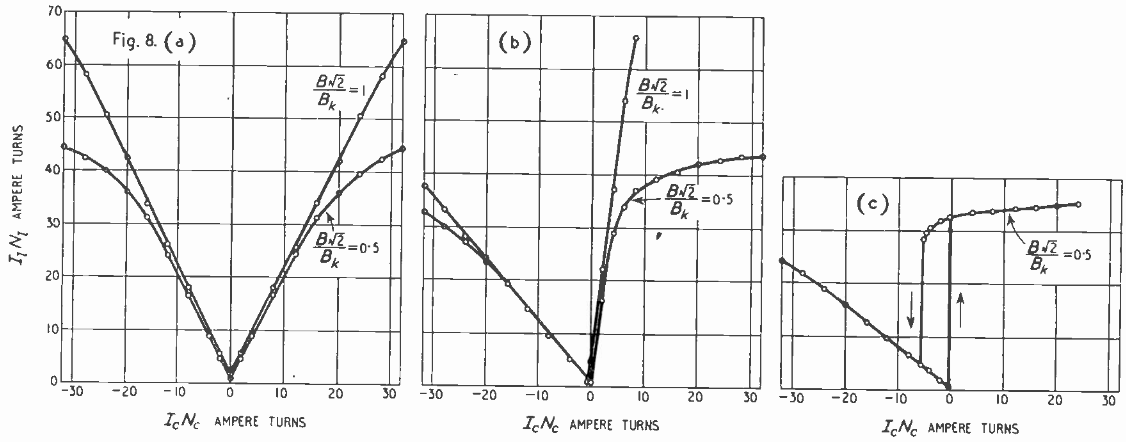
A denotes the area of cross section of a core, N_l the number of turns of a load coil, and ω the angular frequency of the a.c. power supply.

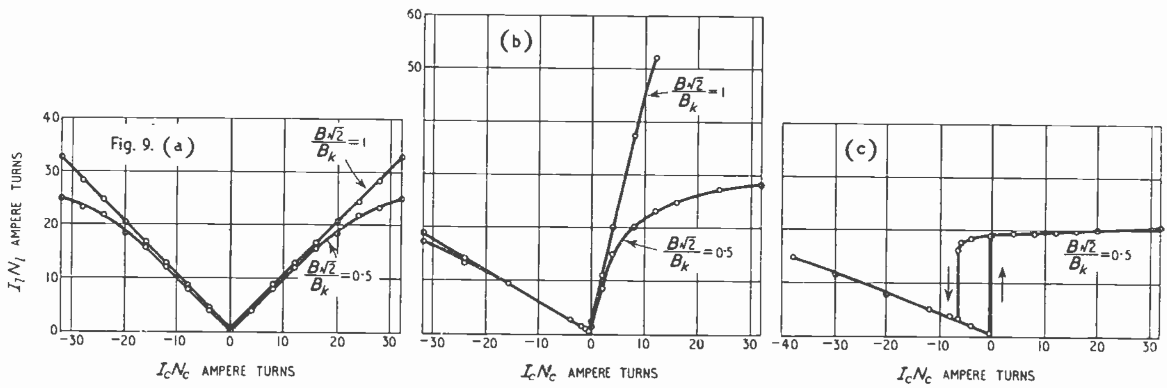
The Transductor Characteristics

A transductor characteristic is calculated with x [equation (22)] as the abscissa and y [equation (20)] as the ordinate, the values of H_0 and I_l









Figs. 6 & 6. *Theoretical characteristics of the transductor with* $\mu_1 \rightarrow \infty$ (Fig. 7) *and with* $\mu_1/\mu_2 \geqslant 100$ (Fig. 7); (*a*) *without self-excitation*; (*b*) *with low self-excitation (stable region)*; (*c*) *with high self-excitation (unstable region).*

Fig. 8. *Experimental characteristics of a parallel-transductor.* (*a*) $N_f = 0$ (*no self-excitation*); (*b*) $N_f/N_l = 0{\cdot}4$ (*stable self-excitation*); (*c*) $N_f/N_l = 0{\cdot}8$ (*unstable self-excitation*).

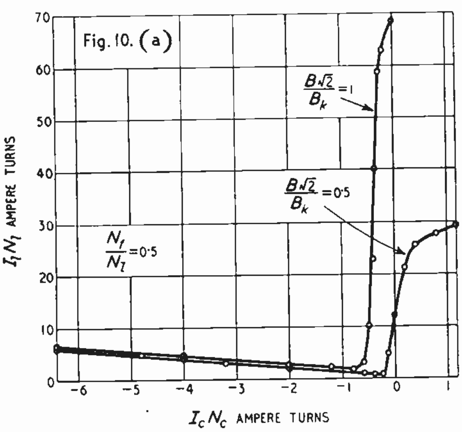
Fig. 9. *Experimental characteristics of a series-transductor.* (*a*) $N_f = 0$ (*no self-excitation*); (*b*) $N_f/N_l = 0{\cdot}8$ (*stable self-excitation*); (*c*) $N_f/N_l = 1{\cdot}6$ (*unstable self-excitation*).

being computed from equations (12) and (13), respectively. The curves of Fig. 6(a) show the characteristics for $N_f = 0$ (i.e., for the transductor without self-excitation) and the curves of Fig. 6(b) for $2k_iN_f/N_l = \frac{1}{2}\pi$, the calculation assuming that $\mu_1 \to \infty$; i.e., that virtually $\mu_2/\mu_1 = 0$. The curves of Figs. 7(a) and 7(b) are calculated with $\mu_1/\mu_2 \geqslant 100$ to show how a finite value of μ_1 affects the bottom section of the corresponding curves of Figs. 6(a) and 6(b), respectively.

The curves of Figs. 6(c) and 7(c) are calculated with $2k_iN_f/N_l = \pi$; i.e., with an increased value of N_f and hence an increased degree of feedback. N_f is so large that the transductor characteristic contains an unstable region, marked by the broken line, causing y to jump at a critical value of x when x goes positive, and to jump down at another critical value of x when x goes negative.

The theoretical curves of Figs. 6 and 7 are verified by corresponding experimental characteristics shown in Fig. 8 for a parallel-transductor and in Fig. 9 for a series-transductor.

The experimental curve of Fig. 10(a) has been obtained with the parallel-transductor to illustrate the large current amplification when N_f is slightly smaller than the critical value $N_{f,crit}$ at which transductor operation becomes unstable, and the experimental curve of Fig. 10 (b) demonstrates the smallness of the instability loop when N_f is slightly greater than $N_{f,crit}$.



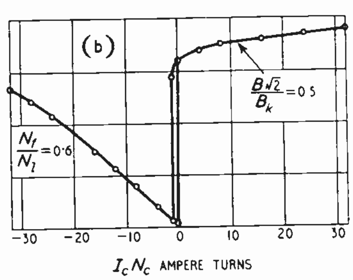


Fig. 10. *(a) Large current amplification of a parallel-transductor with $N_f/N_l = 0{\cdot}5$; (b) small instability loop of a parallel-transductor with $N_f/N_l = 0{\cdot}6$.*

Sensitivity of Control

The current control sensitivity of the self-excited transductor is on the linear part of the characteristic for $\mu_1 \to \infty$ (i.e., $\mu_2/\mu_1 = 0$), see for example Fig. 6(b),

$$S_f = I_l/I_c = k_i \frac{2}{\pi} \frac{N_c}{N_l} \frac{y}{x} \quad \text{(26)}$$

Inverting equation (26) and substituting equation (22) gives

$$\frac{1}{S_f} = \frac{1}{k_i} \frac{\pi}{2} \frac{N_l}{N_c} \frac{x}{y} = \frac{1}{k_i} \frac{\pi}{2} \frac{N_l}{N_c} \frac{\mu_0\mu_2}{B_k} \frac{H_0}{y} - \frac{N_f}{N_c}$$

$$= \frac{1}{S_0} - \frac{N_f}{N_c} \quad \text{(27)}$$

where

$$S_0 = k_i \frac{2}{\pi} \frac{N_c}{N_l} \frac{B_k}{\mu_0\mu_2} \frac{y}{H_0} = k_i \frac{N_c}{N_l} \quad \text{(28)}$$

is the sensitivity of control of the transductor without self-excitation[1]. From equations (27) and (28)

$$S_f = S_0/[1 - S_0N_f/N_c] = \frac{N_c/N_l}{1/k_i - N_f/N_l} \quad \text{(29)}$$

with $k_i = 2$ for the parallel-transductor [equation (14)], and $k_i = 1$ for the series-transductor [equation (15)].

The transductor is in stable operation when $N_f < N_{f,crit}$, but in unstable operation when $N_f \geqslant N_{f,crit}$ where $N_{f,crit}$ is the value of N_f at which $S_f \to \infty$. From equation (29)

$$N_{f,crit} = N_l/k_i \quad \text{(30)}$$

For the linear part of the negative branch of the transductor characteristic we obtain analogously

$$S'_f = S_0/[1 + S_0N_f/N_c] = \frac{N_c/N_l}{1/k_i + N_f/N_l} \quad \text{(31)}$$

For practical evaluation equations (29) to (31) are first approximations because of the assumption $\mu_1 \to \infty$.

Conclusions

Self-excitation of a transductor is an effective means of increasing the control sensitivity and hence the current amplification of the transductor. The degree of self-excitation (i.e., of feedback) is varied in a given transductor by varying the ratio between the number of turns of the self-excitation winding and of the control winding. Transductor operation is stable below a critical value of turns ratio, but unstable at and above the critical value of turns ratio. The transductor with self-excitation in stable condition is usable for a high-gain amplifier circuit. The transductor may be used to function like an on-off trigger relay when self-excitation is in the instability region[2].

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PHYSICAL SOCIETY'S EXHIBITION

THE 35th annual exhibition of scientific instruments and apparatus organized by the Physical Society was held at Imperial College, London, from 6th to 11th April; and this year there was also an overflow into the Huxley Building in Exhibition Road. The number of exhibits was again so large that it is not proposed to describe any that had been shown before or which embodied only insignificant modifications.

It becomes more difficult every year to decide which exhibits fall within the scope of this journal, for many of the techniques which originally were peculiar to radio are now in common use throughout industry and research, and on the other hand there are developments which in themselves could not be described as coming within the field of radio, even in its broadest sense, but which might well interest those concerned with the manufacture of radio and electronic equipment.

Materials

Production of zirconium on a commercial scale was featured by Murex, who claim that it is exceptionally suitable as a getter for valves and other evacuated products. The same firm showed examples of further progress in the production of sintered composite magnets for moving-coil instruments, etc. The range of sintered permanent-magnet alloys now available includes Alcomox III and IV. Information was available on the Telcon stand concerning Permendur, a magnetic alloy notable for its high saturation induction (24,000 gauss) and Curie point (nearly 1,000°C).

The possibilities of high-permittivity ceramics were again demonstrated, this year by Plessey. Capacitors up to 1 μF have been made by stacking rectangular ceramic films. Temperature coefficient characteristics can be controlled to a considerable extent by varying the composition. The piezo-electric properties of these materials are being applied in gramophone pick-ups.

Increasing use of photoelectric devices has created a demand for material transparent to ultra-violet light. Quartz has certain practical disadvantages which are absent in synthetic mica, shown on the Mullard stand. This material is similar to natural mica in appearance and ease of cleavage into thin sheets, but not in chemical composition, and it is more costly. Unlike quartz, it can readily be fusion-sealed to glass.

Various specialized types of tubing were shown by Accles & Pollock, from hair-like gauges of aluminium for instrument pointers, 0·01-in. outside diameter and wall thickness only 0·001 in., to waveguide tubing in copper and steel. Precision tubing for waveguides, etc., is available from Hilger & Watts in solid silver, silver-lined brass, or copper.

Components, including Valves

Most of the components shown were specialized types designed for very high voltages or frequencies or to meet the exacting requirements of Service equipment. The Plessey tantalum-silver electrolytic capacitor, for example, is for use where a working temperature range of $-70°$ to $+200°$C must be maintained, almost regardless of expense. This type, in which the electrolyte is sulphuric acid, is remarkable also for its small size; 22 mm diameter by 6 mm thick for 55 μF at 80 V working. The power factor is less than 5% down to $-20°$C; then rises rapidly to 30% at $-60°$. Vacuum high-voltage r.f. capacitors were shown by G.E.C., the peak rating being 12·5 kV.

Resistors by British Physical Laboratories cover the high range 10^9 to $10^{11}\,\Omega$, with a voltage coefficient less than 0·06% per volt. The Berco hermetically-sealed 3-terminal rheostat, available from 5 Ω to 50 kΩ, was demonstrated under water. The aluminium case has proved to be corrosion-resisting without any special finish. A useful feature of the Plessey ceramic semi-conductor resistors is that they are available with a wide range of temperature coefficients—negative, zero, and positive. The Plessey range of Breeze connectors now includes types permitting cabling to be detached from either side of a pressure bulkhead. Other components were a connector for earthing the electrical systems of aircraft, with a peak-current rating of 1,500 A, and a crimping tool for making reliable solderless connections between cables and tags, etc.

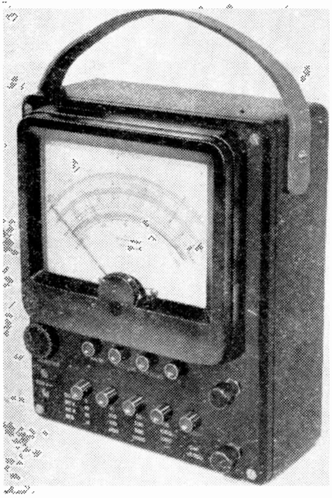
A range of Standard Telephones low-current selenium rectifiers, suitable for television e.h.t. supplies, which was shown in prototype form last year, is now in production. Germanium diodes and triodes continue to be developed; the B.T-H. diodes are now moulded in thermo-setting plastic, and a wire-ended form has been substituted for last year's valve-based G.E.C. triode. The G.E.C. germanium diode is claimed to work satisfactorily as a frequency changer up to 1,000 Mc/s, and a new coaxial silicon type has been developed for use up to 10,000 Mc/s.

The very few new thermionic valves included the Ferranti DP61, a B7G-based high-slope r.f. pentode suitable for frequencies up to 400 Mc/s.

An extensive range of Telecommunications Research Establishment waveguide components was shown by Hilger & Watts. One of the most interesting was an attenuator with micrometer adjustment.

Meters

There were signs of a swing-back of interest from valve-assisted instruments to the "straight" types. The few valve voltmeters to be seen appeared to be mainly conventional in design, but there were some novel features in valveless meters. The non-uniform damping in meters which have been given special (e.g., decibel) scale shapes by pole-piece design has been overcome in Ernest Turner instruments by separating the functions of pointer-drive and damping, using a uniform (or any other desired) magnet gap for the latter. A 14-in. scale is obtained in a 6-in. moving-coil instrument by the same firm, without sacrifice of damping, speed of indication, percentage accuracy, etc, by using a conventional 90° movement coupled by a ligament to the pointer rotating up to 310°. Television has stimulated a demand for e.h.t. voltmeters, and the Electrical Instrument Co. are not convinced that the electrostatic type is necessarily the best. Although a portable 10-kV e.s. instrument is available, it is being supplemented by a moving-coil model, in which a sensitivity of 14 μA full-scale ($>$ 70,000 Ω/V) is achieved by use of the composite Alcomax and iron magnets already mentioned. It has been found essential to embed the resistor chains in



Electrical Instrument Co. Universal test Type 5PB.

polythene to prevent corona and leakage. Besides being more stable than e.s. meters, this type has the advantage of a linear scale. The same firm showed a multi-range test set in which four push-buttons in one row select the type of measurement and five in a second row select the range. A feature of the Salford Instruments precision dynamometers was the method of changing current ranges by loosening or tightening a number of nuts. A new model of the celebrated "Megger," having three voltage ranges (500, 1,000 and 2,500) and six resistance ranges (up to 50,000 MΩ) was shown by Evershed & Vignoles. Another celebrated instrument, the Avometer, is now being provided with a slight modification, enabling it to be used with an accessory for measuring power factor—and hence power—in single-, two-, and three-phase circuits.

Demonstrations on a number of stands testified to the demand for recording instruments. The ingenious Fielden "Servograph" has proved to be particularly valuable, on account of its requiring only a few microwatts from the measured source to give full-scale deflection. The time required for full deflection from zero has been reduced to 3 seconds, and a number of other developments were demonstrated. Its versatility was shown in the aerial polar diagram equipment by Belling & Lee, in which a Servograph has been modified to give straight-line radial deflection, with some slight sacrifice of scale linearity.

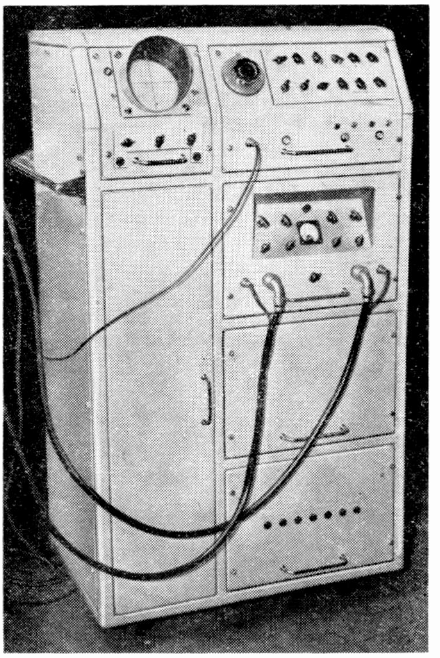
The Standard Telephones maximum-amplitude measuring set is a specialized type of valve voltmeter in which a capacitor is charged through a diode and the indicator continues to read the peak value until either a higher peak comes along or the instrument is reset by a push-button. Two separate indicators are provided so that both positive and negative peaks can be measured simultaneously.

To overcome the disadvantages of sensitive galvanometers for d.c. bridge measurements—risk of damage, slow indication, and susceptibility to vibration—Sunvic Controls have produced an "electronic galvanometer," in which the signal is chopped at 50 c/s by a mains-driven relay and amplified. Signals less than 1 μV can be detected, with input impedance 1,000 Ω, and period less than 0·5 sec. For a.c. bridges, the Baldwin tuned null indicator was shown in improved form.

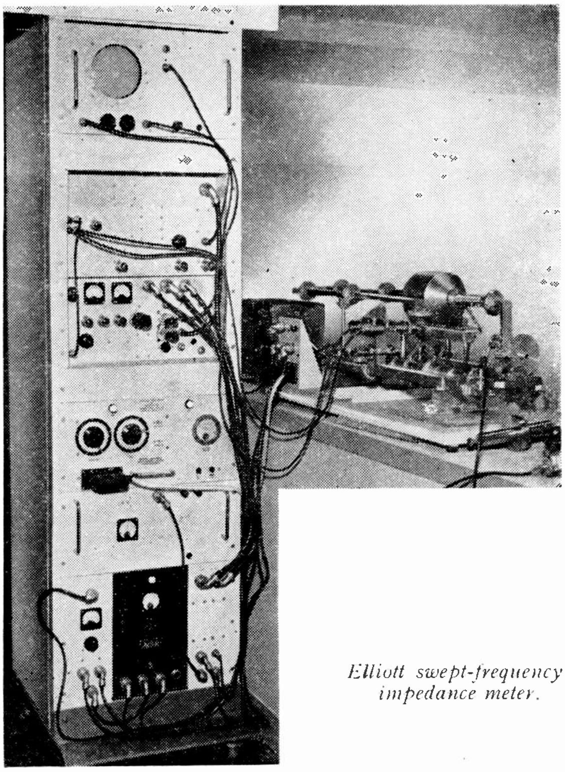
Amplifiers

It is difficult to draw a clear dividing line between this category and the last, because valve-assisted meters incorporate amplifiers, and most laboratory amplifiers incorporate or are intended to be used with indicators. A notable feature of this year's exhibition was the number of z.f. (d.c.) amplifiers. It was interesting to compare the half-dozen or more distinct techniques employed in these amplifiers to combat drift.

The very high voltage gain (500,000 times) obtained by A. E. Cawkell in a physiological amplifier is due to careful balancing of push-pull stages. By adjusting the balancing controls to minimize a 'common mode' test signal, the discrimination against it can be raised to a factor of 20,000. Push-pull connection is also adopted in the 2-stage battery-driven Cossor pre-amplifier, intended mainly for use with their 1049 oscilloscope. Nagard wide-range amplifiers (0–10 Mc/s), with a gain of 300, employ a different balancing technique, in which all the grids are brought to chassis potential on the potential-divider chains between positive and negative lines, and each stage has heavy negative feedback. Pye favour neon-tube couplings, and the Commissariat à l'Energie Atomique use direct couplings and concentrate on stabilization of power supplies. Southern Instruments use a relay system to correct zero drift, and Electronic Instruments incorporate a Servograph in a system that



Metropolitan-Vickers high-speed waveform recorder.



Elliott swept-frequency impedance meter.

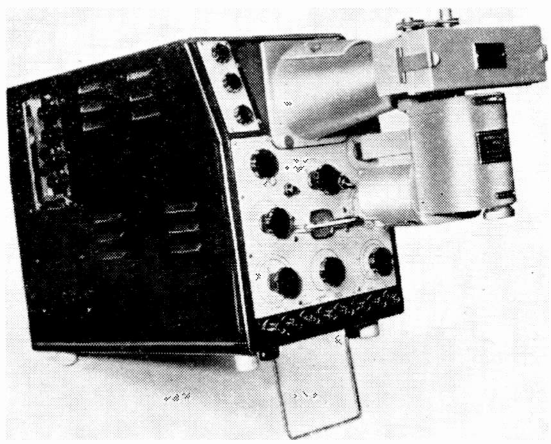
corrects for drift once per hour. Other designers use indirect methods: an example of the chopper method, in which the direct voltage is interrupted by a vibrator so that a 50-c/s amplifier can be used, was shown by Sunvic Controls ; a magnetic type of modulator is used by Elliott Bros. to give an amplifier signal at 50 c/s ; and in the Nagard type 103/1 amplifier, with a gain of 20,000, the carrier frequency is as high as 14 Mc/s in order to cover input frequencies of 0–2 Mc/s. If very high sensitivity is attempted, the contact vibrator type of modulator causes difficulty with contact potentials, so a proximity vibrator is used to vary the input capacitance, as in the electrometers shown by Electronic Instruments, Dawe, E. K. Cole, and the Atomic Energy Research Establishment. Lastly, there is the reflecting galvanometer and photo-cell method adopted by Tinsley and Elliott Bros. The former incorporates a new type of vibration-proof galvanometer, in which the suspension is totally immersed in liquid.

Purely alternating amplifiers were comparatively few, but some interesting examples were to be seen on the stand of Solartron Laboratory Instruments. These are so stabilized as to be suitable for laboratory measurement work with meter or oscilloscope. One model gives a gain of 80 db from 10 c/s to 250 kc/s, and another 50 db from 20 c/s to 20 Mc/s, the latter enabling a rise time of 0·02 μsec to be observed. A valuable piece of e.h.f. equipment is the Wayne Kerr laboratory receiver, covering signal frequencies from 400 to 4,000 Mc/s. It comprises several rack-mounted units—power supply, main amplifier unit, three types of frequency changer and their associated head amplifiers, and two local oscillators with self-contained power supplies.

Oscillators and Signal Generators

Resistance-capacitance frequency-determining circuits are now practically standard for oscillators for audio frequencies and well into the radio frequencies, as in the models by Solartron (25 c/s–250 kc/s) and Standard Telephones (30 c/s–300 kc/s), as well as for sub-audible frequencies (Inter Electron Industries, 2–10,000 c/s). The Muirhead-Wigan decade oscillator (1–111,100 c/s) appeared with important improvements based on experience with the original model, and there is also a variety covering 0·1–20,000 c/s.

Some new signal generators have been produced for television. The Cossor "Tele-Check" is a servicing instrument giving a wobbulated signal for receiver alignment in conjunction with an oscilloscope, and a v.f. pattern for checking time-base linearity. The Standard Telephones waveform generator, on the other hand, is designed to produce accurately-formed synchronizing signals for the testing of television relay systems. General pulse testing is provided for by the Solartron generator, 1–40 μsec, with variable delay with reference to an injected triggering pulse in the range 50–10,000 c/s.



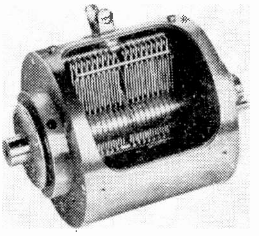
Cossor 1049 *oscilloscope with* 1428 *camera on* 1431 *motor drive for camera.*

A swept-frequency oscillator developed by G.E.C. for the Ministry of Supply is unusual in that it brings to the v.h.f. band (30–140 Mc/s), the cascaded phase-shift technique originally devised for frequencies of the order of 25 c/s. The chief modification is that a cathode-follower is used for each of the four phase-shift stages, and the frequency is varied by a suitable recurring voltage applied to the valve grids, controlling the effective phase-shift resistance. Still higher frequency signals are generated in a neat self-contained mains-driven source shown by Metrovick, covering 8,500–9,700 Mc/s, with a reflex klystron and incorporating a direct-reading wavemeter and calibrated attenuator.

Systems for providing a large number of spot-frequencies per quartz crystal were shown by Marconi's W.T. Co. and Royal Aircraft Establishment. In the Marconi system, the F.M.Q. method of modulating a crystal oscillator is used to provide 25 adjacent controlled frequencies in a band. An example of the R.A.E. system uses 32 crystals to provide 2,000 spot-frequencies at $16\frac{2}{3}$-kc/s intervals throughout the 33–66-Mc/s band.

Bridges and Standards, including Impedance Measurement

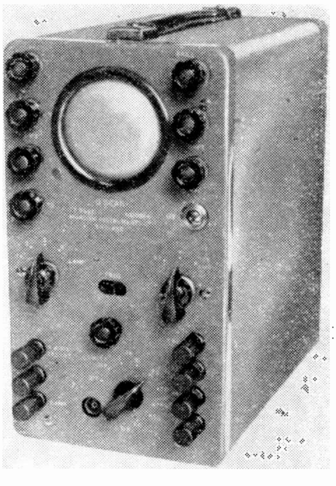
An important introduction is the Mullard variable capacitor, in which precision quality is combined with small size and suitability for quantity production. The 20–340-pF model has a capacitance repeatability to within $\pm$0·01%, temperature coefficient -15 ± 10 in 10^6 per °C, inductance 0·012 μH, and power factor (at 1 Mc/s) 0·00004. The insulation is polytetrafluoroethylene, and the plate shape yields an accurate straight-line-frequency law. A two-gang model is also available in the same external size, but with a lower accuracy.



Mullard precision variable capacitor.

Experimental work with the Wayne Kerr v.h.f. bridge B.901, shown last year, now enables its accuracy up to 250 Mc/s to be specified. Another v.h.f. bridge, shown by G.E.C., uses continuously-variable inductive ratio arms. At still higher frequencies, 300–600 Mc/s, permittivity and power factor of solids and liquids are measured by the resonance-curve method in a cavity resonator, as shown by the N.P.L. An instrument for measuring power up to 60 W in the 100–160-Mc/s band was shown by Plessey; and a milliwatt calibrator by Standard Telephones, used for checking line test sets, employs as the standard a Weston cell. Another instrument using a standard cell for comparison is the electronic reference unit by Sunvic Controls, providing potentials up to 1 V.

The Langham Thompson direct-reading capacitance meter, with four ranges up to 250 pF, is unusual in using a frequency discriminator to give an indication of the frequency shift of a 465-kc/s oscillator when the capacitance to be measured is connected. The B.P.L. interelectrode capacitance meter enables input, output, and grid-to-anode capacitances to be measured at the same time by a resonance method.

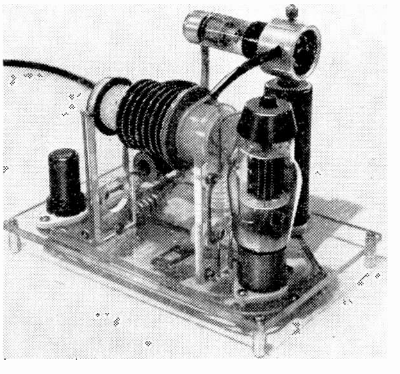


Marconi Instruments Q-scan television alignment oscilloscope.

Primarily for design work on wideband aircraft aerials, but useful for measuring other impedances, the Fairey Aviation automatic impedance plotter covers the wide range 40–600 Mc/s and has a cathode-ray presentation. So has the microwave swept-frequency impedance meter centred on 9,400 Mc/s by Elliott Bros. An impedance grid of the Smith chart type can be placed over the tube screen for directly reading Z/ϕ. Very different in frequency coverage is the network analyser by Sunvic Controls, designed to show the phase characteristics of four-terminal networks at low and very low frequencies. The angle is read from the scale of a control when it has been set to reduce the phase ellipse to a straight line.

Oscilloscopes

Recent progress has been mainly towards higher speed and refinement of auxiliary apparatus. The Elliott exhibits included the ultra-high-speed oscilloscope developed from that shown in 1949 by W. Nethercot, which gives sweep times down to 0·05 μsec. Using a 10-kV c.r. tube, writing speeds of 20,000 km/sec can be photographed. A very clear steady trace of a 0·1-μsec pulse was to be seen on the Radar Research and Development Establishment high-speed oscilloscope. Wideband amplification up to several hundred Mc/s might seem impracticable, but that is in effect what Metrovick have achieved in a c.r. monitor for observing recurrent waveforms having components up to 300 Mc/s and more. The principle is analogous to that used for stroboscopically examining fast-moving machinery—



Brandenberg Designs BG 5130 high-voltage generator.

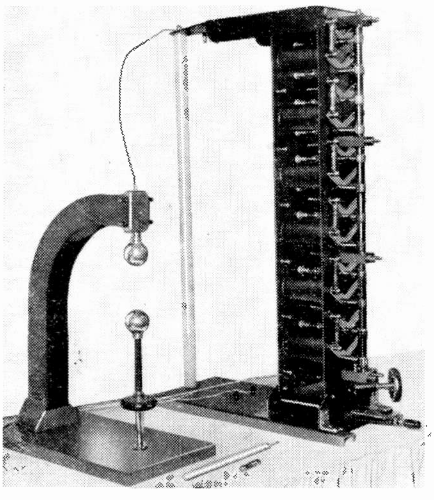
the waveform is plotted at points taken from successive groups of cycles, using a gating pulse of the order of 0·001 μsec.

A considerable amount of photographic apparatus for oscillography was shown, including the Cossor motor drive with an ingenious 9-speed gearbox. An electronic switch for displaying four non-recurrent waveforms on a single-beam oscilloscope was shown by R.R.D.E. Each is plotted in steps of 10-μsec duration.

For servicing television receivers, Marconi Instruments have produced the 'Q-Scan,' a small portable oscilloscope which, in addition to the usual facilities, includes a wobbulator covering the range 10–95 Mc/s.

Power Supplies and E.H.T. Test Gear

Several new stabilized power units were to be seen; the Solartron model was an example of a variable-voltage unit suitable for general laboratory use; the new Airmec unit has both positive and negative stabilized supplies; and Inter Electron showed d.c. units with voltage variable down to zero and a.c. units up to 500-W output controlled by a servo-operated carbon pile. Apart from stabilization, the main trend is in the direction of higher voltages, for purposes such as projection television. A range of models shown by Brandenberg employ rectified r.f. to give various high voltages at a moderate cost. Hazlehurst exhibited a 100-kV impulse generator employing the Marx circuit, and are prepared to supply units up to 250 kV. An ionization test set shown by the same firm has an output variable from 1 to 30 kV. A portable test set shown on the Marconi Instruments stand provides up to 10 kV for ionization and other insulation tests and as an e.h.t. source for testing c.r. tubes. The Cinema Television surge testing equipment uses high-voltage pulse technique for detecting faults in transformer windings, etc, with a c.r. display.



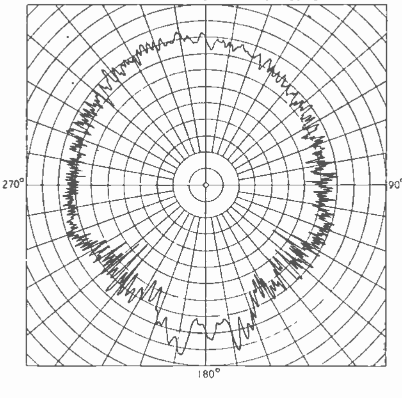
Hazlehurst Designs 100-kV impulse generator.

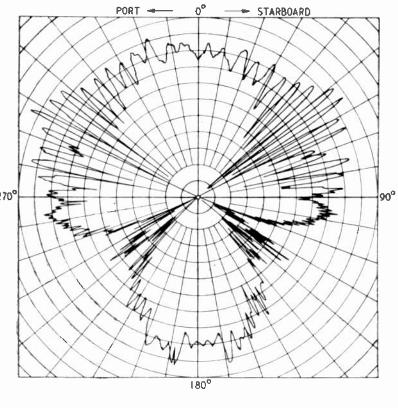
Miscellaneous Apparatus

Three firms demonstrated the automatic plotting of e.h.f. aerial radiation diagrams. The comparatively simple Belling & Lee system, already mentioned, uses scaled-down slowly-rotating models of television aerials, receiving a test signal from a fixed source on a proportionately increased frequency, to predict the performance of full-sized types without the inconvenience of direct testing. In the Ferranti equipment, for plotting the characteristics of aircraft aerials, a scale model of the aircraft is rotated and the signal from it received and plotted against angle. The differences between plots taken with and without propellor disks are remarkable, and provide a measure of propellor modulation. The E. K. Cole equipment plots diagrams for microwave aerials and in this case there is, of course, no need for scaling down. As with the B. & L. apparatus, the signal source is fixed.

A number of electronic computing and counting units were shown, including a small edition of the A.C.E. pilot model (on the N.P.L. stand), a straightforward counter by Panax with binary neon indication up to 64 and mechanical counting above, a Fourier synthesizer by the Birkbeck College Research Laboratory, and a somewhat entertaining set of 'logical' computors by Ferranti, giving the possible solutions to problems such as whether one should go out to the theatre or visit one's uncle, given certain conditions.

The variety of techniques employed in thickness gauges rivalled those for z.f. amplification, including such diverse methods as β-ray absorption (Electronic Instruments and E. K. Cole), magnetism (Salford Electrical Instruments), and oscillator detuning detected by beating (Nash & Thompson and G.E.C.). The latter is used for automatically controlling the aluminizing of c.r. tube screens during manufacture. The tracing of variations in the thickness of a slab of material





Curves drawn by the Ferranti aerial radiation diagram recorder with a model of a Vickers Viscount aircraft using a $\lambda/2$ unipole aerial fitted on the belly centre line 12 ft from the nose. The scaling factor is 1/24 and the frequency 24,000 Mc/s. The curve on the left is without and the one on the right is with propellor discs.

having homogeneous dielectric properties was demonstrated by the Armament Research Establishment with equipment in which the slab is slowly scanned with a search electrode forming part of a capacitance bridge, the variations being recorded to scale on a paper which can then be placed on the dielectric sample to locate the recorded variations.

The Physics Department of Reading University (Mr. M. S. Wills) demonstrated a magnetic field-strength meter in which the output of a rotating search coil is conveyed to the amplifier and indicator via a transformer with a rotating primary and stationary secondary. The range of measurement is from a few oersteds up to 20,000. An entirely different method, based on harmonic distortion caused in Permalloy-cored windings, is employed in the instrument shown by the Office National d'Etudes Aeronautiques, and is sensitive to tenths of a millioersted.

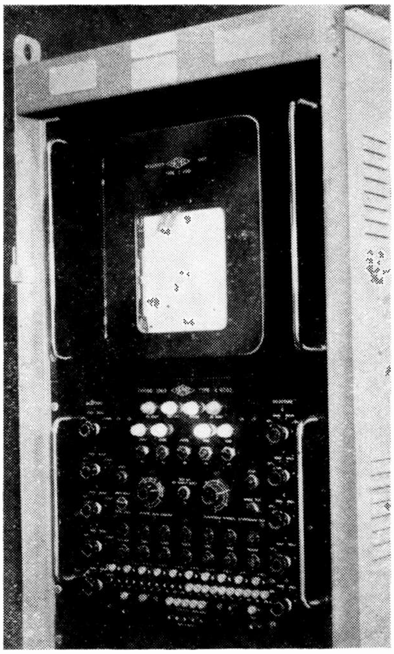
The Marconi Instruments TF.929 instrument measures r.f. field strength from 120–18,000 kc/s by means of the receiver and comparison signal-generator method.

The contact sequence analyser shown by Airmec enables the performance of relays to be thoroughly and rapidly checked. Three successive 150-msec traces are made on a paper roll for each pair of contacts up to a maximum of ten. The connections are made automatically by uniselectors. Another Airmec unit shown was a batch counter operated by signals from a photoelectric set-up or any other suitable pick-up device.

Various other equipments worthy of mention are the A. E. Cawkell valve-noise measuring apparatus for the G.P.O., the Cinema Television adjustable phase correction units and direct-reading electronic frequency meter (seven ranges, up to 150 kc/s), the Barr & Stroud radar chart comparison unit, the W. Edwards vacuum leak detector exploiting the properties of palladium in relation to hydrogen, and the Ediswan polythene welder, which electrically heats a jet of compressed air to a suitable temperature for welding polythene, p.v.c., and similar plastics. The technique of drilling the final conical bore in diamond wire-drawing dies was well staged by G.E.C. And a most interesting demonstration was the 'synthetic' waveguide designed by Prof. Barlow and made by Tinsley, enabling many of the properties and characteristics of propagation to be visualized and studied. Details are given in *J. sci. Instrum.*, November 1950.

Research

The boundary of research is difficult to lay down in this exhibition as now constituted, and a number of the items already mentioned could no doubt be reckoned within it. A particularly interesting system was demonstrated by Dr. Schneider, of Durham University. It enables an input voltage of the order of 10^{-8} to be detected, by synchronizing it with an RC oscillator critically adjusted for amplitude of oscillation. Owing to the feedback characteristics of the system, an appreciable change in amplitude is caused. In effect, noise is reduced to an extremely low level by narrowing the bandwidth almost to zero. The system can be regarded as a development of the homodyne method of reception. It operates at audio or sub-audio frequencies; but the extra-terrestrial noise-measuring apparatus demonstrated by T.R.E. is at the other end of the frequency scale, in millimetre wavelengths. The noise is chopped to 20 c/s and amplified at that frequency after crystal rectification. The properties of 160-Mc/s polarized waves were shown by Marconi's W.T. Co. by means of a receiving unit consisting of three dipoles mutually at right angles and used separately or in various combinations. G.E.C. research exhibits included evidence of work for the Ministry of Supply on quartz crystals, and a demonstration of some of the piezoelectric properties of the high-permittivity barium titanate, enabling it to function as a resonator or oscillator.



Recorder and timing units of the Airmec contact sequence analyser.

CORRESPONDENCE

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

Surface-Wave Transmission Line

SIR,—Professor Barlow in the February number suggests that the use of a smooth bare wire would not be satisfactory because of the wide radius of the field. This, of course, depends on the frequency and the permissible radius of the field.

We have found during recent experiments with a 2·5 mm bare copper wire—50 metres long—using a frequency of 4 700 Mc/s that the field radius, for approximately 90% transmission of power, was about 8 cm, which is less than the calculated radius. This reduction, I suggest, may be due to the thin film which is formed on the copper surface after exposure to the atmosphere.

The measured attenuation was within 10% of the calculated value and the variation over a period of four weeks during varying weather conditions of snow and rain with temperatures ranging between 42°F–14°F was, with one exception, very small, the exception being the formation of water drops following a slow thaw or the formation of frozen drops of water near freezing point. This results in an additional loss of some three to five times in power. It is easily overcome by heating or wiping the wire. I agree with Mr. Rust's comments in the October/November number concerning the use of dielectric loading and would mention that we have found that it is not necessary to use horns for launching the wave. A tapered dielectric rod which forms part of a coaxial/wire or waveguide/wire transition is also satisfactory and unlike a horn does not need special protection against snow or rain when using the wire vertically.

R. H. NELSON.

Telefon AB L M Ericsson,
Stockholm, Sweden.
13th March, 1951.

Twin-T Circuits

SIR,—I find upon enquiry among my colleagues the view is prevalent that the parallel-T RC circuit has a great disadvantage in being unsuitable for applications requiring a variable rejection frequency. It is pointed out that three-ganged elements are required.[1] This is not necessarily the case. In Fig. 1 the condition that there shall be no output at any one frequency is

$$\frac{R_1R_2}{R_1 + R_2} . C_3 = R_3(C_1 + C_2)$$

which may be written

$$T_L = T_H \quad \ldots \quad (1)$$

where T_L is the time-constant of the low-pass component-T and T_H that of the high-pass portion, the ends of the network being shorted. The frequency at which this occurs is given by

$$\omega_0^2 = \frac{1}{C_1C_2R_3(R_1 + R_2)}$$

$$= \frac{1}{T_LT_M} \quad \ldots \quad (2)$$

where T_M is the time-constant of the mesh $R_1R_2C_1C_2$. It will thus be seen that provided equation (1) is satisfied (and this is the more fundamental condition) only one element in each of the component-Ts need be varied. The obvious possibilities are R_s and C_s; R_s and R_p; C_s and C_p; where the suffixes *s* and *p* denote series and parallel elements respectively. Of the other possibilities we shall confine our attention to those shown in Figs. 2 and 4. The former of these has already been pointed out by Stanton,[2] who states that the variable components cannot be ganged. It is easy to show, however, that this view is wrong. If for any position of the series potentiometer its resistance is divided into fractions γ and $1-\gamma$ of the total value, then equation (1) requires

$$\gamma(1-\gamma)R_sC = 2CR_p$$

and this is satisfied in the circuit of Fig. 3 where the shunt resistance is now a shorted potentiometer. With this connection the mesh time-constant T_M necessarily remains constant and the variation in rejection frequency is due to that in the time-constant of the component channels.

In Fig. 4 the series variable is a split-stator or differential capacitor with the usual property of this device that the sum of the capacitances between the rotor and stator halves is a constant. The stator/stator capacitance varies from zero through a maximum and back to zero

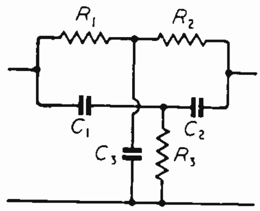


Fig. 1

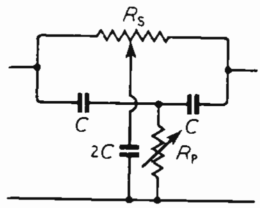


Fig. 2

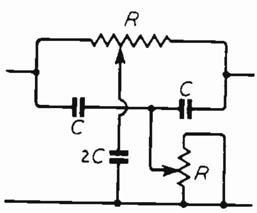


Fig. 3

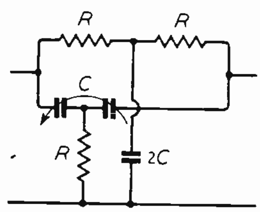


Fig. 4

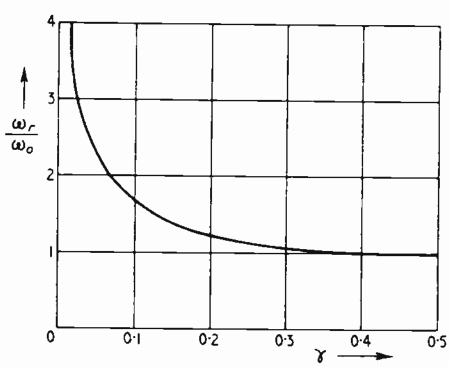


Fig. 5

for 180° rotation. It is thus the capacitance or susceptance analogue of the shorted potentiometer used above. As the capacitor shaft is rotated the value of T_H remains constant since as much is added to as is subtracted from the effective shunt capacitance on R_3. Equation (1) is, therefore, satisfied with a fixed low-pass circuit. The rejection frequency will vary this time because the mesh time-constant T_M is changing. This circuit gives variable rejection virtually by control of a single component. The manner of variation with linear controls is shown in Fig. 5 where the ratio of rejection frequency to that obtaining at centre position of the control is plotted against shaft position. With our simple arrangement the component is only properly used at one end of its travel, but this disadvantage can be overcome by suitable padding.

S. C. DUNN.

Tollington Park,
London, N.4.
27th February, 1951.

[1] S. W. Punnett, "Audio Frequency Selective Amplifiers," *Journ. Brit. Instn Radio Engrs*, February 1950, Vol. 10, pp. 39–59.
[2] L. Stanton, "Theory and Application of Parallel-T Resistance-Capacitance Frequency Selective Networks," *Proc. Inst. Radio Engrs*, July 1946, Vol. 34, pp. 447–456.

Maintaining Oscillations by the Periodic Variation of L or C

SIR,—Your Editorial (March 1951) deals with an interesting subject and reaches conclusions by a purely physical argument.

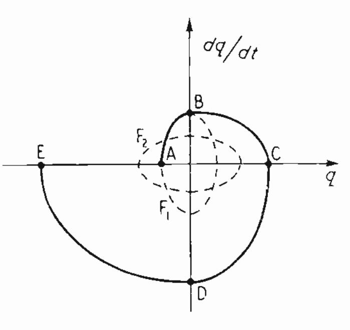
The whole picture of this phenomenon (called sometime *parametric excitation*) can be still further specified by an elementary mathematical argument as follows.

The differential equation (d.e., for short) of an ideal oscillating circuit is $L_0 d^2q/dt^2 + (1/C)\,q = 0$ where q is the charge on the capacitor. It can be reduced (setting $1/L_0C_0 = \omega_0^2$ and using $\tau = \omega_0 t$ as an independent variable) to the form $d^2q/dt^2 + q = 0$. The integral curves of this d.e. in the plane of the variables q, dq/dt (*the phase plane*) are circles concentric with the origin.

Suppose that, instead of the fixed parameters L_0, C_0, we consider that these parameters can vary between the limits $L = L_0 \pm \Delta L$; $C = C_0 \pm \Delta C$ assuming that one of these parameters, say C, is held constant at the value $C_0 + \Delta C$ during the time $\pi/2$ and then suddenly changed to $C_0 - \Delta C$ at which value it is held for another $\pi/2$ interval, etc. Without any modification in the qualitative aspect of the phenomenon, one can assume that these changes occur instantaneously; this simplifies the argument.

Under these assumptions we have two d.e.

$d^2q/dt^2 + (1 + \alpha^2)\,q = 0$ and $d^2q/dt^2 + (1 - \alpha^2)\,q = 0$ replacing each other at the (double) frequency of the parameter variation. In these d.e. $\alpha^2 < 1$ is a constant easy to evaluate in carrying out this calculation. The first of these d.e. is for $C = C_0 - \Delta C$ and the second for $C = C_0 - \Delta C$. The family F_1 of integral curves for the first d.e. are ellipses elongated along the dq/dt axis and the family F_2 for the second are ellipses elongated along the q axis as shown in the figure. Through any point of the phase plane P passes one and only one curve of each family.



Let us assume that we start the investigation from the initial condition represented by the point A of P with the ellipse F_1 (i.e., $C = C_0 - \Delta C$) and at the point B we change the parameter to $C = C_0 + \Delta C$ (family F_2) No impulsive work will be involved in this operation since $q = 0$ at B. We follow now the arc BC and change the parameter at C to $C = C_0 - \Delta C$. This time there will be an impulsive work required since $q \neq 0$ at this point. After that an arc CD will be traversed and at D the parameter will be changed to $C = C_0 + \Delta C$ without any work and we complete the cycle at D, where again an impulsive work will be required, etc.

Since in this mode of representation the square of the radius vector at A, B, C . . . is a measure of the total energy of the system, one sees that the energy steadily increases and, in the linear case, grows beyond any bound. In a non-linear case, on the contrary, it approaches a limit.

An exactly similar argument can be reproduced when the parameter L undergoes variations. The only difference from the preceding case will be that the points at which the parameter changes without involving any work are now on the q axis and those at which the impulsive work is required on the dq/dt axis.

One can extend this argument to the case when the parameter changes occur continuously but this requires a more detailed study of the d.e. of Mathieu.

A quite natural question arises: how can one time these changes of parameter so accurately as to produce precisely the configuration shown, without drifting one way or the other which obviously would in the long run destroy this theoretical picture?

The answer to this question is that this particular configuration is the only one *which is stable*. Even if some disturbing factors appear tending to destroy this configuration, the stability condition will restore this configuration again.

The *natural* parametric effect takes place always in the direction mentioned; viz., the mechanical energy is absorbed in the degree of freedom of the parameter variation and appears (is 'generated') in the electrical form in the oscillating circuit. The inverse parametric effect (the parametric damping) is possible theoretically but it corresponds to the *unstable phase*. Hence, if by some external means this unstable phase could be artificially stabilized, this inverse effect is likely to be detected but, as far as known, no such attempts have been made.*

N. MINORSKY.

Aix en Provence,
B. du. R., France.
8th March, 1951.

* See also a note "Sur l'excitation paramétrique" by the writer in *Comptes Rendus*, 18th December, 1950, Vol. 231, p. 1417.

Television Camera Tubes

SIR,—In Mr. L. H. Bedford's stimulating article in your January* issue he sets out to give a comparison of the performance of various types of television pick-up tubes. However, his conclusions are rendered misleading by reason of several questionable basic assumptions that he makes and his use of out-of-date data.

Thus, on page 5, Mr. Bedford states that the data quoted for the Orthicon in Table 1 are really those for the C.P.S. Emitron. In fact, the figures he quotes for the sensitivity of the C.P.S. Emitron are about 10 times worse than those made public, and demonstrated at the I.E.E.[1] on 12th April, 1950—some nine months before the publication of his article.

In the rest of the paper, the reader is left under the impression that the data quoted under the generic heading of "Orthicon" are, in fact, the data for the C.P.S. Emitron: the only hint to the contrary is in the references in Table 4 where we find that the resolution figures are those for an R.C.A. Orthicon. The reader is referred to the above mentioned papers before the I.E.E., especially the authors' reply to the discussion, for accurate data on the C.P.S. Emitron.

In Table 2, the factor S_3 is used to 'write off' the fact that certain tubes have a limiting signal/noise ratio less than the figure of 100 : 1 which is quite reasonably assumed for the ideal "quanticon". Surely this destroys the basis of comparison, which should be for equal signal/noise ratios? It has the unfortunate effect of making the tube with the lowest signal/noise ratio appear best, other things being equal. It would seem better to calculate a hypothetical $\phi_{100}^{-\frac{1}{2}}$ (where the suffix indicates 100 : 1 signal/noise ratio) from the experimental values of $\phi^{-\frac{1}{2}}$ and signal/noise obtainable; e.g., Image Orthicon Type 5820 would read $\phi_{100}^{-\frac{1}{2}} = \frac{\phi^{-\frac{1}{2}}}{S_3} = \frac{51}{2{\cdot}8} = 18$. No alteration would be necessary for tubes with $S_3 = 1$. Incidentally there is a misprint in Table 2. The column headed "$\phi^{\frac{1}{2}}$" should, of course, be "$\phi^{-\frac{1}{2}}$".

In connection with Table 2, the method of utilization of the factor η_2 (gain of image stage) is erroneous. As there shown, its effect is the same as that of an increase in the primary photo-current, and hence the signal/noise ratio, but this, of course, cannot be the case: no tube can be better than its photo-cathode. What it does do in the case of the Image Orthicon is to reduce the adverse effect of values greater than unity of the factor S_1 (ratio of beam current to signal current), in fact, the signal/noise ratio in the return beam (and thus, neglecting multiplier noise, in the electron multiplier output also) is

$$\frac{1}{\sqrt{(1 + S_1/\eta_2)}}$$

times the signal/noise ratio of the primary photo-emission.

Formula (6) on page 6 for the noise/signal ratio in the general case is incorrect for the same reason, and also—rather less excusably—because the noise for the amplifier is taken [see Appendix (c)] as $\delta \overline{I_n}^2 = 2eI_a\delta f$. This expression is only true for random or temperature-limited emission, not for space-charge-limited current.[2] The temperature-limited formula gives a noise power 6-7 times greater than the measured value for the types of valve commonly employed. It is usual to express the valve noise as that of an equivalent noise resistance R_n considered to be at room temperature and inserted in series with the grid. For valves operated under optimum conditions, the lowest value of R_n seems to be about $3/g$.

Considering for simplicity formula (3) of the Appendix, this becomes Noise/Signal ratio =

$$F^{\frac{1}{2}}I^{-1}\left[2eI\left(1 + \frac{S_1}{\eta_2}\right) + m^{-2}4kTR_n\frac{(2\pi FC)^2}{3}\right]^{\frac{1}{2}}$$

Finally, referring to Section 4, we would point out that variations in landing velocities of the beam electrons in a Cathode Potential Stabilized type of pick-up tube, Orthicon or C.P.S. Emitron, do not cause shading or spurious signals. They merely alter the potentials at which various parts of the mosaic stablize in the dark. Also target secondary emission does not produce spurious signals in this type of tube.

E. L. C. White.
J. D. McGee.

E.M.I. Research Laboratories,
Hayes, Middlesex.
30th March, 1951.

* "Television Camera Tubes," *Wireless Engineer*, Jan. 1951, Vol 28, No. 328, p. 4.
[1] *Proc. Instn elect. Engrs*, Part III, Nov. 1950, pp. 377-413.
[2] Thompson, North and Harris, *R.C.A. Rev*, April 1940, Vol. 4, No. 4, also July 1940, Vol. 5, No. 1.

NEW BOOKS

Semi-Conductors

By D. A. Wright. Pp. 130+viii, with 32 diagrams. Methuen & Co., 36 Essex Street, London, W.C.2. Price 7s. 6d.

The author is a member of the research staff of the General Electric Co., at Wembley, and expresses his indebtedness to numerous members of the staff for their assistance. The book is one of a series of monographs on physical subjects. The title is apt to be misleading; a much better title would have been "The Theory of Electron Flow in Metals and Semi-conductors." This is made clear in the preface where it is stated that the book is specially concerned with the theory of electron flow in semi-conductors, and across the boundary between them and either a metal or a vacuum. It also states that "only brief reference is made to practical applications and experimental results, as the main interest is in the basic theory."

The book is necessarily very theoretical and on page 4 the reader is introduced to the exclusion principle and Fermi-Dirac statistics. There are eight chapters which cover Electrons and Metals, Electrons in Crystals, Electron Emission from Semi-conductors, Determination of Electron Density in Semi-conductors, Secondary Emission, Metal-Semi-conductor Contacts, Thermionic Cathodes, Photo-electric Cathodes. Numerous references are given at the end of every chapter.

This monograph will undoubtedly be of great use to students of the electron physics of solids.

In Fig. 9(c) the point P referred to in the text does not appear, but by way of compensation the point O occurs twice. On page 48, semi-conductors are given a high-temperature coefficient of resistance, and in Chapter IV there is much confusion between N and n. These are minor details, however, and the book can be recommended to anyone seriously interested in the subject.

G. W. O. H.

Oxford Junior Encyclopaedia

Vol. IV, Communications. Pp. 496+xvi; illustrated. Oxford University Press, 12 Warwick Square, London, E.C.4. Price 30s.

Practical Wireless Encyclopaedia (12th Edition)

By F. J. Camm. Pp. 360+xii; with 554 illustrations. George Newnes, Ltd., Southampton Street, Strand, W.C.2. Price 21s.

Vacuum

This is the title of a new quarterly journal intended for the scientist and industrialist. It covers high-vacuum technology generally. Although it is published by W. Edwards & Co. (London), Ltd., Worsley Bridge Road, Lower Sydenham, London, S.E.26, who are manufacturers of vacuum equipment, the publishers state that it is not a 'House' journal. In addition to articles on vacuum matters, an abstracts section is included.

BRITISH STANDARDS

An exhibition showing the benefits derived from standardization is to be held at the Science Museum, South Kensington, London, from 18th to 28th June (except Sunday). Admission will be free and it will be open from 10 a.m. to 7 p.m.

The exhibition is a part of the celebration of the Golden Jubilee of the British Standards movement.

ABSTRACTS and REFERENCES

Compiled by the Radio Research Board and published by arrangement with the Department of Scientific and Industrial Research

The abstracts are classified in accordance with the Universal Decimal Classification. They are arranged within broad subject sections in the order of the U.D.C. numbers, except that notices of book reviews are placed at the ends of the sections. U.D.C. numbers marked with a dagger (†) must be regarded as provisional. The abbreviations of the titles of journals are taken from the World List of Scientific Periodicals. Titles that do not appear in this List are abbreviated in a style conforming to it.

ACOUSTICS AND AUDIO FREQUENCIES

016 : 534 **1042**

References to Contemporary Papers on Acoustics.—A. Taber Jones. (*J. acoust. Soc. Amer.*, Nov. 1950, Vol. 22, No. 6, pp. 888–894.) Continuation of 1 of January.

534 : 061.4 **1043**

The Audio Fair Review.—(*Audio Engng*, Dec. 1950, Vol. 34, No. 12, pp. 24–30 . . 54.) Brief descriptions of loudspeakers, amplifiers, pickups, recorders, etc., exhibited at the Second Audio Fair.

534.21 : 551.463 **1044**

The Dependence of Shadow-Zone Sound on the Surface Sound-Velocity Gradient.—R. W. Morse. (*J. acoust. Soc. Amer.*, Nov. 1950, Vol. 22, No. 6, pp. 857–860.) Sound penetrating the so-called 'shadow zone' of ocean acoustics is investigated by ray-theory methods. On the assumption that the sound-velocity gradient approaches zero in a surface layer a few feet deep, sound levels in the shadow are derived of the same order as those measured at a frequency of 24 kc/s.

534.213.4 : 534.373 **1045**

Boundary-Layer Attenuation of Higher-Order Modes in Rectangular and Circular Tubes.—R. E. Beatty, Jr. (*J. acoust. Soc. Amer.*, Nov. 1950, Vol. 22, No. 6, pp. 850–854.) The boundary-layer attenuation caused by viscous and thermal losses of all higher-order modes of propagation in rigid-walled tubes is calculated, using the concept of boundary-layer admittance introduced by Cremer. The symmetrical $(0, n)$ circular-tube modes are the least attenuated of all higher-order modes in both rectangular and circular tubes; above a certain frequency the attenuation is less than that for a plane wave.

534.23/.24 **1046**

The Equivalent Circuit for the Transmission of Plane Elastic Waves through a Plate at Oblique Incidence.—W. T. Thomson. (*J. appl. Phys.*, Dec. 1950, Vol. 21, No. 12, pp. 1215–1217.) Formulae are derived for the transmission and reflection coefficients of a plate with different liquids in contact with its two faces. A simple equivalent circuit involving impedances which are functions of the angle of incidence is given for a plate transmitting both the dilatation and the shear wave.

534.26 **1047**

On the Diffraction of a Plane Sound Wave by a Paraboloid of Revolution.—C. W. Horton & F. C. Karal, Jr. (*J. acoust. Soc. Amer.*, Nov. 1950, Vol. 22, No. 6, pp. 855–856.) "An infinite set of equations is obtained for the coefficients of the pressure fields inside and outside the paraboloid. For the particular case of a rigid paraboloid the coefficients of the series for the scattered field can be obtained explicitly. The ratio of the total pressure at the nose of the paraboloid to the pressure of the incident wave is shown as a function of the shape of the paraboloid for a plane wave whose wave front is perpendicular to the axis of revolution. The same quantity is plotted for the case of a sphere whose radius of curvature is equal to that of the paraboloid at the nose."

534.321.9 : 534.22 : 577 **1048**

The Velocity of Sound through Tissues and the Acoustic Impedance of Tissues.—G. D. Ludwig. (*J. acoust. Soc. Amer.*, Nov. 1950, Vol. 22, No. 6, pp. 862–866.)

534.321.9 : 534.373 **1049**

Effects of Humidity on Ultrasonic Absorption in Air at 14·6 Mc/s.—G. S. Verma. (*J. acoust. Soc. Amer.*, Nov. 1950, Vol. 22, No. 6, pp. 861–862.) Using the acoustic resonator interferometer developed by Hubbard (1932 Abstracts, pp. 171 and 647), the absorption coefficient (α) was determined for air of relative humidity ranging from zero to 84%. At the frequency used, a maximum value $\alpha = 0{\cdot}52\ \text{cm}^{-1}$ was found for a relative humidity of 46%. This confirms results obtained at lower frequencies by Knudsen, Pielemeier and Richardson.

534.321.9 : 577 **1050**

Physical Factors Involved in Ultrasonically Induced Changes in Living Systems: Part 1. Identification of Non-Temperature Effects.—W. J. Fry, V. J. Wulff, D. Tucker & F. J. Fry. (*J. acoust. Soc. Amer.*, Nov. 1950, Vol. 22, No. 6, pp. 867–876.) Changes in physical variables accompanying high-intensity ultrasonic disturbances in liquid media are discussed. The significance of temperature changes is investigated experimentally for nerve and other tissues of frogs and crayfish, at frequencies of 975–980 kc/s; factors other than absorption of acoustic energy, e.g. cavitation, are found to be operative. The method involves determination of the acoustic absorption coefficient of the specimens from measurements of initial rate of change of temperature.

534.756 **1051**

Theory of Operation of the Cochlea: a Contribution to the Hydrodynamics of the Cochlea.—O. F. Ranke. (*J. acoust. Soc. Amer.*, Nov. 1950, Vol. 22, No. 6, pp. 772–777.) A paper presented at the Speech Communication Conference at Massachusetts Institute of Technology, 1950.

534.756 **1052**

Theory of the Acoustical Action of the Cochlea.—J. Zwislocki. (*J. acoust. Soc. Amer.*, Nov. 1950, Vol. 22, No. 6, pp. 778–784.) A paper presented at the Speech Communication Conference at Massachusetts Institute of Technology, 1950.

534.771 **1053**

The Electrical and Acoustical Performance of some Commercial Audiometers.—K. C. Morrical, R. W. Benson & H. Davis. (*J. acoust. Soc. Amer.*, Nov. 1950, Vol. 22, No. 6, pp. 843–847.)

534.771.089.6 **1054**

Calibration of Audiometers.—E. L. R. Corliss & W. F. Snyder. (*J. acoust. Soc. Amer.*, Nov. 1950, Vol. 22, No. 6, pp. 837–842.) An account is given of the technique developed at the National Bureau of Standards. Standard sources are described, and the physical basis of the method is discussed. Limitations on the validity of the threshold data are pointed out. Location of causes of defective operation in audiometers is facilitated by reference to earphone response data.

534.78 : 519.272.119 **1055**

Correlation Function Analysis.—Kraft. (See 1179.)

534.78 : 519.272.119 **1056**

Autocorrelation Analysis of Speech Sounds.—K. N. Stevens. (*J. acoust. Soc. Amer.*, Nov. 1950, Vol. 22, No. 6, pp. 769–771.) A paper presented at the Speech Communication Conference at Massachusetts Institute of Technology, 1950. The short-time autocorrelation function is defined and is used to analyse various speech sounds. Equipment used at M.I.T. is described.

534.78 : 621.314.26 **1057**

An Apparatus for Speech Compression and Expansion and for Replaying Visible-Speech Records.—F. Vilbig. (*J. acoust. Soc. Amer.*, Nov. 1950, Vol. 22, No. 6, pp. 754–761.) A paper presented at the Speech Communication Conference at Massachusetts Institute of Technology, 1950. See also 2693 of 1950.

534.782 **1058**

An Electrical Vocal System.—L. O. Schott. (*Bell Lab. Rec.*, Dec. 1950, Vol. 28, No. 12, pp. 549–554.) An outline description of apparatus arranged to form an electrical analogue of the human vocal system. The 'vocal tract' comprises an artificial transmission line of 24 L-sections. The 'tongue hump' is an auxiliary inductor of adjustable value ('hump-height') which can be switched in series with any chosen inductive series element of the transmission line. The 'lip opening' is an adjustable inductor which terminates the line and across which the loudspeaker is connected. Separate generators are provided for producing vowels and unvoiced fricatives, while further arrangements enable pitch and inflection to be varied and other vocal features to be reproduced.

534.784 **1059**

System-Function Analysis of Speech Sounds.—W. H. Huggins. (*J. acoust. Soc. Amer.*, Nov. 1950, Vol. 22, No. 6, pp. 765–767.) A paper presented at the Speech Communication Conference at Massachusetts Institute of Technology, 1950. "The analysis of speech sound is facilitated if the sound is considered to be the response of a slow-time-varying linear system to appropriate excitations. The linear system is characterized by one or more system functions which may be represented most naturally as the sum of complex exponential terms having complex frequencies corresponding to the various formants. Thus, a generalized frequency analysis is made not of the speech wave itself but rather of the system function that shaped that wave. On the other hand, the excitation is best analyzed by autocorrelation methods."

534.861 : 534.76 **1060**

Stereophonic Broadcasting of Music.—J. Bernhart & J. W. Garrett. (*Toute la Radio*, Nov. 1950, No. 150, pp. 353–358.) Discussion of methods of obtaining 'relief' effect in sound reproduction, including an analysis of directional discrimination in audition and a description of the principle of 'directed stereophony'. Variation of intensity itself is sufficient to convey a sense of spatial movement of the sound source, particularly for low frequencies. This can be accomplished using a single microphone and two separate modulation channels the levels of which are adjusted independently: a 5-db alteration between the two levels corresponds to a radial movement of 3°, angular displacement being a linear function of the intensity difference. Phase variation has a similar effect but necessitates a more complex system in practice. Displacement in depth is conveyed by use of an echo room. The effects of combination and rapidity of these adjustments are discussed with reference to an experimental broadcast transmission from Paris [see 1061 (Aisberg)].

534.861 : 534.76 **1061**

Improved Stereophony.—E. Aisberg. (*Wireless World*, Sept. 1950, Vol. 56, No. 9, pp. 327–330.) An account of experiments in France, using simultaneous transmissions from the Parisian and Paris-Inter broadcasting stations. For reception two loudspeakers spaced 5–7 ft apart were required, the listeners being 7–10 ft away. The two receivers receive signals from separate channels and by suitably controlling the modulation depth for the two channels a realistic impression of movement of the original source of sound is given to the listeners. Modulation of the two channels was recorded on double-track magnetic tape. See also 1060 above.

621.314.2.029.3 : 621.3.018.78† **1062**

The Calculation of Waveform Distortion in Iron-Cored Audio-Frequency Transformers.—Macfadyen. (See 1087.)

621.392.52 **1063**

Unusual Ladder Filter. Applications in Audio and Radio Circuits.—Davey. (See 1112.)

621.395.623.45 **1064**

An Ultrasonic Underwater 'Point Source' Probe.—L. Fein. (*J. acoust. Soc. Amer.*, Nov. 1950, Vol. 22, No. 6, pp. 876–877.) An ultrasonic receiver uses a conical piezoelectric probe made of ADP. Its sensitivity over the frequency range 1·235–1·280 Mc/s was determined by means of a reciprocity calibration, an average value of 0·012 μV/dyne/cm^2 being obtained. The measured directional pattern does not vary with position in the field, indicating that the probe is effectively a point receiver.

621.395.625.3/.6(41) **1065**

Magnetic-Sound-Film Developments in Great Britain.—O. K. Kolb. (*J. Soc. Mot. Pict. Televis. Engrs*, Nov. 1950, Vol. 55, No. 5, pp. 496–508.) The introduction of this type of recording and reproducing apparatus in Great Britain is reviewed. Types of magnetic film available and their characteristics are described. The general circuit arrangements and details of apparatus are given together with a description of special apparatus that has been used for adding a visible record to the invisible magnetic sound-track. Jointing processes and a method for bulk erasure of records on magnetic-film stock are discussed.

621.396.645.029.3 : 621.385.3/.4 **1066**

A Comparison of Triodes and Beam-Tetrodes as Power Output Valves in Audio Amplifiers.—Bruckmann, Carey & Fuller. (See 1123.)

AERIALS AND TRANSMISSION LINES

621.392 **1067**

Phase Distortion in Feeders.—L. Lewin, J. J. Muller & R. Basard. (*Elect. Commun.*, Dec. 1950, Vol. 27, No. 4, pp. 320–323.) Reprint. See 2125 of 1950.

621.392 : 621.3.012.1/.2 **1068**

Geometrical Solution of some Transmission-Line Problems.—Q. C. Gupta & F. Duerden. (*Electronic Engng*, Dec. 1950, Vol. 22, No. 274, pp. 525–529.) A description of the vector-diagram approach, which is said to provide a better physical picture than analysis based on differential equations. The method is illustrated by an analysis of stub matching and a verification of the principles of the circle-diagram calculator.

621.392 : [621.3.015.7† : 621.314.2 **1069**

Pulse Transients in Exponential Transmission Lines.—In 280 of February, for 'E. R. Schatz' please read 'E. R. Schatz & E. M. Williams'.

621.392 : 621.3.09 **1070**

The Propagation of Transient and Periodic Waves along Transmission Lines.—R. Pélissier. (*Rev. gén. Élect.*, Sept.–Nov. 1950, Vol. 59, Nos. 9–11, pp. 379–399, 437–454 & 502–512.) A general theory of the mode of propagation and of the different causes of attenuation of electric waves is developed, using the symbolic calculus. The theory generalizes those of Carson and Pollaczek. Approximate formulae are derived and the limits of their applicability are discussed. Applications of the theory to cases of practical importance include (*a*) calculation of charge and current distribution along an infinitely long line after application of a step voltage, (*b*) corresponding calculations for a line of finite length, including discussion of reflection at terminations of various types, (*c*) reflection at line discontinuities, (*d*) discussion of causes of wave attenuation, (*e*) propagation of h.f. currents and of surges due to lightning. The theory shows that the explanation of the attenuation of such surges depending on corona effects is inexact. A new explanation is developed and confirmed by experimental results. See also 2972 of 1950.

621.392.011.2 : 621.3.012.3 **1071**

Transmission-Line Impedance Curves.—H. A. Wheeler. (*Proc. Inst. Radio Engrs*, Dec. 1950, Vol. 38, No. 12, pp. 1400–1403.) A universal family of curves gives the wave resistance, inductance and capacitance of transmission lines with 1, 2 or 4 wires. Each curve represents one field pattern. An expanded scale is given for the region of very low impedance. The seven field patterns are for the cases of conductors of circular cross-section in or near shields such as a plane, inclined or parallel planes, a trough, or a coaxial conductor.

621.392.4 **1072**

A Comparison of the Bandwidths of Resonant Transmission Lines and Lumped LC Circuits.—van Weel. (See 1107.)

621.396.67 **1073**

Theory of Collinear Antennas.—R. King. (*J. appl. Phys.*, Dec. 1950, Vol. 21, No. 12, pp. 1232–1251.) The analysis of a system of collinear aerials is practicable only if the transmission lines feeding the system are in the neutral plane of the electromagnetic field. Two cases are considered: a single centre-driven unit (*a*) with two symmetrical collinear parasitic elements, and (*b*) with the outer elements driven from the centre unit by means of phase-reversing stubs. Approximate expressions are obtained for the currents and for the mutual and self impedances of the elements. The particular case of $\lambda/2$ dipoles is treated in detail. The advantages of the collinear array with phase-reversing stubs as compared with arrays in which each element is centre driven are discussed.

621.396.67 **1074**

Elliptically Polarized Waves.—L. Hatkin. (*Proc. Inst. Radio Engrs*, Dec. 1950, Vol. 38, No. 12, p. 1455.) Comment on 3343 of 1948 (Sichak & Milazzo).

621.396.67 **1075**

On the Radiation Patterns of Dielectric Rods of Circular Cross-Section — the TM$_{0,1}$ Mode.—C. W. Horton, F. C. Karal, Jr, & C. M. McKinney. (*J. appl. Phys.*, Dec. 1950, Vol. 21, No. 12, pp. 1279–1283.) Radiation patterns were measured for five rods of diameter $d = 0{\cdot}87\ \lambda_0$ and lengths $2\ \lambda_0$ to $10\ \lambda_0$. The theoretical radiation patterns were computed by means of equivalent-surface electric and magnetic currents. Excellent agreement is obtained when it is assumed that the diameter of the surface on which these currents are distributed is $0{\cdot}65\ d$. Representative experimental and theoretical patterns are shown for the full 360°.

621.396.67 : 621.397.6 **1076**

Temporary Vision Aerial.—F. D. Bolt. (*Wireless World*, Jan. 1951, Vol. 57, No. 1, pp. 13–14.) A brief description of a temporary folded dipole used for radiating the vision signal from the London television transmitter (frequency, 45 Mc/s; mean power, 7·5 kW) while the permanent system was being overhauled. Two coaxial feeders, about 400 ft long, one being a half-wavelength longer than the other, were used as a 'binocular pair' to energize the aerial. A central element, of Al tubing, surrounded by three driven elements, gave a driving-point impedance of $168\,\Omega \pm 10\%$ over the frequency band 42–48 Mc/s. Test transmissions produced pictures nearly as good as those from the original aerial.

621.396.67.029.64 **1077**

Microwave-Aerial Radiation Patterns.—D. G. Kiely & A. E. Collins. (*Wireless Engr*, Jan. 1951, Vol. 28, No. 328, pp. 23–29.) The errors involved in measuring the side-lobe amplitudes of 3-cm fan-beam aerials are discussed. Site errors and instrumental errors are considered in detail. Experimental evidence shows that the largest errors are caused by reradiation from trees, hedges and similar obstacles near the measurement sites. For two typical field sites, when three slightly different frequencies were used, the standard deviation of site errors in measurements of side lobes in the radiation patterns of cheese aerials was 1·4 db. The design of aerials to conform to specifications for side-lobe magnitude must make allowance for such errors.

621.396.677 **1078**

An Achromatic Microwave Antenna.—N. I. Korman & J. R. Ford. (*Proc. Inst. Radio Engrs*, Dec. 1950, Vol. 38, No. 12, pp. 1455–1456.) Comment on 1345 of 1950 (Ruze).

621.396.677 : 538.56 **1079**

Reflection and Refraction of Microwaves at a Set of Parallel Metallic Plates.—Berz. (See 1138.)

621.396.677 : 621.396.932.1 **1080**

Cheese Aerials for Marine Navigational Radar.—D. G. Kiely, A. E. Collins & G. S. Evans. (*Proc. Instn elect. Engrs*, Part III, Jan. 1951, Vol. 98, No. 51, pp. 37–45.) A consideration of the problem of the suppression of side lobes in the horizontal radiation pattern of fan-beam aerials. The theory of the primary feed requirements of cheese and half-cheese aerials is outlined and the effect of the secondary radiator on the radiation of the primary feed, and hence on the distribution of energy across the aperture, is analysed. The side lobes of a symmetrical cheese aerial cannot be reduced much more than 23 db below the main-beam level. The design and performance of a half-cheese aerial having side lobes 30 db below the main-beam level are described.

621.396.67 **1081**

Antenna Theory and Design, Vols. 1 & 2. [Book Review]—H. P. Williams. Publishers: Pitman & Sons, London, 1950, 142 pp. and 522 pp., 21s. and 63s. (*J. Brit. Instn Radio Engrs*, Dec. 1950, Vol. 10, No. 12, p. vii.) The first volume, which is independent of the second, deals with the theory of aerials. The treatment is necessarily mathematical, but numerous diagrams are used in illustration of the theory. The second volume is concerned primarily with aerial design. All types are considered and many curves, diagrams and photographs are given. A full bibliography is included in both volumes.

CIRCUITS AND CIRCUIT ELEMENTS

537.312.6 : 621.315.592† **1082**

Thermistors.—Please note that the above U.D.C. number is now used for thermistors in place of 621.316.86 used hitherto.

537.312.6 : 621.315.592† **1083**

A Relationship between Resistance and Temperature of Thermistors.—G. Bosson, F. Gutmann & L. M. Simmons. (*J. appl. Phys.*, Dec. 1950, Vol. 21, No. 12, pp. 1267–1268.) The equation $\log R = A + B/(T + \theta)$ for the resistance R of a thermistor at T°K is proposed. Least-square analyses of the most precise resistance/temperature data available for three different thermistor materials show that the equation is a considerable improvement over its predecessors; the standard relative errors of fit are 0·4%, 0·17%, and 0·91%.

621.3.011.3/.4 : 621.3.012.3 **1084**

Reactance Chart.—H. A. Wheeler. (*Proc. Inst. Radio Engrs*, Dec. 1950, Vol. 38, No. 12, pp. 1392–1397.) An extension of the range of the normal chart involving reactance, frequency, inductance and capacitance, with added scales for susceptance, wavelength, and time constant. Simple geometrical patterns are given which enable the chart to be used for the direct solution of problems such as the bandwidth of a resonant circuit, wave impedance and delay of a transmission line, and design of constant-*k* filter half-sections.

621.3.015.7† : 621.3.087 **1085**

A Stable Ninety-Nine Channel Pulse-Amplitude Analyser for Slow Counting.—D. H. Wilkinson. (*Proc. Camb. phil. Soc.*, July 1950, Vol. 46, Part 3, pp. 508–518.) The input pulse is converted into a series of pulses of standard height, their number being proportional to the amplitude of the input pulse. Registration is then done by passing the series of pulses into two decade counting rings in cascade, the first ring passing a pulse to the second ring each time it has itself made a complete circuit. Attached to each element of the rings is a telephone message register which operates only when the pulse counting stops at the associated element. The analyser accepts pulses at a maximum rate of 10 per sec. Its calibration is linear to within 1%, and independent of pulse shape provided the pulse rise time is between 1 and 50 μs.

621.3.016.352† **1086**

Relation of Nyquist Diagram to Pole-Zero Plots in the Complex Frequency Plane.—W. W. Harman. (*Proc. Inst. Radio Engrs*, Dec. 1950, Vol. 38, No. 12, pp. 1454–1455.) Two criteria for the stability of a closed-loop system consisting of an amplifier of gain A with fractional feedback β are: (*a*) the complex gain $A/(1 - A\beta)$ must not have a pole in the right half of the complex-frequency plane; (*b*) the Nyquist polar plot of the loop gain $A\beta$ must not enclose the point (1, 0). The equivalence of these conditions is demonstrated by reference to a 4-stage RC amplifier.

621.314.2.029.3 : 621.3.018.78† **1087**

The Calculation of Waveform Distortion in Iron-Cored Audio-Frequency Transformers.—K. A. Macfadyen. (*Proc. Instn elect. Engrs*, Part II, Dec. 1950, Vol. 97, No. 60, pp. 809–811.) "In the design of transformers, the prediction of the amount of waveform distortion from a knowledge of the properties of the core material becomes complicated if the preceding circuit has a high impedance. A theorem is given showing that, if suitable distortion and other measurements are made on a sample core of the material to be used and the distortion factor plotted as a function of two suitably chosen variables, then this information is sufficient to enable the distortion arising from any transformer in any resistive circuit to be calculated exactly."

621.314.22 : 621.395.641.1 **1088**

Double-Humped Resonance Curves of Repeaters.—G. Schmitt & H. Schrag. (*Fernmeldetech. Z.*, Nov. 1950, Vol. 3, No. 11, pp. 422–427.) Experiment shows that the two resonance peaks in the output from the divided secondary winding of a repeater are due to the different winding capacitances of the two windings w_2 and w_3. One peak is eliminated by shunting w_2 with a capacitor, connecting the outer end of w_2 to the inner end of w_3 and earthing the outer end of w_3. Theory of the method is outlined.

621.314.224 **1089**

Dielectric Admittances in Current Transformers.—A. H. M. Arnold. (*Proc. Instn elect. Engrs*, Part II, Dec. 1950, Vol. 97, No. 60, pp. 727–732. Discussion, pp. 733–734.) For a current transformer with appreciable capacitance between windings, five ratios may be defined without knowledge of the external circuits, but only two of these ratios can be measured directly with a simple bridge. Formulae are developed for evaluation of all the ratios from simple bridge measurements, and the formulae are tested experimentally. A simple test is described for determining quickly whether the effect of capacitance is appreciable in a transformer of unknown characteristics, and methods of construction of low-capacitance transformers are described.

621.314.224 **1090**

The Effect of Capacitance on the Design of Toroidal Current-Transformers.—A. H. M. Arnold. (*Proc. Instn elect. Engrs*, Part II, Dec. 1950, Vol. 97, No. 60, pp. 797–808.) For high-accuracy operation at currents below 5 A or frequencies above 50 c/s, transformer performance is dependent not only on the magnitude of the self-capacitances of the windings and their mutual capacitance, but also on the distribution of the capacitance between windings. The normal method of layer winding gives an unsatisfactory concentrated distribution, but a form of bank winding is described which gives a symmetrical distributed capacitance between windings, resulting in improved performance. If the thickness of the insulation between windings is increased to about one-eighth of the mean depth of the secondary winding and an insulant of low permittivity is used, further improvement in performance is obtained. The performance of multi-ratio transformers is considered and good winding arrangements are suggested.

621.314.3† **1091**

Feedback in Magnetic Amplifiers.—A. S. Fitzgerald. (*Elect. Commun.*, Dec. 1950, Vol. 27, No. 4, pp. 298–319.) Reprint. See 2731 of 1949.

621.314.3† **1092**

Magnetic-Amplifier Characteristics.—L. A. Finzi & D. C. Beaumariage. (*Elect. Engng, N.Y.*, Dec. 1950, Vol. 69, No. 12, pp. 1109–1115.) Essential text of A.I.E.E. 1950 Summer and Pacific General Meeting paper. Dimensionless curves are presented for the determination of the steady-state output characteristics of magnetic amplifiers in terms of design parameters. Correction factors for the feedback ratio are calculated. Results obtained by this method agree well with experimental results for amplifiers with cores of gradually varying incremental permeability.

621.314.3† **1093**

A Theoretical and Experimental Study of the Series-Connected Magnetic Amplifier.—H. M. Gale & P. D. Atkinson. (*Proc. Instn elect. Engrs*, Part I, Nov. 1950, Vol. 97, No. 108, pp. 302–303.) Discussion of paper abstracted in 2729 of 1949.

621.314.3† **1094**

Magnetic Amplifiers.—A. G. Milnes. (*Proc. Instn elect. Engrs*, Part I, Nov. 1950, Vol. 97, No. 108, pp. 302–303.) Discussion of paper abstracted in 2728 of 1949.

621.314.3† **1095**

A Theory of the Series Transductor.—C. S. Hudson. (*Proc. Instn elect. Engrs*, Part II, Dec. 1950, Vol. 97, No. 60, pp. 751–755.) "The theory of the series transductor with natural excitation is discussed and general equations for load and control current are derived. The effect of phase shift between the supply voltage and that across the transductor is taken into account, and the transient changes which occur at the moment of transductor cut-off are explained. The effect on the transductor characteristic of the shape of the B/H curve for the core material is considered, and the analysis is extended to enable an experimental B/H curve to be used in the derivation of the transductor characteristic. It is shown that, with 100% self-excitation, the transductor characteristic approximates in shape to the upper part of the B/H curve."

621.318.4 **1096**

Design of Iron-Cored Inductances carrying D.C.—N. H. Crowhurst. (*Electronic Engng*, Dec. 1950, Vol. 22, No. 274, pp. 516–523.) A series of charts based on the properties of typical transformer iron permit the rapid design of inductors for many purposes. From the d.c. polarizing current and the required inductance and resistance, the optimum core size, number of turns and width of air gap may be readily determined. Data are tabulated for laminations of stock sizes.

621.318.423 **1097**

A 100-Kilowatt Water-Cooled Solenoid.—J. M. Daniels. (*Proc. phys. Soc.*, 1st Dec. 1950, Vol. 63, No. 372B, pp. 1028–1034.) Design data and construction details for an air-cored solenoid producing a field of 14·7 kilogauss, uniform to within 0·5% in a cylindrical volume of length 6 cm and diameter 4 cm.

621.318.423.011.3 : 621.3.012.3 **1098**

Inductance Chart for Solenoid Coil.—H. A. Wheeler. (*Proc. Inst. Radio Engrs*, Dec. 1950, Vol. 38, No. 12, pp. 1398–1400.) A simple chart relates inductance, coil over-all dimensions, and winding density. Logarithmic scales covering several decades suffice for almost all practical applications.

621.318.423.015.33 **1099**

The Natural Frequencies of Single-Layer Solenoids subjected to Voltage Surges.—B. Heller, J. Hlávka & A. Veverka. (*Bull. schweiz. elektrotech. Ver.*, 26th Nov. 1949, Vol. 40, No. 24, pp. 951–957. In German.) Formulae for the natural frequencies are developed, taking into account the mutual inductance between turns, for the two cases where the far end of the coil is (*a*) earthed or (*b*) free. In case (*a*) the spatial distribution of voltage and current can be represented to a close approximation by harmonic functions whose wavelengths are integral multiples of π, the hyperbolic terms being negligible. In case (*b*) the spatial fundamental frequency varies within wide limits. For values of $\lambda > 1$, the fundamental frequency is to a first approximation equal to $\pi/2$, but this does not hold if $\lambda < 1$, when the hyperbolic terms must also be taken into account.

621.319.45.025 **1100**

Study of Alternating-Current Electrolytic Capacitors.—W. C. van Geel & A. J. Dekker. (*Philips Res. Rep.*, Aug. 1950, Vol. 5, No. 4, pp. 250–261. In French.) Capacitors consisting of two Al plates coated with Al_2O_3 and immersed in an electrolyte are investigated by analogy with a system comprising two ideal capacitors in series, independently shunted by opposed rectifiers, also in series. The potential differences between the various parts of the actual system are then considered, taking into account the effect of the leakage current. Measurements confirming the theory are reported and oscillograms of voltages and currents are shown.

621.385.3 : 621.315.592 † : 518.4 **1101**

Graphical Analysis of Transistor Characteristics.—L. P. Hunter. (*Proc. Inst. Radio Engrs*, Dec. 1950, Vol. 38, No. 12, pp. 1387–1391.) Graphical methods are described for determining the required values of circuit components of voltage amplifiers, current amplifiers, and oscillators, from the d.c. characteristics of the transistors used.

621.392 **1102**

A Note on the Synthesis of Resistor-Capacitor Networks.—C. Belove. (*Proc. Inst. Radio Engrs*, Dec. 1950, Vol. 38, No. 12, pp. 1453–1454.) Comment on 1605 of 1950 (Bower & Ordung).

621.392 **1103**

Wide-Band Two-Phase Networks.—H. J. Orchard. (*Wireless Engr*, Jan. 1951, Vol. 28, No. 328, p. 30.) Comment on 56 of January (Johannesson). It is suggested that a still better way of treating problems of wide-band networks is to reverse the Landen transformation and work from a suitable hyperbolic function. The advantage of the method is that the larger the bandwidth ratio, the easier does the computation become, only one step being necessary for bandwidths exceeding one decade.

621.392 **1104**

Wide-Band Two-Phase Networks.—W. Saraga. (*Wireless Engr*, Jan. 1951, Vol. 28, No. 328, pp. 30–31.) In recent papers by Darlington (1359 of 1950), Orchard (1356 of 1950) and Saraga (2736 of 1950) on wide-band networks, elliptic functions are used for obtaining a Tchebycheff approximation to the ideal performance. The expressions given for this approximation differ very much in form, although defining the same mathematical relation. Their equivalence is here demonstrated by reference to certain standard formulae in the theory of elliptic functions and their transformation.

621.392 : 621.387.424† **1105**

Auxiliary Electronic Circuits for Geiger Counters.—L. Fontes & R. Moret. (*Radio franç.*, Nov. & Dec. 1950, Nos. 11 & 12, pp. 5–15 & 18–24.) Description of quench circuits, amplifiers, stabilized h.v. supply units, scaling-down circuits, including decade systems, and integrating circuits.

621.392.4/.5 **1106**

Selective Generation, Amplification or Rejection of Alternating Voltages without the aid of Oscillatory Circuits.—H. Pieplow. (*Bull. schweiz. elektrotech. Ver.*, 24th Dec. 1949, Vol. 40, No. 6, pp. 1031–1039. In German.) Starting from the conditions which must be satisfied by feedback networks, rules are established for the construction of oscillator, amplifier or rejector networks not including LC circuits. Examination of the way in which different couplings can be used shows that the sharpness of resonance or rejection of such networks tends toward a lower limit which is appreciably higher than in the case of oscillatory circuits and can only be lowered by special means not dependent on the circuit arrangement. It follows that complex circuits are especially not worth while when the frequency must be controllable.

621.392.4 **1107**

A Comparison of the Bandwidths of Resonant Transmission Lines and Lumped LC Circuits.—A. van Weel. (*Philips Res. Rep.*, Aug. 1950, Vol. 5, No. 4, pp. 241–249.) The ratio of the bandwidths obtainable in the two cases was calculated and found to depend on the product $\omega_0 Z_0 C$ (ω_0 = resonance frequency; Z_0 = characteristic impedance of line; C = terminal capacitance). The bandwidth of Lecher lines is always less than that of an LC circuit; the longer the line, the less the bandwidth. For a short-circuited $3\lambda/4$ line the bandwidth is at least four times less than for an LC circuit. The bandwidths of $\lambda/2$ and $3\lambda/4$ lines can be increased to that of a $\lambda/4$ line by introducing discontinuities in the characteristic impedance at certain points. A qualitative explanation of these effects is given which applies also to cavity resonators.

621.392.41 **1108**

Impedance Synthesis without Use of Transformers.—R. Bott & R. J. Duffin. (*J. appl. Phys.*, Aug. 1949, Vol. 20, No. 8, p. 816.) A method is outlined for the synthesis of an arbitrary impedance by series-parallel combinations of inductors, resistors, and capacitors.

621.392.41 **1109**

Impedance Synthesis distributing Available Loss in the Reactance Elements.—W. Nijenhuis. (*Philips Res. Rep.*, Aug. 1950, Vol. 5, No. 4, pp. 288–302.) A study of the synthesis of 2-pole networks whose elements possess maximal losses. Difficulties arising from mutual inductances are overcome in the way suggested by Bott & Duffin (1108 above). It is shown how losses can be introduced into coils to simulate skin effects or eddy currents. A numerical example is worked out.

621.392.5 **1110**

Influence of Losses on Attenuation of Four-Terminal Impedance Networks.—S. Brimberg. (*Kungl. tekn. Högsk. Handl., Stockholm*, 1950, No. 43, 30 pp. In English.) The influence of losses on complex image attenuation in the case of infinite idealized image attenuation and at band edges is calculated. The effect of losses on complex effective attenuation at band edges in symmetrically designed filters with symmetrical loads is also investigated and theoretical and experimental results are compared for a low-pass filter.

621.392.52 **1111**

Attenuation Equalization within the Pass Band of Ladder Filters.—W. Wolman. (*Arch. Elektrotech.*, 1950, Vol. 40, No. 1, pp. 30–36.) Analysis shows that the frequency-response curve of each section of a filter can be flattened by introducing small resistances in shunt or in series, effectively forming auxiliary meshes with complementary response curves. Measurements are reported in confirmation of the theory, and the effect of the compensation is illustrated in response curves for a band-pass filter of 6 half-sections and for a low-pass filter of 4 sections. Factors affecting the choice of location of the auxiliary resistors, a matter of great practical importance, are discussed.

621.392.52 **1112**

Unusual Ladder Filter. Applications in Audio and Radio Circuits.—F. G. G. Davey. (*Wireless World*, Jan. 1951, Vol. 57, No. 1, pp. 31–34.) The properties are discussed of a class of filters in which the impedances looking forward and backward from the central element bear a constant ratio to each other with change of frequency and maintain equal phase angles. The application is described of adjustable low-pass filters of this type between the speech coil of a loudspeaker and the corresponding transformer winding, and of band-pass versions for use as variable coupling elements at intermediate frequency.

621.392.52 : 621.395.44 **1113**

Loss-Compensated Filters in Carrier-Frequency Systems.—H. Lehmann. (*Fernmeldetech. Z.*, Nov. 1950, Vol. 3, No. 11, pp. 415–417.) Discussion of the advantages of using asymmetric filters in s.s.b. carrier systems, particularly in eliminating the need for l.f. equalizers, in affording a choice of design, and in improving stability. The discussion is based on experimental results for channel filters covering the range 12–24 kc/s.

621.396.611.3 **1114**

Rolling-Ball Analogue of Coupled Circuits.—P. C. Magnusson. (*Elect. Engng, N.Y.*, Dec. 1950, Vol. 69, No. 12, p. 1079.) Summary of A.I.E.E. 1950 Fall General Meeting paper. The system consisting of two inductively coupled LC loops with negligible resistance has as dynamic analogue a ball rolling freely on a specially curved surface. The time integrals of the loop currents are used as displacements along horizontal co-ordinates x_1, x_2, with the loop currents themselves analogous to the velocity of the ball along those co-ordinates. The magnetic-field energy is analogous to the kinetic energy of the ball if the scales of x_1 and x_2 are proportional to the square roots of the corresponding self-inductances and x_1 is inclined to x_2 at an angle β, the cosine of which is equal to the coefficient of coupling. The height of any point of the surface, and hence the potential energy of the ball, is proportional to the dielectric field energy, contours of which are concentric ellipses. The ball path is shown which corresponds to initial circuit conditions where both capacitors are charged and circuit currents are zero. The method of deriving an analogue should be applicable to any dynamically conservative electrical or mechanical system with two degrees of freedom, and should be useful in the interpretation of transients in such systems.

621.396.611.31.001.11 **1115**

Theory of Oscillations in Coupled Electromagnetic Cavity Resonators.—E. Ledinegg & P. Urban. (*Acta phys. austriaca*, Sept. 1948, Vol. 2, No. 2, pp. 198–213. In German.) General formulae are derived by a perturbation method for the frequencies of a system of cavity resonators coupled by similar resonant members. Except for the assumption of weak coupling between the resonators, which in the undisturbed state must give rise to at least one common frequency, no essential limitations are introduced in the calculations. The theory shows that the frequencies of such coupled systems satisfy a secular determinant whose members represent energy terms determined by the individual elements of the cavity-resonator system and which can accordingly be calculated. The theory is applied to the practical cases of (*a*) two concentric transmission-line sections with a coaxial coupling unit, (*b*) two inductively coupled quasistationary circuits.

621.396.615 **1116**

The Self-Blocking Oscillator and its Applications.—J. Moline. (*Radio franç.*, Dec. 1950, No. 12, pp. 11–15, & Jan. 1951, No. 1, pp. 12–16.) The development of the blocking oscillator from one having a sinusoidal waveform, by increasing the coupling between the anode and grid circuits, is described and the blocking action is explained, the effects of the different circuit elements being considered separately. Various typical practical circuits are shown and their particular advantages are mentioned. Applications described include pulse and sawtooth generators, use in counter circuits, etc.

621.396.615 : 621.316.726 : 621.396.4 **1117**

Stabilized Master Oscillator for Multichannel Communication.—E. W. Pappenfus. (*Electronics*, Dec. 1950, Vol. 23, No. 12, pp. 108–113.) Particular attention is paid to systems providing a wide range of operating frequencies but utilizing only a single reference crystal. A commercially available oscillator covering the band 2–4·5 Mc/s is described. Using a 100-kc/s reference crystal, the oscillator provides an output of any desired frequency in the above range stabilized to within $\pm 5 \times 10^{-6}$ for wide temperature and humidity variations. Details of the servomotor a.f.c. circuit are given and discussed. Any one of 10 preset output frequencies can be selected by operation of a remote switch.

621.396.615 : 621.316.726.078.3 **1118**

Frequency Stabilization of Oscillators.—J. Zakheim. (*Radio franç.*, Nov. 1950, No. 11, pp. 1–4.) If the effective output impedance Z of the amplifier section is given by $Z = R + jX$ and the transfer coefficient k of the limiter by $k = \alpha + j\beta$, the phase angle is theoretically constant and independent of the overall slope of the maintaining amplifier when $\beta/\alpha = -X/R$. Practical application of the formulae derived is illustrated by numerical calculation for an oscillator of the Wien-bridge type operating at 3 kc/s.

621.396.615.14.029.63 **1119**

Oscillators for Decimetre Waves with Disk-Seal Valves in Grounded-Grid Circuits.—Ratheiser. (See 1280.)

621.396.615.141.2 **1120**

Some Aspects of Split-Anode Magnetron Operation.—H. J. Reich, J. C. May, J. G. Skalnik & R. L. Ungvary. (*Proc. Inst. Radio Engrs*, Dec. 1950, Vol. 38, No. 12, pp. 1428–1433.) Resonators and valves at present available are unsuitable for use in wide-band magnetron oscillators using butterfly-type resonators, but they could be more successfully used if specially modified for the purpose. The negative slope of part of the anode-current/voltage characteristic, explainable by back heating, is absent when the current is varied rapidly. The best method of anode-current stabilization is that in which the magnetic field is produced by an electromagnet excited by the direct anode current.

621.396.615.17 : 621.385.832 **1121**

A Simple Slow-Running Timebase.—C. J. Dickinson. (*Electronic Engng*, Dec. 1950, Vol. 22, No. 274, pp. 505–506.) A detailed description of a circuit designed to overcome inherent disadvantages of earlier timebases. The circuit requires less critical adjustment and avoids the need for a large negative supply voltage.

621.396.645 : 621.3.018.78† **1122**

Universal Design Curves for Intermediate-Frequency Amplifiers with Particular Reference to Phase and Amplitude Distortion and their Compensation.—W. Hackett. (*Proc. Inst. Radio Engrs*, Dec. 1950, Vol. 38, No. 12, pp. 1408–1417.) Graphs and abacs are provided which enable the phase and amplitude distortion to be determined for various types of coupling network, including compensation combinations. Graphical methods are indicated for assessment of performance in terms of the product of gain and bandwidth, and for the interpretation of equation parameters in terms of circuit parameters.

621.396.645.029.3 : 621.385.3/.4 **1123**

A Comparison of Triodes and Beam-Tetrodes as Power Output Valves in Audio Amplifiers.—D. Bruckmann, W. S. Carey & D. J. Fuller. (*Trans. S. Afr. Inst. elect. Engrs*, Sept. 1950, Vol. 41, Part 9, pp. 258–266. Discussion, pp. 266–267.) Tests made on balanced push-pull amplifiers show that the curve representing the change of harmonic distortion with increase in load resistance is parabolic for beam tetrodes, while for triodes it resembles a rectangular hyperbola. With reactive loading the increase in distortion is greater for the tetrode. As the load resistance rises the intermodulation distortion for triodes drops to a negligible value, while for tetrodes it increases rapidly. The effects with reactive loading are similar to those observed in the case of harmonic distortion. The relative merits of triodes and tetrodes for use in a.f. amplifiers are summarized.

621.396.645.371 **1124**

Amplifiers with Selective Negative Feedback.—H. Boucke & O. Schmidt. (*Fernmeldetech. Z.*, Nov. 1950, Vol. 3, No. 11, pp. 417–420.) Basic circuits are described for suppressing either a harmonic or the fundamental in the output of an amplifier by incorporating a resonant circuit in the feedback line. Tuning this circuit to the first harmonic reduces the distortion appreciably. A typical attenuation curve for a frequency-multiplier circuit shows a drop of 28 db in output level for the 800-c/s fundamental below that obtained with a non-selective cathode circuit.

621.396.662.085.4 **1125**

Bandspreading and Scale Equalization for RC Tuning Networks.—C. F. Van L. Weiland. (*TV Engng, N.Y.*, July, Aug. & Oct. 1950, Vol. 1, Nos. 7, 8 & 10, pp. 12–15, 28–30 & 21, 28.) A continuous-selector type of circuit is described. By use of suitable tuning elements in series and parallel, a threefold scale elongation is achieved, with increased spread at the h.f. end of the scale. Circuits providing partially equalized and completely equalized bandspreading are shown and design formulae given.

621.396.69 + 621.38] : 061.5(058.7) 1126

Alphabetical Listings of All Components, Complete Units, Allied Products, used in Electronic Equipment for All Purposes.—(*Electronics, Annual Buyers' Guide Issue*, Mid-month June 1950, Vol. 23, No. 6A, pp. D1–D164.) A list giving the names of more than 2 500 manufacturers of electronic components and equipment. Products are arranged alphabetically under generic names, with subdivisions. A trade-name index and an index of manufacturers are also included.

GENERAL PHYSICS

519.2 : 621.3.015.2 1127

On a First-Passage-Time Problem.—F. L. H. M. Stumpers. (*Philips Res. Rep.*, Aug. 1950, Vol. 5, No. 4, pp. 270–281.) An investigation of the probability that a function starting at a time $t = 0$ with a value E_0 will not exceed a value E_1 in the time interval $0–t_1$. This probability is calculated (*a*) from the Fokker-Planck equation with suitable boundary conditions, (*b*) from an integral equation derived by Schrödinger. The theory has application to the pulse charging of a capacitor and also to fluctuation problems.

534.1 + 538.56] : 517.93 1128

On the Forced Vibrations of Quasi-Linear Systems.—Friedlander. (See 1177.)

534.26 + [535.42 : 538.56 1129

Theory of Diffraction at a Screen.—W. Franz. (*Z. Phys.*, 3rd Oct. 1950, Vol. 128, No. 3, pp. 432–441.) The author's successive-approximation method for diffraction calculations (3118 of 1949 and 323 of February) yields, for sharp-edged screens, functions which are singular at the edge and in the higher approximations give rise to divergent integrals. The singularities can be eliminated and the method made useful for higher approximations by assuming the integrands of the Kirchhoff integrals to fall to zero gradually, instead of discontinuously, beyond the edge. See also 2183 of 1950 (Braunbek).

535.42 1130

A Note on the Diffraction of a Cylindrical Wave by a Perfectly Conducting Half-Plane.—P. C. Clemmow. (*Quart. J. Mech. appl. Math.*, Sept. 1950, Vol. 3, Part 3, pp. 377–384.) The most easily derived solutions in diffraction theory are for the case of incident plane waves. The solution for a line source can then be obtained by a further integration. This is illustrated by the example of a line source in the presence of an infinitely thin, perfectly conducting half-plane. The introduction of Hankel functions is avoided, and the solution appears in a useful form, equivalent to that given by Macdonald and analogous to Sommerfeld's solution for an incident plane wave.

537.52 1131

Retroaction and Similitude Considerations in Relation to the Starting of the Electrical Gas Discharge.—W. Fucks. (*Arch. Elektrotech.*, 1950, Vol. 40, No. 1, pp. 16–30.)

537.523.5 : 621.317.33.029.64 1132

A Microwave Study of the High-Pressure Arc.—J. D. Cobine, E. P. Cleary & W. C. Gray. (*J. appl. Phys.*, Dec. 1950, Vol. 21, No. 12, pp. 1264–1267.) The impedance of an arc at atmospheric pressure was measured at 1 kMc/s by means of a coaxial line and standing-wave detector. The arc reactance and resistance, essentially a lumped load at the end of the line, increase with arc length. The resistance decreases with increasing current and is approximately the same as the d.c. resistance. The reactance is capacitive and nearly independent of the current. Arcs in air, argon, and helium, with direct currents of 1–4 A, were studied.

537.525 : 621.396.822 1133

Noise Temperature of a D.C. Gas-Discharge Plasma.—S. Kojima & K. Takayama. (*Phys. Rev.*, 1st Dec. 1950, Vol. 80, No. 5, p. 907.) At 14 Mc/s, the temperature of the plasma noise of discharge tubes containing argon at pressures of 1 mm and 3 mm Hg is found to be proportional to the electron temperature. The ratio of plasma-noise temperature to electron temperature is approximately 4·5 : 1 at 1 mm and 2 : 1 at 3 mm Hg. The noise temperature is nearly the same at 600 kc/s as at 14 Mc/s, indicating that the noise observed is mainly thermal.

537.525.029.64 : 538.69 1134

The Effect of Magnetic Field on the Breakdown of Gases at Microwave Frequencies.—B. Lax, W. P. Allis & S. C. Brown. (*J. appl. Phys.*, Dec. 1950, Vol. 21, No. 12, pp. 1297–1304.) Results are given of measurements of the electric field required to produce h.f. breakdown in helium at low pressure in the presence of a magnetic field of 0–3 000 gauss. Two relevant theories, the average electron theory and the Boltzmann theory, are presented, and the correspondence between them is discussed.

537.525.5 1135

Techniques for Measuring the Dynamic Characteristics of a Low-Pressure Discharge.—B. T. Barnes & S. Eros. (*J. appl. Phys.*, Dec. 1950, Vol. 21, No. 12, pp. 1275–1278.)

537.527 1136

The Minimum Voltage and the Discharge Process on Flashover with Alternating Voltage in Humid Compressed Air.—H. Böcker. (*Arch. Elektrotech.*, 1950, Vol. 40, No. 1, pp. 37–44.) Measurements were made of the flashover voltage in air at pressures up to 24 atm with water-vapour content close to saturation, cylindrical glass insulators 0·6 to 3 cm long being inserted between the dished electrodes of the plane parallel gap. The lowest value of the reduced flashover voltage observed in the region of saturation is the 'minimum voltage'. Results are shown graphically and discussed.

538.3 1137

Wheeler & Feynman's Theory of Electromagnetic Interaction.—O. Costa de Beauregard. (*Rev. sci., Paris*, Jan./March 1950, Vol. 88, No. 3305, pp. 34–40.) Analysis in the light of other theories leads to criticism on the grounds that conditions for the irreversibility of the radiation process are contained in the postulates of the theory.

538.56 : 621.396.677 1138

Reflection and Refraction of Microwaves at a Set of Parallel Metallic Plates.—F. Berz. (*Proc. Instn elect. Engrs*, Part III, Jan. 1951, Vol. 98, No. 51, pp. 47–55.) A mathematical analysis of the problem based on Maxwell's equations and on periodicity and continuity considerations. "Formulae are obtained giving the nature, direction, phase and amplitude of the reflected and transmitted waves. The reflected waves follow the usual laws found in the grating theory. In the case when only one non-evanescent plane wave is reflected, the transmission and reflection power coefficients are those obtained at the junction of two semi-infinite transmission lines corresponding to the free-space and the plate medium respectively. The analogy with transmission lines is more complex when two non-evanescent waves are reflected. For the usual angles of incidence and plate spacings the transmission power coefficient is high (maximum 99·5%), and the phase shift on transmission is small."

538.561 **1139**

Radio-Frequency Emission due to the Gyromagnetic Effect in a Discharge.—M. Laffineur & C. Pecker. (*C. R. Acad. Sci., Paris*, 18th Dec. 1950, Vol. 231, No. 25, pp. 1446–1447.) The source of electrons used was a Penning vacuum gauge (1525 of 1937) with two cold cathodes, one on each side of a common ring-shaped anode. This was evacuated, placed in the field of an electromagnet producing fields up to 300 gauss, and a voltage of 2 kV was applied. The e.m. waves emitted were detected by means of a receiver tuned to a wavelength of 55 cm and connected to a pickup loop at the side of the apparatus. The best results were obtained at a pressure of 10^{-4} mm Hg. A curve shows the voltage at the receiver terminals, less the background noise (equivalent to about 50 μV at the input), for fields from 184·5 to 194·5 gauss. This curve has a very sharp peak at 193·5 gauss, the corresponding gyromagnetic frequency being precisely that for which the receiver was tuned. A much smaller peak at about 190 gauss is attributed to an image frequency of the receiving apparatus.

549.321.13 : 537.311.3 : 537.533.9 **1140**

Equilibrium Currents induced in Zinc Blende by Electron Bombardment of Negative Electrode.—M. F. Distad. (*Phys. Rev.*, 1st Dec. 1950, Vol. 80, No. 5, pp. 879–886.)

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.7 : 538.122 **1141**

Visual Investigation of the General Magnetic Field of the Sun.—G. Thiessen. (*Z. Astrophys.*, 31st July 1949, Vol. 26, No. 1, pp. 16–30.) The method of measurement and the results obtained during 1947–1948 with the 60-cm refractor of the Hamburg observatory are described. Measurements on various Cr and Fe spectral lines show that at the time of the observations the pole field strength in the direction given by Hale could not have a value $>$ 5 gauss; the results rather indicate a very weak field in the reverse direction.

523.7 + 523.854] : 621.396.822 **1142**

Electromagnetic Waves radiated by Fast Protons in Strong Magnetic Fields, and Correlation between Cosmic Radiation and Radio Noise from the Sun and the Galaxy.—B. Kwal. (*C. R. Acad. Sci., Paris*, 13th Nov. 1950, Vol. 231, No. 20, pp. 1057–1059.) The classical theory of radiation from a charged particle in a magnetic field shows that, in fields of the order of 10^3–10^4 gauss, protons of energy 10^9 to 10^{10} eV radiate metre and centimetre waves. Hence it seems possible to associate radio noise with the acceleration of cosmic rays in sunspots and in similar spots occurring in stars.

523.752 : [551.510.535 + 537.591 **1143**

The Chromospheric Flare of 19th November 1949 and its Geophysical Consequences.—R. Bureau & A. Dauvillier. (*Ann. Géophys.*, April/June 1950, Vol. 6, No. 2, pp. 77–103.) Close correlation was observed between ionospheric and cosmic-radiation effects associated with this flare. The rarity of such a correlation is emphasized. The two authors describe respectively their detailed investigations of the two effects. Observations from various parts of the world are reported and discussed, especially in relation to possible explanations of the delay of the corpuscular effects with respect to the electromagnetic effects. See also 1911 (Ellison & Conway), 2206 (Müller) and 3039 (Clay & Jongen) of 1950.

550.362 : 621.317 **1144**

Theoretical Basis of a Method for Determining the Specific A.C. Resistivity of the Ground.—Weber. (See 1193.)

550.386/.387] : 621.331 : 625.1 **1145**

Disturbances of Terrestrial Magnetic Field and Currents by Electrified Railways.—G. Dupouy. (*Ann. Géophys.*, Jan./March 1950, Vol. 6, No. 1, pp. 18–50.) Disturbances observed in the magnetograms recorded at Chambon-la-Forêt have been found to be caused by the Paris-Orléans line, 28 km away. This effect is particularly noticeable when the continuity of the electric cables is broken, either accidentally or for purposes of maintenance. For horizontally stratified ground the disturbances can be calculated as a function of the leakage currents along the track. Agreement with measurements is satisfactory.

551.510.534/.535 **1146**

Stratospheric-Ionospheric Relationships.—N. C. Gerson. (*Proc. Inst. Radio Engrs*, Dec. 1950, Vol. 38, No. 12, p. 1456.) Investigations reveal no correlation between either the temperature or pressure at the tropopause and the critical frequency of any of the ionospheric regions, and also no correlation between temperatures at a height of 13 km and critical frequencies. Some slight correlation was found in several instances between the pressure at a height of 13 km and the critical frequencies of the E and F_2 regions, but no correlation with sporadic-E conditions.

551.510.535 **1147**

Magneto-ionic Triple Splitting of Ionospheric Waves.—O. E. H. Rydbeck. (*J. appl. Phys.*, Dec. 1950, Vol. 21, No. 12, pp. 1205–1214.) A full theoretical discussion of the third magneto-ionic component, or *z*-trace, which has been observed frequently at the new Kiruna observatory. Strong coupling often exists between the ordinary and *z* components at a critical level. The *z*-wave becomes strongly excited when the collisional frequency ν approaches the critical value ν_c of Appleton & Builder. When ν is larger than ν_c the ordinary echo rapidly disappears and only the *z* and extraordinary components remain. A useful expression for the *o-z*-wave transmission coefficient is derived. Transmission of appreciable strength is possible only at high latitudes, where ν_c is fairly small. Experimental results obtained at Kiruna (magnetic latitude 65°) are reported.

551.510.535 **1148**

Preliminary Results of Ionospheric Observations made at Dakar.—F. Delobeau, E. Harnischmacher & F. Oboril. (*Rev. sci., Paris*, Jan./March 1950, Vol. 88, No. 3305, pp. 17–20.) Discussion of records obtained during the period May–September 1949. Monthly mean values of f_0F_2 conform to the general shape of curves plotted for comparable latitudes, but extreme values are more pronounced: the highest critical frequencies in May were estimated at about 15 Mc/s; median midday values for June–July were 13·8–13·5 Mc/s and the night value for June was 4·2 Mc/s. This lends support for the revision of zone demarcation proposed by Oboril & Rawer (2794 of 1949). Strong diffusion effects observed cannot be correlated with magnetic disturbances. The presence of an intermediate layer between F_1 and F_2 is indicated; the true height of the F_2 layer, deduced from virtual-height measurements, may reach 500 km. Values of k and n in the formula $f_0E = k(\cos\chi)^n$ are tabulated. Records of E_s-layer characteristics are also shown and discussed.

551.510.535 : 523.745 **1149**

Long-Term Relation between F_2-Layer Ionization and Solar Activity, over the Whole World.—R. Gallet & K. Rawer. (*Ann. Géophys.*, April/June & July/Sept. 1950, Vol. 6, Nos. 2 & 3, pp. 104–114 & 212–213.) Close correlation is found between the monthly mean value of the F_2-layer maximum electron density and the relative sunspot number, on comparing the running means over 13 months. The coefficients of the linear correlation equations for 12 stations in different parts of the world are plotted against (*a*) geographic latitude, (*b*) geomagnetic latitude, and (*c*) magnetic inclination; the last method gives the most satisfactory picture of their systematic variation. Abnormal relations are observed at the magnetic equator; observations from stations in that region are reserved for separate study.

551.510.535 : 551.594.5 **1150**

Ionospheric Disturbance caused by the Aurora Borealis of 20th-21st February 1950.—R. Eyfrig. (*Ann. Géophys.*, Jan./March 1950, Vol. 6, No. 1, pp. 70–73.) A direct observation of an aurora borealis, relatively rare in latitudes as low as 48°, was made at Freiburg, starting at 1845 universal time on the 20th February. Unusually high electron concentration in the F_2 layer preceded and accompanied the observation. Further abnormalities of the E and F layers are shown in a series of *h'f* records for the period from 1830 to 2200. The disturbance is interpreted as a horizontally travelling dip in the ionosphere. f_0F_2 was low throughout the night and until the next afternoon. A marked ionospheric disturbance was also observed in the Antarctic on board the *Commandant Charcot* on 20th February.

551.510.535 : 551.594.51 **1151**

New Results of Importance in the Study of Auroral Spectra and the Physics of the Ionosphere.—L. Vegard. (*Ann. Géophys.*, July/Sept. 1950, Vol. 6, No. 3, pp. 157–163.) Results more complete and more precise than ever before have been obtained at Oslo, using a new spectrograph for analysis of the aurora of 23rd–24th February 1950. Fifty-four new bands or lines were observed. The presence of molecular oxygen was proved for the first time. Many lines are due to nitrogen and oxygen, neutral and ionized, as previously observed. The temperature of the luminous region was estimated to be − 54°C.

551.510.535 : 621.3.087.4 **1152**

An Automatic Ionospheric Recorder for the Frequency Range 0·55 to 17 Mc/s.—Naismith & Bailey. (See 1191.)

551.594.11(988) **1153**

Measurements of the Electric Field of the Atmosphere in Greenland between Sea Level and the Centre of the Ice-Cap.—P. Pluvinage & G. Taylor. (*Ann. Géophys.*, Jan./March 1950, Vol. 6, No. 1, p. 69.)

LOCATION AND AIDS TO NAVIGATION

621.396.9 : 526.9 **1154**

The Application of Radar to Geodetic Surveying.—J. Warner. (*Aust. J. appl. Sci.*, June 1950, Vol. 1, No. 2, pp. 133–146.) A modified shoran radar equipment is used for measuring distances between 160 and 310 miles to an accuracy within about 7 parts in 10^5. The method of measurement, equipment used, reduction of observations, and errors involved are described. The greatest errors were instrumental; it is considered that modifications of the equipment would reduce them to about 2 parts in 10^5.

621.396.9 : 551.578.1/.4 **1155**

Radar Echoes from Meteorological Precipitation.—J. E. N. Hooper & A. A. Kippax. (*Proc. Instn elect. Engrs*, Part I, Nov. 1950, Vol. 97, No. 108, pp. 303–304.) Discussion of paper abstracted in 2222 of 1950.

621.396.9 : 621.396.11 **1156**

Some Adverse Influences of Meteorological Factors on Marine Navigational Radar.—Saxton & Hopkins. (See 1228.)

621.396.932.1 : 621.396.677 **1157**

Cheese Aerials for Marine Navigational Radar.—Kiely, Collins & Evans. (See 1080.)

621.396.933.2 **1158**

Rho-Theta System of Air Navigation.—P. C. Sandretto. (*Elect. Commun.*, Dec. 1950, Vol. 27, No. 4, pp. 268–276.) Description of equipment designed to fulfil I.C.A.O. recommendations for short-range navigational aids. Azimuth (theta) is obtained by reception of signals from a ground v.h.f. transmitter radiating (*a*) a carrier with 30-c/s tone modulation from an aerial having a circular horizontal radiation pattern, and (*b*) an unmodulated carrier of the same frequency radiated from a dipole rotating at 1 800 r.p.m. The phase relations of the two carriers remain rigidly fixed. Azimuth is obtained from the measurement of the phase difference between the tone from the rotating system and that from the omnidirectional system. Distance (rho) is determined from the time taken by a pulse radiated from the aircraft at u.h.f. to travel to a ground station and be returned, at a different u.h.f., to the aircraft.

MATERIALS AND SUBSIDIARY TECHNIQUES

531.788.12.088 **1159**

Reading Errors with the McLeod Gauge.—P. Romann. (*Le Vide*, Nov. 1948, Vol. 3, No. 18, pp. 522–530.)

533.56 **1160**

Study and Realization of a New Rotary Molecular Vacuum Pump.—H. Gondet. (*Le Vide*, Nov. 1948, Vol. 3, No. 18, pp. 513–521.) Description of a pump of the Holweck type in which a disk is used instead of the usual drum, so that an increased output is obtained. The optimum dimensions for the various channels are calculated and pumping rates are given for an experimental pump with a duralumin disk, of diameter 40 cm and thickness 1·2 cm, rotating between two duralumin castings with spiral channels of cross-section rapidly decreasing toward the centre. Clearance between the disk faces and the partitions separating the channels is of the order of 0·03 mm. The driving motor is fitted within the part of the machine connected to the backing pump.

537.311.33 : 546.289 **1161**

Pressure Dependence of Resistance of Germanium.—J. H. Taylor. (*Phys. Rev.*, 1st Dec. 1950, Vol. 80, No. 5, pp. 919–920.) The application of hydrostatic pressure increases the width of the forbidden band and hence increases the resistivity. Measurements, at pressures up to 4 500 lb/in.², of the change of resistivity with pressure for several samples leads to the value $10{\cdot}2 \pm 0{\cdot}4 \times 10^{-5}$/atm for the pressure coefficient of resistance at 300°K. The temperature rate of change of width of the forbidden band then follows as $-0{\cdot}87 \times 10^{-4}$ eV/°C.

537.311.33 : 546.289 : 539.164.9 **1162**

Changes in Conductivity of Germanium induced by Alpha-Particle Bombardment.—W. H. Brattain & G. L. Pearson. (*Phys. Rev.*, 1st Dec. 1950, Vol. 80, No. 5, pp. 846–850.) The bombardment of *n*-type Ge by α particles first removes the conduction electrons at the rate of 78 per α particle. After these electrons are gone, conducting holes are introduced at the initial rate of 8·6 per α particle. 6·6 of these conducting holes decay with a half-life of 13 hours at room temperature and the remaining two holes are permanent at this temperature. Conductivity is changed only to the depth of penetration of the particles, namely $1{\cdot}9 \times 10^{-3}$ cm. The distribution of holes with depth is not uniform.

537.312.5 : 546.46.86 **1163**

Photoconductivity in Magnesium Antimonide Layers.—T. S. Moss. (*Proc. phys. Soc.*, 1st Dec. 1950, Vol. 63, No. 372B, pp. 982–989.) Layers of Mg_3Sb_2 evaporated in vacuo act as semiconductors and are photoconductive.

Current/voltage and radiation-intensity/voltage relations, response time, spectral dependence of photosensitivity and the resistance/temperature relation were investigated. Threshold wavelengths of $1{\cdot}5\,\mu$, $2{\cdot}6\,\mu$ and $3{\cdot}5\,\mu$ were found, the first corresponding to intrinsic conductivity ($\epsilon = 0{\cdot}81$ eV) and the latter two ($\epsilon = 0{\cdot}48$ eV and $0{\cdot}35$ eV) to impurity levels. Recent Russian work is in agreement with these results [*Zh. tekh. Fiz.*, 1949, Vol. 18, pp. 1459–1477 (Boltaks & Zhuze)].

537.312.8 : 669.15 **1164**

A Study of the Magnetoresistance of Silicon-Iron.—R. Parker. (*Proc. phys. Soc.*, 1st Dec. 1950, Vol. 63, No. 372B, pp. 996–1004.) Curves for $\Delta\rho/\rho$ against field strength were obtained for single crystals and polycrystalline specimens, for both longitudinal and transverse magnetization. Strong fields give a linear decrease less than 0·02 times that for Ni and Co. Weak fields give unusual features. It is concluded that the small magnetoresistance of polycrystalline silicon-iron is due to the small contribution of individual crystallites and not to an averaging-out process.

537.312.8.029.63 : 546.87 **1165**

Magnetoresistance of Bismuth at 3 000 Mc/s.—C. W. Heaps. (*Phys. Rev.*, 1st Dec. 1950, Vol. 80, No. 5, pp. 892–893.) Using a resonant re-entrant cavity machined from cast Bi, fed through a slotted coaxial waveguide, s.w.r. measurements indicated that at 3 kMc/s the magnetoresistance of Bi is not more than half the d.c. value. This is possibly due to the skin depth being of the same order as the mean free path of electrons in the metal.

538.221 **1166**

Oxide Ferromagnetic Materials.—E. Flegler. (*Arch. Elektrotech.*, 1950, Vol. 40, No. 1, pp. 4–16.) An account is given of the experimental investigation of various γ-iron oxides and ferrites subjected to alternating fields. The variations with frequency of the conductivity, permittivity, permeability and loss factors are shown and conclusions are drawn regarding the causes of these variations. The different uses of the term 'complex permeability' are briefly discussed.

538.221 **1167**

High-Frequency Permeability of Ferromagnetic Materials.—F. V. Webster & K. S. Driver. (*Proc. phys. Soc.*, 1st Dec. 1950, Vol. 63, No. 372B, p. 1040.) Preliminary results obtained with more accurate apparatus than that previously used [913 of April (Millership & Webster)] are reported. The general conclusions remain unaffected.

538.221 : 538.214 **1168**

Investigations on the Reversible Susceptibility of Ferromagnetics.—R. S. Tebble & W. D. Corner. (*Proc. phys. Soc.*, 1st Dec. 1950, Vol. 63, No. 372B, pp. 1005–1016.) The reversible susceptibility κ_r of long wire specimens was measured by applying small alternating fields and using a mutual-inductometer bridge. Results are discussed. The minimum contribution of reversible processes to the total change in magnetization varies from 10% to 20% for iron and nickel specimens. Gans' statement, that κ_r is a unique function of intensity of magnetization and independent of magnetic history, has definite limitations.

549.514.51 : 621.396.611.21 **1169**

Quartz Crystals free from Harmonics.—J. J. Vormer. (*Tijdschr. ned. Radiogenoot.*, Nov. 1950, Vol. 15, No. 6, pp. 339–346.) Overtones of Y′ oscillations in X-cut crystals can be suppressed by shaping the width of the electrodes to a sine curve; this has the effect that the areas subjected to opposed fields are exactly equal for any harmonic. Experiments made to test the theory indicated that the intensity of all harmonics could certainly be reduced to less than 1% of the fundamental. Alternatively, a field distributed sinusoidally in the Y direction may be used with ordinary electrodes to achieve the same effect.

621.314.634 : 537.311.33 **1170**

Forming Processes in Thallium-Containing Selenium Rectifiers.—A. Hoffmann. (*Z. Phys.*, 3rd Oct. 1950, Vol. 128, No. 3, pp. 414–431.) A new method of measuring the capacitance of a dry rectifier, due to Lehovec, makes it possible to estimate the concentration of impurity centres at different production stages. For a certain Tl content of the counterelectrode there is (*a*) a reduction of concentration of impurity centres in the boundary zone, (*b*) an increase of path resistance, and (*c*) an increase of blocking power; (*a*) and (*b*) are quantitatively related. The influence of the Tl is markedly dependent on whether the counterelectrode is absent or present during the final heat treatment at 210°C. The increase of blocking power cannot be explained simply by the reduction of the boundary field strength, but must be mainly due to other causes.

621.318.2 **1171**

Calculation and Construction of Permanent-Magnet Systems.—E. Steinort. (*Radio Tech., Vienna*, Nov. 1950, Vol. 26, No. 11, pp. 545–548.) The design affording maximum flux in the air gap with minimum use of material is discussed. Formulae are given for optimum length and cross-section of the magnet; the correction factor for length and the leakage flux as a function of air-gap inductance are shown graphically for ticonal and alnico materials. When the leakage characteristic of the material is not known, optimum length and cross-section may be determined by an experimental method using an electromagnet.

621.791.3 **1172**

Solders and Soldering: Some Recent Advances.—H. C. Watkins. (*Metallurgia, Manchr*, Dec. 1950, Vol. 42, No. 254, pp. 372–376.) Survey of materials and techniques, with special reference to wiping solders, the soldering of aluminium, special solders for automatic processing, and flux-cored solders. The physical effects of surface roughening and of absorption of atoms of the liquid metal are briefly considered.

666.1.037.5 **1173**

Glass-to-Metal Seals.—J. A. Pask. (*Product Engng*, Jan. 1950, Vol. 21, No. 1, pp. 129–134.) A discussion of the properties of various metals and special alloys used for sealing to glass. Only direct seals, with no intermediate material other than the metal oxide, are considered. Thermal expansion curves are given and the stresses arising from imperfect matching between metal and glass are analysed. Other important factors are the electrical and thermal conductivities of the metal, and the thickness of the oxide film allowed to form on its surface.

621.317.2 : 537.72 **1174**

Techniques Générales du Laboratoire de Physique: Vol. II. [Book Review]—J. Surugue (Ed.). Publishers: Centre National de la Recherche Scientifique, Paris, & H. K. Lewis & Co., London, 336 pp., paper cover 42s., cloth 44s. (*J. sci. Instrum.*, Dec. 1950, Vol. 27, No. 12, p. 341.) Nine sections, contributed by different authors, deal respectively with: measurement of infrared radiation; technique of thin quartz fibres and Wollaston wires; casting thin cellulose films; electrometers; electrometer valves and circuits; ionization counters; investigation of fields by means of electrolyte tanks; 'microforge'; automatic regulation of temperature.

MATHEMATICS

512.831 : 518.6 : 681.177 **1175**

Calculation of the Eigenvalues of Matrices by means of Punched-Card Machines.—P. Henrici. (*Z. angew. Math. Phys.*, 15th May 1950, Vol. 1, No. 3, pp. 185–189.)

517.9 **1176**

Natural Eigenvalue Problems.—E. Stiefel & H. Ziegler. (*Z. angew. Math. Phys.*, 15th March 1950, Vol. 1, No. 2, pp. 111–138.) Eigenvalue problems, arising from questions of stability or in connection with oscillations, may be approached (*a*) as problems of the calculus of variations, (*b*) by establishing their differential equations, (*c*) by means of their integral or integrodifferential equations. Up to the present, method (*b*) has been developed further than the others, but it has the disadvantage that it is not applicable to problems whose eigenvalue is explicitly involved in the boundary conditions. To overcome this difficulty, it is advisable to use method (*a*) which, from a physical point of view, has the advantage of facilitating the mathematical interpretation of mechanical restrictions. The authors intend to generalize the theory of eigenvalues and to establish relations between the three methods for the general case of so-called 'natural' problems. In the present paper the relation between methods (*a*) and (*b*) is established for the case of one independent and one dependent variable.

517.93 : [534.1 + 538.56 **1177**

On the Forced Vibrations of Quasi-Linear Systems.—F. G. Friedlander. (*Quart. J. Mech. appl. Math.*, Sept. 1950, Vol. 3, Part 3, pp. 364–376.) Discussion of the differential equation $\ddot{x} + \omega^2 x = k f(x, \dot{x}, t)$ where ω, k are parameters, k being small, and $f(x, \dot{x}, t)$ is periodic with period 2π in t. Using an approximate solution due to Kryloff and Bogoliuboff it is shown that, if $f(x, \dot{x}, t)$ is a polynomial in x, $\dot{x}$, cos t, and sin t, the amplitude $(x^2 + \dot{x}^2/\omega^2)^{\frac{1}{2}}$ settles down to an asymptotic range of variation of order k as $t \to \infty$ which depends on the initial conditions, unless ω differs only by order k from any one of a set of critical rational values. For these values, subharmonic resonance occurs. In this case the upper limit for the asymptotic range of variation of the amplitude is of order $k/|\omega-\omega_0|$, where ω_0 is the critical frequency nearest to ω. The extension of the argument to an equation in which $f(x, \dot{x}, t)$ is not a polynomial in x, $\dot{x}$, cos t, and sin t, but still periodic in t, is indicated. The asymptotic behaviour of the phase $\tan^{-1}(x\omega/\dot{x})$ is also considered briefly. The reduction of the case of subharmonic resonance to an auxiliary differential equation of the first order is outlined, if either $\omega_0 = m/n$, where m and n are large relatively prime integers, or $k/|\omega-\omega_0|$ is small, the results obtained in this way are in agreement with those obtained when subharmonic resonance does not occur.

517.942.82 : 517.942.6/.9 **1178**

Determination of Eigenvalues using a Generalized Laplace Transform.—M. D. Friedman. (*J. appl. Phys.*, Dec. 1950, Vol. 21, No. 12, pp. 1333–1337.) The classical eigenvalue problems of mathematical physics are solved by means of a Laplace transform extended, not over the interval $(0, \infty)$ but over the interval of interest for the differential equation. The method is applied to the Hermite, Laguerre and Bessel equations and to the equation for the hypergeometric polynomials which include the Legendre, Tchebycheff, and Jacobi polynomials as special cases.

519.272.119 : 534.78 **1179**

Correlation Function Analysis.—L. G. Kraft. (*J. acoust. Soc. Amer.*, Nov. 1950, Vol. 22, No. 6, pp. 762–764.) A paper presented at the Speech Communication Conference at Massachusetts Institute of Technology, 1950. Discussion of theory on which the M.I.T. digital electronic correlator is based, and presentation of speech correlation curves obtained with it.

519.272.119 : 621.39.001.11 **1180**

The Auto-correlation Function.—D. A. Bell. (*Wireless Engr*, Jan. 1951, Vol. 28, No. 328, pp. 31–32.) An explanation is given of what the auto-correlation function is and what it can be used for, and a list of papers and books bearing directly on the subject is included.

681.142 **1181**

Electronic Computers: General Design and Construction.—L. Kosten. (*Tijdschr. ned. Radiogenoot.*, Nov. 1950, Vol. 15, No. 6, pp. 313–338.) An outline is given of the main parts of digital computers, viz., the input and output units, the memory, the arithmetic unit and the external control. Magnetic, supersonic and electrostatic memory systems are discussed. Two adding units are described, working respectively on a parallel and on a series basis. A simplified block diagram of the external control of the EDSAC is shown, together with a schedule of orders.

681.142 **1182**

Electronic Computers: Mathematical Bases.—I. Boxma. (*Tijdschr. ned. Radiogenoot.*, Nov. 1950, Vol. 15, No. 6, pp. 299–310. Discussion, p. 311.) An examination is made of the conflicting considerations imposed in the choice of a calculating system by considerations of accuracy, speed and scope on the one hand and cost on the other. The advantages and disadvantages of the binary digital system are discussed, and the elementary mathematical operations required in that system are described.

681.142 **1183**

Special Devices for Differential Analyzers.—A. C. Cook, L. K. Kirchmayer & C. N. Weygandt. (*Elect. Engng, N.Y.*, Dec. 1950, Vol. 69, No. 12, p. 1080.) Summary of A.I.E.E. 1950 Fall General Meeting paper. The range of the General Electric Company's differential analyser is being extended by the development of special devices for the generation of auxiliary functions, thus effectively increasing the number of integrators available for the solution of a given problem. Recent additions include (*a*) a gear-ratio selector, (*b*) a remote indicator, (*c*) a transportation-lag device, (*d*) a recording counter, (*e*) dead-band units, (*f*) disconnect units; their functions are described.

681.142 **1184**

The Physical Realization of an Electronic Digital Computor.—A. D. Booth. (*Electronic Engng*, Dec. 1950, Vol. 22, No. 274, pp. 492–498.) Details are given of circuits designed to convert an existing relay computer with magnetic storage (3179 of 1949) to fully electronic operation. The units discussed are the memory control, the digit amplifier, the shifting register, the magnetic gate circuits and the adding equipment.

681.142 **1185**

Marginal Checking as an Aid to Computer Reliability.—N. H. Taylor. (*Proc. Inst. Radio Engrs*, Dec. 1950, Vol. 38, No. 12, pp. 1418–1421.) The presence of deteriorating components, which are liable to fail and cause errors in a digital computer, can be detected by a variation of circuit voltages, thus inducing actual failure, while a test programme is performed. The failing components can then be systematically isolated. Trials have shown the value of this 'preventive maintenance' in improving reliability.

681.142 : 519.272.119 : 621.39.001.11 **1186**

A Digital Electronic Correlator.—Singleton. (See 1237.)

681.142 : 621.383 **1187**

Photoelectric Analog Computer.—E. C. Koenig. (*Electronics*, Dec. 1950, Vol. 23, No. 12, pp. 124, 126.) Description of equipment making use of units comprising a single lamp for excitation of ten photocells arranged radially and having separate apertures and outputs. The photocells used are of the vacuum type with $Ag/Cs_2O/Cs$ cathode. Advantages of the computer are enumerated.

MEASUREMENTS AND TEST GEAR

621.3.018.41(083.7) : 621.396.615.17/.18+621.317.761 **1188**

Frequency Generation and Measurements.—W. S. Mortley: H. J. Finden. (*Electronic Engng*, Dec. 1950, Vol. 22, No. 274, p. 531.) Comment on 1961 of 1950 and author's reply.

621.3.018.41(083.74) + 621.316.726.078.3 **1189**

The Evolution of Frequency Control.—Booth. (See 1271.)

621.3.018.41(083.74) : 621.396.81 **1190**

Coverage of Standard-Frequency Station WWVH.—E. L. Hall. (*TV Engng, N.Y.*, Aug. & Oct. 1950, Vol. 1, Nos. 8 & 10, pp. 16–18 & 20 . . 27.) A report of reception tests at several dozen stations in U.S.A., Alaska, Australia, China, Japan, and islands in the Pacific Ocean. The results are plotted and the resulting curves show reception efficiency as dependent on signal frequency, time of day, season of year, and sunspot conditions. The times of sunrise and sunset at the receiving station and at the transmitter have a definite effect on the percentage usability of the signals; the best reception is obtained for an all-darkness path. This is particularly the case for the 5-Mc/s signals. The 10-Mc/s signals are usable for more time than the 5-Mc/s signals, while the 15-Mc/s signals are usable still longer and may give uninterrupted service throughout the 24 hours at distances up to nearly 4 000 miles. The results in general indicate that for Alaska and the Pacific areas the transmissions from WWVH have resulted in an improvement of 87% in the standard-frequency and time-signal service, though in U.S.A. some confusion or interference with WWV transmissions, amounting to 2·5%, has been caused.

621.3.087.4 : 551,510.535 **1191**

An Automatic Ionospheric Recorder for the Frequency Range 0·55 to 17 Mc/s.—R. Naismith & R. Bailey. (*Proc. Instn elect. Engrs*, Part III, Jan. 1951, Vol. 98, No. 51, pp. 11–18.) A full description of automatic equipment giving a photographic record of ionospheric characteristics in the form of a graph of height of reflection at vertical incidence against transmitted frequency. A pulsed self-oscillator coupled directly to the transmitting aerial is tuned from 0·55 to 17 Mc/s in five bands. The receiver tuning capacitor is driven by a magslip motor whose speed is controlled by a discriminator circuit so as to keep the receiver in tune with the transmitter. The height and frequency calibrations and the date and time are recorded automatically on each record. The equipment has been in service for six years and has recorded 98% of the hourly values of the critical frequency of the F_2 layer.

621.3.087.4 : 551.510.535 **1192**

A Single-Band 0-20-Mc/s Ionosphere Recorder embodying some New Techniques.—T. L. Wadley. (*Proc. Instn elect. Engrs*, Part III, Jan. 1951, Vol. 98, No. 51, pp. 45–46.) Discussion on 392 of 1950.

621.317 : 550.362 **1193**

Theoretical Basis of a Method for Determining the Specific A.C. Resistivity of the Ground.—H. Weber. (*Tech. Mitt. schweiz. Telegr.-TelephVerw.*, 1st July 1950, Vol. 28, No. 7, pp. 257–260. In French and German.) Description of a method, similar to that of Collard (1136 of 1936), making use of a family of curves for the magnetic field produced by current in an overhead line with earth return.

621.317.3 : 621.392.5 **1194**

Automatic Transmission Measuring Set.—J. M. Hudack. (*Bell Lab. Rec.*, Dec. 1950, Vol. 28, No. 12, pp. 538–541.) An outline description of the equipment and its principles of operation. It provides, in about 30 sec, a record of the amplitude transmission characteristic of the apparatus under test as a function of frequency within the alternative bands 20 c/s to 20 kc/s or 100 c/s to 100 kc/s. Portions of these bands may be selected if required. Other uses for the equipment include the measurement of harmonic generation within the equipment under test and examination of its noise spectrum.

621.317.311 + 536.58 : 621.38 + 621.397.62 **1195**

Electronics in Britain.—J. H. Jupe. (*N.Z. elect. J.*, 25th March 1950, Vol. 23, No. 3, p. 235.) Short descriptions are given of a method for measuring extremely small direct currents or voltages, electronic equipment for temperature control, and extension units for television.

The direct current or voltage to be measured is applied to a circuit including a short thin Pt wire whose resistance is varied periodically by application of h.f current modulated at 50 c/s. From the resulting fluctuations of the d.c. an alternating voltage is derived, amplified, and applied to an indicator calibrated in terms of direct current or voltage. See also 1206 and 1266.

621.317.329 **1196**

Rheographic Study of Laplace Fields with Helicoidal Structure.—H. Dormont. (*Philips Res. Rep.*, Aug. 1950, Vol. 5, No. 4, pp. 262–269. In French.) Two methods are described for determining the equipotentials of a spirally wound system of conductors from measurements in an electrolyte tank.

621.317.334/.335.087.3 : 546.3 **1197**

A Method for the Measurement of Small Variations of Inductance and Capacitance (Metal Locator).—A. Hersping. (*Arch. Elektrotech.*, 1950, Vol. 40, No. 1, pp. 57–74.) The method is based on the phase variation within the pull-in range of a pair of coupled valve oscillators. A description is given of a metal locator embodying the principle. The theory of the pull-in action is given briefly; formulae derived are confirmed by experiments.

621.317.725.029.45/.5 **1198**

Study of a H. F. Millivoltmeter.—(*Radio franç.*, Dec. 1950, No. 12, pp. 1–6.) Description of a Philips instrument. See 2285 of 1950 (Lindenhovius et al.).

621.317.755 : 621.3.011.6 **1199**

Measurement of Time Constants.—R. Aschen. (*Radio franç.*, Nov. 1950, No. 11, pp. 16–19.) Description of the principle of operation of a c.r.o. circuit. Two EL41 valves are connected in a multivibrator circuit which provides the horizontal sweep voltage and also controls the current through a third EL41, in the anode circuit of which is the component or circuit under test. When the valve is conducting, the scanning spot is held to the right of the screen; during fly-back the valve anode current is cut off and the discharge characteristic of the component is displayed on the screen. Typical traces and complete circuit details are shown.

621.396.622.63.001.4 **1200**

Mixer-Crystal Checker.—P. D. Strum. (*Electronics*, Dec. 1950, Vol. 23, No. 12, pp. 94–97.) A method for quick determination of conversion loss from d.c. measurements. The theory of the method is outlined and a practical instrument described; its accuracy is estimated by comparison with a series of direct measurements of conversion loss made at wavelengths of 3 and 10 cm. The basis of the method is that minimum conversion loss for a crystal can be predicted from its static voltage/current curve.

621.396.645.001.4 **1201**

Test and Alignment Procedures for Video Amplifiers.—F. E. Cone & N. P. Kellaway. (*Broadcast News*, Sept./Oct. 1950, No. 61, pp. 28–33.) General procedures are outlined and specific alignment methods are described for each component of the R.C.A. television terminal equipment.

621.397.6.001.4 **1202**

Signal Sources for Television Testing.—D. W. Thomasson. (*J. Brit. Instn Radio Engrs*, Dec. 1950, Vol. 10, No. 12, pp. 369–390. Discussion, pp. 391–392.) A detailed discussion of the performance requirements of signal sources. The r.f. source must be within 0·1% of the correct frequency. The video-frequency generator must give an output approximating to the actual television signal. Tolerances for the timing of the various parts of the synchronizing and picture signals are given. Suitable circuits are described for the complete generator.

621.317.3.029.6 **1203**

Micro-Wave Measurements. [Book Review]—H. M. Barlow & A. L. Cullen. Publishers: Constable & Co., London, 339 pp., 30s. (*J. sci. Instrum.*, Dec. 1950, Vol. 27, No. 12, p. 342.) A book of permanent value not only to communication engineers, for whom it is primarily intended, but also to research physicists. Clear explanations are provided of the principles involved in the measurements. The longest chapters are those dealing with standing-wave measurements and with matching and transmission systems.

621.396.6.001.4 **1204**

Testing Radio Sets. [Book Review]—J. H. Reyner. Publishers: Chapman & Hall, London, 5th edn, 215 pp., 22s. 6d. (*J. sci. Instrum.*, Dec. 1950, Vol. 27, No. 12, p. 341.) A practical book which can be recommended.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

621.316.79.076.7 : 621.369.3 **1205**

Electronic Control of Home Heating.—J. M. Wilson. (*Electronics*, Dec. 1950, Vol. 23, No. 12, pp. 84–87.) Description, with complete circuit details, of an a.c. resistance bridge and associated amplifier controlling the heating system of a building, greater heat being switched on when the resistance variation of an outdoor element indicates a drop in temperature.

621.317.311 + 536.58 : 621.38 + 621.397.62 **1206**

Electronics in Britain.—J. H. Jupe. (*N.Z. elect. J.*, 25th March 1950, Vol. 23, No. 3, p. 235.) Short descriptions are given of a method for measuring extremely small direct currents or voltages, electronic equipment for temperature control, and extension units for television.

The temperature-control unit operates by means of a Wheatstone-bridge circuit with a special diagonal branch including a resistor forming part of an amplifier circuit, a galvanometer, and two thermocouples connected in opposition whose action controls a servomechanism. See also 1195 and 1266.

621.317.334/.335.087.3 : 546.3 **1207**

A Method for the Measurement of Small Variations of Inductance and Capacitance (Metal Locator).—Hersping. (See 1197.)

621.365.55† **1208**

Load-Matching Dielectric Heaters.—R. H. Hagopian. (*Electronics*, Dec. 1950, Vol. 23, No. 12, pp. 98–101.) Increased efficiency is obtained by use of transmission lines for coupling r.f. generators to the loads, as matching problems are simplified.

621.38.001.8 **1209**

Electronics Symposium and Exhibition.—(*Nature, Lond.*, 23rd Dec. 1950, Vol. 166, No. 4234, pp. 1062–1063.) Summaries of the lectures at the opening ceremony of the conference arranged by the Scientific Instrument Manufacturers Association, London, Sept. 1950, and of the papers presented at the various sessions, with a note of the exhibits. For another account see 420 of February.

621.38.001.8 : 786.6 **1210**

The Baldwin Electronic Organ.—A. Douglas. (*Electronic Engng*, Dec. 1950, Vol. 22, No. 274, pp. 507–511.) An instrument in which the tones are generated entirely by valve oscillators. The basic waveform is a sawtooth, and square waves are generated by the addition of two sawtooth waves. Instrument tones are simulated by the use of resonant filters. Details are given of the basic circuits and of various refinements designed to provide tone quality closely resembling that of a pipe organ.

621.383 + 621.385.38] : 796.357 **1211**

Thyratron-Controlled Photoelectric Umpire.—R. F. Shea. (*Electronics*, Dec. 1950, Vol. 23, No. 12, pp. 74–77.) Description of an automatic device that indicates passage of a baseball through the strike zone and also the speed of the ball as it crosses the plate. Three sets of photocells look at the sky through narrow slits, thus defining two vertical planes and an inclined plane intersecting the other two at knee and arm-pit height respectively. Passage of the ball through these planes in the correct sequence triggers an interconnected trio of thyratrons and operates an indicator.

621.384.611.2† **1212**

The Design of Electron Synchrotrons.—F. K. Goward, J. D. Lawson, J. J. Wilkins & R. Carruthers. (*Proc. Instn elect. Engrs*, Part I, Nov. 1950, Vol. 97, No. 108, pp. 320–333. Discussion, pp. 334–339.)

621.384.611.2† **1213**

The Design and Operation of a 30-MeV Synchrotron.—D. W. Fry, J. Dain, H. H. H. Watson & H. E. Payne. (*Proc. Instn elect. Engrs*, Part I, Nov. 1950, Vol. 97, No. 108, pp. 305–319. Discussion, pp. 334–339.)

621.384.612.1† **1214**

Beam Oscillations in a F.M. Cyclotron.—T. Teichmann. (*J. appl. Phys.*, Dec. 1950, Vol. 21, No. 12, pp. 1251–1257.)

621.385.1.001.8 : 531.768.087 1215

The Improved Ramberg Vacuum-Tube Accelerometer.—(*Tech. Bull. nat. Bur. Stand.*, Dec. 1950, Vol. 34, No. 12, pp. 180–182.) Improved performance is obtained by (*a*) an aging treatment resulting in more constant emission, (*b*) double gettering, (*c*) a 25-fold increase of sensitivity. See also 812 of 1949 and 2528 of 1947.

621.385.832 : 621.3.012 : 518.4 1216

The Graph-Scope, an Electronic Graph Plotter and Graphical Computer.—A. L. Thomas, Jr. (*Elect. Engng, N.Y.*, Dec. 1950, Vol. 69, No. 12, pp. 1097–1100.) Description of an instrument for displaying on the screen of a c.r. tube a graph corresponding to any set of data, together with co-ordinate scales which can be either linear, logarithmic or hyperbolic. A keyboard like that of a computer is used for setting the point-plotting circuits.

621.385.833 1217

A New Ratiometer for the Electron Microscope.—R. Strauss. (*Arch. Elektrotech.*, 1950, Vol. 40, No. 1, pp. 49–56.) The object of this ratiometer is to make possible the full exploitation of the high resolving power of electron microscopes with magnetic lens systems, by accurately measuring and maintaining the ratio of the accelerating voltage to the square of the lens current, which ratio determines the image definition. The ratiometer comprises a long magnetic coil, within whose field the path of an electron beam is a helix; variations of the path are displayed on a fluorescent screen, and adjustment can be performed automatically or by hand. The accuracy attained for voltage variations is to within 2 parts in 10^4.

621.385.833 1218

Focusing in Electron Microscopy.—R. S. M. Revell & A. W. Agar. (*J. sci. Instrum.*, Dec. 1950, Vol. 27, No. 12, p. 337.) In high-resolution work, personal errors in judging optimum focus can be avoided by taking a through-focus series of micrographs.

621.385.833 1219

The Correction of the Spherical Aberration of Electron Lenses using a Correcting Foil Element.—U. F. Gianola. (*Proc. phys. Soc.*, 1st Dec. 1950, Vol. 63, No. 372B, pp. 1037–1039.) A physical explanation of the action of foil elements, introduced into the active region of an electron lens, in correcting the positive aberration of the lens, is confirmed by rigorous analysis.

621.387.4† 1220

An End-Window Alpha Scintillation Counter for Low Counting Rates.—C. W. Reed. (*Nucleonics*, Dec. 1950, Vol. 7, No. 6, pp. 56–62.)

621.387.424† : 621.392 1221

Auxiliary Electronic Circuits for Geiger Counters.—Fontes & Moret. (See 1105.)

778.37 : 537.523.4 1222

A High-Intensity Short-Duration Spark Light Source.—J. A. Fitzpatrick, J. C. Hubbard & W. J. Thaler. (*J. appl. Phys.*, Dec. 1950, Vol. 21, No. 12, pp. 1269–1271.) Summary noted in 2604 of 1950 (Fitzpatrick & Thaler).

778.37 : 621.397.331.2 1223

An Iconoscope Electro-optical Shutter for High-Speed Photography.—H. A. Prime & R. C. Turnock. (*Proc. Instn elect. Engrs*, Part II, Dec. 1950, Vol. 97, No. 60, pp. 793–796.) The system described makes use of an iconoscope in which the photo-emission is controlled by the application of suitable voltages to the collector electrode. The charge image, formed during a selected interval, is stored on the mosaic and subsequently scanned. Examples of exposures of duration ranging down to 20 μs are shown. The mechanisms of spurious-image formation during the 'shutter-closed' period are discussed. Performance is compared with that of the Kerr-cell type of shutter.

621.387† 1224

Ionization Chambers and Counters. [Book Review]—D. H. Wilkinson. Publishers: Cambridge University Press, New York, 1950, 255 pp., $4.50. (*Proc. Inst. Radio Engrs*, Dec. 1950, Vol. 38, No. 12, p. 1465.) The treatment is focused "upon the principles of instrumentation, particularly upon the Geiger counter and ion chamber, to the exclusion of any description of the crystal counter and some of the more recent nuclear detection devices. None the less, the book is very complete and well organized with respect to the more conventional counting devices."

PROPAGATION OF WAVES

621.396.11 + 535.222 1225

The Velocity of Propagation of Electromagnetic Waves derived from the Resonant Frequencies of a Cylindrical Cavity Resonator.—L. Essen. (*Proc. roy. Soc. A*, 7th Dec. 1950, Vol. 204, No. 1077, pp. 260–277.) The length of the resonator was varied by means of a piston whose change of position was measured by calibrated slip gauges. This change was measured for the H_{01n} resonance modes in vacuo near 9 and 6 kMc/s up to $n = 8$ and $n = 2$ respectively. The frequency was measured by comparison with a quartz-crystal standard, using an interpolation oscillator. Measurements made at different frequencies allowed the diameter of the cavity to be determined without knowing c; the value obtained was greater by 2×10^{-4} cm than that determined by metrological methods; the difference is attributed to a film of AgS, which would also account for the 25% difference between the measured and calculated values of Q.

From a survey of all possible sources of error the maximum estimated error is $\pm$ 3 km/sec and the final value of $c = 299\,792{\cdot}5$ km/sec. See also 1751 of 1950, and 324 (Bergstrand) and 430 (Bol) of February.

621.396.11 : 551.510.535 1226

The Path of a Ray in a Curved Ionosphere Layer.—K. Bibl. (*Rev. sci., Paris*, Jan./March 1950, Vol. 88, No. 3305, pp. 27–29.) Calculation of the effective ground-path length (D_1) for a reflected ray can be simplified by considering, not the angle of incidence on the lower surface of the layer, but the angle which the non-refracted ray makes with the median plane of the layer, and by starting integration from this plane. The expression derived for D_1 indicates that the plane-layer formula is applicable in the case of a curved layer, the maximum error not exceeding 25 km. For nearly horizontal propagation the second-order (frequency correction) terms cannot be neglected.

621.396.11 : 621.396.812.4 1227

Comparison of Tropospheric Reception at 44·1 Mc/s with 92·1 Mc/s over the 167-Mile Path of Alpine, New Jersey, to Needham, Massachusetts.—G. W. Pickard & H. T. Stetson. (*Proc. Inst. Radio Engrs*, Dec. 1950, Vol. 38, No. 12, p. 1450.) Summary only of paper abstracted in 977 of April.

621.396.11 : 621.396.9 1228

Some Adverse Influences of Meteorological Factors on Marine Navigational Radar.—J. A. Saxton & H. G. Hopkins. (*Proc. Instn elect. Engrs*, Part III, Jan. 1951, Vol. 98, No. 51, pp. 26–36.) The effect of absorption and scattering of centimetre radio waves by atmospheric gases, uncondensed water vapour and various forms of precipitation on the range obtained by marine radar equipment is discussed. For the wavelength of 3·2 cm now used for marine radar, the attenuation by atmospheric gases and uncondensed water vapour is hardly significant; absorption and scattering by rain are likely to cause most of the appreciable reductions in range of detection. It appears that, at a wavelength of 3·2 cm, for targets having echoing areas greater than 2 000 m^2 (e.g., ships of more than about 10 000 tons), attenuation in rainstorms will cause reduction in range, whereas for smaller targets masking by echoes from the rain will often be the more serious factor. Although very intense snowstorms can produce echoes sufficiently strong to be troublesome, the rate of precipitation required is such that its frequency of occurrence is unlikely to be great. In dense fogs reduction in detection range may be appreciable. Tables and graphs are given showing the effect on detection range of various types of precipitation in polar, temperate and tropical regions for ships of 10 000 tons and of 1 000 tons, and for small boats or buoys.

621.396.11.029.62 **1229**

The Propagation of Metre Waves.—F. C. Saic. (*Elektrotech. u. Maschinenb.*, Nov. 1950, Vol. 67, No. 11, pp. 325–332.) Critical discussion of the formulae of van der Pol & Bremmer and of Eckersley for propagation over land and over sea, both within and beyond the 'visibility' range, taking account of transmitter height. Experimental results for wavelengths in the range 3–8 m are shown graphically.

621.396.81 : 621.3.018.41(083.74) **1230**

Coverage of Standard-Frequency Station WWVH.—Hall. (See 1190.)

621.396.812 **1231**

Radio Propagation between Noumea and Adélie Land.—M. Barré, K. Rawer & É. Argence. (*Rev. sci., Paris*, Jan./March 1950, Vol. 88, No. 3305, pp. 21–26.) A series of graphs shows the relative strengths at various times of day of communication signals at different frequencies received at Noumea and in the s.s. *Commandant Charcot* about 5 500 km away in the Antarctic. The optimum time for communication was about 1100 G.M.T. (2100 local time), using frequencies in the range 8–17 Mc/s. Conditions generally remained good during the night but deteriorated abruptly at dawn. Factors used in predicting usable frequencies for the voyage are considered; due account was taken of auroral effects and of records formerly obtained at stations in the corresponding northern latitudes. The predictions show remarkable agreement with observations.

RECEPTION

621.396.812 **1232**

Diversity Effects in Spaced-Aerial Reception of Ionospheric Waves.—E. N. Bramley. (*Proc. Instn elect. Engrs*, Part III, Jan. 1951, Vol. 98, No. 51, pp. 19–25.) A mathematical analysis is given of the phase and amplitude of the signals induced in two spaced aerials by a continuous angular spectrum of downcoming rays, all lying in the vertical plane through the two aerials. It is assumed that the amplitudes of the individual rays are constant and that their phases vary in a random manner. If the energy is concentrated in a small range of angles the extent of the angular distribution can be estimated from observations of the variation of either the amplitudes or the phase difference of the signals in the spaced aerials. A similar analysis is made for a specularly reflected component superimposed on an angular distribution of waves having random phase. If observations are made of both the amplitudes and the phase difference of the signals in the two aerials, the width of the angular distribution and the ratio of the amplitude of the specularly reflected component to that of the angular distribution (the signal/noise ratio) may be found.

Some daytime measurements of first-order reflections from the ionosphere at nearly vertical incidence at frequencies between 4 and 7 Mc/s are most satisfactorily explained in terms of a steady specularly reflected component superimposed on a noise background. Values of signal/noise ratio of about 2 or 3 have been obtained with a noise background having an angular spread of about 1°.

621.396.823 **1233**

Car Ignition Radiation.—C. C. Eaglesfield. (*Wireless Engr*, Jan. 1951, Vol. 28, No. 328, pp. 17–22.) Theory previously given (3741 of 1946) for the radiation from car ignition systems is developed, assuming that the impulse voltage is applied to the sparking plug through a radiating inductor. Allowance is made for resonances in the ignition system and for the addition of suppression resistors. Satisfactory agreement is obtained with the experimental results of Pressey & Ashwell (1779 of 1949) for the latter case.

621.396.822 **1234**

Threshold Signals. [Book Review]—J. L. Lawson & G. E. Uhlenbeck. Publishers: McGraw-Hill, New York, 1950, 388 pp., 42s. 6d. (*Electronic Engng*, Dec. 1950, Vol. 22, No. 274, p. 532.) Volume 24 of the Radiation Laboratory Series. ". . . presents a connected account of the work at the Radiation Laboratory on the discrimination of signals in the presence of noise, with particular emphasis on pulsed signals, visually displayed in the presence of random noise. . . . The volume can be highly recommended both to the newcomer and to the specialist."

STATIONS AND COMMUNICATION SYSTEMS

621.39 : 001.8 **1235**

Storage Devices for Communications.—A. J. Lephakis. (*Electronics*, Dec. 1950, Vol. 23, No. 12, pp. 69–73.) The advantages to be gained by systematically recording messages prior to transmission over a communications link are considered. For such purposes storage systems are required in which the rate of extraction of data is not at all times equal to the input rate; possible systems are indicated. Other uses of storage devices considered include (*a*) radar display gear designed to respond only to moving targets and (*b*) equipment in which a desired relation between input and output signals (transfer characteristic) is synthesized.

621.39.001.11 : 519.272.119 **1236**

The Auto-correlation Function.—Bell. (See 1180.)

621.39.001.11 : 681.142 : 519.272.119 **1237**

A Digital Electronic Correlator.—H. E. Singleton. (*Proc. Inst. Radio Engrs*, Dec. 1950, Vol. 38, No. 12, pp. 1422–1428.) 1950 I.R.E. National Convention paper. "The relation between correlation functions and the general theory of communication is presented, and this relation leads to a technique for electronic computation of correlation functions and to the design of a machine for carrying out the computation. Because of the requirements of great accuracy and long storage, the machine makes use of binary digital techniques for storage, multiplication, and integration. Descriptions of the more unusual circuits in the machine are given, and circuit diagrams are included. A number of experimental results obtained by the machine are presented."

621.392.001.11 **1238**

The Information-Theory Point of View in Speech Communication.—R. M. Fano. (*J. acoust. Soc. Amer.*, Nov. 1950, Vol. 22, No. 6, pp. 691–696.) A paper presented at the Speech Communication Conference at Massachusetts Institute of Technology, 1950. A non-mathematical discussion of the Wiener-Shannon theory of information, which is used to estimate the rate of transmission of information in speech communication.

621.395.44 : 621.392.52 **1239**

Loss-Compensated Filters in Carrier-Frequency Systems.—Lehmann. (See 1113.)

621.396.4 : 621.396.615 : 621.316.726 **1240**

Stabilized Master Oscillator for Multichannel Communication.—Pappenfus. (See 1117.)

621.396.41 : 621.396.619.13 **1241**

The Application of Frequency Modulation to V.H.F. Multichannel Radiotelephony.—J. H. H. Merriman & R. W. White. (*Proc. Instn elect. Engrs*, Part III, Jan. 1951, Vol. 98, No. 51, p. 56.) Discussion on 3513 of 1948.

621.396.619.13 : 621.3.018.42† **1242**

Bandwidth of a Sinusoidal Carrier Wave, Frequency-Modulated by a Rectangular Wave with Half-Sine-Wave Build-Up.—W. A. Cawthra & W. E. Thomson. (*Proc. Instn elect. Engrs*, Part III, Jan. 1951, Vol. 98, No. 51, pp. 69–74.) "The problem of determining the frequency spectrum of a sinusoidal carrier, frequency modulated by a modified rectangular wave, is considered and a general solution obtained. The modification of the rectangular wave consists in replacing the vertical sides by a non-instantaneous transition in the form of a half-cycle sine wave. It is shown that for the determination of bandwidth it is reasonable to use, instead of the spectrum proper, a function which forms an envelope to it; this function is simpler and, in particular, it is independent of the on/off ratio of the rectangular wave. Curves are given for various values of the parameters involved."

621.396.65 : [621.317.083.7 + 621.396.5 + 621.398 **1243**

Field Testing a Microwave Channel.—D. R. Pattison, M. E. Reagan, S. C. Leyland & F. B. Gunter. (*Elect. Engng, N.Y.*, Dec. 1950, Vol. 69, No. 12, pp. 1092–1097.) Essentially full text of A.I.E.E. 1950 Summer and Pacific General Meeting paper. A description is given of microwave-channel and a.f. multiplex terminal equipment installed by the Pennsylvania Electric Company for simultaneous relaying, telemetering, supervisory control and voice communication. A reflector is used to provide a continuous microwave path between terminal stations where direct line-of-sight communication is not possible.

621.396.65 : 621.396.828 **1244**

Reflected-Ray Suppression.—H. E. Bussey. (*Proc. Inst. Radio Engrs*, Dec. 1950, Vol. 38, No. 12, p. 1453.) Destructive interference between the direct and ground-reflected rays may impair line-of-sight microwave communication. The reflected signal may be removed by means of an opaque screen on the ground at the geometrical reflection point which blocks out one half of the first Fresnel zone. Confirmatory tests for an 800-ft transmission path operating on a frequency of 4·5 kMc/s are described.

621.396.65.029.63 **1245**

A Simplified Method of Planning Decimetre-Wave Line-of-Sight Links, taking into account the Earth's Curvature and the Vision Ellipse.—K. O. Schmidt. (*Fernmeldetech. Z.*, Nov. 1950, Vol. 3, No. 11, pp. 408–414.) Description, with relevant curves and abacs, of a geometrical method based on constructing the ellipse representing the first Fresnel region. The method is applied to a 50-km link using a wavelength of 12·5 cm.

621.396.712 **1246**

WBSM, WBSM-FM, New Bedford, Massachusetts.—O. F. A. Arnold. (*Broadcast News*, Sept./Oct. 1950, No. 61, pp. 9–15.) Illustrated description of studios, control room and transmission system.

621.396.712 **1247**

WKPT and WKPT-FM.—T. Phillips. (*Broadcast News*, Sept./Oct. 1950, No. 61, pp. 22–27.) Illustrated description of studios, control rooms and transmission system at Kingsport, Tennessee.

621.396.712(94) **1248**

Engineering Aspects of the National Broadcasting Service.—R. J. Boyle. (*Telecommun. J. Aust.*, June 1948, Vol. 7, No. 1, pp. 16–30.) An outline of the establishment of the National Broadcasting Service in Australia, and discussion of the problems involved and the methods used to solve them. The planning of the transmitter locations, site selection, and field-strength surveys with test transmitters, are first considered. Typical transmitter equipment is then described, with construction details of the top-loaded vertical radiators used at Cumnock, N.S.W., Dooen, Victoria, and Wagin, Western Australia. As standard equipment, all transmitting stations are provided with a beat-frequency oscillator, noise and distortion meter, and modulation monitor; all have stand-by power plant.

Practically all the programme material broadcast from National stations originates in, or is routed through, the studios in the capital cities. A general account of the studio and programme-switching arrangements is illustrated by reference to the Sydney studios. Relay systems for inter-state and national broadcasting, and the arrangements at Liverpool, N.S.W., for the reception and rebroadcasting of overseas programmes, are also described.

SUBSIDIARY APPARATUS

621-526 **1249**

Relay Servomechanisms: The Shunt-Motor Servo with Inertia Load.—T. A. Rogers & W. C. Hurty. (*Trans. Amer. Soc. mech. Engrs*, Nov. 1950, Vol. 72, No. 8, pp. 1163–1172.) The theory is developed in terms of dimensionless motor parameters. Operating curves are drawn in the phase plane illustrating the effect of parameter changes on the stability of the servo for three forms of input signal.

621.314.634.011.4 **1250**

The Capacitance of Selenium Rectifiers.—R. E. Burgess. (*Proc. phys. Soc.*, 1st Dec. 1950, Vol. 63, No. 372B, pp. 1036–1037.) Comment on paper abstracted in 744 of March (Cooper).

621.316.722 **1251**

Stability of Voltage Regulators.—F. E. Bothwell. (*Elect. Engng, N.Y.*, Dec. 1950, Vol. 69, No. 12, p. 1090.) Summary of A.I.E.E. 1950 Fall General Meeting paper. The presence of nonlinear control elements in many voltage-regulator circuits has prevented a complete mathematical analysis. Theory of Liapounoff, showing that stability of the continuous periodic solutions of a set of nonlinear differential equations is determined by the stability of its linear perturbation equations, is used to solve the problem of stability of electric circuits subjected to small displacements from equilibrium. As an example, characteristic equations are developed for a voltage-regulated n-stage d.c. generator, and the stability boundary is obtained for the case of a step-resistor control element.

621.316.722.1 1252

Stabilized Power Pack.—R. E. Harvison. (*Electronic Engng*, Dec. 1950, Vol. 22, No. 274, pp. 512–513.) An analysis of the operation of two stabilizing valves in parallel to provide high current-output.

621.316.722.1.076.7 1253

Precision A.C. Voltage Stabilizers.—G. N. Patchett. (*Electronic Engng*, Sept.–Dec. 1950, Vol. 22, Nos. 271–274, pp. 371–377, 424–428, 470–473, 499–503.) Critical study of different types and circuit arrangements providing stabilization to within about 0·01%. See also 227 of January.

621.319.45.025 : 621.314.64 1254

Rectifier Properties of the System Al-Al$_2$O$_3$-Electrolyte subjected to an Alternating Voltage.—A. J. Dekker & W. C. van Geel. (*Philips Res. Rep.*, Aug. 1950, Vol. 5, No. 4, pp. 303–314. In French.) An account is given of an experiment showing that the static and dynamic current/voltage characteristics of this system are different. A bridge method is described for measuring the dynamic characteristic; the curve exhibits a loop in the region corresponding to forward current. The effect on the loop of frequency and temperature variations is investigated experimentally, and it is concluded that the structure of the oxide coating is not stable but depends on the magnitude and direction of the applied voltage.

621.316.722 1255

Voltage Stabilizers. [Book Review]—F. A. Benson. Publishers: Electronic Engineering, London, 1950, 125 pp., 12s. 6d. (*Electronics*, Dec. 1950, Vol. 23, No. 12, pp. 154, 158.) "A concise monograph . . . this book will be an extremely handy reference for an engineer or technician."

TELEVISION AND PHOTOTELEGRAPHY

621.397 1256

G.E.C. Picture-Telegraph Equipment.—(*Electronic Engng*, Dec. 1950, Vol. 22, No. 274, pp. 503–504.) A general description, with diagrams, of the principles of operation of transmitting and receiving equipment providing the highest possible definition within C.C.I.T. limits. At normal speed a 10 × 8 in. picture is transmitted in 14 minutes; this time can be halved at the expense of picture quality.

621.397.24/.26 : 621.396.65 1257

Portable Microwave Television Link.—(*Elect. Commun.*, Dec. 1950, Vol. 27, No. 4, pp. 295–297.) For another account see 459 of February.

621.397.331 : 535.623 1258

Photoelectric Method of Scanning and Chromatic Selection for Designs in Colour.—M. Lange. (*Toute la Radio*, Nov. 1950, No. 150, pp. 343–348.) The pattern from which the various colours are to be selected is fixed to a revolving cylinder moving along its own axis, and is scanned by a brightly illuminated spot. An optical system making use of the toothed-wheel principle derives two rays of light, each of which is directed on to a prism spectroscope. Parts of the spectra so obtained are applied to three (panchromatic) photocells, arrangements being made for varying the parts selected. The outputs from the three cells are amplified and mixed in such a way that the resultant voltage is zero for the colour to be selected, increases as this colour fades, and reaches a threshold value for white; black and any colour other than that required produce a voltage above the threshold value. This voltage actuates a galvanometer mirror reflecting a pencil of light of diamond-shaped cross-section through an aperture of the same shape. A beam of light of width depending on the position of the mirror is directed on to a light-sensitive film carried by the receiving drum. As the drum rotates a band is recorded on the film. An interposed revolving slotted disk transforms this band into a series of dots. Various refinements affecting the reproduced pattern are described.

621.397.5 : 535.624 1259

Subtractive Color TV.—I. Kamen. (*TV Engng, N.Y.*, Oct. 1950, Vol. 1, No. 10, pp. 12–13, 28.) The image is produced by passing light from a projection lamp through three dark-trace tubes in succession, the screens being of different materials corresponding to absorption of the three primary colours. A sequential-frame transmission system is used and the receiver tubes are fed with the appropriate colour signals. A greater luminous efficiency is claimed for the system.

621.397.5(494) 1260

Present State of Television in Switzerland.—(*Bull. schweiz. elektrotech. Ver.*, 10th Dec. 1949, Vol. 40, No. 25, pp. 1001–1003. In German.) General review, with discussion of the technical and financial problems involved.

621.397.6 : 621.396.67 1261

Empire State Television.—(*Broadcast News*, Sept./Oct. 1950, No. 61, pp. 41–45.) A description, with many photographs, of the replacement of the old 67-ft aerial surmounting the Empire State Building by a 217-ft multiple-aerial system for stations WCBS-TV, WABD, WJZ-TV, WPIX and WNBT.

621.397.6 : 621.396.67 1262

Temporary Vision Aerial.—Bolt. (See 1076.)

621.397.6.001.4 1263

Signal Sources for Television Testing.—Thomasson. (See 1202.)

621.397.611.2 1264

Television Camera Tubes.—L. H. Bedford. (*Wireless Engr*, Jan. 1951, Vol. 28, No. 328, pp. 4–16.) Paper read at the International Television Congress, Milan, 1949. The effect of a high-sensitivity tube in a television camera is to make it possible to use a smaller lens aperture for the same object illumination and thus obtain a greater depth of field. Curves showing depth of field against illumination are drawn for present-day camera tubes, the minimum possible illumination being governed by the largest practical aperture which can be used with the tube. The minimum illumination is a criterion of sensitivity. A dimensionless form of the criterion, independent of arbitrarily chosen factors, can be found by taking the ratio of this illumination to that required by an ideal tube, the 'Quanticon'. Other properties of camera tubes discussed and tabulated are colour response, resolution, spurious signal generation, signal/noise ratio, contrast, image persistence, geometrical accuracy, stability and life.

621.397.611.2 : 778.2 1265

Flying-Spot Camera, Type TK-3A.—C. R. Monro. (*Broadcast News*, Sept./Oct. 1950, No. 61, pp. 16–21.) Equipment is described for the production of video signals from 35-mm slides by the flying-spot method. The control and operating features are explained and typical applications are illustrated by photographs.

621.397.62 + 621.317.311 + 536.58 : 621.38 1266

Electronics in Britain.—J. H. Jupe. (*N.Z. elect. J.*, 25th March 1950, Vol. 23, No. 3, p. 235.) Short descriptions are given of a method for measuring extremely small direct currents or voltages, electronic equipment for temperature control, and extension units for television.

The Cossor extension unit for television comprises a coupling unit which reduces the modulation from the receiver to 0·5 V, together with the main unit which includes a loudspeaker and c.r. tube. The unit is designed to give perfect linearity, so that the extension signal is not affected by any nonlinearity existing in the main receiver. The picture may be much larger than that of the main set and the unit can be made to cover more than one type of television system, with negligible increase in cost. For example, units made for use with the British 405-line 50-frames sec system can be adjusted in a few minutes to work with an American 525-line 60-frames/sec receiver. See also 1216 of 1949 (Zeluff) and 1195 and 1206 above.

621.397.62 **1267**

The Simplification of Television Receivers.—W. B. Whalley. (*Proc. Inst. Radio Engrs*, Dec. 1950, Vol. 38, No. 12, pp. 1404–1408.) Reprint. See 1271 of 1950.

621.397.62 **1268**

The Video Output Stage.—E. T. Emms & E. Jones. (*Electronic Engng*, Oct. & Nov. 1950, Vol. 22, Nos. 272 & 273, pp. 408–413 & 454–460.) Values of t, the signal build-up time for the various elements of the video stage, are derived from the required overall build-up time T estimated from the build-up time τ for the transmitted wave, picture definition being classified broadly in terms of T/τ. Different anode compensation circuits are shown and the indicial response (response to unit-step input) is plotted for each circuit. The design of a critically damped cathode compensation circuit is indicated. For equal output and equal t values, the input of such a stage must be 1·4 times that of a first-order shunt-compensated stage; but since the build-up time of the input signal is finite, this is not necessarily a disadvantage. Required characteristics of the video valve are tabulated. Class-A and Class-B modes of operation of this valve as an anode-bend demodulator are discussed and theoretical efficiency curves are derived. Secondary factors in design are considered and a design procedure is described for each of the first two types of circuit, component values being calculated for circuits using Type-EF80 valves.

621.397.62 : 621.396.662 **1269**

A Variable-Inductance TV Tuner.—D. R. DeTar & H. T. Lyman, Jr. (*Electronics*, Dec. 1950, Vol. 23, No. 12, pp. 102–106.) Description of a tuner covering all v.h.f. television channels without the use of switches or sliding contacts. It comprises a r.f. amplifier, mixer-oscillator and first detector. The wide frequency range is obtained by the use of only four variable inductors, each of which consists of a specially wound solenoidal coil with fixed core, the tuning being effected by sliding an Al sleeve between coil and core. Performance details are given and comparison is made with conventional designs.

621.397.7 **1270**

WOR's TV Studios.—N. F. Smith. (*Broadcast News*, Sept./Oct. 1950, No. 61, pp. 46–73.) A very comprehensive and detailed description of the New York studios and control rooms, including switching arrangements, film-projection equipment, master control system and monitoring facilities.

TRANSMISSION

621.316.726.078.3 + 621.3.018.41(083.74) **1271**

The Evolution of Frequency Control.—C. F. Booth. (*Proc. Instn elect. Engrs*, Part III, Jan. 1951, Vol. 98, No. 51, pp. 1–10.) A historical survey is made of the frequency stability of radio transmitters and the frequency tolerances specified by successive international conferences. The theoretical bandwidth needed by a single-channel telegraph transmitter at the higher end of the 4–30-Mc/s band is a small fraction of the tolerance (± 0·003%) assigned to it at the Atlantic City conference. Any improvement in stability will thus increase the number of stations possible. Simple crystal-controlled drive units capable of achieving the present tolerances are described and drive units for synchronized m.f. transmitters, frequency standards and quartz clocks capable of much higher stability are discussed.

VALVES AND THERMIONICS

535.215.1/.2 : 621.383.2 **1272**

Some Properties of Complex Photoelectric Layers.—A. Lallemand & M. Duchesne. (*Z. angew. Math. Phys.*, 15th May 1950, Vol. 1, No. 3, pp. 195–201. In French.) From experimental results on photoelectric layers of the type $Ag\text{-}Cs_2O\text{-}Cs$, it is concluded (*a*) that such complex layers, although not obeying Richardson's law, have a well defined work function which can be determined from Fowler's theory; (*b*) that the layers are sensitive to wavelengths in the neighbourhood of 15 000 Å, but that the photoelectric emission, though not completely explained by Fowler's theory, is not appreciably greater than that to be expected from the theory; (*c*) that the observed thermionic emission can be explained on the assumption that the layers consist of a large number of elements whose work function can be determined from Fowler's theory, or of a very small number of elements with a much smaller work function of the order of 0·4 V.

535.215.4 : 621.383.4 **1273**

A Relationship between the Refractive Index and the Infrared Threshold of Sensitivity for Photoconductors.—T. S. Moss. (*Proc. phys. Soc.*, 1st March 1950, Vol. 63, No. 363B, pp. 167–176.) "It is shown that for photoconductive compounds the long-wavelength 'threshold' of sensitivity λ should be related to the refractive index n of the photoconductor, and analysis of the available data shows that for the more highly refractive compounds n^4/λ always has a value close to 77. As a consequence of this relation, an explanation can be given for the change of the threshold wavelength with temperature which is observed for lead sulphide and similar materials. Suggestions are made as to what materials ought to be suitable for photoconductive detectors at longer wavelengths in the infra-red, and some confirmation of the ideas is shown by results obtained with cadmium arsenide, the properties of which are briefly described."

537.525.92 : 537.533.7 : 621.396.822 **1274**

Experimental Verification of Space-Charge and Transit-Time Reduction of Noise in Electron Beams.—C. C. Cutler & C. F. Quate. (*Phys. Rev.*, 1st Dec. 1950, Vol. 80, No. 5, pp. 875–878.) A cavity, resonant at 4·2 kMc/s, was arranged to be moved axially along a parallel electron beam from a Pierce gun. The observed variation of noise power in the cavity with distance along the beam is in fair agreement with the sinusoidal variation predicted theoretically. The greatest reduction of noise power obtained was 20 db below shot noise. That the minimum noise power is not zero is largely due to residual partition noise.

537.525.92 : 621.385 **1275**

Congruent Space-Charge Flow.—G. B. Walker. (*Proc. phys. Soc.*, 1st Dec. 1950, Vol. 63, No. 372B, pp. 1017–1027.) For irrotational flow, propositions are established regarding (*a*) rectilinear motion, (*b*) motion with constant current density along lines of flow, and (*c*) representation of lines of flow by the level lines of a harmonic function. Two cases of curvilinear flow from a unipotential cathode are discussed and shown to possess important features regarding magnification and transit time.

621.385.029.63/.64 1276

Interaction between a Travelling Electromagnetic Wave and a Beam of Electrons moving in a Cylindrical System Perpendicular to Steady Crossed Electric and Magnetic Fields.—R. Warnecke & O. Doehler. (*C. R. Acad. Sci., Paris*, 20th Nov. 1950, Vol. 231, No. 21, pp. 1132–1134.) A travelling-wave valve is described in which a steady magnetic field is produced by the passage of a current I_m through the axial element of a coaxial delay line, and a steady electric field by applying a constant voltage V_0 between the inner and outer elements of the line. Relatively high gain and high efficiency are claimed to result. Two or more waves are propagated, depending on the operating conditions. For a particular system operating at about 1·5 kMc/s, with $I_m = 500$ A and $V_0 = 1{\cdot}2$ kV, calculation gives a useful power of 100 W, efficiency of 40%, and gain of the order of 2·5 db/cm.

621.385.3 : 537.212 1277

On the Electric Field in a Single-Grid Radio Valve.—G. B. Walker. (*Proc. Instn elect. Engrs*, Part III, Jan. 1951, Vol. 98, No. 51, pp. 57–63.) The problem of calculating the e.s. field in a planar triode is solved by assuming a system of line charges at the centre of each grid wire and hence deriving a potential function which satisfies all the boundary conditions. This function is expressed in series form, but can be readily used to compute field characteristics to any required accuracy. Tables of the more important characteristics are provided.

621.385.3 : 621.315.592† : 518.4 1278

Graphical Analysis of Transistor Characteristics.—Hunter. (See 1101.)

621.385.5 : 537.212 1279

On the Electric Field in a Multigrid Radio Valve.—G. B. Walker. (*Proc. Instn elect. Engrs*, Part III, Jan. 1951, Vol. 98, No. 51, pp. 64–67.) The question of the extent to which field disturbances caused by the individual wires of a grid penetrate an adjacent grid is examined by considering interelectrode-capacitance relations. The effect of such disturbances can be measured, using an electrolyte tank, and can also be computed. Only space-charge-free fields are considered, and the electrode system is assumed either planar or concentric. It is concluded that Dow's simplified treatment (1904 of 1941) may always be used, at least as a first step, in the analysis of a valve containing two or more grids, the field in the region of one particular grid being afterwards investigated by a more rigorous triode analysis.

621.396.615.14.029.63 1280

Oscillators for Decimetre Waves with Disk-Seal Valves in Grounded-Grid Circuits.—L. Ratheiser. (*Radio Tech., Vienna*, Nov. 1950, Vol. 26, No. 11, pp. 519–524.) The special features and the use of the disk-seal triode with glass envelope and of the Telefunken metal/ceramic type are described. Advantages of the grounded-grid circuit are stressed and methods of overcoming the low inter-electrode capacitance in this case are considered. These are (*a*) 'disk-edge' feedback, in which the output and input circuits are coupled capacitively to the grid through an insulating layer; (*b*) feedback windows, which are recesses in the wall of the grid cylinder; these give stability but limit bandwidth: (*c*) additional external capacitance feedback as in the Philips Type-EC55 valve. Drawings of different assemblies are shown and their construction is discussed, particularly the methods used for obtaining good mechanical and electrical contacts.

621.396.615.141.2 1281

Some Aspects of Split-Anode Magnetron Operation.—Reich, May, Skalnik & Ungvary. (See 1120.)

621.396.615.141.2 : 537.525.92 1282

Effects of Space Charge on Frequency Characteristics of Magnetrons.—H. W. Welch, Jr. (*Proc. Inst. Radio Engrs*, Dec. 1950, Vol. 38, No. 12, pp. 1434–1449.) "Properties of the magnetron space-charge swarm which affect the propagation of electromagnetic waves are defined in terms of an effective dielectric constant. The space charge is found to be doubly refractive in nature, the velocity of propagation depending on the direction of propagation of the wave, polarization of the wave, and frequency. Effects of synchronism of the rotating space-charge swarm are discussed qualitatively. Experimental results which check parts of the theory are presented. The relationship of the space charge to the circuit is discussed in terms of the nonoscillating and the oscillating magnetron."

621.396.822 1283

Threshold Signals. [Book Review]—Lawson & Uhlenbeck. (See 1234.)

MISCELLANEOUS

621.3.012.3 1284

Reference Sheets.—(*Electronics, Annual Buyers' Guide Issue*, Mid-month June 1950, Vol. 23, No.6A, pp. R1–R40.) A collection of 21 of the graphical and tabular sheets that have appeared in recent years in *Electronics*, with a complete index of all published since April 1930. Many of those now included have been revised.

621.396.69 + 621.38] : 061.5(058.7) 1285

Alphabetical Listings of All Components, Complete Units, Allied Products, used in Electronic Equipment for All Purposes.—(See 1126.)

621.38/.39 1286

Electronic Engineering Master Index, 1949. [Book Review]—Publishers: Electronics Research Publishing Co., New York, 1950, 296 pp., $17.50. (*J. Franklin Inst.*, Dec. 1950, Vol. 250, No. 6, p. 587.) The third supplement, with coverage increased to nearly 400 periodicals, etc. Asterisks indicate Russian articles of which an English translation is available at Brookhaven National Laboratory.

621.396 1287

Einführung in die Funktechnik (Introduction to Radio Technology). [Book Review]—F. Benz. Publishers: Springer Verlag, Vienna, 4th edn, 736 pp., 78s. (*Wireless Engr*, Oct./Nov. 1950, Vol. 27, Nos. 325/326, p. 273.) "First published in 1937; it has been much enlarged and is a very comprehensive and up-to-date text-book of radio engineering."

621.396 1288

Fortschritte der Funktechnik und ihrer Grenzgebiete, Band 7/8 (Progress in Radio Technology and Related Fields. Vol. 7/8). [Book Review]—H. Richter. Publishers: Franckh'sche Verlagsbuchhandlung, Stuttgart, 1950, 387 pp., DM.60. (*Arch. elekt. Übertragung*, Aug. 1950, Vol. 4, No. 8, p. 340.) Includes surveys of the following subjects, contributed by different authors:—broadcasting receiver design, radio wave propagation, f.m., television, h.f. and industrial measurements, radio prospecting, c.r. oscillography, and magnetic-tape recording.

621.396 1289

Radio Engineering Handbook. [Book Review]—K. Henney. Publishers: McGraw-Hill, New York, 4th edn 1950, 1197 pp., $10.00. (*Electronics*, Dec. 1950, Vol. 23, No. 12, pp. 136..142.) "A good reference [book] for radio engineers."